Calibration and Measurement Guide

VectorStar™ MS464xB Series Microwave Vector Network Analyzer

MS4642B VNA, 10 MHz to 20 GHz, K Connectors

MS4644B VNA, 10 MHz to 40 GHz, K Connectors

MS4645B VNA, 10 MHz to 50 GHz, V Connectors (No longer available)

MS4647B VNA, 10 MHz to 70 GHz, V Connectors



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Chapter 1 — Calibration and Measurement Overview

1-1 General Information

This chapter discusses general calibration requirements and the benefits of different calibration types, algorithms, and routines. General calibration setup and measurement procedures are described. Technical references to Anritsu and other calibration-related articles are also presented where appropriate. Some sections provide cross-references to more detailed explanations and procedures in subsequent chapters.

Manual Scope

The purpose of this Calibration and Measurement Guide is to introduce the basic calibration and operation of the VectorStar™ MS464xB Vector Network Analyzer (VNA) System and reduce the time required to become proficient at performing basic measurements. The procedures in this manual assume a working knowledge of, and a familiarity with, vector network analyzers.

Related Documentation

The latest product information and documentation can be found on the VectorStar product web page:

https://www.anritsu.com/test-measurement/products/ms4640b-series

On this web page, you can select various tabs for more information about your instrument. Included is a "Library" tab which contains links to all the latest technical documentation related to this instrument.

Product Information, Compliance, and Safety

Refer to the *VectorStar Product Information, Compliance, and Safety* (PICS)—10100-00063 for applicable product information, compliance statements, and safety information, including links to applicable product web pages.

Contacting Anritsu

To contact Anritsu, please visit:

https://www.anritsu.com/contact-us

From here, you can select the latest sales, service and support contact information in your country or region, provide online feedback, complete a Talk to Anritsu form to get your questions answered, or obtain other services offered by Anritsu.

1-2 Calibration Kits, Verification Kits, and Components

Anritsu and other vendors provide calibration kits for a variety of algorithms and circumstances. In all cases, certain information must be provided to the VNA in order to complete the calibration, but the nature of that information varies by kit and application.

Calibration kits contain the precision components and tools required to calibrate the VectorStar MS464xB Series VNA for up to a complete 12-term error-corrected measurement. The calibration kits are available as automatic calibrator units or as manual calibration kits. A set of verification kits is also available to verify the accuracy of the calibration kits and the instrument settings.

Calibration/Verification Equipment Part Numbers and General Specifications

The following table summarizes the equipment related to calibration procedures.

Table 1-1. Calibration Equipment Listing (1 of 3)

Part Number	Name	Specifications	Connectors
Precision AutoCal Modules			
36585K-2M	Precision AutoCal™ Module, K	70 kHz to 40 GHz	K (m) to K (m)
36585K-2F			K (f) to K (f)
36585K-2MF			K (m) to K (f)
36585V-2M	Precision AutoCal Module, V	70 kHz to 70 GHz	V (m) to V (m)
36585V-2F			V (f) to V (f)
36585V-2MF			V (m) to V (f)
Standard AutoCal Module			
36581KKF	Standard AutoCal Module, K	40 MHz to 20 GHz	K (m) to K (f)
Matched Adapters for 36581KKF AutoCal Modules			
36583S	Matched Adapters Set, K	K (f) to K (f) (2 each) K (f) to K (m) (2 each)	
36583L	Matched Adapters Set, 3.5 mm	K (f) to 3.5 mm (f) (2 each) K (f) to 3.5 mm (f) (2 each)	
36583SMA	Matched Adapters Set, SMA	K (f) to SMA (m) (2 each) K (f) to SMA (f) (2 each)	
SMA Connector Manual Calibration Kits			
3650A	SMA/3.5 mm Calibration Kit	Without sliding loads	SMA/3.5 mm
3650A-1	SMA/3.5 mm Calibration Kit	With sliding loads	SMA/3.5 mm
K Connector Manual Calibration Kits			
3652A	K (2.92 mm) Calibration Kit	With pin depth gauge	K
3652A-1	K (2.92 mm) Calibration Kit	With pin depth gauge and sliding loads	K
3652A-2	K (2.92 mm) Calibration Kit	No pin depth gauge or sliding loads	K
3652A-3	K (2.92 mm) Calibration Kit	With pin depth gauge and .s1p characterization	K
3652A-4	K (2.92 mm) Calibration Kit	With .s1p characterization	K

Table 1-1. Calibration Equipment Listing (2 of 3)

Part Number	Name	Specifications	Connectors
Type N Connector Manual Calibration Kit			
3653A	Type N Calibration Kit	With fixed loads	Type N
V Connector Manual Calibration Kits			
3654D	V (1.85 mm) Calibration Kit	With pin depth gauge	V
3654D-1	V (1.85 mm) Calibration Kit	With pin depth gauge and sliding loads	V
3654D-2	V (1.85 mm) Calibration Kit	No pin depth gauge or sliding loads	V
3654D-3	V (1.85 mm) Calibration Kit	With pin depth gauge and .s1p characterization	V
3654D-4	V (1.85 mm) Calibration Kit	With .s1p characterization	V
V Connector MultiLine Calibration Kits			
3657	V (1.85 mm) Multi-Line Calibration Kit	Without shorts	V
3657-1	V (1.85 mm) Multi-Line Calibration Kit	With shorts	V
W1 Connector Calibration/Verification Kit			
3656B, 3656C, 3656C-5	W1 (1 mm) Calibration Kit	With fixed loads	W1 (1 mm)
3656B-3, 3656C-3, 3656C-6	W1 (1 mm) Calibration Kit	With .s1p characterization	W1 (1 mm)
0.8 mm Connector Calibration/Verification Kit			
3659	0.8 mm Calibration Kit	With fixed loads	0.8 mm
Verification Kits			
3661-1	3.5 mm Connector Verification Kit	NIST traceable standards with two attenuators, an airline, and a stepped impedance airline Beatty Standard. Includes the 2300-579 Performance Verification Software application and related documentation.	
3663-1	Type N Connector Verification Kit		
3668-1	K Connector Verification Kit	Used with Short-Open-Load-Thru (SOLT) with Sliding Load (Fixed Loads for Type N) using separately purchased 365x-1 Cal Kit and 36585X-2MF Series Precision AutoCal Module.	
3669B-1	V Connector Verification Kit		
Test Port Cables			
3670K50-1	Ruggedized, Semi-Rigid, DC to 40 GHz	30.5 cm (12 in)	K (f) to K (m)
3670K50-2		61.0 cm (24 in)	K (f) to K (m)
3670V50A-1	Ruggedized, Semi-Rigid, DC to 70 GHz	30.5 cm (12 in)	V (f) to V (m)
3670V50A-2		61.0 cm (24 in)	V (f) to V (m)
3670.850-1	Ruggedized, Semi-Rigid, DC to 145 GHz	10 cm (3.94 in)	0.8 mm (f) to 0.8 mm (m)

Table 1-1. Calibration Equipment Listing (3 of 3)

Part Number	Name	Specifications	Connectors
3670.850-2	Ruggedized, Semi-Rigid, DC to 145 GHz	16 cm (6.30 in)	0.8 mm (m) to 0.8 mm (f)
3671KFS50 -60	Flexible, Phase Stable, DC to 26.5 GHz	60 cm (23.5 in)	K* (f) to 3.5 mm (m)
3671KFK50 -60	Flexible, Phase Stable, DC to 40 GHz	60 cm (23.5 in)	K* (f) to K (m)
3671KFK50 -100	Flexible, Phase Stable, DC to 40 GHz	100 cm (39.3 in)	K* (f) to K (m)
3671KFKF50 -60	Flexible, Phase Stable, DC to 40 GHz	60 cm (23.5 in)	K* (f) to K (f)
3671VFV50B -60	Flexible, Phase Stable, DC to 70 GHz	60 cm (23.5 in)	V* (f) to V (m)
3671VFV50B -100	Flexible, Phase Stable, DC to 70 GHz	100 cm (39.3 in)	V* (f) to V (m)
3671W1-50 -1	Flexible, Phase Stable, DC to 110 GHz	10 cm (3.94 in)	1 mm (f) to 1 mm (m)
3671W1-50 -2	Flexible, Phase Stable, DC to 110 GHz	13 cm (5.12 in)	1 mm (f) to 1 mm (m)
3671W1-50 -3	Flexible, Phase Stable, DC to 110 GHz	16 cm (6.30 in)	1 mm (f) to 1 mm (m)
* Ruggedized style for VNA test ports. Does not fit standard K (m) or V (m) connectors.			
Universal Test Fixtures (UTF) and Right Angle Launchers			
3680-20	UTF, DC to 20 GHz	0.5 cm (min) to 10 cm (max)	3.5 mm (f) to (f)
3680K	UTF, DC to 40 GHz	0.5 cm (min) to 5 cm (max)	K (f) to K (f)
3680V	UTF, DC to 60 GHz	0.5 cm (min) to 5 cm (max)	V (f) to V (f)
36801K	Right Angle Launcher, DC to 40 GHz	1 cm (min) to 4 cm max	
36801V	Right Angle Launcher, DC to 60 GHz	0 cm (min) to 2 cm (max)	

AutoCal Automatic Calibration Modules

The auto calibration process represents both a calibration kit and an algorithm that can be used to speed up the calibration process with extremely high accuracy and a minimal number of manual steps.

The 36585X AutoCal module calibrates the VNA by a process known as “transfer calibration.” There are a number of impedance and transmission states in the module designed to be extremely stable in time and these states are carefully “characterized,” generally by the Anritsu factory but also in a customer laboratory in certain cases. When the same states are re-measured during an actual calibration, and the results compared to the characterization data, an accurate picture can be generated of the behaviors and error terms of the VNA and setup being calibrated.

A very high calibration accuracy is maintained through the use of certain principles:

- The use of many impedance and transmission states.
- The creation of very stable states that are further enhanced with a constant-temperature thermal platform inside the module.
- The use of very reliable and repeatable solid-state switching constructed to provide a great variety of state impedances (for better calibration stability) and clean transmission paths.

- The use of a very careful characterization process that can generate excellent starting data.

The resulting accuracy can exceed that obtained with a typical mechanical calibration performed in a laboratory. The AutoCal results may not be better than that of an exotic manual calibration (such as is performed during factory characterization) but will be far better than that done typically. Since it is a very high performance calibration, connector care is of the utmost importance and is discussed in detail later in this chapter.

The AutoCal consists of the calibration module itself, a separate external power supply, a control cable that runs to the VNA, and the characterization data provided on a USB memory device.

AutoCal Calibrations Available

Of the various calibration types described for the instrument, most are available with the AutoCal unit. Frequency response calibrations are omitted since the AutoCal unit would provide no benefit in these cases (since only a thru or high reflect standard is needed for these calibrations). Available calibrations include:

- Full 2-port calibration
- Reflection only (full 1-port calibration); either port or both can be specified
- 1-path 2-port (1p2p); either direction can be specified (i.e., calibrate S_{11} and S_{21} or calibrate S_{22} and S_{12})

SOLT Kits (3650x and 3654x)

These kits, all based on short-open-load-thru, require data describing all of the reflection standards (provided by the factory) be loaded into the instrument on a serial number basis. If this media (a USB key) is not available, average default coefficients are available within the VNA that may suffice for some measurements.

Typically these calibration kits are loaded using the CAL KIT/AUTOCAL utility menu (Figure 1-1) but user-defined kits can also be created using the parameters described above. If calibration kits from another manufacturer are used or to create a calibration kit, the following parameters are typically entered into one of the user-defined kits:

- Open definition (M and F typically)
- Short definition (M and F typically)
- Load definition (M and F typically)

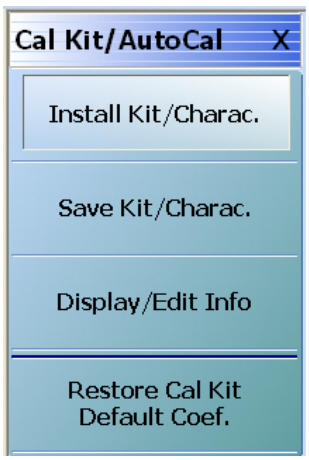


Figure 1-1. CAL KIT/AUTOCAL Utility Menu

The CAL KIT/AUTOCAL utility menu is shown here. This menu allows calibration kit details to be loaded from external files (as provided with Anritsu calibration kits), saved to a file (for user-defined cal kits), defaulted (for standard connector types), or simply displayed.

An option on the cal kit is available that provides .s1p data for all of the standards in the cal kit. A calibration using the .s1p definitions can provide enhanced accuracy over that generated using the modeled coefficients.

LRL Kits (3657/3657-1)

These airline-based kits use the LRL algorithm so much less definition of components is required. Reflects may be part of the kit but the only piece of information here is an offset length that is used to help with root selection (and is hence somewhat non-critical). The line lengths are the other parameters and they mainly serve for reference plane placement. All of these parameters must be entered manually since there are a large number of lines in the kit and usually only 2 or 3 will be used per calibration. Details on line selection and the trade-offs involved are discussed in the LRL/LRM section to follow.

Items as part of the definition:

- Line lengths (at least 2)
- Reflect offset length

Broadband SOLR/SSST kits (3656x and 3659x)

These kits, intended for use with the broadband ME7838X systems (incorporating the MS464XX), allow defined standard calibrations to 110 GHz and above. The 3656X kit is based on the 1mm connector and the 3659X kit is based on the 0.8 mm connector (for calibrations to 145 GHz). These kits combine open/short/load standards with multiple offset short standards to cover the wide frequency range with excellent performance. The Broadband calibration type is available that automatically combines the use of the standards but separate SOLT and SSST calibrations can also be merged. More information is available in Chapter 17 of this guide.

As with the SOLT kits, a USB stick is provided that contains coefficients for the kit and an option with .s1p definitions is available for the 3656X kit. If the data for the particular kit is not loaded, average coefficients will be used by default.

Offset Short Waveguide Kits (3655X)

Waveguide calibration kits based on offset short calibrations are also provided for different waveguide bands. Here two different offset-length shorts (accomplished with flush shorts and two different insert lengths), loads, and a thru must be specified. Some of the standard kits are pre-defined and user-defined kits are possible as usual. Additional pieces of information here due to line type are the cutoff frequency and dielectric constant. Items as part of the definition:

- Load definition
- Short 1 definition
- Short 2 definition
- Waveguide cutoff frequency

Microstrip/Coplanar Waveguide Kits for the UTF (36804-XXX)

For certain microstrip and coplanar waveguide measurements, the Universal Test Fixture (UT) can accommodate a range of substrate sizes and thicknesses (see 3680 brochure for more information). The 36804 series of calibration kits provide opens, shorts, loads and a variety of transmission line lengths on alumina that can be used for different calibration algorithms. User-defined kits must be generated based on the information provided with the kits.

On-Wafer Calibration Kits

A variety of calibration standard substrates or impedance standard substrates are available from other vendors that contain opens, shorts, loads, and transmission lines for on-wafer calibrations. A variety of calibration algorithms may be used depending on the application. For the defined-standards calibrations, a user-defined kit will have to be generated.

1-3 Calibration Algorithms Available

The VectorStar MS4640B Series VNAs provide for the following calibration methods:

AutoCal

- See Chapter 2, “AutoCal Calibration”

SOLT/SOLR

- Short-Open-Load-Thru (SOLT) with Fixed or Sliding Load
  - See Chapter 3, “SOLT/SOLR Calibration”
- Short-Open-Load-Reciprocal (SOLR)
  - See Chapter 3, “SOLT/SOLR Calibration”

SSLT

- Offset-Short (SSLT)
- See Chapter 4, “Offset Short (SSLT) Calibration”

SSST

- Triple-Offset-Short (SSST)
- See Chapter 5, “Triple Offset Short (SSST) and Multiple Offset Short (mSSS) Calibrations”

Broadband SOLT\R-SSST\R

- Broadband Cal (SOLT\R-SSST\R)—Appears only with VNA in Broadband Mode
- See Section “Broadband Cal” in Chapter 17, “Broadband/mmWave Measurements (Option 7, Option 8x)”

LRL/LRM/ALRM

- Line-Reflect-Line (LRL) / Line-Reflect-Match (LRM)
- Advanced-LRM (A-LRM™) for improved on-wafer calibrations
  - See Chapter 6, “LRL/LRM Calibration”

mTRL

- Multiline TRL
  - See Chapter 7, “mTRL Calibration”

Reciprocal Measurements or Unknown Thru

- Reciprocal or Unknown Thru Method
  - See Chapter 16, “Reciprocal Measurements”

1-4 About Calibration

The most important central concept to making good VNA S-parameter measurements is the calibration of the instrument. The background on calibration mathematics and theory will only be lightly covered in this section; more information is available in Anritsu Application Notes and in the reference literature. While the VNA is a highly linear receiver and has sufficient spectral purity in its sources to make good measurements, there are a number of imperfections that limit measurements done without calibrations:

Match

Because the VNA is such a broadband instrument, the raw match can be good but not excellent. Even a 20 dB match, which is physically very good, can lead to errors of greater than 1 dB. By correcting for this raw match, the potential error can be greatly reduced.

Directivity

A key component of a VNA is a directional coupler that allows separation of the signal incident on the DUT from the reflected back from the DUT. While the couplers used in the VNA are of very high quality, there is an amount of coupled signal even when a perfect termination is connected. This is related to directivity and can impact measurements of very small reflection coefficients.

Frequency Response

While the internal frequency response of the VNA could be calibrated at the factory, any cables connected externally will have some frequency response that must be calibrated out for high quality measurements.

The calibration is a method of correcting for these and other defects. There are an enormous number of possible calibration algorithms and many of them are implemented within the MS464xB Series VNAs. The choice between them is largely determined by the media used, the calibration standards available, and the desired accuracy/effort trade-off. Some of the choices to be made are:

- **Calibration Type**

  Which ports are being corrected and to what level are they being corrected?

- **Calibration Algorithm**

  How is the correction being accomplished?

Calibration Types

Table 1-2. MS464xB Series VNA Calibration Types

VNA Mode	Type	Parameters Calibrated	Uses
2-port VNA Mode	Full 2-port	S_{11}, S_{12}, S_{21}, and S_{22}	Most complete calibration
	Full 1-port	S_{11} or S_{22}	Reflection calibration only
	1-path 2-port	S_{11} and S_{21} or S_{22} and S_{12}	1-port reflection plus simple transmission (faster, lower transmission accuracy unless DUT very lossy)
	Frequency Response	Any one parameter (or pairs of symmetric parameters such as S_{12} and S_{21})	Normalization only. Fast, lower accuracy
4-port VNA Mode	When in 4-port mode, additional calibration types of 4-port and 3-port become available in the CALIBRATION menus with an associated expansion of S-Parameters, User-Defined, and Mixed-Mode options in the RESPONSE menus.		See Chapter 24, “Multiport Measurements”

Calibration Algorithm Descriptions

The use of acronyms for the various calibration algorithms is often inconsistent. The following table presents calibration algorithm acronyms as used in Anritsu documentation.

Table 1-3. Calibration Algorithms

Calibration Algorithm	Description	Advantages	Disadvantages
SOLT (Short-Open-Load-Thru)	Common coaxially	Simple, redundant standards; not band-limited	Requires very well-defined standards, poor on-wafer, lower accuracy at high frequencies.
SSLT (Short-Short-Load-Thru, also called Offset Short), shorts with different offset lengths	Common in waveguide	Same as SOLT	Same as SOLT and band-limited.
SSST (Short-Short-Short-Thru, also called Triple Offset Short), all shorts with different offset lengths	Common in waveguide or high frequency coax	Same as SOLT but better accuracy at high frequencies	Requires very well-defined stds, poor on-wafer, band-limited.
SOLR/SSLR/SSSR, like above but with “Reciprocal” instead of “Thru”	Like the above but when a good thru is not available	Does not require well-defined thru	Some accuracy degradation but slightly less definition, other disadvantages of parent cal.
Broadband Calibration SOLT/SOLR-SSST/SSSR	Combines SOLT/R and SSST/R in a single calibration (SOL below an entered frequency and SSS above it).	Saves time and connections when doing a broadband cal (no need to merge). Simple, redundant standards. Not as band-limited as constituents.	Requires very well-defined standards. Generally poor on-wafer.
LRL (Line-Reflect-Line; also called TRL)	High performance coax, waveguide, or on-wafer	Highest accuracy, minimal standard definition	Requires very good transmission lines, less redundancy so more care is required, band-limited
ALRM, advanced line-reflect-match, simplifies to LRM (Line-Reflect-Match) or TRM (Thru-Reflect-Match)	Relatively high performance	High accuracy, only one line length so easier to fixture/on-wafer, not band-limited usually	Requires load definition. Reflect standard setup may require care depending on load model used.
mTRL (Multiline TRL)	High performance coax, waveguide, or on-wafer	Highest accuracy, minimal standard definition	Requires very good transmission lines, uses up to 10 lines and all lines are used at all frequencies in a least-squares sense so it more seamlessly deals with band limitation. Usually requires more lines than LRL for best performance.

For those calibrations with defined reflection standards (the first four lines of the table), there are a number of variations on how the reflection coefficients of those standards are defined:

- Non serialized models describing the reflection coefficients (used on some lower cost calibration kits).
- Serialized simple polynomial models describing the reflection coefficients (standard on Anritsu kits for many years).
- Serialized simple polynomial models for the open and short supplemented by better raw match sliding loads for the load standard (Option 1 available on most Anritsu calibration kits).
- Serialized .s1p files describing the reflection coefficients [available as -3 or -4 options on certain Anritsu calibration kits (and are listed in Table 1-1 on page 1-2)]. These descriptions can provide improved performance since they capture all of the frequency response of the standard.

Table 1-4 relates the Calibration Types to the Calibration Algorithms.

Table 1-4. Calibration Types and Calibration Algorithms

VNA Mode	Type	SOLT	SSLT	SSST	SOLR/SSLR /SSSR	Broadband SOLT/R-SS ST/R	LRL	ALRM	mTRL	mSSS
2-port Mode	Full 2-port	YES	YES	YES	YES	YES	YES	YES	YES	YES
	Full 1-port	YES	YES	YES	Can be selected for these types, but the reciprocal nature is not used and will function like the base calibration.	YES	–	–	–	YES
	1-path 2-port	YES	YES	YES		–	–	–	–	YES
	Frequency Response	YES	YES	YES		–	–	–	–	–
4-port Mode	The calibration types and modes change when the VNA is in multiport (4-port) mode. See Chapter 24, “Multiport Measurements” for a discussion of multiport calibration methods and types.									

1-5 Calibration Setup

Before proceeding to the calibrations and some of the alternatives available, there are certain instrument setup issues that must be discussed first since they will affect the performance of all calibrations. In almost all cases, the current VNA settings will be used during the calibration so setting up the VNA as desired beforehand will help.

Frequency Start, Stop, and Number of Points

The Start Frequency, Stop Frequency, and number of points should be decided and set in the VNA before performing a calibration. Segmented sweeps should also be set up in advance if a more custom frequency list is desired.

Harmonic Sweep Setup

For certain applications using low pass time domain transforms (requires Option 2; see Chapter 14 of this guide), there is a specific type of frequency list that is required called a 'harmonic sweep' and there is a setup tool under the Calibration menu to help with this. This frequency setup, where the start frequency is equal to the step frequency, allows higher resolution time domain impulse response plots and step response plots without using interpolation. The tool dialog is shown below and allows one to choose the dominant parameters among start frequency, stop frequency and number of points. Once entered and the calculation performed, the non-dominant parameter (and then the other parameters as necessary) will be adjusted to create the appropriate harmonic sweep. One can then proceed with the calibration as usual.

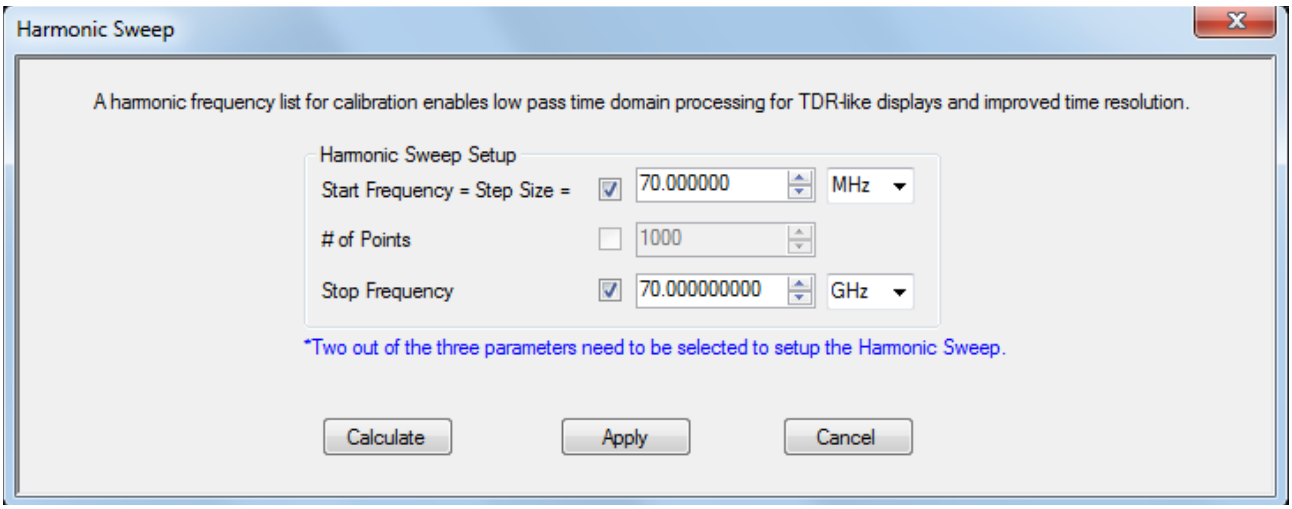


Figure 1-2. HARMONIC SWEEP Setup Dialog

Example 1:

One selects start and stop frequencies to be dominant and enters a 70 MHz start and 70 GHz stop. The system will select 1000 points and this creates a list of 70 MHz, 140 MHz,...70 GHz.

Example 2:

One selects start and stop frequencies to be dominant and enters a 27 MHz start and 20 GHz stop frequency. Since the stop frequency is not an integer multiple of the start, these constraints cannot hold. The system will assign super-dominance to the stop frequency and try to get as close as possible on start. The result would be:

- Start 27.100271 MHz
- Number of points 738
- Stop 20 GHz

There are cases when both start and stop will move slightly due to frequency resolution limits and the requirement for an integer number of points.

In other dominance cases, the frequency selections will win out over number of points. If another priority is desired, the parameters can always be entered manually on the Frequency menu.

Note	If a calibration is applied at the time the setup is run or a different sweep type is in use (e.g., segmented sweep), those will have to change before the harmonic sweep can be applied and a warning dialog will appear to allow the user to choose if that should happen.

IF Bandwidth, Averaging and Power

These parameters control the digital filtering and post-processing that determine the effective noise floor, the amount of trace noise and, in some special cases, the immunity to interfering signals. The trade-off for improved noise performance is slower sweep speed.

- **IF Bandwidth (IFBW)**

  Settings of 1 Hz to 1 MHz are allowed with the RMS trace noise ranging from <1 mdB at the low end to a few hundred mdB at the high end (for high level signals, more for lower level signals). Sweep time will be roughly proportional with the reciprocal of IFBW once below 100 kHz IFBW.

- **Point-by-Point vs. Sweep-by-Sweep Averaging**
  - Point-by-Point averaging incurs additional measurements at each given frequency point and will increase sweep time roughly proportionally. Because the additional measurements are taken at once, the effect is similar to proportional change in IFBW.
  - Sweep-by-sweep averaging acquires additional measurements on subsequent sweeps and is better at removing lower frequency variations than point-by-point averaging or IFBW reduction. Sweep-by-sweep averaging is a rolling average so the time to fully stabilize from a sudden DUT change is roughly proportional to the average count.

- **Power**

  Port power is somewhat less critical because of the excellent linearity of the MS464xB Series VNA receivers, but any step attenuator settings must be selected before the calibration. Because changing the step attenuator settings alters the RF match in the measurement paths as well as insertion loss, changing them will invalidate the calibration.

1-6 Types of Calibrations

The types of calibrations are defined by what ports are involved and what level of correction is accomplished (see Table 1-2, “MS464xB Series VNA Calibration Types” on page 1-8).

Full 4-Port

This is the most commonly used and most complete calibration involving four ports. All sixteen S-parameters (S11, S12, S21, S22, S13, S23, S33, S31, S32, S14, S24, S34, S41, S42, S43, and S44,) are fully corrected.

Full 2-Port

This is the most commonly used and most complete calibration involving two ports. All four S-parameters (S_{11}, S_{12}, S_{21}, and S_{22}) are fully corrected.

Full 1-Port

A single reflection parameter is fully corrected in this case (S_{11} or S_{22}). Both ports can be covered but only reflection measurements will be corrected. This calibration type is useful for reflection-only measurements including the possibility of doing two reflection-only measurements at the same time.

1-Path 2-Port (Forward or Reverse)

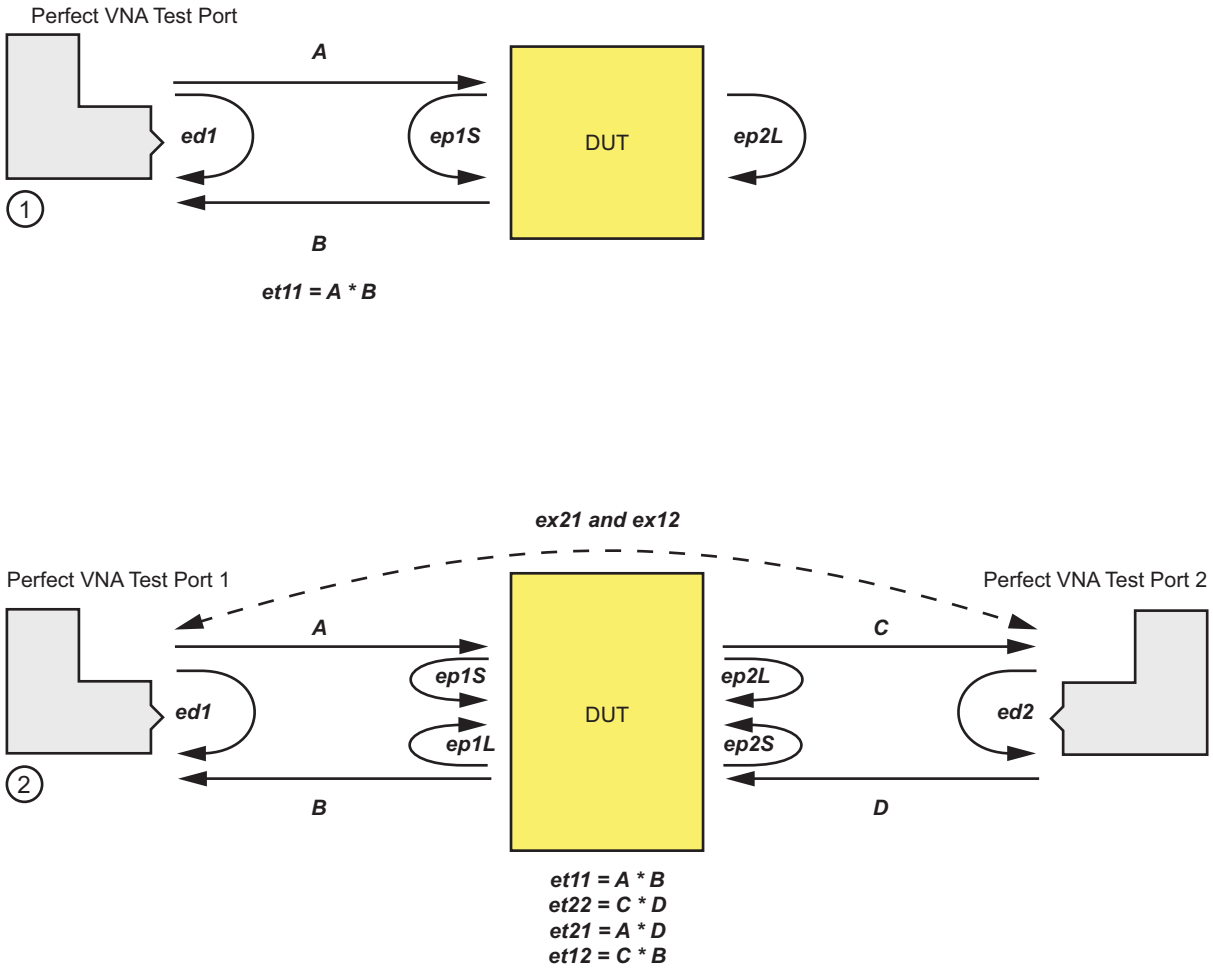
In this case, reflection measurements on one port are corrected and one transmission path is partially corrected (load match is not). Here forward means S_{11} and S_{21} are covered while reverse means S_{12} and S_{22} are covered. This technique may be used when speed is at a premium, only two S-parameters are needed and either the accuracy requirements on the transmission parameter are low or the DUT is very lossy (@ > 10–20 dB insertion loss).

Frequency Response (Transmission and Reflection Frequency Response)

This calibration is essentially a normalization and partially corrects one parameter (although two can be covered within the cal menus). Only the frequency response, or tracking slope, of the parameter is corrected so directivity and match behaviors are not taken into account. This technique is valuable when accuracy requirements are not at a premium and a quick measurement is all that is needed.

Each of these calibrations has an associated error model that describes what is being corrected. The error coefficients used fall into several categories that roughly describe the physical effect that they are responsible for correcting. To establish a context for these error terms, consider the usual model where all of the VNA/setup errors are lumped into error boxes (that act like S-parameters) between a perfect VNA and the DUT reference planes (see Figure 1-3).

Note	The calibration types and modes change when the VNA is in multiport (4-port) mode. See Chapter 24, “Multiport Measurements” for a discussion of multiport calibration methods and types.



1. At top, a classic 1-port error model.	2. At bottom, a classic 2-port error model.

Figure 1-3. Classic 1- and 2-Port Error Models

Two slightly different error models are used: one where each port is considered to be driving separately (so one can clearly delineate source match from load match) and one where both ports are present and no driving distinction is made (requiring some preprocessing to take care of source match-load match differences).

- **Directivity**

  Directivity (ed1 and ed2) describes the finite directivity of the bridges or directional couplers in the system. Partially includes some internal mismatch mechanisms that contribute to effective directivity.

- **Source Match**

  Source match (ep1S and ep2S) describes the return loss of a driving port.

- **Load Match**

  Load match (ep1L and ep2L) describes the return loss of a terminating port. In the 8-term error models used as a basis for the LRL/LRM and other calibration families, this is treated as the same as source match but the incoming data is precorrected to take into account the (measured) difference in match between driving and terminating states.

- **Reflection Tracking**

  Reflection tracking (et11 and et22) describes the frequency response of a reflect measurement including loss behaviors due to the couplers, transmission lines, converters, and other components.

- **Transmission Tracking**

  Transmission tracking (et12 and et21) is the same as above but for the transmission paths. The tracking terms are not entirely independent and this fact is used in some of the calibration algorithms.

- **Isolation**

  Isolation (ex12 and ex21) takes into account certain types of internal (non-DUT dependent) leakages that may be present in hardware. It is largely present for legacy reasons and is rarely used in practice since this type of leakage is typically very small in modern VNAs.

1-7 Line Types (Transmission Media)

Part of the calibration definition is the selection of line type. The main purpose of this is to assign a dispersion characteristic that will be needed later. Dispersion is the dependence of the phase velocity on the line with frequency. Media such as coax and coplanar waveguides are largely dispersion-free; that is, we can define phase velocity by a single number:

$$v_{ph} = \frac{c}{\sqrt{\varepsilon_r}}$$
$$= \textit{Phase velocity for coaxial and non-dispersive media}$$

Equation 1-1.

Where:

- c is the speed of light in a vacuum (~2.9978108 m/s) and
- E_r is the relative permittivity of the medium involved.

Coaxial cable has its own selection since it is intrinsic to the instrument while other non-dispersive media can be selected separately.

One type of dispersive media is regular waveguide. The phase velocity here is defined by:

$$v_{ph} = \frac{c}{\sqrt{\varepsilon_r} \times \sqrt{1 - \left(\frac{f_c}{f}\right)}}$$
$$= \frac{c}{\sqrt{\varepsilon_r - \left(\frac{f_{c0}}{f}\right)^2}}$$
$$= \textit{phase velocity for waveguide}$$

Equation 1-2.

Where:

- E_r is the dielectric constant
- f_c is the cutoff frequency of the waveguide (with dielectric) and
- f_{c0} is the cutoff frequency of the waveguide in a vacuum (which is what is entered).

The system will compute the required values. This information is needed for computing distances when in time domain and when adjusting reference planes.

Microstrip lines are another example of dispersive media that can be selected. Here the dimensions of the line together with the dielectric material determine the phase velocity behavior. An intermediate quantity, called the effective dielectric constant ($\mathcal{E}_{r,eff}$), is used and a suggested value computed by the VNA but this value can be overridden. At low frequencies, the structure can be considered non-dispersive (like coax) with a phase velocity given by:

$$v_{ph} = \frac{c}{\sqrt{\varepsilon_{r,\,eff}}}$$

$$= \textit{low frequency limit}$$

Equation 1-3.

At higher frequencies when additional mode behavior becomes important, dispersion must be handled. The dielectric constants (media-based and effective) together with a transition frequency f_t are used to compute this effect which is heavily dependent on the dielectric thickness:

$$v_{ph} = \frac{c}{\sqrt{\dfrac{\varepsilon_{r,\,eff} + \varepsilon_r \cdot \left(\dfrac{f}{f_t}\right)^2}{1 + \left(\dfrac{f}{f_t}\right)^2}}} \qquad \textit{where} \qquad f_t = \frac{Z_0 \varepsilon_0 c^2 \sqrt{\varepsilon_r}}{2t\sqrt{\varepsilon_{r,\,eff}}}$$

Equation 1-4.

Where:

- Z_0 is the characteristic impedance of the microstrip line
- t is the dielectric thickness

1-8 Preparing a Sliding Load (Termination) Calibration Kit

Sliding terminations (loads) are the traditional Z_0 calibration-reference devices for vector network analyzer calibration. When correctly used and perfectly aligned, they can be more accurate than precision fixed loads. However, sliding terminations have a low frequency limit dependent on connector type and must be used with a fixed load for full frequency-range coverage. The connector frequency limits are:

- 2 GHz for 3.5 mm Connector sliding loads
- 2 GHz for K Connector sliding loads
- 4 GHz for V Connector sliding loads

Sliding terminations consist of a connector, a long section of precision transmission line, and a microwave load that is movable within the transmission line. Pin depth is the relationship between the interface positions of the outer and center conductors and is the most critical parameter that you can control in a sliding termination. An example of its criticality is that an incorrect pin depth of 0.001 inch can cause a reflection return loss of 44 dB. Since you are usually calibrating to accurately measure a greater than 40 dB return loss, correct pin depth is essential.

Since setting an accurate pin depth is so important, this discussion centers on describing how to set the pin depth for male and female sliding terminations. Calibration with the sliding termination is essentially the same as described below for the broadband load.

The procedure below uses the Model 3652 Calibration Kit and its 17KF50 and 17K50 Sliding Terminations. Calibration is similar for the Model 3650 SMA/3.5mm, and Model 3654 V connector kits.

Procedure

1. Remove the **Pin Depth Gauge** from the kit, place it on the bench top.

Note	The gauge is convertible between male and female. The following procedure describes the zeroing process for the female fitting. The procedure for the male fitting begins with Step 16 below.

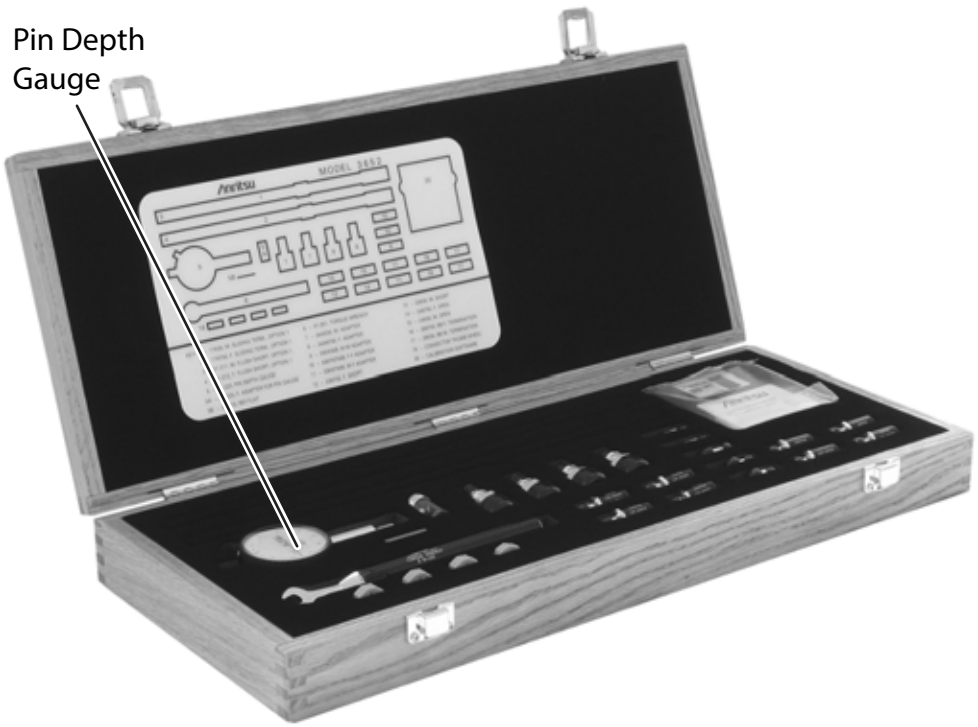


Figure 1-4. Calibration Kit

2. Push the outer locking ring towards the gauge to expose the center pin.



Figure 1-5. Exposing the Center Pin

3. Take the **01-210 Reference Flat (Ref Flat)** from the kit.

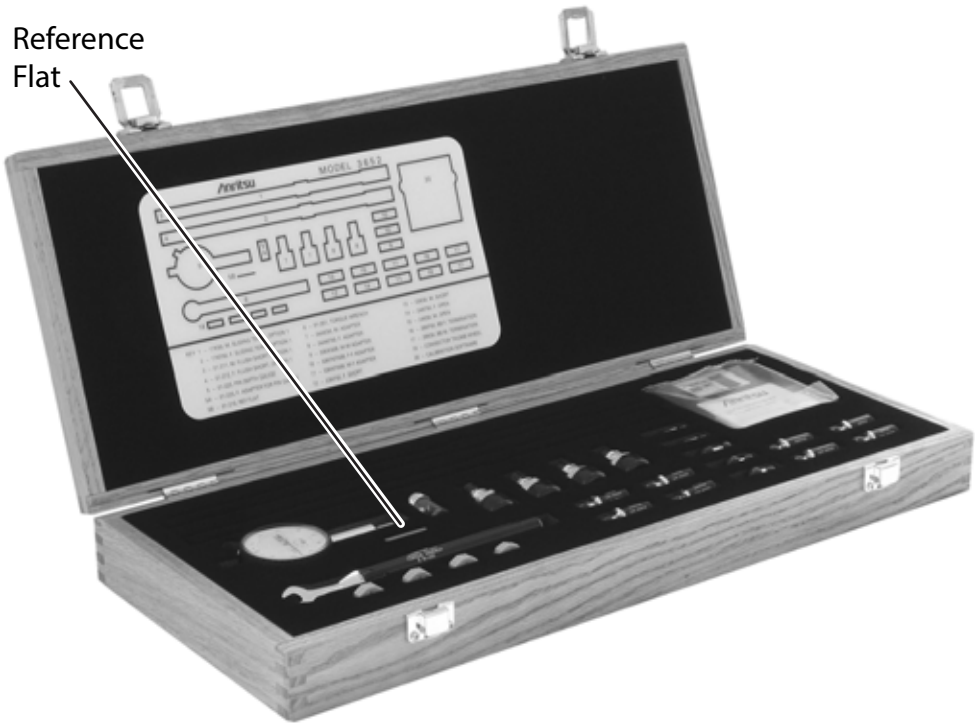


Figure 1-6. Reference Flat in Case

4. While holding the gauge as shown, press the **Ref Flat** firmly against the end of the exposed center pin.



Figure 1-7. Setting Pointer to “0”

5. While pressing the **Ref Flat** against the center pin, check that the pointer aligns with the “**0**” mark. If it does not, loosen the bezel lock screw and rotate the bezel to align the pointer with the “**0**” mark. Tighten the **bezel lock screw**.

Note	Gently rock the Ref Flat against the center pin to ensure that it is fully depressed and you have accurately set the gauge for zero.



Figure 1-8. Dial Bezel and Pointer at “0”

6. Remove the **Sliding Termination** with the female-connector (17KF50, for this example) from the kit, and slide the load all the way toward the end closest to the connector.



Figure 1-9. F-F Sliding Terminator, Towards End Closest to Connector

Caution	The center pin of the sliding load is floating. The load should be sighted from the connector end and rotated to allow gravity to pull the center pin such that it is concentric with the outer housing. Note the orientation of the sliding load assembly and ensure this orientation is preserved when making connections to the sliding load to avoid damaging the center pin.

7. With either hand, pick up the sliding termination near its connector end.



Figure 1-10. Holding F-F Sliding Terminator

8. Cup the **Sliding Termination** in your palm, and support the barrel between your body and crooked elbow.



Figure 1-11. Holding F-F Sliding Terminator, Ready to Measure

9. Remove the **Flush Short** by holding its body and unscrewing its connector.



Figure 1-12. F-F Sliding Terminator, Removing Flush Short

10. Ensure that the orientation of the sliding load is such that the center pin is concentric with the outer housing and install the gauge onto the end of the sliding termination.



Figure 1-13. F-F Sliding Terminator, Installing Gauge

11. If the factory-set **COARSE SET** adjustment has not moved, the inner dial on the gauge will read "**0**." If it does not, perform the **Coarse Set Adjustment** in Step 15 below.



Figure 1-14. Gauge Reading, Outer Dial Reading, Inner Dial Reading

12. Place the sliding termination, with the gauge attached, on the bench top.



Figure 1-15. F-F Sliding Terminator with Attached Dial Gauge

13. Loosen the **FINE LOCK** ring and turn the **FINE ADJ** ring to position the gauge pointer 0–3 small divisions on the "**–**" side of zero. The Sliding Load Pin Depth specification is from 0.000 inches to –0.0003 inches (0.000 in to –0.0003 in).

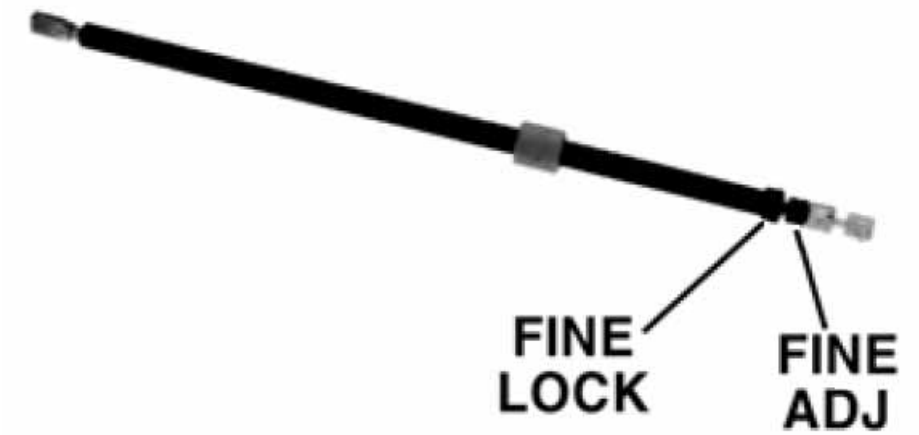


Figure 1-16. Using the Fine Lock Ring and Fine Lock Adjustment

14. Turn the **FINE LOCK** ring clockwise to both tighten the adjustment and place the pointer exactly to "0." The Sliding Termination is now ready to use.

Note	Insure that the inner dial reads "0." The following step is not normally necessary and it needs to be performed only if the adjustment has changed since it was set at the factory.



Figure 1-17. Fine Lock Adjustment with Gauge Inner Dial Reading “0”

15. With the **01-211 Flush Short** installed, loosen the **COARSE LOCK** and gently push the **COARSE SET** adjustment rod in as far as it will go. This coarsely sets the center conductor to be flush against the attached short. Return to Step 2 above.

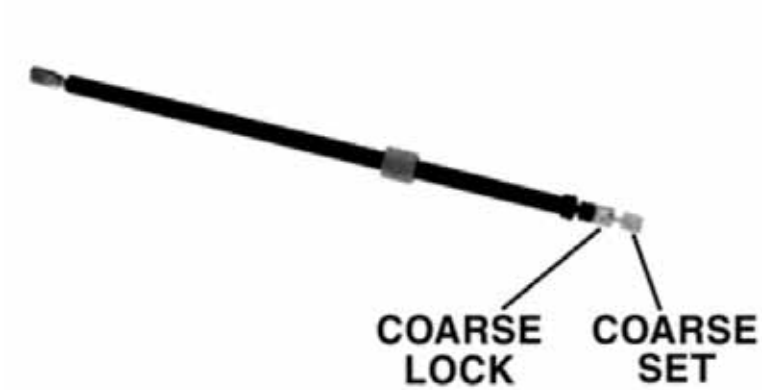


Figure 1-18. M-M Sliding Terminator

16. The procedure for adjusting the male-connector sliding termination is essentially the same as that described above. The only difference is that you must install the female adapter on the end of the gauge shaft, over the center conductor. To install this adapter, proceed as follows:

a. Zero-set the gauge as described in Step 2 through Step 5 above.

b. Push the outer locking ring back toward the gauge and turn it clockwise onto the exposed threads.

c. Loosen the lock ring one turn in a counterclockwise direction.



Figure 1-19. Preparing the Gauge for the Female Adapter

17. Remove the **01-223 Female Adapter** ("F ADAPTER FOR PIN GAUGE") from the kit.

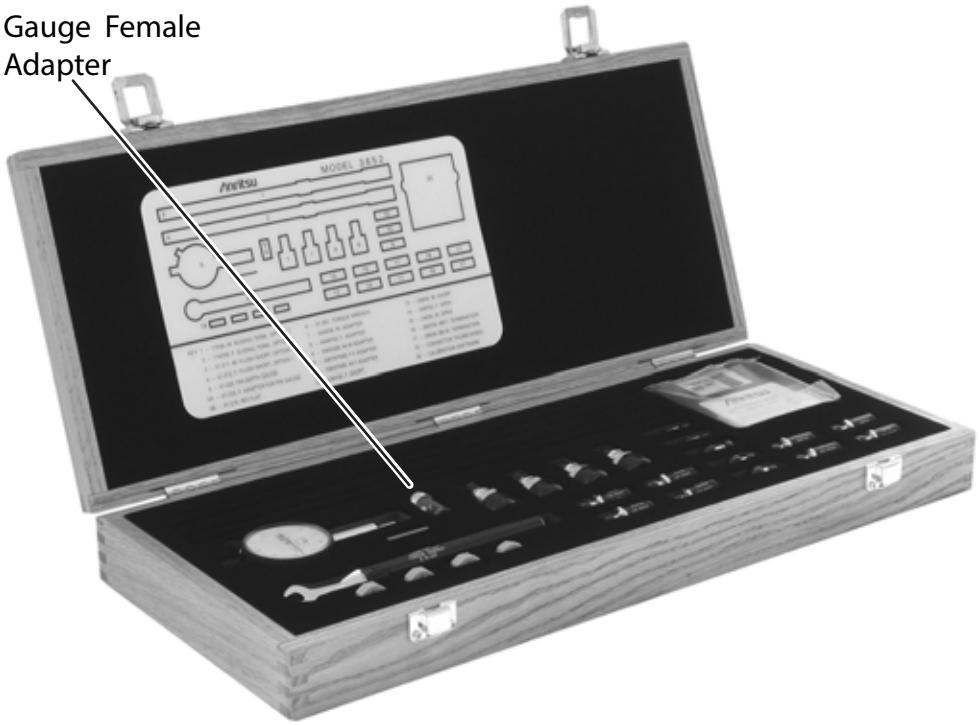


Figure 1-20. Female Adapter Location in Case

18. Install the female adapter over the center pin and screw it into the locking ring, and tighten the outer ring until it is snug against the housing.



Figure 1-21. Installing the Female Adapter on the Gauge

19. Inspect the end of the adapter, you should see no more than two exposed threads. If so, repeat Step 7 through Step 10 above.

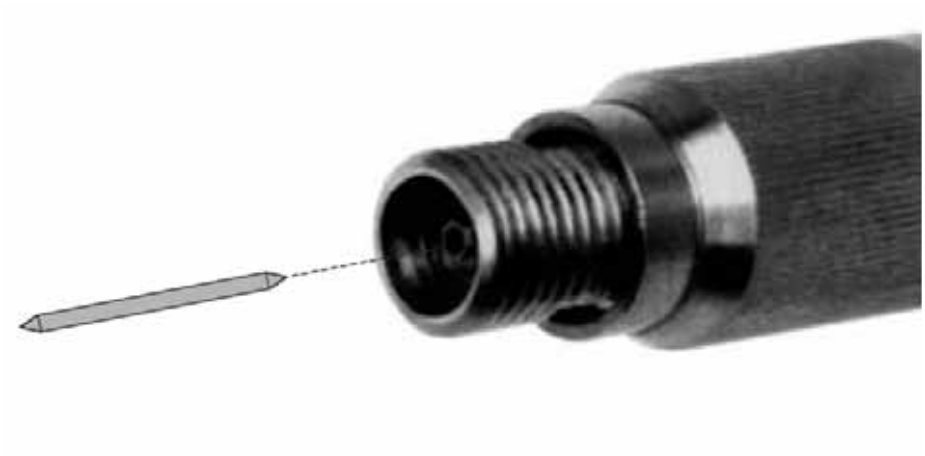


Figure 1-22. Inspecting the End of the Female Adapter

20. Insert the **Female Alignment Pin—7862** as shown in Figure 1-22 to aid in center-pin alignment with the pin-depth gauge.

21. Connect the gauge to the sliding termination and zero set the center pin using the **FINE ADJ** as previously described in Step 2 through Step 5 above.

1-9 Connector Care

This section describes proper use and care of coaxial RF connectors.

Note	The components in Anritsu calibration kits are of the highest quality and accuracy. All components are NMI (National Metrology Institute) and NIST (National Institute of Standards and Technology) traceable. Mechanical shock can damage them. Handle with care.

Common Causes of Connector Failure

Following are common causes of connector-related instrument failures and effects on measurement accuracy and repeatability.

- Connectors contaminated with material such as metal debris and dust particles will increase the risk of damaging connector. A dusty connector affects the measurement accuracy due to lack of repeatability.
- Worn or damaged threads can damage mating connectors.
- Over torquing connectors will introduce excessive stress to connector interface. Under torquing will increase repeatability uncertainty.
- Improper pin depth of the center conductor can permanently damage the mating connector.
- ESD—A connector is a perfect conductor for electrostatic discharge (ESD). Excessive charge accumulation will damage expensive and sensitive electronic devices residing in the instrument.

Coaxial Connector Care

Most coax connectors are assembled into a system and forgotten, but some, especially on test equipment are used almost continuously. The care and cleaning of these connectors is critical to maintain accurate and reliable performance. Good connector performance can be achieved with the following:

- Periodic visual inspection
- Routine cleaning
- Proper connection and disconnection techniques using a torque wrench
- Appropriate gauging techniques

Visual Inspection

Connectors contaminated with material such as metal debris and dust particles will increase the risk of damaging the connector. A dusty connector affects the measurement accuracy due to lack of repeatability. Careful visual inspection with a magnification aid and lighting should be performed at least once per day before connecting. A “good” connector may get damaged if it is mated with a “bad” one.

Any connector with the following defects should be repaired or discarded:

- Plating concerns that include bubbles, blisters, and deep scratches showing bare metal on the mating surfaces.
- Center conductors with bent, broken or damaged contacts.

Magnification

Use a device with a magnification range of 2X to 10X.

Table 1-5. Magnification for Connector Inspection

Connector Type	Minimum Magnification for Inspection
Type N	2X
3.5 mm	7X
2.4 mm	
K (2.92 mm)	
V (1.85 mm)	
W (1.0 mm)	10X
0.8 mm	

Outer Conductor and Thread Inspection

When connector threads are clean and free from defects, coupling nuts should move freely. Inspect for defects such as:

- Dirt, dust, metal particles and oil
- Damaged threads (distorted, dented, or crushed)
- Dents or raised edges on the mating plane

Note	Connectors may lose some gloss over time due to normal usage. Light scratches, marks, and other cosmetic imperfections found on the mating plane surfaces are signs of normal wear and should not be considered cause for replacement. Regular cleaning and proper connection techniques will minimize wear on the plating due to abrasion from debris.

Pin Depth

Pin depth dimensions are critical for the mechanical integrity, non-destructive mating and electrical performance of coaxial RF connectors. Before mating RF connectors, the pin depth of each device should be measured. This chapter provides instructions for properly measuring pin depth using the pin depth gauge provided in each applicable Anritsu calibration kit.

Mechanical gauging of coaxial connectors will detect and prevent the following problems:

Positive Pin Depth

- Positive pin depth or pin protrusion can result in buckling of the fingers of the female center conductor or damage to the internal structure of a device due to the axial forces generated.

Negative pin depth

- Excessive negative pin depth or pin recession can result in poor return loss, unreliable connections, and could even cause breakdown under peak power conditions.

A connector should be checked before it is used at a minimum of once per day when in use. If the connector is to be used on another item of equipment, the connector on the equipment to be tested should also be gauged.

Pin depth gauges should be checked for cleanliness before they are used at a minimum of once per month. The connector cleaning procedures in Section 1-10 can be used to clean the pin depth gauges.

Pin Depth Defined

Pin depth is the distance between the outer conductor reference plane and the plane that intersects the shoulder of the male or female center conductor.

- For a female connector, it is the plane that intersects the shoulder (face) of the female center conductor.
- For a male connector, it is the plane that intersects the shoulder of the male center conductor.

The pin depth definition for 3.5mm, K, and V female and male connectors is illustrated in Figure 1-23.

The pin depth definition for Type N female and male connectors is illustrated in Figure 1-24.

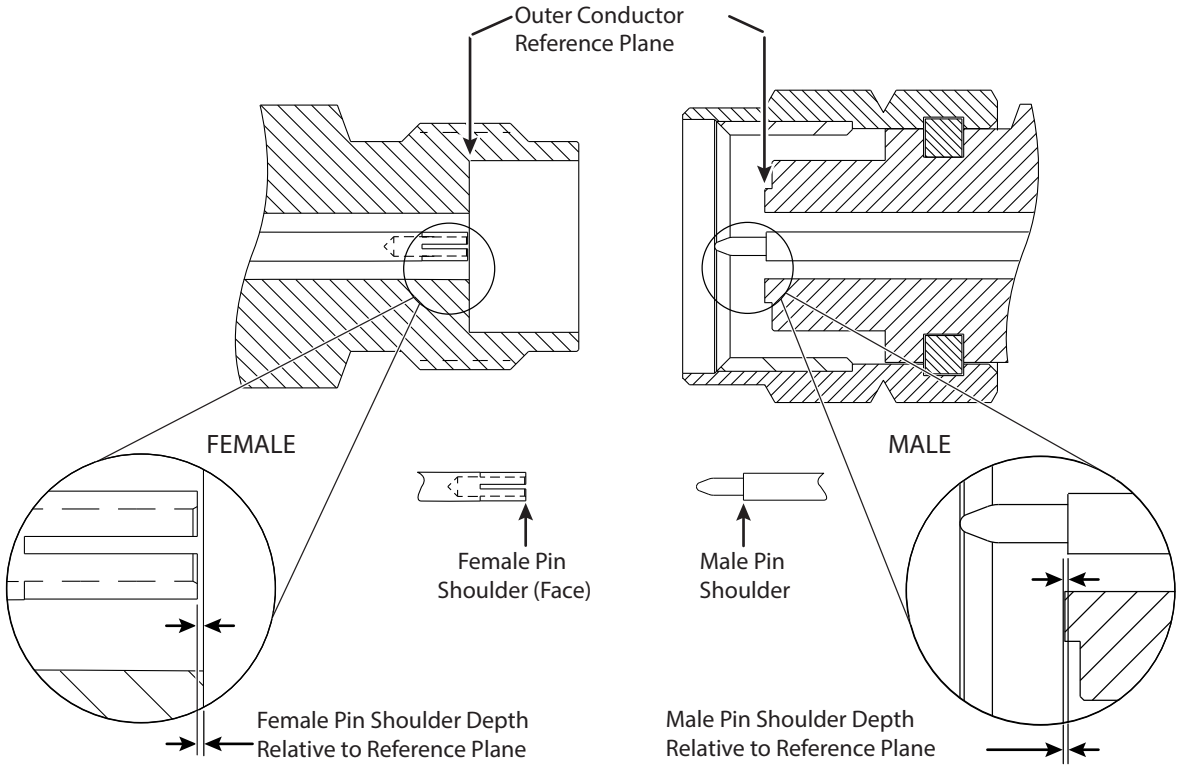


Figure 1-23. 3.5 mm, K, and V Connector Pin Depth Definition

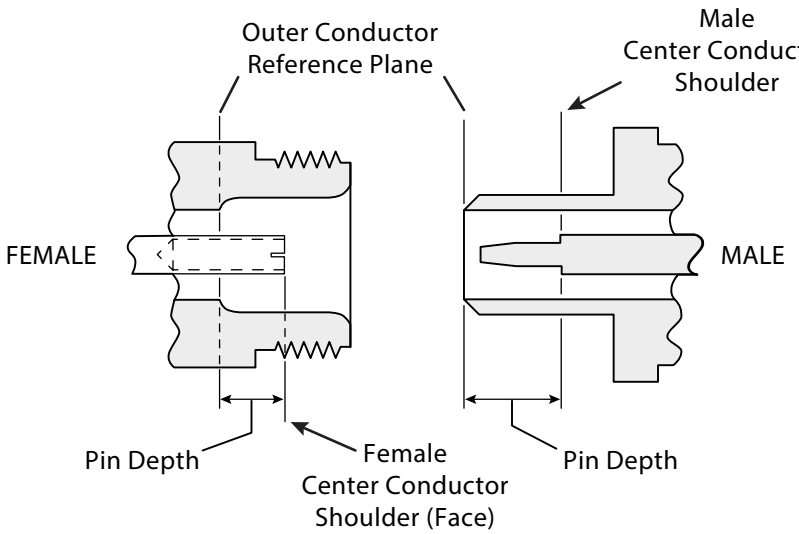


Figure 1-24. Type N Connector Pin Depth Definition

Pin Depth Limits

Pin depth is not a performance specification but it provides an indication of the mechanical condition of the calibration kit components. By verifying whether the pin depth is within the expected range of limits, one can determine if the component is in good mechanical condition. Table 1-6 lists the range of the observable limits of each calibration kit component as shown on the pin depth gauge that is supplied with the calibration kit.

Note

Standards and government committees have created pin depth limits on the most common connectors to ensure the ability of mating with connectors of the same type. Components in the calibration kit might have a narrower acceptable range of pin depth limits due to electrical performance optimization.

Connectors on devices that mate with connectors of calibration kit components may not have the proper depth. They must be measured before mating to ensure suitability and to avoid connector damage to the components.

Anritsu Pin Depth Gauge accuracy specification is ±0.0002 in.

Table 1-6. Pin Depth Limits (1 of 2)

Connector Type/Device Type	Observed Pin Depth Limits (Inches)	
	Male	Female
GPC-7		
Terminations/Loads	–0.0002 to –0.0007 (GPC-7 has no gender)	
Shorts	Not applicable due to design	
Opens	–0.0005 to –0.0010	
Type N		
Note: For Type N connectors, the pin depth gauge (P/N 01-224) is zeroed on the +0.207 or –0.207 inch reference block depending on the gender of the N connector, so the gauge reading is always +0.000 to –0.002 inches for the 34 Series adapter, and +0.000 to –0.001 inches for Loads/Opens/Shorts.		
34 Series Adapter (N to GPC-7)	Type N: –0.207 to –0.209 GPC-7: –0.0002 to –0.0007	Type N: +0.205 to +0.207 GPC-7: –0.0002 to –0.0007
Terminations/Loads	–0.207 to –0.208	+0.206 to +0.207
Shorts	–0.207 to –0.208	+0.206 to +0.207
Opens	–0.207 to –0.208	+0.206 to +0.207
3.5 mm		
33 Series Adapter	–0.0002 to –0.0010	–0.0002 to –0.0010
34 Series Adapter (3.5 mm to GPC-7)[a]	–0.0002 to –0.0007 (3.5 mm and GPC-7)	–0.0002 to –0.0007 (3.5 mm and GPC-7)
Terminations/Loads	–0.0002 to –0.0007	–0.0002 to –0.0007
Shorts	–0.0002 to –0.0007	–0.0002 to –0.0007
Opens	–0.0002 to –0.0007	–0.0002 to –0.0007
K Connector		
33 Series Adapter	–0.0002 to –0.0010	–0.0002 to –0.0010
34 Series Adapter (K to GPC-7)	–0.0002 to –0.0007 (K and GPC-7)	–0.0002 to –0.0007 (K and GPC-7)
Terminations/Loads	–0.005 to –0.040	–0.005 to –0.040
Shorts	–0.0002 to –0.0007	–0.0002 to –0.0007
Opens	–0.0002 to –0.0007	–0.0002 to –0.0007

Table 1-6. Pin Depth Limits (2 of 2)

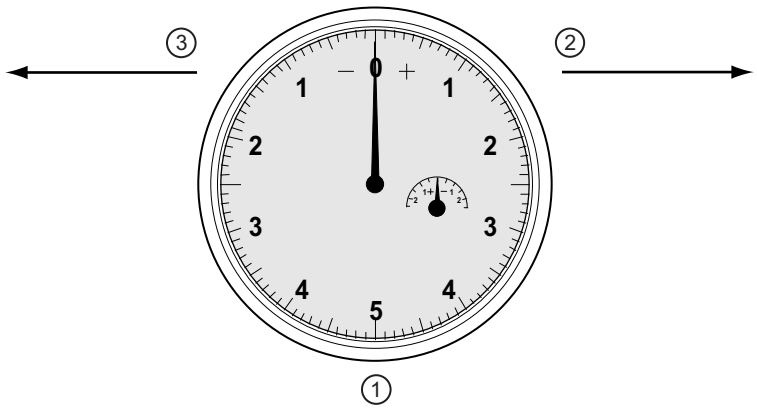
Connector Type/Device Type	Observed Pin Depth Limits (Inches)	
	Male	Female
Test Port Adapter (34YK50C)	–0.0002 to –0.0010	NA
V Connector		
33 Series Adapter	–0.0002 to –0.0010	–0.0002 to –0.0010
Terminations/Loads	–0.0002 to –0.0020	–0.0002 to –0.0020
Shorts	–0.0002 to –0.0010	–0.0002 to –0.0010
Opens	–0.0002 to –0.0010	–0.0002 to –0.0010
Test Port Adapter (34YV50C)	–0.0002 to –0.0010	NA
2.4 mm Connector		
Test Port Adapter (34YQ50A)	–0.0002 to –0.0010	NA

a.Starting with kit serial number 1644002, part numbers 34AS50-2 and 34ASF50-2 (3.5 mm to GPC-7 Adapters) are not supplied with the 3650A/3650A-1 calibration kits.

Reading Pin Depth

When gauging pin depth, if the connector measures out of tolerance in the "+" region of the gauge, the center pin too long or is protruding. Mating under this condition can damage or break female pin contact fingers.

If the test device connector measures out of tolerance in the "–" region, the center pin is too short or is recessed. While this will not cause any damage, it will result in a poor connection and a consequent degradation in performance.



1. Pin Depth Gauge with needle setting at zero.
2. Positive needle direction clockwise to right. Center pin is long or protruding.
3. Negative needle direction counter-clockwise to left. Center pin is short or recessed.

Figure 1-25. Pin Depth Gauge

1-10 Connector Cleaning Instructions

With repeated connections and disconnections, the threads and outer conductor mating interface builds up a layer of dirt and metal chips which can severely degrade connector electrical and mechanical performance. This debris can also increase the coupling torque required for a good connection which can then damage the mating interfaces. Cleaning of connectors is essential for maintaining good electrical performance. Therefore, connectors should be checked for cleanliness before making any measurements (or calibration).

See the *RF and Microwave Connector Care Instruction Sheet*—10100-00031.

Caution	Use the correct sized cotton swabs that are made specifically for cleaning small areas. Oversized cotton swabs can put lateral pressure on the center pin and damage it. Inspect after cleaning to be sure that no cotton strands get caught in the connector.

Required Cleaning Items

- Low-pressure dry, compressed air (oil free, solvent free), maximum pressure: 40 PSI
- Lint-free cotton swabs
- Isopropyl alcohol (IPA), >90 %
- Microscope Cleaning Procedure

	Teflon Tuning Washers: The center conductor on some RF components contains a small teflon tuning washer located near the point of mating (interface). This washer compensates for minor impedance discontinuities at the interface. The washer's location is critical to the RF component's performance. Be careful not to disturb it with a cotton swab or compressed air and don't apply alcohol to it.
Caution	**Compressed Air:** Do not use compressed air on power sensors. Some power sensors have components internally located near the connector that can be damaged from excessive air pressure.
	Solvents: Do not use industrial solvents or water to clean the connector. Use 90% Isopropyl Alcohol (IPA) only. Do not spray alcohol directly onto connector surfaces. Do not saturate the swab with alcohol. Instead, lightly dampen it by touching the tip onto a bead of alcohol formed at the bottle tip as shown in Figure 1-26.

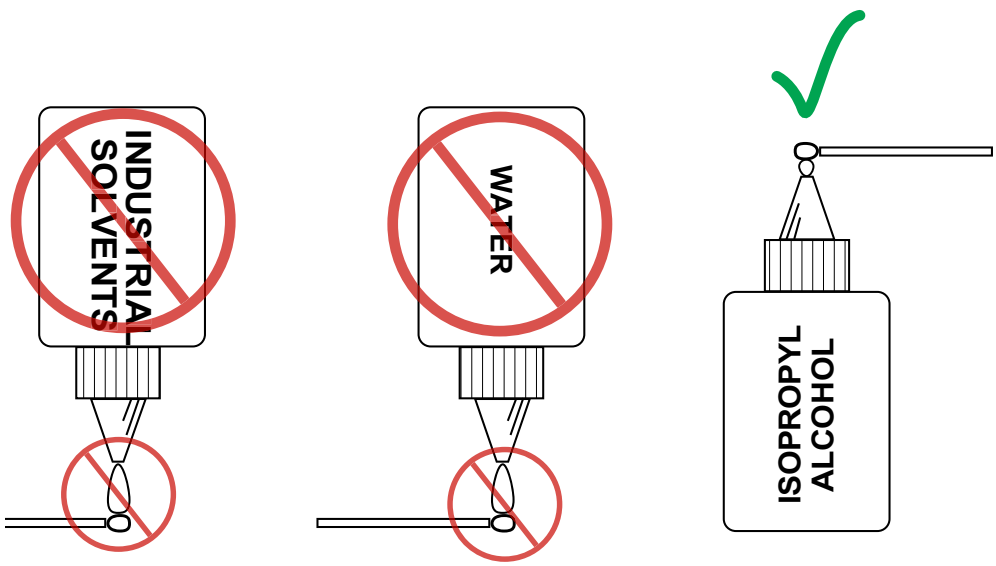


Figure 1-26. Isopropyl Alcohol Only

Procedure

1. Remove loose particles on the mating surfaces, threads, and similar surfaces using low-pressure clean (oil free, solvent free) compressed air (40 PSI max) applied at a shallow angle so foreign particles are not forced down into the connector.

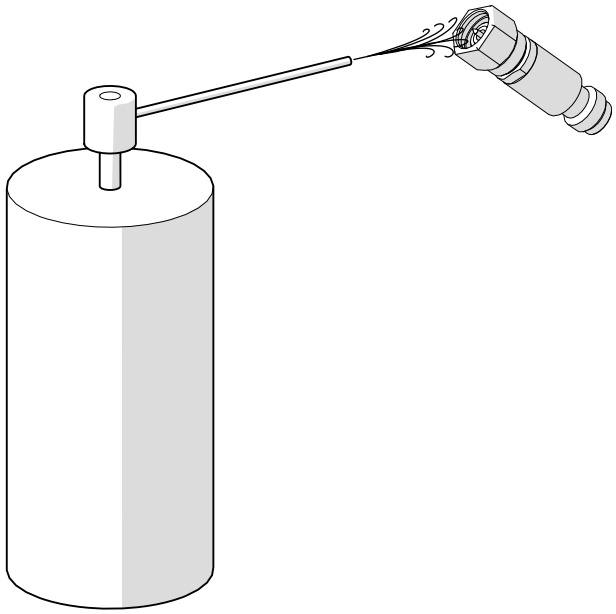


Figure 1-27. Low Pressure Compressed Air Cleaning

2. Clean the threads of the connector with a lint-free cotton swab dampened with IPA.

Caution	Do not insert cotton swabs at an angle or use a swab that is too large (Figure 1-28). Doing so can damage the center pin.



Figure 1-28. Avoid Angled or Large Swab

3. Clean the mating surfaces and connector threads by gently moving the cotton swab around the center pin. Avoid applying pressure to the center pin. See Figure 1-29 on page 1-36.

   When the connector is clean, you should be able to hand tighten the connector to within approximately one half turn of the specified torque.

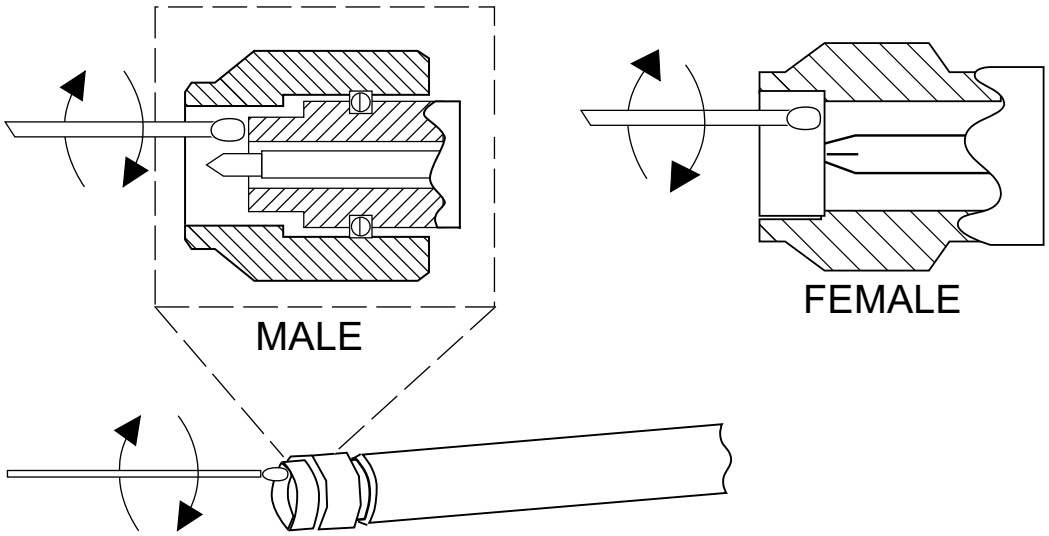


Figure 1-29. Cleaning Connector with Cotton Swabs

4. After cleaning with swabs, again use low-pressure compressed air to remove any remaining small particles and dry the connector surfaces.
5. Using magnification and adequate lighting, perform a final inspection of the connectors for damage or debris.

1-11 Torque Specifications and Tools

Torque values are listed below for connectors used with the 365xx-x mechanical calibration kits. If you are not sure which connector types are available on your instrument, consult the instrument Technical Data Sheet available on the web at www.anritsu.com.

Table 1-7. Connector Torque Settings and Recommended Tools

Connector Type/Size	Wrench Size	Torque Setting	Recommended Tools
3.5 mm[a b]	8 mm 5/16 in	0.9 N·m 8 lbf·in	01-201 5/16 in Torque End Wrench 01-204 5/16 in or 8 mm Open End Wrench
GPC-7	19 mm 3/4 in	12 lbf·in 1.36 N·m	01-200 3/4 in Torque End Wrench
K (2.92 mm)	8 mm 5/16 in	0.9 N·m 8 lbf·in	01-201 5/16 in Torque End Wrench 01-204 5/16 in or 8 mm Open End Wrench
V (1.85 mm)	8 mm 5/16 in	0.9 N·m 8 lbf·in	01-201 5/16 in Torque End Wrench 01-204 5/16 in or 8 mm Open End Wrench
K, V, 2.4 mm (Test Port Adapter)	1/2 in	4.1 N·m 36 lbf·in	1/2 in Torque End Wrench 1/2 in Open End Wrench
W (1 mm)	6 mm	0.45 N·m 4 lbf·in	01-504 6 mm Torque End Wrench and 7 mm Open End Wrench
0.8 mm	6 mm	0.45 N·m 4 lbf·in	01-504 6 mm Torque End Wrench and 7 mm Open End Wrench

a.3.5 mm, 2.92 mm and K connectors are electrically compatible with 3.5 mm and have the same connector nut size (8 mm) and torque requirement. Refer to Table 1-8 for connector compatibility information.

b.Although the 2.4 mm, 1.85 mm, and V connectors have the same connector nut size (8 mm) and torque requirement, they are not mechanically compatible with 3.5 mm. Refer to Table 1-8 for connector compatibility information.

1-12 Connector Types and Cross-Mating Compatibility

The following table lists RF connector characteristics and cross-mating compatibilities.

Table 1-8. RF Connector Characteristics

Connector Type/ Parameter	7/16 DIN	20 GHz Type N	Type N	7 mm	SMA	3.5 mm	K (2.92 mm)	2.4 mm	V (1.85 mm)	W1 (1 mm)	0.8 mm
Upper Frequency Limit (GHz)	7.5	20	18	18	18	26.5	40	50	70	110	145
Mating Type	M/F	M/F	M/F	Sexless	M/F	M/F	M/F	M/F	M/F	M/F	M/F
Cross-mating Compatibility	7/16 DIN only	Type N	20 GHz Type N	7 mm only	3.5 mm, K,	K, SMA,	3.5 mm, SMA,	V	2.4 mm	W1 (1 mm) only	0.8 mm only
Dielectric	Air	Air	Air	Air	Teflon	Air	Air	Air	Air	Air	Air
Thread	M29 x 1.5	5/8–24	5/8–24	0.6785-24	1/4–36	1/4–36	1/4–36	M7 x 0.75	M7 x 0.75	M4 x 0.7	M3.5 x 0.35
Outer Conductor (mm)	16	7	7	7	4.2	3.5	2.92	2.4	1.85	1	0.8

Chapter 2 — AutoCal Calibration

2-1 Chapter Overview

This chapter describes common procedures for both Internal Thru and True-Thru calibration using Anritsu AutoCal modules including the **36585-Series Precision AutoCal Modules** and the **36581KKF Standard AutoCal Module**. This chapter provides two examples of using a precision adapter with an AutoCal unit to create difference-gender reference planes as is adapter removal with an AutoCal unit. The chapter concludes with user-based AutoCal module characterization and working with older Anritsu AutoCal units.



Figure 2-1. MS4647B VNA with 36585V-Series Precision AutoCal Module

2-2 Automatic Calibrator (AutoCal) Terms

This section defines various terms used with AutoCal modules and the calibration procedures used with the VectorStar.

Insertable and Non-Insertable Devices

Insertable devices have an insertable connector pair such as a male input connector and a female output connector. These devices can be measured after a thru calibration. A non-insertable device has a non-insertable pair of connectors such as female connectors on both ports or different connector types such as a DUT with both a K and V connector. Non-insertables cannot be connected directly into the measurement path without an adapter.

- **Through ("Thru")**

  A *thru* (properly called a "through") is a connection of the two test ports. Two kinds of through connections, Internal Through and True Through, are defined for the AutoCal calibration.

- **Internal Through ("Internal Thru")**

  An Internal Through is an internal path through the calibrator requiring no operator involvement in the AutoCal setup.

- **True Through ("True Thru")**

  A True Through is a direct cable connection between the test ports, with no intervening connectors. An internal through is not as accurate as a true through, but requires more operator involvement and takes more time to accomplish. The true through can be more accurate than the internal through (if very well-matched and of very low loss) but the length and loss must be entered. The return loss of the adapters must be very good to avoid added error.

For the 36585 series of AutoCal, both the true thru and the internal thru can be processed as 'reciprocals'. This concept is discussed in more detail in Chapter 3 and Chapter 16 of this guide but the basic principle is that calibration will not use knowledge of the thru characteristics, but will place more weight on the accuracy of the reflection standards. In the case of the true thru, the reciprocal approach is advisable if the characteristics of the thru are not well-known and the insertion loss/mismatch are substantial. Reciprocal processing is also usually advised for the internal thru path, but defaults to using the characterization data for backwards-compatibility.

Auto-Sense

A selectable function of the AutoCal module where, if enabled, the AutoCal module decides the left/right test cable assignment. If not enabled, the operator defines the port number to left/right assignment.

Characterization File

Each calibrator module has a file containing data which characterizes each standard in the calibrator. This file also contains information (identification number, start and stop frequencies) concerning the capabilities of the calibrator. Each characterization file has the extension "`.acd`." When modules are changed, the appropriate new characterization file must be installed. The files are typically stored on a Characterization Memory Device (USB memory device). Insert the **Characterization Memory Device** into a **VectorStar USB** port. This file can be installed by navigating to the INSTALL (AUTOCAL CHARACTERIZATION/CAL KIT) dialog box:

- MAIN MENU | Calibration | CALIBRATION | Cal. Kit/AutoCal Characterization | CAL KIT/AUTOCAL | Install Kit/Charac. | INSTALL (AUTOCAL CHARACTERIZATION/CAL KIT) dialog box

From the dialog box, navigate to the location of the Characterization Memory Device. Select the file, and copy it to a location within the `C:\AnritsuVNA` folder. In addition, each AutoCal module can be re-characterized using the VNA, although specifications are only valid if Anritsu has performed the characterization (recommended re-characterization interval is 12 months). A valid 12-term calibration must be active, which is used to characterize the standards within the module.

2-3 AutoCal Introduction

Note	In some menus and screens, S11 is used to represent the S_{11} calibration parameter, S12 for S_{12}, S21 for S_{21}, and S22 for S_{22}, and so on.

The auto calibration (AutoCal) process represents both a calibration kit and an algorithm that is used to speed up the calibration process with extremely high accuracy, minimizing the number of manual steps and test operator involvement.

Transfer Calibration

The **36585x-Series Precision AutoCal** and the **36581KKF Standard AutoCal** module calibrates the VNA by a process known as "transfer calibration." There are a number of impedance and transmission states in the module designed to be extremely stable in time and these states are carefully "characterized", generally by the Anritsu factory but also in a customer laboratory in certain cases.

When the same states are re-measured during a calibration, and the results compared to the characterization data, an accurate picture can be generated of the behaviors and error terms of the VNA and setup being calibrated.

Calibration Accuracy

A very high calibration accuracy is maintained through the use of certain principles:

- The use of many impedance and transmission states
- The creation of very stable states that are further enhanced with a constant-temperature thermal platform inside the module.
- The use of very reliable and repeatable solid-state switching constructed to provide a great variety of state impedances (for better calibration stability) and clean transmission paths.
- The use of a very careful characterization process that can generate excellent starting data.

The resulting accuracy, for some parameters, such as residual directivity, can exceed that obtained with typical mechanical calibration typically performed in a metrology laboratory. The variances, particularly on tracking, are related to pin depth variances on the test ports being calibrated and the practical difficulty in controlling those precisely.

The reflectometer portion of the calibration is based on the use of five reflection standards in an overdetermined algorithm. The standards were constructed to ensure that the reflection coefficients of at least three of them are far enough apart at every frequency to ensure an independent set of data. This was done through engineering of electrical lengths and added compensation. The transmissive part of the calibration is based on three transmission standards of varying insertion loss and return loss to add another level of overdetermined analysis. Finally, an assurance standard (yet another transmissive standard but with different parameters) is measured after the user calibration and the S-parameters compared against values stored at the factory. If these parameters all agree to within the measurement uncertainty, the message 'Assurance Passed' is generated and can give some confidence that the calibration was successful to within expected tolerances.

AutoCal Components

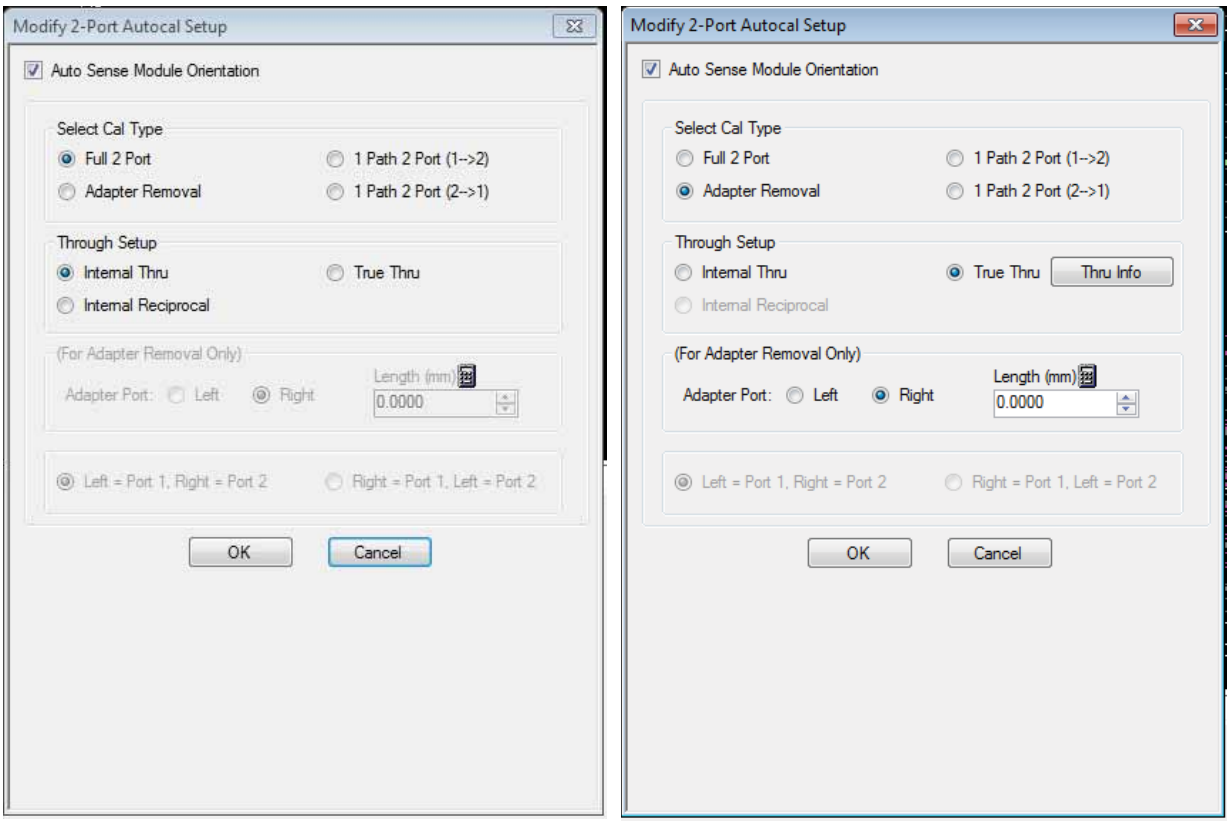
The AutoCal unit consists of the calibration module itself, a separate external power supply, a control cable that runs to the VNA, and the characterization data that is initially provided on a USB memory device. The AutoCal unit should be powered up and allowed to warm up prior to use (typically a few minutes, the blue **Operate LED** will illuminate when the unit is at temperature). Allow 90 minutes warm-up time for calibrations. The control cable should be connected to the serial port on the back of the VNA.

Connector Care

Since an AutoCal calibration is a very high performance calibration, connector care is of the utmost importance.

AutoCal Calibration Parameters and Types

AutoCal for 2-port requires four general parameters be defined, as shown in the AutoCal dialog box:



1. Settings (at left) for Full 2-port Calibration with Internal Thru.
2. Settings (at right) for Adapter Removal and True Thru.

Figure 2-2. MODIFY 2-PORT AUTOCAL SETUP Dialog Box

AutoCal for 1-port and 4-port is discussed in Section 24-6 “AutoCal Calibrations” on page 24-37.

The parameters are:

- the calibration type required
- the thru type to be used
- whether to automatically or manually define the left/right test port assignments
- whether an adapter will be used with the AutoCal module.

Most standard calibration types are available with the AutoCal units. Frequency response calibrations are the only ones omitted since the AutoCal unit would provide no benefit in these cases since only a thru or high reflect standard is needed for these calibrations. The AutoCal modules support:

- Full Two Port (also called Full 2-port) calibration
- Full Port 1 (S_{11}) reflection only calibration
- Full Port 2 (S_{22}) reflection only calibration

- 1-Path 2-port Forward (S_{11}, S_{21}) calibration
- 1-Path 2-port Reverse (S_{22}, S_{12}) calibration

Adapter Removal

The DUT may require a different connector configuration than is on the current AutoCal unit (e.g., F-F DUT and the auto cal is M-F). AutoCal adapter removal is one way of handling this situation by performing two auto calibrations with an adapter connected to the AutoCal. The instrument will guide the user through the required connections.

Thru Options

For each calibration type above, there are three thru AutoCal calibration options:

- **Internal Thru**

  The AutoCal module provides the internal thru connections without the user having to move or reconnect the test cables. The benefit is speed of calibration balanced against potentially less accuracy and accommodates inexperienced operators, such as in an assembly line testing station.

- **Internal Reciprocal**

  The Internal Reciprocal selection uses the same internal path as the Internal Thru selection, so there are the same advantages in terms of reduced connections required, but the path is treated as an unknown thru or reciprocal and does not use the thru path characterization data. As such, it can provide improved accuracy in cases where the characterized thru does not perform as well (larger pin depth variations in the reference plane (cable-end) connectors, more environmental variation, etc.). The best-case-situation accuracy may not be quite as good relative to Internal Thru

- **True Thru**

  All AutoCal modules can also be configured to allow True Thru where the user is prompted when and how to connect the test cables to complete the calibration. The benefit is a higher accuracy calibrations balanced against a longer calibration time and more operator involvement.

Test Port Left/Right Naming Assignments

The AutoCal module setup allows the user to specify the left/right identification of the test port cables as either:

- **Auto Sense Module Orientation = ON**

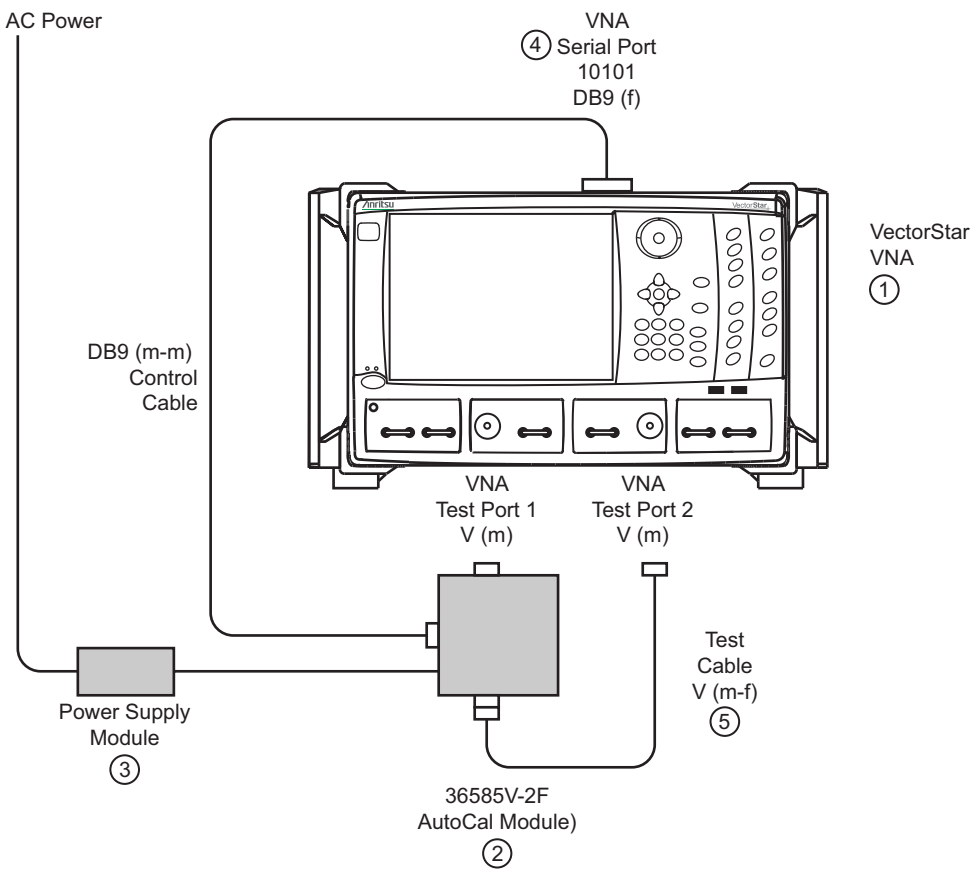
  Where the module determines the left/right identification of the test port cables

- **Auto Sense Module Orientation = OFF**

  Where the user defines either Left = Port 1 and Right = Port 2, or Left = Port 2 and Right = Port 1. This setting is useful where the VNA orientation is different for the operator.

Typical AutoCal Module Connections

The typical **F-F AutoCal** module connections are shown below in Figure 2-3.

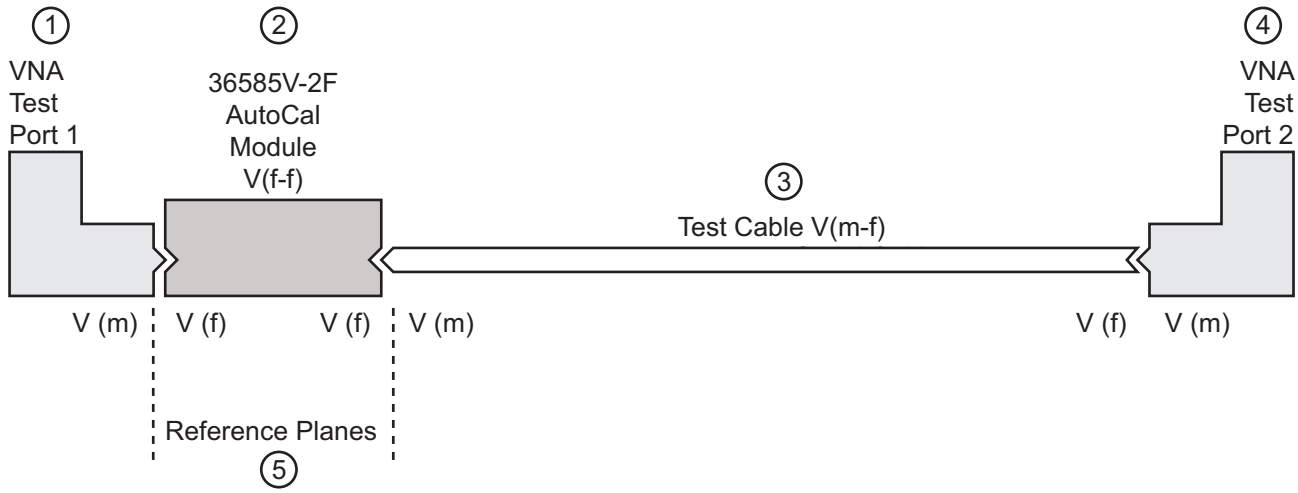


1. MS4645B or MS4647B VNA with V connectors. 2. 36585V-2F AutoCal Module 3. AutoCal Power Supply Module with extension able to local AC Mains power.	4. AutoCal DB9 (m-m) Control Cable connected between AutoCal Module and VNA rear panel Serial 10101 Port. 5. Test Cable V (m-f)

Figure 2-3. AutoCal Module Connections for Internal Thru

1. Prepare the **VectorStar MS4645B or MS4647B VNA** and power it up. The instrument has two V (m) test ports.
2. Connect the **366565V-2F Precision AutoCal Module V (f)** connected directly to **VNA Test Port 1 V (m).**
3. The AutoCal module is then connected to its **AC Power Supply Module** and it to AC power
4. Connect the **DB-9 (m-m) Serial Control Cable** to the AutoCal module and the VNA rear panel **Serial Port (10101)**.
5. The second **AutoCal V (f)** connector is connected to the **V (m)** end of the **Test Cable**. The test cable can be changed depending on measurement requirements. Cables on both VNA ports may be used, different types of cables may be used, or other configurations established. Then connect the **V (f)** end of the **Test Cable** to the **VNA Test Port 2 V (m).**

Schematically, this setup is shown in Figure 2-4.



1. VNA Test Port 1 V (m)	4. VNA Test Port 2 V (m)
2. 36585V-2F AutoCal Module	5. Resultant reference planes
3. V (m-f) Test Cable	

Figure 2-4. Schematic of AutoCal 36585V-2F Module Connections for Internal Thru

For optimal results, use the shortest cable lengths that do not require excessive bending when performing calibration or measurements. Results will be improved using the most practical phase- and amplitude-stable cables.

The power supply and control cables may be bundled for ease of routing or may be separated for convenience in some cases.

Using the AutoCal Module

The calibration procedure can be broken up into several simple steps:

1. Power up the VNA and connect the control cable to the AutoCal unit.
2. Install the characterization file if not already done. Copy the AutoCal Characterization File from the Characterization Memory Device (a USB memory device) onto the VectorStar hard disk drive.

Note	The AutoCal Characterization File (ACD file) is provided on a USB memory device and can be installed directly from that device. These files can also be copied from the USB memory device to the VectorStar hard drive and installed from there. The preferred method is to copy all ACD files onto the hard drive, and then install the file for the specific AutoCal being used.

3. Connect the AutoCal module to the VNA:
   - AutoCal Module directly to one **VNA Test Port**
   - A test cable between the AutoCal and **VNA Test Port**
   - The Serial Control Cable between the AutoCal module the VectorStar back panel **Serial 10101** port
4. Connect the AutoCal Module to its power supply and AC power. The green **Power LED** illuminates immediately. When the module is at operating temperature, the blue **Operate LED** will illuminate.
5. Setup the VNA instrument for the desired calibration:
   - The required minimum settings are Frequency Start, Frequency Stop, and Number of Points.
   - If required, optional settings for Segmented Sweep, IF Bandwidth, and/or Averaging are applied.

6. Select the AutoCal parameters of interest and connect the test port cables to the AutoCal unit. Make the required selections for Cal Type, Thru Select, Adapter Removal, and Auto Sense Module.
7. Perform the calibration by clicking Begin Cal to start the auto calibration.
8. A status dialog box with a progress bar appears after an AutoCal sequence has started. The status messages define has far the program has progressed and if any user actions are required.
9. At any time, the AutoCal sequence can be canceled by clicking the dialog box Abort button.
10. If any manual steps are requested such as specifying a true-thru, a dialog box will prompt for the action.
11. When the auto calibration is complete, a status message appears with a statement about assurance passing or failing. On the Calibration menu, the Cal Status field button shows ON.

Each procedure is described in greater detail in the following sections.

2-4 Available AutoCal Modules and Adapters

The Precision and Standard AutoCal modules are show in the figure below.



1. 36581KKF Standard AutoCal Module (at left).	2. 36585-Series Precision AutoCal Module (at right).

Figure 2-5. AutoCal Automatic Calibrator Calibration Modules

Anritsu has both Precision AutoCal modules as well as a Standard AutoCal module. The table below shows the available modules.

Table 2-1. Precision and Standard AutoCal Kits and Adapters

Part Number	Name	Frequency Range and Connectors
Precision AutoCal Kits		
36585V-Series Precision AutoCal Modules can be used with MS4642B or MS4644B VNAs with adapters. The calibration range is limited to that of the VNA. Only the 36585V-2MF and 36585K-2MF Precision AutoCal Modules can be used with the 366xX-1 Verification Kits. The 36585K-Series Precision AutoCal Kits can be used with MS4645B or MS4647B VNAs with adapters. The calibration range is limited to 40 GHz.		
36585K-2M	Precision AutoCal Kit, K (m-m), 2-port	70 kHz to 40 GHz, K (m) to K (m)
36585K-2F	Precision AutoCal Kit, K (f-f), 2-port	70 kHz to 40 GHz, K (f) to K (f)
36585K-2MF	Precision AutoCal Kit, K (m-f), 2-port	70 kHz to 40 GHz, K (m) to K (f)
36585V-2M	Precision AutoCal Kit V (m-m), 2-port	70 kHz to 70 GHz, V (m) to V (m)
36585V-2F	Precision AutoCal Kit V (f-f), 2-port	70 kHz to 70 GHz, V (f) to V (f)
36585V-2MF	Precision AutoCal Kit V (m-f), 2-port	70 kHz to 70 GHz, V (m) to V (f)
Standard AutoCal Kit		
The 36581KKF Standard AutoCal Module can be used with any MS464xB VNA with adapters as required on the Test Ports and/or Module. The calibration range is limited to 20 GHz.		
36581KKF	Standard AutoCal Kit	40 MHz to 20 GHz, K (m) to K (f)
Standard AutoCal Kit Matched Adapters		
Matched adapters are recommended only for use with the 36581KKF Standard AutoCal Module. The standard AutoCal module can be ordered with connector adapters which assume the use of K (m) test port cables. These adapters can be also used on the Precision AutoCal modules, but with some degradation of precision. See Chapter 1, "Calibration and Measurement Overview" for available calibration modules, tools, and accessories.		
36583K	Matched K Adapters Set, 2 adapters	40 MHz to 20 GHz, K (f) to K (m) adapters
36583L	Matched 3.5 mm Adapters Set, 4 adapters	40 MHz to 20 GHz, K(f) to 3.5 mm (m) adapters, 2 each K (f) to 3.5 mm (f) adapters, 2 each
36583S	Matched SMA Adapters Set, 3 adapters	40 MHz to 20 GHz, K (f) to SMA (m) adapter, 1 each K (f) to SMA (f) adapters, 2 each

2-5 Copying the AutoCal Module Characterization File to the VectorStar

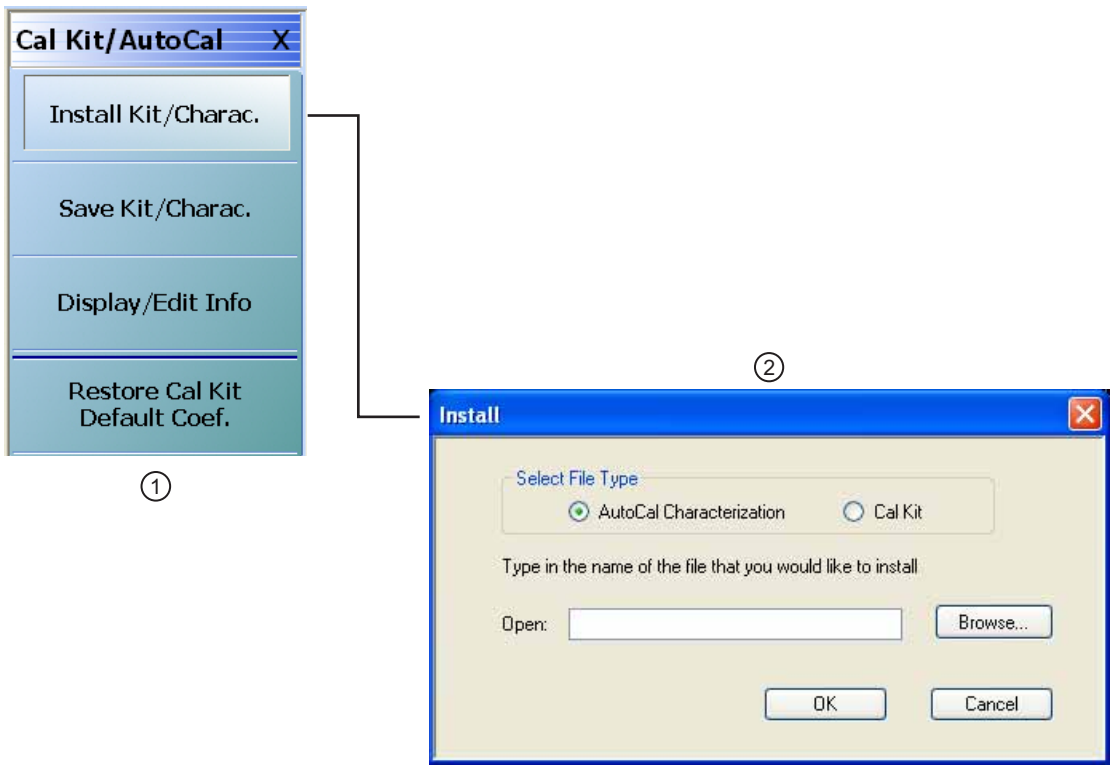
The AutoCal Characterization File (.acd file) is provided on a USB memory device and can be installed directly from that device. The files can also be copied from the USB memory device to the VectorStar hard drive and installed from there. The preferred method is to copy all ACD files to the hard drive, and then select the file for the specific AutoCal being used.

Procedure

1. Inspect the AutoCal kit and make a note of its serial number. Typical AutoCal Kit serial numbers are six-digit integers such as "123456."
2. Insert the AutoCal Characterization Memory Device (USB memory device) into one of the VectorStar USB Ports.

Note	Alternatively, Windows Explorer can also be used to move files from the USB memory device to the VectorStar hard drive.

3. Navigate to the CAL KIT/AUTOCAL menu:
   - MAIN | Calibration | Cal Kit/AutoCal Characterization | CAL KIT/AUTOCAL



1. At left, the CAL KIT/AUTOCAL Characterization Coefficients Utility Menu

2. The Install Kit/Charac. (Characterization) button links to the INSTALL AUTOCAL CHARACTERIZATION/CAL KIT Dialog Box shown here with the AutoCal Characterization button selected.

Figure 2-6. CAL KIT/AUTOCAL Utility Menu—INSTALL CHARACTERIZATION FILE Dialog Box

4. On the CAL KIT/AUTOCAL menu, select Install Kit/Charac. The Install (AutoCal Characterization/CalKit) dialog box appears.
5. Select the AutoCal Characterization radio button and then click Browse. The Open (AutoCal ACD File) dialog box appears.

6. Navigate to the USB port location holding the USB memory device and select the file "V123456.ACD," where "123456" is the serial number of the AutoCal kit.
7. Click the dialog box Open button. The Open dialog box closes; the Install dialog box re-appears.
8. Click OK in the Install dialog box. Another install dialog box appears (this one showing the AutoCal controller serial number, which is not the module serial number). Click Install in this box close it.
9. On the Cal Kit/AutoCal menu, select Save Kit/Charac. The Save (AutoCal Characterization) dialog box appears. Click OK.
10. Navigate to the required storage directory for the characterization file. The recommended directory destination on the VectorStar is C:\Anritsu\VNA\AutoCal
11. Click Save. The Save dialog box closes, the CAL KIT/AUTOCAL menu reappears.
12. Repeat the steps above to copy each new AutoCal Characterization File to the VectorStar.

2-6 Loading a Previously Stored AutoCal Characterization File

Use this procedure if the AutoCal Characterization File has already been copied onto the VectorStar hard drive.

Procedure

1. Navigate to CAL KIT/AUTOCAL menu.
   - MAIN | Calibration | CALIBRATION | Cal Kit/AutoCal Characterization | CAL KIT/AUTOCAL
2. On the CAL KIT/AUTOCAL menu, select Install Kit/Charac.
3. Select the AutoCal Characterization radio button and then click Browse.
4. Navigate to the hard drive (or USB memory device) location of the AutoCal Characterization file for the AutoCal module in use. The recommended directory destination on the VectorStar is `C:\Anritsu\VNA\AutoCal`
5. Select the AutoCal Characterization file, such as "`V123456.ACD,`" where "`123456`" is the serial number of the AutoCal kit. Click Open.
6. Click OK in the Install dialog box. Another install dialog box appears (this one showing the AutoCal controller serial number, which is not the module serial number). Click Install in this box to close it.
   - The Install dialog box closes and the Cal Kit/AutoCal menu is again available.

2-7 Pre-Calibration Instrument Setup

Use this procedure to setup the minimum required instrument configuration parameters:

- Frequency Start
- Frequency Stop
- Number of Points

Any other required measurement parameters must be defined and applied before the AutoCal procedure. This section provides a highlight of typical additional measurement parameters.

Segmented Sweep

If required, segmented sweep must be setup in advance if the calibration needs a custom frequency list. See "Frequency-Based Segmented Sweep" on page 15-18.

IF Bandwidth (IFBW) and Averaging

IF Bandwidth and Averaging control the digital filtering and post-processing that determine the effective noise floor, the amount of trace noise, and, in some cases, the immunity to interfering signals. The trade-off for improved noise performance is slower sweep speed.

Port Power

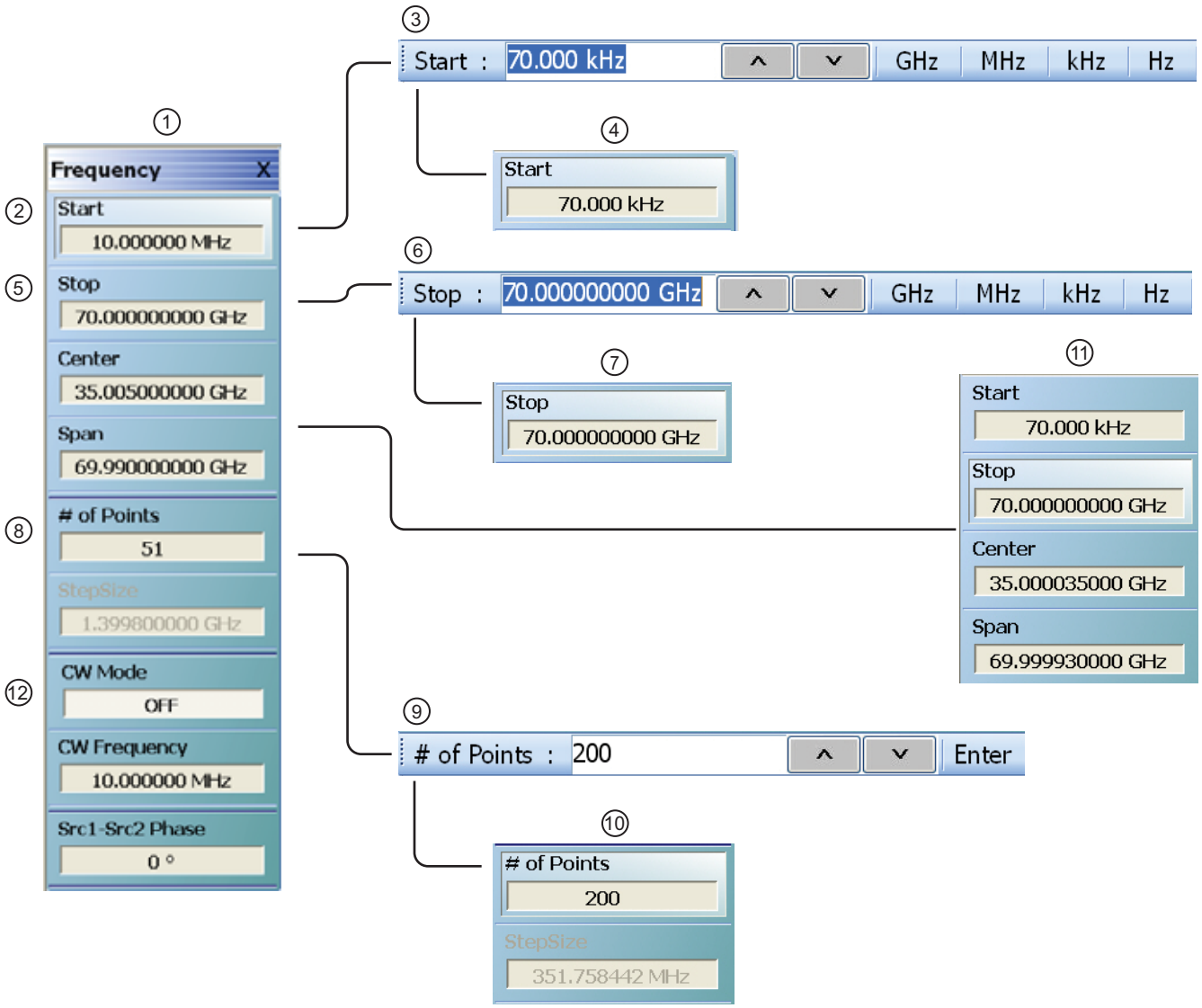
Port power is less critical than IFBW or Averaging due to the excellent linearity of the VectorStar MS4640B Series VNA receivers. The AutoCal unit has an absolute maximum power limit of +10 dBm. The preferred calibration power is –10 dBm for improved accuracy at frequencies <100 MHz.

If power adjustments are required, any step attenuator settings must be selected before the calibration. Changes in the step attenuator settings alters both the RF match and the insertion loss in the measurements paths. Power can be changed after a calibration, but an attenuator change after a calibration invalidates the prior calibration.

Example Procedure

This example procedure assumes the VectorStar MS464xB Series VNA is equipped with the Option 70—70 kHz Low Frequency Extension and that only Frequency Start, Frequency Stop, Number of Points, and CW Mode settings are required.

1. Determine the values for the minimum setup parameters:
   - Frequency Start: 70 kHz
   - Frequency Stop: 70 GHz
   - Number of Points: 200
   - CW Mode: OFF
   - Segmented Sweep: Not required
   - IFBW: Defaults to 1 kHz
   - Averaging: Defaults to no averaging
2. Power up the VectorStar and allow it to stabilize its internal temperature.
3. Navigate to the FREQUENCY menu:
   - MAIN | Frequency | FREQUENCY



1. FREQUENCY Menu	7. Stop Button after set to 70 GHz
2. Start (Frequency) Button before.	8. # (Number) of Points (Frequency) Button before
3. Start Field Toolbar with 70.000 kHz set	9. # of Points Field Toolbar with 200 points set
4. Start Button after set to 70.000 kHz	10. # of Points Button after set to 200 points. Below, Step Size automatically calculated to 351.758442 MHz.
5. Stop (Frequency) Button before	11. Center (Frequency) Button automatically calculated to 35.00003500 GHz. Below, Span Button automatically calculated to 69.999930000 GHz.
6. Stop Field Toolbar with 70 GHz set	

Figure 2-7. FREQUENCY Menu—Setting Initial AutoCal Parameters

4. On the FREQUENCY menu, click the Start frequency button (Figure 2-7—#2).
   - The frequency Start toolbar appears just below the icon toolbar (Figure 2-7—#3).
5. Enter the required start frequency number value.
   - For this example, set the start frequency value at 70.
6. Enter the required start frequency units.

- For this example, set the units to kilohertz by clicking kHz.
- The Start frequency field button now shows 70.000 kHz (Figure 2-7—#4).

Note	If the MS464xB Series VNA is not equipped with Option 70—70 kHz Low Frequency Extension, the lowest available starting frequency is 10 MHz.

7. On the FREQUENCY menu, click the Stop frequency field button (Figure 2-7—#5).
   - The frequency Stop toolbar appears (Figure 2-7—#6).
8. On the Stop frequency toolbar, enter the required frequency number value and units.
   - For this example, set the stop value at 70 and click the GHz units button.
   - The Stop frequency button now shows 70.000000000 GHz (Figure 2-7—#7).
9. The Center and Span display buttons show calculated values based on settings made above (See Figure 2-7—#7 above).
   - The Center frequency field button shows a value of 35.000035000 GHz.
   - The Span frequency field button shows a value of 69.9999930000 GHz.
10. On the Frequency menu, click the # of Points (Number of Points) button (Figure 2-7—#8).
    - The # of Points toolbar appears (Figure 2-7—#9).
11. On the # of Points toolbar, enter the required number of points.
    - In this example, set the # of Points to 200 and then click Enter on the toolbar.
    - The # of Points field button shows 200 (Figure 2-7—#10).
    - The Step Size display button automatically calculates the step size of 351.758442 MHz.
12. Click the CW Mode toggle button so it is set to OFF.
13. The FREQUENCY menu appears with the new settings as shown below.



Figure 2-8. FREQUENCY Menu—After Settings

14. If optional parameters are required, do any of the following optional procedures described elsewhere in this manual:

- “Frequency-Based Segmented Sweep” on page 15-18.
- “IFBW” on page 13-40.
- Averaging
- Power and Attenuation

15. If no optional parameters are required, the AutoCal is ready to proceed.

- “AutoCal—Full Two Port—Internal Thru” on page 2-17.

  OR

- “AutoCal—Two Port Cal—True Thru” on page 2-21

2-8 AutoCal—Full Two Port—Internal Thru

This procedure performs an AutoCal calibration using a full 2-port calibration with an internal thru, which is sufficiently accurate for most DUTs. The AutoCal Characterization file has already been loaded onto the VectorStar.

Required Equipment

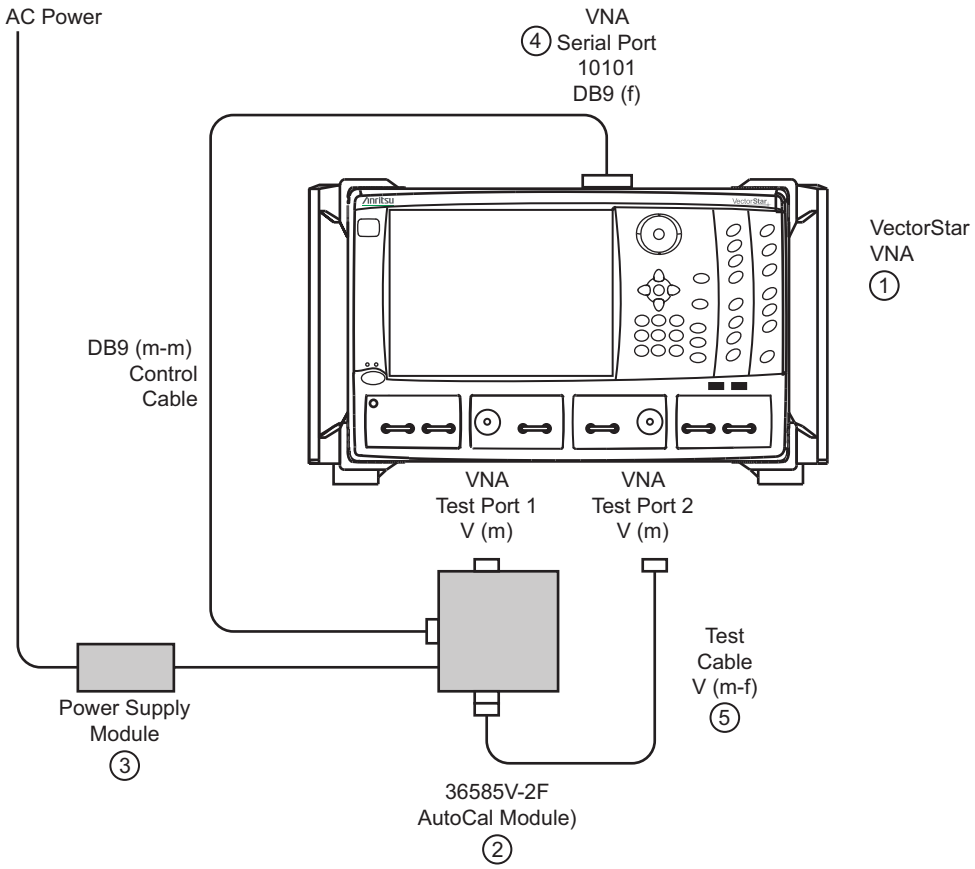
- VectorStar MS4645B or MS4647B VNA with V (m) test port connectors, with Option 70—70 kHz Low End Frequency Extension.
- 36585V-2F Precision AutoCal Module, with V (f) connectors and required power and control cables. The AutoCal Characterization File has been loaded onto the VectorStar VNA.
- Test port cable, V (f) to V (m)

Procedure

1. Power up the VectorStar VNA.
2. Set the required Frequency Start, Frequency Stop, and Number of Points parameters.
   - See "Pre-Calibration Instrument Setup" on page 2-13.
3. If required, set up other parameters as required such as Segmented Sweep, IFBW, Averaging, and Power and Attenuation.

AutoCal Module Connections

4. Make the necessary cable connections between the **AutoCal Module**, its **Power Module**, and the **VectorStar** rear panel. See Figure 2-9 on page 2-18.
5. Connect the **AutoCal V (f)** connector directly to the VectorStar left side **V (m) Test Port 1**.
6. Connect the **Serial Control Cable** between the DB-9 connector on the top of the AutoCal Module and the **Serial (10101) Port** on the VectorStar rear panel.
7. Connect the coaxial power plug from the **Power Supply Module** to the **AutoCal Module**. Connect the other end to AC power.
8. Once connected to power, the **AutoCal Module Power LED** is illuminated green. When the module has warmed up to operating temperature, the **LED** illuminates as blue.
9. Connect the **test cable** between the **V (m) Test Port 2** and the remaining **AutoCal V(f)** port.



1. MS4645B or MS4647B VNA with V Test Port Connectors 2. 36585V-2F AutoCal Module 3. AutoCal Power Supply Module	4. AutoCal DB9 (m-m) Control Cable connected between AutoCal Module and VNA rear panel Serial 10101 Port. 5. Test Cable V (m-f)

Figure 2-9. Precision AutoCal 36585V-2F Cable Connections for Internal Thru

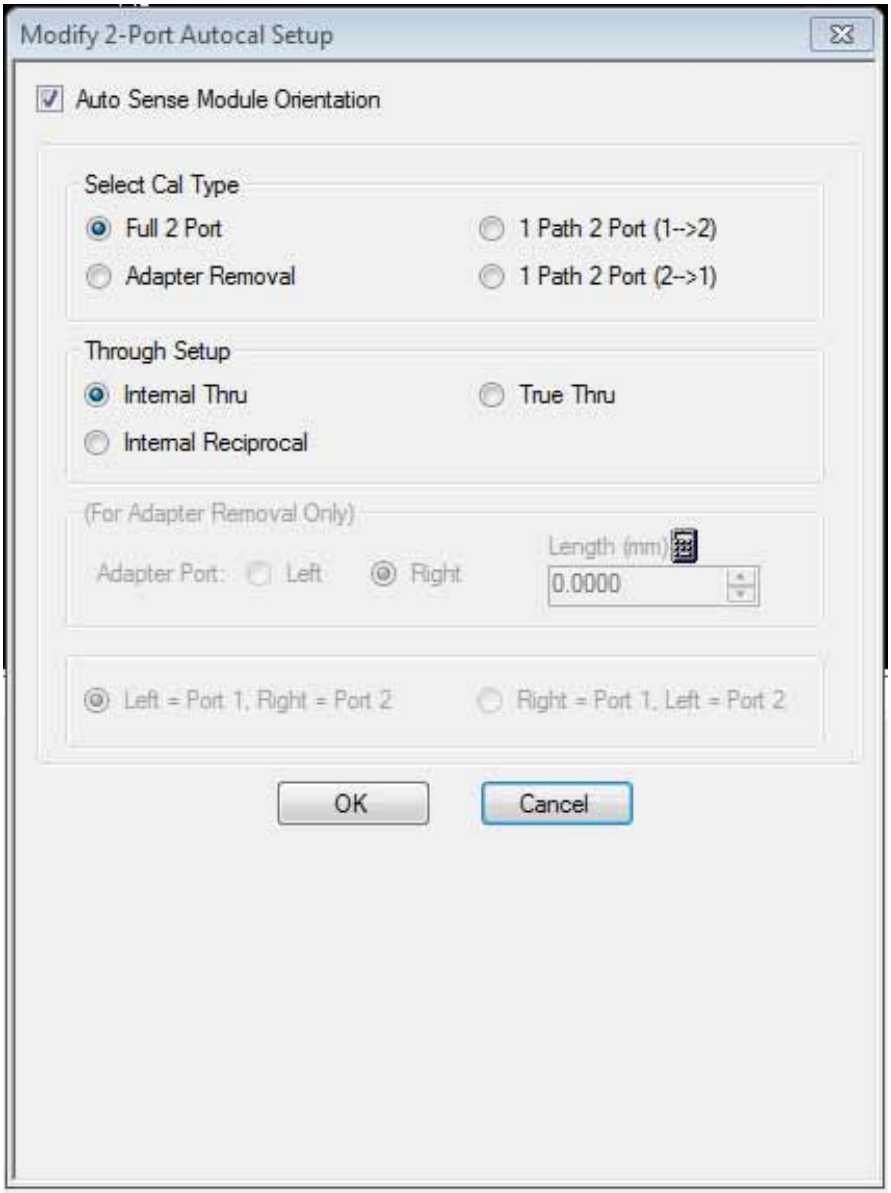
10. Navigate to the AUTOCAL SETUP (2-port) menu:

- MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | AutoCal | AUTOCAL PORTS | 2-PORT Cal | AUTOCAL SETUP (2-PORT)

11. On the AUTOCAL SETUP (2-PORT) menu, if the Port Selection, Cal Type, Thru Type, and Module Orientation display buttons do not show the correct values, click the Modify Cal Setup button.

- The MODIFY AUTOCAL SETUP dialog box appears. The exact name depends on the VNA mode and the user selections for the number of ports. The dialog box can be named:
  - MODIFY 1-PORT AUTOCAL SETUP dialog box
  - MODIFY 2-PORT AUTOCAL SETUP dialog box
  - MODIFY 4-PORT AUTOCAL SETUP dialog box

12. In this example, the required settings are for a Full 2-port Calibration, with Auto Sense Module Orientation ON, and Internal Thru while running on a VNA in 2-port Mode. The resultant configuration dialog box is named MODIFY 2-PORT AUTOCAL SETUP, and is shown in Figure 2-10 on page 2-19.



Settings for 2-port AutoCal Calibration • Auto Sense On	• Full 2-port • Internal Thru

Figure 2-10. MODIFY 2-PORT AUTOCAL SETUP Dialog Box

13. On the MODIFY 2-PORT AUTOCAL SETUP dialog box, select the settings:

a. Cal Type Select area: Select the Full Two Port radio button.

b. Thru Select area: Select the Internal Thru radio button.

c. Auto Sense Module Orientation check box selected: Allows the AutoCal module determine left/right cable identification.

d. When the settings are complete, select OK to close the dialog box.

14. The AUTOCAL SETUP (2-PORT) menu reappears with new values for Cal Type, Thru Type, and Module Orientation (see Figure 2-11 on page 2-20).

15. The window area at the bottom of the instrument display area appears with general instructions:

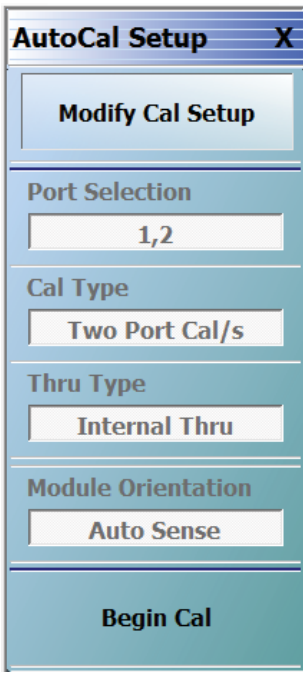
a. Ensure correct cable connections to AutoCal module.

b. Ensure that the **Power** and **Operate LEDs** are both illuminated.

c. Ensure characterization file is loaded before starting Cal. To load characterization file, go to the INSTALL (AUTOCAL CHARACTERIZATION) dialog box.

- MAIN | Calibration | CALIBRATION | Cal Kit/AutoCal Characterization | CAL KIT/AUTOCAL | Install Kit/Charc. | INSTALL (AUTOCAL CHARACTERIZATION)

d. Existing system setups such as averaging, power level, etc. will be applied during the cal.



Settings for Ports 1 and 2—Full 2-port—Internal Thru—Auto Sense On

Figure 2-11. Configured AUTOCAL 2-Port Menu

16. When ready, click the Begin Cal button.

17. If the AutoCal module is connected incorrectly, the AUTOCAL MODULE NOT DETECTED warning message appears.



- Correct the connections as required and click Retry.

Note	A different dialog box may appear if the RF cables are connected incorrectly (with Autosense on) that states that Autosense was unable to determine the orientation of the module. If this dialog appears, and a large amount of loss is not present between the port and the AutoCal module, check the connections. If a known large amount of loss is present (from the test fixtures or the use of very long cables), the orientation should be manually entered.

18. When the calibration is complete, the Status Message dialog box will close and the display will return to the CALIBRATION menu with the Cal Status button set to ON. The assurance dialog will remain up with the pass/fail message, and must be manually closed.

2-9 AutoCal—Two Port Cal—True Thru

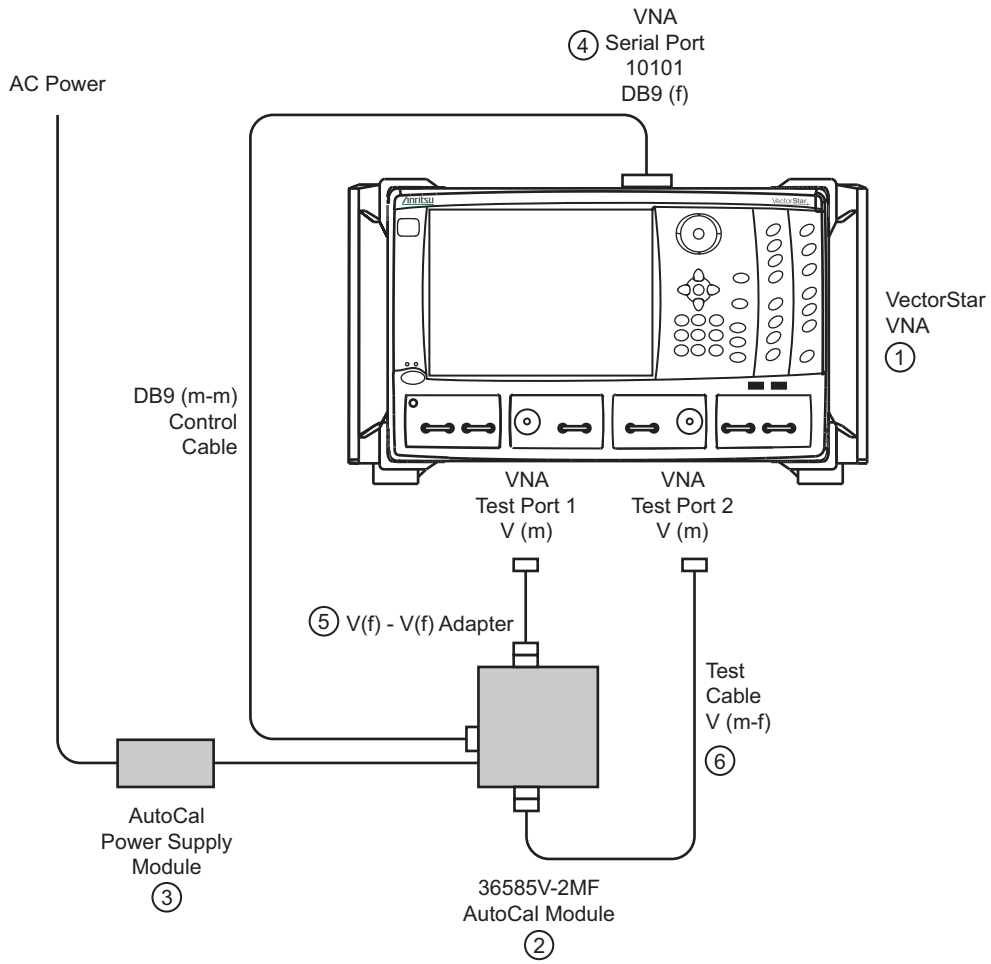
This procedure performs an AutoCal procedure using a full 2-port calibration where a True Thru (or external thru) is required.

Required Equipment

- VectorStar MS4645B or MS4647B VNAs with V(m) test port connectors, with Option 70—70 kHz Low Frequency Extension.
- 36585V-2MF Precision AutoCal Module, with V(f) and V(m) connectors and required power and control cables. The AutoCal Characterization file has been loaded onto the VectorStar.
- Test port cable with V(f) to V(m) connectors.
- Test port cable with V(f) to V(f) connectors, or a V(f) to V(f) adapter.

Procedure

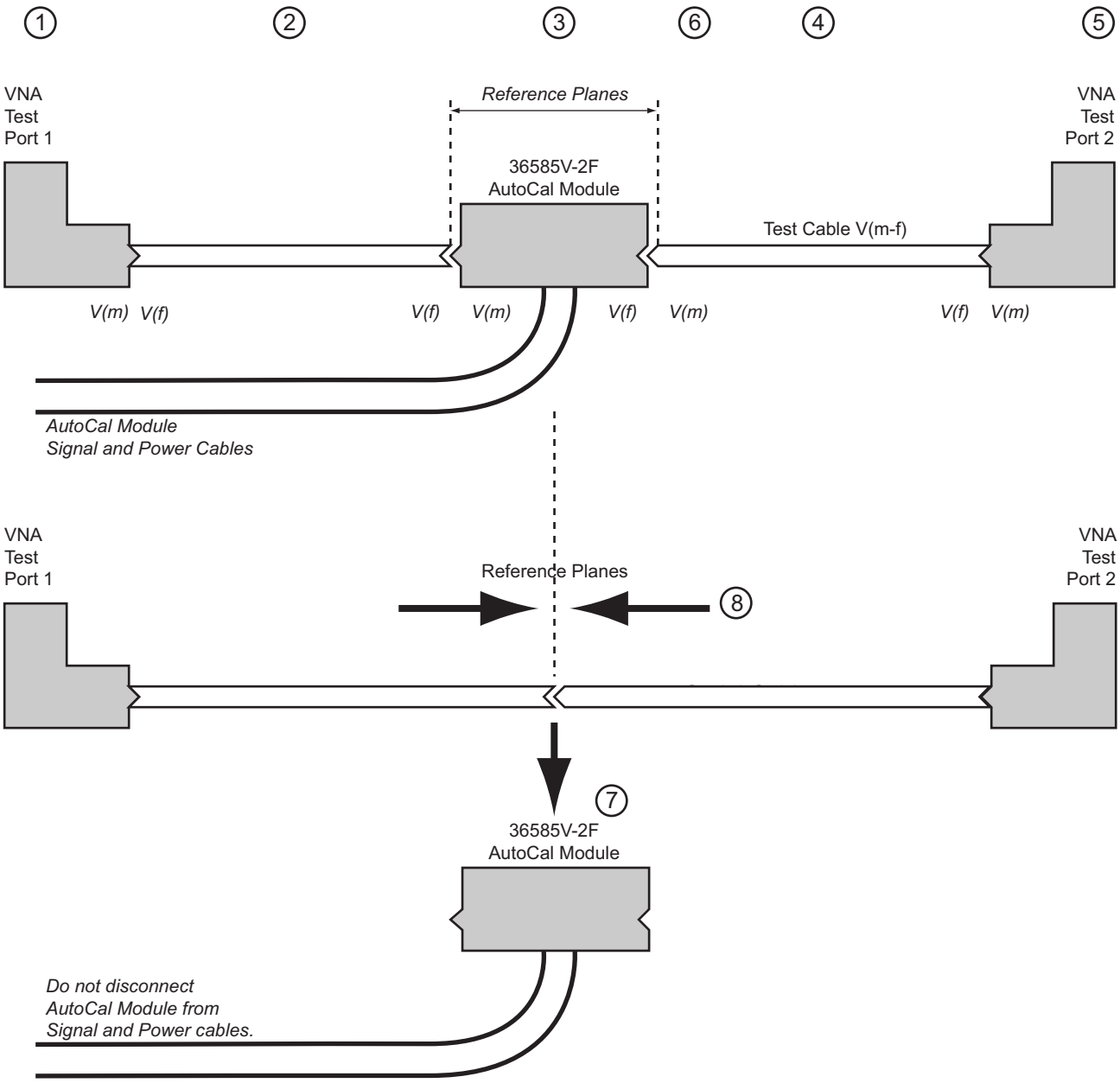
1. Power up the VectorStar.
2. Set the required Frequency Start, Frequency Stop, and Number of Points parameters.
   - "Pre-Calibration Instrument Setup" on page 2-13.
3. If required, set up other parameters as required such as Segmented Sweep, IFBW, Averaging, and Power and Attenuation.
4. Make the necessary cable connections between the **AutoCal Module** and the **VectorStar**:
   a. Connect the **Test Cable V(f)–V(m)** between Test Port 2 V (m) and the AutoCal V(f) port.
   b. Connect the **Test Cable V(f)–V(f)** between Test Port 1 V (m) and the AutoCal V(f) port.
   c. Connect the **Serial Control Cable** between the DB-9 connector on the AutoCal Module and the Serial (10101) Port on the VectorStar back panel.
   d. Connect the **coaxial power plug** on the Power Supply Module to the AutoCal Module and the other end to AC power.
5. Once connected to power, the **Power LED** illuminates as green. When the AutoCal module has warmed up, the **LED** illuminates as blue.



1. VectorStar MS4645B or MS4647B VNA with V Test Port Connectors	4. AutoCal DB9 (m-m) Signal Cable connected to VNA Rear Panel Serial 10101 Port
2. 36585V-2F Precision AutoCal Module	5. V (f-f) Adapter connected to VNA Test Port 1
3. AutoCal Power Supply Module connected to AC Power	6. V (m-f) Test Cable connected to VNA Test Port

Figure 2-12. Precision AutoCal 36585V-2MF Cable Connections for True Thru

Schematically, the connections shown in Figure 2-12 above are shown in Figure 2-13.



Part 1—AutoCal True Thru Procedure

1. VNA Test Port 1 V (m)
2. V(f-f) Test Cable
3. 36585V-2MF AutoCal Module
4. V (m-f) Test Cable
5. VNA Test Port 2 V (m)
6. Resultant reference planes during first AutoCal procedure.

Part 2—AutoCal True Thru Procedure

7. After initial calibrations, user is directed to remove AutoCal module and connect test cables. At all times during the calibration procedure, the AutoCal module must be connected to the VNA and to AC power.
8. Connect the two Test Cables together to complete the thru calibration.

Figure 2-13. AutoCal True-Thru Connections

6. Navigate to the AUTOCAL SETUP menu:
   - MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | AutoCal | AUTOCAL PORTS | 2-Port Cal | AUTOCAL SETUP
7. On the AUTOCAL SETUP menu, if the Cal Type, Thru Type, and Module Orientation display buttons do not show the correct values, select the Modify Cal Setup button.
   a. The MODIFY 2-PORT AUTOCAL SETUP dialog box appears.
   b. Change the settings as required.
   c. In the Cal Type Select area: Select the Full Two Port radio button.
   d. In the Thru Select area, select the True Thru radio button and the THRU INFO dialog box appears.
   e. Enter information about the thru line. Enter 0 length and 0 dB/mm loss (and a reference frequency of 0, which forces it to use that loss at all frequencies) and 50 ohm impedance. Select OK to close the dialog box.
   - Select the Auto Sense Module Orientation check box which allows the AutoCal module determine the left/right cable identification.
   - Click OK to close the dialog box.

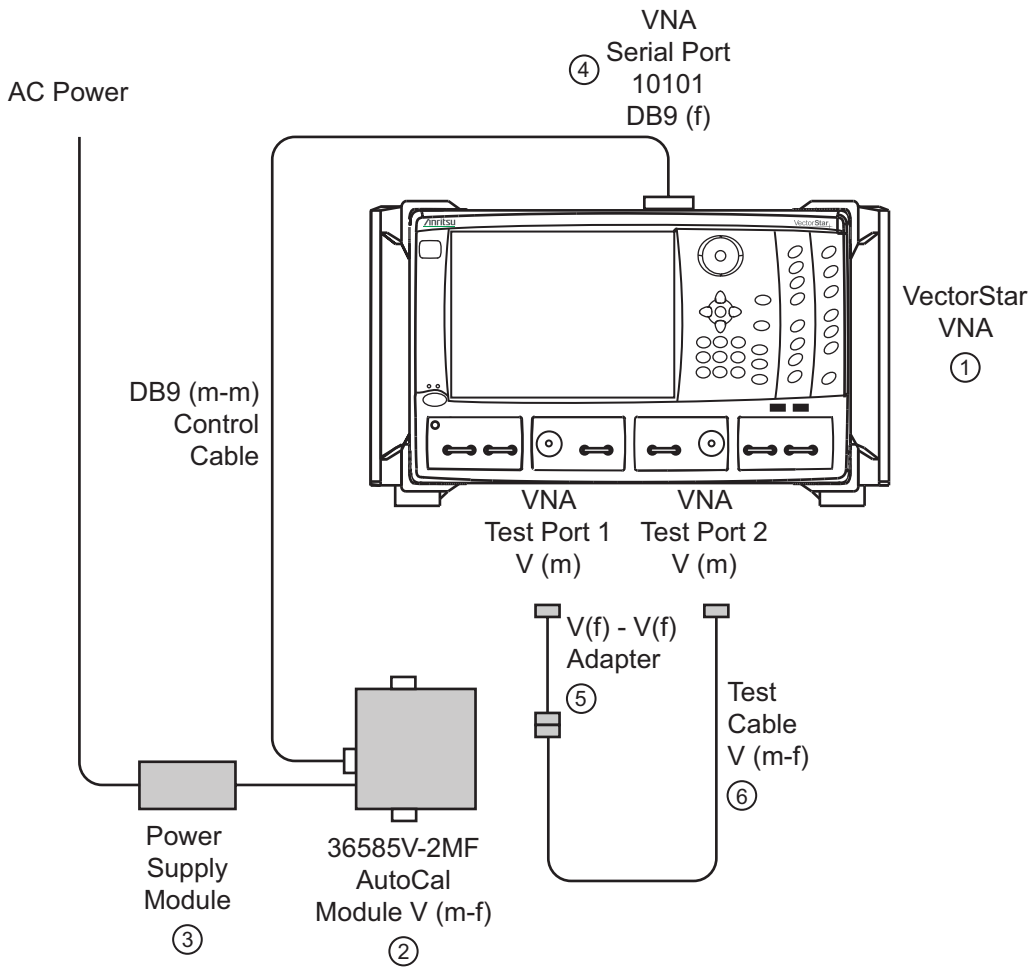
Note	All existing system setups such as IF Bandwidth, Averaging, and Power Level will be applied during the calibration procedure.

8. When ready, click the Begin Cal button.
9. A status dialog box with a progress bar appears after an AutoCal sequence has started. The status messages define has far the program has progressed and if any user actions are required.
10. At any time, the AutoCal sequence can be cancelled by clicking the dialog box Abort button.
11. A dialog box will appear when the true thru is to be connected.
12. Connect the **Test Port 1 test cable** to the **Test Port 2 test cable**.



If the AutoCal module is connected incorrectly, the AUTOCAL MODULE NOT DETECTED warning message appears. Correct connections as required and click the Retry button.

Figure 2-14. AUTOCAL MODULE NOT DETECTED Warning Message



1. MS4645B or MS4647B VNA with V Test Port Connectors	4. AutoCal DB9 (m-m) Signal Cable connected to VNA Rear Panel Serial 10101 Port
2. 36585V-2MF AutoCal Module	5. V (f) to V (f) Adapter connected to Test Port 1 V (m)
3. AutoCal Module Power Supply connected to AC Power. AutoCal module and connect test cables. At all times during the calibration procedure, the AutoCal module must be connected to the VNA and to AC power.	6. V (m-f) Test Cable connected to V (f-f) adapter and Test Port 2 V (m)

Figure 2-15. AutoCal True-Thru Connections—Module Removed

13. After connecting the thru, select Continue.

14. When the auto calibration is complete, a status message appears with a statement about assurance passing or failing. On the CALIBRATION menu, the Cal Status field button shows ON.

15. Closing the dialog box returns to the CALIBRATION menu.

2-10 AutoCal Module Characterization

Typically, characterization is performed by Anritsu since the process can be very carefully controlled for maximum accuracy. In certain cases, the customer may wish to perform the characterization themselves, but it is important to note that all specifications for the calibration are void and the customer takes responsibility for performing a characterization of adequate quality. With that caveat, the process for performing a characterization is as follows.

Characterization Procedure

1. Setup the instrument for the frequency range, point count, power level, and IFBW desired. It is particularly important to use as many points as reasonable in order to reduce interpolation needs.
2. Perform as high a quality manual calibration as possible at the reference planes that will be connected to the AutoCal module. LRL/LRM is recommended if possible.
3. Connect the AutoCal module to the reference planes, apply power, and connect the control cable to both the AutoCal module and to the MS464xB Series VNA. Allow the AutoCal module to reach operating temperature so the blue **Operate LED** illuminates.
4. Select the AUTOCAL CHARAC. menu (shown below) to start characterization. The VNA will automatically switch the unit through its various states and characterize them. Alternatively, the connected test port order may be specified and may be necessary if there is substantial loss in the test setup (note that this may indicate an accuracy hazard).
   - MAIN | System | SYSTEM | Utility | UTILITY | AutoCal Characterization | AUTOCAL CHARAC.
5. Save the characterization file as appropriate. Note that the characterization file will be tied by serial number to the particular AutoCal module that was characterized.

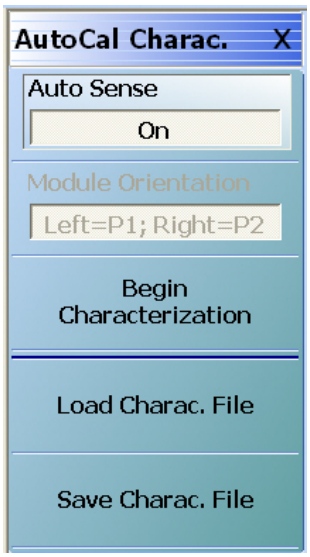


Figure 2-16. AUTOCAL CHARAC. (AUTOCAL CHARACTERIZATION) Menu

2-11 Working with Older AutoCal Modules

Auto Calibration Units 36581x and 36582x

Older Anritsu auto calibration units are supported by the MS464xB Series VNAs but with somewhat lower performance and different functionality. Fewer reflection states are available in these older models so an over-determined algorithm is not employed and a different characterization process is typically employed; both of which result in lower accuracy (see data sheets for specific figures).

These differences are practically less significant at 20 GHz and below where one may use the 36581X; they may be more significant at higher frequencies. The 36582X is an electromechanical switch-based unit, unlike the other modules, so it may suffer from additional degradation due to switch repeatability.

The characterization load process is similar. It may be required to transfer the characterization file from a floppy to a USB stick for transfer to the VNA (or transfer by network or GPIB).

A maximum of 1601 characterization points are possible on the older platforms so interpolation errors may be higher. These AutoCal units may be re-characterized with the MS464xB Series VNA to increase point count with the caveats discussed in "AutoCal Module Characterization" on page 2-26.

Autosensing of the AutoCal orientation is not possible. The current orientation must be entered correctly or large errors will result.

Adjustable switch averaging is no longer supported. A factory recommended level will automatically be invoked for electromechanical units.

2-12 Adapter Removal—M-M or F-F Reference Plane with a M-F AutoCal

Adapter Removal Overview

Adapter removal for AutoCal modules primarily refers to the case of connector gender incompatibility when it is not desired to use test port converters such as when the user has a M-F AutoCal module and M-M reference planes are required. A separate menu item is provided for AutoCal adapter removal to speed up the process since fewer manual steps are needed. In this calibration sequence, one uses an adapter (that can mate the desired reference plane connectors) as part of the calibration.

Two possible scenarios are covered; both use a pair of calibration sequences to remove the effects of the adapter. In both of the cases below, it is assumed all connectors are from the same family. If not (e.g., one is using a special inter-series AutoCal unit), then this AutoCal-specific adapter removal technique may not be applied; see the standard adapter removal section.

This procedure performs an AutoCal procedure when an adapter is required to accommodate either the AutoCal unit or the DUT connector genders and a M-M reference plane is required.

Required Equipment

- **VNA**

  MS4642B or MS4644B VNAs with K Test Port Connectors

- **AutoCal Module**

  Precision AutoCal Module, 36585K-2MF, with K (f-m) connectors. Includes the necessary Power Supply Module with cords to AutoCal Module and to AC power

- **Test Cable**

  A test port cable with K (m-f) connectors

- **K (f-f) Adapter**

  A Matched K Adapter, 36583K with K (f-f) connectors.

Prerequisites

The following prerequisite procedures have already been accomplished:

- AutoCal Characterization file previously loaded.
- VectorStar powered up.
- Required settings for Frequency Start, Frequency Stop, Number of Points, and CW Mode configured.
- Optional settings as required for Segmented Sweep, IF Bandwidth, Averaging, and Port Power configured.

Procedure

1. Make the necessary cable connections between the AutoCal Module, its Power Module, AC power, and the VectorStar back panel.
   - When the blue **Operate LED** is illuminated, the module has warmed up and is ready for calibration operations.
2. Connect the **36583K (f-f) Adapter** to the **AutoCal K (m) connector**.

- Consider the Adapter and AutoCal module as an "assembly" for the duration the AutoCal procedure.

Note	Once assembled, do not break the connection between the adapter and the AutoCal module, do not disconnect the assembly from the **VNA Serial 10101 Port**, and do not disconnect the assembly from AC power. If the connection between the adapter is broken before the AutoCal procedure is completed, the entire calibration is invalidated and must be repeated.

3. Connect the remaining **36583K (f) connector** to the VectorStar **Test Port 1 K (m) connector**.
4. Connect the **Test Cable K (f) connector** to the VectorStar **Test Port 2 K (m) connector**.
5. Connect the **Test Cable K (m) connector** to the **AutoCal Module K (f) connector**.

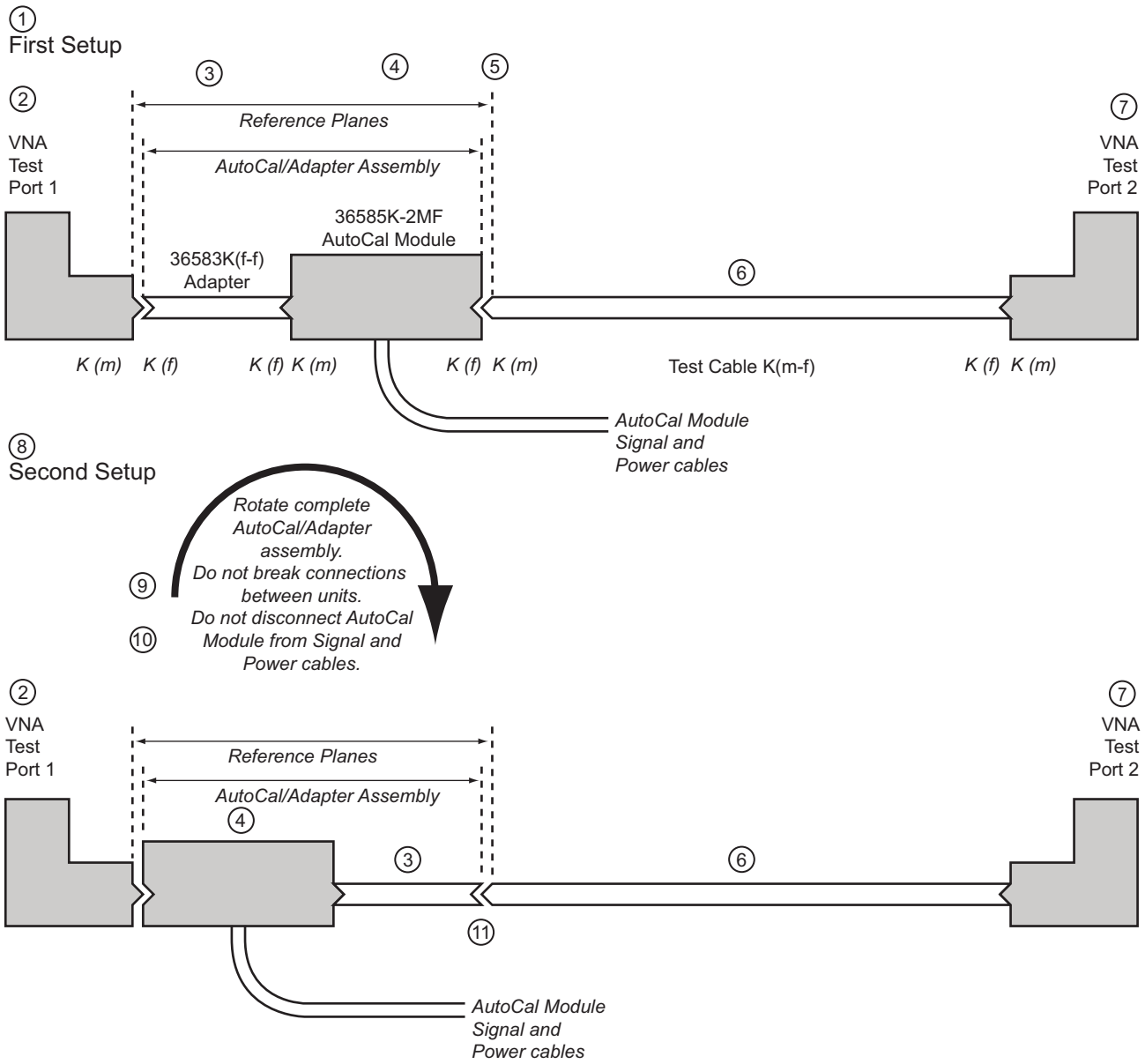


Figure 2-17. Precision MF AutoCal 36585K-2MF, Adapter Removal, Internal Thru (1 of 2)

Part 1—AutoCal Adapter Removal Procedure	Part 2—AutoCal Adapter Removal Procedure
1. First setup for the Adapter Removal procedure. 2. VNA Test Port 1 3. 36583K (f-f) Adapter 4. 36585K-2MF AutoCal Module—For the duration of the calibration, the K (f-f) Adapter and the AutoCal Module must be connected as an assembly. Do not disassemble or disconnect from its Power and Signal cables. 5. Resultant calibration reference planes. 6. K (m-f) Test Cable 7. VNA Test Port 2	8. Second setup for the Adapter Removal procedure. 9. After the first calibration, rotate the complete assembly so that the AutoCal K (f) connector is connected to VNA Test Port 1. Do not disconnect the Adapter from the AutoCal module. 10. Do not disconnect the AutoCal Module from its Power or Signal cables. 11. Connect the 36583 K (f-f) adapter to the Test Cable K (m-f). The reference planes remain in place. The user is guided through the remaining steps of the procedure.

Figure 2-17. Precision MF AutoCal 36585K-2MF, Adapter Removal, Internal Thru (2 of 2)

6. Navigate to the AUTOCAL SETUP menu.

- MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | AutoCal | AUTOCAL PORTS | 2-Port Cal | AUTOCAL SETUP

7. If the Cal Type, Thru Type, and Module Orientation display buttons do not show the correct values, select the Modify Cal Setup button.

- The Modify AutoCal Setup dialog box appears.

8. Make the following changes to the Modify AutoCal Setup dialog box settings:

a. Select the Auto Sense Module Orientation check box which allows the AutoCal module determine the left/right cable identification.

b. In the Select Cal Type area, select the Adapter Removal radio button. The For Adapter Removal Only area becomes available.

c. In the Thru Setup area, select the Internal Thru radio button and Adapter Removal.

d. In the For Adapter Removal Only area, enter the estimate of electrical length (in mm) and select which AutoCal port the adapter is attached to.

e. Select OK to close the dialog box. The AUTOCAL SETUP menu reappears with new values for Cal Type, Thru Type, and Module Orientation.

Note	All existing system setups such as IF Bandwidth, Averaging, and Power Level will be applied during the calibration procedure.

9. When ready, click the Begin Cal button.

10. If the AutoCal module is connected incorrectly, the AutoCal Module Not Detected warning message appears. Correct connections as required and click Retry.

11. A status dialog box with a progress bar appears after an AutoCal sequence has started. The status messages define has far the program has progressed and if any user actions are required.

12. At any time, the AutoCal sequence can be cancelled by clicking the dialog box Abort button.

- When the first AutoCal process is complete, the user is prompted to reverse the Adapter/AutoCal assembly.

Note	Once assembled, do not break the connection between the adapter and the AutoCal module. If the connection between the adapter is broken before the AutoCal procedure is completed, the entire calibration is invalidated and must be repeated.

13. The instrument will prompt the user to reverse the module-adapter assembly.
14. As shown above in Figure 2-17, “Precision MF AutoCal 36585K-2MF, Adapter Removal, Internal Thru” on page 2-29 above, disconnect the **Adapter** from **Test Port 1** and the **AutoCal Module** from the **Test Cable** on Test Port 2.
15. Reverse the **Adapter/AutoCal Module assembly** so that the Adapter end is pointing towards Test Port 2.
16. On the AutoCal module, connect the free **AutoCal Module K (f) connector** to Test Port 1 K (m).
17. On the Adapter, connect the free **Adapter K (f) connector** to the Test Cable connected to Test Port 2.
18. When the auto calibration is complete, a status message appears with a statement about assurance passing or failing. Closing the dialog will return to the regular menu system. On the CALIBRATION menu, the Cal Status field button shows ON.

Chapter 3 — SOLT/SOLR Calibration

3-1 SOLT/SOLR Introduction

This chapter describes calibration procedures using the SOLT/SOLR calibration algorithms. One of the more common calibration algorithms is based on Short-Open-Load-Thru. This is a defined-standards calibration meaning the behavior of all of the components is specified in advance via data or models.

Since the behaviors of all standards are known, by measuring them with the VNA we can define all of the error terms. The load behavior largely sets the directivity terms, the short and open together largely determine source match and reflection tracking and the thru largely determines transmission tracking and load match.

Related Chapters

The following contains content related to SOLT/SOLR Calibration:

- Chapter 16, “Reciprocal Measurements”

3-2 Definitions

Shorts

Shorts can be defined by a model consisting of a transmission line length and a frequency-dependent inductance or with a .s1p file describing the reflection coefficient.

Opens

Opens can be defined by a model consisting of a transmission line length and a frequency-dependent capacitance or with a .s1p file describing the reflection coefficient.

Loads

Loads can be defined by a model consisting of a transmission line length, a shunt capacitance, a resistance and a series inductance or with a .s1p file describing the reflection coefficient.

Note that a sliding load can be used in lieu of a fixed load. The sliding load is based on a sliding termination embedded in an airline and the transmission line properties of that airline are used to deduce a more nearly perfect synthetic load. Because of the transmission line dependence, a fixed load is also needed at low frequencies (below 4 GHz for V connectors (shorter sliding load) and below 2 GHz for others).

The .s1p-based load definition can perform as well as a sliding load in most cases and will almost always outperform a simpler modeled definition.

Thru

Modeled as a transmission line length with some frequency-dependent loss. A root-f frequency dependence of that loss is assumed. A .s2p-defined thru is also possible where loss and mismatch are used. Interpolation and extrapolation of the .s2p data will be used to complete the calibration. A *characterize thru* option is also available to help generate such an .s2p file based on a network extraction method (Type B, see Chapter 9 of this guide for more information).

Reciprocal

The thru can sometimes be replaced by a unknown but reciprocal network (like an adapter or a fixture) when an actual thru connection is not practical. The accuracy will be somewhat less than if an actual thru could have been used but will be better than assuming a poor thru is a good one.

Notes on .s1p Cal Kit Definitions:

- These are provided with some Anritsu calibration kits when Option 3 or Option 4 is ordered. The data files shipped with those calibration cal kits also include a .lst file that allows loading of .s1p files for all calibration kit components simultaneously.
- The user can also generate their own .s1p files for calibration kit components based on measurement, electromagnetic modeling or some other means. If based on measurement, it is recommended that a higher accuracy calibration form the basis of the characterization (e.g., LRL based on beadless airlines for coaxial situations).
- The use of a .s1p-based calibration can substantially improve uncertainties in measurement. As an example, consider the use of an Anritsu 3652A K calibration kit to perform a reflection measurement between 20 and 40 GHz. The uncertainties are plotted below both using the old polynomial model and using the .s1p definition available with Option 3 or Option 4 on that cal kit. Consult the *MS464xB Technical Data Sheet*—11410-00611 for a more complete comparison.

  The uncertainties for a K-calibration-kit-based reflection measurement are plotted in Figure 3-1 for different calibration kit definitions (the same calibration kit components),

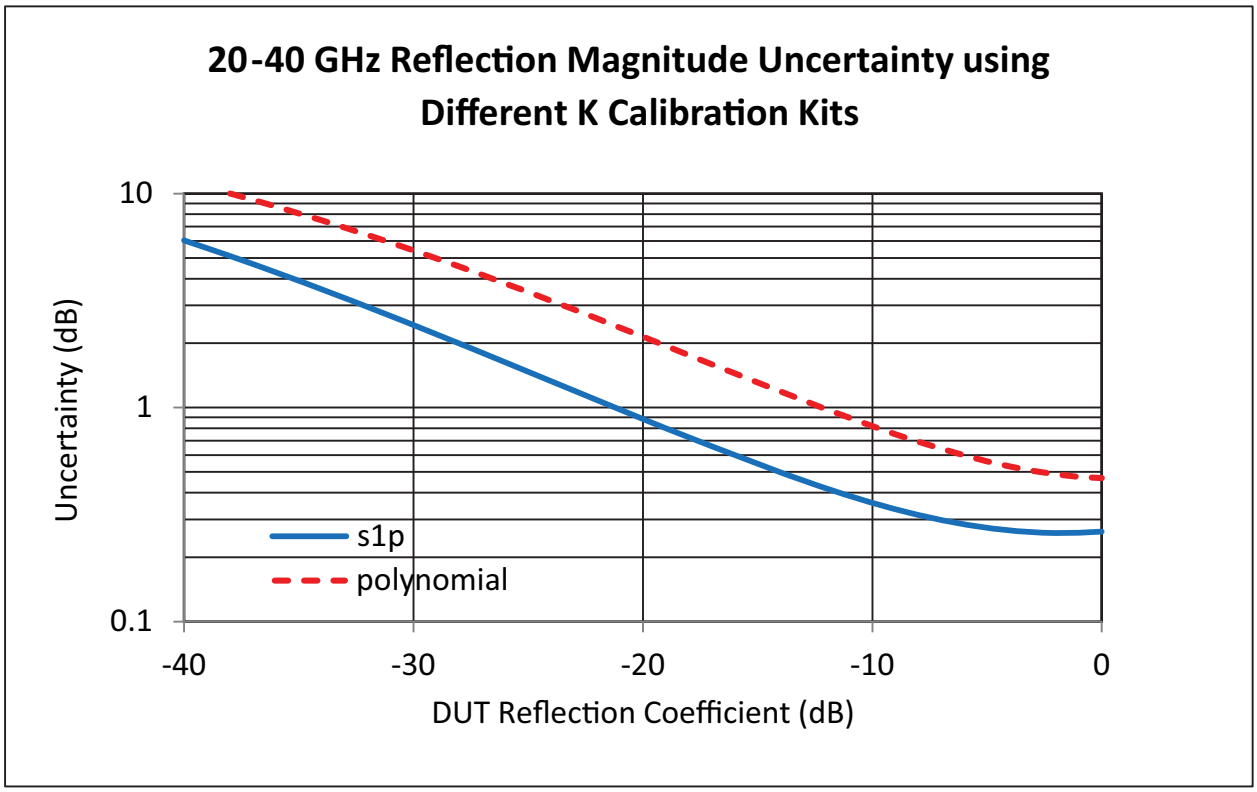
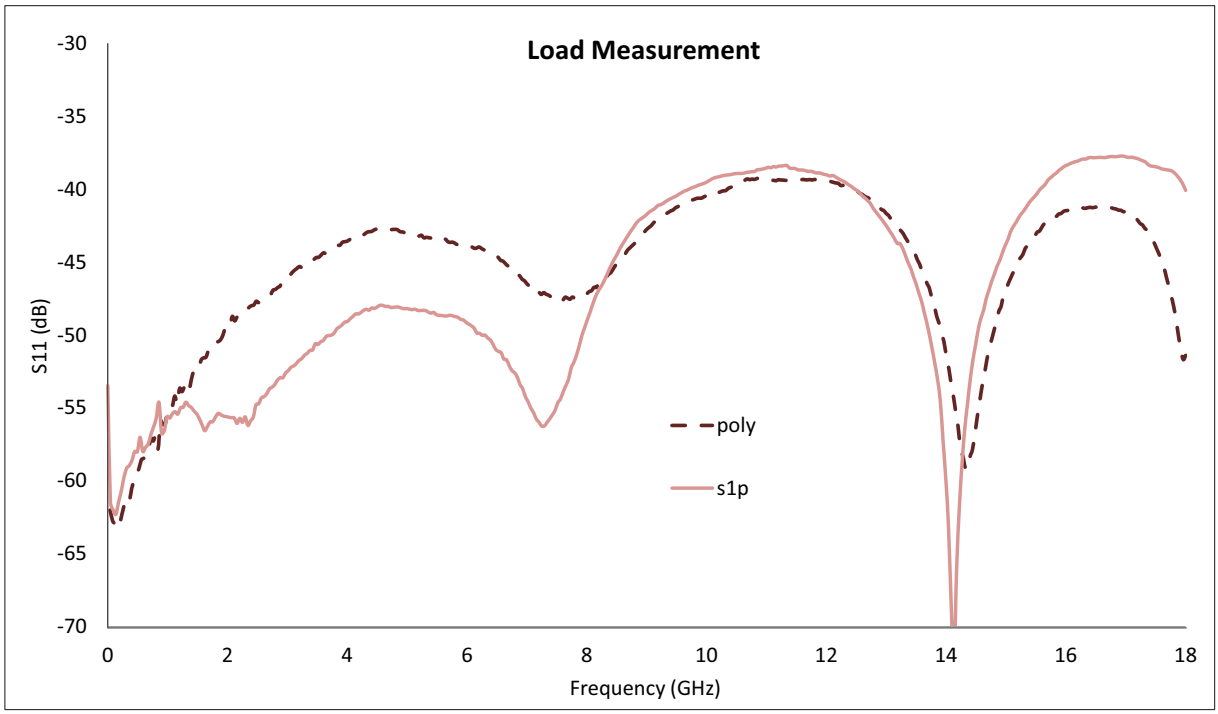


Figure 3-1. Comparison of Reflection Magnitude Uncertainty with s1p vs. Polynomial Definitions

The results can get more confusing when measuring very small reflections since the uncertainties are becoming proportionately larger in both cases. Consider the measurement of a load (not used during calibration). The measurements using two different calibration kit definitions are shown in Figure 3-2 on page 3-3 below along with the uncertainties for those measurements.



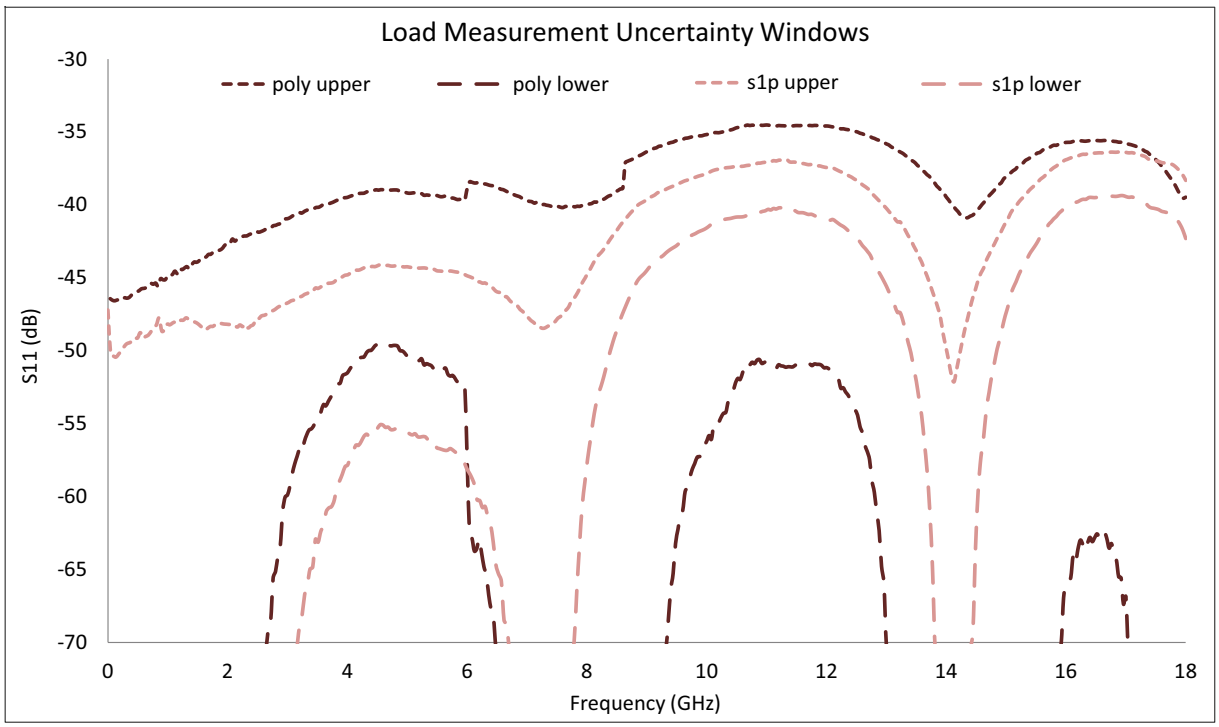
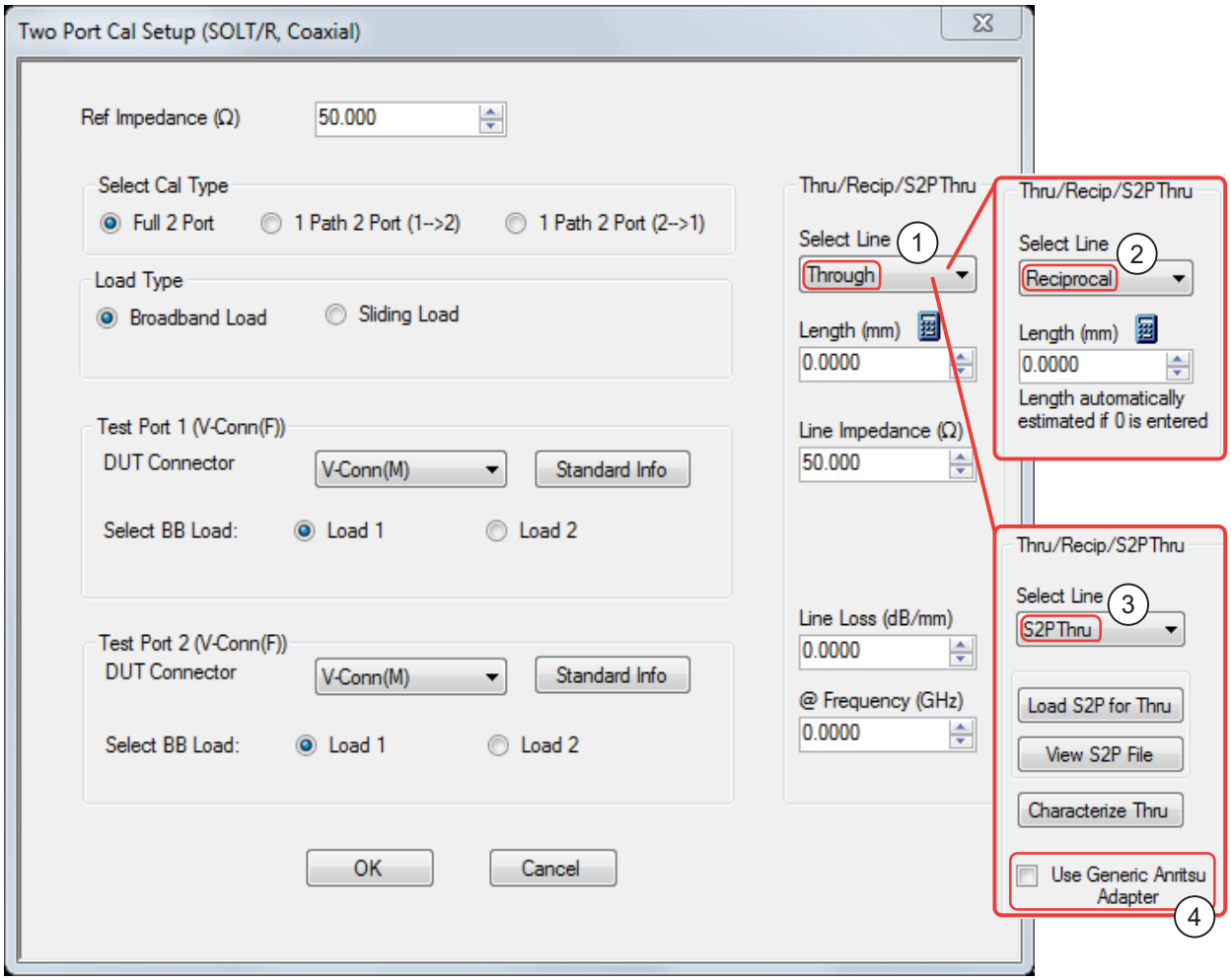


Figure 3-2. Load Measurement Comparing Two Different Calibration Kit Definitions (Polynomial vs. s1p)

Note that the absolute return losses differ considerably, particularly at low frequencies, but because the reflections are so low, these differences are within the uncertainties. Note also (very clear at higher reflection levels) how much narrower the uncertainty windows are for the .s1p-based calibration. Below about –50 dB reflection coefficients, the uncertainties are quite high for both methods.

3-3 SOLT/SOLR Calibration Setup

The coaxial setup dialog for SOLT (and SOLR), full 2-port calibration is shown in below.



1. Thru selected allows user entries for length, line impedance, line loss and frequency.
2. Reciprocal selected allows user entry for length.
3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).
4. Checking the Use Generic Anritsu Adapter checkbox will automatically load a default .s2p file for a generic Anritsu adapter. Note that selecting the checkbox disables the Load S2P for Thru and Characterize Thru buttons.

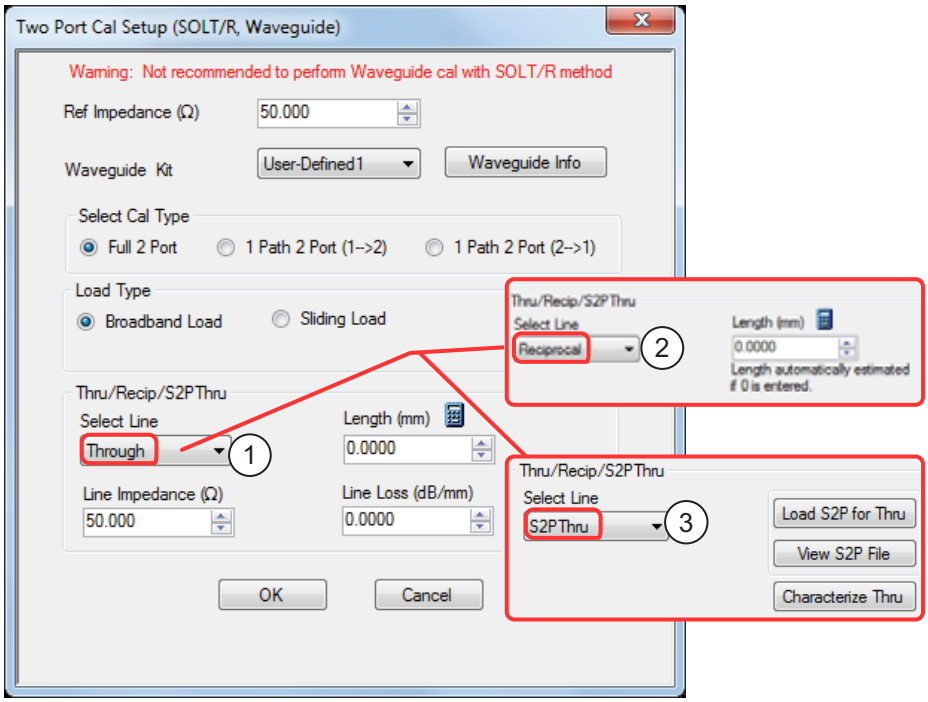
Figure 3-3. TWO PORT CAL SETUP (SOLT/R, COAXIAL) Dialog Box

In the coaxial setup dialog, the connector types for both ports are selected as well as the thru details and the type of load to be used. For one port calibrations, only one of the port definitions (unless reflection-only calibrations are being performed for both ports 1 and 2) will be present. For a 1-path 2-port cal, one of the Test Port definition sections will not be shown.

The setup dialog above is for coaxial and non-dispersive line types. For waveguide and microstrip, a few things change:

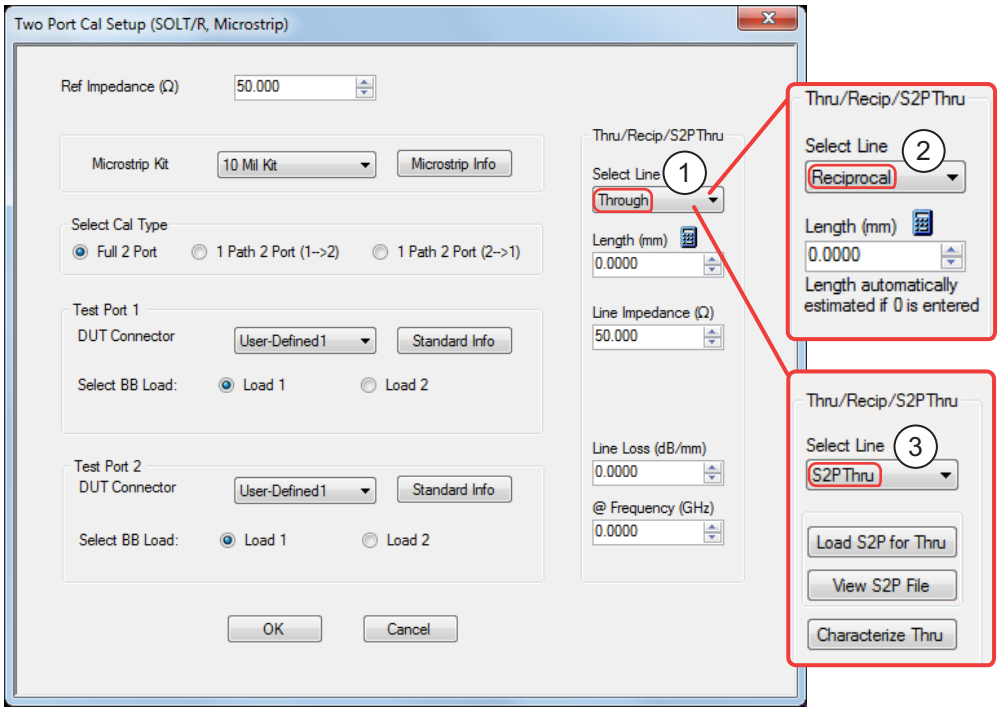
- Fewer cal kits are factory-defined and more are user-defined
- The media must be part of the definition (cutoff frequency and dielectric constant for waveguide; line width, substrate height, and substrate dielectric constant for microstrip)
- SOLT is not recommended for waveguide due to the difficulty in modeling and open standards

A waveguide SOLT setup is shown in Figure 3-4 and a setup for microstrip is shown in Figure 3-5.



1. Thru Line selected allows user entries for length, line impedance, line loss, and frequency.
2. Reciprocal selected allows user entry for length.
3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 3-4. TWO PORT CAL SETUP (SOLT/R, WAVEGUIDE) Dialog Box



1. Thru selected allows user entries for length, line impedance, line loss and frequency.
2. Reciprocal selected allows user entry for length.
3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 3-5. TWO PORT CAL SETUP (SOLT/R, MICROSTRIP) Dialog Box

The STANDARD INFO dialog for SOLT (and SOLR) is shown in Figure 3-6, using a V-connector as an example.

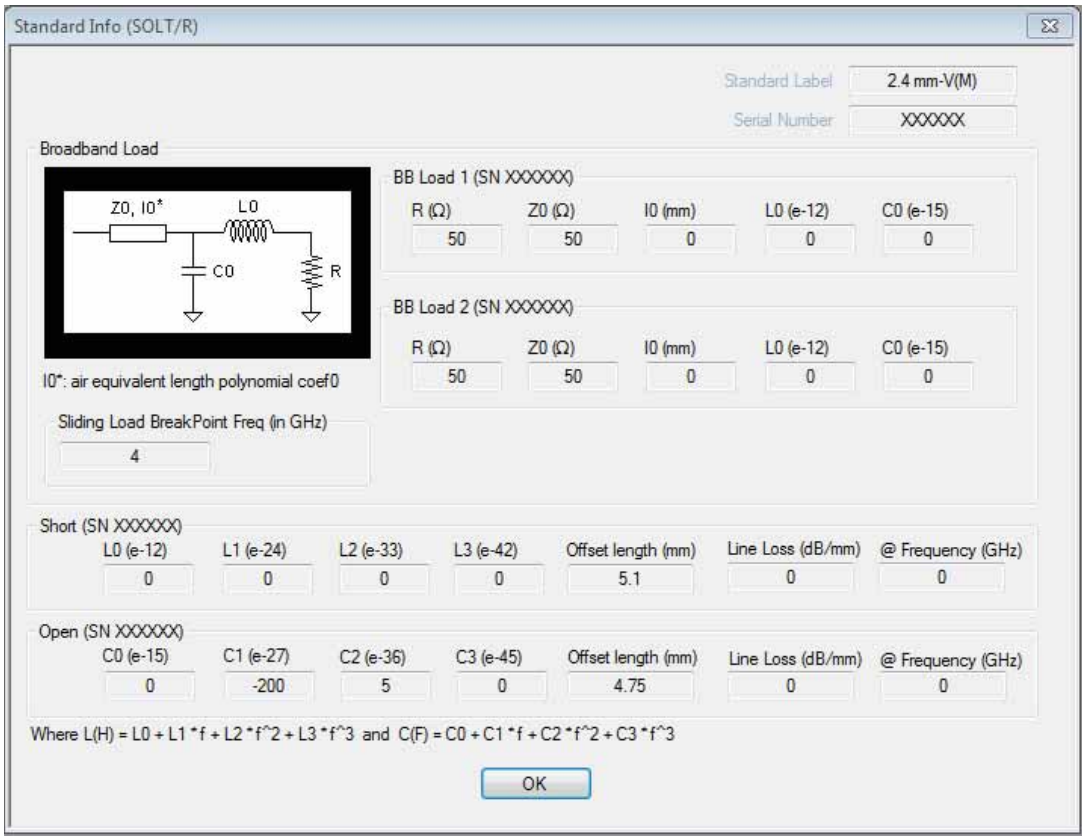


Figure 3-6. STANDARD INFO (SOLT/R)—V-Connectors (M)

For cal kits loaded from Anritsu cal kit files, the model terms are not editable. When using user-defined cal kits, the model terms can be edited.

The insertion loss of the opens and shorts can be defined. A loss variable is entered in dB/mm (which will be applied against the offset length) and the frequency (f_{ref}) at which that loss value applies. For other frequencies, the loss is scaled by sqrt(f/f_{ref}). If a 0 reference frequency is entered, the loss value is assumed to be constant with frequency. Sometimes, the loss of reflect standards is expressed as gigaohm/second at 1 GHz.

To convert those values to dB/mm, use this equation:

$$\text{dB/mm} = \frac{\text{loss in gigaohms/s} \times 0.014476483}{Z_0}$$

Equation 3-1.

Where Z_0 is the characteristic impedance (usually 50 ohms). Also, enter 1 GHz for the reference frequency as the scaling is the same.

The standards information for microstrip does not change but the microstrip media information must be either user-defined (Figure 3-7) or selected from an Anritsu microstrip cal kit (Figure 3-8, typically used with Anritsu Universal Test Fixtures).

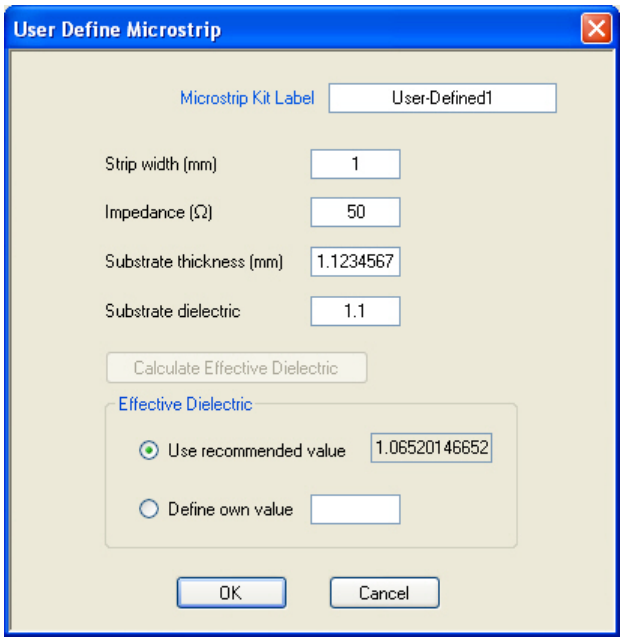
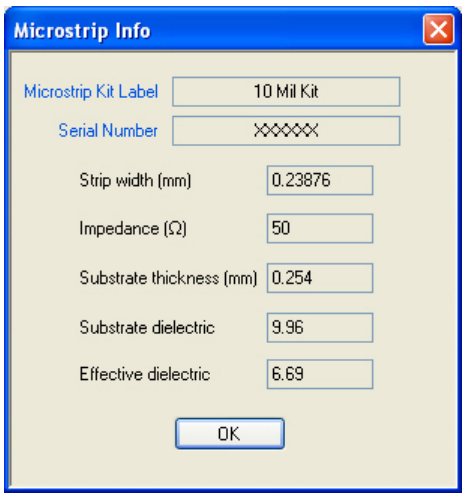


Figure 3-7. USER-DEFINED MICROSTRIP Data Input Dialog Box

Effective dielectric constants can be entered, or the recommended value can be selected.



Typical values for an Anritsu-defined 10 Mil Microstrip Kit

Figure 3-8. MICROSTRIP INFO 10 Mil Kit Dialog Box

For waveguide, the model parameters and the media parameters are combined in one dialog (Figure 3-9).

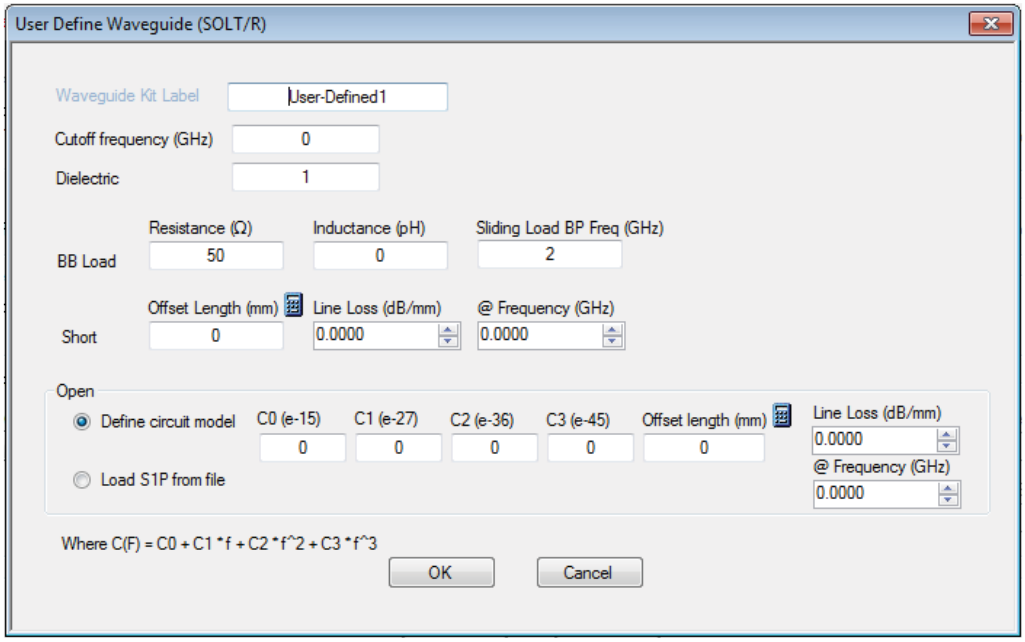


Figure 3-9. USER-DEFINED WAVEGUIDE (SOLT/R) Information Dialog Box

Note	Reciprocal measurements (SOLR vs. SOLT) are covered in more detail in Chapter 16, “Reciprocal Measurements”.

3-4 SOLT/SOLR Calibration Example

The following example presumes an MS4647B (70 GHz unit) with V connectors. A different connector can be selected in step 4 if a different model/configuration is being used. It is assumed that a M-F cable is connected to port 2 so that a M (port 1) and F (port 2) reference plane pair is available. In this example, a full 2-port SOLT calibration will be performed although a number of other options are discussed along the way. The implications of these options are discussed in the calibration overview section.

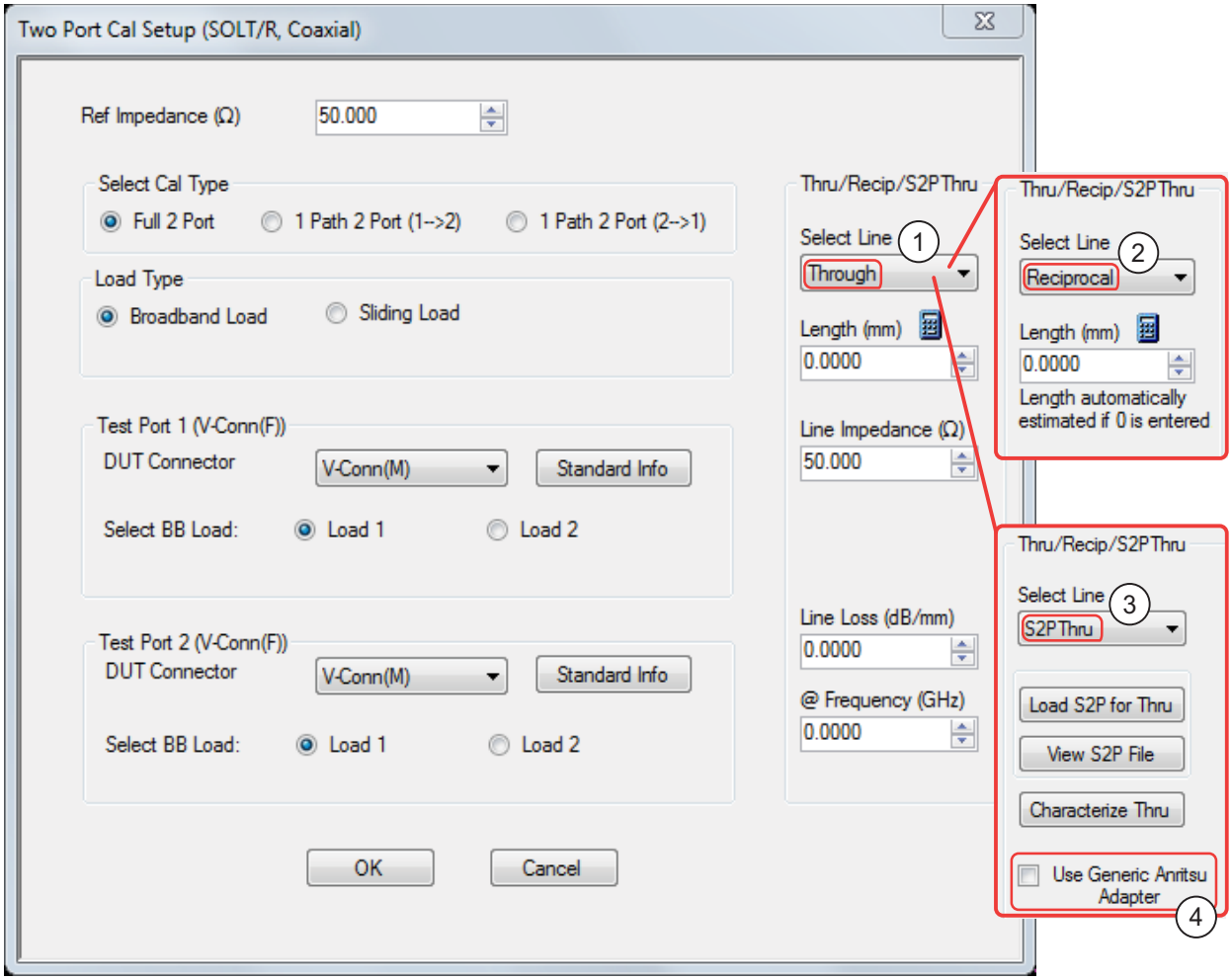
1. Setup the desired frequency range (Frequency menu), power (Power menu) and IFBW/averaging (Averaging menu). As a default, the IFBW will be 1 kHz and the averaging will be off which is adequate for many applications. The default power level will vary depending on instrument model and options but will often be adequate for all passive and many active device measurements.
2. Navigate to the TWO PORT CAL menu.
   - MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | Manual Cal | MANUAL CAL | 2-Port Cal | TWO PORT CAL

Note	If a previous calibration exists, the Thru Update button will be active. See "Through (Thru) Update" on page 10-1 for more information.

3. Select Modify Cal Setup. The CAL SETUP menu appears. On this level, select a cal method of SOLT/SOLR and a line type of Coaxial.
4. Select Edit Cal Params; the TWO PORT CAL SETUP (SOLT/R, COAXIAL) dialog box appears (Figure 3-10 on page 3-11) which describes the calibration components:
   a. Reference impedance defaults to 50 ohms and can be left there commonly. This value is used for referencing the standards reflection coefficients and for reference plane shift and Smith chart calculations. The latter two items can be handled later using a per-trace definition of reference impedance. The standards definition process is not affected by that later per-trace reference impedance change.
   b. In the Load Type area, select broadband load. A sliding load can be used for better performance if one is available in the calibration kit. If low frequencies are included in the frequency range (< 4 GHz for V, < 2 GHz for K or GPC3.5), then a broadband load will be used in addition to the sliding load.
   c. In the Thru/Reciprocal/S2P Thru area, a zero-length (or mating) thru will be used. Set the Select Line field to Thru, 0 mm for the length and 0 dB/mm loss. Zero can also be entered for the reference frequency. When 0 is entered for this value, no loss scaling is employed and the entered loss value is used for all frequencies. If a reciprocal network was being used instead of a thru, use Reciprocal for the Select Line field and the length entered would serve as an estimate for root choice purposes.

      If s2p thru was selected, the .s2p file describing the thru s-parameters would be loaded and used during the calibration. A *characterize thru* option is also available using two previous reflection-only calibrations (with and without the thru element present) to calculate the .s2p file.
   d. For **Port 1**, select a DUT connector of V(f). Note the dialog will then indicate that port connector is V (m).
   e. For **Port 2**, select a DUT connector of V(f). The dialog will indicate a port connector of V(f).
   f. For both ports, select a BB load of Load 1. This selection is for certain firmware/cal kit versions where modeled loads are available and the distinction between loads is important.
   g. Select OK to close the dialog. Select Back at the bottom of the menu to return to the previous level and the TWO PORT CAL menu where two Reflective Devices buttons appear with six reflection standards on the two submenus.

Note	The menu calibration steps can be performed in any order. For this example, a top to bottom menu approach is assumed.



1. Through selected allows user entries for length, line impedance, line loss and frequency.
2. Reciprocal selected allows user entry for length.
3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).
4. Checking the Use Generic Anritsu Adapter checkbox will automatically load a default .s2p file for a generic Anritsu adapter. Note that selecting the checkbox disables the Load S2P for Thru and the Characterize Thru buttons.

Figure 3-10. TWO PORT CAL SETUP (SOLT/R, COAXIAL) Dialog Box

5. Select Port 1 Reflective Devices and the REFL. DEVICES PORT 1 menu. On the menu, measure the three reflection standards of Open, Short, and Load. Connect each standard in turn, and THEN, click the corresponding button. When all are done, click the Back button to return to the TWO PORT CAL menu.
6. Then select Port 2 Reflective Devices and the REFL. DEVICES PORT 2 menu where next three reflection standards are listed. When all six are done (and six check marks appear), click Done to return to the previous level and the TWO PORT CAL menu.
7. After measuring all six reflection standards, connect the cable to **Port 1** to complete the zero length thru. Now click on the Thru/Recip button where a check mark should appear after the sweep pair. Note that the displayed graphs may change during this step as the instrument must measure all four S-parameters of the thru line.

8. An optional isolation step using the Isolation (Optional) button and the linked ISOLATION menu is available but is generally not recommended. If desired, terminate **Port 1** and the end of the cable attached to **Port 2** before clicking on the Isolation (Optional) button if needed.

9. Click on Done. The calibration is now completed and turned on where the Cal Status button on the CALIBRATION menu is set to ON.

Chapter 4 — Offset Short (SSLT) Calibration

4-1 Introduction

This chapter describes calibration methods and procedures using the SSLT calibration algorithm.

Related Chapters

The following contains content related to Offset Short (SSLT) Calibration:

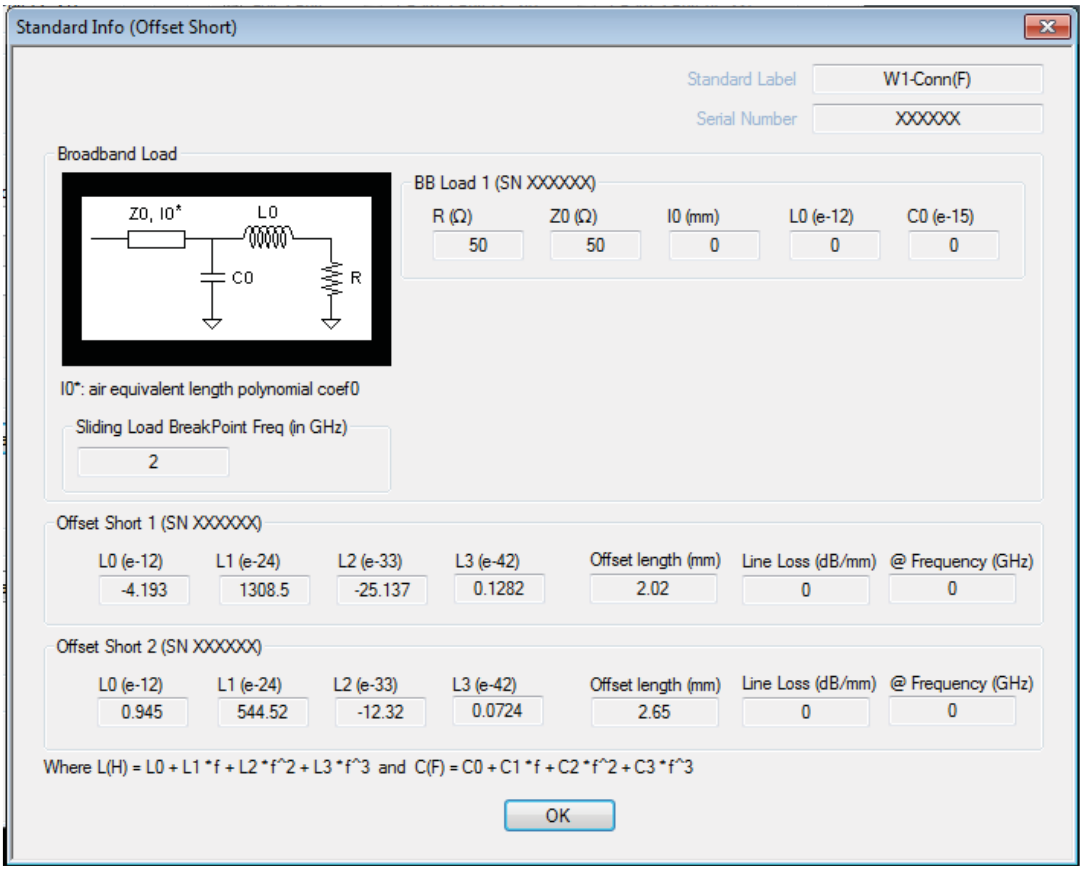
- Chapter 16, “Reciprocal Measurements”

4-2 SSLT Overview

The SSLT calibration differs from an SOLT calibration by the differing offset lengths between two shorts which are used to help define reflection behavior instead of an open and short. Because of this, the frequency range is limited since, at DC and at higher frequencies, these reflect standards will look the same. This method is most commonly used for waveguide problems where creating a stable, high reflection open standard is difficult, but there are certain coax and board-level or wafer-level situations where it is useful. The modeling constructs are about the same as for an SOLT calibration. From an error term perspective, the only difference is that the two shorts together now largely determine source match and reflection tracking behavior.

The electrical length difference between the shorts should be between 20 and 160 degrees over the frequency range of interest.

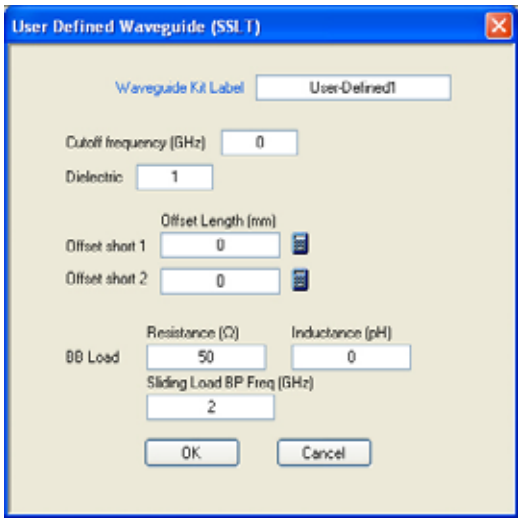
The top calibration kit definition dialog box for SSLT calibration is identical to the SOLT dialog (Figure 3-3, “TWO PORT CAL SETUP (SOLT/R, COAXIAL) Dialog Box” on page 3-4). The standards information dialog box is different and is shown in Figure 4-1.



Typical parameters for Offset Short Calibrations

Figure 4-1. STANDARD INFO (OFFSET SHORT) Information Dialog Box

Variations for other line types (waveguide or microstrip) are similar to those for SOLT. For waveguide, the media and standards information are combined as shown in Figure 4-2. Defined, un-editable values would be present for Anritsu-defined cal kits.



Typical parameters for an Offset Short Calibration

Figure 4-2. Waveguide Media and Standards Information

A simplified short model is used for waveguide, with only an offset length and no inductance terms since usually those terms are small.

In SSLT, it is also possible to define the reflection standards with .s1p files as discussed in Chapter 3 for SOLT. Anritsu calibration kits for SSLT are generally for waveguide and do not include these .s1p files but the user can create them for any calibration kit based on measurement, electromagnetic modeling or other means. If based on measurement, it is recommended that a higher accuracy calibration form the basis of the characterization (e.g., LRL based on beadless airlines for coaxial situations).

As with SOLT, an .s2p-defined thru is also an option for SSLT. A file describing the 'thru' can be loaded to account for its loss and mismatch. Interpolation and extrapolation will be used if the file frequency list does not match the current calibration requirement.

As with SOLT, and even without .s1p definitions, the insertion loss of the opens and shorts can be defined. A loss variable is entered in dB/mm (which will be applied against the offset length) and the frequency (f_{ref}) at which that loss value applies. For other frequencies, the loss is scaled by sqrt(f/f_{ref}). If a 0 reference frequency is entered, the loss value is assumed to be constant with frequency. Sometimes, the loss of reflect standards is expressed as gigaohm/second at 1 GHz.

To convert those values to dB/mm, use this equation:

$$\text{dB/mm} = \frac{\text{loss in gigaohms/s} \times 0.014476483}{Z_0}$$

Equation 4-1.

Where Z_0 is the characteristic impedance (usually 50 ohms). Also, enter 1 GHz for the reference frequency, as the scaling is the same.

4-3 SSLT Calibration Example

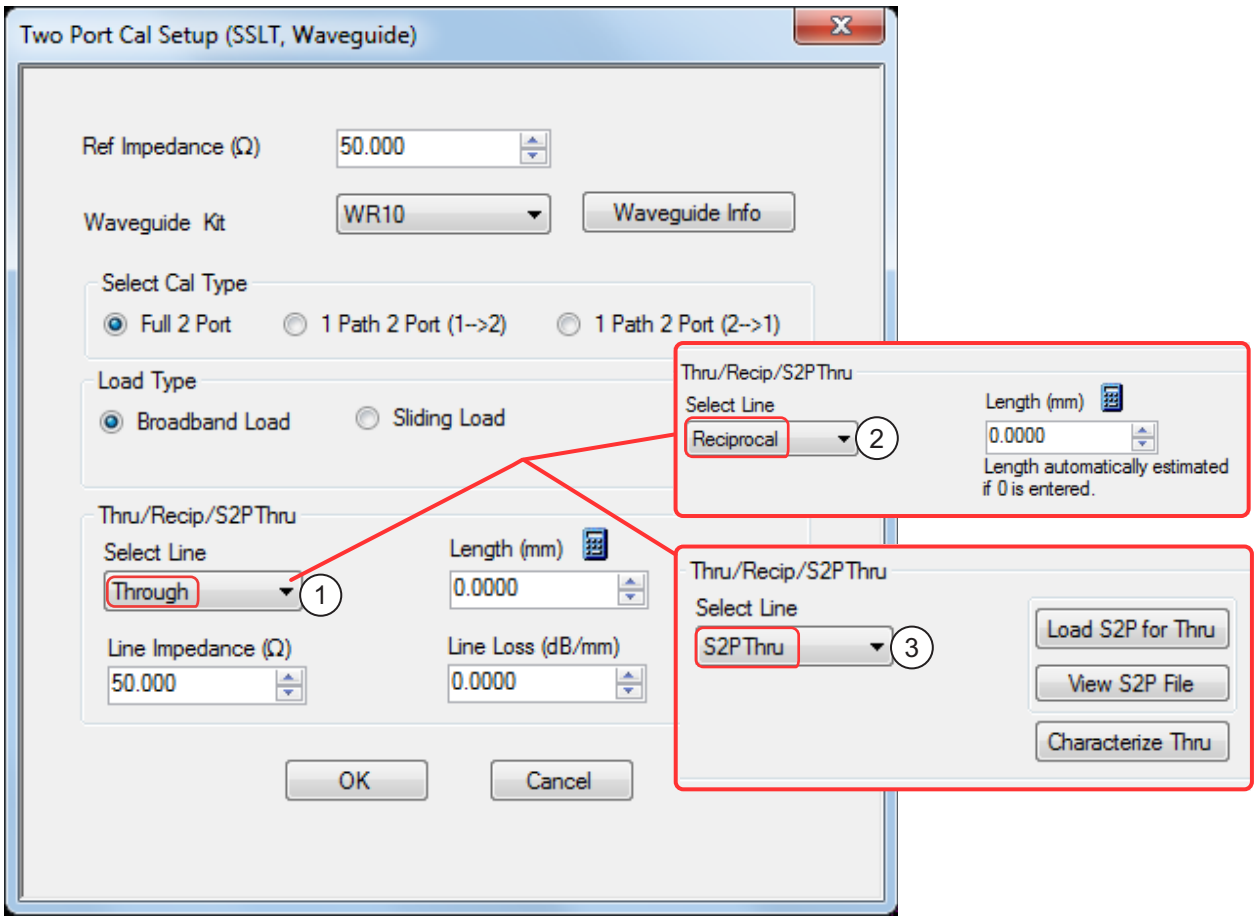
The following example presumes a MS464xB Series VNA is being used with waveguide adapters (since SSLT is commonly used with waveguide). WR-10 is the waveguide size used in the example, which presumes the use of mmWave extensions, but other waveguide kits can be used. In this example, a full 2-port SSLT calibration is performed although a number of other options are available. The implications of these options are discussed in Section 4-2 “SSLT Overview”.

Set the desired frequency range (Frequency menu), power (Power menu) and IFBW/Averaging (Averaging menu). The default power level varies depending on the instrument model and options, but will often be adequate for all passive and many active device measurements. The default IFBW is 1 kHz with averaging off, which is adequate for many applications.

1. Navigate to the TWO PORT CAL menu.
   - MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | Manual Cal | MANUAL CAL | 2-Port Cal | TWO PORT CAL

Note	If a previous calibration exists, the Thru Update button will be active. See “Through (Thru) Update” on page 10-1 for more information.

2. Select Modify Cal Setup, then on the CAL SETUP menu, select a Cal Method of SSLT and a Line Type of Waveguide.



1. Thru Line selected allows user entries for length, line impedance, line loss and frequency. 2. Reciprocal selected allows user entry for length. 3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 4-3. TWO PORT CAL SETUP (SSLT, WAVEGUIDE) Dialog Box

3. Select Edit Cal Params which will displays the TWO PORT CAL SETUP (SSLT, WAVEGUIDE) dialog box describing the calibration components (shown above).

   a. Reference impedance defaults to 50 ohms. Although this does not represent the waveguide impedance, it is commonly used for conventional Smith chart referencing. Reference impedance can be changed here for certain waveguide applications or it can be changed later on a per trace basis from the DISPLAY menu.

   b. Select Broadband Load. A sliding load can be used for better performance if one is available in the calibration kit. If low frequencies are included in the frequency range (< 2 GHz), then a broadband load will be used in addition to the sliding load.

   c. A zero-length (or mating) thru will be used, so enter Through for the line, 0 mm for the length and 0 dB/mm loss. Zero can be entered for the reference frequency when no loss scaling is employed and the entered loss value is used for all frequencies. If a reciprocal network was being used instead of a thru, the line selection would be Reciprocal and the length entered would serve as an estimate for root choice purposes.

      If s2p thru was selected, the .s2p file describing the thru s-parameters would be loaded and used during the calibration. A *characterize thru* option is also available using two previous reflection-only calibrations (with and without the thru element present) to calculate the .s2p file.

   d. For a waveguide kit, select WR10 (for this example). A user-defined kit can be selected from the pull-down menu, but the cutoff frequency, the dielectric, the short offset lengths, and the load impedance (relative to the reference impedance) will have to be specified.

   e. Select OK to close the dialog, then select BACK at the bottom of the menu to return to the previous level and the TWO PORT CAL menu.

4. The Port 1 Reflective Devices and the Port 2 Reflective Devices buttons display the REFL. DEVICES PORT 1 and PORT 2 menus. The six reflection standards to be measured (Short 1, Short 2, Load on each menu) will be displayed here along with the read-only fields showing the port connector choices. Connect each standard in turn, and THEN, click the corresponding button. When all six are done (and six check marks appear), click Done to return to the previous level and the TWO PORT CAL menu.

5. After measuring all six reflection standards, connect the test cable to **Port 1** to complete the zero length thru. Click on the Thru/Recip button to display the THRU/RECIP menu where a check mark should appear after the sweep pair. Note that the displayed graphs change during this step as the instrument must measure all four S-parameters of the thru line.

6. An optional isolation step is available but is not recommended. If desired, terminate **Port 1** and the end of the cable attached to **Port 2** before clicking on the Isolation (Optional) button and displaying the ISOLATIONS menu.

7. When all steps are successfully completed, the Done button is available where select returns to the CALIBRATION menu where the Cal Status button is set to ON.

8. The calibration is now completed and turned on.

Note	Much like SOLT/SOLR, there is another version of SSLT using a reciprocal in place of a thru called SSLR. Reciprocal measurements (using SOLR vs. SOLT as an example) are covered in more detail in Chapter 16, “Reciprocal Measurements”.

Chapter 5 — Triple Offset Short (SSST) and Multiple Offset Short (mSSS) Calibrations

5-1 SSST Introduction

This chapter describes calibration using the triple offset short (SSST) algorithm. The next step in this progression is to remove the load so that the entire reflection space is defined by three shorts of varying offset lengths. The individual short definitions are the same as for an SOLT calibration.

- SSST and Reflectivity Error Terms
- SSST/SSSR Calibration Example

A generalization of this approach is multiple offset short (mSSS) where up to 10 different shorts can be used per-portper-port to extend bandwidth and improve immunity to repeatability challenges. This mSSS calibration uses a least-squares solving approach to optimize the weighting of the calculation at each frequency.

- mSSS Calibration
- mSSS Calibration Example

Related Chapters

The following contains content related to Triple Offset Short (SSST) Calibration:

- Chapter 16, “Reciprocal Measurements”

5-2 SSST and Reflectivity Error Terms

With an SSST calibration the three shorts together determine all of the reflectivity error terms (directivity, source match and reflection tracking). This calibration is more band-limited than the double offset short method. If short1 is defined as having the smallest offset length, and short3 to the longest offset length, then two variables can be defined:

$$A = L_{offset_short2} - L_{offset_short1}$$
$$B = L_{offset_short3} - L_{offset_short2}$$

Equation 5-1.

The electrical length equivalents of A and B should generally be between 20 and 85 degrees over the frequency range of interest. This is not sufficient in itself since one will also require that A+B (which represents the difference between short1 and short3) also be constrained:

$$40 < \frac{720 \cdot f \cdot B}{v_{ph}} < 180$$

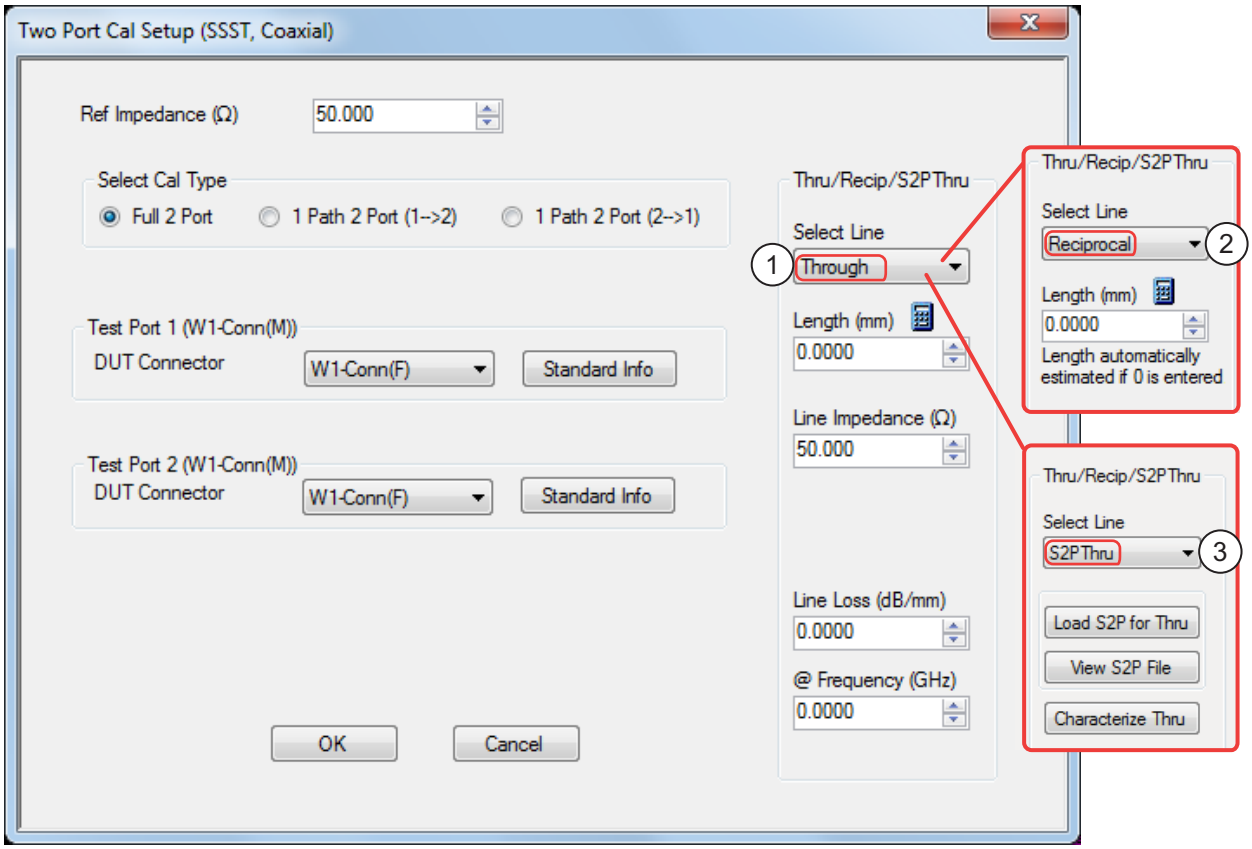
$$40 < \frac{720 \cdot f \cdot A}{v_{ph}} < 180$$

$$40 < \frac{720 \cdot f \cdot (A + B)}{v_{ph}} < 320$$

Equation 5-2.

Since the only standards needed are shorts, this method is attractive for millimeter wave applications and for certain board-level and wafer-level calibrations where other types of standards are difficult to manufacture.

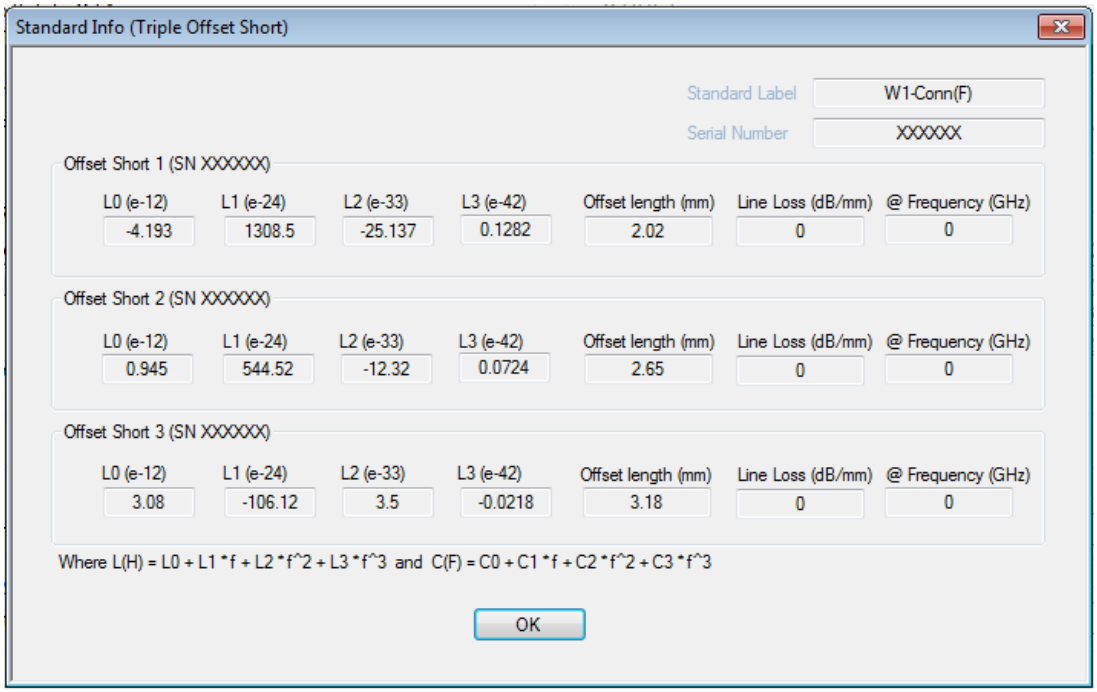
The setup and standards information dialogs for SSST are shown in Figure 5-1 and Figure 5-2 for a 2-port calibration.



1. Through selected allows user entries for length, line impedance, line loss and frequency.
2. Reciprocal selected allows user entry for length.
3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 5-1. TWO PORT CAL SETUP (SSST, COAXIAL)

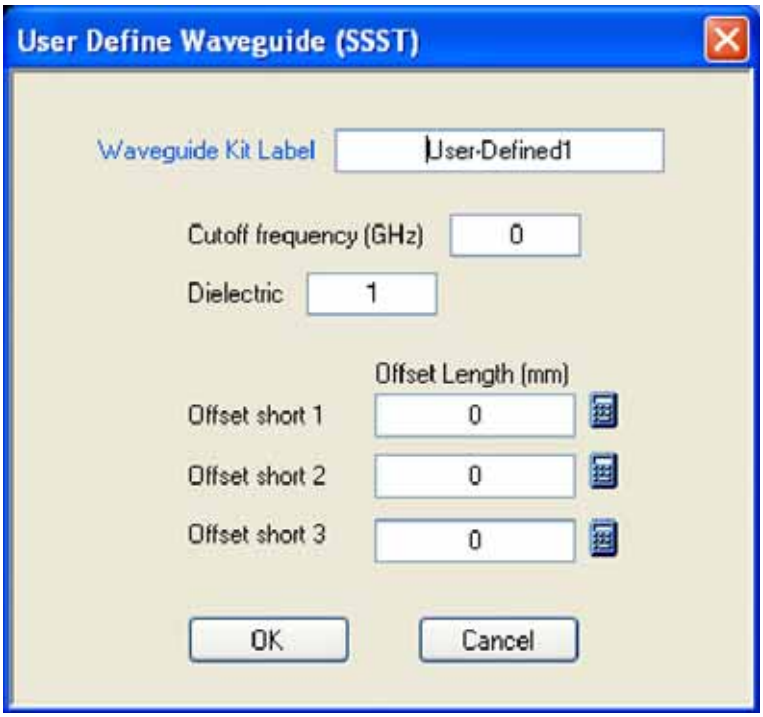
For one port calibrations, only one of the port definitions (unless reflection-only calibrations are being performed for both ports 1 and 2) will be present and the thru line section will not be present. For a 1-path 2-port calibration, one of the port definition sections will not be present.



Typical parameters for SSST calibration—W1 (F) Connectors

Figure 5-2. STANDARD INFO (TRIPLE OFFSET SHORT) Information Dialog Box

Variations for other line types are similar to those for SOLT. For waveguide, the media and standards information are again combined into one dialog (Figure 5-3).



Typical parameters for SSST calibration—Waveguide—User-Defined 1—Triple-Offset Short Calibrations

Figure 5-3. USER DEFINED WAVEGUIDE (SSST) Information Dialog Box

Note	Much like SOLT/SOLR, there is another version of SSST using a reciprocal in place of a thru called SSSR. Reciprocal measurements (using SOLR vs. SOLT as an example) are covered in more detail in Chapter 16, “Reciprocal Measurements”.

In SSST, it is also possible to define the reflection standards with .s1p files as discussed in Chapter 3 for SOLT. The user can create these files for any calibration kit based on measurement, electromagnetic modeling or other means. If based on measurement, it is recommended that a higher accuracy calibration form the basis of the characterization (e.g., LRL based on beadless airlines for coaxial situations).

As with SOLT, an .s2p-defined thru is also an option for SSST. A file describing the 'thru' can be loaded to account for its loss and mismatch. Interpolation and extrapolation will be used if the file frequency list does not match the current calibration requirement.

As with SOLT, and even without .s1p definitions, the insertion loss of the opens and shorts can be defined. A loss variable is entered in dB/mm (which will be applied against the offset length) and the frequency (f_{ref}) at which that loss value applies. For other frequencies, the loss is scaled by sqrt(f/f_{ref}). If a 0 reference frequency is entered, the loss value is assumed to be constant with frequency. Sometimes, the loss of reflect standards is expressed as gigaohm/second at 1 GHz.

To convert those values to dB/mm, use this equation:

$$\text{dB/mm} = \frac{\text{loss in gigaohms/s} \times 0.014476483}{Z_0}$$

Equation 5-3.

Where Z_0 is the characteristic impedance (usually 50 ohms). Also, enter 1 GHz for the reference frequency, as the scaling is the same.

5-3 SSST/SSSR Calibration Example

The following example presumes an MS464xB Series VNA in a millimeter-wave or broadband setup so that W1 connectors can be used (since they represent a common triple offset short example). A different connector can be selected in Step 4 below if a different model/configuration is being used. It is assumed that a M-F cable is connected to Port 2 so that a M (port 1) and F (port 2) reference plane pair is available.

In this example, a full 2-port SSST calibration will be performed although a number of other options are discussed along the way. These options are discussed in Section 5-4 “mSSS Calibration”.

1. Set the desired frequency range (FREQUENCY menu), power (POWER menu) and IFBW/Averaging (AVERAGING menu). The default power level varies depending on the instrument model and options, but will often be adequate for all passive and many active device measurements. The default IFBW is 1 kHz with averaging off, which is adequate for many applications.
2. Navigate to the TWO PORT CAL menu.
   - MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | Manual Cal | MANUAL CAL | 2 -Port Cal | TWO PORT CAL

Note	If a previous calibration exists, the Thru Update button will be active. See the Calibration Overview in Chapter 10 — Other Calibration Procedures for more information.

3. Select Modify Cal Setup, and on the CAL SETUP menu, select a Cal Method of SSST/SSSR and a Line Type of Coaxial.
4. Select Edit Cal Params, the TWO PORT CAL SETUP (SSST, COAXIAL) dialog box appears (shown below).

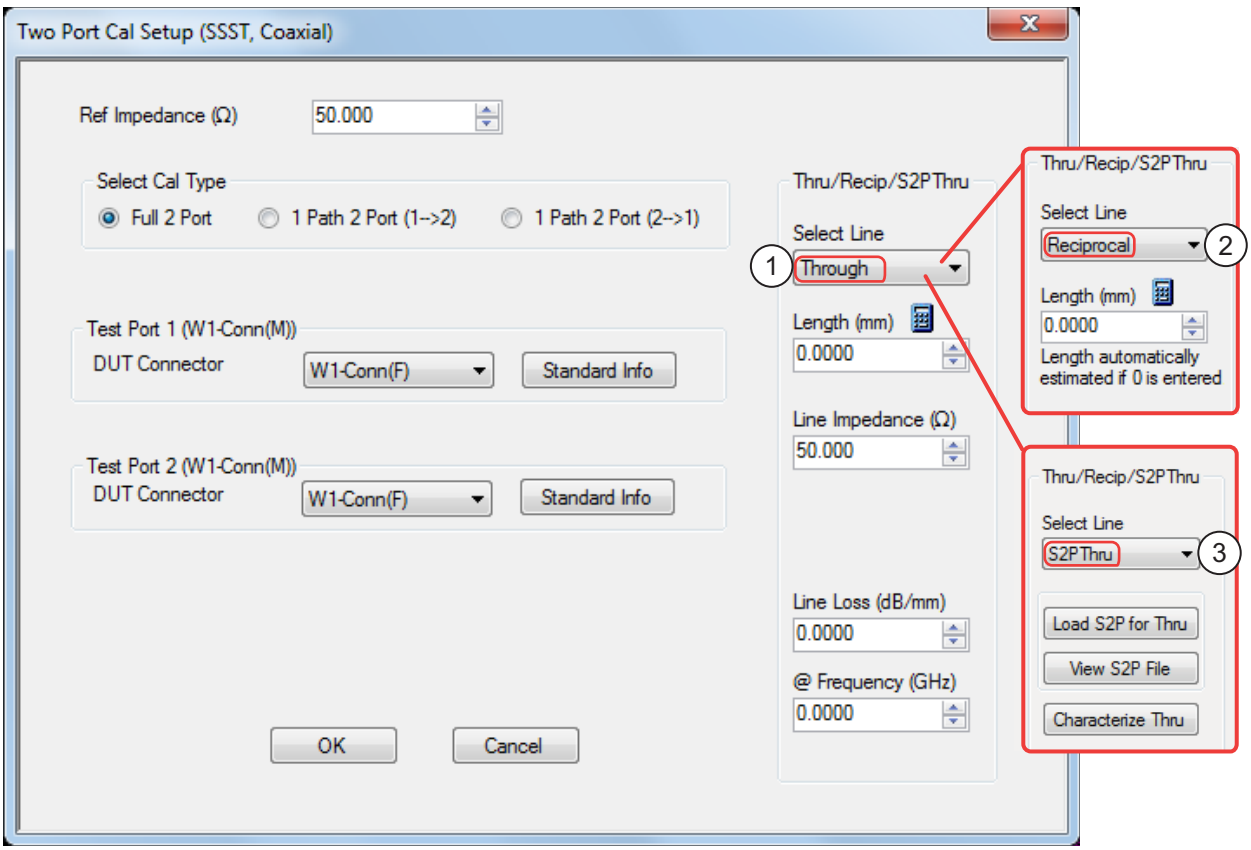


Figure 5-4. TWO PORT CAL SETUP (SSST, COAXIAL) Dialog Box (1 of 2)

1. Through selected allows user entries for length, line impedance, line loss and frequency.
2. Reciprocal selected allows user entry for length.
3. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 5-4. TWO PORT CAL SETUP (SSST, COAXIAL) Dialog Box (2 of 2)

5. In the TWO PORT CAL SETUP (SSST, COAXIAL) dialog box, set the following calibration components:
   a. Reference impedance defaults to 50 ohms. This establishes the reference impedance for reference plane changes and Smith chart plotting but can be changed later on a per-trace basis.
   b. In the Select Cal Type area, select the Full 2 Port radio button.
   c. A zero-length (or mating) thru will be used so enter Through for the line, 0 mm for the length and 0 dB/mm loss. Zero can be entered for the reference frequency when no loss scaling is employed and the entered loss value is used for all frequencies. If a reciprocal network was being used instead of a thru, the line selection would be Reciprocal and the length entered would serve as an estimate for root choice purposes.

      If s2p thru was selected, the .s2p file describing the thru s-parameters would be loaded and used during the calibration. A *characterize thru* option is also available using two previous reflection-only calibrations (with and without the thru element present) to calculate the .s2p file.
   d. For **Test Port 1**, select a DUT connector of W1-Conn (F). The dialog box then displays a port connector as W1-Conn (M).
   e. For **Test Port 2**, select a DUT connector of W1-Conn (M). The dialog box then displays a port connector as W1-Conn (F).
   f. Click OK to close the dialog. Select Back at the bottom of the menu to return to the previous level and the TWO PORT CAL menu.
6. The Port 1 Reflective Devices and the Port 2 Reflective Devices buttons display the REFL. DEVICES PORT 1 and PORT 2 menus. The six reflection standards to be measured (Short 1, Short 2, Short 3 on each menu) will be displayed here along with the read-only fields showing the port connector choices. Connect each standard in turn, and then click the corresponding button. When all six are done (and six check marks appear), click Done to return to the previous level and the TWO PORT CAL menu.
7. After measuring all six reflection standards, connect the test cable to **Port 1** to complete the zero length thru. Click on the Thru/Recip button to display the THRU/RECIP menu where a check mark should appear after the sweep pair. Note that the displayed graphs change during this step as the instrument must measure all four S-parameters of the thru line.
8. An optional isolation step is available but is not recommended. If desired, terminate **Port 1** and the end of the cable attached to **Port 2** before clicking on the Isolation (Optional) button and displaying the ISOLATIONS menu.
9. When all steps are successfully completed, the Done button is available where select returns to the CALIBRATION menu where the Cal Status button is set to ON.
10. Click Done. The calibration is now completed and turned on.

5-4 mSSS Calibration

While the T (for thru, or R for reciprocal) has been dropped from the acronym, this calibration method is a logical extension of SSST/R discussed earlier in that now more than 3 shorts may be employed. Much like mTRL (discussed in Chapter 7 — mTRL Calibration), the concept is to use an overdetermined set of standards and then select the optimal (in the sense of minimum least-squared error) set of error coefficients. In practice, this can help minimize the impact of a single (or small number of) repeatability problem(s) during the calibration or the impact of a poor definition/characterization of a single calibration standard. Also, it can avoid the need to merge a number of SSS calibrations in order to achieve a wider bandwidth as the larger starting set can automatically cover a wider frequency range.

This method is particularly attractive for higher frequency on-wafer measurements since a match is not required (although it can be used within mSSS particularly to help low frequency measurements) as match models can sometimes become complicated above 110–150 GHz. mSSS also can sometimes be preferred to mTRL on-wafer since less wafer space is required and there is less sensitivity to launch and impedance variations. This method is not a panacea however as it is a defined standards approach and, as such, does require accurate knowledge of the shorts' reflection coefficients.

This calibration approach has been discussed extensively in the literature and, if one is creating a calibration kit for this method (e.g., for millimeter wave on-wafer measurements), some of these papers may be of assistance in selecting line lengths. This list is not meant to be exhaustive and many papers in this area exist.

- A. Lewandowski, W. Wiatr, L. J. Opalski and R. Biedrzycki, "Accuracy and bandwidth optimization of the over-determined offset-short reflectometer calibration," *IEEE Trans. Micr. Theory Techn.*, vol. 63, pp. 1076–1089, Mar. 2015.
- W. Wiatr and A. Lewandowski, "Multiple reflect technique for wideband one-port VNA calibration," *Proc. 16th Int. Microw., Radar, Wireless Commun. Conf.*, pp. 37–40, May 2006.
- J. Hoffmann, P. Leuchtmann, and R. Vahldieck, "Over-determined offset short calibration of a VNA," *IEEE MTT-S Int. Microw. Symp. Dig.*, pp. 47–50, June 2008.

The calibration still solves for one reflectometer at a time and there are still the three independent variables. This equation can be cast in the form:

$$\begin{bmatrix} C_1K_1 & K_1 & -C_1M_1K_1 \\ C_2K_2 & K_2 & -C_2M_2K_2 \\ \cdots & \cdots & \cdots \\ C_NK_N & K_N & -C_NM_NK_N \end{bmatrix} \times \begin{bmatrix} x \\ y \\ z \end{bmatrix} = \begin{bmatrix} M_1 \\ M_2 \\ \cdots \\ M_N \end{bmatrix}$$

Equation 5-4.

Where the C_i are the defined reflection coefficients for the different shorts (usually $\Gamma_i e^{-j2\beta L_i}$; the Γ_i are nominally –1, but more on that later), the M_i are the measured values, and the K_i are the weights (more on that later also). The row indices represent the offset short number (N total). The reflectometer terms then come from x, y, and z:

$$edj = y$$
$$epjS = -z$$
$$etjj = x - yz$$

Equation 5-5.

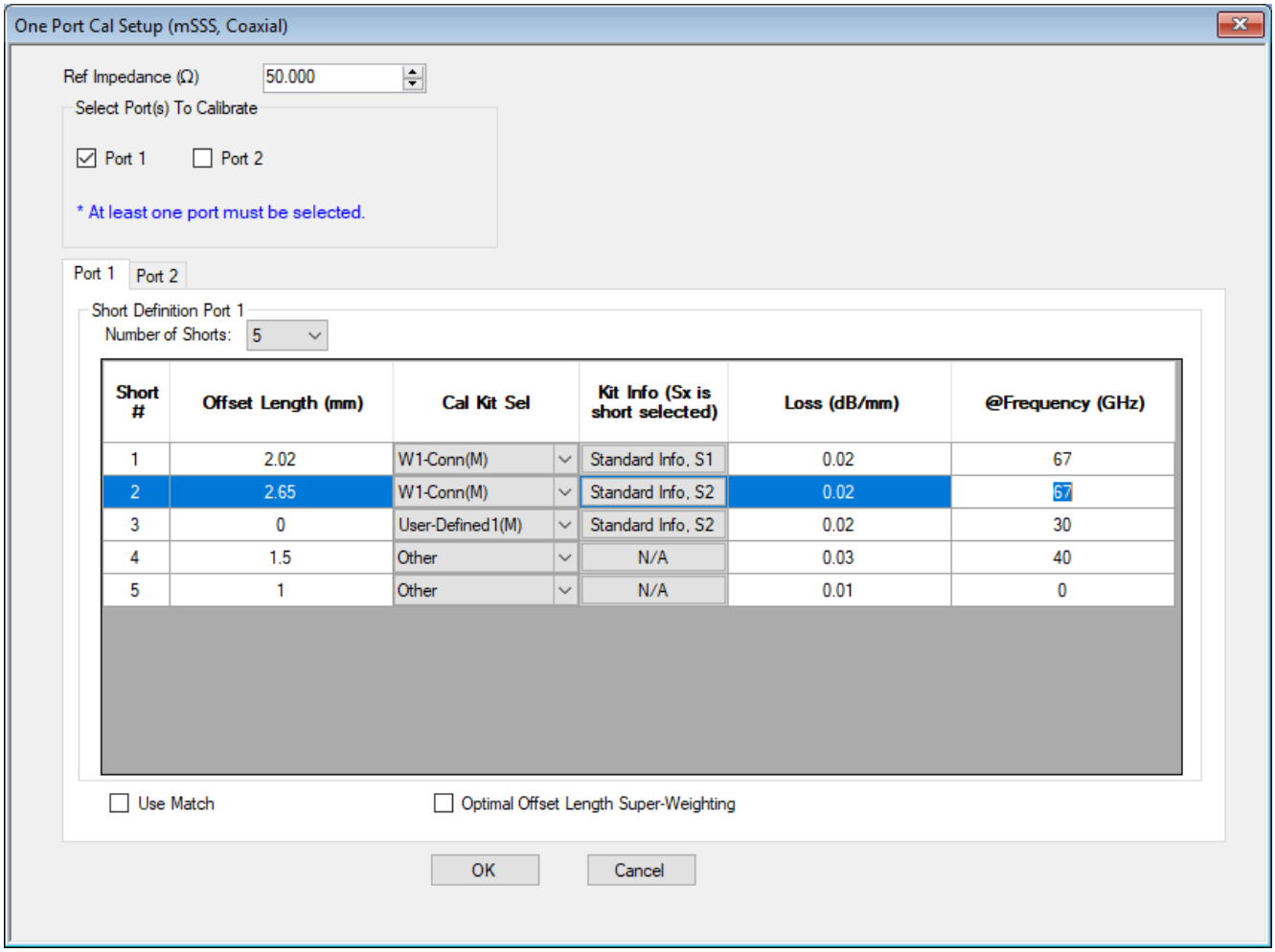
As can be deduced from other chapters in this guide, *edj* is the directivity error coefficient, *epjS* is the source match error coefficient and *etjj* is reflection tracking error coefficient for the j^{th} port. The thru line/reciprocal choice processing (and the added-line processes for 4-port) are the same as for SSST/R and all of the other defined standards calibrations. The above matrix equation is solved in a least-squares sense to generate an optimal (in the LSQ sense) set of reflectometer error coefficients independently at each frequency.

The user has some flexibility in how the shorts (the C_i) are defined.

- A standard Anritsu calibration kit can be used (the 3656x 1 mm kit and the 3659 0.8 mm kit both have sets of offset short standards) and any subset of the shorts in those kits can be employed.
- A user-defined calibration kit can be employed with either modeled or .s1p definitions. Again, these kits generally contain a number of offset shorts and any subset can be employed.

- A selection of Other can be employed for a one-off definition of a short based on offset length and a loss profile (X dB/mm at a reference frequency f_r and the loss defined to be flat with frequency or scale as sqrt(f/f_r)).

A setup dialog (for coax, 1-port calibrations on a 2-port VNA in this case) is shown in Figure 5-5.

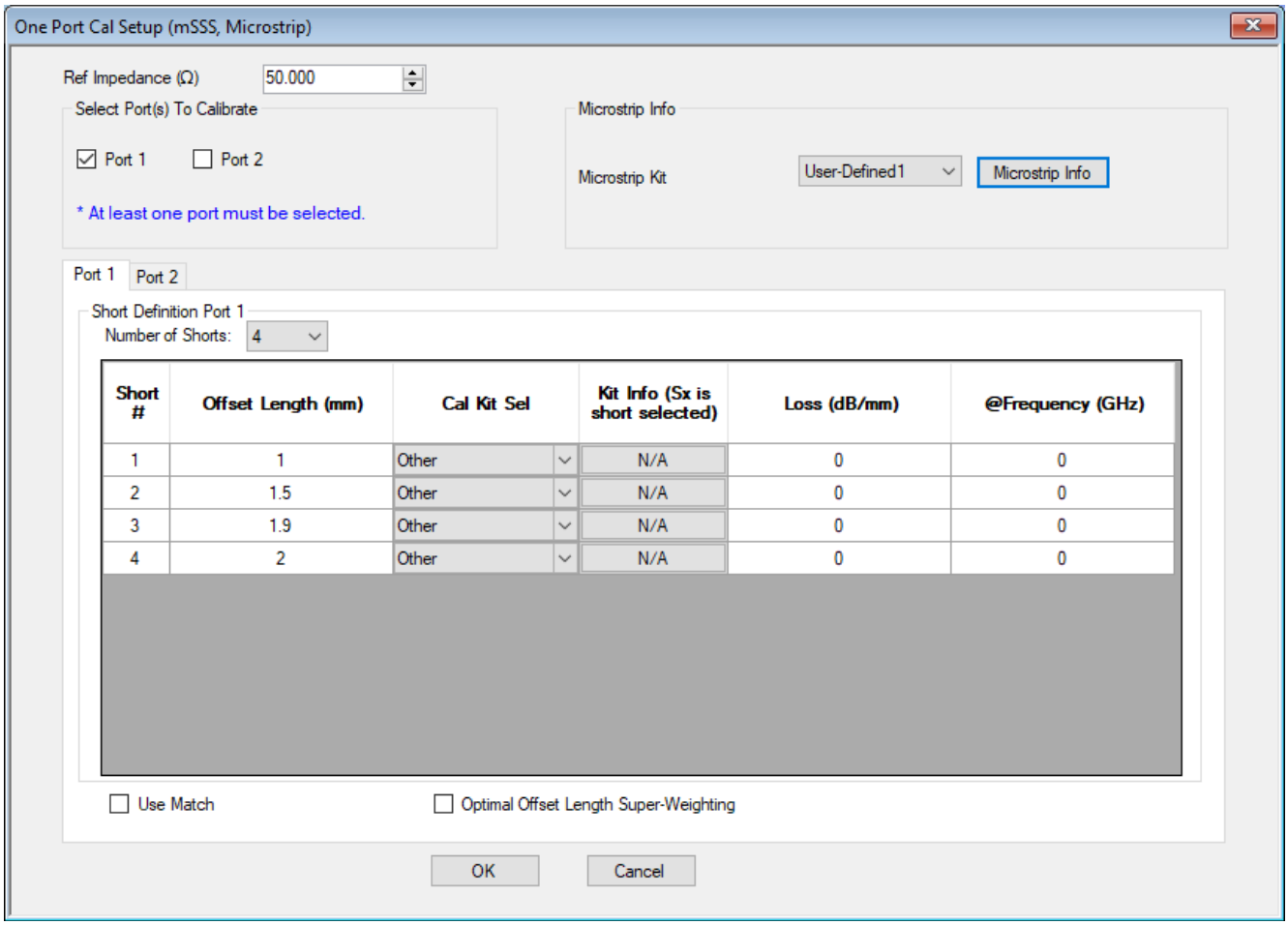


Note the variety of calibration component definition formats that are possible. Up to 10 shorts can be defined.

Figure 5-5. mSSS Configuration for an Example 1-Port Coaxial Calibration

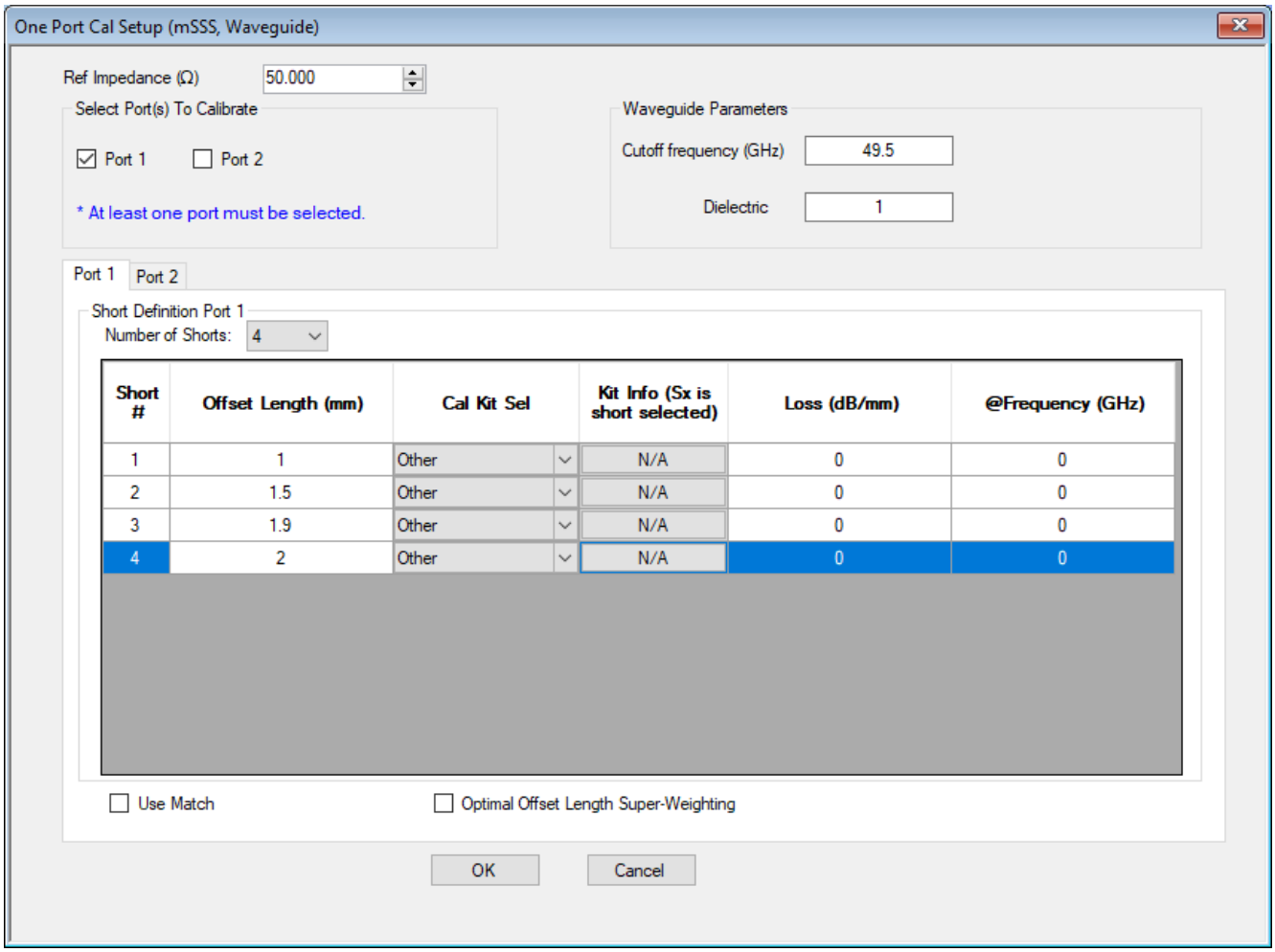
The dialog will check that two shorts are not identical but will not directly assess the suitability of the set for the frequency range being calibrated. Warning messages will be issued if the condition number of the matrix in Equation 5-4 is extreme or if a near-singularity is encountered. Generally, the largest 2-3 offset length deltas will establish the lower frequency limit and the shortest 2-3 will establish the higher frequency limit as one might have guessed from the SSST sections earlier in this chapter.

Coaxial mSSS can be useful in millimeter wave contexts as coaxial load performance may not be optimal and the related non-dispersive selection is applicable to coplanar waveguide and stripline systems for on-wafer and/or fixtured environments. Waveguide and microstrip contexts are also useful and they have very similar dialogs except the media parameters (cutoff frequency and dielectric in the case of waveguide and substrate parameters in general in the case of microstrip) must be specified globally for all of the shorts to be used. As with the other calibrations, these media parameters will propagate to the Measurement menu for use in reference plane extension and time domain processing. The waveguide and microstrip versions of the mSSS configuration dialog are shown in Figure 5-6 and Figure 5-7 (for 1-port calibrations; similar dialogs exist for 2+ port calibrations except they also have fields for thru/reciprocal information).



Note the specification of the global media parameters.

Figure 5-6. Microstrip mSSS Configuration Dialog



Note the specification of the global media parameters.

Figure 5-7. Waveguide mSSS Configuration Dialog

The weights K_i (in Equation 5-4) start off with a value of 1 so, unless the super-weighting box is checked, Equation 5-4 simplifies to Equation 5-6.

$$\begin{bmatrix} C_1 & 1 & -C_1M_1 \\ C_2 & 1 & -C_2M_2 \\ \cdots & \cdots & \cdots \\ C_N & 1 & -C_NM_N \end{bmatrix} \times \begin{bmatrix} x \\ y \\ z \end{bmatrix} = \begin{bmatrix} M_1 \\ M_2 \\ \cdots \\ M_N \end{bmatrix}$$

Equation 5-6.

By checking the super-weighting box, weights are applied to the shorts based on a prediction of which ones are most useful to the calibration. That weight, in an approach similar to that discussed in the literature, is obtained by looking at the error coefficient sensitivity for all possible permutations of three shorts (in an SSS sense) and then mapping each individual short's contributions to those sensitivities. Geometrically, each combination of three shorts (at each frequency) can be thought to define a triangle on the complex plane and (generally) the larger the area of that triangle, the less sensitive the calibration will be. Those shorts that are part of the most large-area triangle combinations will receive the highest weight. As an example, consider a set of five lossless shorts with offset lengths of 0 mm, 1 mm, 7 mm, 3 mm, and 12 mm and this kit will be assessed at three frequencies: 1 GHz, 10 GHz, and 40 GHz. The weights chosen are shown in Figure 5-8.

At 40 GHz, one would expect the smaller length deltas to get higher weights and indeed that happens as shorts 1, 2, and 4 (with differences of 1, 2, and 3 mm) get the higher weights. At the low frequency extreme, one would expect the opposite. This also happens as shorts 3 and 5 (in conjunction with short 1) are favored for that frequency.

The use of super-weighting does accentuate the importance of the most bandwidth-friendly offset lengths but it does suppress, to some extent, the algorithm's ability to underweight repeatability-induced outlier measurements. Thus, super-weighting is attractive when trying to absolutely maximize calibration bandwidth (for a given set of shorts) but is not attractive in a repeatability-challenged environment (e.g., in a fixture).

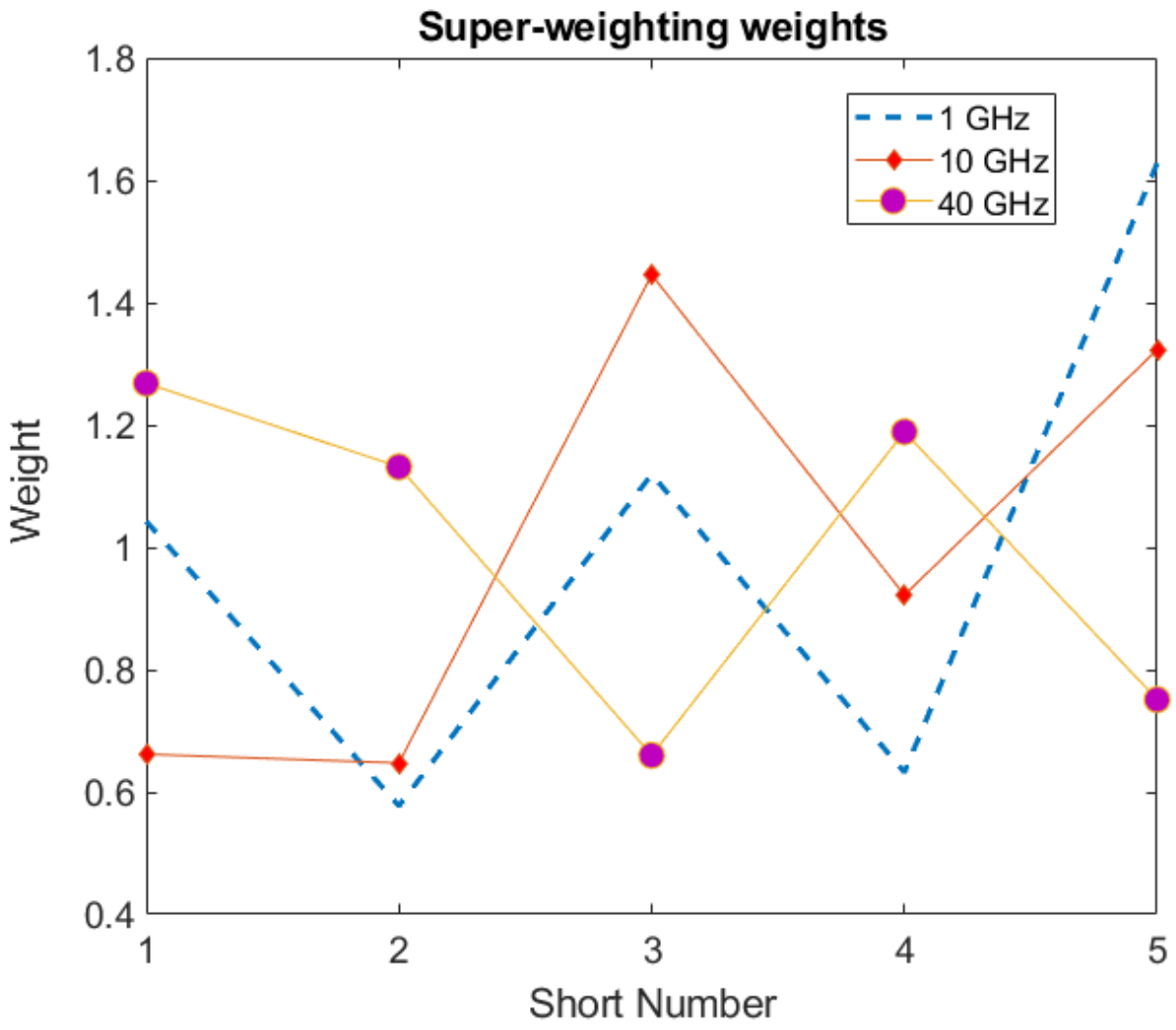
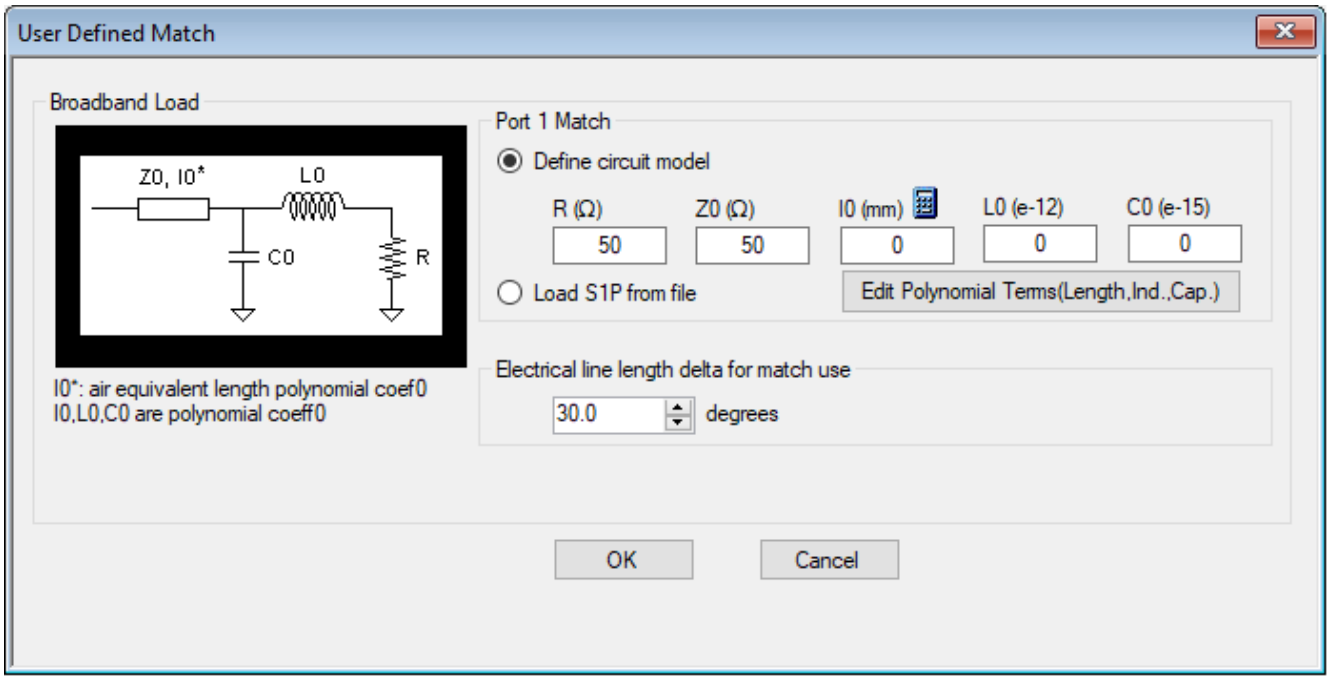


Figure 5-8. Super-weighting Results (1 of 2)

The super-weighting option provides additional emphasis to those shorts that contribute to larger reflection coefficient coverage and hence reduced calibration sensitivity. Weights are applied to the equations in a linear sense with a nominal value of unity.

Figure 5-8. Super-weighting Results (2 of 2)

The mSSS calibration also allows the incorporation of a match standard (we realize this does not fit into the acronym) to (a) help with low end frequency coverage as offset short calibrations are always inherently somewhat bandpass in nature and (b) provide additional sensitivity suppression if some of the short standards become electrically indistinct over other parts of the bandwidth of interest. As with some other calibrations in the MS464XX system, a full load model or an .s1p file can be entered as suggested in Figure 5-9. This dialog is only available if the Use Match box is checked.

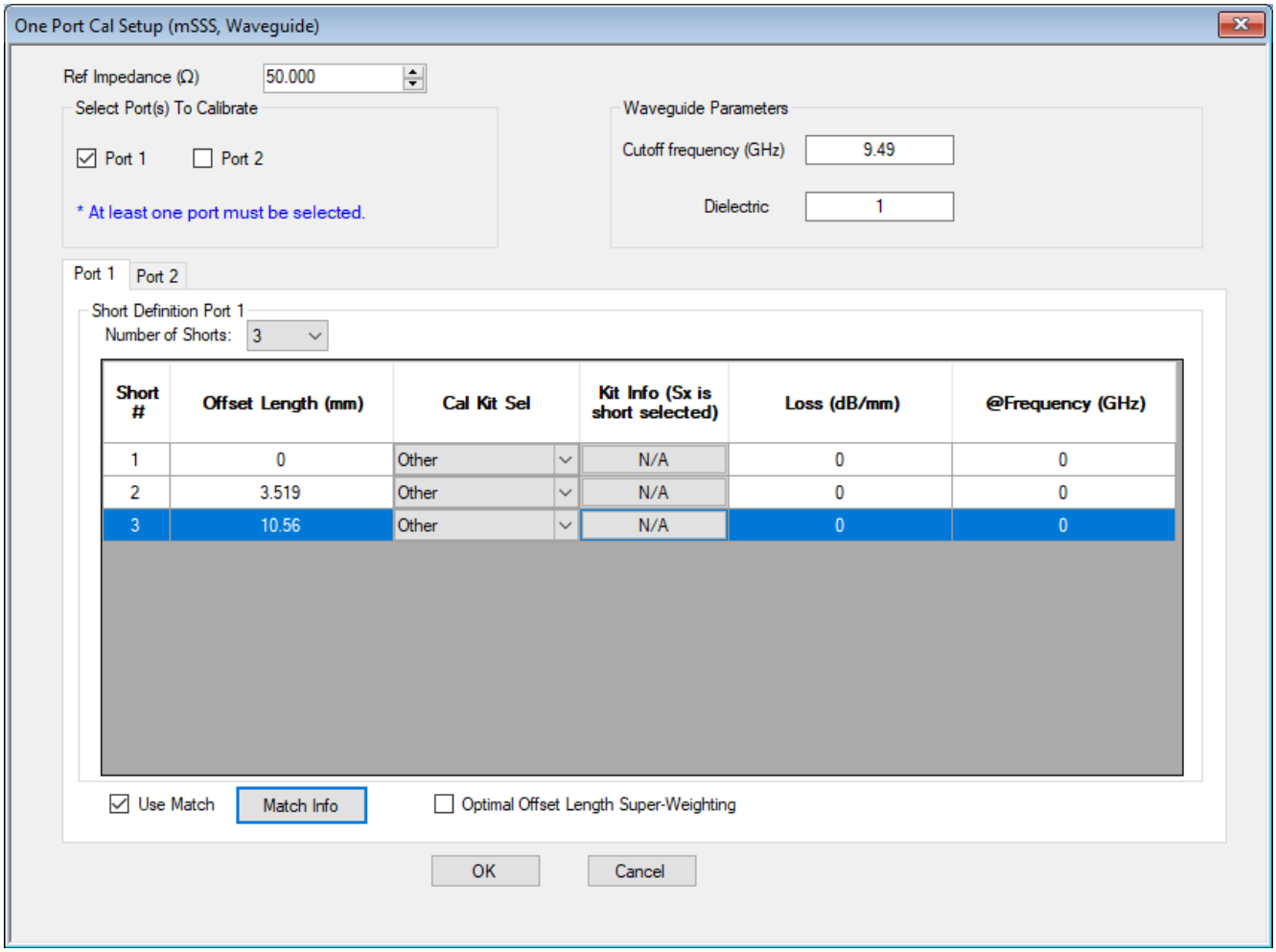


This dialog is available to define a match standard should its use be desired within the mSSS calibration.

Figure 5-9. Match Info Dialog

The decision to use the match measurement as part of the solution (it is added as another row in Equation 5-4 with a weight set to twice the maximum of the weights amongst the shorts) is based on the offset length differences between the shorts. A minimum delta (expressed in terms of difference in reflection coefficient phase) is computed for every permutation of three shorts among the defined shorts. If the maximum of those minimum deltas (amongst all of the permutations) is less that the entered 'Electrical line length delta for match use', then the match measurement will be employed at that frequency. This decision is made on a frequency-by-frequency basis.

As a simplified example of this process, consider an intentionally inappropriate set of 3 offset shorts for a calibration and the resulting measurement of a DUT. The measurements will be done in Ku-band waveguide (~12 –18 GHz) with offset short lengths of 0 mm, 3.519 mm and 10.56 mm using a simple 1-port calibration and the setup information shown in Figure 5-10. The problem with this kit is the larger delta between the 0 mm and the 10.56 mm standards. At 17.08 GHz, those two shorts will have essentially the same reflection coefficient (ignoring losses) so a near-singularity is likely.

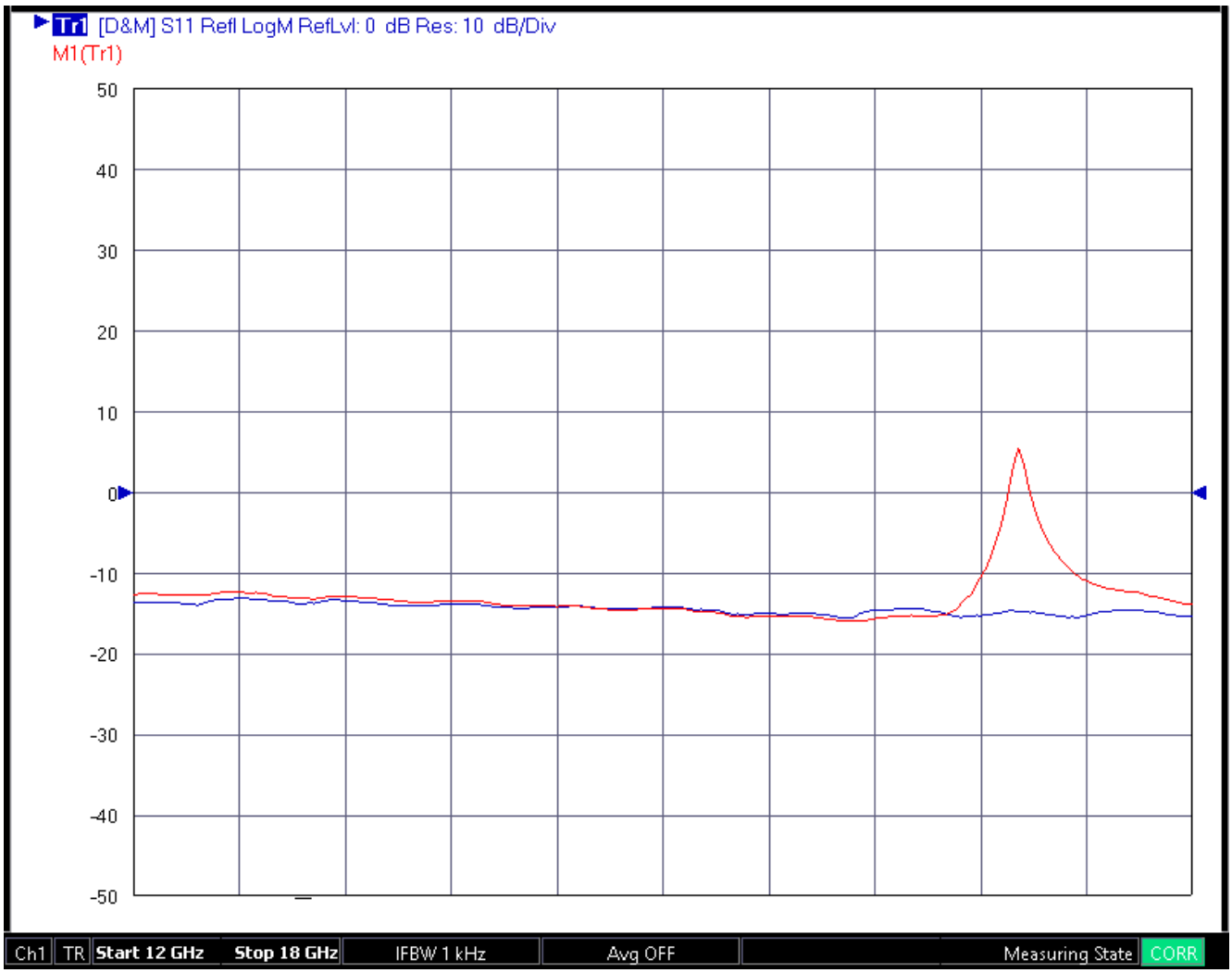


The medium is Ku-band waveguide.

Figure 5-10. Example Setup: Use of the Match Standard

An open flange will be used as the DUT which, as is well-known, should have a reflection coefficient somewhere in the vicinity of –15 dB (ignoring flange imperfections, the effects of mounting holes and pins, etc.). Without using the match measurement, the calibration with the 3 shorts results in the red curve in Figure 5-11. The effects of the near singularity should be obvious (resulting in a value above 0 dB at some frequencies).

The calibration was then re-done using the match standard (box checked as in Figure 5-10). The default 'perfect' load model was used for the standard waveguide termination and the phase delta threshold was left at 30 degrees as in Figure 5-9. The measurement of the open flange using this calibration is shown in blue in Figure 5-11 and is much closer to the expected result.



Measurements of an open waveguide flange using a defective 3-short calibration (red) and the 3-short calibration augmented by the match measurement (blue). The near-singularity in the red curve is due to the inappropriate choice of offset lengths for this frequency range.

Figure 5-11. Measurements of an Open Waveguide Flange Using a Defective 3-short Calibration

5-5 mSSS Calibration Example

Consider a broadband coaxial example where it is desired to measure over as wide a frequency range as possible. Four shorts are available with three of them being from a standard 3659 0.8 mm calibration kit (those three shorts are normally used to cover 80 –145 GHz as part of a broadband SOLT/SSST combined calibration). The basic setup is shown in Figure 5-12.

The VNA system is a ME7838D (70 kHz- 145 GHz) based on the MS464XX and these broadband systems are covered in more detail in Chapter 17 — Broadband/mmWave Measurements (Option 7, Option 8x) of this guide. The DUT for this experiment will be a stepped impedance line terminated in a pseudo-broadband load (thus creating a scalloped appearance of the reflection coefficient). The basic 4-standard mSSS calibration will be compared against variants with super-weighting applied and with the match measurement applied in addition using the protocol discussed in the previous section.

The fourth short's offset length is quite disparate from those of the other shorts so one might expect low frequency coverage being possible but potentially with some holes where near-singularities could occur.

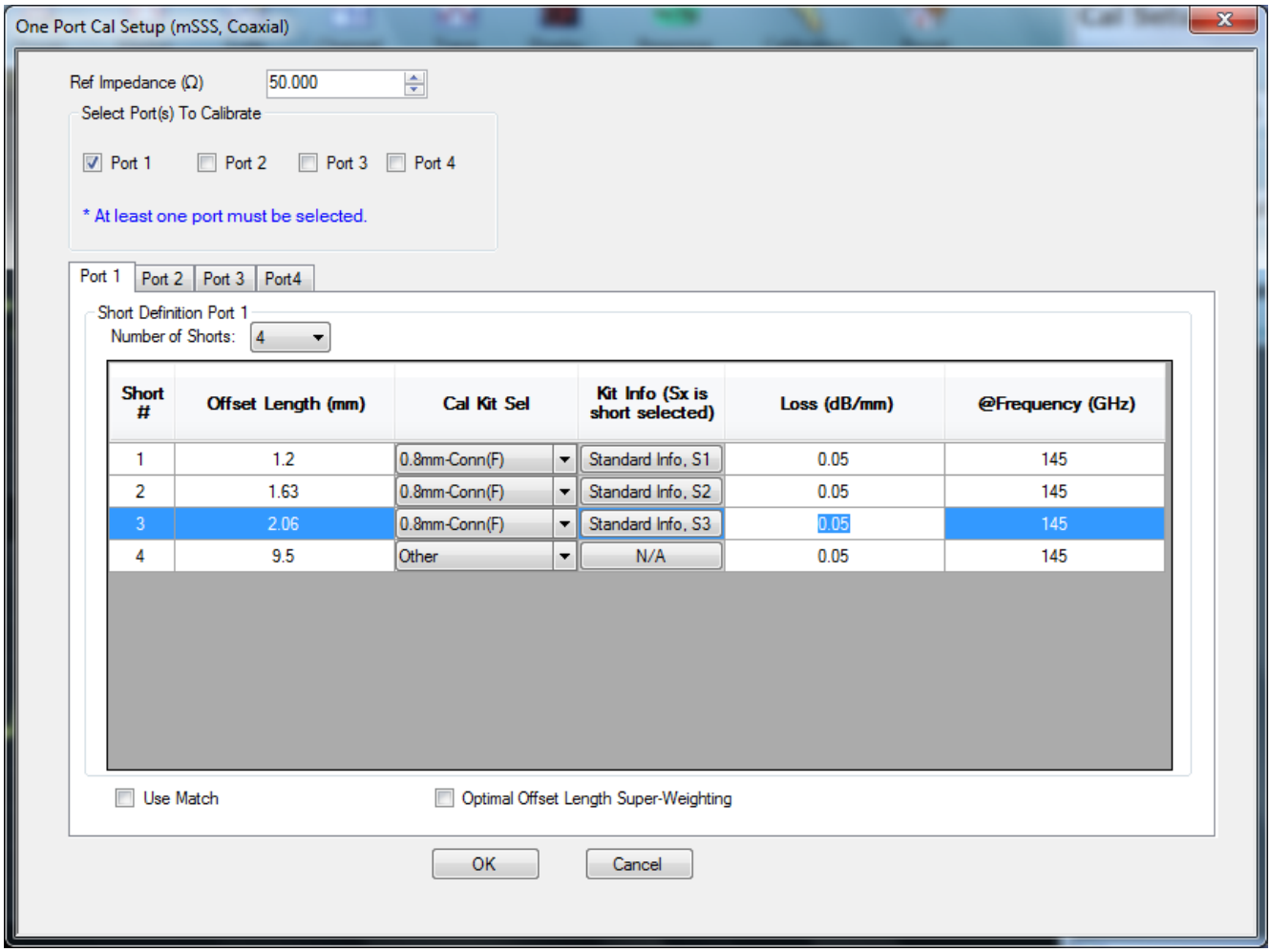


Figure 5-12. Example Setup Dialog: Measurement using a broadband ME7838D system based on the MS464XX.

The measurement results are shown in Figure 5-13 and none of the variants are completely surprising. The match-enabled measurement helps resolve a near-singularity around 24 GHz and shows reasonable results (agreeing with expectations) all the way down to 10 GHz. The super-weighting helps resolve a smaller issue near 60 GHz and helps with some of the deep return loss areas at higher frequency (where it helps to de-emphasize the long offset short a bit more). One should not read too much into the deep notch measurements as repeatability is a contributor but it does illustrate a scenario where an unusual combination of shorts can be aided by weight adjustment.

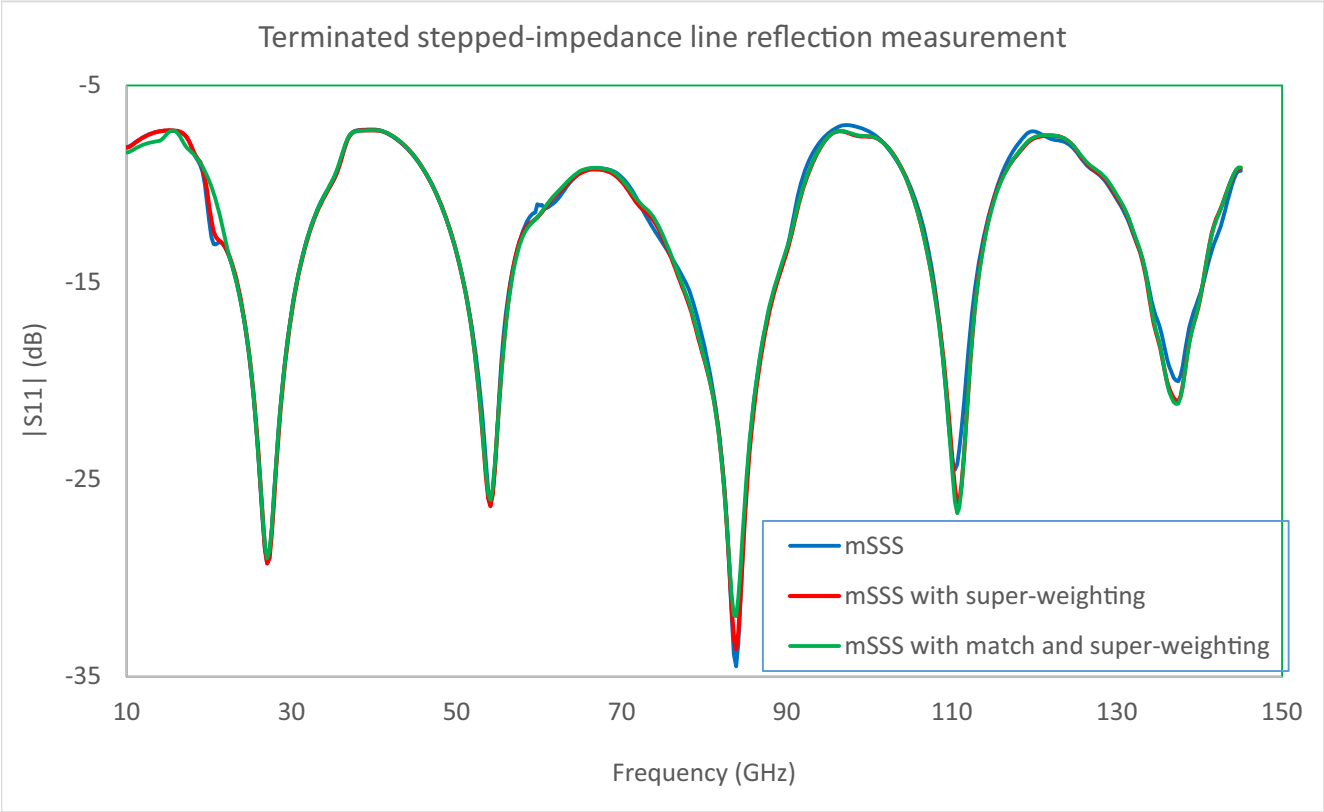


Figure 5-13. Measurements of the example DUT for the three permutations of the mSSS calibration.

Chapter 6 — LRL/LRM Calibration

6-1 LRL/LRM Introduction

This chapter describes LRL/LRM calibration algorithms and procedures. The LRL/LRM/ALRM family of calibrations relies more on the fundamental behavior of certain components (primarily transmission lines) than it does on characterized/modeled behaviors of components. It makes less use of redundancy, so fewer measurements are needed to complete a calibration, but it is also less tolerant of poor or non-repeatable measurements.

Related Chapters

The following contains content related to LRL/LRM Calibration:

- Chapter 16, “Reciprocal Measurements”
- Chapter 24, “Multiport Measurements”

6-2 LRL/LRM Comparison

LRL—Line-Reflect-Line

LRL (Line-Reflect-Line) uses two (or more) transmission lines and a reflect standard (for each port). The line lengths are important as it is required that the two lines look electrically distinct at all times (meaning it will not work at DC nor at a frequency where the difference in length is an integral number of half wavelengths). The reflect standard is assumed to be symmetric and without a high return loss. The lines are assumed perfect (no mismatch), and are usually airlines for coaxial calibrations, although other structures can be used. On-wafer transmission lines can be very good and this calibration approach will work well if the required probe movement can be managed.

LRM—Line-Reflect-Match

LRM (Line-Reflect-Match) and ALRM (Advanced Line-Reflect-Match) calibrations have one of the lines above replaced with a match (or load). The load is modeled/characterized (or assumed perfect). Since only one line is involved, this calibration can work down to DC and up to very high frequencies (practically limited by the match knowledge/characterization). Variations allow one of the match measurements to be traded for a pair of additional reflect measurements (a second reflect standard is needed). Because of the requirement that the reflect standards be distinct, the calibration may become band limited.

In the limiting case of a match that is assumed perfect, or at least assumed symmetric, this calibration reduces to the classical LRM. The added flexibility is in the ability to define asymmetric load models and to use multiple reflect standards as discussed above. The double reflect methodology allows one to feed into a load modeling utility where the load model can be further optimized.

Some parameters to keep in mind:

Line Lengths

In addition to the LRL frequency limits, the line length is used for some reference plane tasks. The fundamental reference plane of an LRL/ALRM calibration is in the middle of the first line. If the reference plane is required at the ends of this line, the line length (and loss which can also be entered) is used to rotate the reference planes to the desired location. The line length delta is also used for some root choice tasks, although the accuracy required on this entry is less. It is allowed to enter a 0 length for line 1 even when the length is physically non-zero as another way of forcing the reference planes to the middle. There are some caveats:

- All other line lengths must be relative to line 1.
- Loss of line 1 cannot be corrected.
- The offset lengths of the reflect standards must take into account the physical length change by zeroing out line 1's length so the offsets may be negative.

Line Length Delta

As mentioned above, the usable frequency range for LRL is set by the line length delta. Strictly speaking, the electrical length should be between 0 and 180 degrees for all frequencies of interest although some margin is usually desired to account for line parasitics, spurious mode launches and other problems. In general, the delta should be kept between 10 and 170 degrees or 20 and 160 degrees. Practically speaking, one can usually be more aggressive on the lower number and will want to be less aggressive on the upper number:

$$10 < \frac{360 \cdot f \cdot \Delta L}{v_{ph}} < 160$$

Equation 6-1.

Where ΔL is in meters, v_{ph} is the phase velocity of the line (= 2.9978 10^8m/s = c for air dielectric) and f can be any frequency in the range of interest, expressed in Hz.

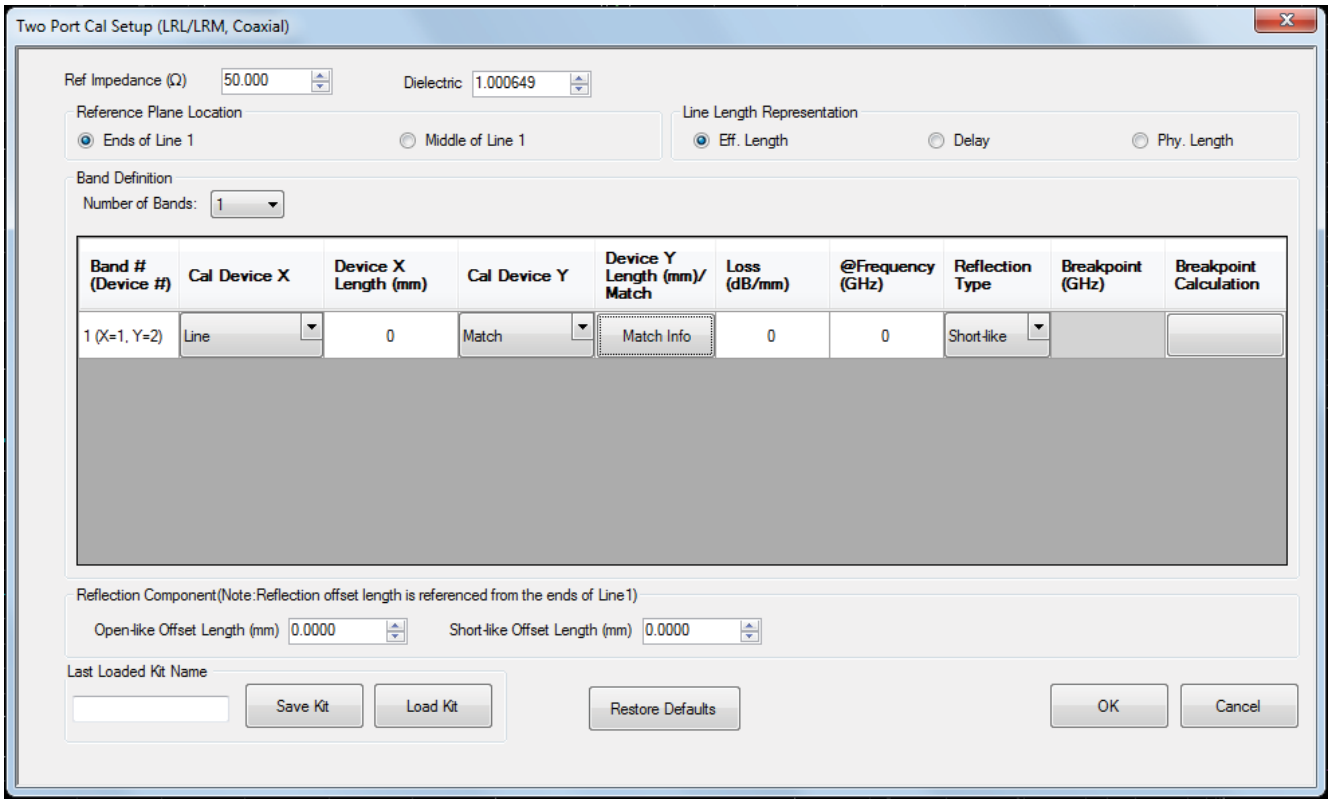
If this range is too small for the application, multiple lines and multiple bands can be used which will be discussed shortly. The single-band version of the dialog is highlighted in Figure 6-1 on page 6-3. Two devices must be defined and the first (Cal Device X) must be a line which has a length associated with it in millimeters (air equivalent) or in picoseconds of delay. Note that the Cal Device X in the first band has an added role when the reference plane choice 'Ends of Line 1' is selected. The length of this Cal Device X will be rotated out from the final error coefficients to place the reference planes at its ends. The basic LRL/LRM/ALRM algorithm places the reference plane in the middle.

A loss and frequency dependence for the line can be optionally specified (loss defaults to 0 dB). The loss is in per-unit-length terms and can be assumed flat with frequency by entering 0 in the frequency field. If a non-zero value is entered, the loss value entered is assumed to be at that frequency and a square-root-of-f scaling will be used at other frequencies. In other words, if the loss K is entered at frequency f_0 then the value at frequency f_1 will be computed as:

$$Loss_at_f_1 = K\sqrt{\frac{f_1}{f_0}}$$

Equation 6-2.

The second device (Cal Device Y) can either be another line or a match. If another line, its length/delay should be entered although this value will be primarily used to help with root choice. The loss values for Cal Device X will be used for Cal Device Y. If the second device is a match, the model for that match may be entered using the sub-dialog (assumed infinite return loss by default). As discussed previously, if both reflects are selected and a match is being used for Cal Device Y, then the ALRM algorithm will be used. The entered load model is only then used for the first part of the calculation to help the optimizer more quickly generate a better fit for the simplified match model. Also, only one of the match measurements is needed for this algorithmic variant and the match to be used is selected in the sub-dialog.



Typical parameters—One-Band LRM calibration

Figure 6-1. TWO PORT CAL SETUP (LRL/LRM, COAXIAL) Dialog Box

Because of the phase restrictions discussed above, each LRL calibration is fundamentally band-limited to something on the order of an 8:1 to 17:1 frequency range (and some users may restrict it further for measurements requiring very low uncertainties when the line losses are low). To cover wider frequency ranges and to, for example, use LRM/ALRM for part of the frequency range, the concept of Multiband LRL was generated. Various combinations of standards (staying within the LRL/LRM/ALRM family) can be used to cover multiple bands and hence cover the extended range. The concept of the 'breakpoint frequency' is needed that defines when the calibration coefficients are taken from the lower band measurements and when they are taken from the upper band measurements. This process does not affect how the measurements are done. It just changes which data (from an over-determined set) is used to compute the calibration coefficients. The calibration process is optimized so that if a particular standard is used in multiple bands, it only needs to be measured once.

The setup dialog is shown in Figure 6-2 on page 6-5 with the maximum of 5 allowed bands selected. In some sense, each band is a new LRL/LRM/ALRM calibration covering some subset of the desired frequency range. All of the bands share the same open-like and short-like reflects although one can choose between those two independently in each band (or use both for ALRM in any band). In each band, one must choose two lines or one line and a match and those devices can be shared amongst the bands. As an example, it is not unusual to share one line in all of the bands and then the 2^{nd} line in each band is chosen to create an ever-decreasing ΔL as the band number increases (thus each subsequent band is to be used in a higher frequency sub-range). In the higher numbered bands, any devices in the previous bands can be used. Note that loss entries are local to a band in order to allow the user to correct for deviations from the square-root-of-*f* loss frequency dependence model.

To illustrate a multi-band setup, we will work through a design example where the calibration range is to be 200 MHz to 70 GHz and we wish to not exceed 20 to 160 degree relationships and do the calibration with only LRL using air dielectric lines. Since the high-to-low ratio is less than 8^3 the calibration can be accomplished in 3 bands under these conditions. It is desired to use a common line of length 10 cm for all bands.

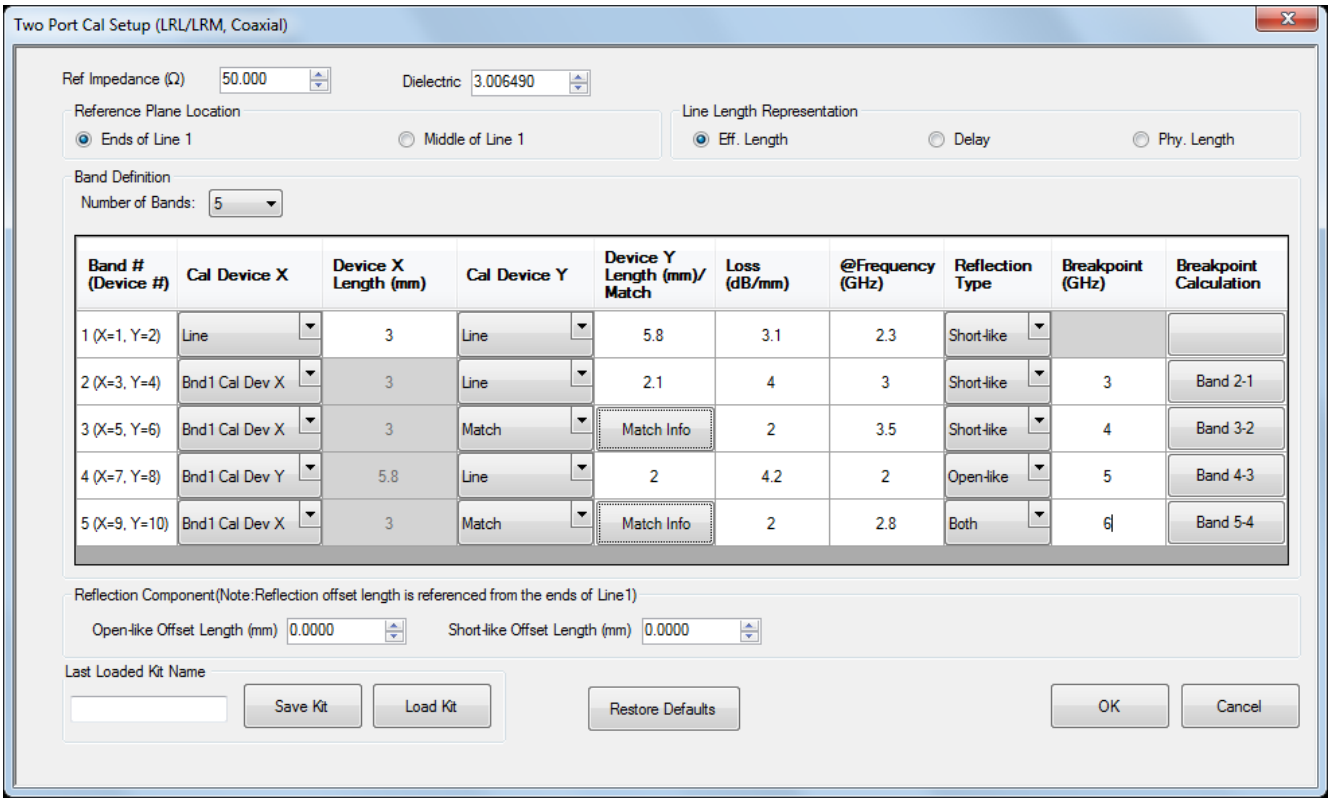
If the first band must work down to 200 MHz with a 20 degree ΔL limit, then the first ΔL=8.333 cm so the second line should be 18.333 cm long. This first band, using the 160 degree limit, should be able to reach 1.6 GHz. To allow some band overlap (and improve uncertainties a bit more), we will choose the 20 degree lower limit of the second band to be a bit lower than 1.6 GHz and select 1.4 GHz. This implies a second ΔL of 1.19 cm. Since we wish to re-use the 10 cm line in this band, the second line in the second band should be 11.19 cm long with a 160 degree-based frequency limit of 11.2 GHz. For the third band, we will again aim for a value under the previous band's 160 degree limit and select 10 GHz. This leads to a third ΔL of 1.67mm or a final line length of 10.167 cm.

Summarizing:

BAND 1:	Cal Device X 10 cm line	Cal Device 18.333 cm line	
BAND 2:	Cal Device (use band 1 Dev X)	Cal Device Y 11.19 cm line	Breakpoint: 1.497 GHz
BAND 3:	Cal Device X (use band 1 Dev X)	Cal Device Y 10.167 cm line	Breakpoint: 10.58 GHz

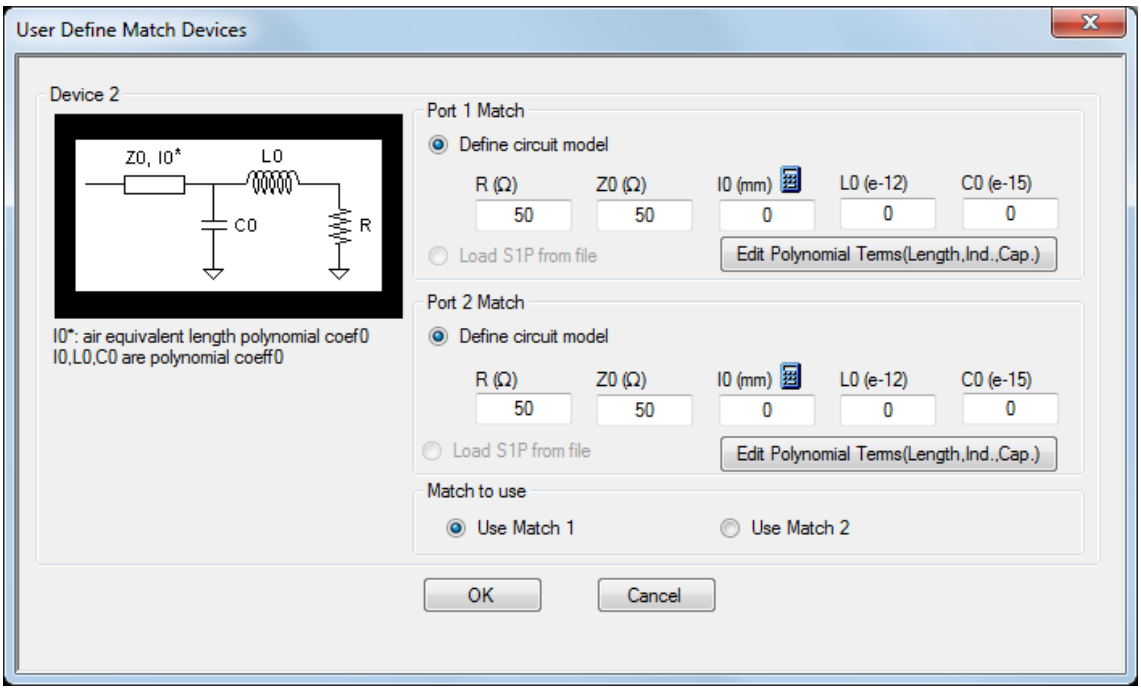
In this full calibration example, there are a total of four lines to be measured (along with whatever reflect standard(s) was (were) selected). Although not all of the frequency range data for each device will be used (e.g., band 2 Cal Device Y data for frequencies below 1.497 GHz or above 10.58 GHz will not be used), all devices will be swept over the whole calibration frequency range in order to simplify sweep control and minimize overhead.

Here the breakpoints were calculated off-line using the geometric mean of the upper limit of the lower band and the lower limit of the upper band. The recommended values that would be generated by the instrument would be different since it uses the 10 to 160 degree process instead of the 20 to 160 degree limit we selected for this example. These slight shifts in breakpoint will generally have limited effect of the results (<0.03 dB error generally at the breakpoint in this case) but if one gets more aggressive with the angle limits, errors can increase. This is less true if the lines are lossy (as in a PC board environment) than if the losses are very low.



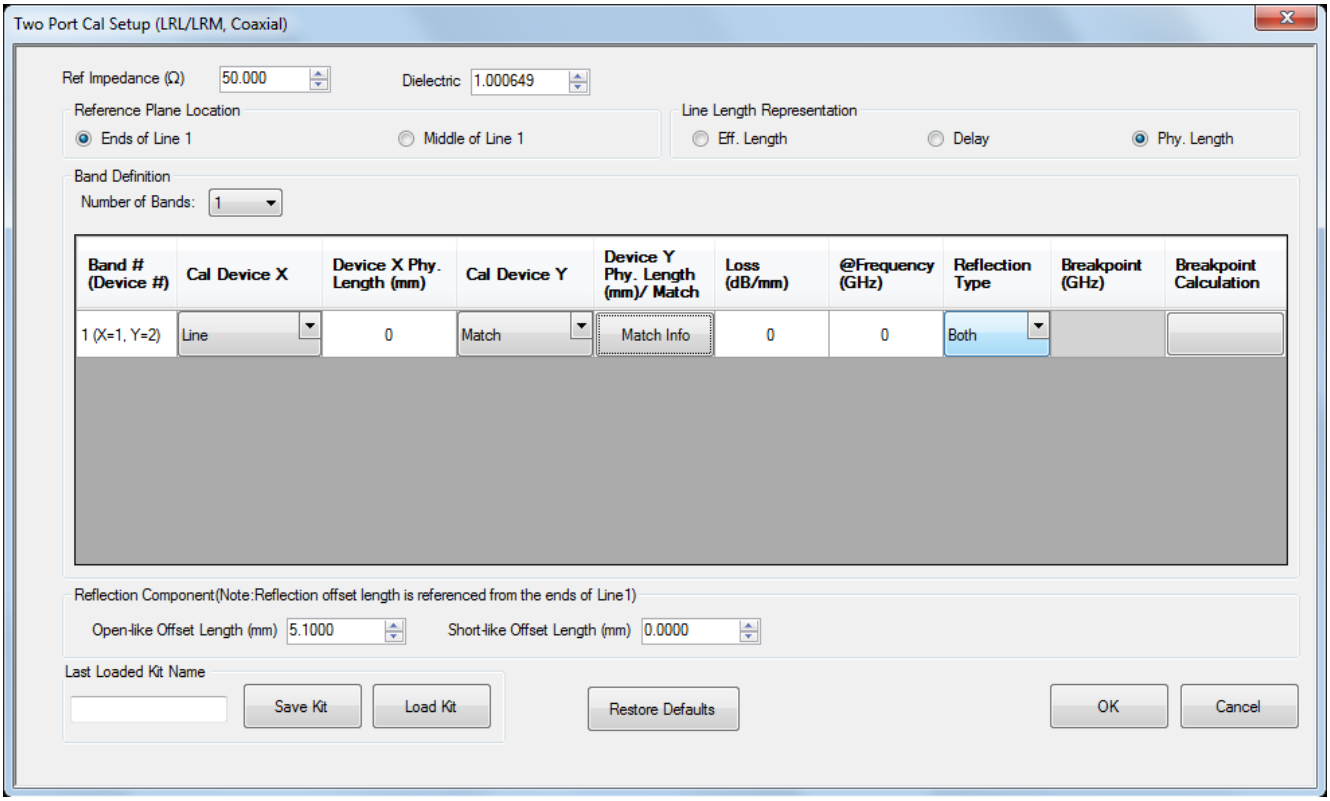
Typical parameters—Bands 1, 2, and 4: LRL, Bands 3 and 5: LRM

Figure 6-2. TWO PORT CAL SETUP (LRL/LRM, COAXIAL) Dialog Box



Typical parameters—Defining the Load for ALRM (match info)

Figure 6-3. USER DEFINED MATCH DEVICES Dialog Box



Typical parameters—One Band ALRM Using Two Reflects

Figure 6-4. TWO PORT CAL SETUP (LRL/LRM, COAXIAL) Dialog Box

The Cal Merge (concatenation) utility can also be used with any other calibration types in order to cover a wider frequency range.

Reflection Offset Length and Reflection Type

Some information is requested about the reflection although a full characterization is not needed. The information is used in some root-choice activities and it only needs to be known if the reflect behaves more like an open or a short (since typically opens and shorts are used as the reflect standard). The offset length is used to dynamically move the reference planes around so the algorithm will know what the reflect looks like at any given frequency. Normally, the reflection offsets are always entered relative to the ends of line 1 (even if a different final reference plane is selected). If a non-zero length line 1 is used but 0 is entered for the length (another approach to get reference planes in the middle), the reflection offsets must still be relative to the ends of the physical line 1, so the offset entries may be negative in this case.

In the double reflect ALRM methodology it is important that the reflect standards be distinct. More specifically, they must be distinct when rotated to the reference plane at the center of line 1. Since large offset lengths will lead to many more degeneracies, this *double reflect* option will generally be used when offset lengths are smaller (such as in on-wafer of fixtured calibrations).

Load Model/Characterization for ALRM

When a single reflect approach is taken within ALRM, it behaves like a classical LRM. For slightly more advanced use, complete load models can be entered for the two matches independently. The same model as described for SOLT applies.

At the highest level, two reflects are measured per port to allow more optimized information to be obtained. When the double reflect methodology is selected, an optimization routine can be selected which can lead to a load model.

The structure in Figure 6-5 below is used (similar to that for the general model except no capacitance). The resistance element is assumed known (whether from DC measurements or other parametric data). The inductance and transmission line parameters can be optimized over given ranges.

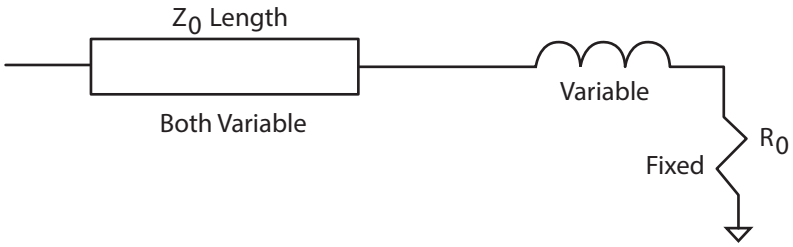
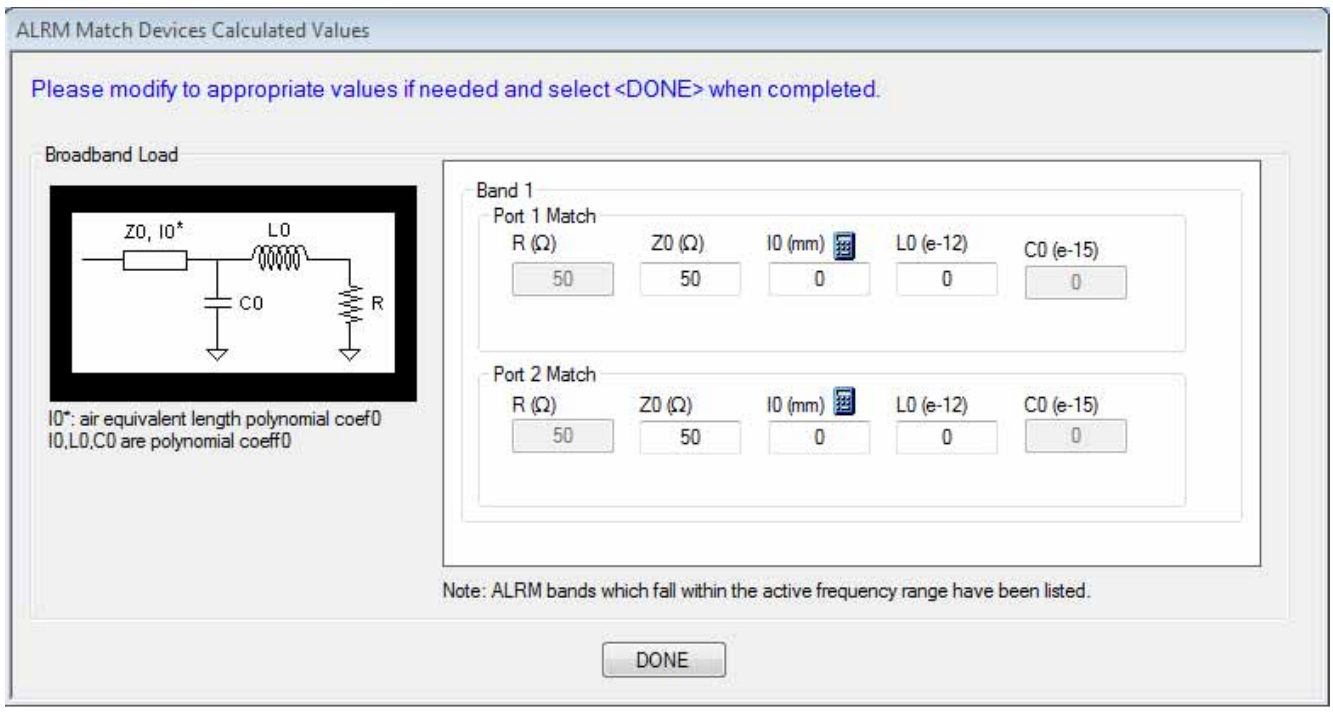


Figure 6-5. Load Model/Characterization for ALRM

The dialog box shown in Figure 6-6 below pertaining to this model will appear after the main calibration steps are complete. At that point, the fit model can be used (default) or modified values can be entered. A model will be suggested by the algorithm but can be overridden in this dialog.



Typical parameters—Load Model Selection
If multiple ALRM bands are in use, entry fields will appear for each additional band.

Figure 6-6. ALRM MATCH DEVICES CALCULATED VALUES Dialog Box (BAND 1)

The dialog boxes shown above are for coaxial or non-dispersive line types. For waveguides and microstrips, the only change is the addition of items at the top of the dialog to specify the media (see Figure 6-7 and Figure 6-8 below).

An example LRL/ALRM setup dialog for a waveguide line type is shown in Figure 6-7. Note the cutoff and dielectric constant line items that are added.

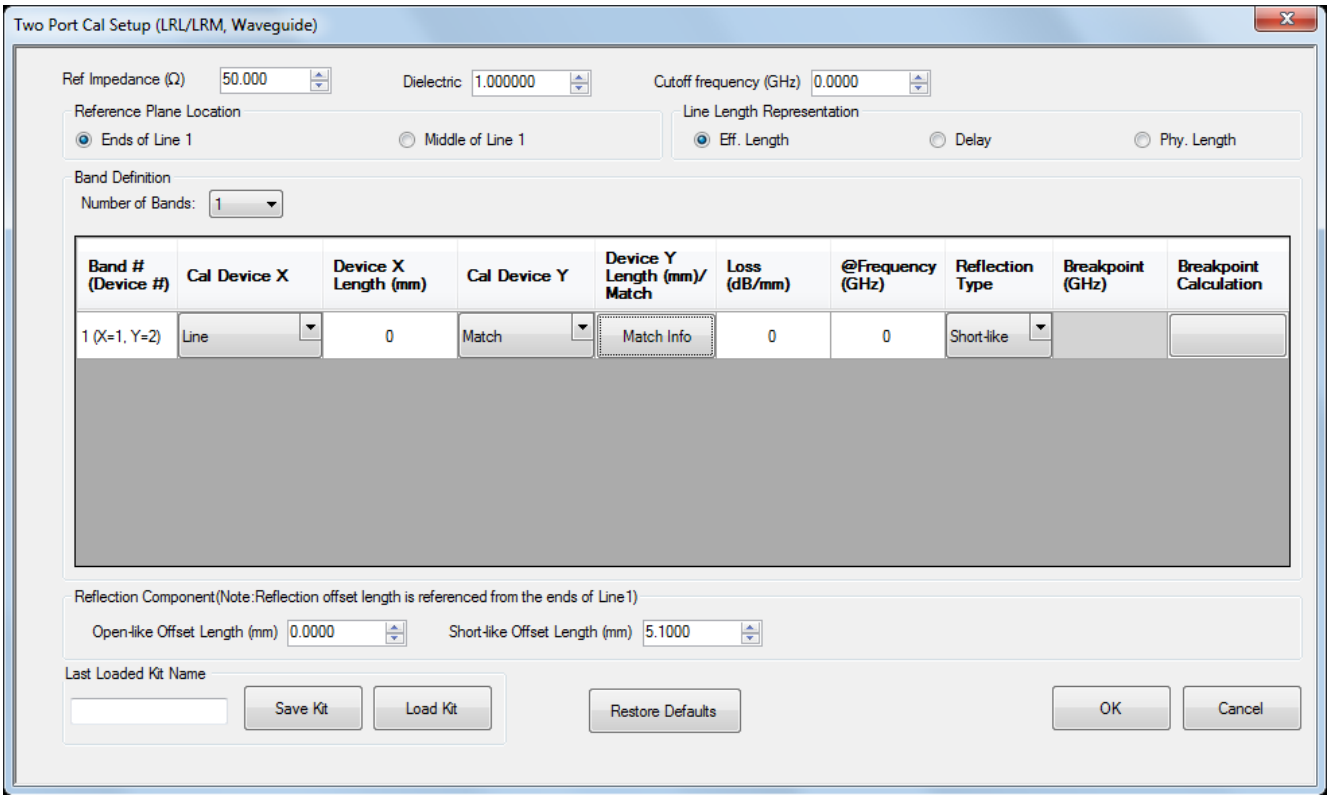


Figure 6-7. Example: LRL/ALRM Setup Dialog for a Waveguide Line Type

An example LRL/ALRM setup dialog for a microstrip line type is shown Figure 6-8. Note the media line items that have been added (the microstrip info sub-dialog is described earlier).

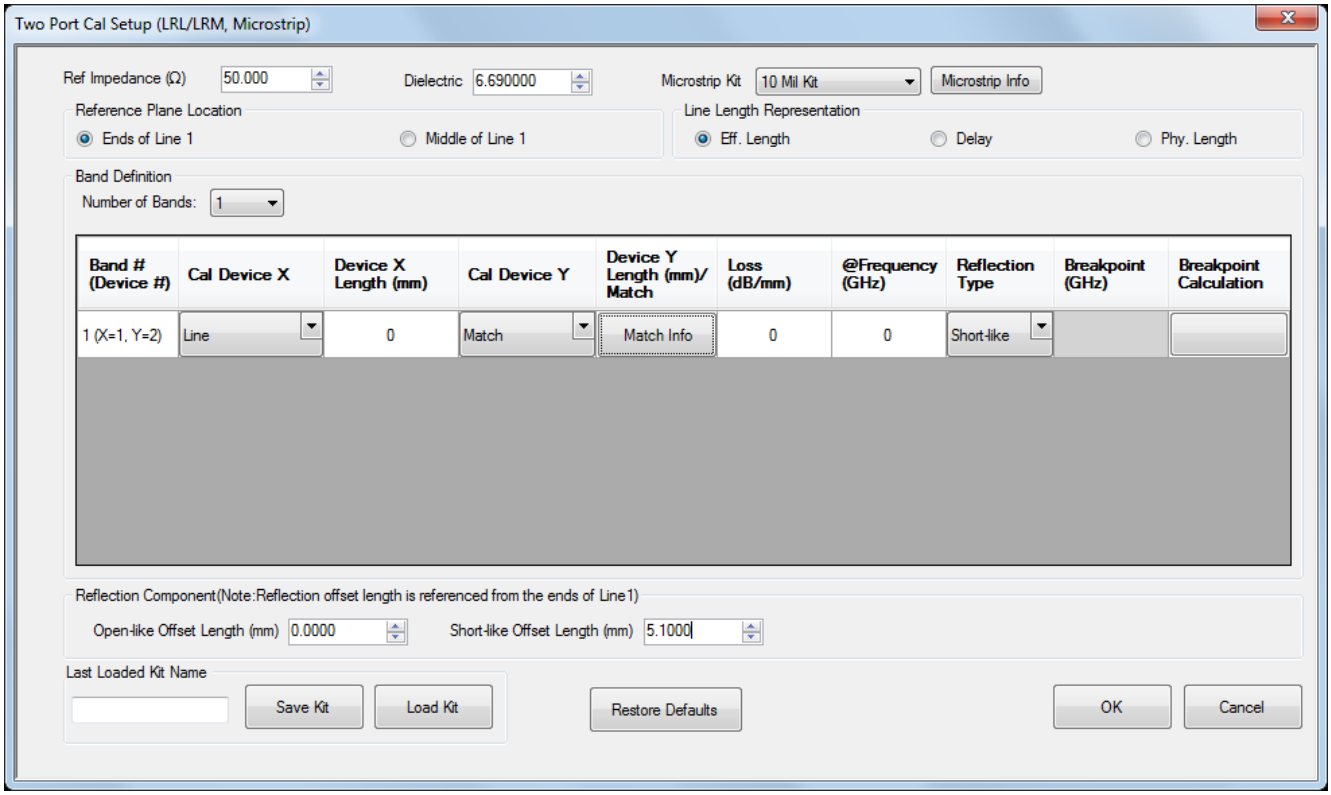


Figure 6-8. Example: LRL/ALRM Setup Dialog for a Microstrip Line Type

When using the double-reflect ALRM method, it is important to note that the reflections must produce distinct reflection coefficients when rotated to the central reference plane. When the reflect offset lengths start to become large, this gets to be more difficult over large frequency ranges. In an on-wafer environment when the offset lengths are typically very short, this does not present a problem but it can be an issue in coax. Since the load modeling is most commonly an issue in high frequency on-wafer measurements, this behavior is usually consistent with the applications.

6-3 LRL/LRM Calibration Example

The following example presumes an MS464xB Series VNA on a coaxial setup. On-wafer scenarios can be accommodated by modifying the entries in Step 4 below. It is assumed that a mating reference plane pair can be created (either MF in coax or zero length-thru compatible). In this example, a full 2-port LRM calibration is performed, although a number of other options are discussed along the way. The implications of these options are further explained in Chapter 1, "Calibration and Measurement Overview".

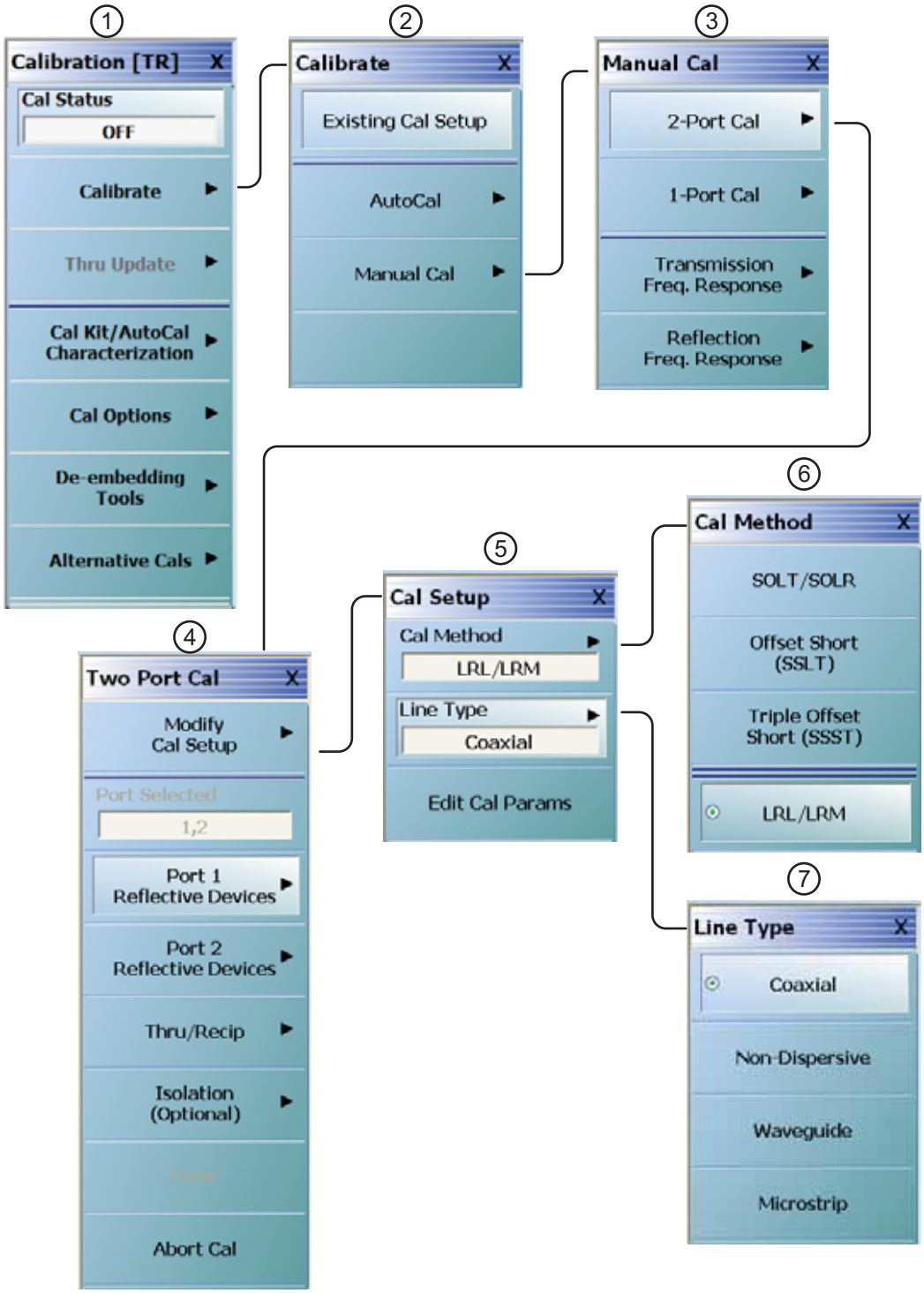


Figure 6-9. Calibration Menu Set for LRL/LRM Coaxial (1 of 2)

1. CALIBRATION menu	5. CAL SETUP menu
2. CALIBRATE menu	6. CAL METHOD menu
3. MANUAL CAL menu	7. LINE TYPE menu
4. TWO PORT CAL menu	

Figure 6-9. Calibration Menu Set for LRL/LRM Coaxial (2 of 2)

Procedure

1. Setup the desired frequency range (**FREQUENCY** menu), power (**POWER** menu) and IFBW/averaging (**AVERAGING** menu). As a default, the IFBW will be 1 kHz and the averaging will be off which is adequate for many applications. The default power level will vary depending on instrument model and options but will often be adequate for all passive and many active device measurements
2. Navigate to the TWO PORT CAL menu.
   - MAIN | Calibration | CALIBRATION | Calibrate | CALIBRATE | Manual Cal | MANUAL CAL | 2 -Port Cal | TWO PORT CAL

Note If a previous cal exists, the Thru Update button will be active. See "Through (Thru) Update" on page 10-1 for more information.

3. Select Modify Cal Setup and on the CAL SETUP menu, select a Cal Method of LRL/LRM and a line type of Coaxial.
4. Select Edit Cal Params to open the TWO PORT CAL SETUP (LRL/LRM, COAXIAL) calibration components dialog box.

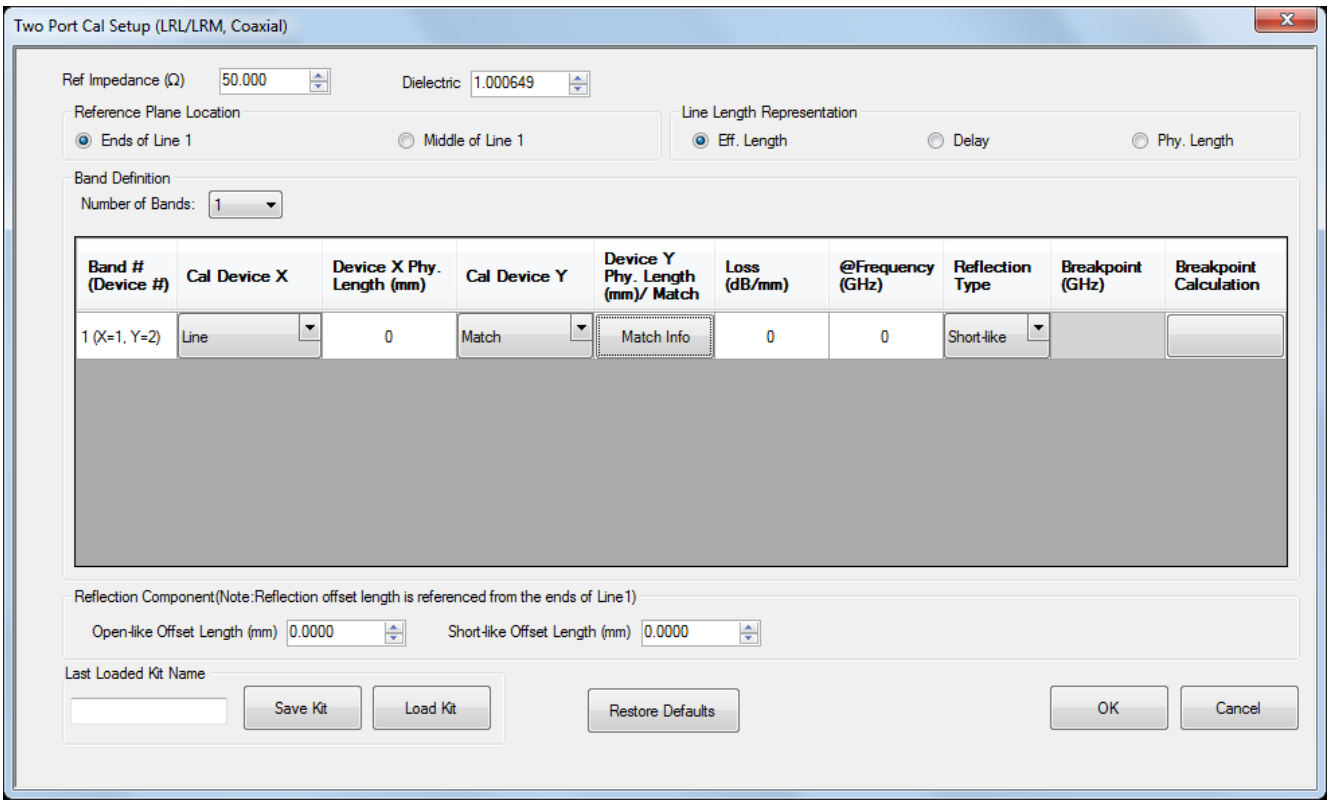


Figure 6-10. Calibration Setup Dialog—TWO PORT CAL SETUP (LRL/LRM, COAXIAL) Dialog Box

5. On the TWO PORT CAL SETUP (LRL/LRM, COAXIAL) dialog box, set the following:
   a. The Reference Impedance establishes the impedance for the load definition, reference plane changes and Smith chart plotting. The default is 50 ohms.
   b. Select Ends Of Line 1 as the Reference Plane Location. Since this example uses a zero length line, the choice makes no difference.
   c. For Number of Bands, select 1. For an LRL calibration, instead of LRM as in this example, the bandwidth of the calibration will be limited by the difference in lengths between the lines. Two bands can be used (with 3 or 4 lines) to cover a larger bandwidth.
   d. For this example, set Band 1 Device 1 to Line with a Line Length of 0 and Line Loss of 0 dB/mm. The reference frequency for the loss can be entered as 0. In general, this forces no loss scaling with frequency and the line loss value entered will be used at all frequencies.
   e. Set Band 1 Device 2 to Match since we are performing an LRM calibration. A specific match (or load) model may be entered under Match Info, but this example uses the default of 50 ohms.
   f. For the Type of Reflection select Use Short-like Component. This implies the impedance is usually less than the reference impedance but it need not be precise. This information is used to help in root selection. Enter the Short-like Offset Length in the Reflection Component section at the bottom of the dialog box.
   g. Click OK to accept the entries and close the dialog box. Select Back at the bottom of the Cal Setup menu to return to the previous level TWO PORT CAL menu.
6. The menu-by-menu procedure is shown in the figure below.

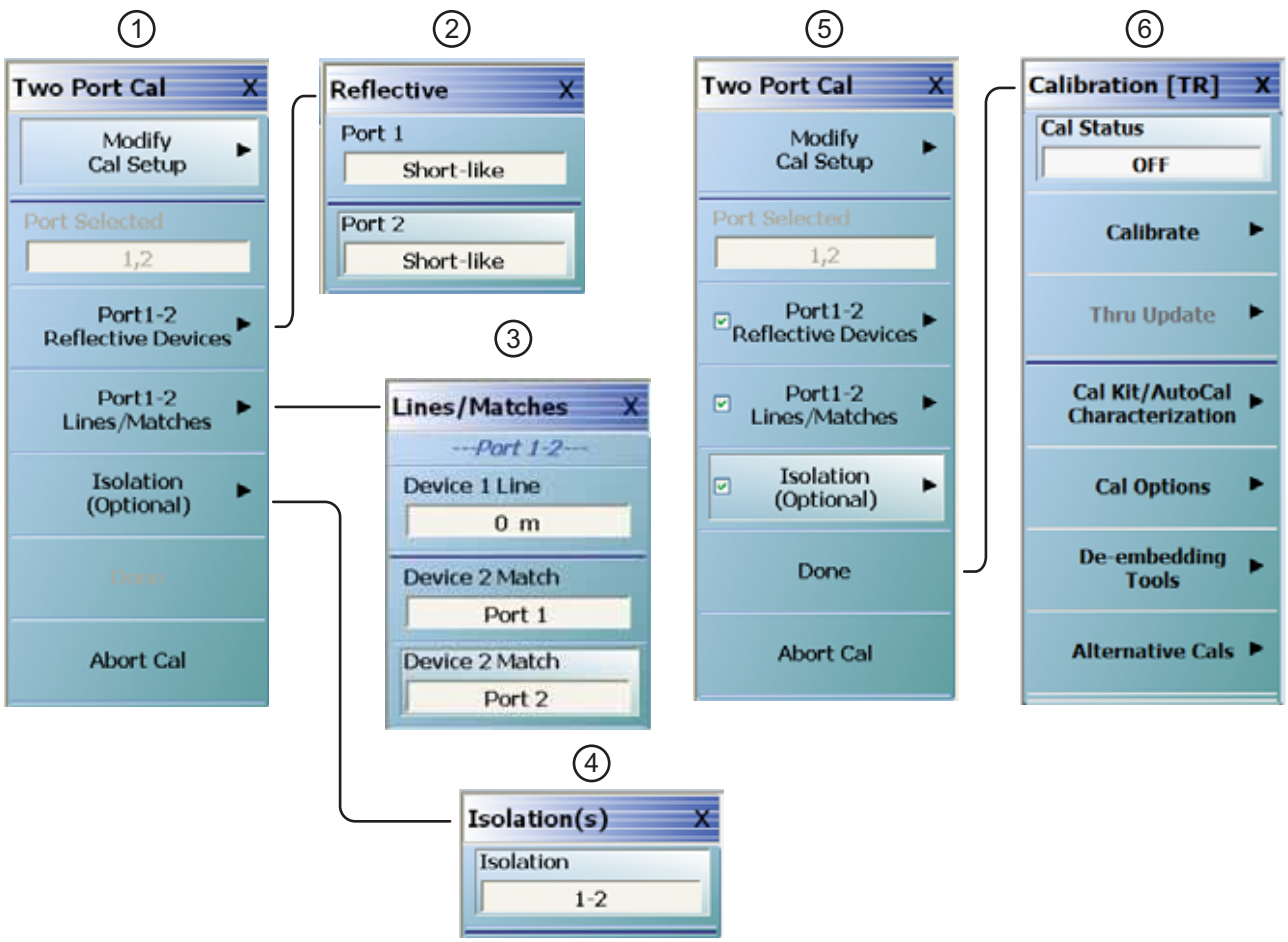


Figure 6-11. Two-Port LRL/LRM Coaxial Calibration Procedure Menus (1 of 2)

1. The TWO PORT CAL menu with link/completion buttons for Port 1-2 Reflective Devices, Port 1-2 Lines/Matches, and Isolation (Optional). The Done button is not available. 2. REFLECTIVE Port 1-2 Menu 3. LINES/MATCHES Menu	4. Optional ISOLATIONS Port 1-2 Menu 5. TWO PORT CAL menu after all calibration procedures successfully completed. Note Done button is available. 6. CALIBRATION Menu with Cal Status set to ON.

Figure 6-11. Two-Port LRL/LRM Coaxial Calibration Procedure Menus (2 of 2)

7. On the TWO PORT CAL menu (above in Figure 6-11—#1), select the Port 1-2 Reflective Devices button and its linked REFLECTIVE menu (Figure 6-11—#2). In this example, the short-like reflective device must be measured at both ports. After the device is connected to a given port, click the corresponding button (a check mark will then appear after the measurement). Click on Back when done with this step to return to the TWO PORT CAL menu.

8. Connect the reference planes together to form the 0 (zero) length line. Select Port 1-2 Lines/Matches and on the LINES/MATCHES menu (Figure 6-11—#3), click on Device 1 Line (0m). Sequentially connect the load (or loads if two models were entered for two physically separate loads) and click the corresponding buttons. For this example, repeat for **Device 2 Match** at **Port 1** and then Device 2 Match at **Port 2**. As before, check marks will appear when a given step is completed. Click BACK when completed and then DONE.

9. An optional isolation step using the Isolation (Optional) button and the linked ISOLATIONS menu (above Figure 6-11—#4) is available but is generally not recommended. If desired, terminate **Port 1** and the end of the cable attached to **Port 2** before clicking on the Isolation 1-2 button.

10. When all procedures are complete (at Figure 6-11—#5), the Done button on the TWO PORT CAL menu is available. Click on Done. The focus returns to the CALIBRATION menu (at Figure 6-11—#6) where the Cal Status button is set to ON.

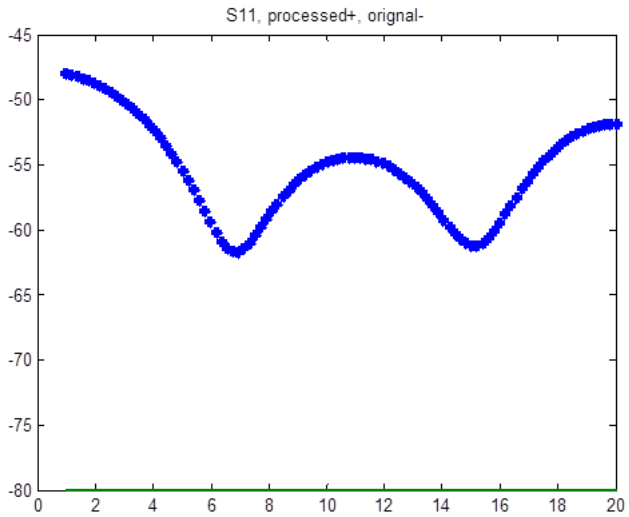
6-4 Hints and Suggestions

Since there are a number of choices involved in setting up the LRL/LRM family of calibrations, some additional hints and points of emphasis may be of assistance:

- Reflect offset lengths are referenced to the ends of Line 1. These lengths are all air-equivalent lengths. The line length entries for the transmission lines are also air-equivalent. If the lengths are known in terms of time delay, the air-equivalent length is given by:

  $\text{time delay (seconds)} \times 2.9978 \times 10^8 \text{ (m/s)}$

- The frequency breakpoint for multi-band calibrations is calculated from the geometric mean of the theoretical upper limit of the low band and the theoretical lower limit of the upper band. The former is calculated based on the 160 degree delta frequency and the latter is calculated from the 10 degree delta frequency. This decision is somewhat arbitrary and is heavily dependent on the materials involved and uncertainties required. With relatively lossy lines, one can approach the limits more closely than with low-loss lines. The impact is also dependent on transmission line quality. If both lines have an impedance of, for example, 50.1 ohms and one measured a theoretical –80 dB termination, the result would be predictable limited by the line impedance imperfection to about –61 dB at all frequencies. If instead, one line was 50.1 ohms and one was 50.2 ohms, then one gets progressively more effect near the theoretical edge frequencies as shown in the figure below.

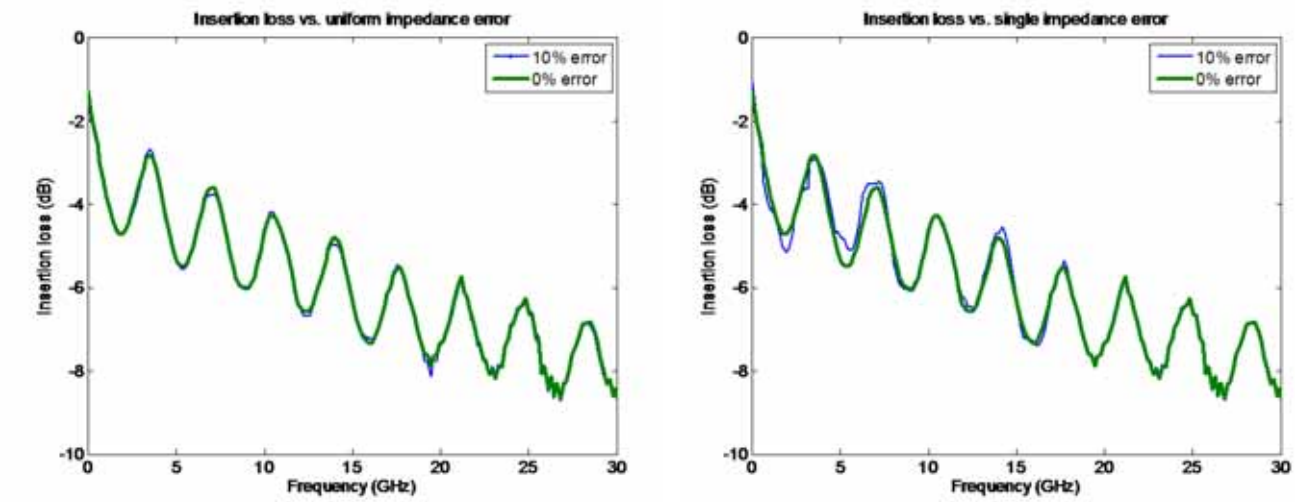


The effect of line impedance problems gets magnified as the line length delta approaches the theoretical band edges (the result for a high RL termination measurement when calibrated with unequal line impedances is shown here). This should be considered when deciding on the 2-band breakpoint frequency.

Figure 6-12. Effect of Line Impedance Problems as Line Length Delta Approaches Theoretical Band Edges

- When doing two band TRL/LRL calibrations, the orientation of the lines between bands can sometimes be confusing. The larger line delta should always be in band 1 (the lower frequency section).
- The TRL family of calibrations is more sensitive to asymmetries in standards (for example: different reflects on the two ports, lines of different impedance) than to problems in the standards themselves. When creating a custom calibration kit, this can be an important point. As an example, the effect on the measurement of a mismatched delay line is shown in Figure 6-13 when there was a global 10% impedance error on the LRL calibration lines and when the error was on only one line.
- While the reflect standard asymmetry can have an effect (particularly on reference plane positioning), this effect is usually weaker than proportionately similar errors on other standards.

- In LRM, the agreement between the actual match standard and the model used will directly impact residual directivity and the measurement of high return losses. Also, since this impedance sets the actual reference impedance of the calibration, if the characteristic of the impedance deviates from that, errors occur. In general the line impedance (for LRL) or the match impedance (for LRM and ALRM) sets the calibration reference impedance. If this value is known, the calibration can be transformed to a different reference impedance using the impedance transform function under the Measurements menu. In ALRM, the accuracy of the net match model will largely determine residual directivity. If the match standard being used is topologically complex for the frequency range being used, the simplified model being optimized for the match may not be adequate and higher residuals will result. As mentioned earlier, ALRM is most successful with physically small and simple matches as might be found in an on-wafer calibration environment.
- While line impedance is central, another critical assumption is the how well the lines approximate ideal transmission lines (i.e., transmission acts like exp(-γ L) where γ is the complex propagation constant). Where the most problems occur is when there are significant launch admittances associated with the line that vary from line to line (non-repeatable connectors, non-repeatable probe placement, etc.). If the admittances were constant, they would just get built into the calibration. If they are large and variable, these can be the largest error source. Care is advised particularly in LRL line sets on a PC board that the connectors are setup as identically as possible, or if probing, that the probe placement is as consistent as possible.
- At the bottom of the dialog is a 'save kit' and 'load kit' capability. This allows one to save and recall an entire LRL/LRM/ALRM definition like a calibration kit independently of the instrument setup. There is, of course, a correlation to the rest of the instrument setup because of breakpoint frequencies in multiband configurations but this allows one to keep a number of kits describing line sets (or lines plus matches) in a central location. The media type is stored as part of the LRL/LRM calibration kit.



TRL/LRL is more sensitive to differences between the line impedances than to the absolute line impedance (although problems on that will shift the reference impedance of the calibration).

Figure 6-13. TRL/LRL Sensitivity to line impedances

- When using microstrip media, it is easy to confuse entries with respect to physical and effective lengths. Make sure the media parameters and radio-button selections are set as needed. The reflection offset entries are always in physical length for microstrip (note the message in the dialog) unlike for the other media types.
- TRL and TRM dialog options are also available. Algorithmically, these are the same as LRL and LRM except line length 1 must be zero. Thus, the dialogs are somewhat simplified and reference plane considerations are less complex. The other topics discussed in this chapter still apply.

Chapter 7 — mTRL Calibration

7-1 Introduction

Multiline TRL (mTRL) is a distinct calibration method from the LRL/LRM family discussed in the previous chapter, but many aspects are shared. All are based on fundamental ideality assumptions about transmission lines (difference between S-parameters of two lines are based solely on the complex propagation constant and line length difference, all lines in the calibration have the same characteristic impedance and propagation constant, etc.). If a match component is used in either family, its measurement data together with any model information will be used explicitly to set directivity.

The LRL family will only use two lines at any given frequency and the user selects which ones through the band definitions (see Chapter 6, "LRL/LRM Calibration" of this guide). mTRL uses all lines (up to 10 are allowed) at all frequencies with their relative contributions determined in a least-squares sense which will de-emphasize those lines with non-optimal lengths (for each frequency) and those that have anomalous measurement data (to help reduce connection repeatability effects). The principles behind the method have been covered extensively in the literature. See, for example,

- R. B. Marks, "A multiline method of network analyzer calibration," *IEEE Trans. Micr. Theory Techn.*, vol. 39, pp. 1205–1215, Jul. 1991.
- D. C. DeGroot, J. A. Jargon, and R. B. Marks, "Multiline TRL revealed," *60th ARFTG Conf. Dig.*, Dec. 2002, pp. 131–155.
- D. F. Williams, C. M. Wang, and U. Arz, "An optimal multiline TRL algorithm," *IEEE MTT-S Int. Micr. Symp. Dig.*, June 2003, pp. 1819–1822.
- Y. Rolain, M. Ishteva, E. Van Nechel, and F. Ferranti, "Multi-line TRL revisited," *85th ARFTG Conf. Dig.*, May 2015, pp. 1–3.

7-2 Setup

Although many implementations are possible, the algorithm here makes use of an optimal line at each frequency (discussed in the first reference above) which has the best line length deltas for that frequency (recall the discussion in the previous chapter that line length deltas should not be close to 0 or 180 degrees). The pairing of this line with all of the other lines forms the set of measurements used in the least-squares analysis. There is considerable freedom in how many lines are used but the shortest line length delta in the set should satisfy the highest frequency to be measured (~<160 degrees of difference at the highest frequency) and the longest line length delta should satisfy the lowest frequency to be measured (~>20 degrees of difference at the lowest frequency although much lower differences are tolerable if the lines have any noticeable loss). Adding additional lines can help further reduce repeatability effects or mitigate the effects of characteristic impedance or propagation constant variation amongst the lines.

As can be seen in the setup dialog in Figure 7-1, the number of lines is selected and then the line lengths entered. The line lengths entered are used for the optimal line selection and the line length delta weighting so the values are important but extreme precision is not. An exception is the line 1 length if the 'ends of line 1' reference plane position is selected since, in that case, line 1 length will be used to rotate out to the selected reference plane. As with the LRL family, the reference planes inherent to the calibration are in the middle of the line space.

Note	The line loss and line frequency are **only** used as part of this reference plane rotation step.

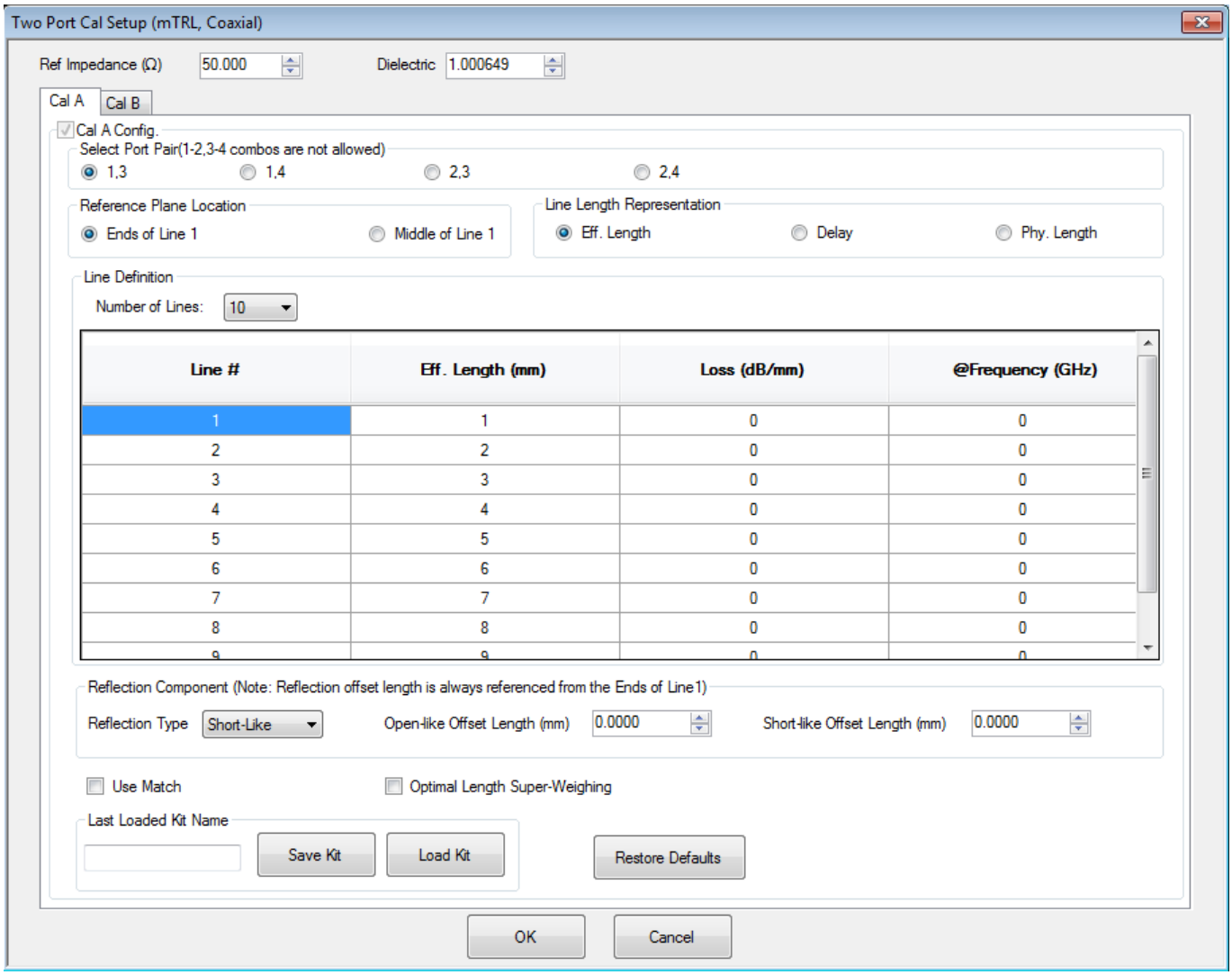


Figure 7-1. mTRL Setup Dialog Box

As with the LRL family, the line length presentation can be selected (and values will be recalculated on the fly using the current media definition) as effective length, physical length or delay. The reflection type is also selected (and offset lengths entered) just as in the LRL family. The calibration kit can be saved and recalled as with the LRL family but the file name extensions are different (.mlcf vs. .lcf; both are text-based XML formats though). The calibration kit files are not interchangeable between the LRL/LRM family and mTRL.

The match element is used quite differently from in LRM (where it was assigned to be used in a specific band or bands) and is slightly distinct from the least-squares solving process used for the rest of mTRL. The premise is if none of the line length deltas are entirely satisfactory for a given frequency point, the match measurement (referenced to the match model provided by the user) will be used for that frequency point. This criterion is set with a minimum electrical length delta entry (expressed in degrees) as shown in the Match Info dialog in Figure 7-2. If the line length deltas fall below that value at a given frequency point, the match data will be used (in an LRM-sense with line 1 data) at that frequency. The default value is 20 degrees. Checking the 'Use Match' box enables the Match measurement and access to the Match Info dialog box (USER DEFINED MATCH DEVICES).

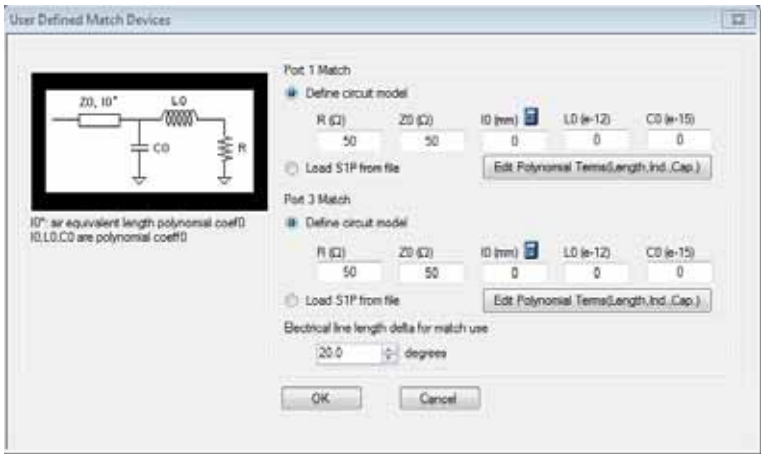


Figure 7-2. Match Information Dialog Box

The match information dialog (Figure 7-2) allows the match model (or .s1p file) to be entered for the standard to be used at each port. The electrical line length delta entry determines when the match data will be used.

The Optimal Length Super-Weighting checkbox applies a weighting to the least-squares computation process to further emphasize the contributions of lines whose deltas relative to the optimal line are closest to 90 degrees at each frequency. While there is already a certain amount of weight applied to this case in the basic method, the selection of Super-Weighting further increases that emphasis. Outlying measurements will be less under-weighted than in the basic method so the use of this check box provides less repeatability immunity but increased line singularity immunity. It may be most useful when the line length collection is skewed so the length delta choices are unfavorable at some parts of the frequency span.

7-3 Measurement Examples and Observations

As an example, four coaxial lines were used in an mTRL calibration (with and without Super-Weighting) and the same four lines were used (in a banded sense) in an LRL calibration (labeled 'normal TRL/LRL' in the plot). These various calibrations were used to measure yet another airline and the return loss results are shown in Figure 7-3.

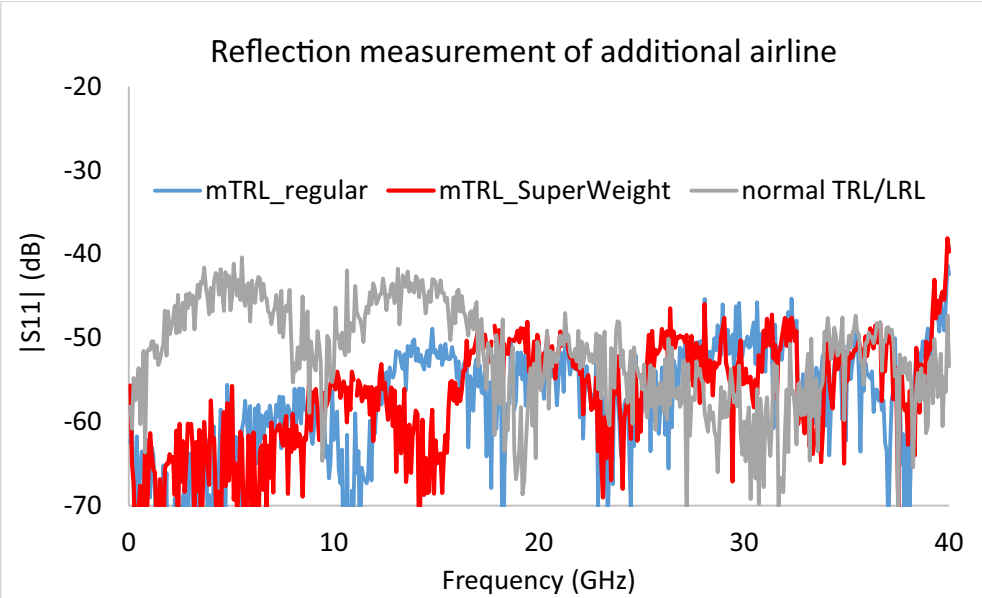
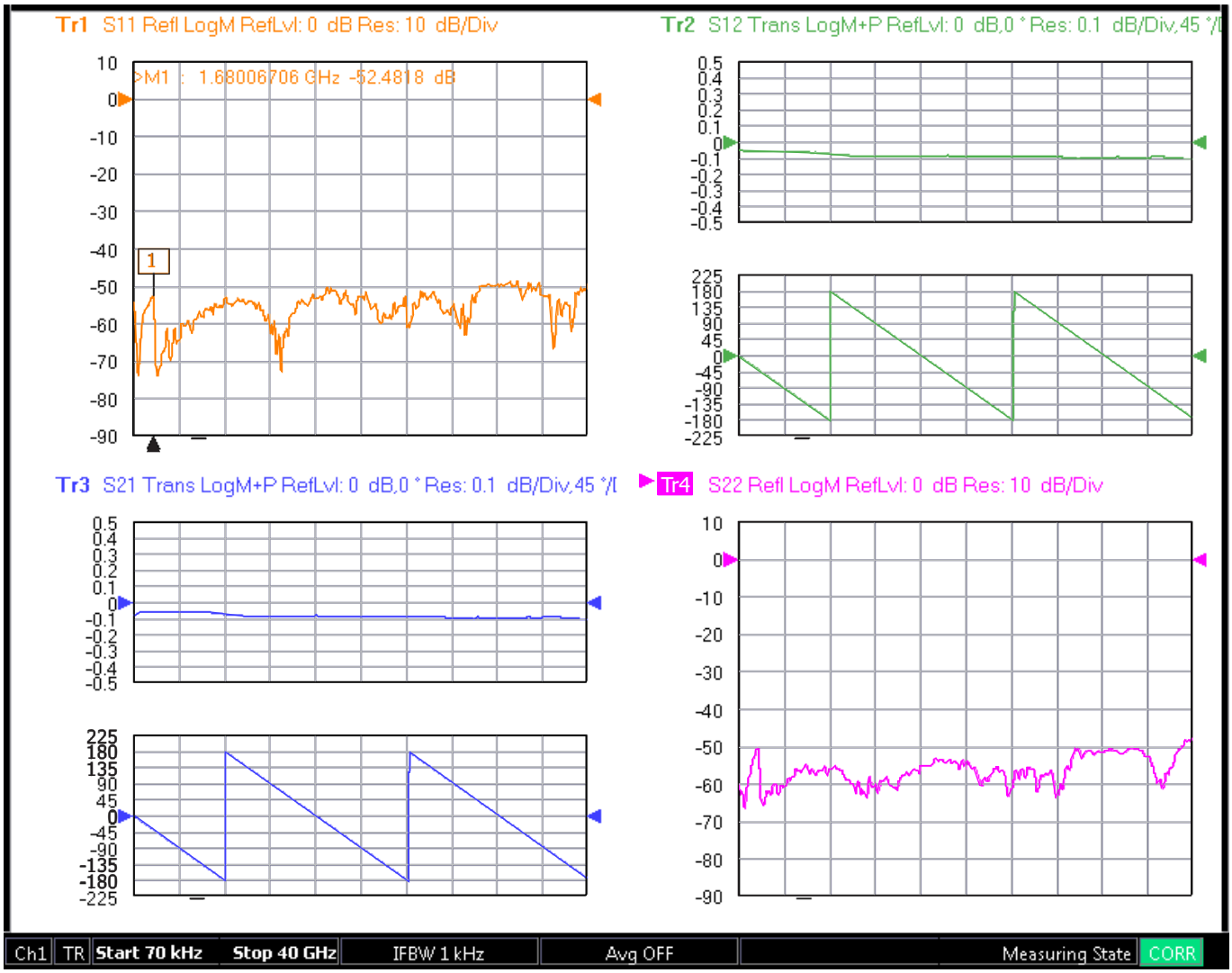


Figure 7-3. Measurements of Return Loss of an Airline After Various mTRL and TRL/LRL Calibrations

The same set of four lines was used for each calibration.

While all of the results are quite good (return loss was expected to be in the area of 60 dB at low frequency and 50 dB near 40 GHz based on dimensional measurements of the DUT and materials knowledge), the mTRL calibration is capable of handling minor differences in characteristic impedances of the calibration lines and differences in connection parasitics. Thus the mTRL calibrations were able to produce results closer to those expected for the DUT. The Optimal Length Super-Weighting did not make a large difference in this calibration since the line length delta distribution was fairly uniform but it did improve the measured return loss values slightly below 20 GHz (making the result even closer to what was expected for the DUT).

As another example, consider a three-line mTRL calibration where match use is enabled with the threshold electrical length delta set to 10 degrees. The calibration was used to measure an additional coaxial airline and the results are shown in Figure 7-4. When the available electrical length difference falls below 10 degrees at low frequency, one can see a transition in return loss (near 1.68 GHz in this example). The return loss is still extremely high but since the match standard and the lines create different reference impedances, seeing a transition should not be unexpected. The match standard used in this measurement was not characterized (except for DC resistance being 50 ohms +/- 0.1 ohms). A smaller transition may be possible with a well-characterized match standard.



Used to Measure an Airline; the Results Show the Match Use Transition Near 1.68 GHz.

Figure 7-4. A Three Line-Plus-Match mTRL Calibration

Other observations:

- As with regular LRM, when a match is used, the agreement between the standard and the model/.s1p file will directly affect residual directivity and the measurement of high return losses. The match impedance sets the calibration reference impedance when it is used (and the 'mean' line characteristic impedance sets the reference impedance when it is not; here 'mean' is in the context of the least-squares solution process).
- As with regular LRL, a central assumption is that the lines are ideal (no launch-related admittances, transmission acts like exp(-γ L) where γ is the complex propagation constant, …).

- Reflect offset lengths are referenced to the end of line 1 and are air-equivalent lengths (= 2.9978 x10^8 (m/s) * Delay (s)). Asymmetry of the reflect standard (between ports) is much more important than the absolute reflection coefficient (which is solved for during the calibration anyway). Still, the sensitivities to reflect standards problems are much weaker than those for the lines (and match, if used).
- Generally, the more repeatability problems there are, the more lines that will be needed for an equal-quality calibration. If the repeatability is extremely poor (~< 20–30 dB), the mTRL/LRL/LRM families may not be the best choice for the planes in question. Doing some full calibration at a stable plane and then using one of the partial information network extraction techniques (see Chapter 9, "Adapter Removal Calibrations and Network Extraction" of this guide for more information) to get to the final reference plane may be a better choice.

Chapter 8 — Receiver Calibrations

8-1 Receiver Calibration Overview

The purpose of this section is to show how receiver calibrations can be set up, and how some of the additional receiver calibration utilities can be used.

Unlike conventional VNA RF calibrations (like SOLT, LRL/LRM, and others) that are used to calibrate the VNA for S-parameter measurements, the receiver calibration is more of an absolute power calibration to help with measurements such as:

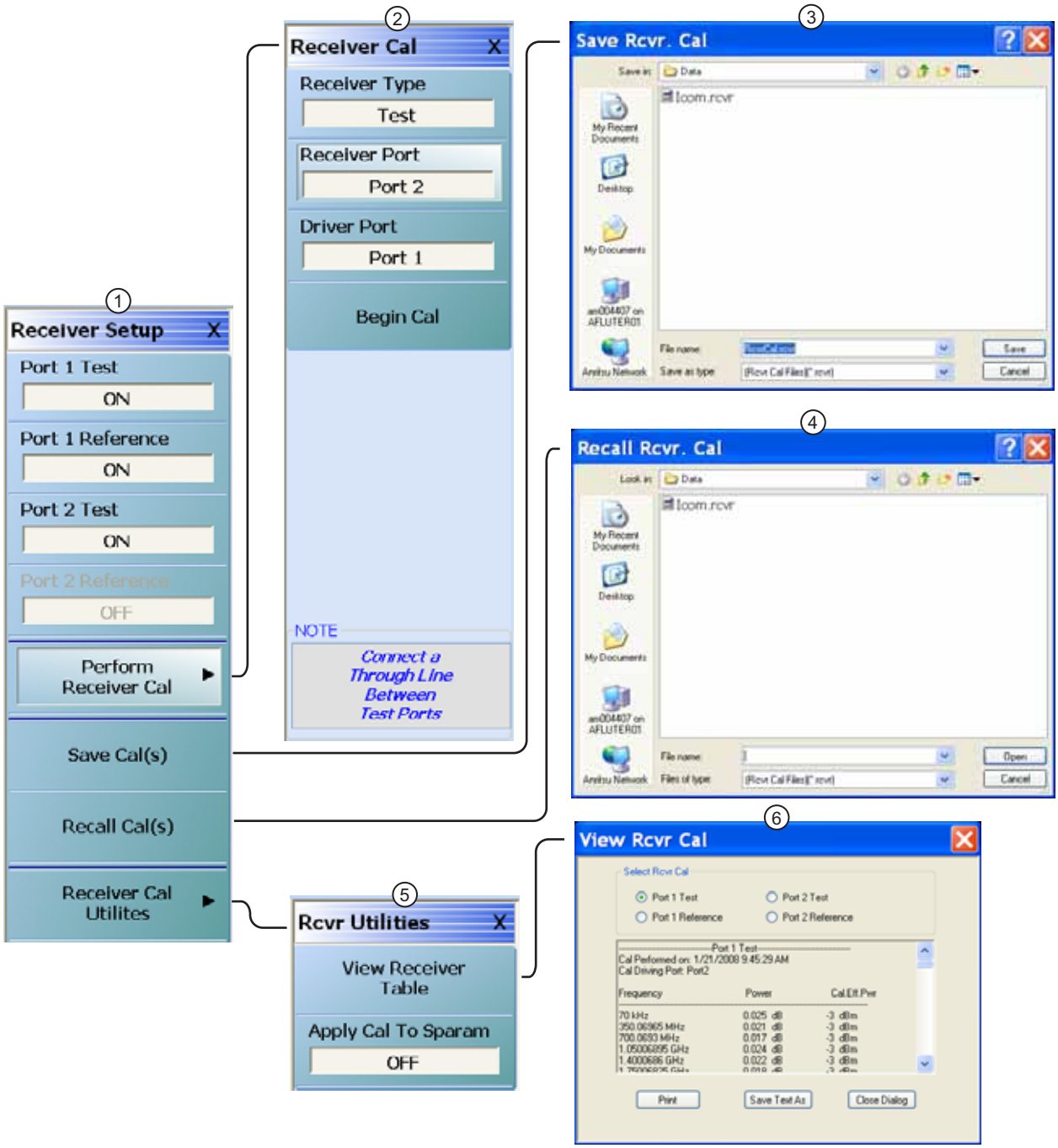
- Harmonics
- IMD and IP3, and other multi-tone distortion measurements
- Mixer conversion loss (in the simpler scalar cases)
- Noise figure measurements
- Other times when the VNA is just used as a channelized receiver

8-2 Receiver Calibration Concepts

The concept of the receiver cal is to take a known source power at some source reference plane and transfer that knowledge to the receiver at a desired receiver reference plane. If it is convenient to use the test port as the source reference plane, the built-in factory ALC calibration can be used to establish the power knowledge. If this is not convenient (because of frequency translation or some other network is required, or greater accuracy is needed), then a power calibration can be performed with the help of a GPIB-controlled power meter to better establish that power knowledge. Power calibrations are covered in more detail in the *Sweep Types* portion of this *Measurement Guide* (Chapter 15, "Measurement—Sweep Types") and in the *Operation Manual*.

In all of these discussions, power refers to signal amplitude at the fundamental frequency. Since the receivers are all tuned, they are not measuring full integrated power as in a power meter. This can be important, as will be discussed in Section 8-6 "Receiver Calibration Uncertainties".

The receiver calibration menu system begins under the Power menu (since it is associated with power reference planes rather than S-parameter calibration). The Receiver Setup menu is shown in Figure 8-1 on page 8-2.



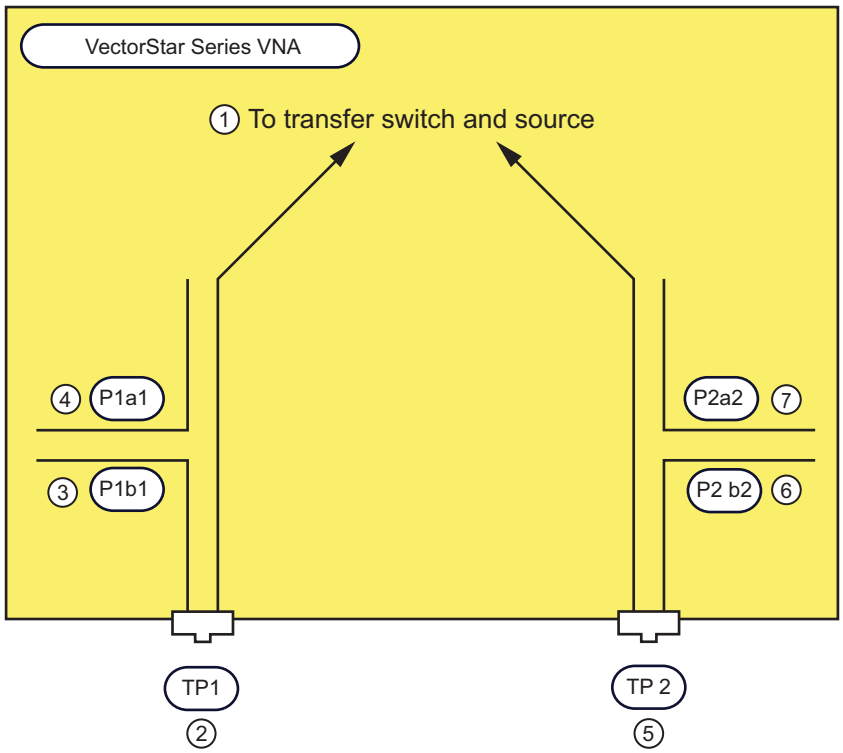
1. RECEIVER SETUP menu MAIN | Power | POWER | Receiver Cal | RECEIVER SETUP	3. SAVE RCVR. CAL dialog box
2. RECEIVER CAL (CALIBRATION) menu	4. RECALL RCVR. CAL dialog box
	5. RCVR UTILITIES menu
	6. VIEW RCVR CAL dialog box

Figure 8-1. RECEIVER SETUP Menu

There are four different receiver calibrations possible (since there are four receivers in the MS464xB Series VNAs) and they can be activated at the top of this menu. A menu item will be active only if that calibration exists. The calibrations can be saved and recalled from this menu. Calibrations can also be saved and recalled as part of the global setup save using the commands located under the File menu.

8-3 Setting Up Receiver Calibrations

To get started, it will help to take a closer look at the architecture of the MS464xB Series VNA as suggested by Figure 8-2.



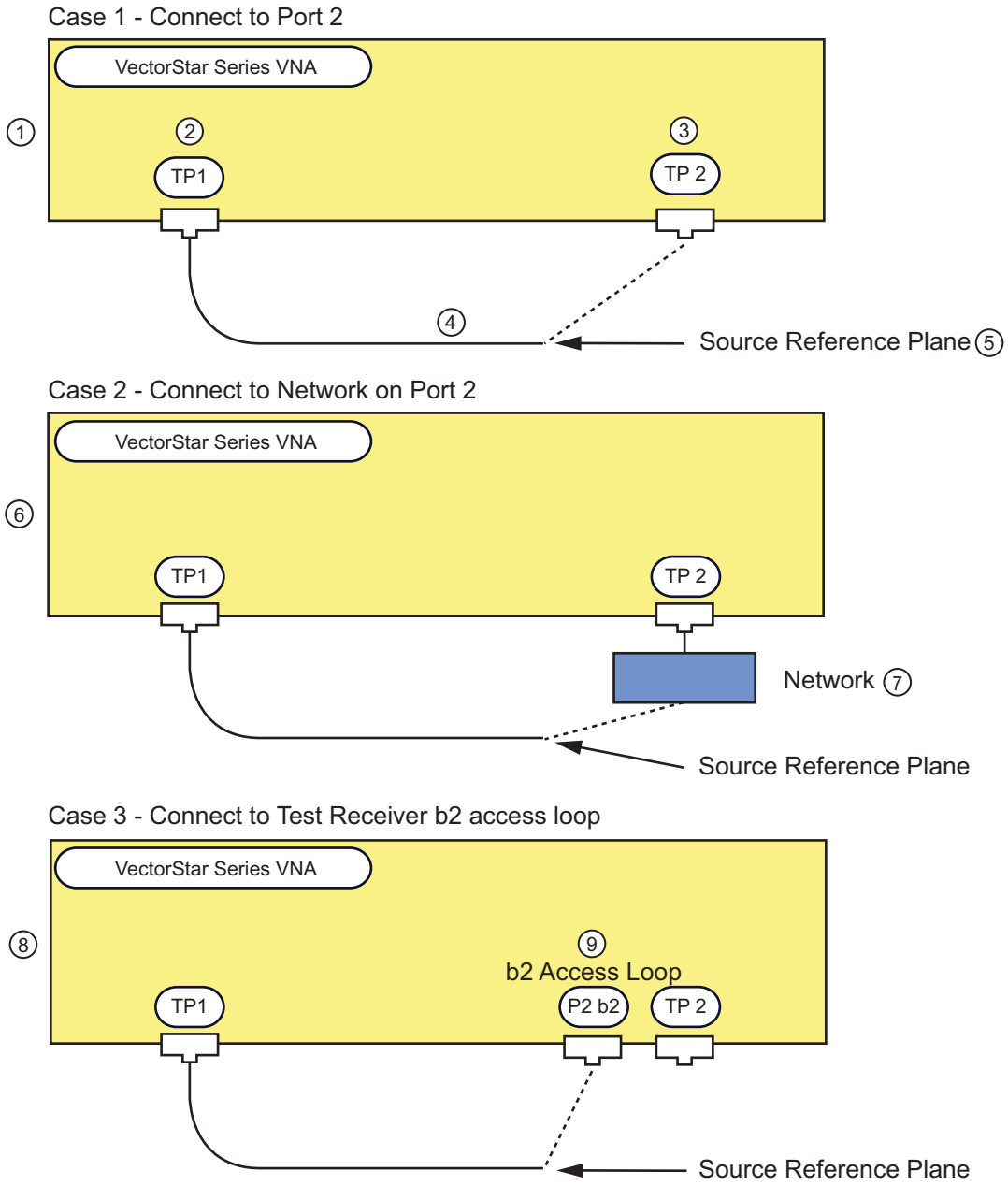
1. To transfer switch and source.	5. TP2—Test Port 2
2. TP1—Test Port 1	6. P2b2—Reference Receiver Port a2
3. P1b1—Reference Receiver Port a1	7. P2a2—Test Receiver Port b2
4. P1a1—Test Receiver Port b1	

Figure 8-2. Simplified Block Diagram of the MS464xB Series VNAs

As this is a 2-port VNA, there are four unique receivers that are performing the measurements: a reference receiver associated with each port (a1 and a2) and a test receiver associated with each port (b1 and b2). An absolute power calibration, which is the receiver calibration, can be associated with each of these receivers.

Note that although the receivers are shown here as associated with couplers, this need not always be the case. With Option 51, 61, and 62, loops are in place between the coupler arms and the receivers, allowing direct access to those receivers. This brings up the point of receiver reference planes. The receiver calibration will establish an absolute power reference, but how the calibration is setup will establish where the reference is being established.

Consider the three cases shown in Figure 8-3. The use of three different receiver reference planes is shown. Where the known source power reference plane is connected for the receiver calibration establishes where that power knowledge is transferred to.



Case 1—Port 2 Direct

1. Case 1—Connect to Port 2
2. Test Port 1
3. Test Port 2
4. Test Cable (f-m)
5. Source Reference Plane

Case 2—Network Connected to Port 2

6. Case 2—Connect to Network Connected to Port 2
7. Network attached to Port 2

Case 3—Port 2 b2 Access Loop

8. Case 3—Connect to Port 2 b2 Access Loop
9. Port 2 b2 Access Loop—Requires VNA equipped with Option 51, 61, or 62.

Figure 8-3. Three Different Receiver Reference Planes

In all of the cases, the source reference plane (where the power is accurately known) is at the end of the dashed cable. This will be connected in different places in the three cases thus establishing the receiver reference plane in three different places. In this case, the absolute power reference has been transferred to (all referring to the Port 2 Test, or b2 receiver calibration in this example set):

- **Case 1**

  To Port 2

- **Case 2**

  To the input of a network connected to Port 2 (could be a pad, a cable assembly, a switch matrix)

- **Case 3**

  To the direct access loop input to the b2 receiver

All of these may be valid receiver reference planes depending on where one wants to connect the DUT (and, more precisely, where one wants to measure power).

The source reference plane is any plane where one accurately knows the signal level. The factory ALC calibration establishes that knowledge at the test ports with moderate accuracy (on the scale of 1 dB). It may be that greater accuracy is needed or it may be that this plane is inconvenient:

- **Cable**

  A cable is needed to reach the receiver reference plane and it is not desired that this cable loss be neglected

- **Preamplifier**

  A preamplifier is needed before the DUT and it is desired that this power level form the reference (if, for example, the receive-side network has very large loss and one wants a better signal-to-noise ratio for the receiver cal)

- **Other**

  Other reasons

For these cases, a power calibration using a GPIB-controlled power meter can be performed at the desired source reference plane prior to performing the receiver calibration. Over reasonable periods of time, the power calibration accuracy can be on the order of 0.1 dB. Aspects of the power calibrations are discussed in the *Sweep Types* section of the *Measurement Guide***,** and in the *Operations Manual* (10410-00317).

A key point in establishing the source reference plane is that the receiver calibration needs to be informed of what that power level is. This is done through the POWER menu. The value entered there is used as the power reference. When using the factory ALC calibration at the port, this link is obvious. When using a power cal at some other reference plane, the power entry field on the POWER menu is also now linked to that reference plane. Thus in both cases, the correct power value is transferred to the receiver cal.

The relationship between the power entry fields and the Target Power field of the power calibration is important. First a few definitions:

ALC entry field:

Where the requested power is entered on the main Power menu for Port 1, Port 2, etc. (to include the >54 GHz fields for certain mmWave modes when present).

Effective power field:

The read-only fields below the ALC entry fields. These normally reflect the settings of source attenuators (if present) but play an additional role when power calibrations are active.

Target power field:

The entry area on the Power Cal submenu where the objective of the power calibration is entered.

The normal procedure is to enter the desired power at the user reference plane (where the power sensor will be connected for the power calibration). If the gain/loss between the VNA port and the user reference plane is small, the ALC entry field can be set to the same value as the Target Power. If the gain/loss is significant, the ALC field should be offset by an estimate of that gain/loss (ALC entry field should be higher if it is a loss, and lower if it is a gain). This offset need not be extremely accurate; it serves to inform the system of where roughly to set the internal sources in order to achieve the desired user reference plane power. After the power calibration is completed, the Effective power field will match the Target power. If the ALC entry field is later changed (from the value at the time of the calibration), the Effective power field will also change to reflect an estimate of the new user reference plane power. This new value is only an estimate since the calibration was not performed at that power level and the accuracy will decrease as the change from the calibrated level increases.

8-4 Receiver Calibration Example

Target power is –20 dBm and there is about 8 dB of loss from the VNA port to the desired reference plane.

1. Set Target Power to –20 dBm and set the ALC entry field (for that port) to –12 dBm. Perform the power calibration.
2. After the calibration is successfully completed, the Effective power will read –20 dBm and the ALC entry field will remain at –12 dBm.
3. Change the ALC entry field to –15 dBm. The Effective power field will change to –23 dBm (the estimate of the new power at the reference plane).
4. Turn the power calibration off. If the ALC entry field is still at –15 dBm, the Effective power field will change to also be at –15 dBm. With the power calibration no longer applied, the system's best estimate of the power being applied is that based on the internal power calibration.

Note that the source attenuator settings (if Option 61 or 62 is installed) will be reflected in the Effective power value even if no power calibration is applied.

Now that the setup is physically ready, the execution of the receiver calibration is performed from the menu in Figure 8-4.



Figure 8-4. RECEIVER CAL (CALIBRATION) Menu

The first two menu items describe which receiver is being calibrated and the third describes the driving port and, hence, the source reference plane. Once the source and receiver reference planes are connected, the Begin Cal button can be selected. The note at the bottom of the menu indicates that a thru line should be connected between ports but what is really meant is that the source and receiver reference planes should be connected together.

During the calibration, different parameters will be displayed as the system collects the appropriate receiver data. Once complete, the calibration will be activated as shown in the Figure 8-1 menu.

8-5 Receiver Calibration Utilities

The RCVR UTILITIES (Receiver Utilities) menu is shown in Figure 8-5.

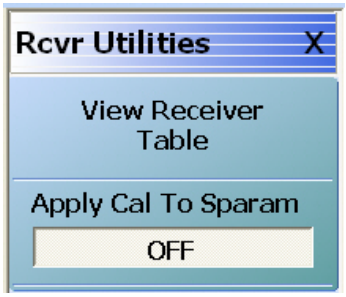


Figure 8-5. RCVR (RECEIVER) UTILITIES Menu

The View Receiver Table button displays the calibration values. This is sometimes valuable as a troubleshooting aid or to gain some more insight into the signal levels in a particular setup. An example Receiver Calibration Table is shown in Figure 8-6. In this example, the table refers to the port 2 test calibration (b2).

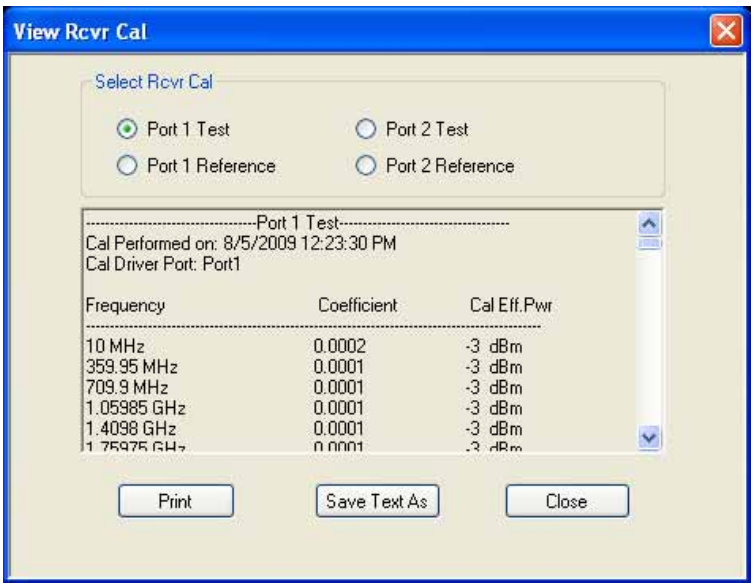
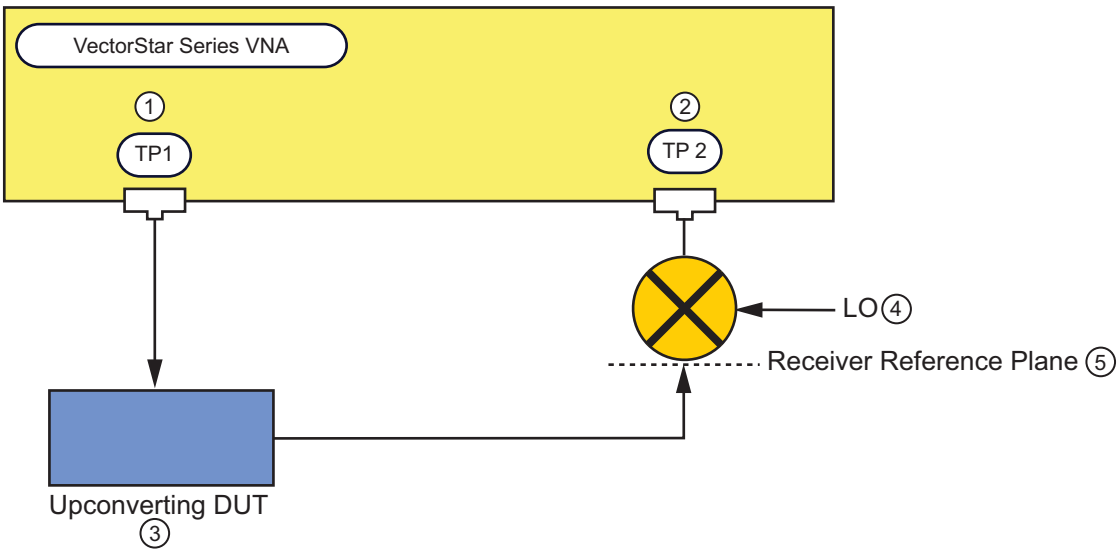


Figure 8-6. VIEW RCVR CAL (RECEIVER CALIBRATION) Dialog Box—Typical Receiver Calibration Table

The receiver calibration is indexed to frequency at the receiver so two vectors of data are key (first two columns). The power level used in the calibration is also shown in the third column as a reference.

The frequency index vector brings up an important point when using receiver calibrations with multiple source mode and certain other frequency conversion modes. It is important that the frequency index match the frequencies present at the receiver reference plane so that the computations can be performed correctly. Consider the setup shown in Figure 8-7 that uses multiple source control (see Chapter 18, "Multiple Source Control (Option 7)" in this *Measurement Guide* and *Multiple Source Application Notes* for more information).



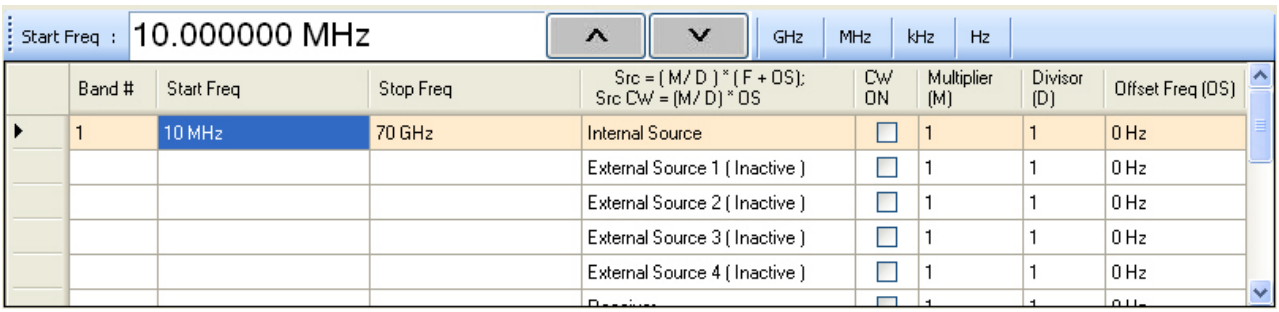
1. Test Port 1	4. LO (Local Oscillator)
2. Test Port 2	5. Receiver Reference Plane
3. Upconverting DUT	

Figure 8-7. Example of a Frequency-Translated Receiver Reference Plane

The DUT in this case is up converting (for example, to millimeter wave frequencies) and a down converter is used to bring the DUT output back to within range of the MS464xB Series VNA. To measure the DUT output power, the receiver reference plane must be in the millimeter wave zone and the frequency list should be referenced here. The receiver source equation in multiple source mode can be used to help here.

The proper indexing can be established using the Receiver Source equation in the multiple source table (Figure 8-8).

- MAIN | Application | APPLICATION | Multiple Source Setup | MULTIPLE SOURCE



Band #	Start Freq	Stop Freq	Src = (M / D) * (F + OS); Src CW = (M / D) * OS	CW ON	Multiplier (M)	Divisor (D)	Offset Freq (OS)
1	10 MHz	70 GHz	Internal Source	☐	1	1	0 Hz
			External Source 1 (Inactive)	☐	1	1	0 Hz
			External Source 2 (Inactive)	☐	1	1	0 Hz
			External Source 3 (Inactive)	☐	1	1	0 Hz
			External Source 4 (Inactive)	☐	1	1	0 Hz

Figure 8-8. Receiver Source Equation From Multiple Source Mode

The purpose of this equation is to provide the proper frequency index for the receiver calibration. See Chapter 18, "Multiple Source Control (Option 7)" for more details.

A final utility controls the interaction between the receiver calibration and ratioed S-parameters. Typically, the receiver calibration pertains to non-ratioed measurements so the correction is not applied to the non-ratioed variables prior to forming the S-parameter ratios. This is done to avoid the confusion of the receiver calibrations interacting with the regular S-parameter calibrations (SOLT, LRL/LRM). The application of the receiver cal is, by default, off for ratios. If desired, the application of the receiver cal to the S-parameter ratios can be turned on (to allow off-line ratio comparison for example) but performing a receiver calibration after an S-parameter calibration will invalidate the S-parameter calibration (as well as trace memory comparisons and other measurements).

8-6 Receiver Calibration Uncertainties

The uncertainty in a receiver calibration is important since it forms a bound on possible uncertainty in many of the measurements listed in the introduction. Since there are many possible setups, we cannot provide a blanket result but can show how this uncertainty can be computed.

Source Side

- Power calibration accuracy (assuming it is used): set by the power meter/sensor and mismatch between the power sensor and the source reference plane.

Receiver Side

- Mismatch between source reference plane and receiver reference plane
- Linearity of the receiver
- Trace noise of the receiver

Receiver linearity is typically better than 0.05 dB as long as one is a few dB away from the compression limits. See the VectorStar VNA data sheet for more information:

- *VectorStar MS464xB Series VNA Technical Data Sheet* (TDS)—11410-00611

Trace noise is typically less than 0.01 dB as long as one is above –10 dBm port equivalent power. The mismatch contribution depends on where the reference planes are. If both return losses are on the order of 20 dB, there could be a contribution of 0.08 dB (best case scenario). The source side uncertainty can be, at best, 0.1 dB but is commonly several tenths.

Typically, these components are combined in an RMS sense and may be as low as 0.5 dB composite. As suggested above, the actual value will be a strong function of the setup being used. The power meter and VNA uncertainty application notes should be consulted for additional information.

The receiver calibration is automatically interpolated and extrapolated (always and independent of the 'interpolation' switch available on the CALIBRATION menu). Generally, the receiver response is a smoothly varying function of frequency so this (linear in magnitude) interpolation does usually not introduce substantial additional uncertainty. If the initial frequency list at the time of receiver calibration is, however, sparse relative to the electrical length of the setup, the errors will increase. Above a few GHz with a simple 1 m cable setup, a transmission receiver response may vary only by 1 dB over a few GHz, so a sparse density will suffice. With a mismatched fixture and multi-meter long cable runs, the step size may need to be in the 10s to 100s of MHz if substantial interpolation is needed and receiver calibration uncertainty needs to remain well under 1 dB.

Another interpolation exception occurs at low receiver frequencies (< 50 MHz) where the system switches to alternate IF frequencies for spur avoidance in certain sub-bands. Interpolating across these edges will result in distortion. In general, major bandswitches (2.5, 5, 10, 20, 38, and 40 GHz for base VNAs) can result in similar issues.

Chapter 9 — Adapter Removal Calibrations and Network Extraction

9-1 Introduction

This chapter describes various methods for handling cases of non-insertable DUTs. In some coaxial cases, this can be handled with a special class of adapter removal calibrations. More generally, de-embedding can be used to remove the effects of fixtures or adapters required to execute the measurement. The de-embedding process itself will be covered in a later chapter, but the means of evaluating the network to be de-embedded will be addressed here.

While it is usually desired to perform a 2-port calibration with mating connectors of the same type, this is sometimes not possible based on the connectors of the device to be tested. Examples of this include:

- The DUT has one N port and one GPC–3.5 port
- The DUT has two female SMA ports and it is not desired to use a non-zero length thru
- The DUT has one fixtured port and one SMA port
- The DUT has one waveguide port and one coaxial port

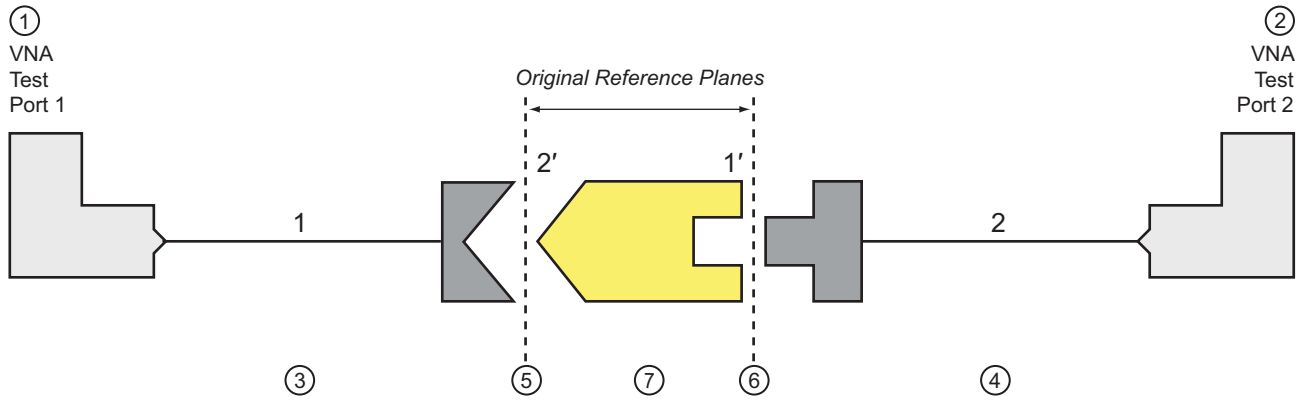
What these examples share is that completing the thru line between the effective test ports requires some kind of adapter or fixture. Since the adapter has some phase length, loss, and mismatch, its effects should be removed for a high quality calibration. Adapter removal is a utility to characterize this adapter and remove its effects from a calibration.

In the more general case of de-embedding, a means of determining the network to be de-embedded must be available. Techniques similar to adapter removal, that is a measurement using one or more calibrations, can be used to extract these parameters. As such, network extraction can be viewed as a generalization of adapter removal.

9-2 Two Related Sets of Reference Planes

The concept of the adapter removal relies on the existence of two related sets of reference planes with one set on either side of the adapter (see Figure 9-1 below). Assuming one can perform a full calibration at each set of reference planes, there is enough information to extract the behavior of the adapter itself. When the calibration is being performed at the reference planes on the left (between Ports 1 and 2'), the adapter behavior is embedded in the characteristics of Port 2'.

Similarly, when the calibration is being performed between Ports 1' and 2, the adapter behavior is embedded in that of Port 1'. Since each of these two calibrations involve mating connector types, these are far easier to perform than the direct 1–2 calibration. It will not be shown here, but the use of the two calibrations provide nearly enough information to extract the parameters of the adapter itself. Figure 9-1 shows the structure of the adapter removal calibration. Two calibrations are performed at the two sets of reference planes shown (between Ports 1 and 2', and between 1' and 2), which allows a determination of the adapter behavior. After the adapter removal, the resulting calibration will be between Ports 1 and 2.



1. Test Port 1	5. Original Reference Plane 2', when adapter is connected to Port 2 Test Cable
2. Test Port 2	6. Original Reference Plane 1', when adapter is connected to Port 1 Test Cable
3. Port 1 Test Cable	7. Adapter to be calibrated
4. Port 2 Test Cable	

Figure 9-1. Adapter Removal Block Diagram

Caveats and Limitations

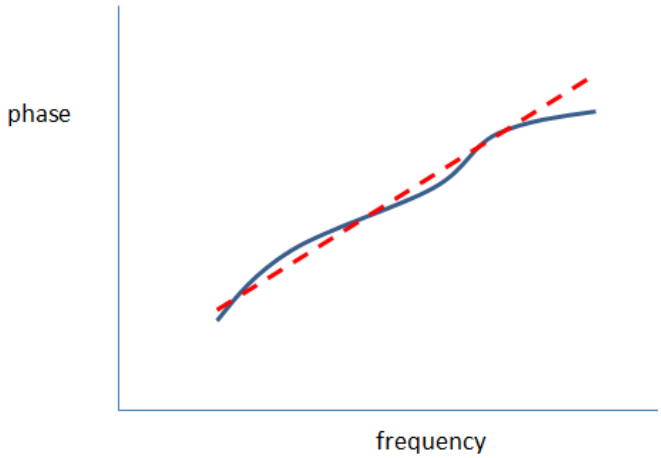
There are two caveats to this procedure.

First, only the $S_{12}S_{21}$ products of the adapter can be determined from this procedure, not the two transmission terms individually. However, since only the product is needed to de-embed the adapter effects, this is not much of a problem. Most adapters are passive and reciprocal anyway, so the individual terms could be determined if necessary.

Second, there is a complex square root operation involved, so a root determination is necessary. To help this, the user must enter a guess as to the electrical length of the adapter (in ps of delay). The guess need not be very accurate, just within the correct half plane. At 2 GHz, this means the error in delay entry should be less than 125 ps to ensure the correct root is selected.

In general the error must be less than $\frac{1}{(4f)}$ where f is the highest frequency being used.

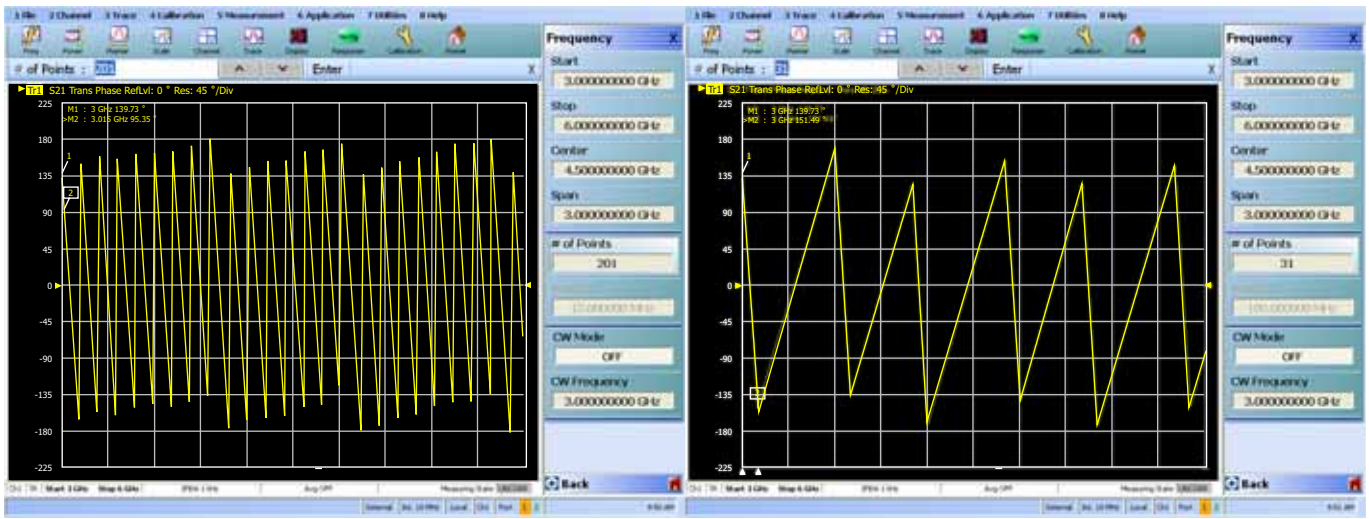
One can enter 0 for the length estimate to force the software to do length estimation internally. This calculation is based on the phase change between frequency points towards the lower end of the sweep range. A linear fit to the phase function is performed and the slope is used to estimate the electrical length as suggested by Figure 9-2. This procedure is quite accurate unless the frequency step size is large relative to the phase change in the measurement. Thus, if the setup uses very long cables, it may help to increase the number of frequency points or at least look at a raw S_{21} phase display (no calibration applied with something of modest insertion loss connected) and see how often the phase wraps.



By entering 0 as the length estimate, the software calculates a length estimate for you. If the frequency step size is large relative to the electrical length of the setup, this may not be advisable.

Figure 9-2. Auto Calculating Length Estimate Issues

In Figure 9-3, one can see how the first frequency step size would be adequate for this setup. The phase change between points is well below 180 degrees so linear fitting will not run into aliasing problems. In the second setup, the phase change between points is nearly 360 degrees so one may start to run into aliasing issues with the 0-entry-length-estimation. For this case, the point count should be increased at least somewhat or a manual length entry can still be used in Adapter Removal.



Uncalibrated phase plots from an example setup are shown here for two different step sizes. In the second case, the automatic length estimation may run into aliasing problems if much more length is added.

Figure 9-3. Uncalibrated Phase Plot Examples

9-3 Performing an Adapter Removal

Two full 2-port calibrations must be performed and those calibrations (plus front panel setups) must be stored to the current directory on a USB memory device or hard disk. The setups for the two calibrations should be the same in terms of frequency range and number of points. Upon entering the adapter removal utility, an estimate for the electrical length of the adapter must be entered, as well as the location of the two calibrations. Once this is done, the utility will generate and apply a new calibration by removing the adapter effects. The menu and help screen for this procedure are shown in Figure 9-4.

Note	ONLY AVAILABLE FOR 2-PORT. For 4-port Adapter Removal, use network extraction for the .snp file of the adapter, and use de-embedding to de-embed it from the calibration.

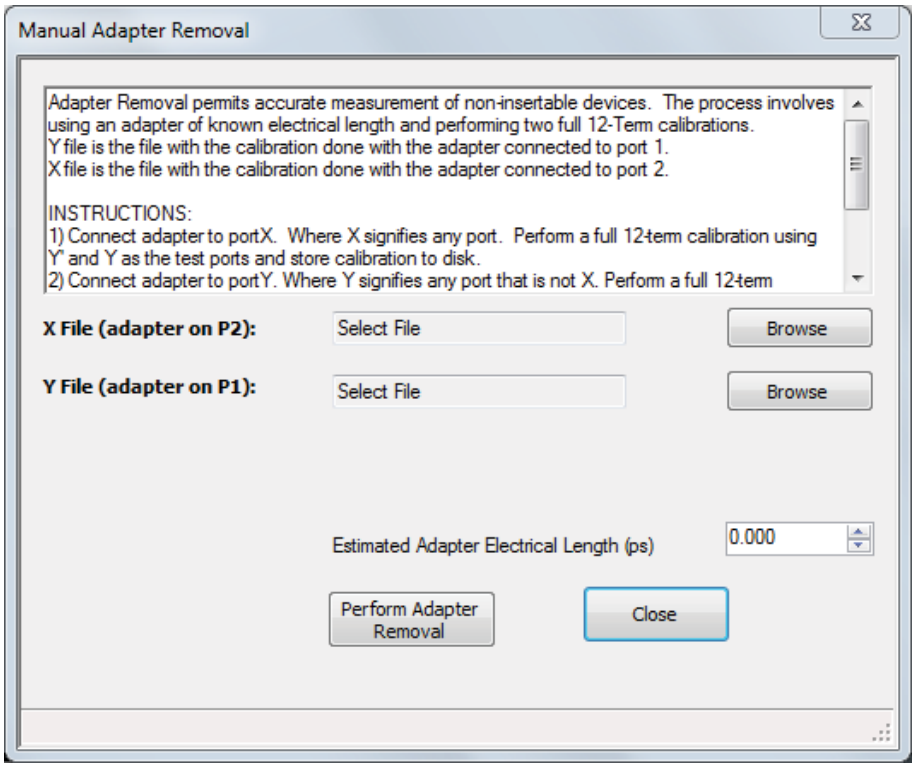


Figure 9-4. MANUAL ADAPTER REMOVAL Dialog Box—2-Port VNAs

Example Adapter Removal

The following example should help illustrate the use of the adapter removal utility. An adapter was constructed with about 3 dB of loss and 180 degrees of phase shift at 3 GHz. This leads to an estimate of the delay length of:

$$\phi = \omega\tau \qquad \tau = \frac{\theta}{\omega} = \frac{\pi}{2\pi(3\times 10^{9})} \approx 167ps$$

Equation 9-1.

Since the loss of this adapter is substantial, one could not simply use reference plane extensions to remove the phase shift and hope for an accurate result.

The two calibrations described earlier were performed and stored to the hard disk and the adapter removal was executed. A thru was then connected without the adapter in place. Normally this would not be possible (since the whole reason for using the adapter removal was for situations when a thru would be difficult), but this example an adapter was constructed just to show that the algorithm functions correctly. The results are shown in below (Figure 9-5 on page 9-5).

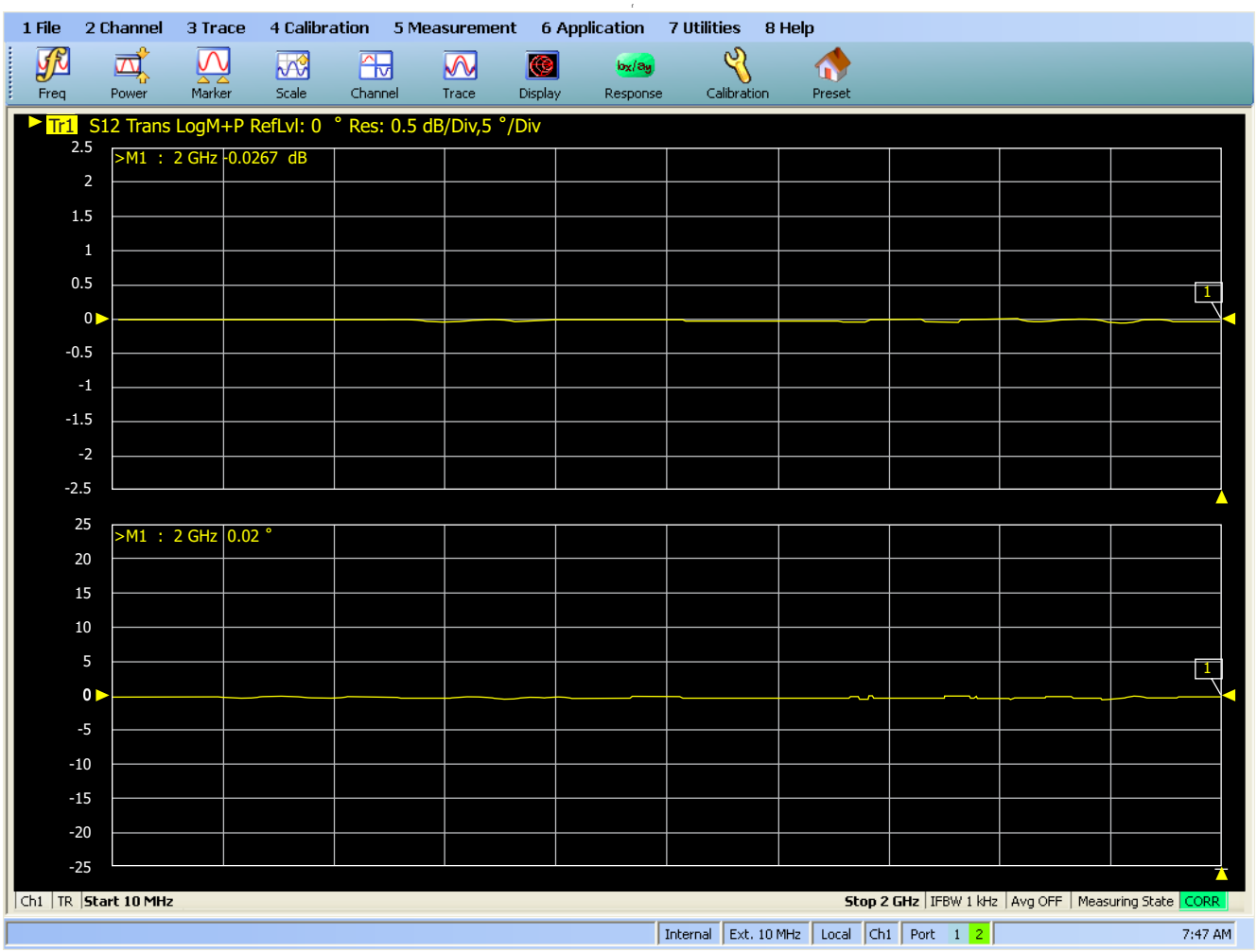


Figure 9-5. VNA Trace Display with Result of Adapter Removal

The thru without the adapter was connected after executing the adapter removal utility and the near-perfect thru values for S_{21} show that the algorithm successfully removed the adapter from the calibration. As expected, the thru without the adapter shows nearly zero insertion loss and phase shift, and a very good match. Any residuals are largely due to cable flex. Had this connection been made with one of the initial two calibrations applied, S_{21} would have shown about 3 dB of gain since the adapter had been built into each calibration.

9-4 Adapter Removal with AutoCal

In the AutoCal-specialized version, adapter removal primarily refers to the case of a gender incompatibility when it is not desired to use test port converters (for example, an M-F AutoCal unit used to establish M-M reference planes). A separate menu item shown below (see Figure 9-7) is provided for AutoCal adapter removal to speed up the process since fewer manual steps are needed. In this calibration sequence, an adapter that can mate the desired reference plane connectors is used as a part of the calibration.

There are two possible ways to accomplish this, both using a pair of calibration sequences to remove the effects of the adapter. In both of the cases below, it is assumed all connectors are from the same family. If not, as with a special inter-series AutoCal unit, this AutoCal-specific adapter removal technique cannot be used. Instead, see the standard adapter removal section.

For an example of the mechanics of the AutoCal module and adapter calibration process, see the figure below (Figure 9-6). Other example figures are available in Chapter 2, “AutoCal Calibration”:

- Figure 2-12, “Precision AutoCal 36585V-2MF Cable Connections for True Thru” on page 2-22
- Figure 2-13, “AutoCal True-Thru Connections” on page 2-23
- Figure 2-15, “AutoCal True-Thru Connections—Module Removed” on page 2-25
- Figure 2-17, “Precision MF AutoCal 36585K-2MF, Adapter Removal, Internal Thru” on page 2-29

MF AutoCal Unit With M-M or F-F Reference Planes

The figure below (Figure 9-6) shows a typical **M-F AutoCal Module** with M-M reference planes.

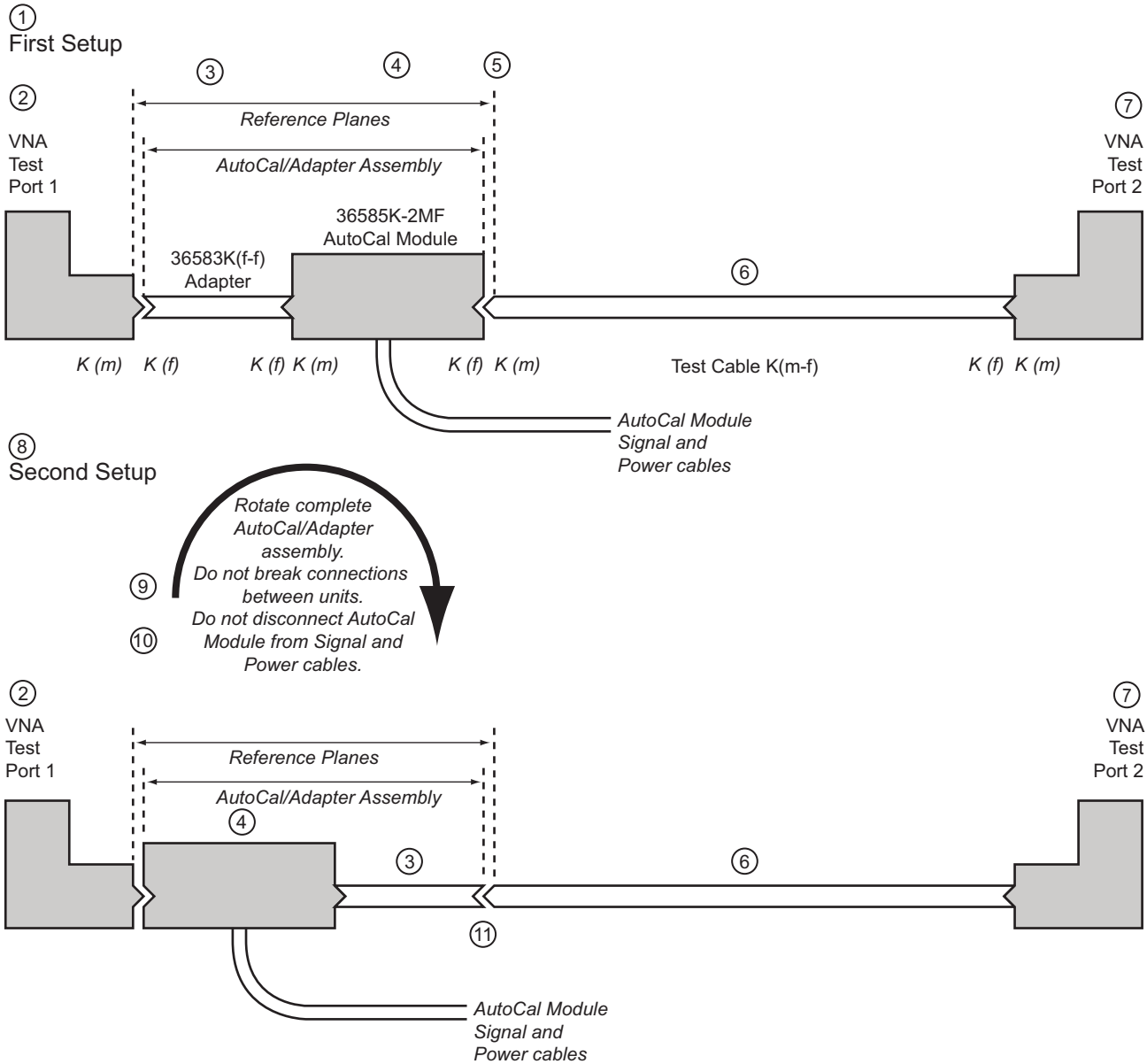


Figure 9-6. AutoCal M-F Module, M-M Reference Planes, and Adapter Removal Calibration (1 of 2)

Part 1—AutoCal Adapter Removal Procedure	Part 2—AutoCal Adapter Removal Procedure
1. First setup for the Adapter Removal procedure. 2. VNA Test Port 1 3. 36583K (f-f) Adapter 4. 36585K-2MF AutoCal Module—For the duration of the calibration, the K (f-f) Adapter and the AutoCal Module must be connected as an assembly. Do not disassemble or disconnect from its Power and Signal cables. 5. Resultant calibration reference planes. 6. K (m-f) Test Cable 7. VNA Test Port 2	8. Second setup for the Adapter Removal procedure. 9. After the first calibration, rotate the complete assembly so that the AutoCal K (f) connector is connected to VNA Test Port 1. Do not disconnect the Adapter from the AutoCal module. 10. Do not disconnect the AutoCal Module from its Power or Signal cables. 11. Connect the 36583 K (f-f) adapter to the Test Cable K (m-f). The reference planes remain in place. The user is guided through the remaining steps of the procedure.

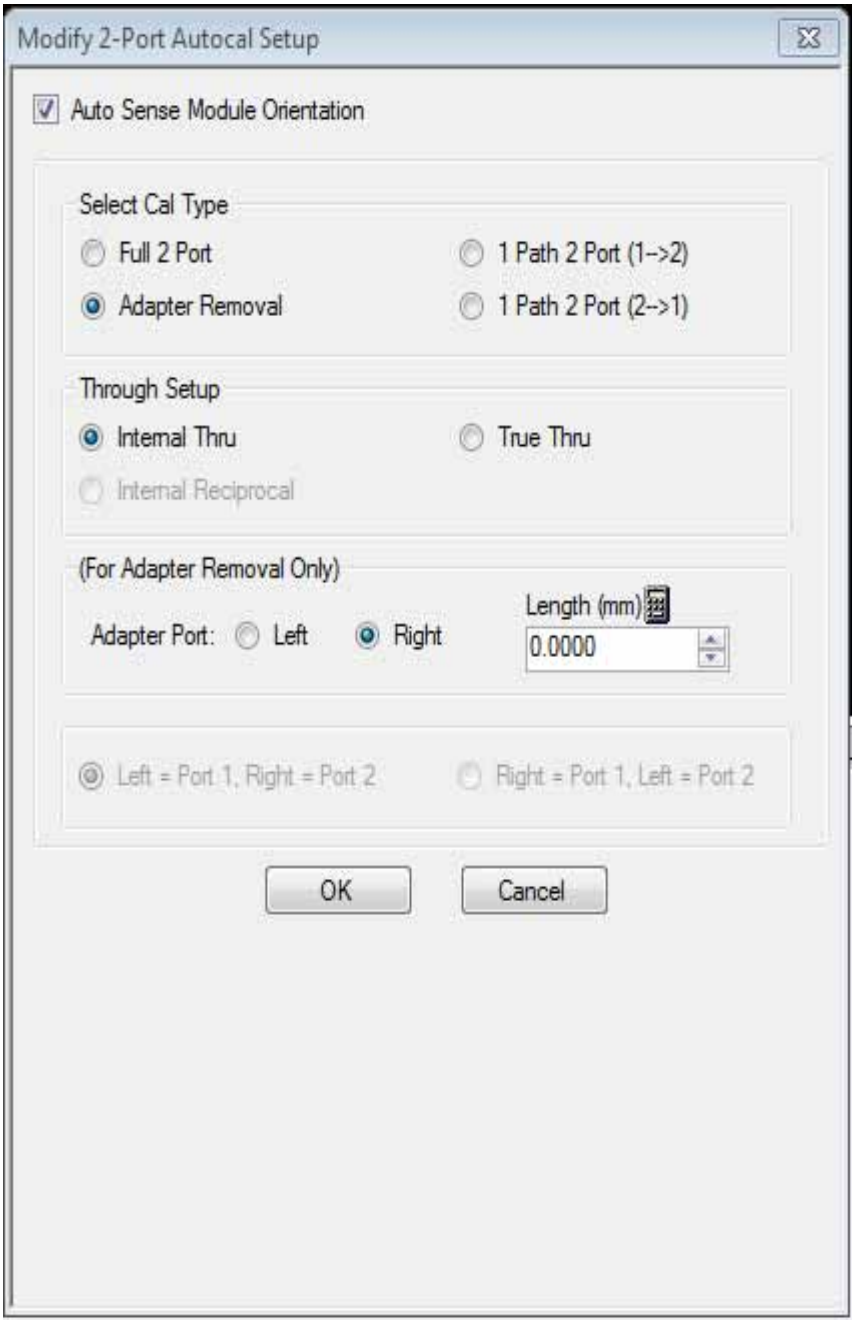
Figure 9-6. AutoCal M-F Module, M-M Reference Planes, and Adapter Removal Calibration (2 of 2)

1. Place the **adapter** on the port of the **AutoCal** unit. The instrument will ask which **port** the adapter is on
2. Connect the **AutoCal** unit and the system will perform the first auto calibration
3. Flip the **AutoCal and adapter assembly** around so that the VNA port connections have changed (do not remove the adapter from the AutoCal assembly).
4. The system will perform a second auto calibration.

M-M or F-F AutoCal Unit With MF Reference Planes

1. Place an **adapter** on the port of the **AutoCal** unit. The instrument will ask which port the adapter is on.
2. Connect the **AutoCal** unit and the system will perform the first auto calibration.
3. Direct-connect the desired MF reference plane (which is, of course possible, by definition here).
4. The system will perform some additional measurements on this thru connection and complete the calibration.

Note	For the 36581x and 36582x AutoCal modules, which only come in M-F versions, only the first type of AutoCal adapter removal will be available.



MODIFY 2-PORT AUTOCAL SETUP Dialog Box—Auto Sense ON—Adapter Removal—Internal Thru

Figure 9-7. AutoCal Setup Dialog with Adapter Removal Selected

9-5 Network Extraction

Overview

De-embedding is covered in detail in Chapter 11, "Calibration and Measurement Enhancements" but since the generation of files for some de-embedding exercises is so closely tied to adapter removal; it will be briefly discussed now. De-embedding is the removal of the effects of a network from a set of data. This network could represent an adapter or fixture, among other things. To perform the de-embedding, the parameters of this network must be known. While there are many methods of deriving these parameters (including simulation), measurement in some way is often preferred. Because of the complex and incompatible media that may be involved, techniques using multiple calibrations (in different connectors or different media) or techniques using a pair of adapters/fixtures back-to-back are sometimes employed.

For the 2-port VNA, there are **four** types of extraction techniques available and are shown in the dialog in Figure 9-8.

For the 2-port VNA, when Option 21—Universal Fixture Extraction is added, there are **six** types of extraction techniques available and are shown in the dialog in Figure 9-9.

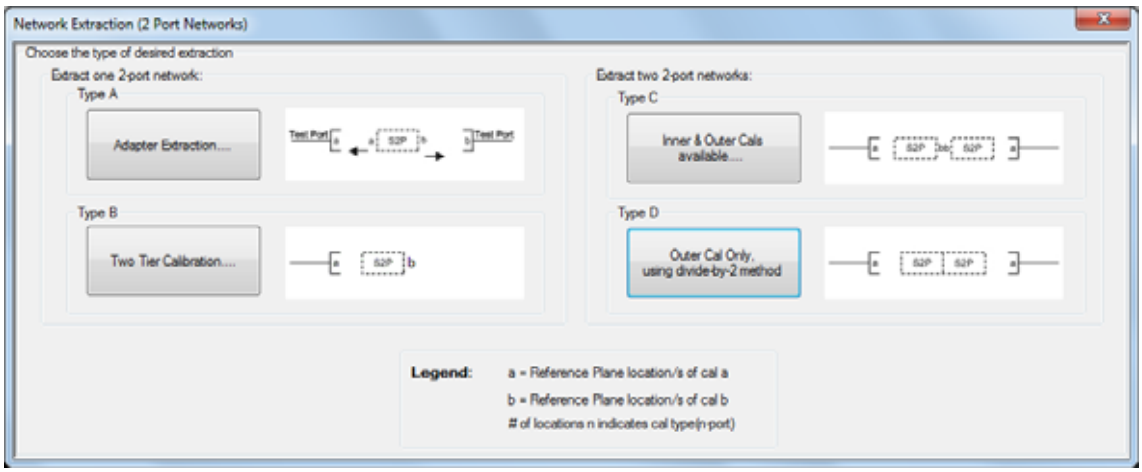


Figure 9-8. NETWORK EXTRACTION Dialog Box—2-Port VNAs

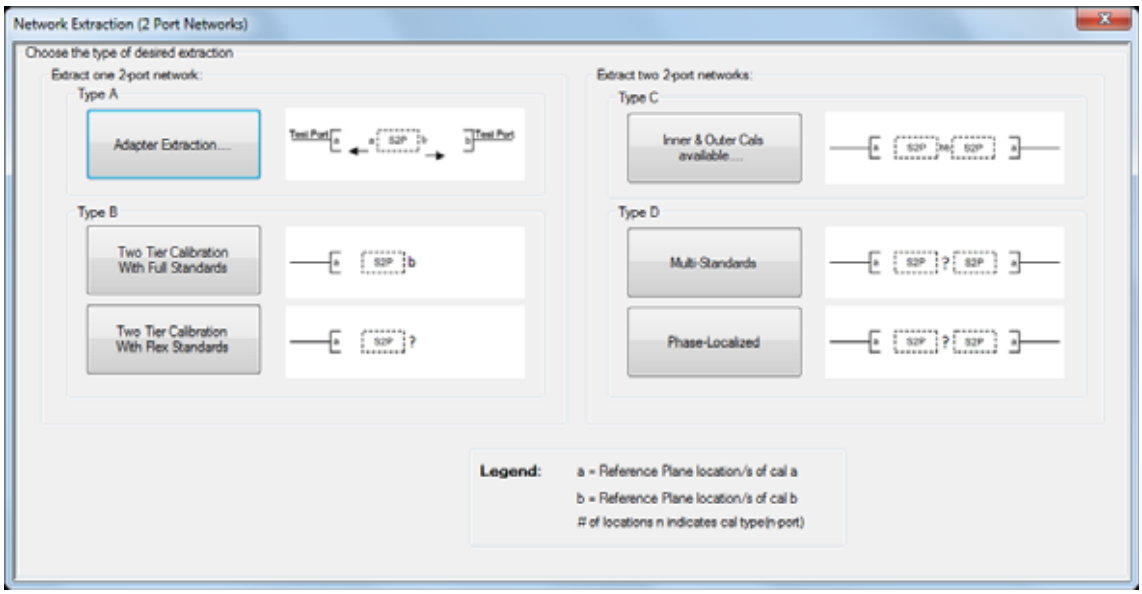


Figure 9-9. NETWORK EXTRACTION Dialog Box—2-Port VNAs (Option 21 Enabled)

Type A—Adapter Extraction

Two full 2-port calibrations are performed; one each with the adapter/fixture attached to one port, then the other. A single S2P file describing the adapter/fixture is generated. This is directly the method of adapter removal except the parameter file is generated explicitly rather than the calibration being directly modified. Refer to "Type A Network Extraction".

Type B—Two Tier Calibration With Full Standards

A two tier calibration, sometimes called 1-port de-embedding or the Bauer-Penfield technique. Here a one port cal is performed, and then additional standards are measured with the adapter/fixture in place. A thru connection is not required, which can be convenient in many cases, and a single .s2p file is generated. In the full standards case (an entire second calibration is required at the end of the adapter/fixture), the calculation is similar to Type A, except the outer match is handled differently. Refer to "Type B Network Extraction".

Type B—Two Tier Calibration With Flex Standards

Algorithmically, this is similar to the full standards case, but a different, or incomplete, calibration may be performed at the fixture output plane. Additional assumptions are made as the standards count dropped (e.g., with one standard, fixture match is neglected).

Type C—Inner and Outer Cals Available

This is the network extraction method available in earlier generations of Anritsu VNAs where full 2-port calibrations are performed at the outer plane (often coaxial or waveguide) and at the inner plane (often a fixtured environment). Two S2P files are generated in this case. Refer to "Type C Network Extraction".

Type D—Outer Cals Using the Divide-By-Two Method (Multi-Standards)

This simplified method is used when standards at the inner plane are difficult to create (as in a complicated fixture structure). Two adapter/fixture "halves" are connected back-to-back and/or with some standards between them and the combination measured using a single outer cal. Assuming the interconnect between the two halves is well-matched and the two halves are identical, S-parameters can be extracted. At least one thru interconnect between halves is needed and an additional (different length) interconnect can be used or high reflection standards can be used at the inner plane. If only one interconnect is used, inner plane match is assumed perfect. Additional information is obtained and accuracy generally improves as more standards are added. Two S2P files are generated. Refer to "Type D Network Extraction—Multi Standards (with Option 21)".

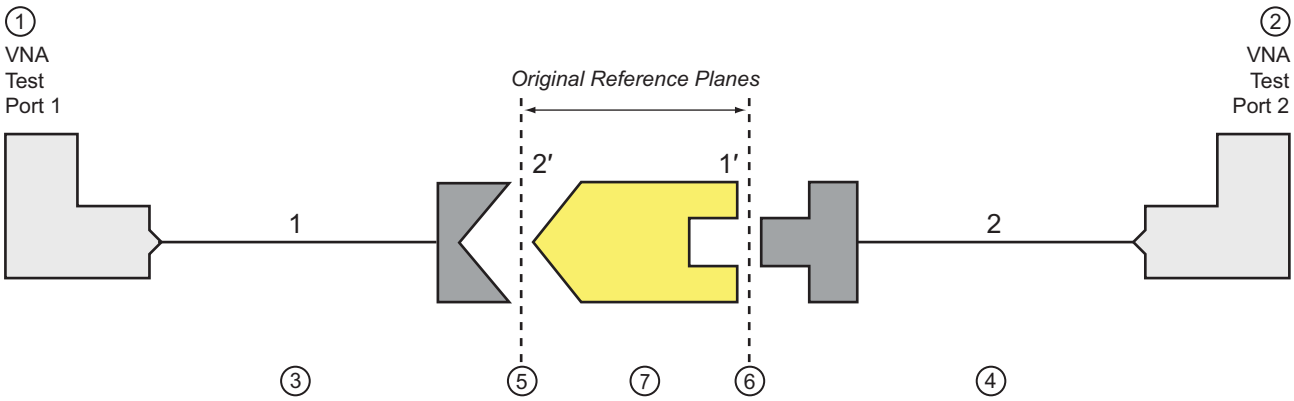
Type D—Phase Localized

A variation of type D makes use of knowledge of fixture length (through user entry or model fitting) to better localize mismatch and enable a more accurate extraction if the fixture is electrically long enough. Refer to "Type D Network Extraction—Phase Localized (with Option 21)".

Note	For multiport systems, In Chapter 24, "Multiport Measurements" in the section titled "Network Extraction", extraction types E, F, and G are discussed.

Type A Network Extraction

As discussed above, the Type A extraction uses exactly the same procedure and algorithm as adapter removal. Instead of directly modifying the calibration to remove the effects of the adapter/fixture, however, the S-parameters of the adapter/fixture are exported to an S2P file for later de-embedding or other uses. The reference plane diagram is repeated below for convenience. Two full 2-port calibrations are required, one with the adapter on **Port 1** (so the cal is between 1' and 2) and one with the adapted on **Port 2** (so the cal is between 1 and 2').



1. Test Port 1
2. Test Port 2
3. Port 1 Test Cable
4. Port 2 Test Cable
5. Original Reference Plane 2', when adapter is connected to Port 2 Test Cable
6. Original Reference Plane 1', when adapter is connected to Port 1 Test Cable
7. Adapter to be calibrated

Figure 9-10. Adapter Removal Block Diagram

The calibrations are performed and then the setups saved typically as an active channel CHX file type. The extraction dialog is shown below where these files are retrieved to perform the process. After "Perform Extraction" is selected, a new dialog will appear asking for the file name where the S2P data should be saved.

As with adapter removal, a few caveats apply

- The two calibration files must have the same frequency lists (i.e., same frequency range and same number of points).
- The cal algorithms and media types may be different but they must both be full 2-port cals
- The adapter is assumed to be reciprocal ($S_{12} = S_{21}$).

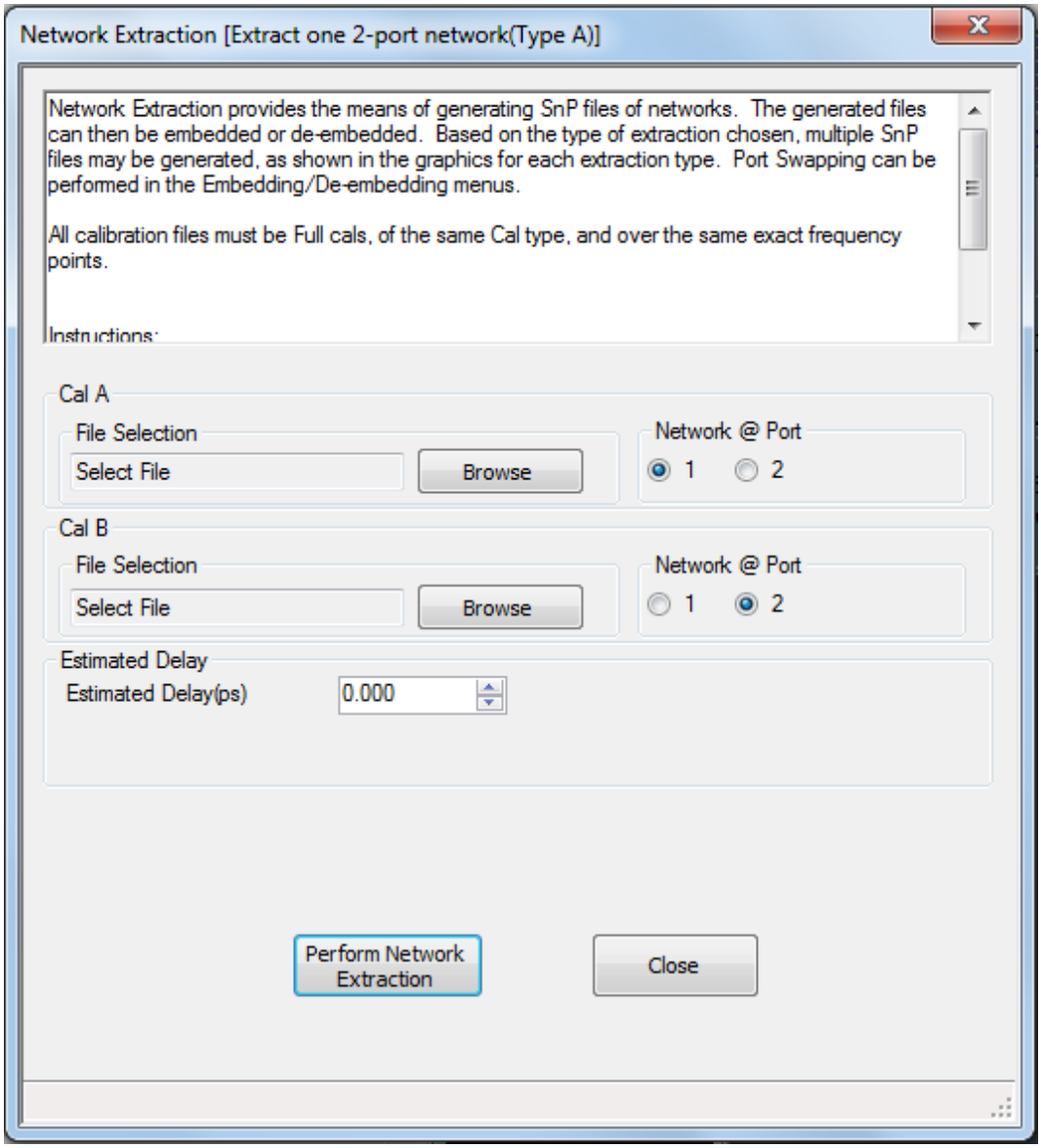


Figure 9-11. NETWORK EXTRACTION Dialog Box—Extract One 2-Port Network—Type A—2-Port VNAs

Type B Network Extraction

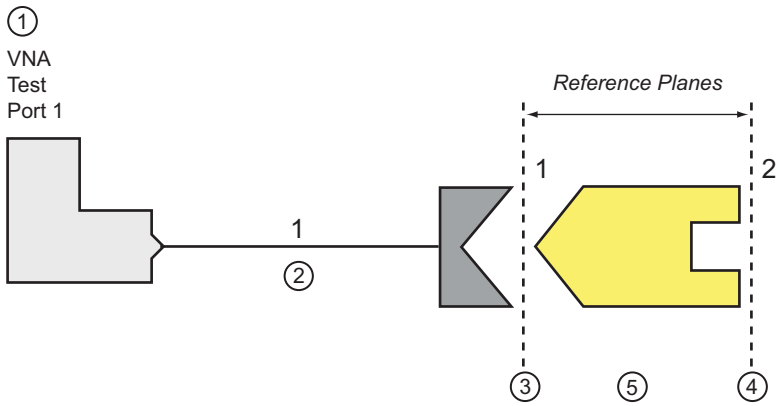
The Type B extraction is a simplification of Type A in that it only requires one port standards. There are two variations of Type B: full standards and flex standards. Both variants can be useful if a thru connect is difficult to implement because of adapter/fixture configuration issues. Below are descriptions of the two versions:

Full Standards: Requires three reflect standards (a full 1-port calibration) at the far end of the fixture arm. This completely solves the reciprocal error box describing the fixture but does require three known standards.

Flex Standards: This allows 1, 2 or 3 reflection standards to be used at the far end of the fixture arm for cases when a variety of different standards may be available. The three-standard case is simply a generalization of the 'full calibration' case where a variety of different standards can be easily tried if independent (e.g., 2 offset opens and a load, 2 offset shorts and an offset open, etc.). The one and two standards cases make assumptions about the fixture (partial information techniques) for cases when few known standards are available. An additional option will be required (Option 21) for this choice to be available.

Type B Network Extraction—Full Standards

The full calibration algorithm has a long history and is covered in the literature extensively (for example, R. Bauer and P. Penfield, "De-embedding and unterminating," *IEEE Trans. Micr. Theory Tech.*, vol. 22, pp.282–288, Mar. 1974.). As suggested by the figure below, a calibration is performed at Plane 1 (often a coaxial or waveguide calibration) and a second calibration is performed at Plane 2 (could also be coaxial or waveguide in the case of an adapter or could be more complicated in the case of a fixture).



1. Test Port 1	4. Second Reference Plane at end of Adapter DUT
2. Port 1 Test Cable	5. Adapter DUT
3. Original Reference Plane for Port 1	

Figure 9-12. Network Extraction Type B Reference Planes

As before, the cals are performed and the setups saved, typically as an active channel CHX file type. The cal files are retrieved using the dialog shown below.

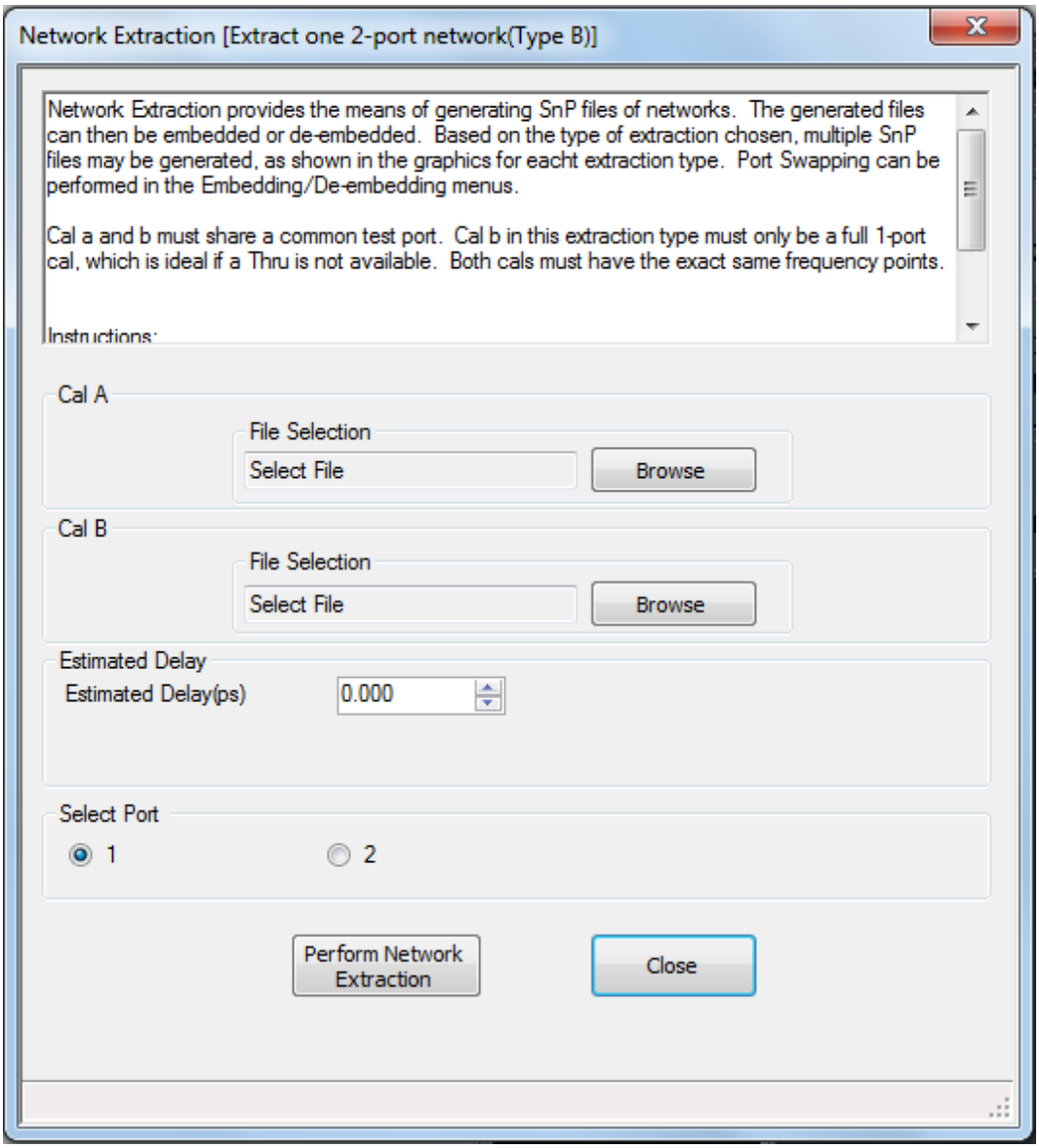


Figure 9-13. NETWORK EXTRACTION Dialog Box—Extract One 2-Port Network—Type B—With Full Standards

After Perform Network Extraction is selected, another dialog appears asking for the file name where the resulting S2P data should be saved. Different cal algorithms and media types may be selected but at least Cal B must be 1-port only. Cal A can be a full 2-port cal or a double 1-port cal (this way it is known how to compute the extraction). As with Type A, the adapter/fixture is assumed to be reciprocal and the frequency lists must be the same.

The results obtained with Type B may be somewhat different from those obtained with Type A since the algorithms are not the same. The main differences will be with respect to outboard match (at Plane 2 in Figure 9-12 above). In Type A, this match is determined with a full reflectometer solve while in Type B, it is determined with a source-match like extraction on the error X of the second cal. As a result, the Type B extraction of this match will be somewhat more sensitive to cal quality than will Type A (particularly with regard to the source match-determining cal components: O and S in OSL or the two shorts in SSL). The trade-off is simplicity and, in some cases, practicality.

Type B Network Extraction—Flex Standards (with Option 21)

There are a number of cases where it is not practical to perform a conventional 1-port calibration at the far end of the fixture arm. An unusual set of three standards may be all that is available or only one or two reasonably-well-known standards exist. The 'flex standards' variant handles these cases. As with 'full standards' the user will select the port in question, an estimate of the length of the arm (and use 0 for an automatic estimate) and where to save the .s2p file for the fixture arm. Also as in the 'full standards' case, the fixture will be assumed to be reciprocal but not necessarily symmetric. In this variant, the user must also select the number of standards as shown in the below dialog.

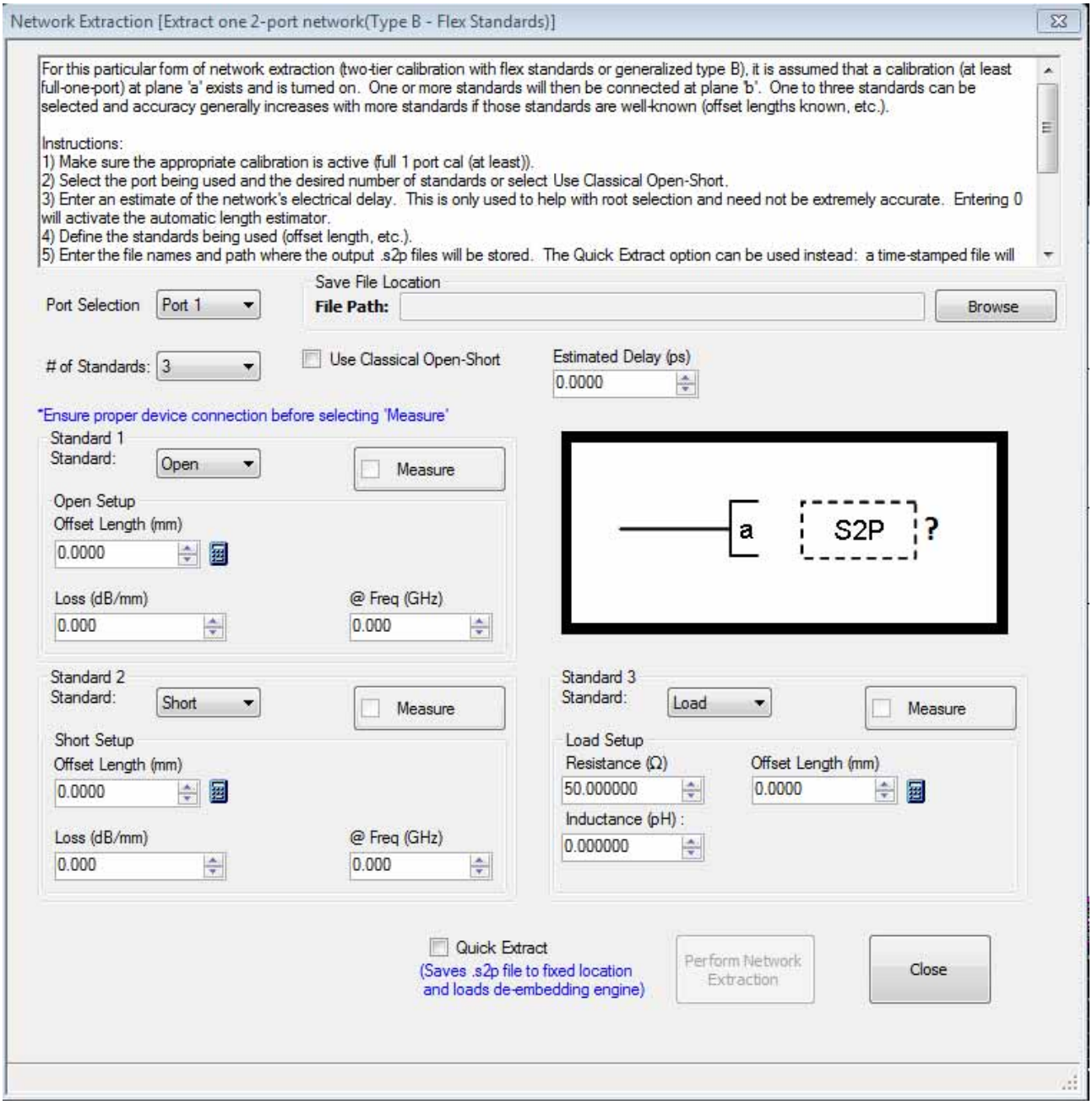


Figure 9-14. NETWORK EXTRACTION Dialog Box—Extract One 2-Port Network—Type B—Flex Standards Variant

When using one standard, only the insertion loss and phase of the fixture arm will be solved for and the match will be assumed to be perfect. If the fixture is well-matched anyway (e.g. the return loss is >15 dB and no deep return loss DUTs need to be measured), this can be a good choice. If only one defined standard is available (e.g., only an open-ended fixture is available), this may be the only choice. Obviously there will be some added uncertainty as fixture mismatch increases. Almost always a high-reflect standard (e.g., open or short) is used for this method.

When using two standards, the mismatch of the fixture arm is assumed to be symmetric but not perfect. If the fixture loss is low, this can be a reasonable approach and is particularly popular in on-wafer applications (also termed an open-short de-embedding method). Most commonly, two high reflect standards are used for this method but a higher return loss device can be substituted to tilt the uncertainty picture more in favor of low reflection DUTs.

The three standards case is very similar to the 'full standards' variant but allows additional flexibility on what those standards are. Examples could include multiple offset open standards used with shorts or loads, a variety of .s1p defined standards, etc.

Four types of standards are available:

Open: Defined only by an offset length relative to the desired reference plane. Loss of the open can be entered that scales with offset length and frequency (in a sqrt(f) sense relative to the entered reference frequency; if a zero reference frequency is entered, the loss will be assumed constant with frequency).

Short: Defined only by an offset length relative to the desired reference plane. Loss of the short can be entered that scales with offset length and frequency (in a sqrt(f) sense relative to the entered reference frequency; if a zero reference frequency is entered, the loss will be assumed constant with frequency).

Load: Defined by a resistance, an inductance and an offset length relative to the desired reference plane. The resistance and inductance are in series as shown in the diagram below.

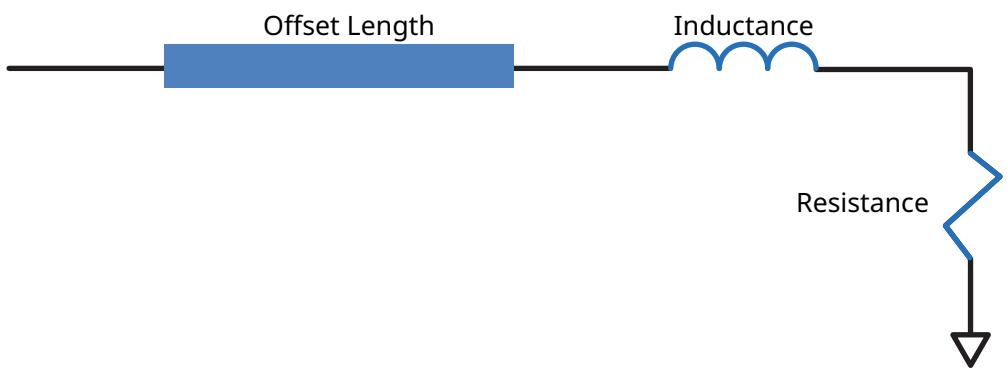


Figure 9-15. Flex Standards Load Model

.s1p-Defined: Defined by an S-parameter file for the standard supplied by the user. The values will be interpolated to match the current frequency list. If the file frequency list and the current list are incompatible (too small an amount of overlap), an error will be generated.

Considerable flexibility is given to the user on selecting the standards to employ but not all combinations will work. On 2 and 3 standards cases, the selections must produce different reflection coefficients at every frequency in the sweep range or singularities will occur. Checking is only done by the system to see if two identical standards are entered but even different entries can produce the same reflection coefficient. An example is a 2 standard, short open case where the short and open have different offset lengths. If, for example, the short offset length is 1mm and the open offset length is 2mm, the two standards produce the same reflection coefficient at 75 GHz. For lower frequency sweeps, this combination will work. The multiple offset short (or open) case also follows these rules as is evident from earlier chapters in this guide on SSST and SSLT calibrations.

The use of the load standard raises somewhat more complicated issues and is often only used in the 3-standards case. In the one standard case, it will raise uncertainties in the fixture insertion loss estimate if the load reflection coefficient is even close to as small as the mismatch of the fixture. In the two-standard case, similar issues happen except both the fixture match and insertion loss extractions can be imperiled. If the fixture is well-matched, the use of a load standard can improve the extraction of that mismatch.

.s1p files cover the gamut of the above possibilities but concept of independence still holds. The reflection coefficients of the various standards (in a complex sense) should be as far apart as possible for maximum accuracy.

The 'Classical open-short' selection is a special case of the 2-standards scenario where a zero offset open and short are used and the fixture arm is assumed to be electrically short. This assumption, most often used in on-wafer de-embedding scenarios, enables more detailed match information about the fixture/lead-in to be determined at the expense of generality.

In the space of this chapter, it would be difficult to cover the sensitivities for all possible combinations of fixture parameters and standards. We can, however, offer some examples that might illustrate some of the trends.

Consider the one standard case first where the fixture has 25 dB return loss (flat with frequency) and has 5 dB of insertion loss at 40 GHz (going to 0 dB at DC with a square-root-of-f dependence). In the first experiment, an open will be used and a Monte Carlo simulation run where the reflection coefficient magnitude (which should be 1) is allowed to vary +/–5% and the phase is allowed to vary +/–10 degrees. The errors on extraction look like that below. The width of the apparent trace gives the sensitivity to the standard changes which is quite slight. The macroscopic curve is the error for ignoring the fixture match (by using 1 standard only). Clearly the standard sensitivity is not a huge problem. Since the loss of the fixture gets closer to the mismatch at higher frequencies, the error increases.

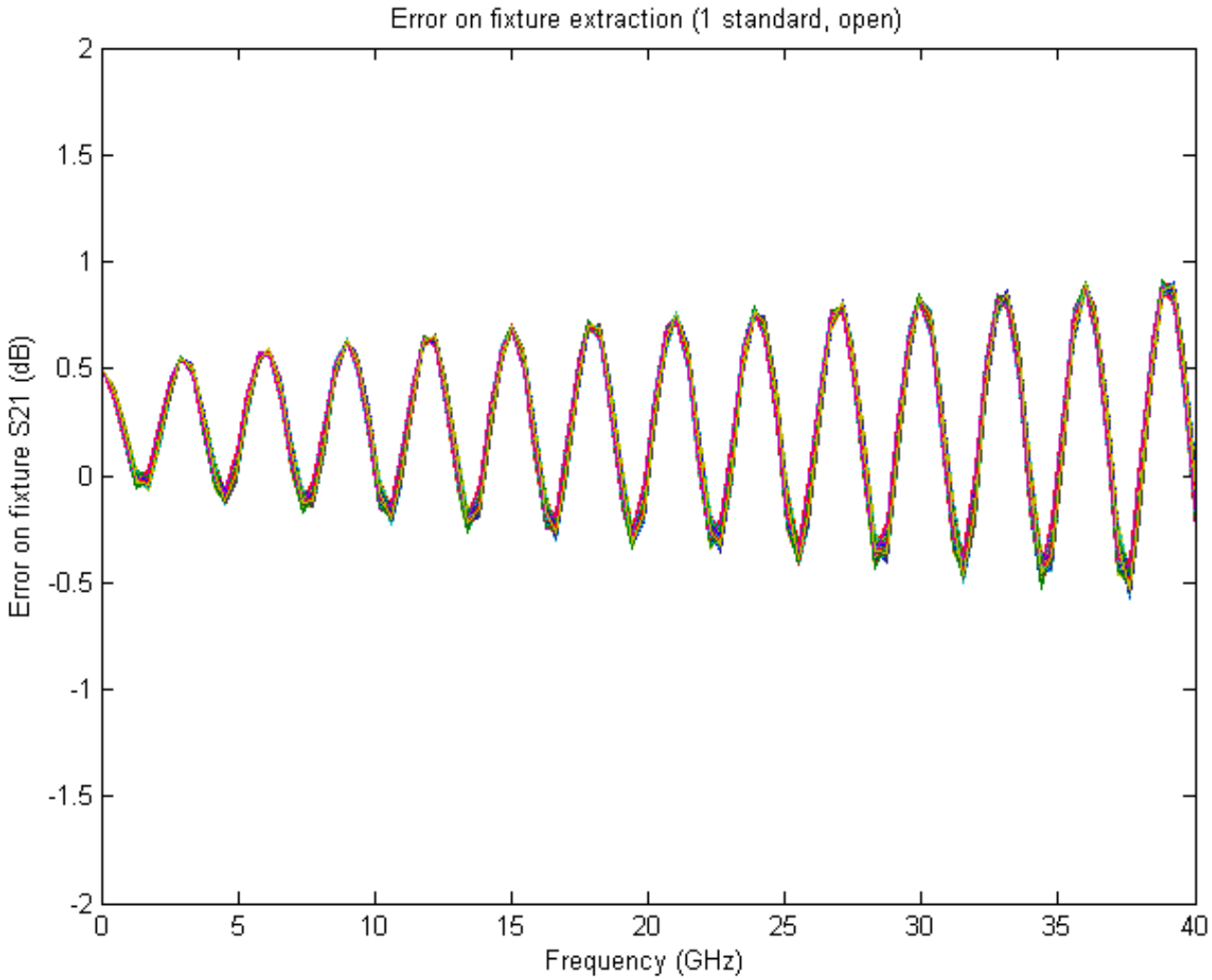


Figure 9-16. Error On Fixture Extraction (1 Standard, Open)

Now consider the case where the standard is a 20 dB return loss device and we run the same simulation. The standards sensitivity has increased in this case (width of the composite trace) although the scaling of the magnitude variations used may not have been comparable. The overall errors increased since the measured insertion loss contribution to the reflection measurement is now even smaller relative to the (ignored) fixture mismatch.

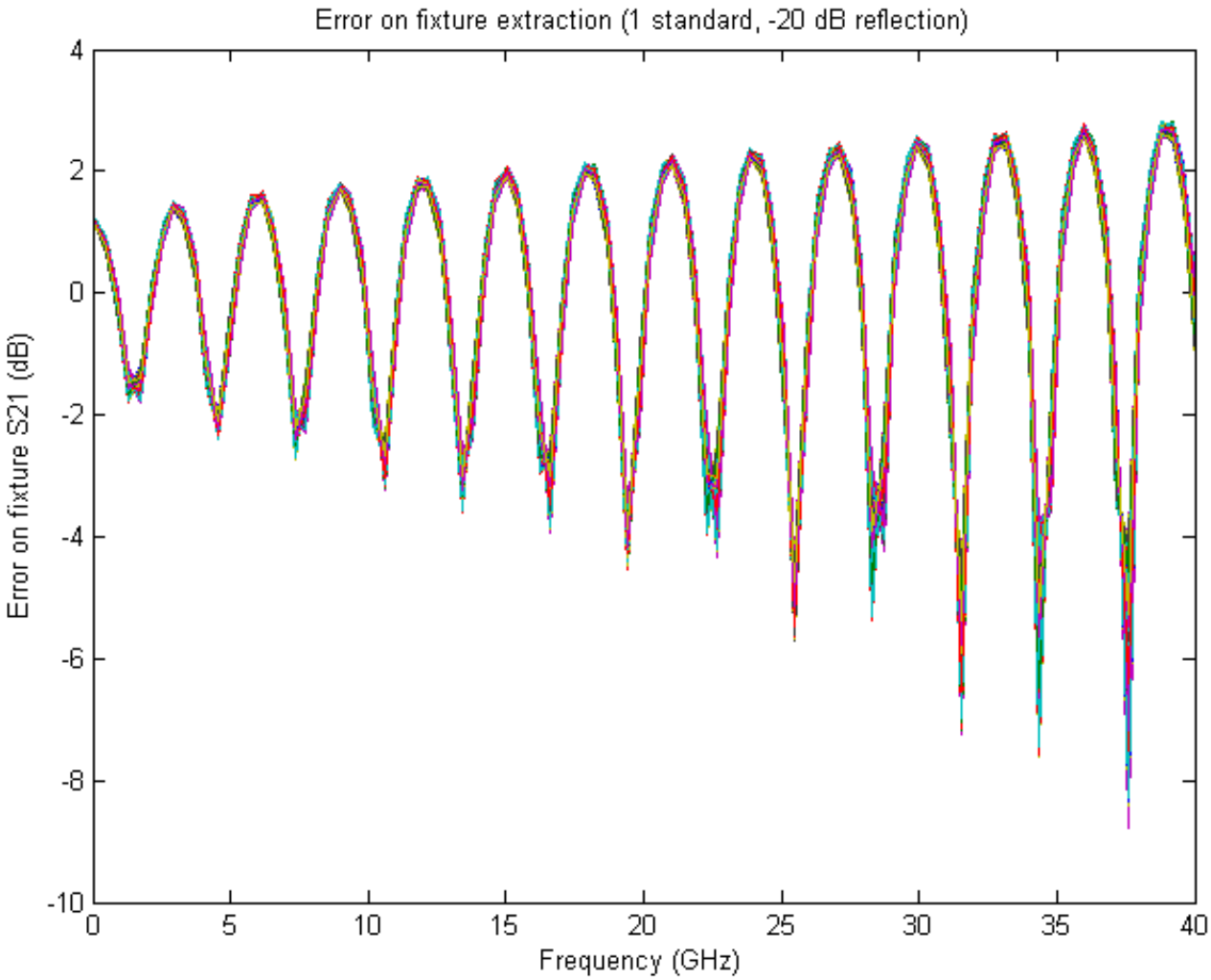


Figure 9-17. Error On Fixture Extraction (1 Standard, –20 dB Reflection)

Moving to the two standard case with roughly the same fixture (except now $|S_{11}| = -25$ dB and $|S_{22}| = -20$ dB). First, an open and a short were used with the same 5%, 10 degree parameter variation. The resulting errors are now much smaller since some mismatch is being accounted for. The standards sensitivity is now a much larger fraction of the error. The errors are also now larger at lower frequencies since the lower loss exposes the mismatch asymmetry of the fixture more to the measurement.

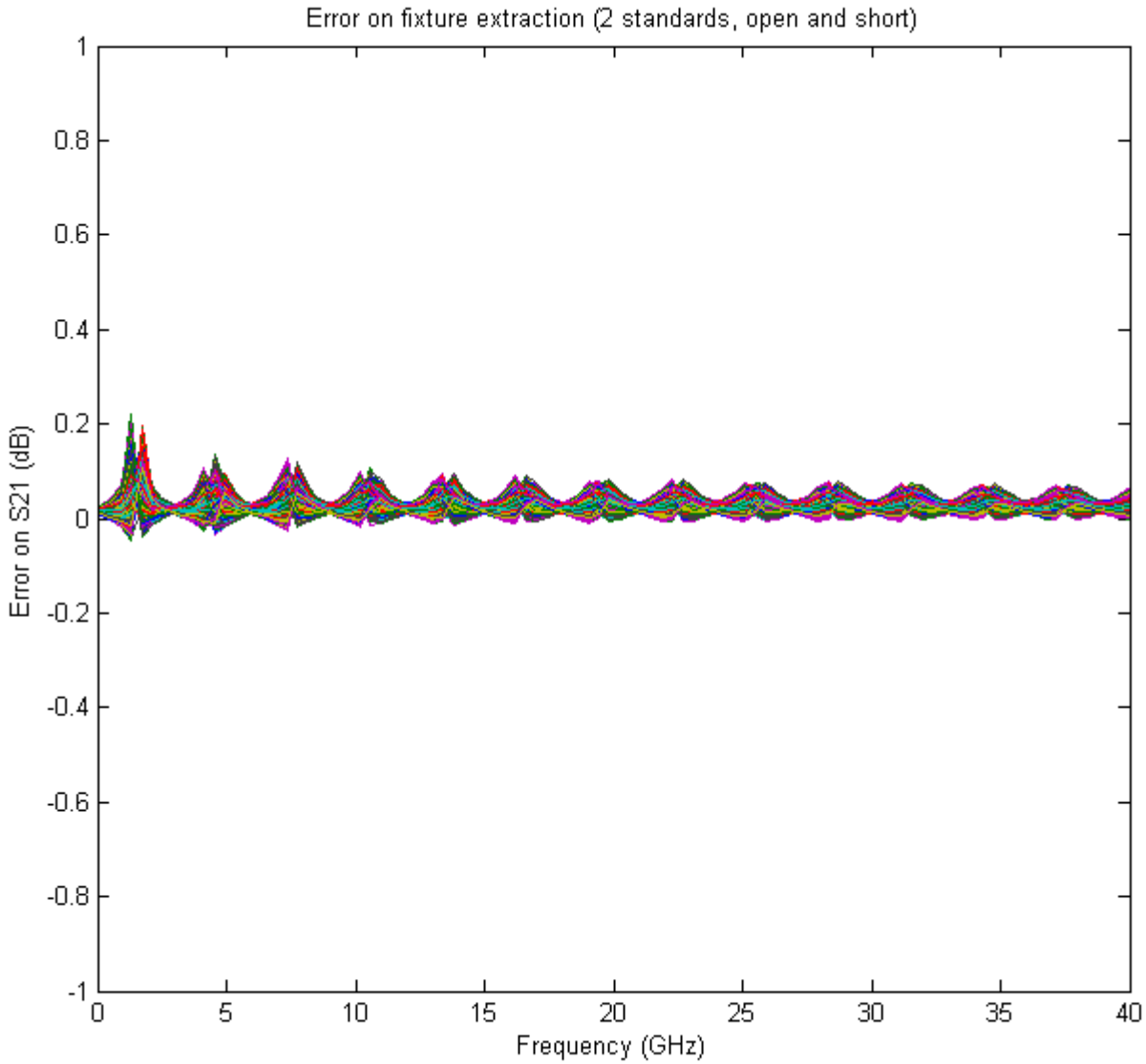


Figure 9-18. Error On Fixture Extraction (2 Standards, Open and Short)

If the 20 dB return loss device is substituted for the short, the profile changes dramatically. Since one of the standards is similar to the fixture match, the deconvolution of loss and mismatch extraction in this method breaks down somewhat so the error increases (although it is still much better than with one standard).

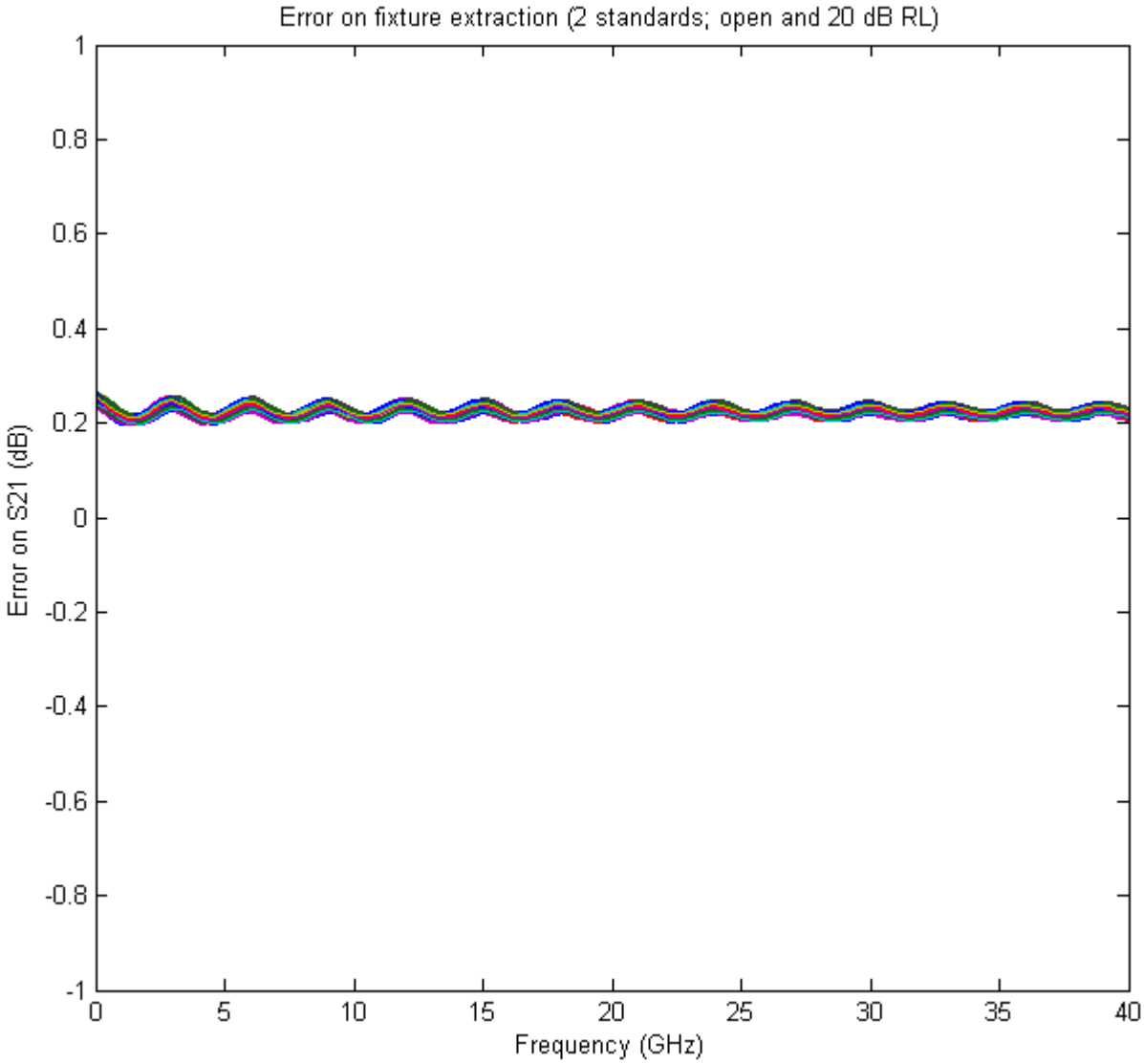


Figure 9-19. Error On Fixture Extraction (2 Standards, Open and 20 dB RL)

In the three-standard case, there are an even larger number of variations but, since the input and output match are being solved for independently, the sensitivity and errors will decrease as long as the standards level-of-knowledge remains constant. One example comparing one, two and three standards is shown below. The mismatch of this fixture is relative symmetric and below –20 dB until about 10 GHz and degrades to about –10 dB by 20 GHz (and becomes less symmetric). One can see the one-standard approach starting to diverge relatively early as the fixture mismatch becomes increasingly significant relative to the insertion loss. The two-standard approach starts deviating later only when the match asymmetry becomes more significant.

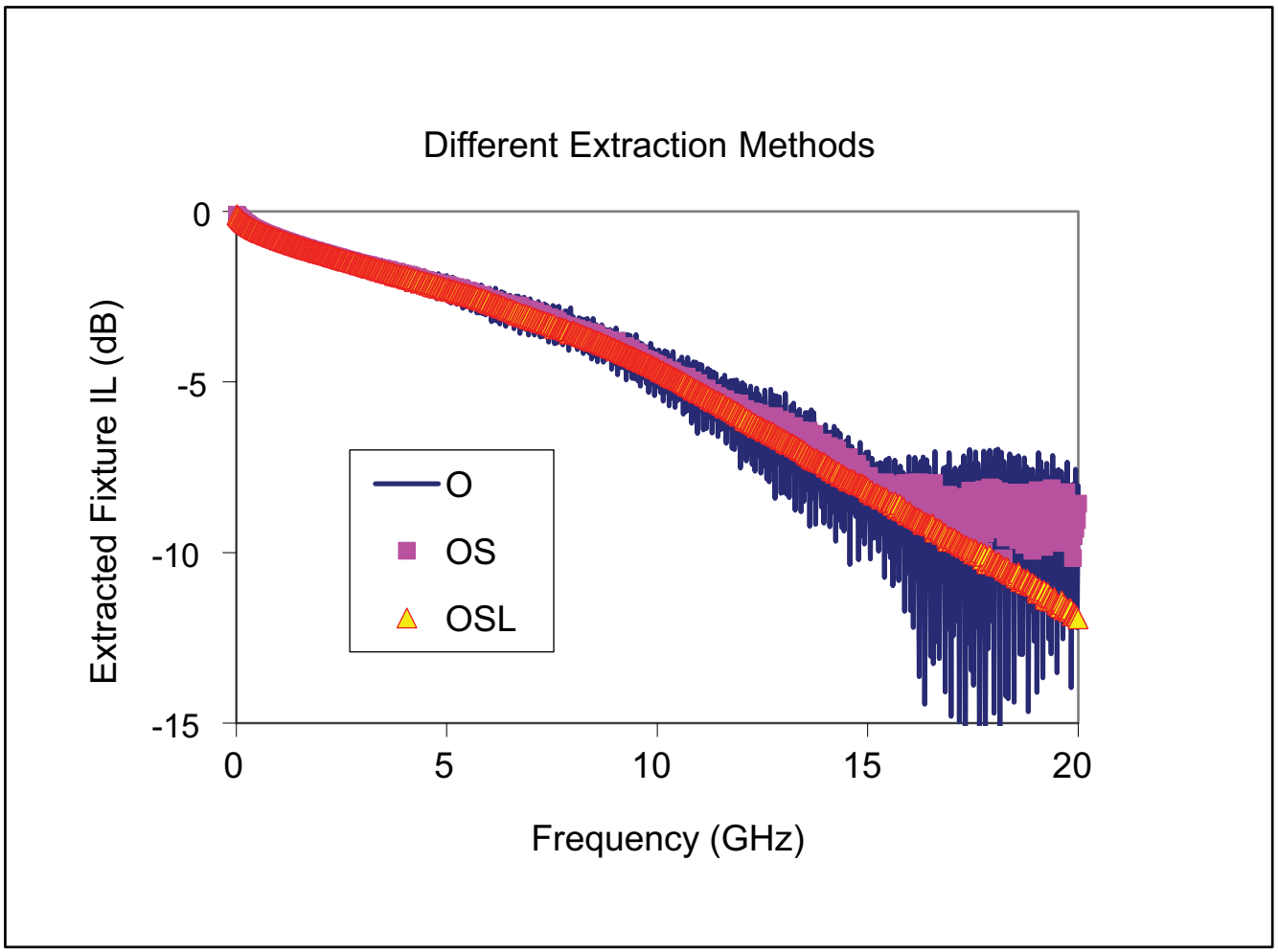


Figure 9-20. Extracted Fixture Example Comparing One, Two, and Three Standards

As a summary of all of the Type B permutations:

- A full standards version (or flex standards with 3 independent standards) can produce the best possible accuracy and lowest sensitivity but only if the standards are well-known. To put it another way, the 'performance ceiling' is the highest.
- Flex standards with two independent standards can do very well if the fixture arm is relatively symmetric and does particularly well when the insertion loss is low. An open-short pairing is the most common and is a favorite in on-wafer and micro-fixture applications. This approach fares less well with higher loss, asymmetric fixture arms.
- Flex standards with one standard is sometimes the only practical method due to the interface on the far end of the fixture. A high reflection is the best choice for that standard usually unless the fixture is exceptionally well-matched. This approach is most accurate when the fixture mismatch is very low.

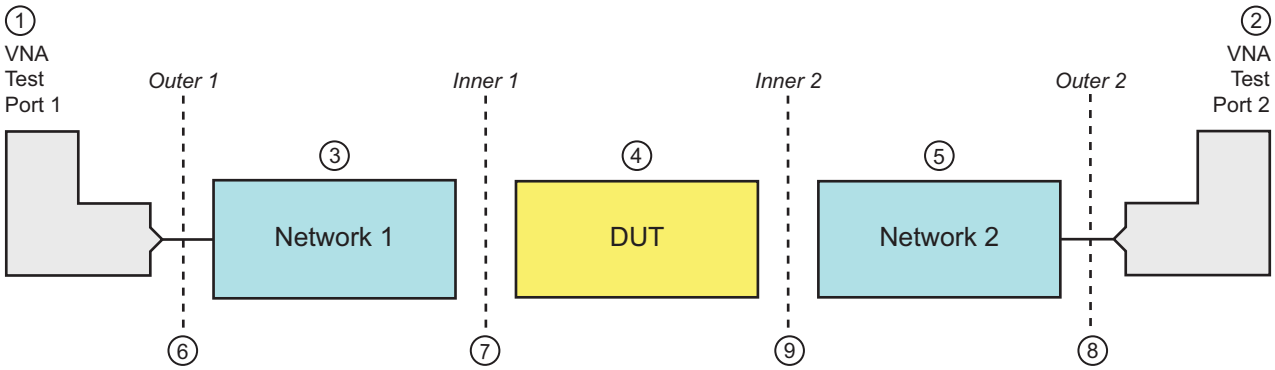
Quick Extract

The Quick Extract check box disables the file entry fields and instead saves the output file to a pre-determined location and automatically starts the de-embedding engine. The file just saved will automatically be loaded into the de-embedder (where it can be edited). This process can help save time if the desire is to immediately de-embed a fixture that was just extracted. Note that any de-embedding in place prior to the extraction will be cleared (and the system will warn if this is about to happen). If de-embedding was on when extraction was run, those de-embedded values will be used during extraction so some caution is advised as it is possible to partially negate an extraction by using already partially-de-embedded data.

Type C Network Extraction

Type C is the most complete, dual fixture extraction approach offered in the VNA. It requires full 2-port calibrations at two sets of reference planes but can fully determine the S-parameters of two networks independently.

Consider the diagram in NETWORK EXTRACTION dialog box (above in Figure 9-9). A calibration is required at the outer reference plane set and the inner reference plane set. The outer calibration can usually be done coaxially (or some other well-defined media) depending on the networks involved. The inner calibration is often more complicated and may be board- or wafer-level (and may require the user create calibration standards). Assuming these calibrations are possible, then the S-parameters of Network 1 and Network 2 can be found.



1. Test Port 1
2. Test Port 2
3. Network 1 attached to Test Port 1
4. DUT attached to Network 1 (on left) and Network 2 (on right)
5. Network 2 attached to Test Port 2
6. Outer Reference Plane 1
7. Inner Reference Plane 1
8. Outer Reference Plane 2
9. Inner Reference Plane 2

Figure 9-21. Network Extraction Process Diagram for Type C Networks

Two-port calibrations at two different reference plane sets are used to extract the S-parameters of the intervening networks (often test fixtures). The dialog for loading the two calibrations is shown in NETWORK EXTRACTION [Extract Two 2-Port Networks (Type C)] dialog box below in Figure 9-22 on page 9-26.

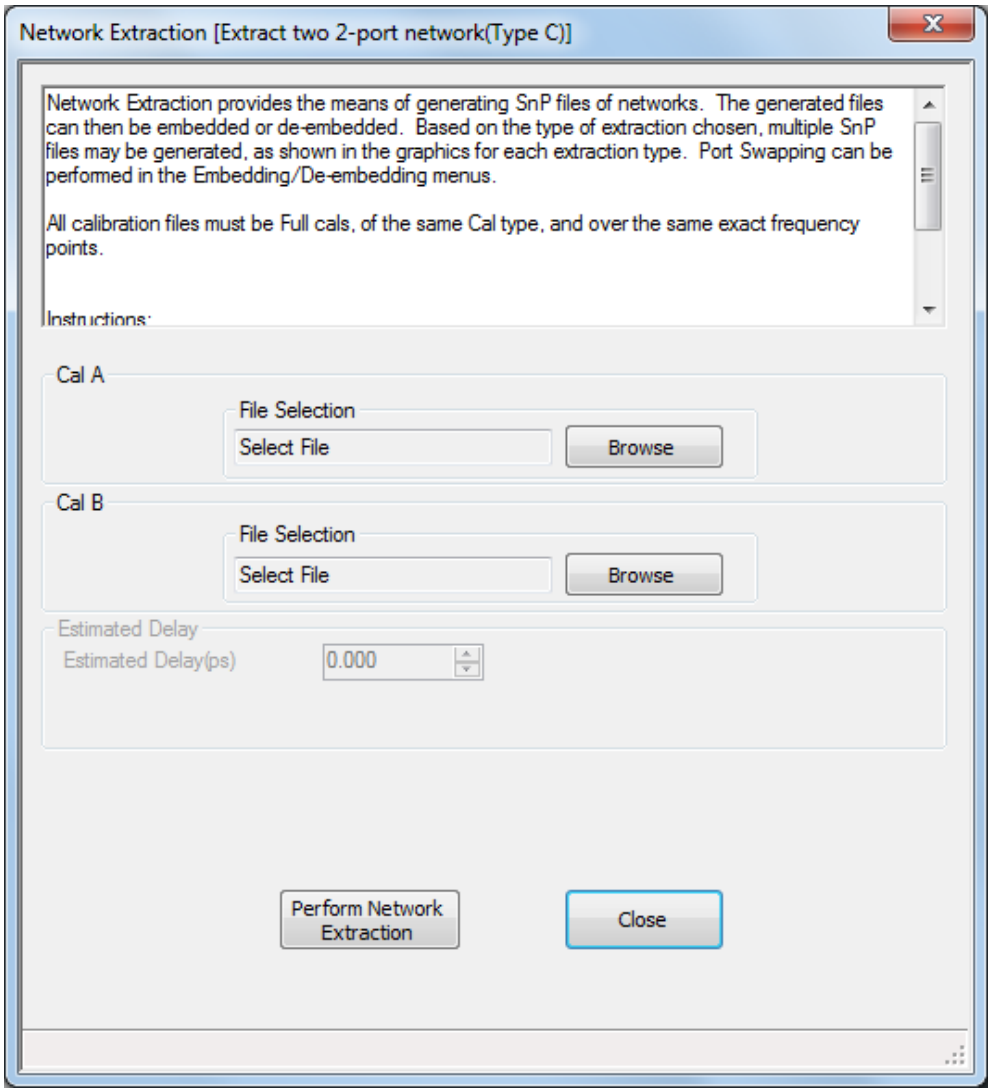


Figure 9-22. NETWORK EXTRACTION Dialog Box—Extract Two 2-Port Networks—Type C

As before, the two calibrations are performed and the setups saved, typically as an active channel CHX file type.

Some conditions:

- The two calibrations must be full 2-port cals and must have the same frequency lists.
- After extraction is performed, a file dialog will appear allowing the user to indicate where the S2P files should be stored.
- The networks are assumed to be reciprocal.

Unlike Types A and B, this method determines the two fixture halves completely and independently. As a trade-off, a complete set of standards at the inner plane are now required. Algorithmically, this type is very similar to Type A except two networks are processed simultaneously. If the inner cal standards can be successfully made/acquired, the inner match values extracted will typically be more stable than those acquired with a Type B analysis for the reasons discussed in the previous section.

Type D Network Extraction

The dialog is a very simple one and is shown in Figure 9-23. The outer calibration should be active when this procedure is called since files are not recalled as with the other techniques. Because it is sometimes difficult to allocate or interpret the match terms, a check box is provided to ignore those terms altogether. In this case, $S_{11} = S_{22} = 0$ (linear) in the exported S2P file. As with the other techniques, a dialog will appear upon execution to allowing the naming of the destination S2P file.

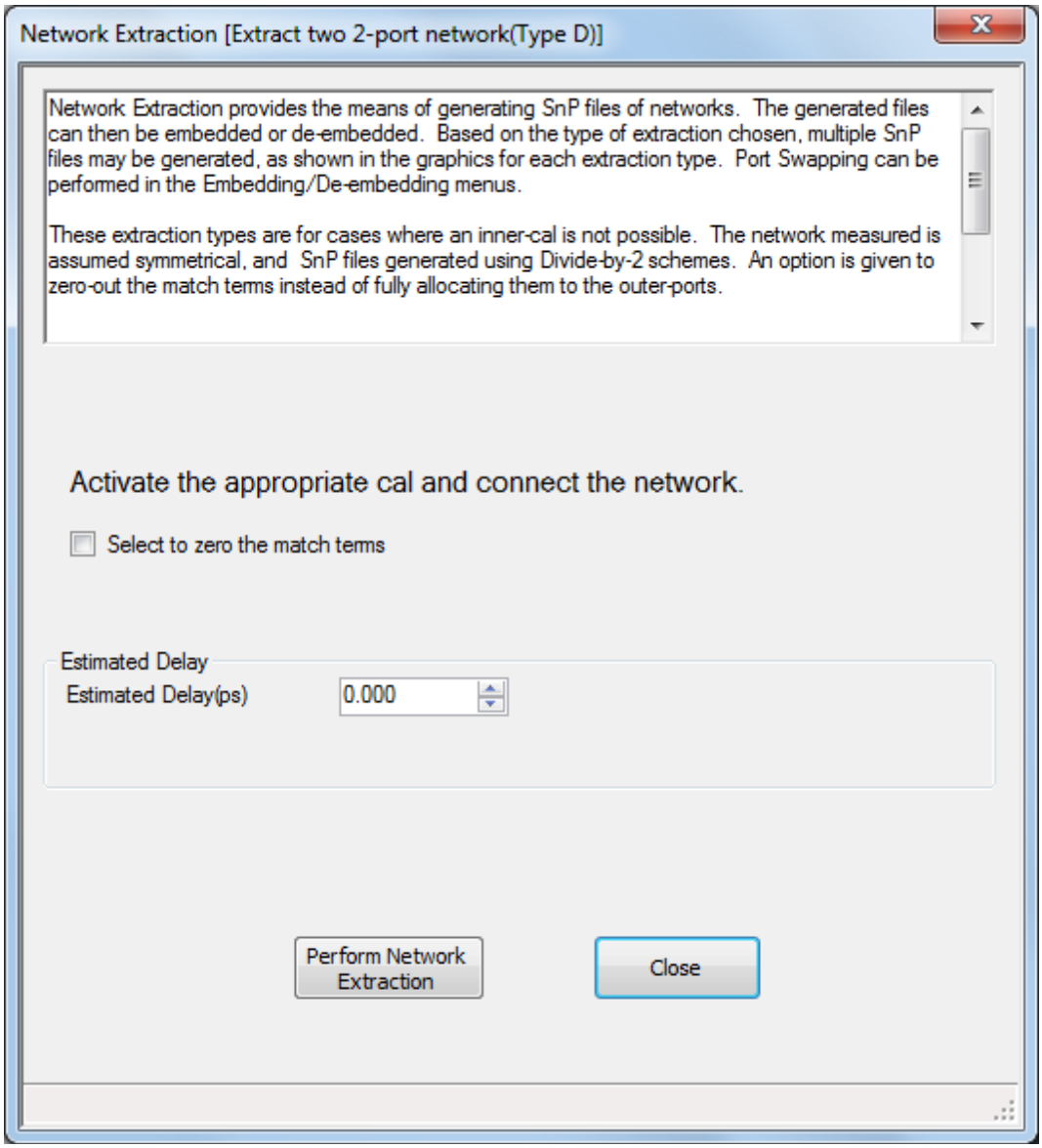


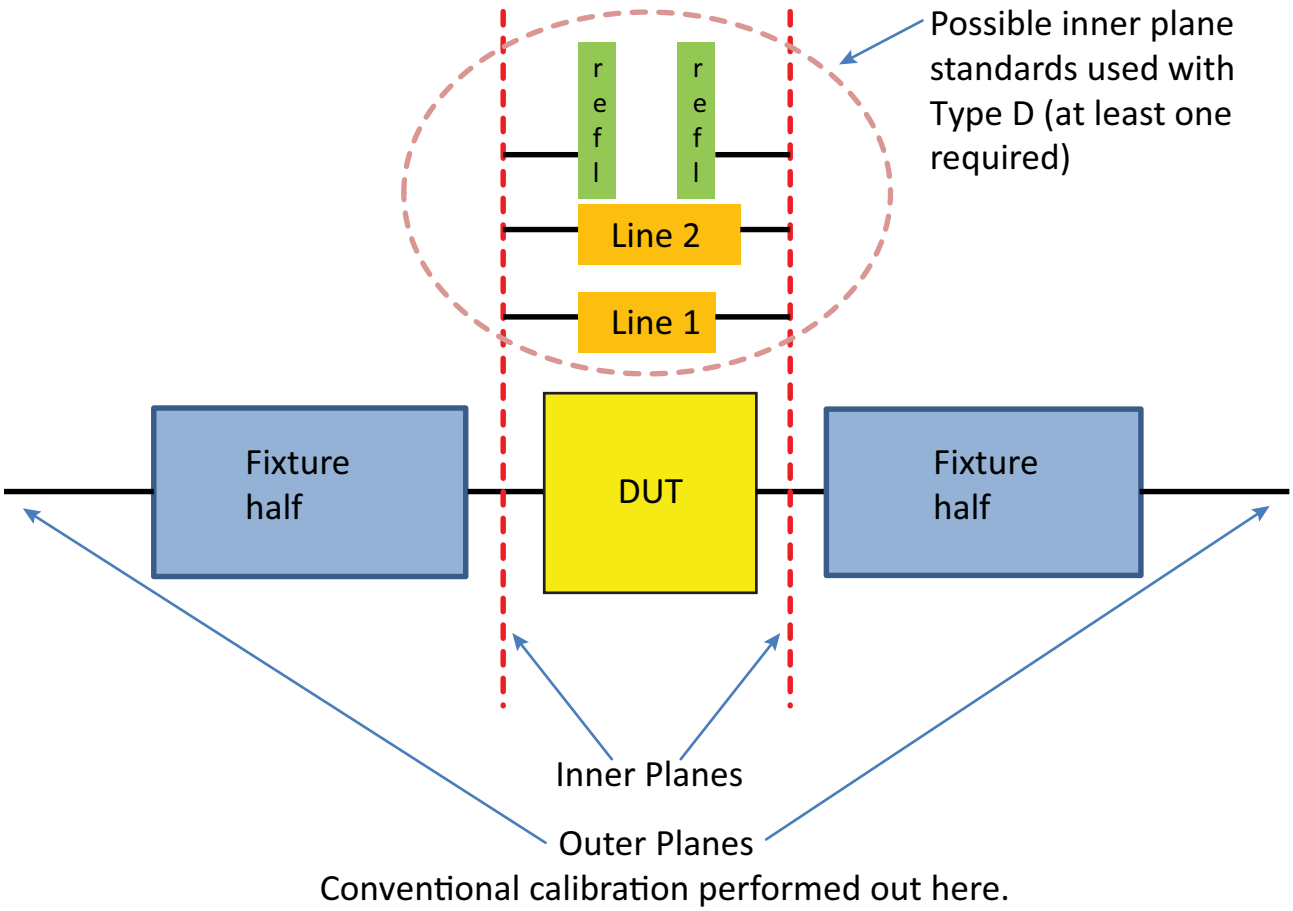
Figure 9-23. NETWORK EXTRACTION Dialog Box—Extract Two 2-Port Networks—Type D (without Option 21)

Type D Network Extraction—Multi Standards (with Option 21)

Type D is considerably different from the other techniques (except some configurations of flexible standards in Type B) in that it relies only on a limited number (1–3) of measurements to extract parameters rather than relying on the manipulation of a pair of calibrations. A full 2-port calibration is performed at the outer planes (often in coax or waveguide or with on-wafer probes) and then simple standards (one or more lines and possibly a single high-reflect standard) are connected between the fixture halves as suggested in Figure 9-24. The non-Option 21 version of Type D differs in that only a single line standard is allowed and its length must be 0.

This technique belongs to a class of approaches that have been termed 'partial information techniques' since they make additional assumptions about the fixture to avoid the necessity of a full calibration at the inner plane. As such, these techniques are particularly attractive when the inner plane has a complex structure or geometry that makes it difficult to create many standards for that plane or difficult to accurately model those standards. There are also cases where such methods are useful because the repeatability of connection at the inner plane is degraded. By de-emphasizing inner plane match in those cases, sensitivity to repeatability issues can be reduced.

There are a number of different ways to use Type D and this section will explore the differences and how one might choose the sub-approach to take. In earlier versions of the VectorStar software (prior to 2017), only one choice was available: the use of a zero-length thru between back-to-back fixture halves which assumed that the halves were identical and had perfect match at the inner plane. That sub-approach is still available, as will be seen, but there is now more flexibility.



The basic structure of Type D extraction is shown here. There is considerable choice in the standards used at the inner plane, but the combinations all share the fact that the set is not 'complete' in the sense of a full calibration at that plane. Some fixed error is accepted in exchange for simpler standards and more immunity to repeatability issues.

Figure 9-24. Network Extraction Type D

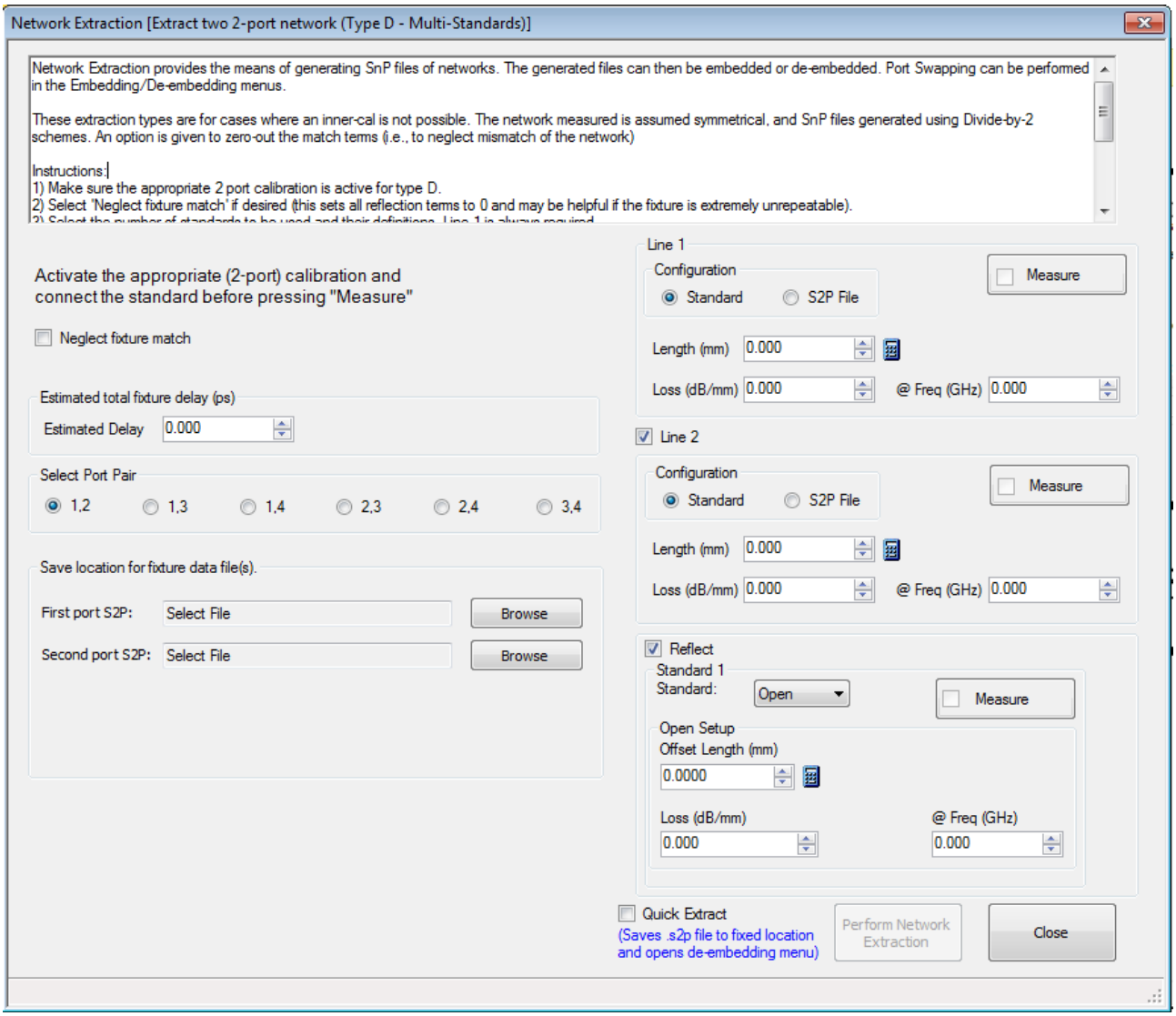


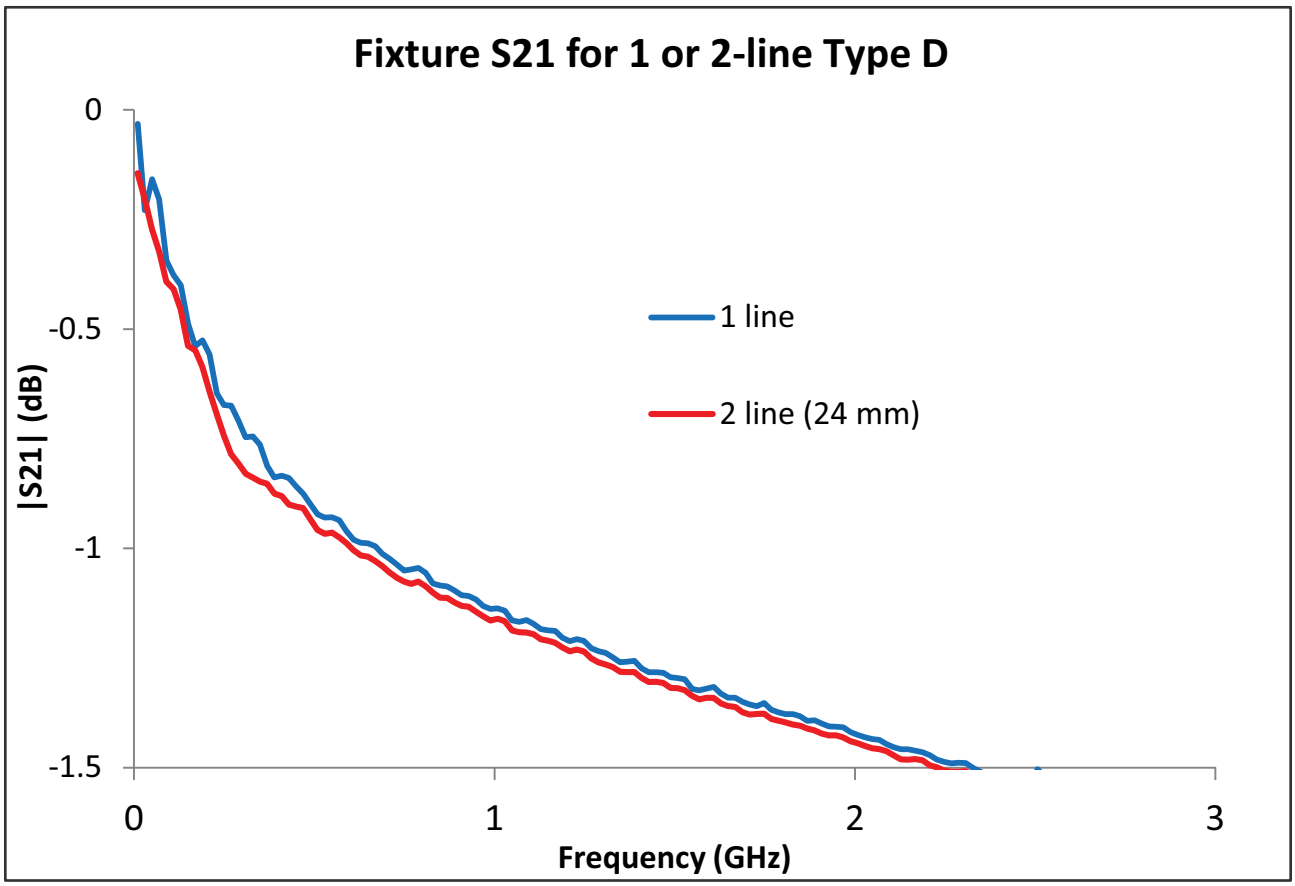
Figure 9-25. NETWORK EXTRACTION Dialog Box Type D Multi-Standards (With Option 21)

The basic Type D extraction starts with a line between fixture halves as mentioned and will assume symmetry of the fixture halves. This line can have any length but that length must be specified and any errors in that specification will map through to the phase lengths of the extracted fixtures. If one stops here, the mismatch at the inner planes of the fixture will be ignored (S_{22} will be zero in the extracted files). One can also elect to set all match terms to zero and that will force both S_{11} and S_{22} to zero no matter how many standards are used. This zero-match choice can be useful if repeatability at the inner plane is particularly poor and insertion loss/phase correction for the fixture is the primary concern (doing a closer-to-full match correction with a very non-repeatable interface can often further reduce the transmission extraction accuracy). A fixture length entry is requested (represents both halves together) and this is used just for root selection so precision is not normally required. If zero is entered for the fixture length, an automatic routine (similar to auto reference plane extension discussed in Chapter 11 of this guide) is used to estimate the length.

One can also add a second line of some different length (and its transmission amplitude can be entered independently) and inner plane match will no longer be ignored. Note, however, that the accuracy of the entered line lengths is more important in this case. Also, the line length difference between the first and second lines should not approach 180 degrees within the frequency range of interest (or be too close to 0). Generally, the line length difference should be between ~10 and 160 degrees over the frequency range of concern.

The use of reflection standards (which must be placed on the inner planes of both fixture halves) will also allow for solving for inner plane match. Finally, one can use all three standards which will generally improve accuracy on both insertion loss and inner plane match. The choice on how many of the standards to use should depend on how well those standards can be made (e.g., can a second line length be made that is still relatively well-matched as a transmission line, can reflect standards be made that have relatively uniform reflection magnitude over the frequency range of interest, etc.). Implicit in this is that the measured characteristics going from the first standard to the second standard do not change for other reasons (e.g., if the structures being measured are different implementations of the same fixture, then they must be quite identical). Note that while fixture symmetry is implicit for all of type D, the sensitivity to asymmetries is heightened with the use of the reflect standard. Generally, if additional standards perform well in this sense, using them will improve the overall extraction.

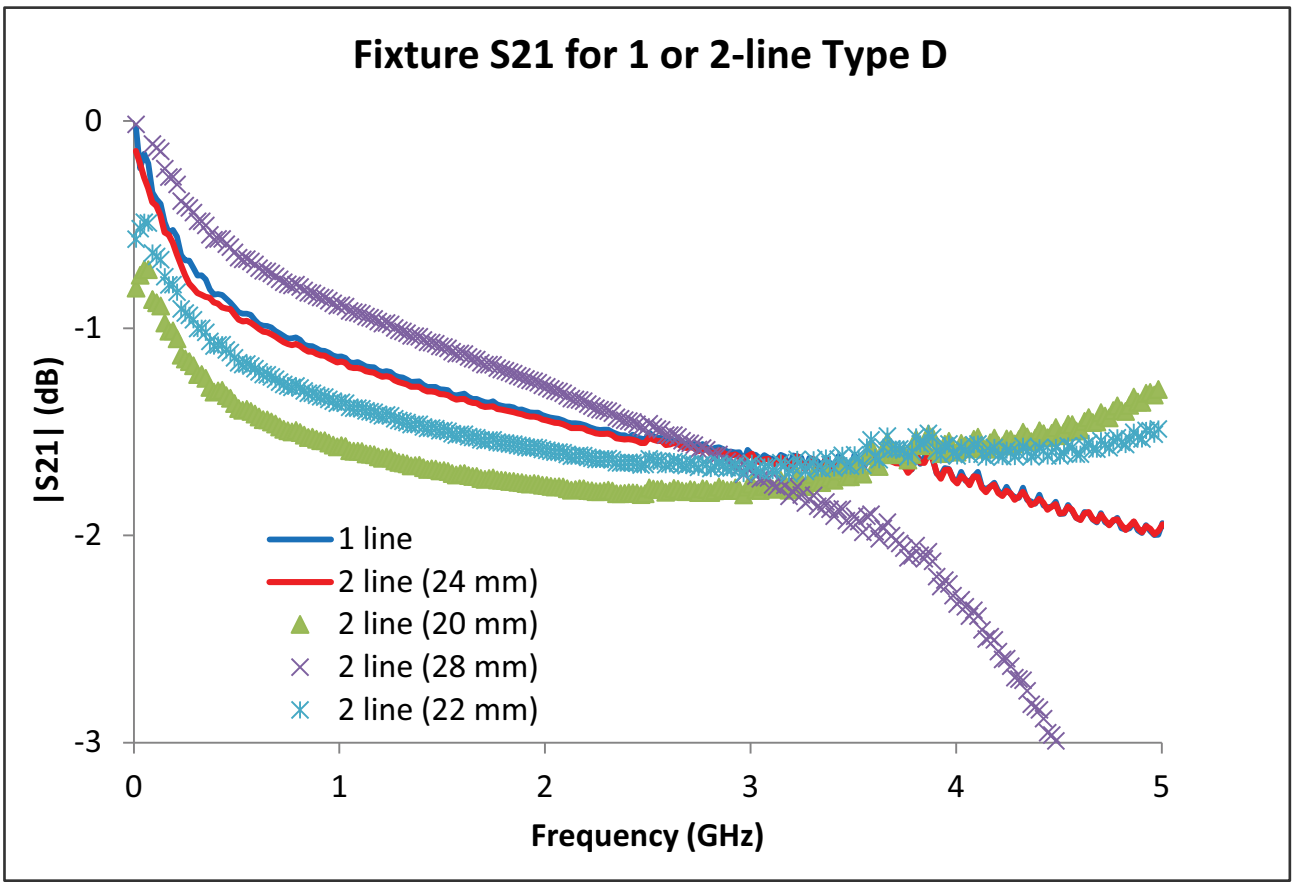
As an example, consider the measurement of a back-to-back cable assembly where two different thru lengths are possible at the inner plane (0 and 24 mm). If one compares the single line approach to the two line approach on extracted insertion loss, one can see some differences (Figure 9-26). The single line approach produces an insertion loss with slightly more ripple and a slightly more optimistic overall value (although errors in either direction are possible).



A comparison of single and double line Type D extractions of insertion loss are shown here. With a correct standard length entry and sufficient repeatability, the double-line method can increase accuracy of the extraction.

Figure 9-26. Comparison of Single and Double Line Type D Extractions of Insertion Loss

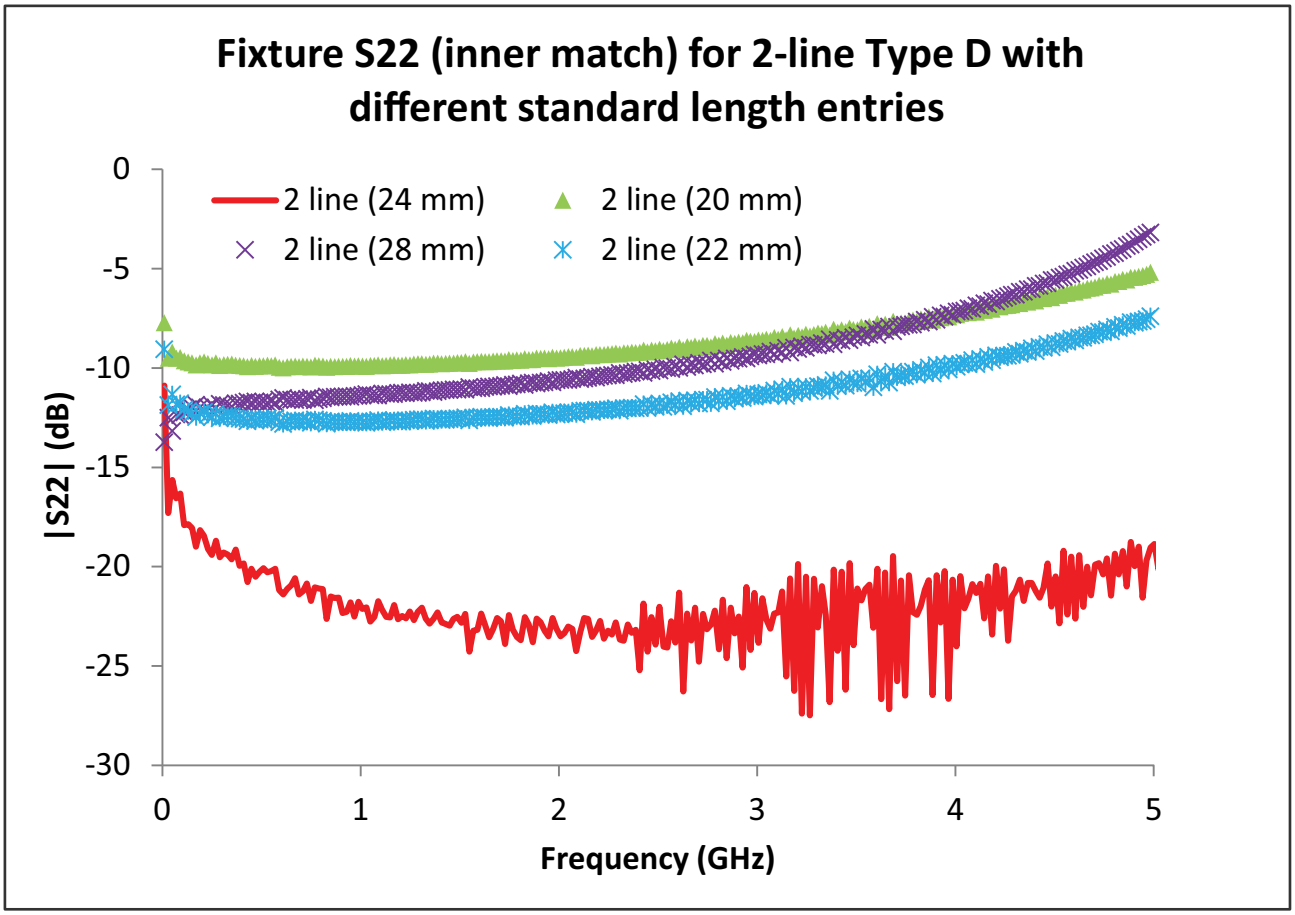
If the entry of the standard line length is not accurate, however, substantial errors can result. The extraction of the previous figure is repeated in Figure 9-27 for the added cases when the 2nd line length is off by ~10 or 20 %. The errors are a few tenths of a dB at low frequency but grow larger at high frequencies as the standing wave that is being corrected grows more dense. Also, in this case, the 28 mm entry brings a singularity to lower frequencies and this has an even more substantial effect on the error.



The results of Figure 9-26 are augmented here with two line cases when the entered length of the second line is incorrect (correct value is 24 mm).

Figure 9-27. Resulting Error with Inaccurate Standard Line Length Entry

One reason for using the two line approach (or line+reflect) is to get more reasonable values for inner plane match which can be important for sequential de-embedding and modeling. Again, the parameter entry accuracy is important for the inner plane match as it was for insertion loss extraction. The inner plane match values for the fixture of Figure 9-27 are shown in Figure 9-28 for the same length entries.



The inner plane match values for the experiment of Figure 9-27 are shown here. Even a 10% length entry error causes ~10 dB errors in return loss.

Figure 9-28. Inner Plane Match Values

The single-line version of Type D is best obviously for well-matched fixtures relative to loss (i.e., very well matched if low loss and moderately well-matched for moderate loss). With a 20 dB return loss, there will be generally 0.3 dB or more of insertion loss uncertainty with this technique (as opposed to ~<0.1 dB with other techniques if good standards are available). With a 10 dB return loss fixture, it will be several dB of uncertainty. With additional standards (assuming accuracy of length entries and sufficient repeatability), the method becomes more mismatch tolerant—often keeping errors under 1 dB for a 10 dB return loss fixture—but the results will still be worse in an absolute sense than with a complete method (assuming the latter was possible).

Additional Notes Regarding Type D—Multi-standards

- A full 2-port calibration must be active and the extraction will be run over that frequency range. For 4-port systems, at least a full 2-port calibration must be active (more details on the 4-port cases are covered in Chapter 24 of this measurement guide).
- The line and reflect offset lengths are entered in millimeters, although a calculator is available if values are in picoseconds. If the material type is set up (from the current calibration or manually thereafter), that and any active dispersion relations will be used in the calculations. In two-line cases, the first and second lines must be electrically distinct (i.e., phase length differences not too close to 0 or 180 degrees within the frequency range of interest).
- -The Quick Extract feature (allowed on all type D variations with Option 21) suppresses the file name entry fields and instead saves the files to a predetermined location, automatically opens the de-embedding engine, and loads the just-generated files into the de-embedder. These entries can be changed at that time if desired, but this process can save time if the desire is to immediately de-embed a fixture that was just extracted.
- The Measure buttons are used to measure the two fixture halves with the appropriate standard connected and those standards must be connected when the relevant Measure button is pressed. The system will trigger a channel sweep at that time to acquire new data.
- Loss of the line and reflect standards can be entered that scales with length and frequency (in a sqrt(f) sense relative to the entered reference frequency; if a zero reference frequency is entered, the loss will be assumed constant with frequency).

Type D Network Extraction—Phase Localized (with Option 21)

Another variation of Type D is termed 'Phase Localized' where a single standard (either a line or a reflect/reflect pair, or some combination) is used, along with the assumption that the fixture is electrically long enough (based on the frequency range being used) and the bulk of the fixture mismatch is not too close to the inner plane. The dialog for setting up phase localized extraction is shown in Figure 9-30.

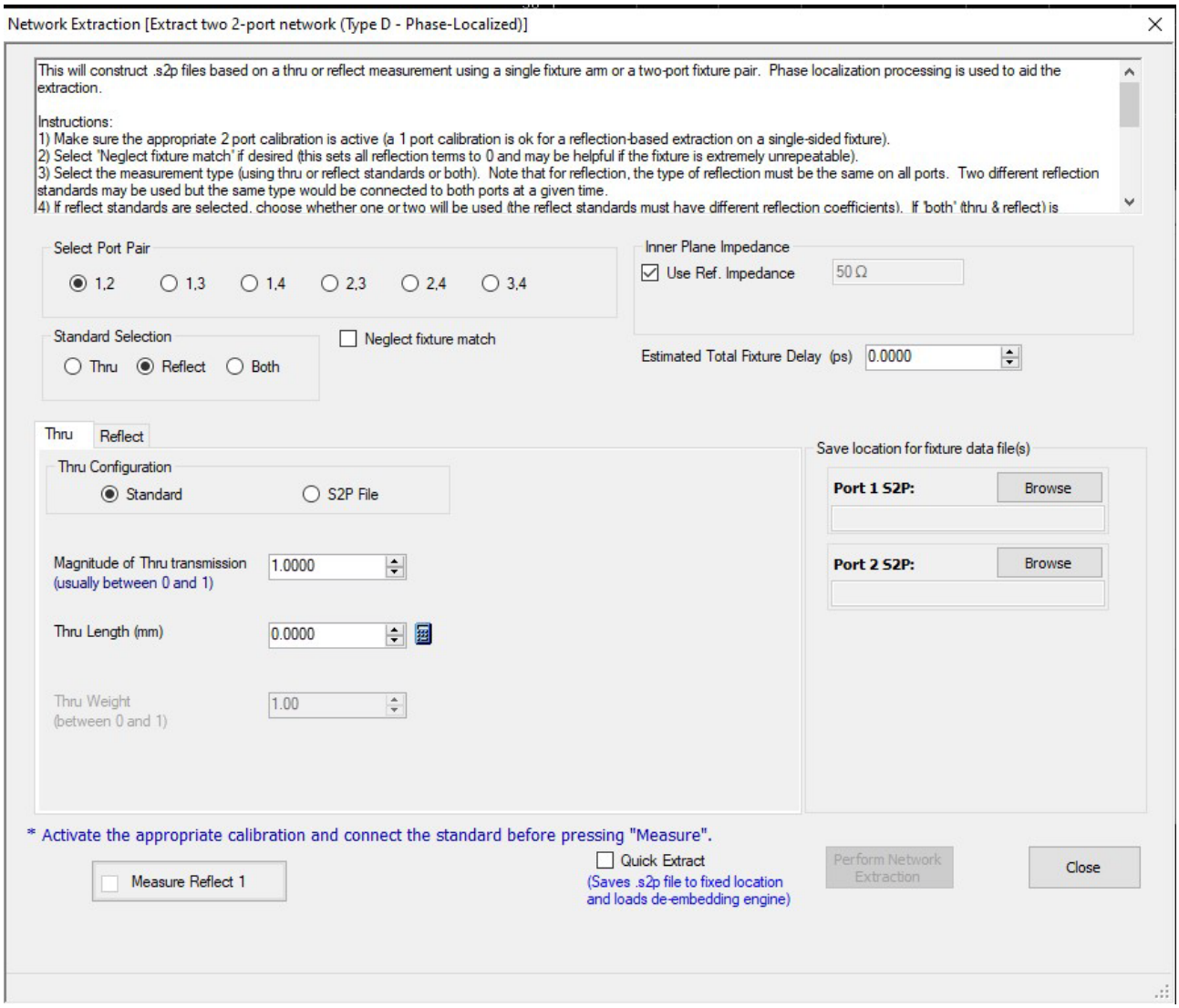


Figure 9-29. NETWORK EXTRACTION Dialog—Type D—Phase Localized (With Option 21) – Thru Tab

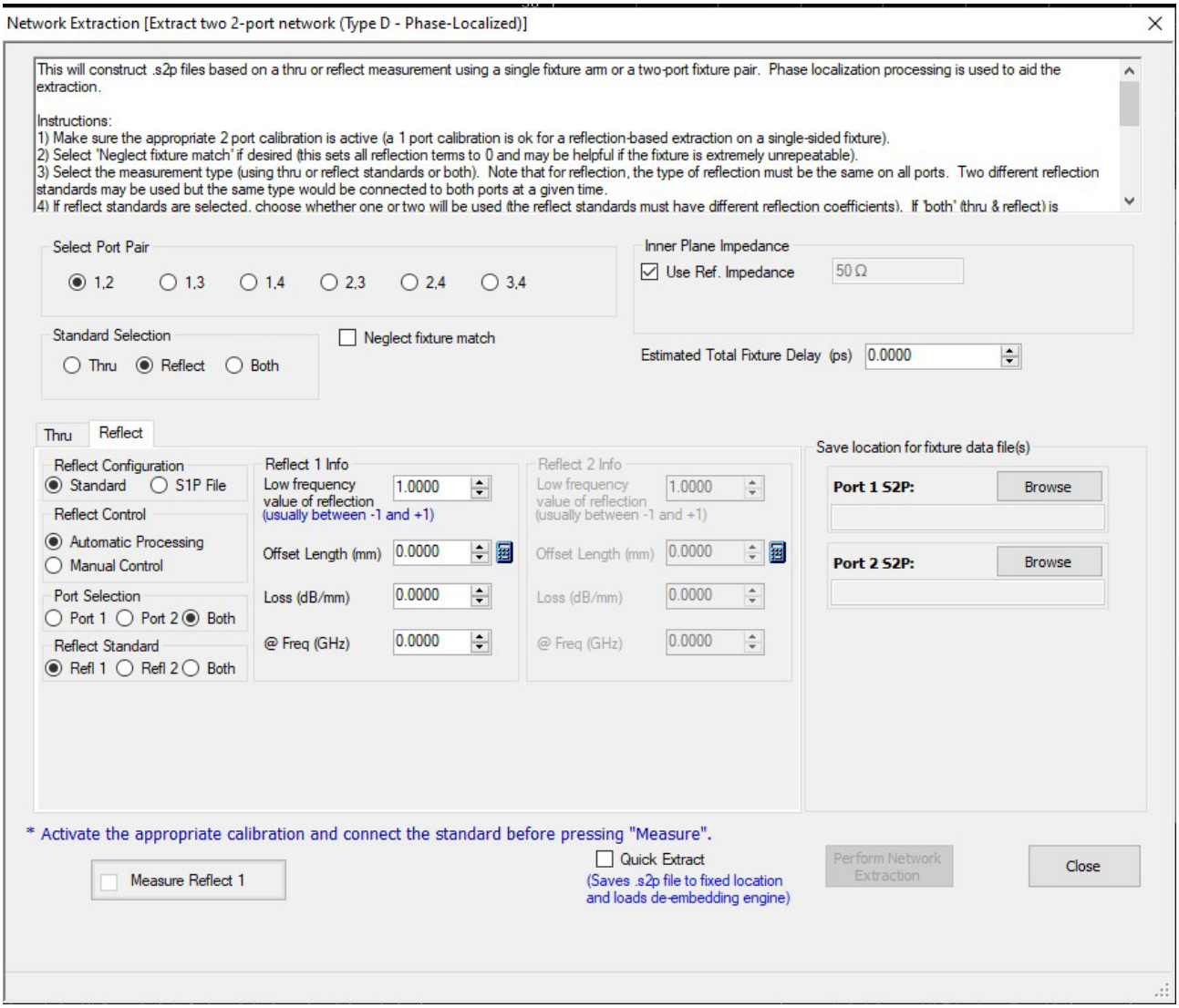


Figure 9-30. NETWORK EXTRACTION Dialog—Type D—Phase Localized (With Option 21) – Reflect Tab

If the assumptions are met, this method can outperform the previously discussed Type D variations. More central to this method is the fixture length as transmission-line-like: functions are cross-correlated with the measured data to better isolate insertion loss and reflection coefficients of the fixture halves. If the fixture length entry is set to zero, an automatic process (much like auto reference plane delay discussed in Chapter 11 of this guide) will estimate the length. As before, entries for the line length or reflect offset length are required and any errors in those values will translate to extracted parameter phase. If the line is chosen as the standard, symmetry between the fixture halves in terms of insertion loss will be assumed. If the reflect standard is chosen, no symmetry is assumed and only one half of the fixture can be extracted if desired. If both halves are to be extracted, length estimates for the individual arms can be entered. In this variation of Type D, there is no Measure button and the measurement is executed when Perform Network Extraction is selected.

Finite transmission of the thru standard (or .s2p definition) can be entered as can loss of the reflect standard. As with some other network extraction methods, the reflect loss is entered in dB/mm terms and the value will scale with frequency (as sqrt(f/fref); if fref=0, the loss will be treated as constant with frequency).

As an example, consider a fixture consisting of a ~50 mm microstrip line, a coaxial launcher on one end and a DUT-local launcher on the other end (for each arm of the composite fixture). Suppose only an open standard is available and one would like to use the phase-localized approach since there was only the one standard and it was believed that most of the mismatch was away from the DUT interface. In this case, it was possible to do a complete calibration at the inner reference plane so a comparison was possible. The extracted vs. nominal insertion loss is shown in Figure 9-31. One can see pretty good agreement until about 30 GHz when it did work out that DUT-plane mismatch on the fixture was getting large. Further, the fixture started having significant radiation above about 35 GHz which further complicated the extraction. The return loss (extracted and nominal again) values are also plotted in Figure 9-31 and again show reasonable agreement until the very high frequencies. Recall that uncertainty in return loss in dB terms gets much larger as the match gets very good just based on a residual directivity argument (a few dB at the –20 dB level for a decent coaxial calibration).

This example does reinforce a couple of points:

- Partial information methods do have some fixed error because of the incomplete 'calibration' at the inner plane
- The further the fixture deviates from the ideal aspects assumed by the method (where mismatch is located in this case), the larger those errors become.

Still, if it was indeed only possible to have an open standard for this fixture, the results shown here are better than one could achieve with a single-standard generalized B method or with a simple normalization.

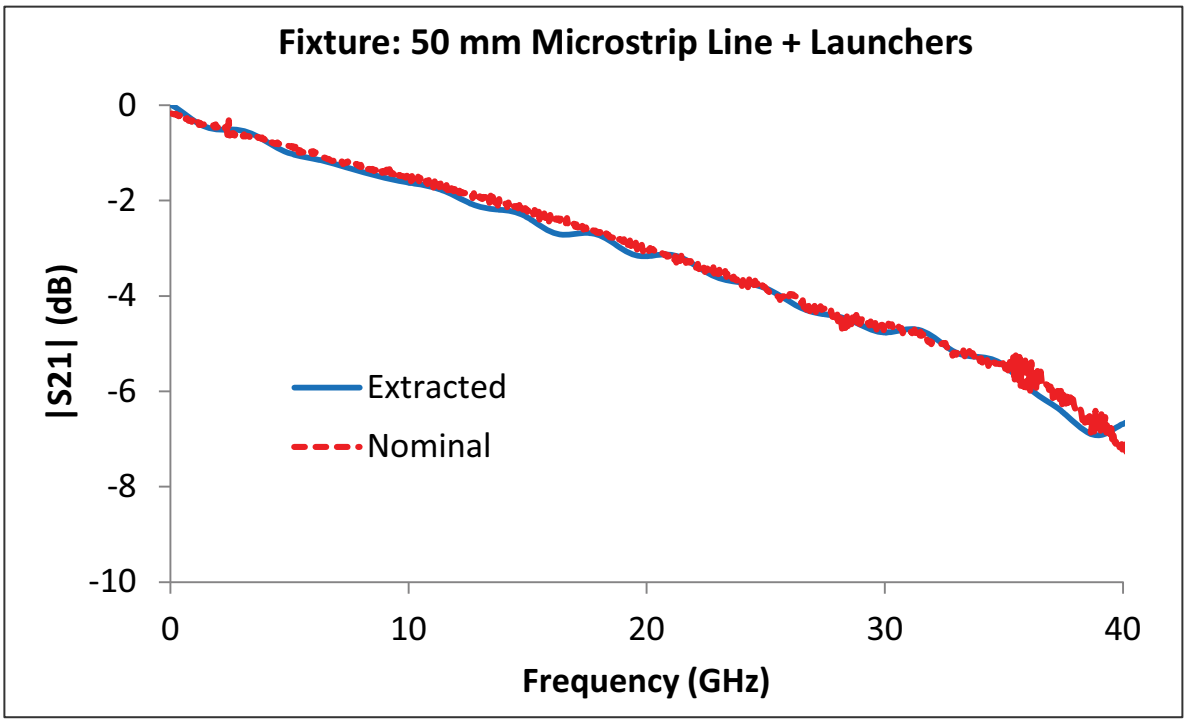
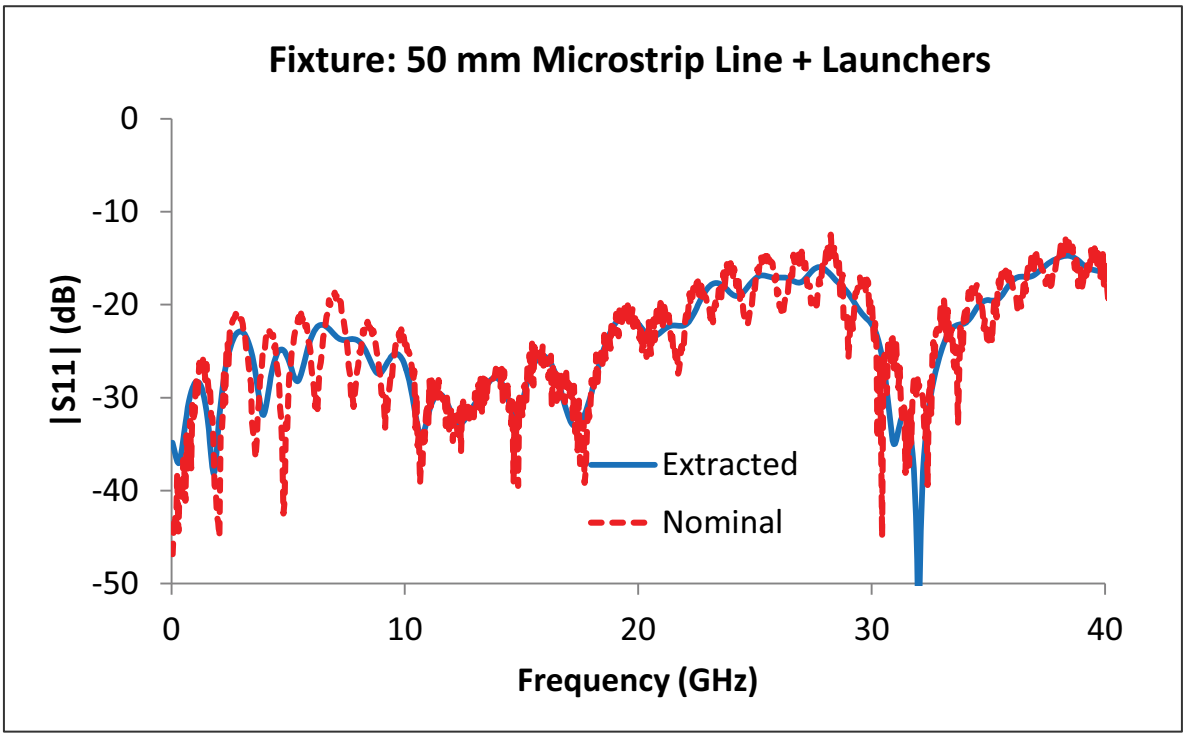


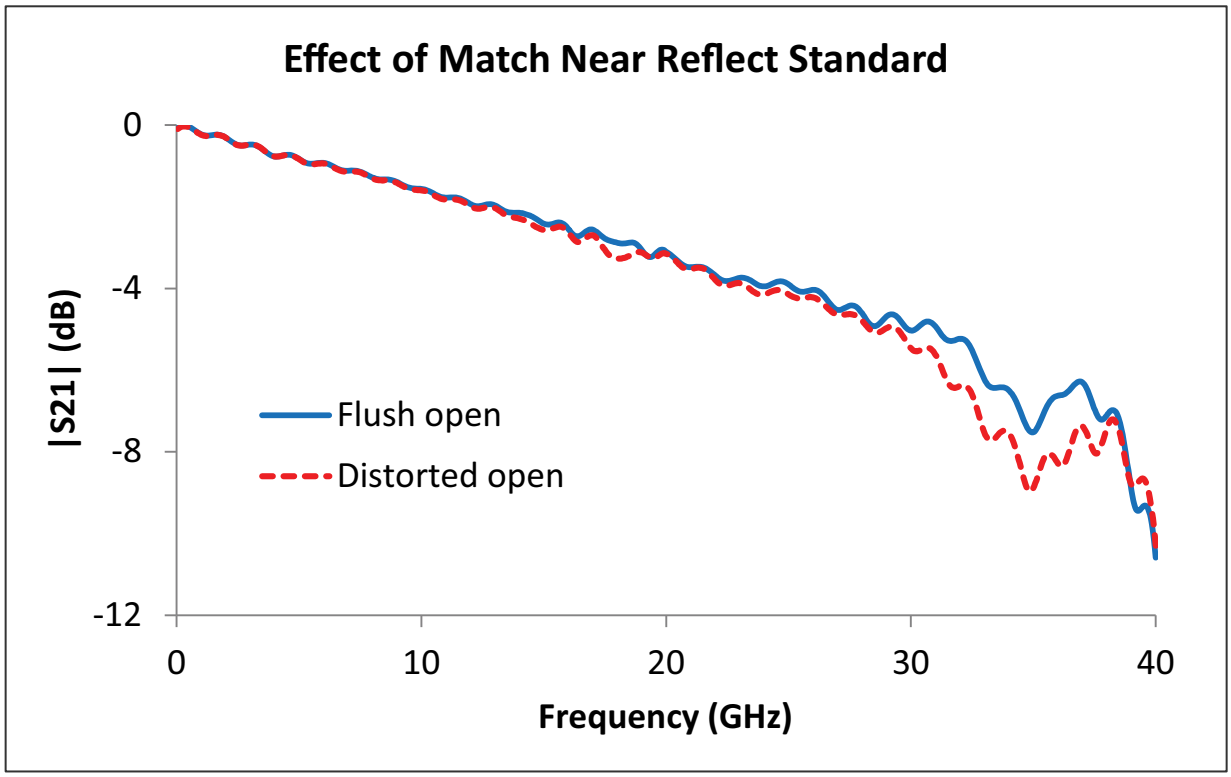
Figure 9-31. Microstrip Line Example (1 of 2)



The results for a phase-localized Type D extraction are shown here along with nominal results for a special case when both port of the fixture arm were connectorized (to allow for comparison). The extraction degrades at higher frequency as fixture radiation and inner-plane mismatch hamper the partial information technique. Only an open reflection was used for the extraction process.

Figure 9-31. Microstrip Line Example (2 of 2)

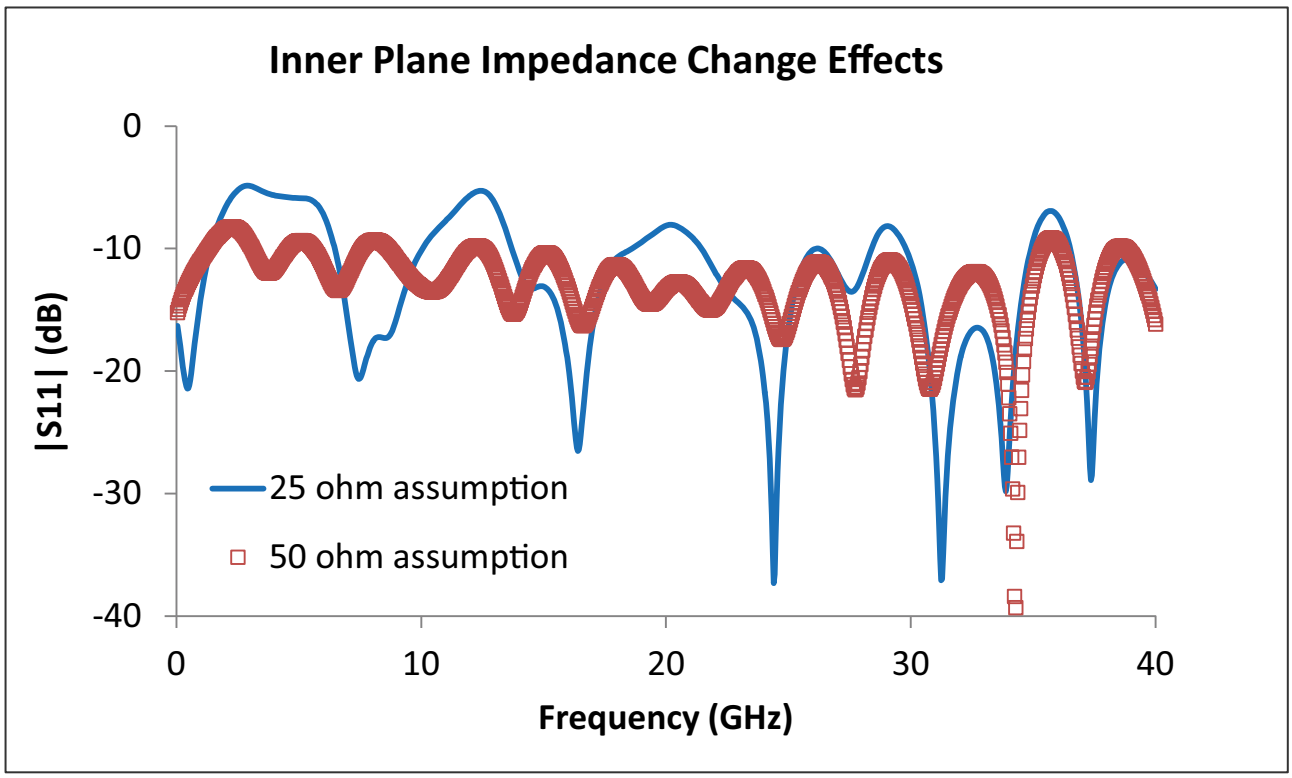
The direct sensitivity to standards definition defects are fairly straightforward since the reflection or transmission coefficient entered is applied multiplicatively to the data prior to a square-root operation. More subtle are non-idealities in the standards (such as mismatch of the line/thru standard). This will have impact through its influence on actual insertion loss during the measurement (in terms of an offset loss and in terms of ripple). Also on the subtle side are the effects of an incorrect fixture length entry. While in normal Type D this is mainly used for root choice, it is used in phase-localized D to determine which phase signatures to correlate against so entering a significantly incorrect value (10s of mm generally) can cause added ripple and, eventually, drop outs in insertion loss extractions as well as incorrect return loss values. The internal length estimate approach (entering 0 in the fixture length estimate field triggers this) can reduce the issues and is recommended unless the fixture phase response is very resonant, in which case a proper manual estimate will yield better results. The issue of where the mismatch is predominantly located was touched on earlier. This has somewhat more of an effect when using the reflect standard (since the fixture mismatch and the reflect standard are almost co-located so phase localization becomes difficult) than using the thru/line standard. As an example, consider the extraction of a microstrip section using a flush open reflect standard native and when the mismatch near that standard has been distorted from the original ~ –15 dB to ~–5 dB at high frequencies. The effects on fixture $|S_{21}|$ are shown in Figure 9-32: added ripple and some substantial differences above 30 GHz (where the mismatch change was the largest).



To further explore the sensitivity to inner-plane mismatch, a (reflect-based) phase-localized D extraction was performed on an original fixture and again after additional mismatch was introduced. In the range of additional mismatch addition, the discrepancies increased as expected.

Figure 9-32. Inner Plane Match Values

Fixtures with distinct impedance changes, intentionally constructed or otherwise, present an additional challenge. Consider a fixture where the inner plane is at 25 ohms and a thru standard is used for the extraction. If the inner plane impedance is ignored, then one is essentially treating the half fixture as being terminated in a 25 ohm impedance (low reflection for that zone) but that is implicitly changing the reference impedance of the S-parameter matrix. While this may sometimes be desired, it will cause errors if not anticipated when using that file for later de-embedding of modeling. More conventional is to keep the reference impedance (of the matrix) consistent at 50 ohms to facilitate later processing. In this case, the difference in the match parameters is substantial as shown in Figure 9-33.



In a (thru-based) phase-localized D extraction, inner plane impedance deviations can create issues if large enough. In this case, the inner plane was at 25 ohms while the launch (and reference impedance for the calibration) was 50 ohms. If the impedance change was ignored (red squares in the plot), the extracted return loss for the fixture half can be significantly in error.

Figure 9-33. Phase Localized Type D Extraction—Effects of Inner Plane Impedance Change

While the above discussion focused on the use of a single thru line or a reflect standard, there are some standards permutations allowed. Two reflect standards (different from each other, but the same reflects are used on both ports if both ports are being analyzed) can be used where the added information is used to refine fixture transmission results (and inner plane match to some degree). Also, a thru can be used in conjunction with a reflect standard where the additional information is used to refine parameters particularly in the case where the thru is not particularly well-characterized. A weighting term is provided (range 0 to 1, where 1 indicates the thru measurement is used more heavily and a 0 means the reflect measurement is used more heavily). It is not allowed to use a thru with two reflects as the amount of information increase was found to be diminishing.

Additional Notes:

- A full 2-port calibration must be active and the extraction will be run over the current frequency range (which is a subset usually of the calibration frequency range). For 4-port systems, at least a full 2-port calibration must be active (more details on the 4-port cases are covered in Chapter 24 of this measurement guide). There is a requirement that the frequency list have nearly uniform frequency steps

(an individual step size cannot deviate from the mean by more than 5%) so some segmented sweep setups (and all log sweep and CW setups) will not be accepted.

- The frequency range of the sweep should be large enough that the total fixture length (ns)>5/(frequency range (GHz)). The frequency step should be small enough that the total fixture length (ns) < 0.3/(Frequency step (GHz)). This helps avoid insufficient phase slope or phase-wrap-aliasing (respectively) that would complicate phase localization.
- The line and reflect offset lengths are entered in millimeters although a calculator is available if values are in picoseconds. If the material type is setup (from the current calibration or manually thereafter), that and any active dispersion relations will be used in the calculations.
- The extracted results are stored as .s2p files with port 1 of each file being the outer plane. Details of the file format options (frequency units, etc.) are set by the entries on the sNp setup dialog. See Table 9-1 for recommendations of extraction types and standards needed for various fixture behaviors.

Table 9-1. Standards Requirements for Generalized B and Type D Extractions

Method	Standards Needed	Best for Fixtures
Generalized B	Open Open/Short	Well-matched fixtures with very well-matched inner plane. Reflection-only standards possible
Existing D	Thru	Well-matched fixtures with very well-matched inner plane. Thru-line standard possible
Multi-standard D	2 lines Line + (Open OR Short)	Moderately-matched fixtures without structural assumptions other than symmetry. At least one line standard possible. Good for high loss fixtures.
Phased-localized D	Line Or (Open OR Short)	Moderately mismatched fixture assuming most mismatch not at inner plane. No symmetry necessary. One standard only

Another way of looking at method choice is with the flow chart shown in the Figure 9-34.

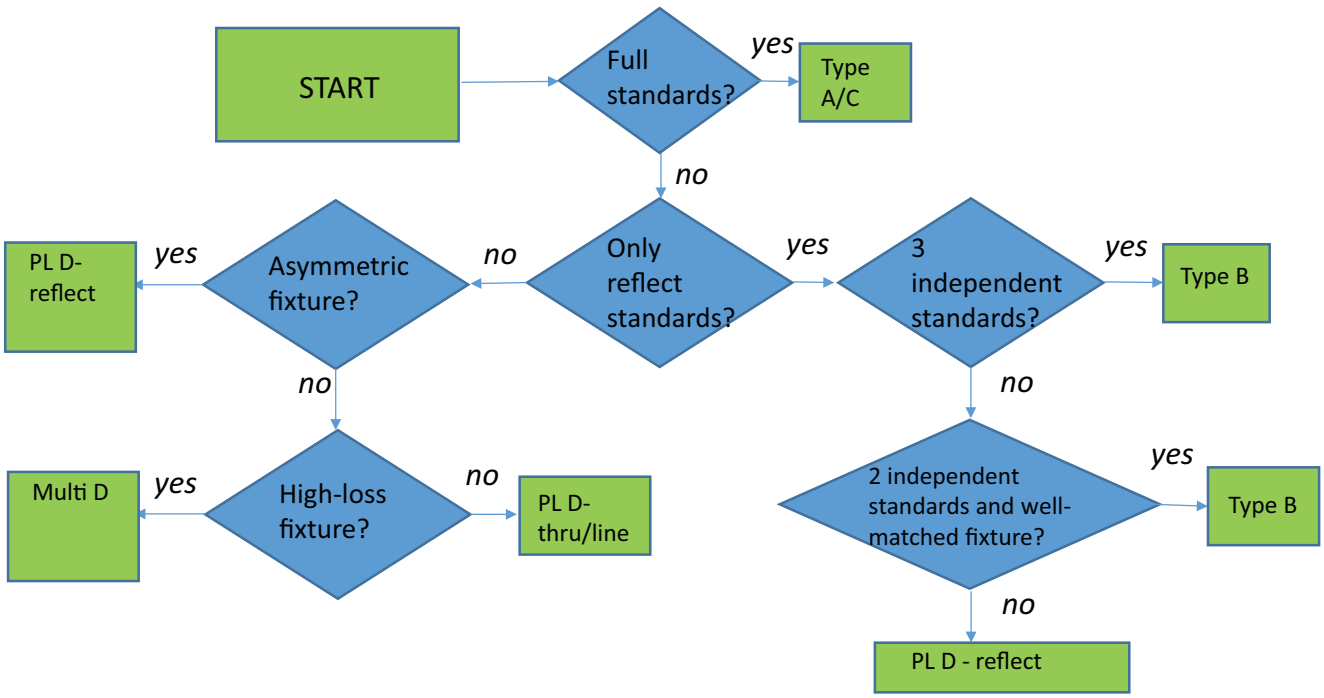


Figure 9-34. Standards Requirements for Generalized B and Type D Extractions Flow Chart

Sequential Extraction—Peeling (with Option 21)

Another method of network extraction involves modeling the network as a collection of lumped elements. This is particularly popular for electrically small structures (e.g., on-wafer) of those with runs of transmission line punctuated by electrically small structures (e.g., PC boards with isolated vias in transmission lines). Procedurally, this method works on one lumped element at a time. For each element, a .s2p file is generated that can be de-embedded to allow one to get at the next element. Also transmission line segments can be separately de-embedded to get between lumped defect areas. The process is based on reflection measurements only and a full calibration incorporating that port must be in force.

The basic method accepts as input the location (in time from the reference plane) of the defect area of interest and the type of element to model the structure: shunt admittance (Y) or series impedance (Z). A 0 can also be entered as the position and, in that case, an automatic calculation will be performed to select the largest remaining defect. A differential pair can also be selected in which case the model element is a crossbar impedance (between the two ports) and a .s4p file will be generated. The dialog is shown in Figure 9-35.

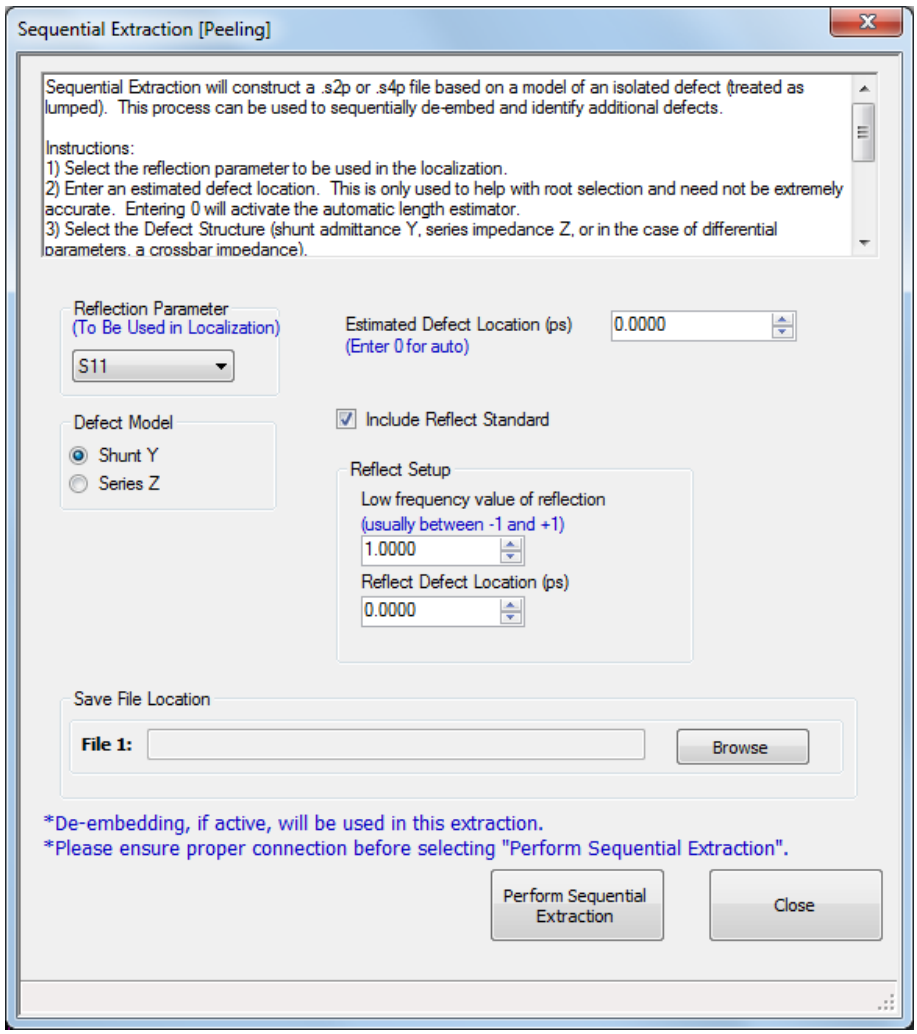
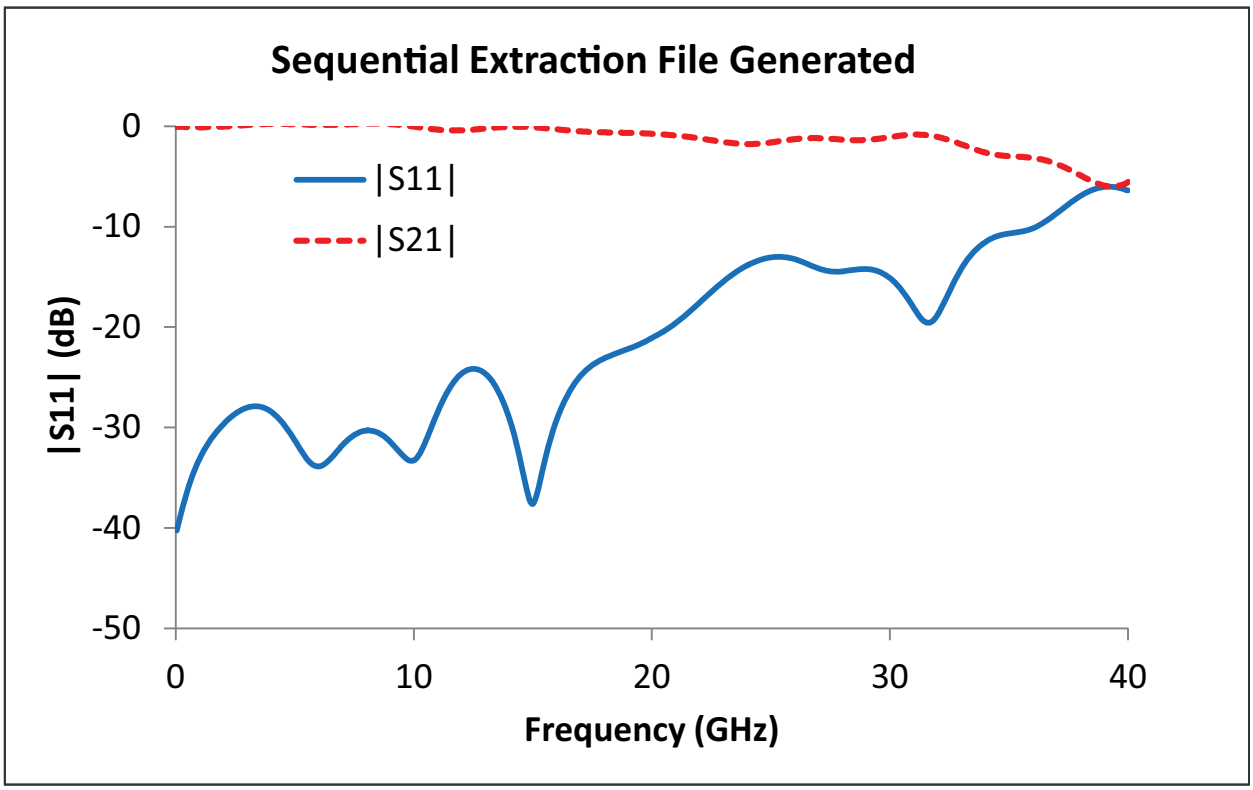


Figure 9-35. Sequential Extraction Dialog

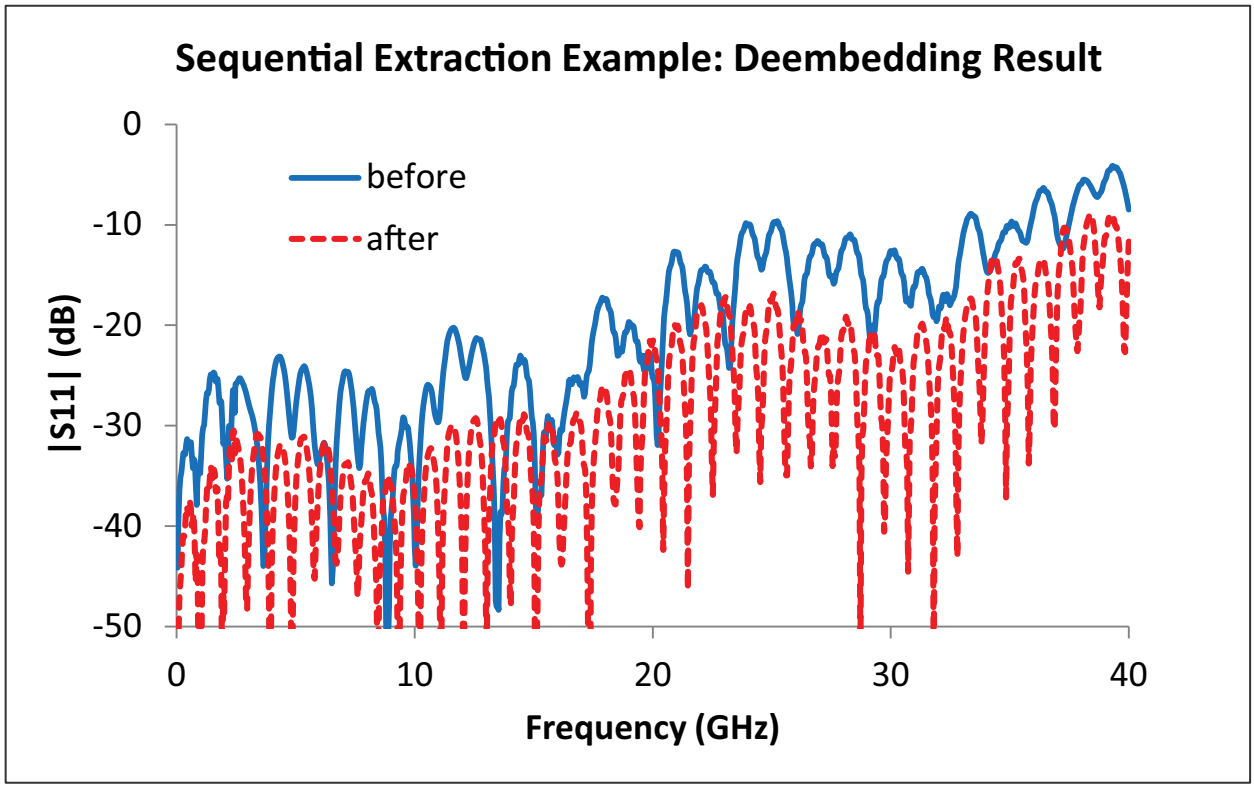
As an example, consider a fixture with multiple transitions but one is known to be particularly inductive (in a series sense). It might be interesting to look at the return loss of this fixture if that particular transition could be improved. It is known that this launch is about 120 ps in from the reference plane so the sequential extraction tool was used with that defect position entry and a Z-series element selection. A 2-port calibration had already been performed. The extracted .s2p file for the element shows the result in Figure 9-36. As might be expected, the return loss of the model element is indeed degrading with frequency and the insertion loss increases.



Values from the series network extracted for the example fixture are plotted here.

Figure 9-36. Sequential Extraction Plot

To remove this from the fixture result, two steps are needed: de-embed the 120 ps (air equivalent 36 mm) of transmission line and de-embed the file just generated. This was done and the before and after de-embedding results are shown in Figure 9-37. Indeed, the inductive transition was responsible for a fairly large share of the overall fixture mismatch and improving that one transition could have significant benefits. The final result is not perfect since there are additional transitions in the fixture but also because this extraction is somewhat model-like: it is only valid if the defect is indeed series in nature (or shunt if the admittance version had been selected). In this particular case, the series model was a very reasonable choice.

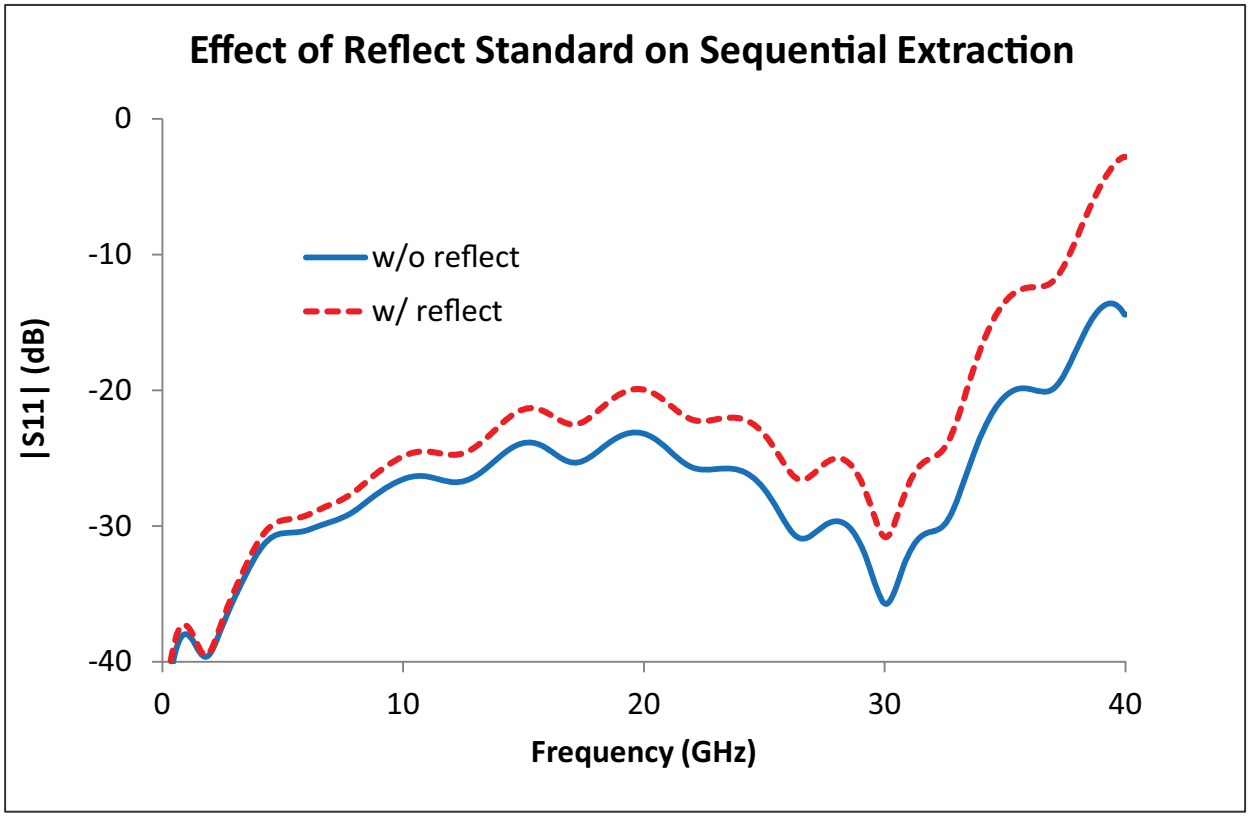


The before and after de-embedding return loss values of the example fixture are shown here.

Figure 9-37. Sequential Extraction Example—Deembedding Result Comparison

The basic method is adequate for electrically small structures but inaccuracies can enter the picture if there are un-de-embedded losses between the reference plane and the defect center. An extension to this method can help compensate for these losses by using analysis of another, known reflection center. If there is a known reflection at the end of the fixture (e.g., an open or a short), the analysis of that response can be used to compute a loss estimate and correct the extraction of a lumped element representing a defect between the reference plane and the reflection standard. The reflection coefficient of the standard must be entered (must be real and normally between –1 and +1) and its position relative to the reference plane is usually entered. A zero (0) can be entered for the location and an automatic routine will be used to locate the largest response which is presumed to be due to the reflection standard. If an auto routine (0 entry) is also used for the defect location, the next larger response (distance less than that of the reflect standard) will be used for that item.

To see the effect of the reflect standard, consider the effort to extract the .s2p file for a via structure in a transmission line. The defect is located ~170 ps in from the reference plane but the transmission line is of fairly high loss. It is possible to place an open reflect standard at about 400 ps from the reference plane and that open is electromagnetically well-behaved at least to 40 GHz (our range of interest). The via structure will be modeled as a series impedance and the simple sequential extraction produces a file with the return loss shown as the solid curve in Figure 9-38. If one also uses the reflect standard (position 400 ps, reflection coefficient = +1), one gets the dashed line instead. Ignoring the line loss up until the defect would have cause one to underestimate the defect return loss by nearly 10 dB at 40 GHz.



As shown with this example, using the reflect standard to compensate for line loss can be useful if the defect is far from the reference plane in a lossy medium.

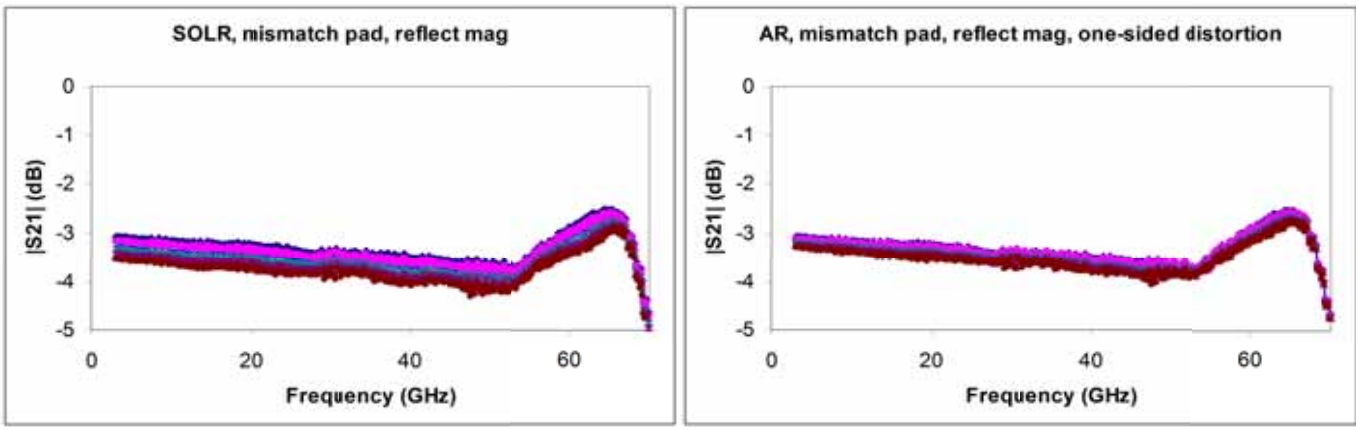
Figure 9-38. Sequential Extraction—Effect of Reflect Standard

9-6 Uncertainty and Sensitivity

The selection of a network extraction method is heavily dependent on the standards that are available at the inner plane and the results can vary wildly depending on the quality of and knowledge about those standards. While we cannot be complete in this guide in a discussion on uncertainties and sensitivities, we can give some general thoughts. Many publications exist on the topic that may provide more information, such as *J. Martens, "Common adapter/fixture extraction techniques: sensitivities to calibration anomalies, 74th ARFTG Conf. Dig., Dec. 2009,* and references therein.

The Type C extraction is, in some sense, the most complete since it uses all of the error terms in two full port calibrations to extract the S-parameters of the individual fixture arms. For that reason, Type C is also the most sensitive to standards quality at the inner plane. Uncertainties will follow those of the underlying calibrations quite closely, subject to a repeatability penalty associated with the two calibrations.

Type A is a bit different in that there is only one fixture arm, but it does still use two full-term calibrations. The computation difference is that it does not use transmission or load match terms to calculate the adapter/fixture S-parameters and hence is more immune to problems with those calibration steps than is Type C. Even if there is a problem with a reflect standard, Type A may be less sensitive since the reflect errors propagate to load match in the standard calibrations (particularly true for reciprocal methods as is discussed in Chapter 16). As an example, a mismatched pad was evaluated using both Type A extraction and an SOLR calibration, but a perturbation was introduced in one of the reflect calibration standards. As shown in Figure 9-39, the sensitivity with Type A (labeled AR for adapter removal) is lower. The comment 'one-sided-distortion' is included in the plot since symmetric errors in a Type A setup tend to cancel (another advantage in some cases), so the introduced error was made on only one side in this measurement to create a worst-case result. Each trace in the plots represents a different assumed reflect magnitude value for the standard (over a 10 % interval).



A comparison of the effects of a reflect standard problem on the transmission of an example DUT are shown here for both type A extraction (AR, adapter removal) and a SOLR calibration. The various traces represent the result with different assumed reflect standard offset lengths.

Figure 9-39. Comparison of Reflected Standard Problem for Type A Extraction and SOLR Calibration.

Although less sensitive to some effects, it may not always be possible to do reasonable full port calibrations to support Type A. The Type B method requires only one port standards, which can be far more convenient. The internal computations are similar between Type A and Type B (in the full-standards case) except in how output match is determined. Type B is a more source match-intensive computation, so there is greater sensitivity to the high reflect standards' behavior with Type B than with Type A (and less dependence on low reflect behavior). A comparison is shown in Figure 9-40 for Type A (again labeled AR for adapter removal) and Type B (labeled BP for Bauer-Penfield, two of the original authors to work on this method class). The sensitivity of output match to a reflect standard problem (a short in this case) is indeed greater. Again, each trace represents a different reflect magnitude value. Type B with flexible standards falls into the category of 'partial information methods' like Type D. With the partial information techniques, it is useful to distinguish between sensitivity to standards problems from absolute error (even if the standards were perfect). With network extraction methods A, B (in its full-standards form) and C, the absolute error is zero. Type B when using one or two standards allows one to accept some non-zero absolute error in exchange for reduced sensitivities to standards errors and repeatability. These sensitivities for Type B were discussed in the section describing "Type B Network Extraction".

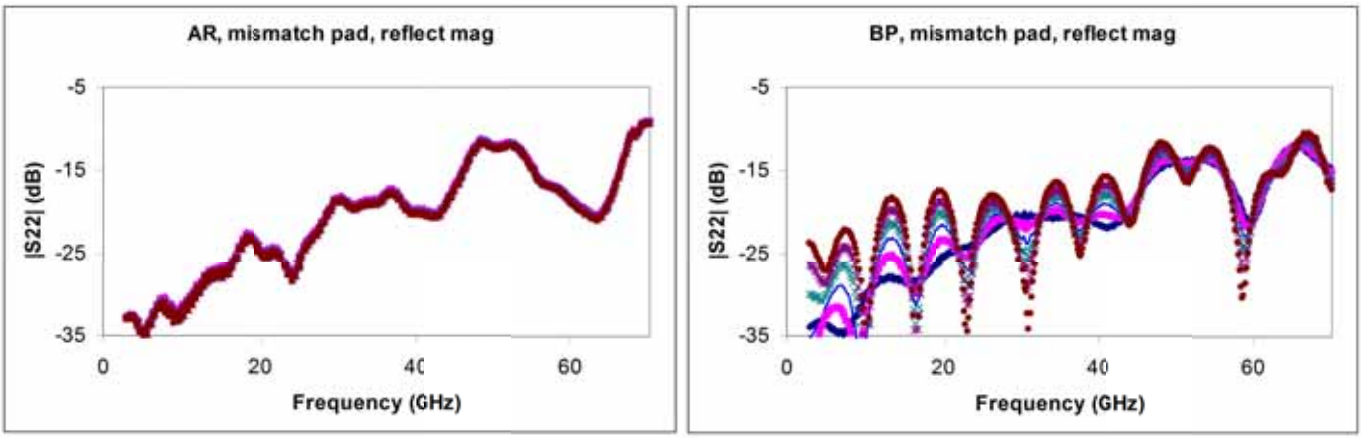


Figure 9-40. Comparison of Output Match Sensitivity for Type A (AR) and Type B (BP) Methods

Correspondingly, the load sensitivity is somewhat reduced, but it will have an impact on high RL adapters and high IL adapters, as suggested in Figure 9-41.

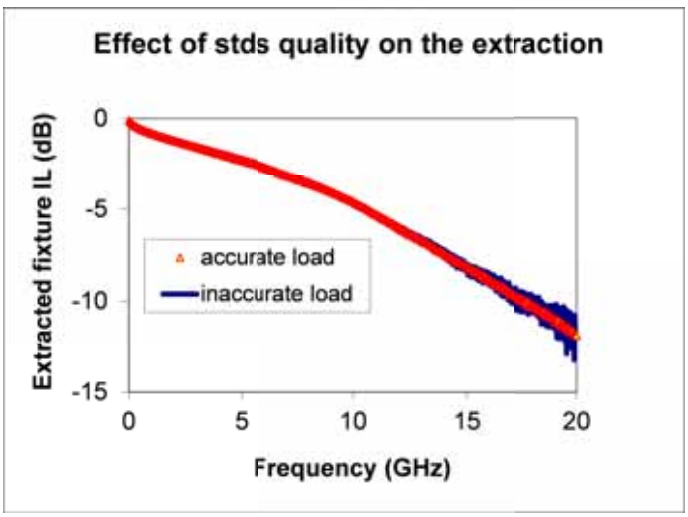
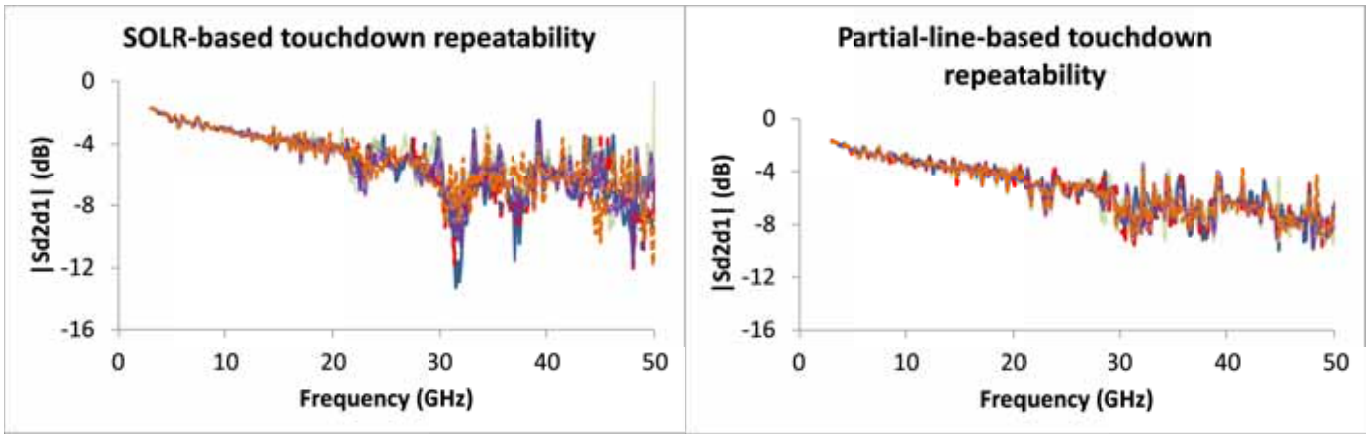


Figure 9-41. Effect of Load Standard Problem on Extracted Insertion Loss (1 of 2)

The effect of a load standard problem on extracted insertion loss with Type B. Eventually, the load accuracy does matter.

Figure 9-41. Effect of Load Standard Problem on Extracted Insertion Loss (2 of 2)

Type D is another approach entirely and is appropriate when it is difficult to create any reasonable standards at the inner interface. Also, when repeatability is problematic (for example, spring contacts, poorly positioned probes, etc.) trying to achieve knowledge of fewer parameters can be useful. The concept is to reduce weight on the inner plane match and/or make additional structural assumptions and focus on improving the accuracy of the insertion loss extraction. The absolute accuracy will be degraded from what one could get from either of the other methods (if good standards were available), but repeatability sensitivity can be greatly improved and the net practical uncertainty can actually be better. This follows from the 1/(1-x) kind of behavior of match terms in the standard methods. If the measurements are not terribly repeatable and the 'x' term is moving near unity, the uncertainty on the final parameter in practical terms can balloon. One can see this in a wafer probing example where probe placement was not that accurate. A series of calibrations were done using SOLR and using Type D extraction with a single thru (labeled partial information in Figure 9-42). The removal of match dependence lowered the scatter in the final insertion loss values. While repeatability sensitivity is better, there is sensitivity to inner plane match and to problems with the underlying calibration at the outer planes (although reduced).

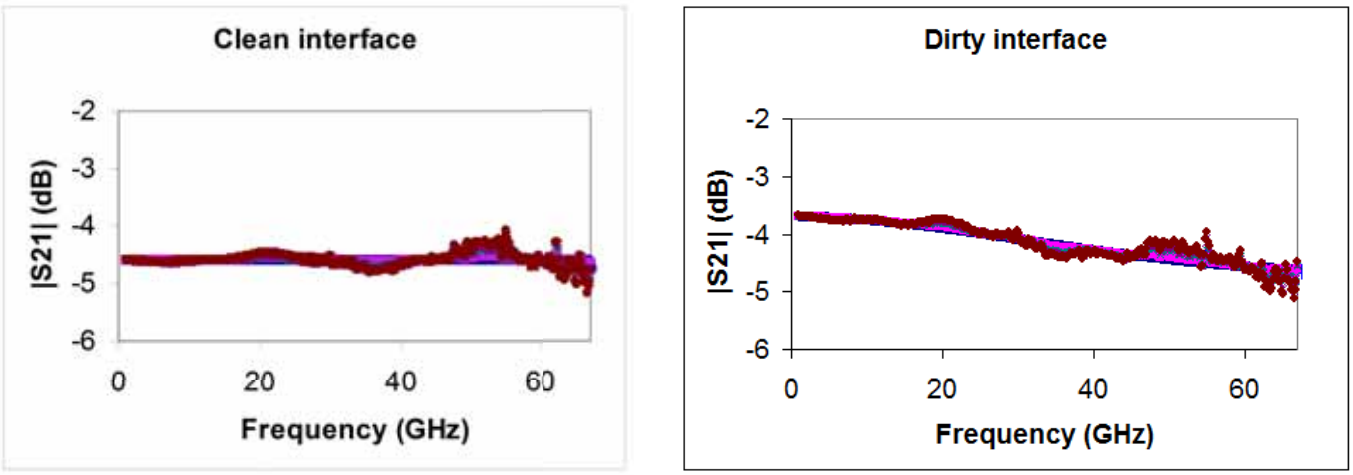


A comparison of repeatability effects is shown here for two methods (standard calibration and Type D) when that repeatability was not good.

Figure 9-42. Comparison of Repeatability Effects

While repeatability sensitivity is better, there is sensitivity to inner plane match and to problems with the underlying calibration at the outer planes (although reduced).

The effect of an inner plane distortion is shown in Figure 9-43. The effects can be substantial, but the errors in this particular example would have exceeded 4 dB with a Type C methodology because the media was pin-based and not repeatable.



The sensitivities of a Type D extraction to underlying calibration issues (multiple traces per plot, reflect magnitude variations) and introduction of an inner plane match and position problem (the dirty interface plot).

Figure 9-43. Sensitivities of a Type D Extraction to Underlying Calibration Issues

Some of the Type D sensitivities were covered in the previous section. As with the one- and two-standards variants of Type B, Type D allows one to accept some non-zero absolute error in exchange for reduced sensitivities to standards errors and repeatability. Within Type D, there are a number of choices depending on which standards are available. If a thru is available and the fixture halves are relatively symmetric, phase localized D with a thru standard can do very well unless the fixture is very mismatched at the inner plane. Multi-standard variations of D (using two lines or a line and a reflect) can take over in the latter case. If only a reflect standard is available, reflect-based phase-localized D will usually outperform Type B with one standard.

The various extraction methods discussed here present a variety of choices dependent on standards quality, the media involved, and the possible measurement repeatability. In some sense it is a continuum (C->A->B->D) of choices more appropriate as the environment becomes decreasingly metrology-friendly. This is an oversimplification, but some of the sensitivities presented may help in making a good choice for a given measurement setup.

9-7 Enforced Passivity and Causality

When S-parameter data from network extraction or just from measurement are used in external simulators, there are sometimes requirements on the data in terms of passivity or causality for simulation convergence or for other performance requirements. It is possible to enforce these characteristics on .sNp saved data (either measurement or network extraction) although changes are made to the data in that process. This section will discuss the enforcement procedures and some of the implications. Figure 9-44 shows the sNp setup dialog for enforcing Passivity and Causality, which can be accessed via:

Main | System | Setup | Misc Setup | SnP Files Setup | SNP SETUP dialog

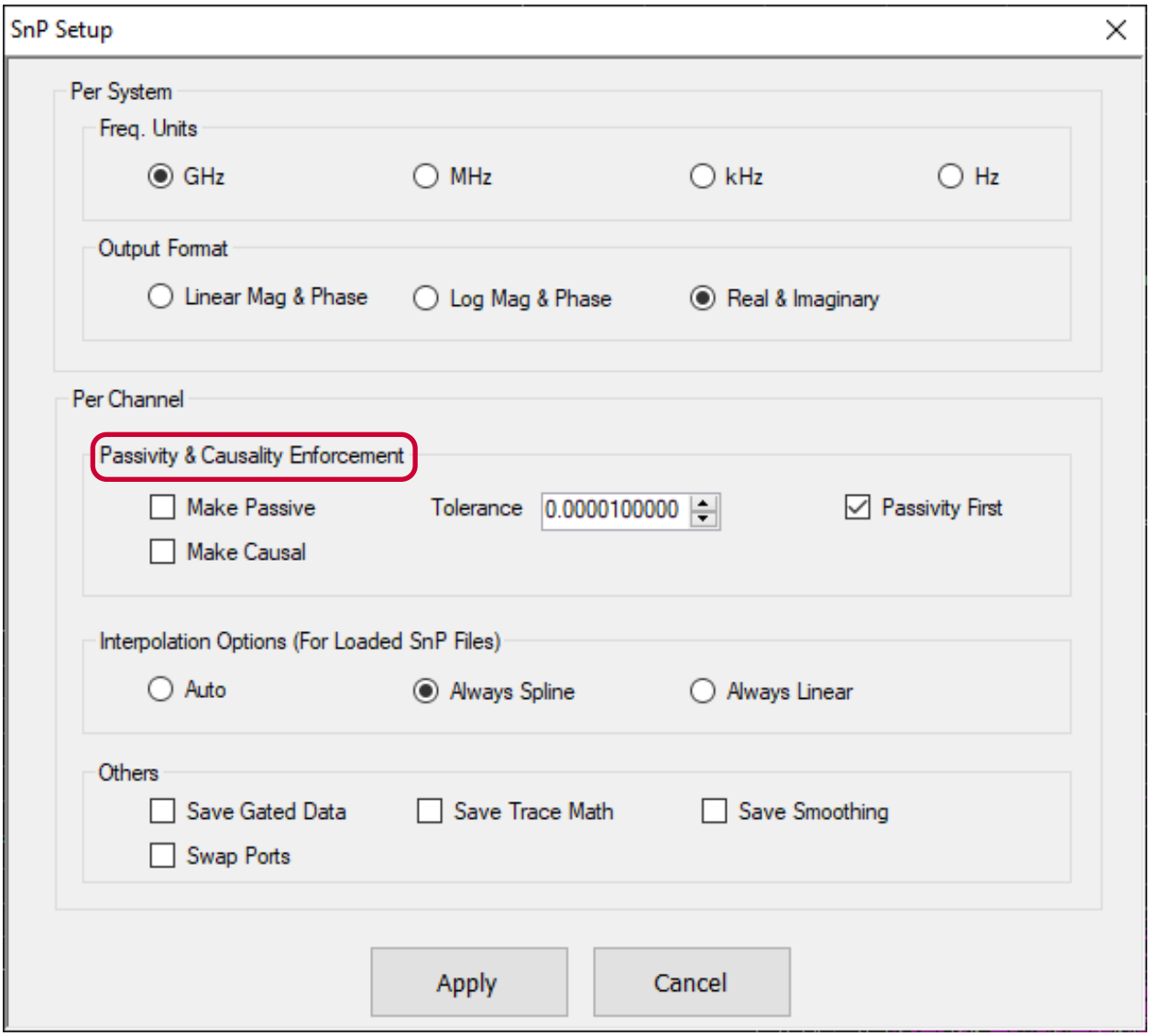


Figure 9-44. SnP Setup Dialog

Passivity

Here, this is defined as the 2-norm of the S-parameter matrix being less than or equal to one. This is equivalent to the eigenvalues of the S-parameter matrix all being less than or equal to one in magnitude. The reader may associate passivity with $|S21| \leq 1$ but that is not a sufficient requirement for most situations where convergence is an issue.

In practice, many passivity violations (on actually passive devices; those not capable of energy conversion) arise from calibration problems (standards issues), cable or drift issues, or repeatability variances (connections or probe touchdowns). Ideally, those issues should be addressed to improve the basic data quality. However, for very low loss devices/fixtures, the data may only be marginally non-passive and it may be expedient to coerce that data to proceed with a simulation, system design, etc. The Make Passive selection will do this on all .sNp saves but it is up to the user to assess if the changes being made to the data are acceptable.

Typically the changes are quite small if the calibration quality is good. As an example, the measurement of an adapter chain is shown in Figure 9-45 and the passivity-enforced result is overlaid on top (aside: as stated elsewhere, this can be done by placing the instrument in Hold and recalling the .s2p file that was saved (with passivity enforcement in this case); the actual measured data was saved to trace memory). One can see some slight changes under 2 GHz where the insertion loss is the lowest. The 'enforced' insertion loss at very low frequencies is increased by ~0.005 dB and the return loss is increased by about 1 dB at the –40 dB level. This level of change is far below the uncertainty levels and is likely to be acceptable.

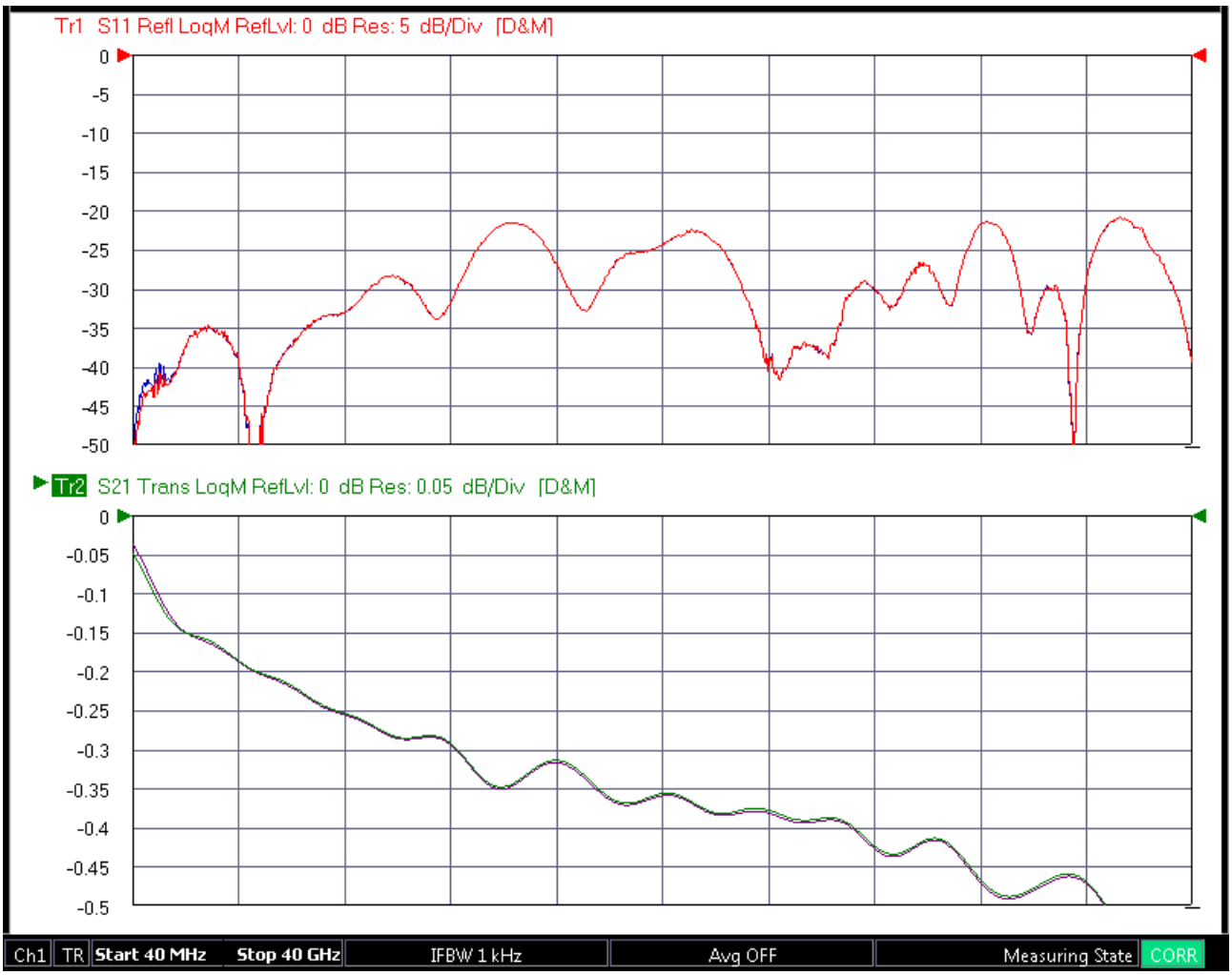
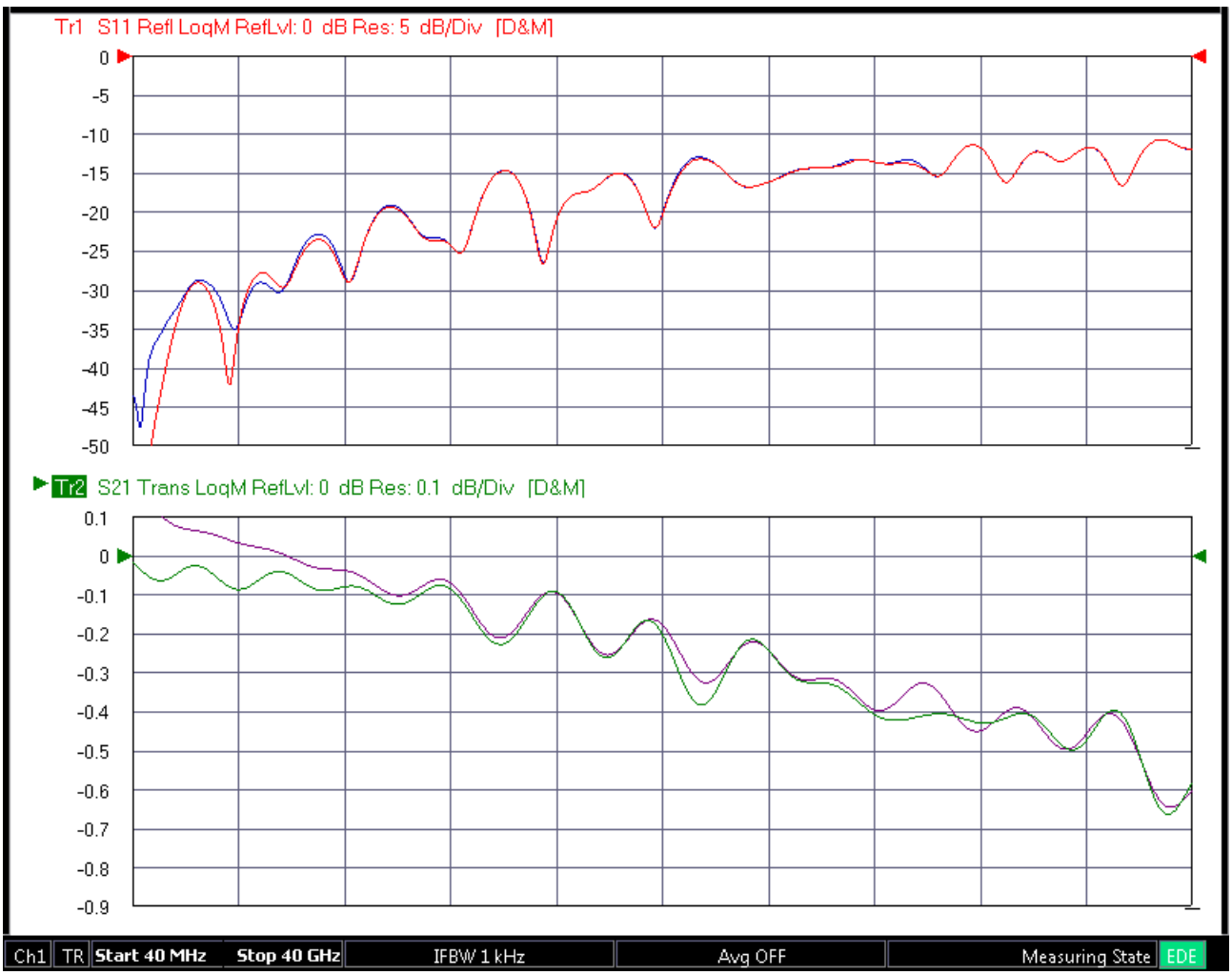


Figure 9-45. Measurement of an Adapter Chain and an Enforced-passivity Version of that Data (1 of 2)

The measurement of an adapter chain and an enforced-passivity version of that data are shown here overlaid. There is only a difference at very low frequencies.

Figure 9-45. Measurement of an Adapter Chain and an Enforced-passivity Version of that Data (2 of 2)

If the calibration quality is poor, the changes can be more dramatic. To simulate this, a non-existent transmission line was de-embedded from the original calibration so insertion loss reads artificially low (and indeed shows insertion gain for the adapter chain at low frequency). The change from passivity enforcement in this case exceeds 0.1 dB at low frequencies and is noticeable at higher frequencies (see Figure 9-46). This happens because the introduced calibration degradation also caused apparent mismatch levels to be much higher. This is important since now the 2-norm can exceed unity even if |S21| is ~–0.4 dB. The changes made in enforcement here are significant and, if they occurred with a real calibration, one might want to investigate the calibration and measurement (cables, connectors, standards, etc.) before continuing with the enforced data.



With calibration problems (actual ones or those introduced artificially like in this example), the passivity enforcement can have more dramatic effects.

Figure 9-46. Passivity Enforcement with Calibration Problems

The passivity enforcement works by re-scaling the eigenvalues of the S-parameter matrix in a self-consistent way and then regenerating the S-parameters from the new eigenvalues. The consistency of this approach is important since one wants to maintain the relationship between reflection and transmission parameters as much as possible to lessen the effect on any other system calculations.

The user is given some control over how much less than unity the 2-norm must be since some simulators screen for passivity using tests that are more-or-less stringent. A 'tolerance' variable is provided and if the entry is *x*, the maximum eigenvalue magnitude will be *1-sqrt(x)*. The default value for *x* is 0.00001 and is allowed to go as high as 0.001 for the more stringent tests.

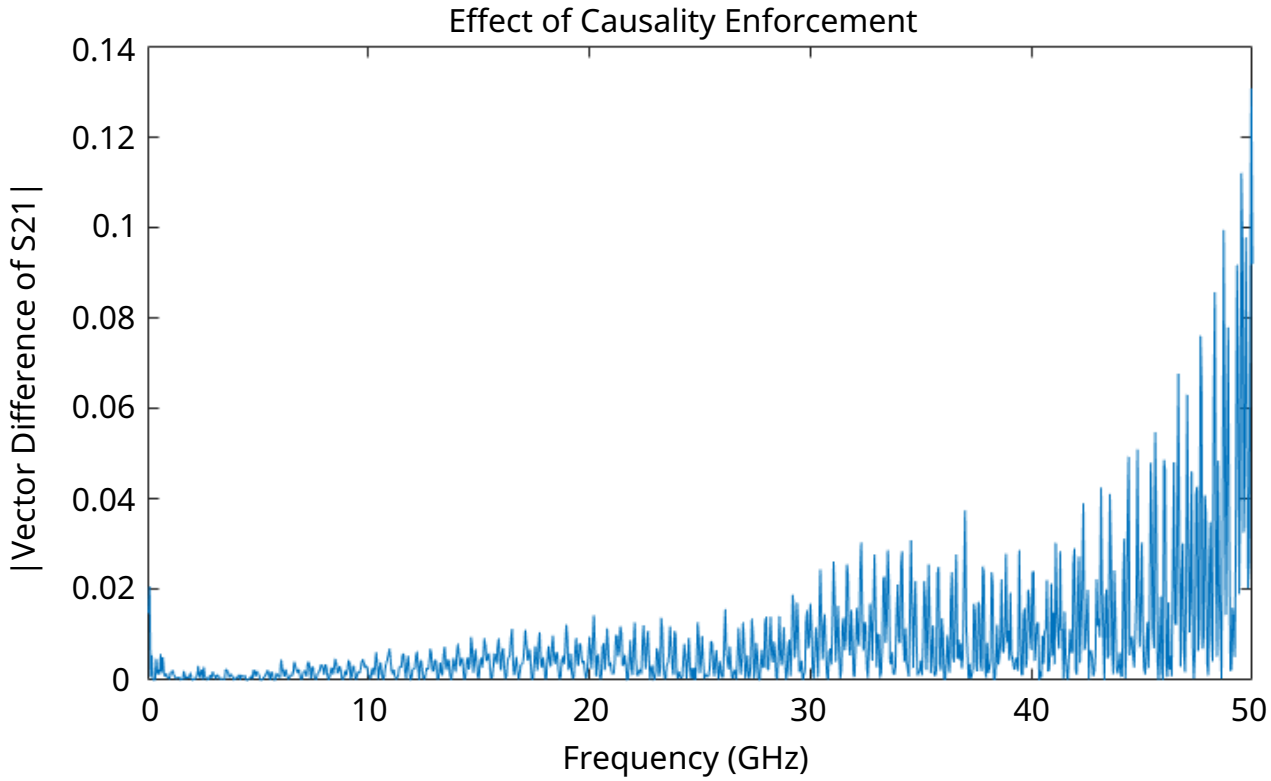
Causality

Causality is the state of not being dependent on the given parameter value at an earlier point in time. This has a number of implications:

- The time domain representation should have no energy before time t=0 (assuming the reference planes have not been shifted and are in a physical location).
- Often it implies the parameter is analytic on the upper half of the complex plane. This conclusion does require that the parameter falls off fast enough with frequency along with a few other conditions.
    - This in turn implies that Kramers-Kronig relations apply. Importantly, this means that the real and imaginary parts (or, equivalently, the magnitude and phase) of the parameter are not independent.

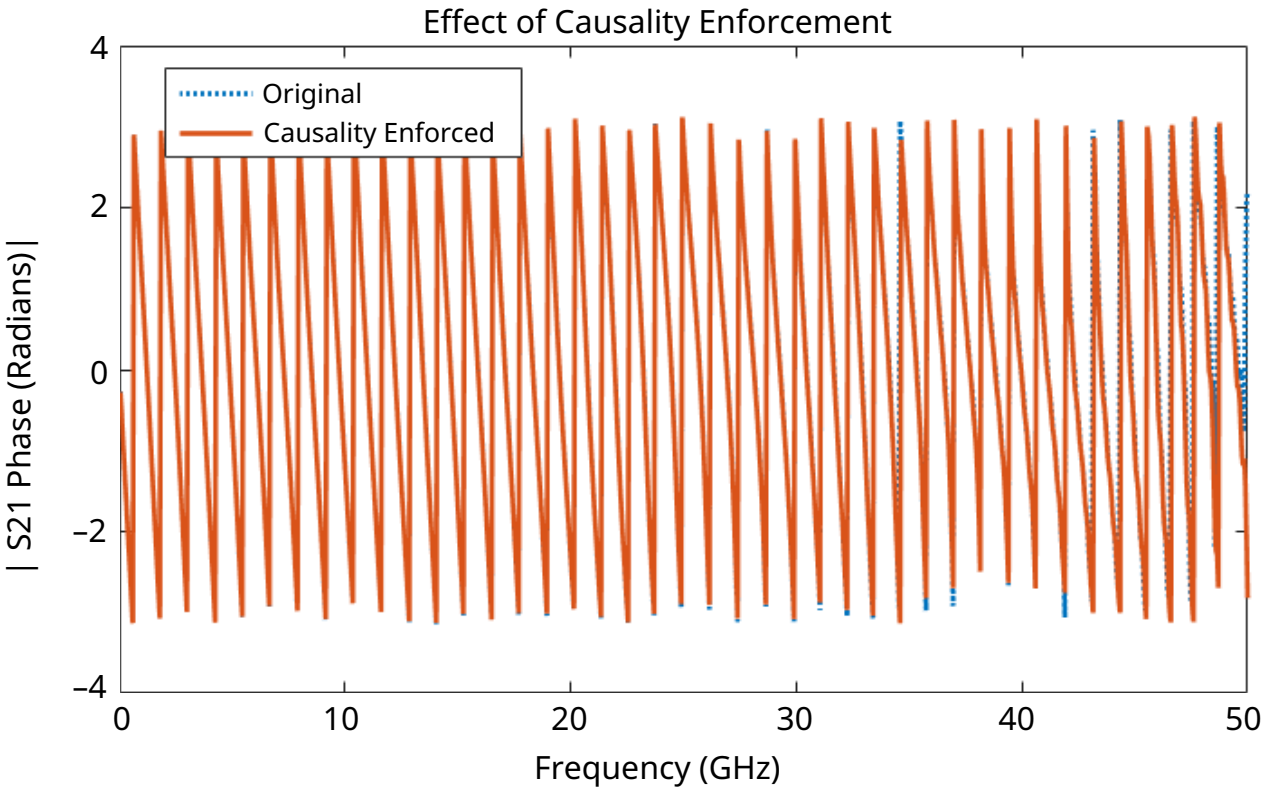
As with passivity, this condition may be required of S-parameter data that is to be used in certain simulations and, again, there is a way to coerce data to make this happen. In practice, causality violations can occur from calibration and drift issues (particularly nonlinear phase drift) and from measurements of very broadband devices with a small measurement bandwidth (so there is little roll-off). This last statement follows from the observation that any limited bandwidth set of data, when transformed to the time domain, will have energy for t<0. This energy becomes significant if there is little parameter roll-off at the highest measurement frequency. Thus, using more measurement bandwidth can help, up until the point where DUT or fixture radiative behaviors make the data unstable.

The corrections in this category also tend to be small and they are largest often at the higher frequencies. The plot in Figure 9-47 shows the vector difference of S21 between original and enforced data on a mismatched line of modest insertion loss (~8 dB at 50 GHz and 50 GHz of measurement bandwidth). The change is not visible on the magnitude of S21 in this example but the alteration can be seen in phase Figure 9-48 as the distortions in phase have more of an effect on the temporal distribution of energy.



The vector difference caused by causality enforcement is plotted here. The corrections are often largest at the higher frequencies.

Figure 9-47. Vector Difference Caused by Causality Enforcement



The actual S21 phase data (with and without causality enforcement) is plotted here for the example. The vector difference in S21 was shown in Figure 9-47.

Figure 9-48. Actual S21 Phase Data With and Without Causality Enforcement

The enforcement process starts by coercing the imaginary part to have the desired relationship to the real part (there is some flexibility on what is a 'fixed' property that we will return to). To have zero energy before t=0, the time domain (impulse) function must have an odd part that will cancel the even part correctly:

$$g_{odd}(t) = \mathrm{sgn}(t) \bullet g_{even}(t)$$

So

$$g(t) = g_{even}(t) + \mathrm{sgn}(t) \bullet g_{even}(t)$$
$$G(f) = F\{g_{even}(t)\} + F\{\mathrm{sgn}(t) \bullet g_{even}(t)\}$$
$$G(f) = G_{even}(f) - jH_{even}(f)$$

Equation 9-2.

Where H_{even} denotes the Hilbert transform has been applied (the Fourier transform of a time function with the sign function applied is a working definition).

Now, another conclusion of causality is that the negative frequency components of the response are the conjugates of the corresponding positive frequency components. This implies that the real part of the frequency response is exactly the even portion. Thus the enforced result is simply related to the Hilbert transform of the real part of the data.

From an energy conversation point-of-view, it may be preferable to keep the magnitude of the parameter fixed instead of the real part and this is presented as the default configuration. This fixing of the magnitude is easy to accomplish since we can simply scale the result above by a real function (normalizing magnitude).

As in passivity enforcement, the coercion process does change the data so if the changes are large, it is generally advised to look into possible measurement execution issues (use a VNA with wider bandwidth, calibration components or method choice, cable drift, connection repeatability, etc.) to see if the measurement could be improved instead.

9-8 Summary

A series of techniques have been presented for handling and studying the problem of non-insertable DUTs. Adapter removal is a 2-calibration technique for removing the effects of an adapter from a given calibration setup (e.g., when the DUT has one coax port and one waveguide port). Network extraction is somewhat more separable in that it tries to extract the S-parameters of the complicating adapter/fixture so that it can be de-embedded later. Seven different types of extraction (with a number of sub-types with different standards or with localization techniques) were presented with various trade-offs in calibration complexity, simplicity, and uncertainty. A modeling-based extraction technique using sequential localization or peeling was also presented.

Although not specifically part of network extraction, sometimes trace math (Data, Memory Math on the Display | View Trace menu) is used for normalization and other forms of pseudo-de-embedding. The ability to save memory data is covered elsewhere in this guide, but an additional tool to help is the ability to save math-modified data to .sNp files. This selection is available in the sNp Setup dialog (see System | Setup | Misc Setup | SNP Files Setup). When Save Trace Math is selected and one or more traces have trace math turned on, the modified data for the parameters with math applied will be saved to the .sNp file. This can be useful for saving normalized results. If trace math is not applied for every parameter (which would only happen for a .s4p file if 16 traces were active and all had trace math applied and each trace had a different S-parameter), the unmodified data for the uncovered parameters will be saved instead.

Chapter 10 — Other Calibration Procedures

10-1 Overview of Other Calibration Procedures

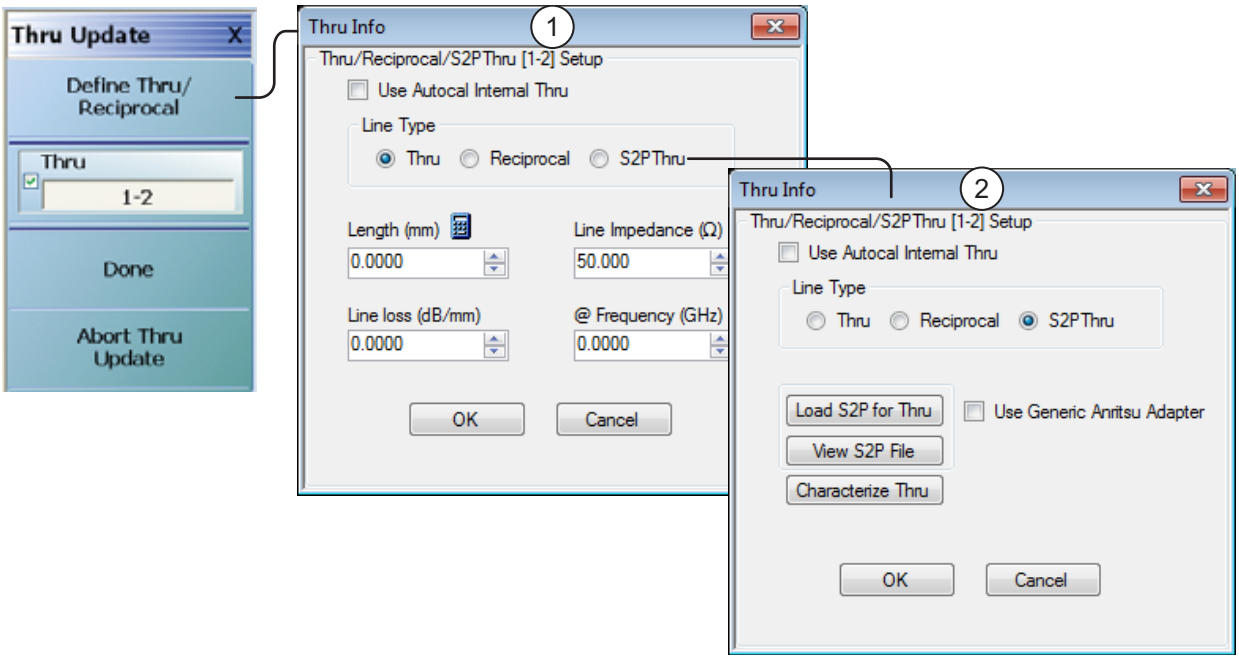
This chapter discusses other calibration procedures not covered in the previous material. These include Thru Update, Cal Merge, Hybrid Cal, Flexible Calibration, Display Cal Data, Display Cal Data, and Secondary Match Correction.

10-2 Through (Thru) Update

A common question related to calibrations is how often the calibration must be redone. The frequency of re-calibration depends on the environment (both in terms of temperature stability and in terms of the cable/fixture construct that is being used). A calibration lifetime is often limited by the stability of the test port cables through drift or motion. Drift defects directly affect transmission tracking and load match, so if those terms could be easily refreshed, the time before the entire calibration would have to be redone could be extended. This is the concept of a Thru Update.

The idea is to connect just a thru line and quickly refresh the transmission tracking and load match terms without great effort. A Thru Update is a one-step calibration that can be used to refresh a current full 2-port or 1 path 2-port calibration.

Like the thru step in other calibrations, the length of the line being used as well as its loss can be specified to ensure minimal disruption of the reference planes (see Figure 10-1). The entry methods for these parameters are the same as in the normal calibration procedures.



1. Thru Line selected allows user entries for length, line impedance, line loss and frequency.
2. S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 10-1. THRU UPDATE Setup Menu and THRU INFO Dialog

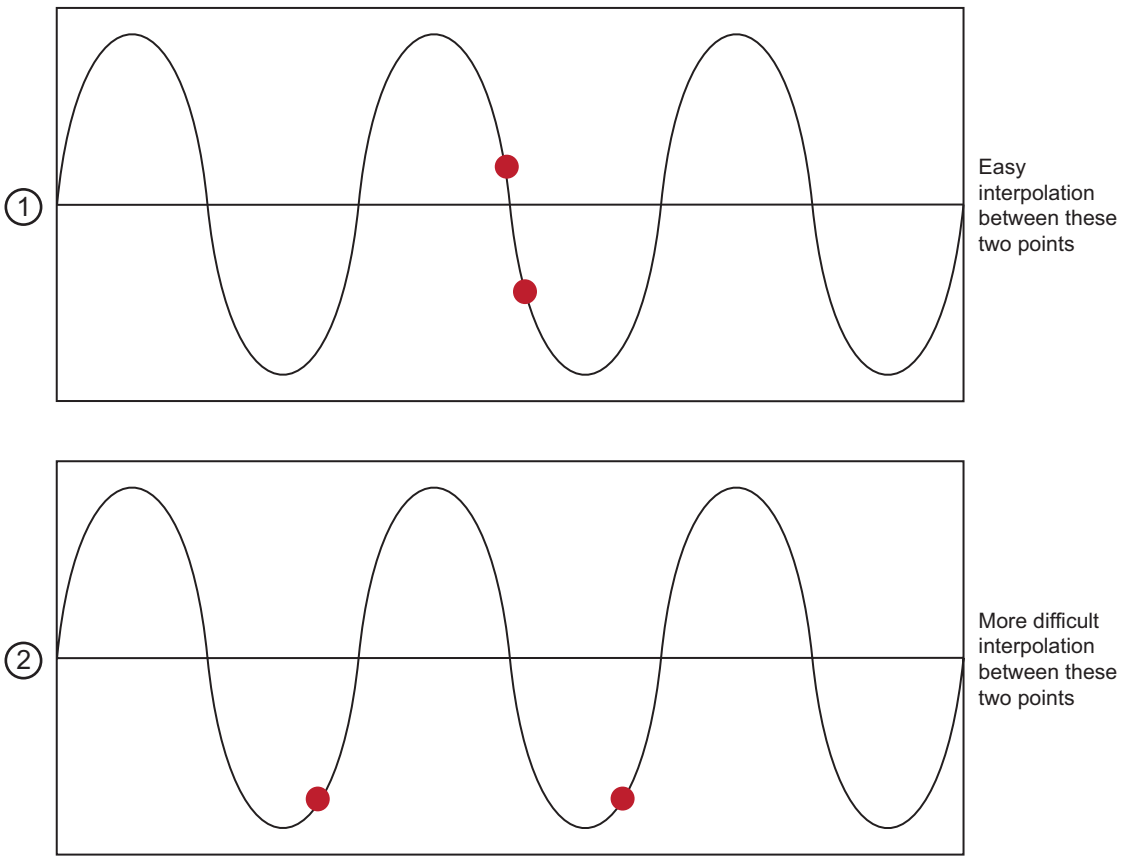
Also, as with the manual calibrations, a .s2p file describing the thru S-parameters can also be loaded and will be used during the update. Depending on the calibration, the reflection parameters in the .s2p file may or may not be used ('not' for a transmission frequency response calibration). A *characterize thru* option is available to calculate the thru's .s2p file using two previously stored reflection calibrations (one with and one without the thru present).

The thru path of an AutoCal module (see Chapter 2 of this guide) can also be used for thru update and this can be useful at times if an adapter is not readily available and the AutoCal module has an appropriate connector combination. Only reciprocal processing is available when using this path for thru update as the loss levels are non-negligible.

10-3 Interpolation

Typically, calibration is done for a specific list of frequencies and then measurements are made over that same list of frequencies. While this is most accurate, it is not necessarily convenient. If, for example, one is measuring a variety of narrow bandpass filters of different center frequencies, it may be useful to be able to zoom in to look at the passband of each filter without re-calibrating. Interpolated calibrations are allowed for purposes like this one. The theory is that the error coefficients are all carefully interpolated between calibration points to minimize possible error.

To see the cause of error, note that the cable runs within the instrument and the cables that the user provides typically result in a large electrical length. Thus the error coefficient magnitude versus frequency is often periodic in shape. If the interpolation is not performed with care, large errors can result (see Figure 10-2).



Step size, interpolation effects, and possible errors.

1. Easy interpolation between these two points.
2. More difficult interpolation between these two points.

Figure 10-2. Effect of Step Size on Interpolation

As a rule, the smaller the step size used during the calibration, the more successful the interpolation will be. It is desirable to keep the step size smaller than the ripple period of the coefficients which will typically range from 50 MHz to 500 MHz. The smaller number is for setups with very long test port cables and fully optioned systems; the larger numbers for setups with short test port cables and with Options 51, 61, and 62. The menu to select interpolation is shown in Figure 10-3.



Figure 10-3. CAL OPTIONS Menu

The calibration interpolation menu will default to OFF where points used during measurement must correspond to calibration frequency points. When interpolation is ON, other points may be used. In neither case may frequencies outside of the calibration frequency range be used.

To gain a better understanding of the uncertainty implications of interpolation, it may be helpful to consider how the error terms of a typical calibration vary over frequency. A large scale and zoomed-in version of one parameter (source match) is shown in Figure 10-4. Even with relatively low point densities, the large scale variation in the large-scale plot would be captured by interpolation. The fine scale variation shown in the zoomed-in plot is, however, another matter.

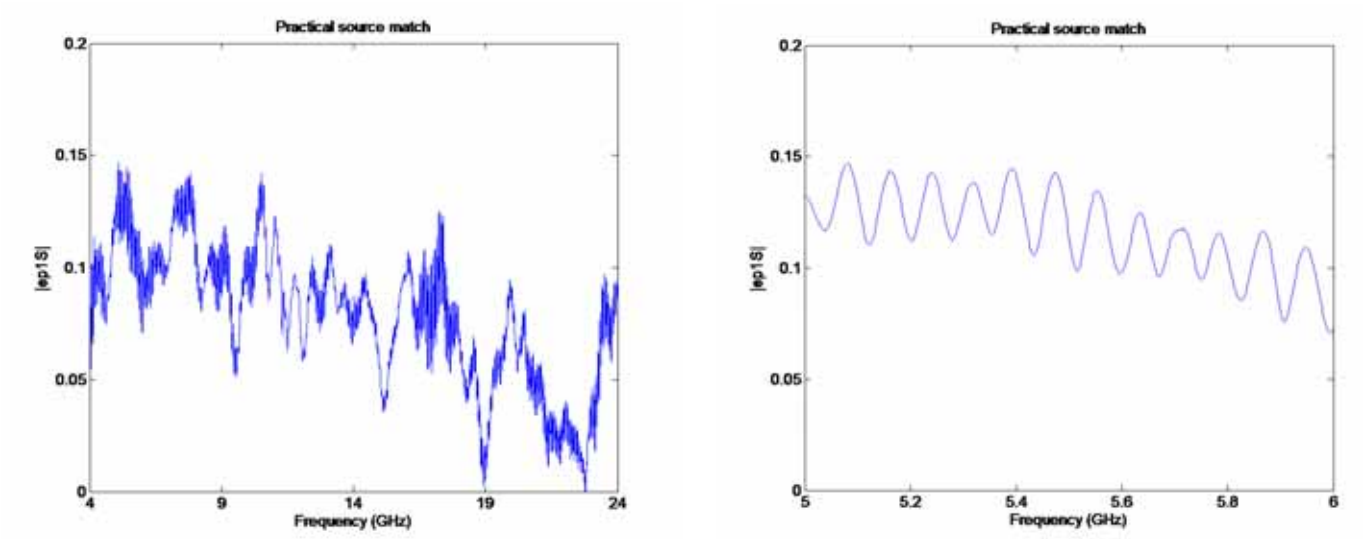
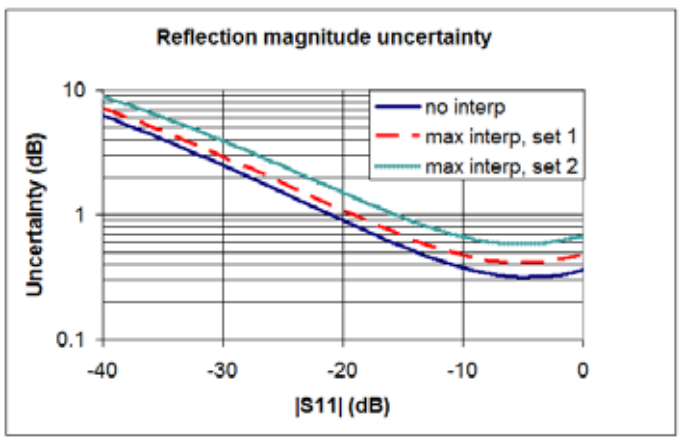


Figure 10-4. Large Scale and Zoomed-in Source Match Variation Over Frequency.

Now, consider two different sets of hardware, both with a nominal tracking coefficient of unity (for simplicity). At lower frequencies, set 1 may be readily achieved, but at millimeter wave frequencies, set 2 may be more practical. The important point is that the interpolation effects are setup-dependent.

- Set 1: Raw directivity ~ 0.02 (–34 dB) and raw source match ~0.1 (–20 dB)
- Set 2: Raw directivity ~ 0.10 (–20 dB) and raw source match ~0.3 (–10 dB)

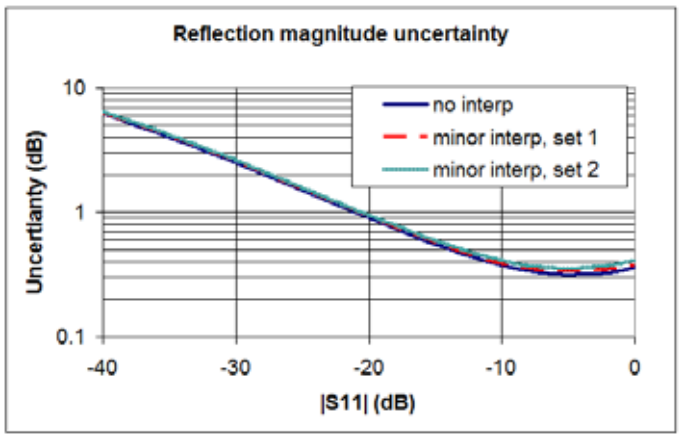
If one calculated the uncertainties for no interpolation and for worst-case interpolation for both of these sets, one would get results like that shown in Figure 10-5 (for a standard SOLT coaxial calibration).



The uncertainty penalties possible with worst-case interpolation are shown here for some example setups.

Figure 10-5. uncertainty Penalties Possible with Worst-case Interpolation

As suggested above, one can get around this problem by increasing the frequency point density. In Figure 10-6, the point count was increased so that there were 10 calibration points per period of the ripple (a proxy for the electrical length of the setup, much like that discussed in Chapter 9).



An increase in frequency point density can reduce interpolation effects in all setups, as would be expected.

Figure 10-6. Increase in Frequency Point Density Can Reduce Interpolation Effects

10-4 Cal Merge

When multiple calibration algorithms are needed (covering different frequency ranges) and the usual calibration approaches are not adequate, any calibrations of the same type may be combined. Historically, a common example was the broadband coaxial W1 calibration where SOLT is used at low frequencies and SSST is used at high frequencies. In more recent software versions (2.3.1 and later), that particular case is handled with an integrated calibration for the 3739-based broadband modes but there are still a number of examples where a particular calibration algorithm may not be appropriate over the entire frequency range of interest. Some examples include:

- LRL/LRM calibrations where more than 5 bands are of interest.
- The use of a defined standards calibration (e.g., SOLT) at low frequency and transitioning to LRM at high frequency (an on-wafer scenario that is sometimes employed).
- A broadband calibration where it may be necessary to transition between a number of SSST or SSLT kits.
- A 3738-based broadband SOLT/SSST calibration in the 1 mm coaxial connector.

This starts off as merging a pair of calibrations but the operation may be repeated any number of times. The Cal Merge utility dialog (shown in Figure 10-7 on page 10-7) simply combines the error coefficient arrays of the various calibrations while rationalizing the frequency lists as best as possible. Some points to keep in mind:

- **Frequencies**

  If all frequencies are distinct, all points will be maintained.

- **Frequency Points**

  If any frequency points are in both calibrations, the coefficients from the first calibration (see Figure 10-7) will be used.

- **Number of Points**

  The combined number of frequency points may not exceed the current limit of the instrument which is either 25,000 or 100,000 points depending on the VNA mode.

- **Calibration Small Steps**

  Since different calibration algorithms can be used, small steps may appear in the data where the calibrations overlap (assuming the cals were all of good quality).

- **Segmented Sweep**

  The combined calibration will have its frequency list and other setup information loaded in a segmented sweep structure but it is not editable.

In the case of waveguide calibrations, an overriding cutoff frequency should be specified (for reference plane rotations and other dispersion processing events). When the merge is complete, the combined calibration will be resident in memory but will not automatically be saved. The user must save the combined calibration manually if desired.

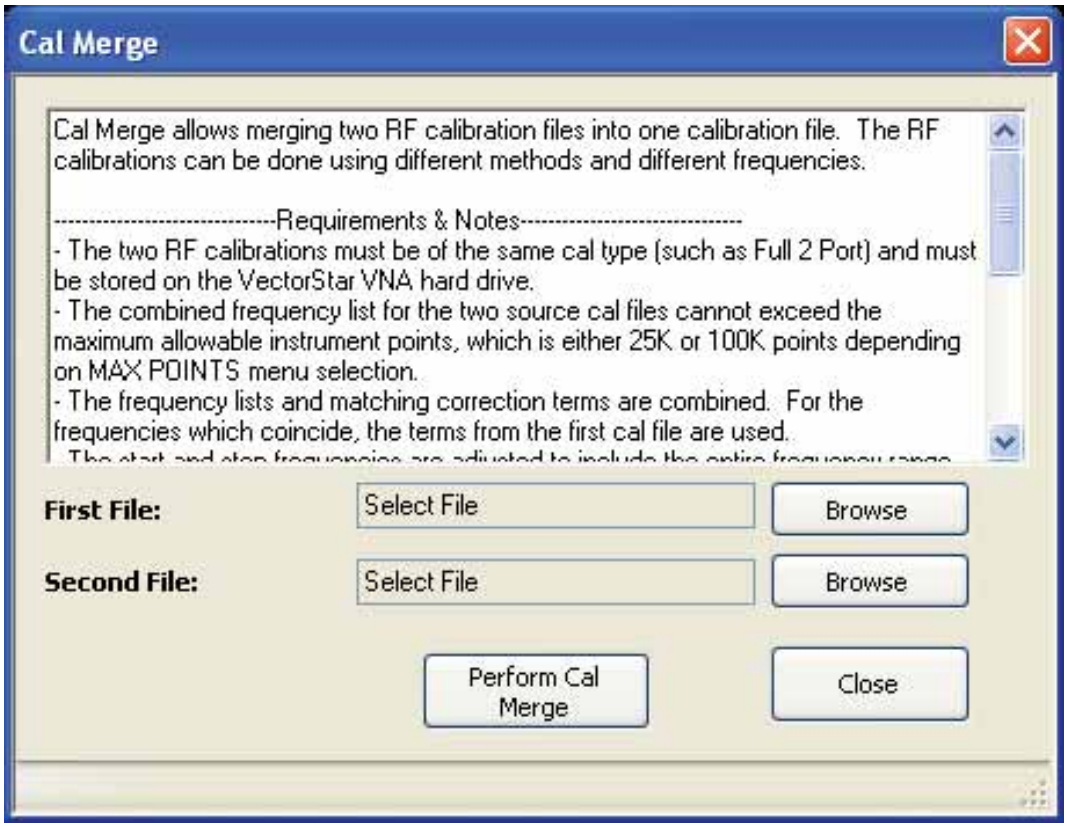


Figure 10-7. CAL MERGE Dialog Box

10-5 Hybrid Calibrations

The hybrid calibration is a method of taking a pair of distinct 1-port calibrations, together with some additional measurements, to create a new full 2-port calibration. The "*hybrid*" part of the definition comes in that the two 1-port calibrations may be with completely different connector types, media types, and/or cal algorithms. One example may be a case where it is desired to have Port 1 in coax but Port 2 in waveguide. This is conceptually similar to some of the adapter removal and network extraction discussions in Chapter 9, but here a little more flexibility is offered on the calibration side. In particular, the through completion step may be a reciprocal device only (thus bringing in SOLR concepts).

Generally the two 1-port cals are performed in advance and the setups saved, typically as an active channel CHX file type. One enters the dialog below and specifies those cal files. Then the "*thru*" between the two ports is specified using the usual THRU INFO dialog format. In the case of a physical thru, length and loss information is specified. In the case of a reciprocal, a length estimate is requested to simplify root choice.

Also as with the manual calibrations, a .s2p file describing the thru S-parameters can also be loaded and will be used during the hybrid cal. A *characterize thru* option is available to calculate the thru's s2p file using two previously stored reflection calibrations (one with and one without the thru present).

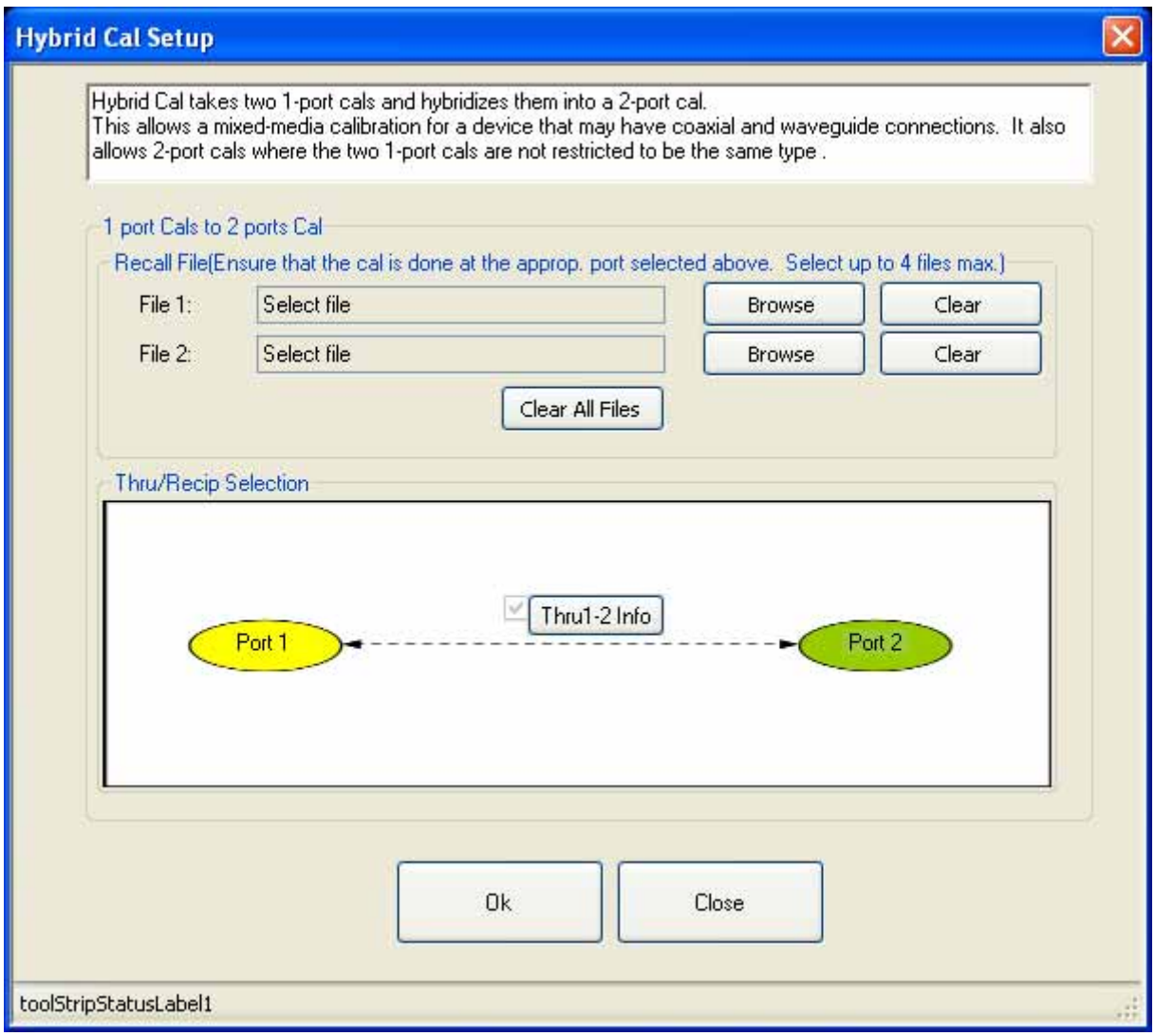
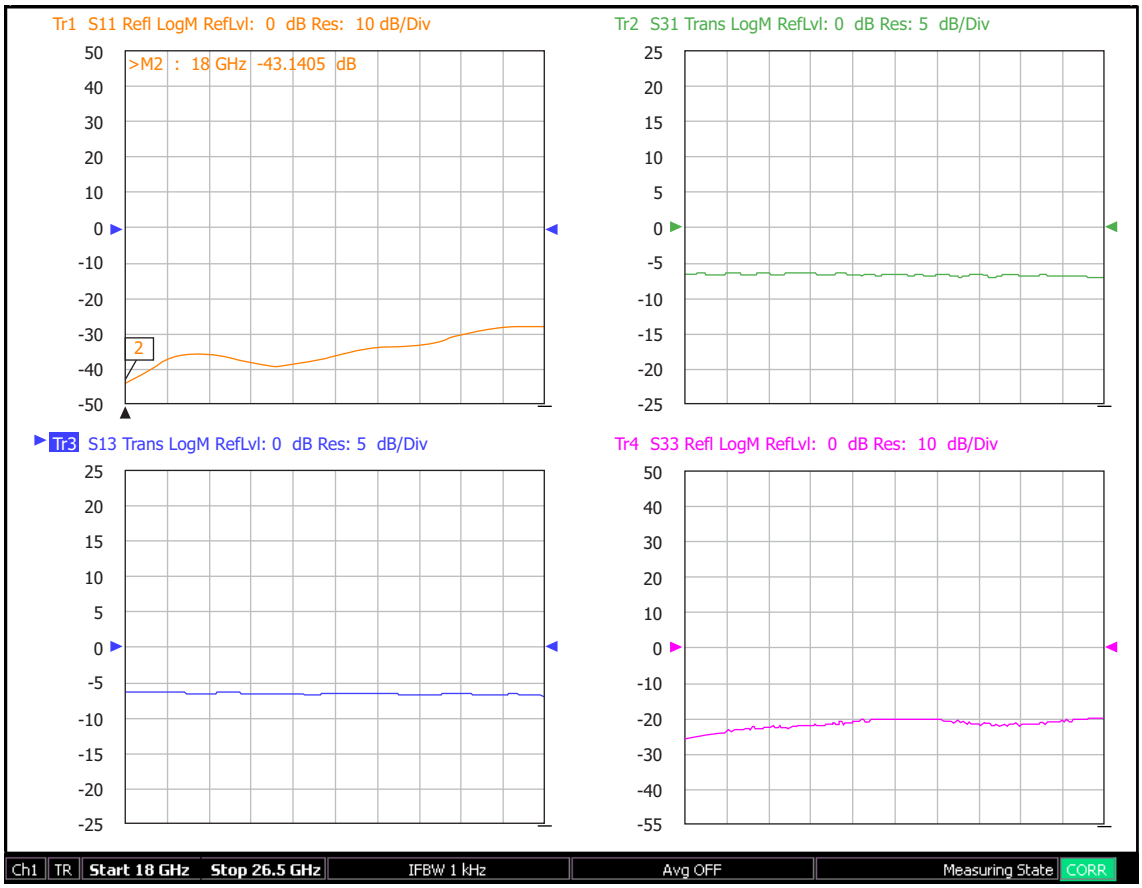


Figure 10-8. HYBRID CAL SETUP Dialog Box

As with the adapter removal and network extraction procedures, the cal files used here must be based on the same frequency lists (same frequency range and same number of points at least). In this case, the files must both describe full 1-port calibrations.

As an example of the mixed-media case where this type of hybrid calibration is helpful, consider a desired coaxial-waveguide combination reference plane. One could perform adapter removal processing as discussed in Chapter 9, but there are occasions where in one of the media planes, one can only perform one port calibrations (due to physical arrangement of the hardware, calibration kit availability, or other reasons). For this example (in WR-42 waveguide), a 1-port K calibration was done for one file and a 1-port WR-42 waveguide calibration was done for the other file. These two files were combined using the hybrid calibration dialog and the resulting calibration was used to measure the hybrid device of Figure 10-9. Note that this figure refers to a port 1-3 calibration associated with multiport measurements (see Chapter 24), but the algorithm is exactly the same for 2-port systems.



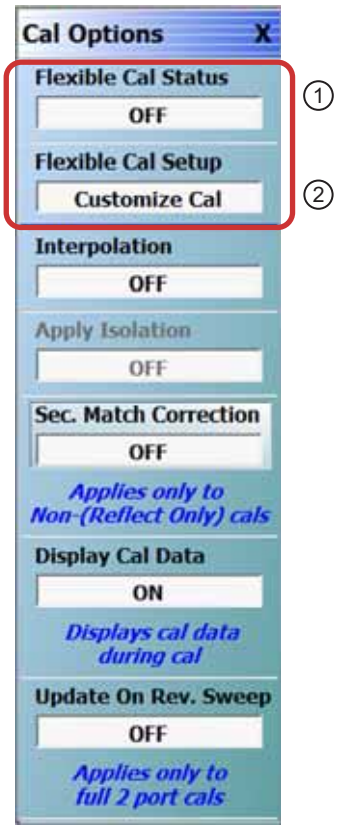
An example hybrid waveguide-coaxial measurement is shown here that was enabled by the hybrid calibration dialog.

Figure 10-9. Example Hybrid Waveguide-coaxial Measurement

10-6 Flexible Calibration

Another selection under the CAL OPTIONS menu is a utility called Flexible Cal. The purpose of this utility is to speed measurements if the full correction capability of a given calibration is not needed and it is desired to not perform and recall a simpler calibration (to simplify a manufacturing test situation for example). For example, one may have performed a full 2-port cal but then, for a period of time, need only measure S_{11} for a device with little transmission to Port 2. If one stayed with the parent cal, two sweeps would always be performed (Port 1 driving and Port 2 driving) even though only one is needed for the measurement. Flexible cal allows one to temporarily and quickly suspend the unneeded measurements to speed up the process.

This utility is only available when a full 2-port cal has been performed (otherwise a subset makes little sense). If a lower order calibration has been performed, the feature will be labeled as Not Available. If a full 2-port cal exists, the CAL OPTIONS menu will appear with the Flexible Cal Terms button highlighted. The buttons will only be available if a full 2-port cal exists.



1. Flexible Cal Status Button—Only available when a full 2-port calibration exists.	2. Flexible Cal Setup Toggle Button—Only available when a full 2-port calibration exists.

Figure 10-10. CAL OPTIONS Menu—Flexible Cal Status Button Highlighted

The Flexible Cal Setup button will bring up a dialog like that shown in FLEXIBLE CAL SETUP Dialog Box.

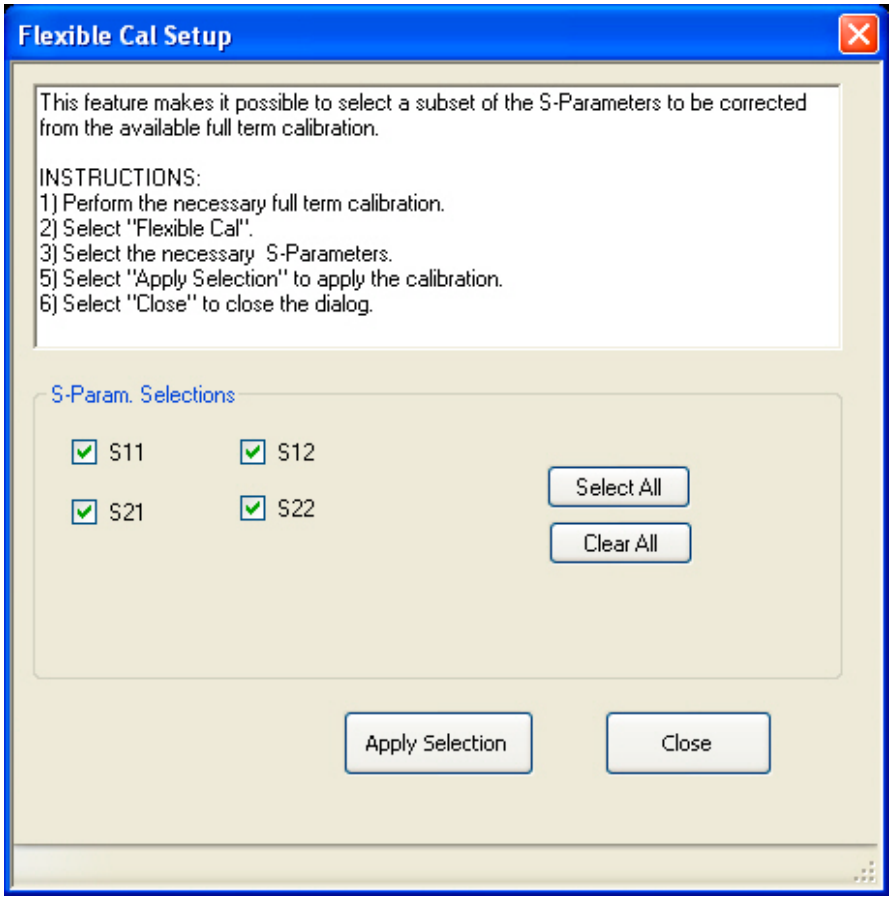


Figure 10-11. FLEXIBLE CAL SETUP Dialog Box—Full Term Cal

Instructions

Perform the necessary full term calibration. On the CAP OPTIONS menu, select Flexible Cal Setup. Select which S-parameters you wish to be measured. Unneeded sweeps will not be performed even though the parent calibration is a full 2-port (for example, if S_{11} and S_{21} only are selected, only the Port 1-driving measurements will be performed). The highest level calibration possible based on the selected parameters will be performed. When ready, select Apply Selection then Close to close the dialog.

Examples:

- S_{11} only is selected: execute as a full 1-port cal.
- S_{22} and S_{12} only are selected: execute as a reverse 1-path–2-port cal.
- S_{12} and S_{21} only are selected: execute as a transmission frequency response cal.
- All four are selected: execute as the parent full 2-port cal.

10-7 Display Cal Data and the Real-Time Calibration Display

When calibrating in an on-wafer or fixtured environment (and sometimes even in connectorized environments), it is sometimes difficult to tell if one has made good contact with the standard. Rather than completing the entire calibration and then seeing if the result was acceptable, it may be useful to get a quick indication of what is being measured, in a quasi-corrected sense, before moving on to the next standard.

The Display Cal Data function is accessible from the CAL OPTIONS menu.

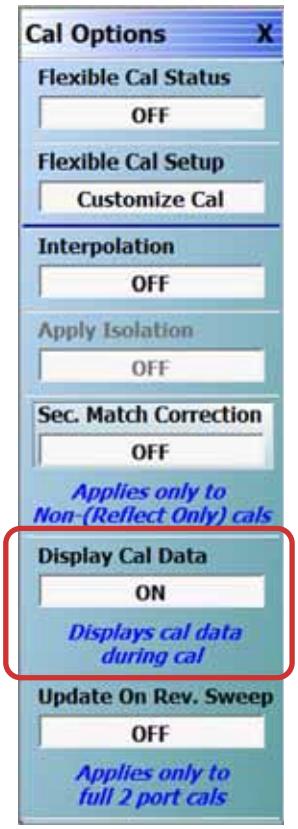


Figure 10-12. Display Cal Data in CAL OPTIONS Menu

When turned on (default is on), Display Cal Data enables the use of a previous calibration to provide a real-time quasi-corrected view of the current standard being measured. After one clicks the standard button (during which the real data for the new calibration is collected), the display reverts to one where the swept data is corrected by the previous calibration in the system. Assuming the previous calibration was good, a load standard should then appear to have high return loss if connected properly, a thru should have low insertion loss and high return loss, etc. Some things to keep in mind:

- The display correction will only be applied if the previous calibration is of the same type as the current calibration (e.g., full 2-port…) and consists of the same frequency list or a superset of the current frequency list. If both of these statements are not true, then the data displayed will only be corrected by a factory RF calibration (should it exist).
- The previous calibration correction is not applied to the data being acquired for the new calibration directly. That is, this is not a second tier calibration; it is the newly acquired raw data that is fed into the calibration algorithm when Done is selected.
- If the previous calibration was poor (e.g., a standard was not connected at some point), the displayed values during the current calibration may appear erroneous. The Display Cal Data should be turned off if the previous calibration is known to be bad. If Display Cal Data is left on in this case, however, it will not affect the result of the new calibration. Only the appearance of the value during the current calibration will be affected.

10-8 Secondary Match Correction

Secondary match correction (SMC) is about improving measurement performance and reducing uncertainties mainly for measurements of very low insertion loss devices. At the frequency of interest, if the DUT loss is more than a few dB, this function will usually not have significant impact but it can offer improvements when DUT losses are smaller than that. Since the function slows down sweep updating, it is turned off by default. The function can be employed after any calibration (excluding reflection-only calibrations) is applied. This section will explore the SMC process and how it can be useful in a number of measurement applications.

Suppose one is measuring a very low loss device (a precision adapter for example). One may get an insertion loss plot like the below when using a precision calibration kit in good condition. One may wonder about the ripple in the data in Figure 10-13. It is only about 0.05 dB peak at worst which is well below the measurement uncertainty of ~0.1 dB to 0.12 dB (peak) for this particular measurement but it still may be undesirable. One may notice that residual error terms are on the order of 30 dB to 35 dB and the DUT match may be on the order of 25 dB so one could even rationalize that multiple reflections between those (effective) interfaces may explain the result.

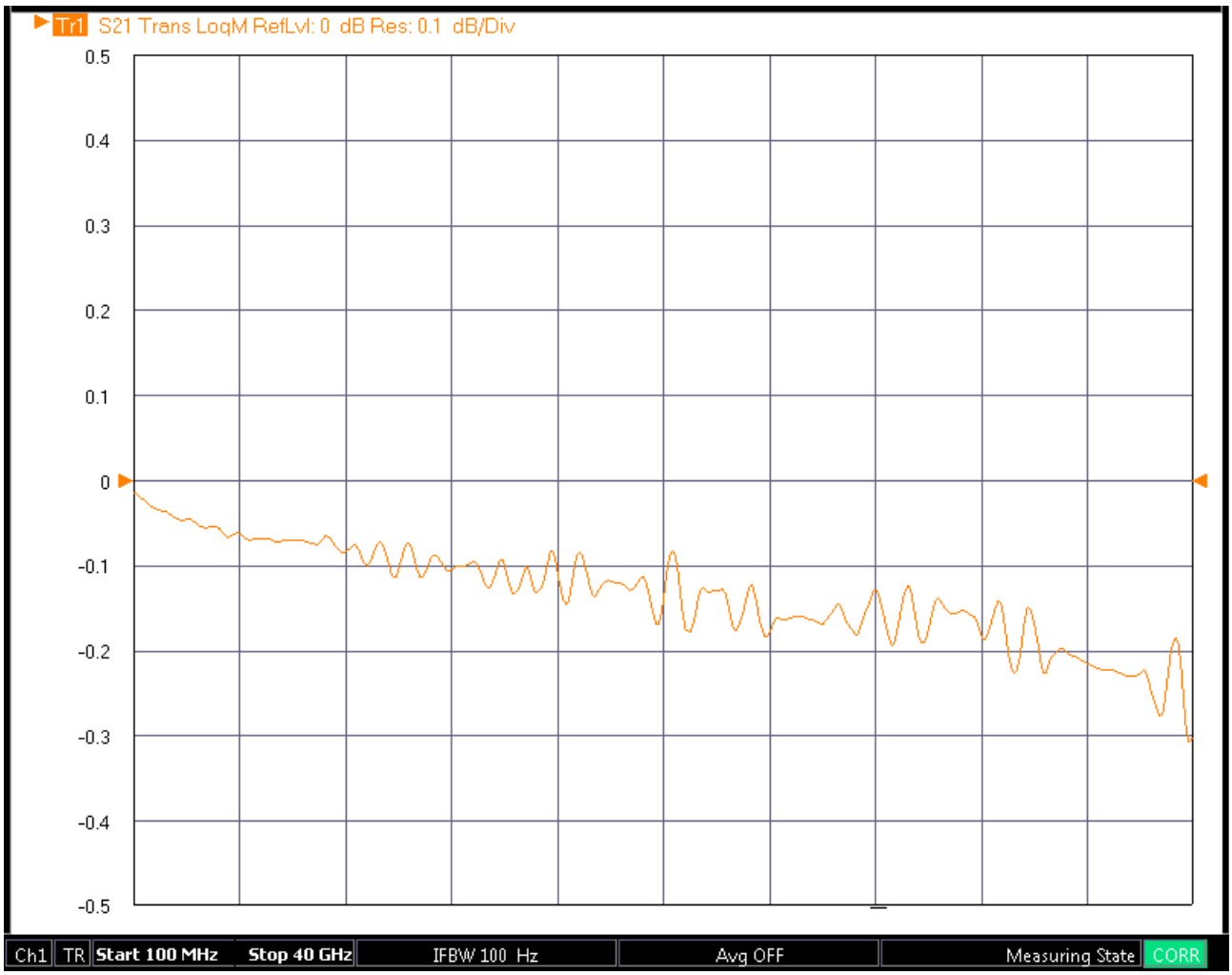
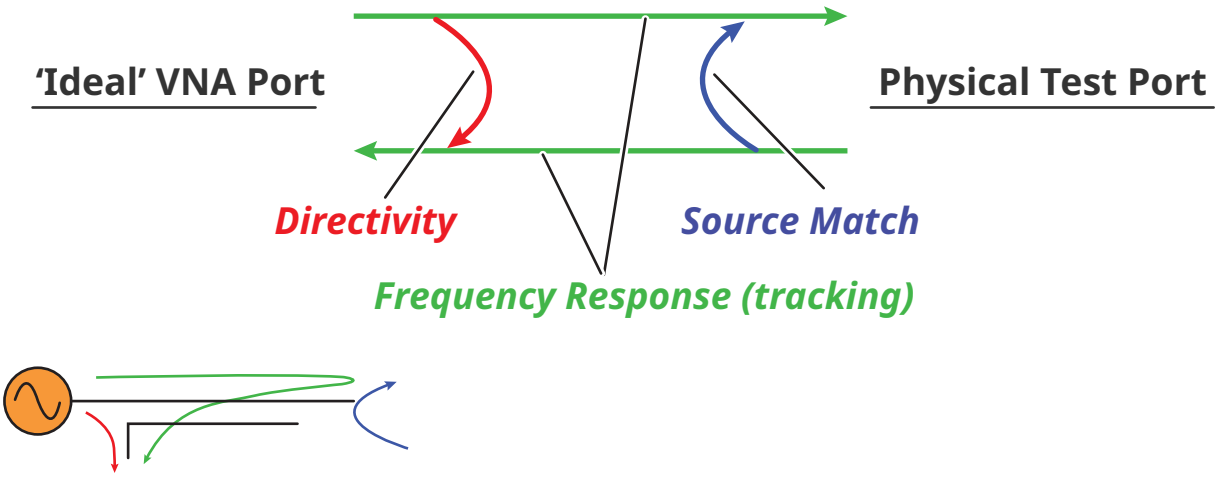


Figure 10-13. Example Low Insertion Loss Measurement

Some of this ripple can arise from high DUT reflections, from pin depth issues in the mating reference planes or with the DUT (or with the calibration kit!), or there may be other explanations. One source of that ripple, however, can be a residual effect related to how match is corrected in the basic VNA calibration. In other chapters of this guide, the basic VNA error model was discussed and part of it is the simple reflectometer structure which is shown in Figure 10-14.



The simple reflectometer error model is sketched here along with the mapped effects to a physical reflectometer.

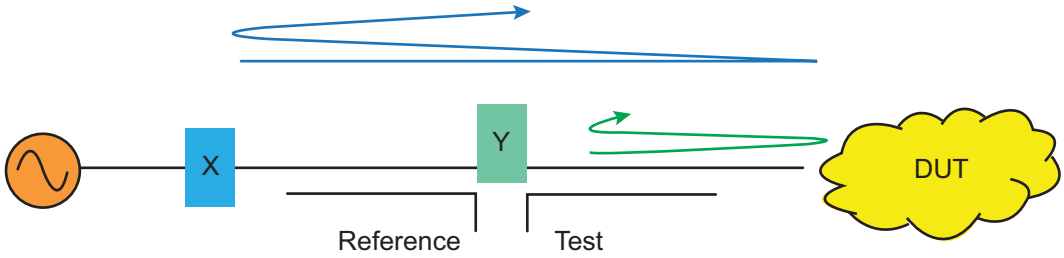
Figure 10-14. Reflectometer Error Model

The error model arose from actual defects that occur in a physical reflectometer: finite directivity of a coupling structure, finite match of the coupler and nearby components, and a non-flat frequency response of the signal chain. Of particular interest to this discussion is match and, like with all models, how it is treated here is something of a simplification. Consider a slightly more complete model diagram in Figure 10-15 on page 10-15. Suppose the dominant source of actual mismatch is at position Y in the figure. When the signal propagates from the source, some energy is reflected back to the source and some is transmitted. Of the portion transmitted, some reflects off of the DUT and then re-reflects off of Y. The product of those reflections (~eps*S11 where eps is the source match term in the calibration terminology discussed earlier in this guide and S11 is the reflection coefficient of the DUT) is important since the product repeats on multiple re-reflections forming a geometric series. Thus when one looks at the basic reflectometer equation, the tell-tale result of an infinite geometric series is present in the denominator.

$$S_{11}^{m} = ed + \frac{et \cdot S_{11}^{act}}{1 - eps \cdot S_{11}^{act}}$$

Equation 10-1. Basic Reflectometer Equation

Here $S11^{act}$ is the actual reflection coefficient of the DUT and the 'm' superscript on the left side denotes 'measured'. Port 1 was used for this equation but any port could have been employed. In the case of Y being dominant, everything works out since the test coupler sees all of the multiple reflections and the actual measurement is very close to what the simple error model would arrive at.



An expanded sketch of the reflectometer is illustrated here.

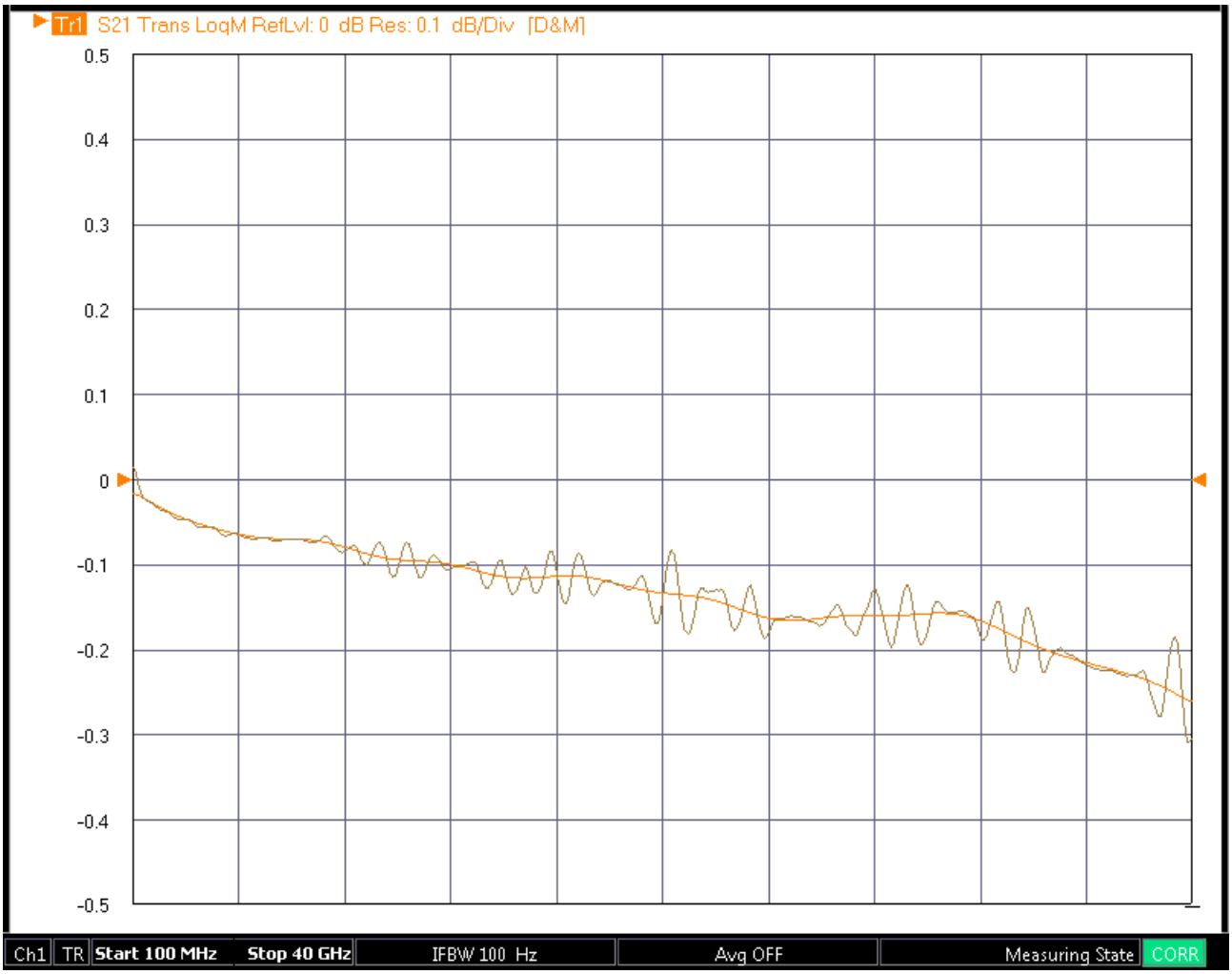
Figure 10-15. Reflectometer Expanded Model

Something different can happen if a significant reflection occurs at X. Now the reference coupler sees the mismatch as does the test coupler so the ratio that forms S11, test/reference, sees a distorted picture of the reflection product and the reflectometer equation is really not as simple as that shown above. The test coupler will indeed see the effective series of reflections but the reference coupler sees some of it as well and the ratioing produces a partial cancellation or amplification depending on phasing.

The significance of these X and Y reflection locations can vary greatly from setup to setup but there are cases where the X location is non-negligible. It should be emphasized that these effects are smaller than quoted uncertainties as the uncertainty component evaluation process is also affected by the above mechanism. It should also be mentioned that this represents a fairly small perturbation on the effective port match so would not be visible on higher DUT loss measurements (more than a few dB) and, even for low loss DUTs, is generally only visible on a fine scale.

One could add additional calibration standards to solve for this more elaborate match model but this would lengthen the calibration process. One can also use the phase information in the calibration residuals to localize where the mismatch elements are. This is the principle behind secondary match correction: use the residual phase information to process a 2^{nd} tier correction that primarily impacts the match terms.

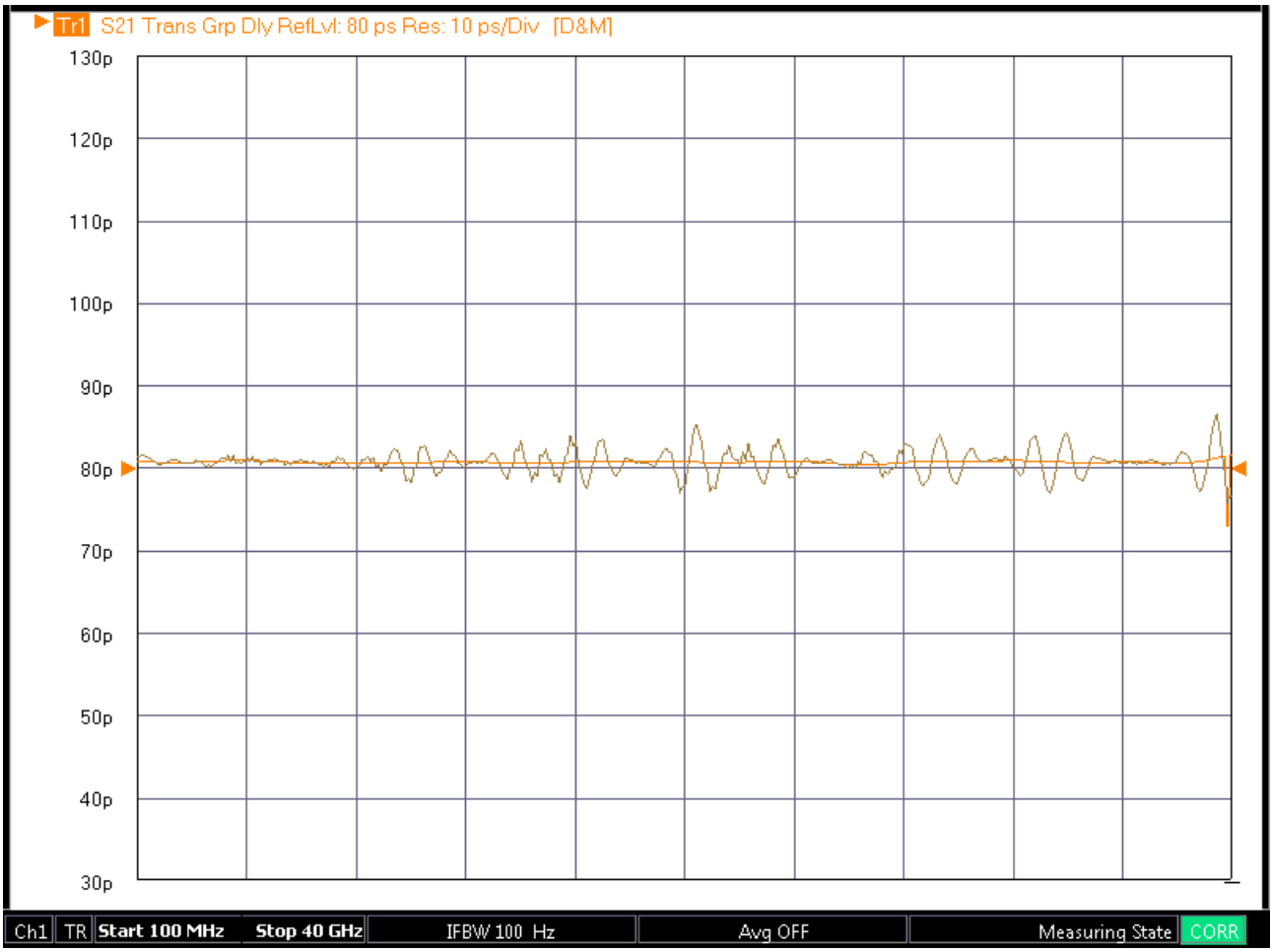
Applying this to our previous example of the adapter measurement, one can see a significant reduction in the ripple (Figure 10-16 on page 10-16) that was not part of the DUT behavior after all (lighter trace is with SMC applied).



An adapter insertion loss measurement without SMC (darker trace) and with SMC (lighter trace) is shown here.

Figure 10-16. Adapter Insertion Loss Measurement—Without/With SMC

A similar effect can be seen in the group delay measurement of the adapter (see Figure 10-17 on page 10-17). Although the effect is only a few ps of delay, it may be important in some applications such as careful delay-matching exercises or fine modulation distortion calculations based on measurements of a transceiver's group delay.



A group delay measurement of an adapter without SMC (darker trace) and with SMC (lighter trace).

Figure 10-17. Adapter Group Delay Measurement—Without/With SMC

Using secondary match correction only requires turning the selection ON under the Cal Options Menu as shown in Figure 10-18. Whenever an appropriate calibration is applied, then SMC will also be applied to the parameters. The appropriate calibration types include full 2-, 3-, and 4- port calibrations, transmission frequency response calibrations, and 1-path 2-port calibrations. SMC will not be applied to 1-port (reflection only) or reflection response calibrations. Since this is part of the correction engine, user-defined parameters (b2/1, b2/a1...) are not affected.

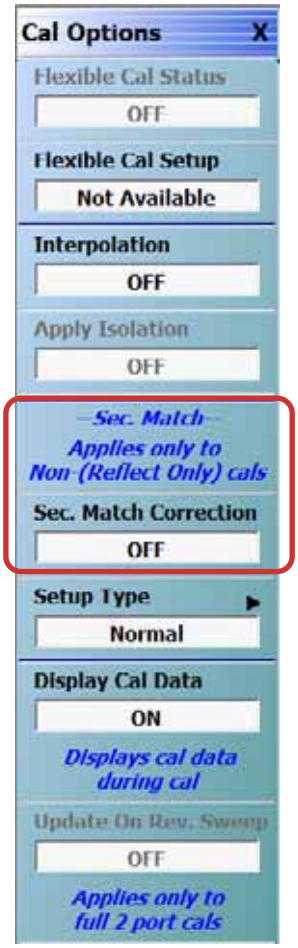


Figure 10-18. Cal Options Menu Showing Secondary Match Correction Button

There are few things to keep in mind about the behavior of this correction:

- If the frequency list is such that the phase residuals provide little information, then the correction will not be applied. This will happen if the step size is very large (~>1 GHz will limit the method's value) or if the frequency range is extremely small (~< 2 GHz will limit the value). In segmented sweeps with very irregular steps (step sizes deviating more than about 2% from the mean step size in the frequency list), the correction will not be applied.
- This method works by identifying residual mismatch centers that are not part of the DUT and further suppressing their contribution. If the DUT is very close to the port or if the DUT is very long electrically, this process can be compromised so a 'setup type' selection is provided to inform the system of these situations. The 'direct connect' choice is for devices connected directly to the test port (or very close if the sweep range is small) as the distance to the residual reflection centers can be very small in those cases. The 'electrically long' case is for extended devices (very long cables or, in particular, long fiber lengths) so the system knows where to search for extraneous selections.
- If SMC is ON and one saves .sNp files, the SMC correction will be applied to relevant parameters. SMC will always flow through to .txt and .csv files.

In summary, the secondary match correction process allows one to optimize measurements of low insertion loss devices by correcting for a simplification made in the standard error model. The improvements are usually on the scale of hundredths of a dB in insertion loss and picoseconds in group delay but, particularly for low loss adapter and fixture characterization, those enhancements can be valuable.

Chapter 11 — Calibration and Measurement Enhancements

11-1 Chapter Overview

This chapter provides a description of functions that provide additional calibration, post-processing, and display options that increase the usefulness of the instrument data. The topics described include: embedding/de-embedding, reference-plane control and modification, and impedance transformations.

A number of functions are provided beyond the basic calibration and display tools to help post-process the data in a way that is useful. The topics described relate to virtually modifying the environment in which the DUT resides.

These topics include:

- Embedding/De-embedding (E/DE)

  This is the virtual removal or insertion of networks or circuits around a DUT that may represent fixtures, launching structures, tuning elements, or other items. E/DE sub-topics are:

- Reference Plane Control

  This can be thought of as a simpler subset of de-embedding in which transmission line lengths and loss are removed from the measured data. Reference Plane sub-topics are:

- Impedance Transformation

  When calibration components are not available in impedances other than 50 ohms, it is possible to view the data as if the VNA had been calibrated in some other impedance.

In addition, there are some clerical tasks to describe including the order of virtual operations and some conversions to other parameter formats (impedances and admittances for example). The measurements menu that contains the majority of these functions is shown in Figure 11-10, “REFERENCE PLANE Control Menu” on page 11-12. Parameter conversions are a per-trace function (as opposed to the others which are per channel) and is listed under the DISPLAY menu.

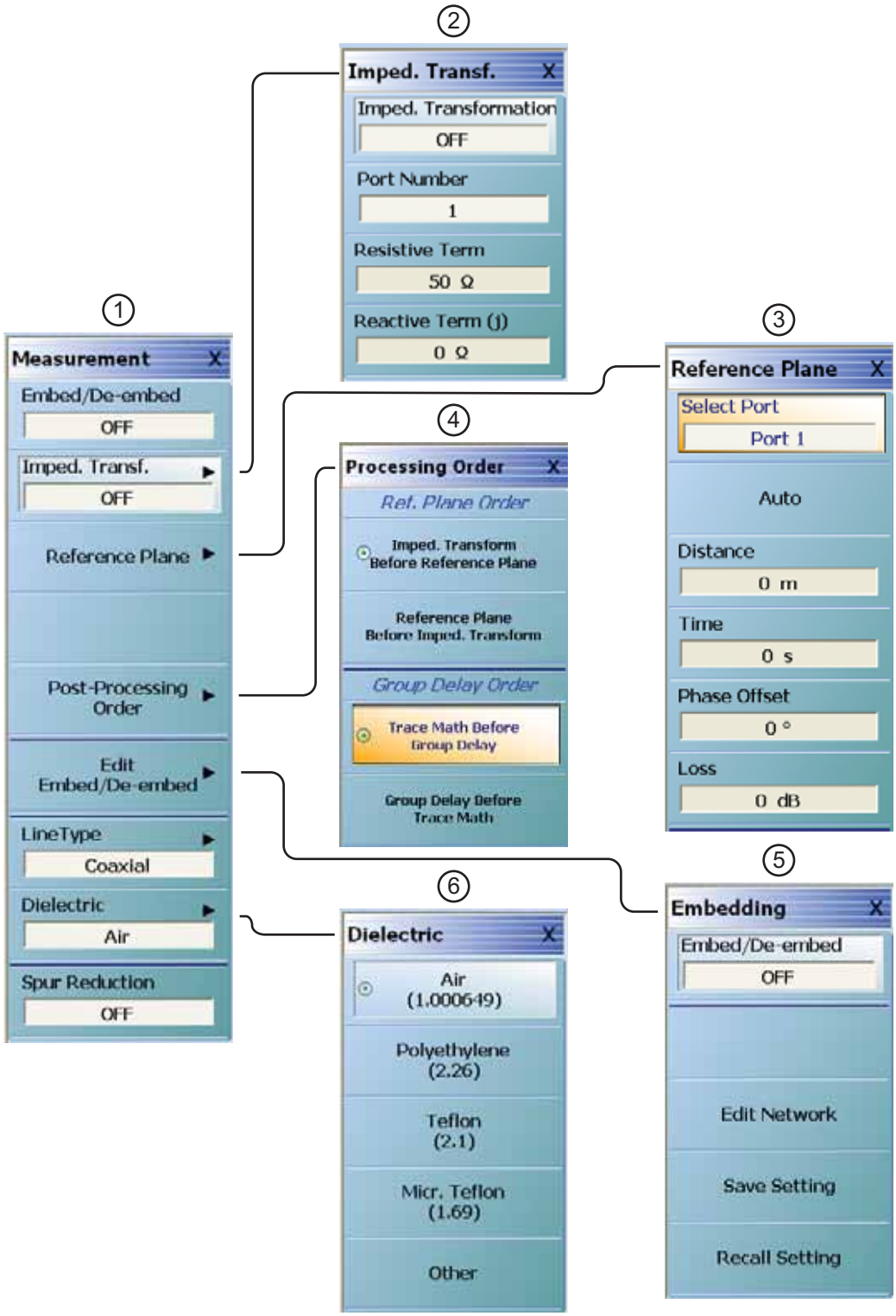


Figure 11-1. The MEASUREMENT Menu Set (1 of 2)

Use menus #1, #3, #4, and #5 for Embedding/De-embedding. 1. MEASUREMENT menu 2. IMPED. TRANSF. menu—Impedance Transformation	3. REFERENCE PLANE menu 4. PROCESSING ORDER menu 5. EMBEDDING menu. 6. DIELECTRIC menu

Figure 11-1. The MEASUREMENT Menu Set (2 of 2)

11-2 Embedding/De-embedding (E/DE)

The MS464xB Series VNAs are equipped with an embedding/de-embedding system. De-embedding is generally used for removal of test fixture contributions, modeled networks, and other networks described by S-parameters (s2p files) from measurements. Similarly, the embedding function can be used to simulate matching circuits for optimizing amplifier designs or simply adding effects of a known structure to a measurement. Multiple networks can be embedded/de-embedded (E/DE) and changing the port and network orientations is handled easily. An extraction utility is part of this package that allows the easier computation of de-embedding files based on some additional calibration steps and measurements.

It is extremely valuable to be able to virtually remove or add networks to the measured data as described above. The process of adding network data to measured data is termed "*embedding*" while the process of removing network data is termed "*de-embedding*."

Embedding Tasks

Common embedding tasks are to:

- View results as if a different launch structure was present
- View results as if a new matching circuit was being used
- View results as if an added cable length or transmission line length was needed

De-Embedding Tasks

Common de-embedding tasks are to:

- Remove the effects of a test fixture
- Remove the effects of a launch or launching transmission line
- Remove the effects of a test matching circuit that will later be physically removed

The MS464xB Series VNA embedding/de-embedding engine (E/DE) is a flexible tool for performing tasks of this type. A number of different circuit element primitives are available and full S2P files can also be loaded.

Note	Circuit parameters for embedding-de-embedding network elements are stored in S2P files that can be loaded into the VectorStar.

Embedding On/Off Control

E/DE can be turned on and off with the Embed/De-embed toggle button at the top of the main MEASUREMENT menu as shown above (Figure 11-1) or with a duplicate toggle button at the top of the EMBEDDING control menu as shown below (Figure 11-2).

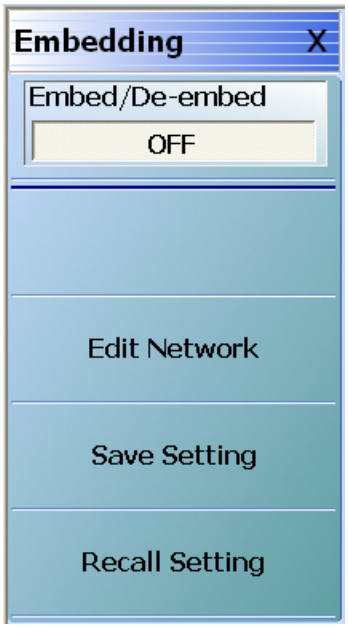
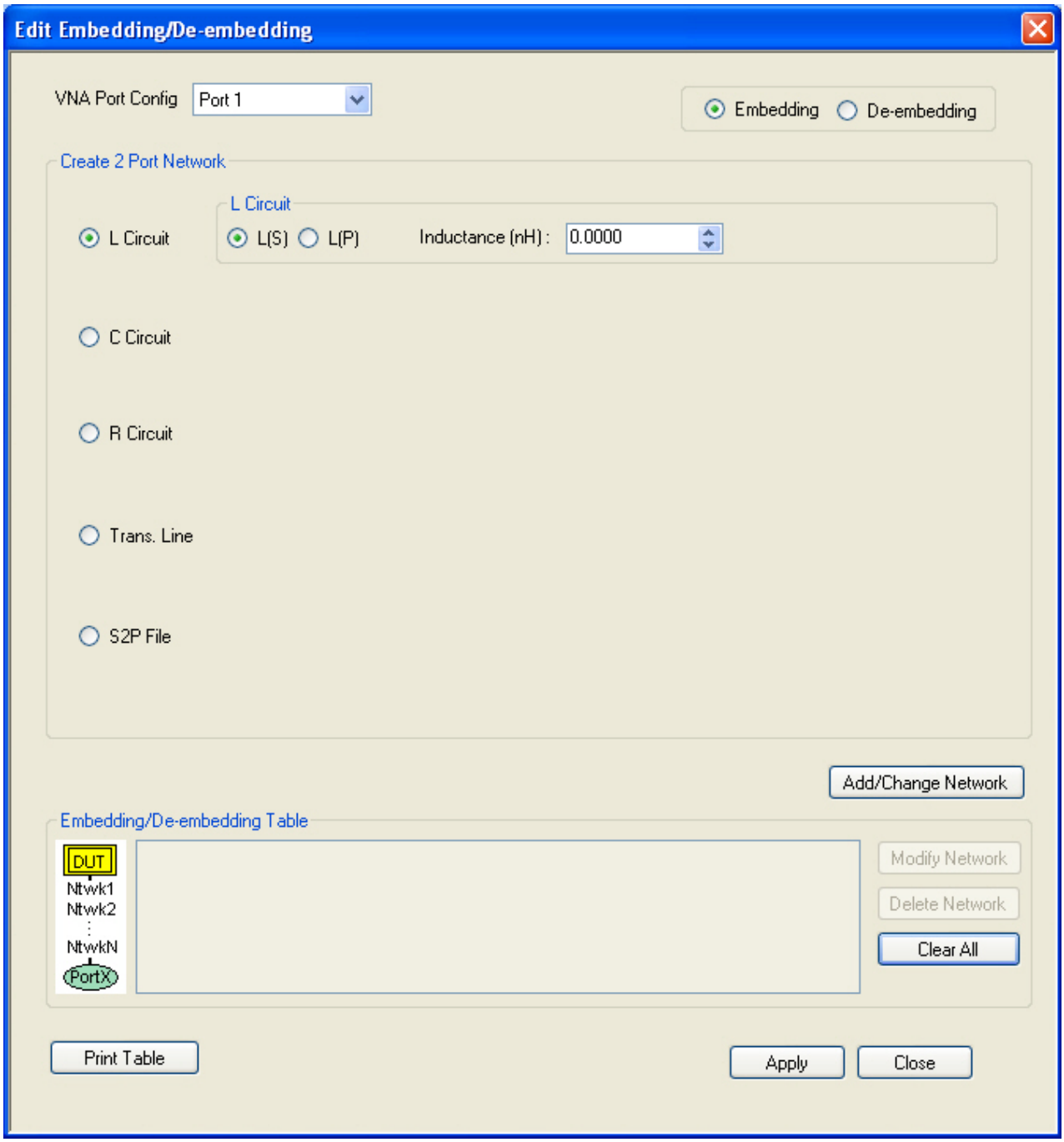


Figure 11-2. EMBEDDING Menu

Clicking on Edit Network displays the main EDIT EMBEDDING/DE-EMBEDDING dialog box. An example with Embedding, L Circuit, and L(S) selected, but with no network information entered is shown in the figure below.



L Circuit—L(S)—Embedding Selected

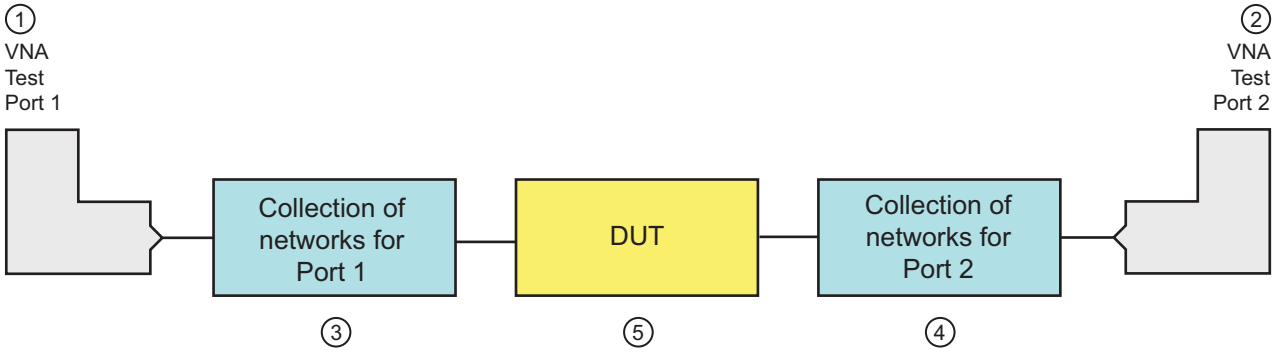
Figure 11-3. EDIT EMBEDDING/DE-EMBEDDING (2-Port DUT) Dialog Box—L Circuit

Note	Embedding and de-embedding is set up for each port and the networks used on the two ports are entirely independent. Also, any number of networks can be cascaded at a given port and the first network entered is always nearest the DUT for embedding. Sequential de-embedded networks are used in the reverse order.

The key concepts for embedding and de-embedding are:

- Networks are setup on a per-port basis
- The networks used on the two ports are entirely independent
- Any number of networks can be cascaded at a given port

The first network entered is always nearest the DUT. The pull-down menu at the top of Figure 11-3 shows which port's networks are currently being edited. The diagram in Figure 11-4 illustrates the independence concept.



1. Test Port 1
2. Test Port 2
3. Collection of networks attached to Port 1.
4. Collection of networks attached to Port 2.
5. DUT attached to Port 1 networks (at left) and Port 2 networks (at right).

Figure 11-4. Global EDE Diagram Showing Independence of Port Networks

Types of E/DE Networks

There are five types of networks that can be entered:

- Inductive elements
- Capacitive elements
- Resistive elements
- Transmission lines
- .S2P-defined, file-based networks

In the Edit Embedding/De-embedding dialog box in Figure 11-3 above, the radio button for entering an LC network has been selected. An additional version of the dialog box is shown in the figure below where a C(S) network has been selected.

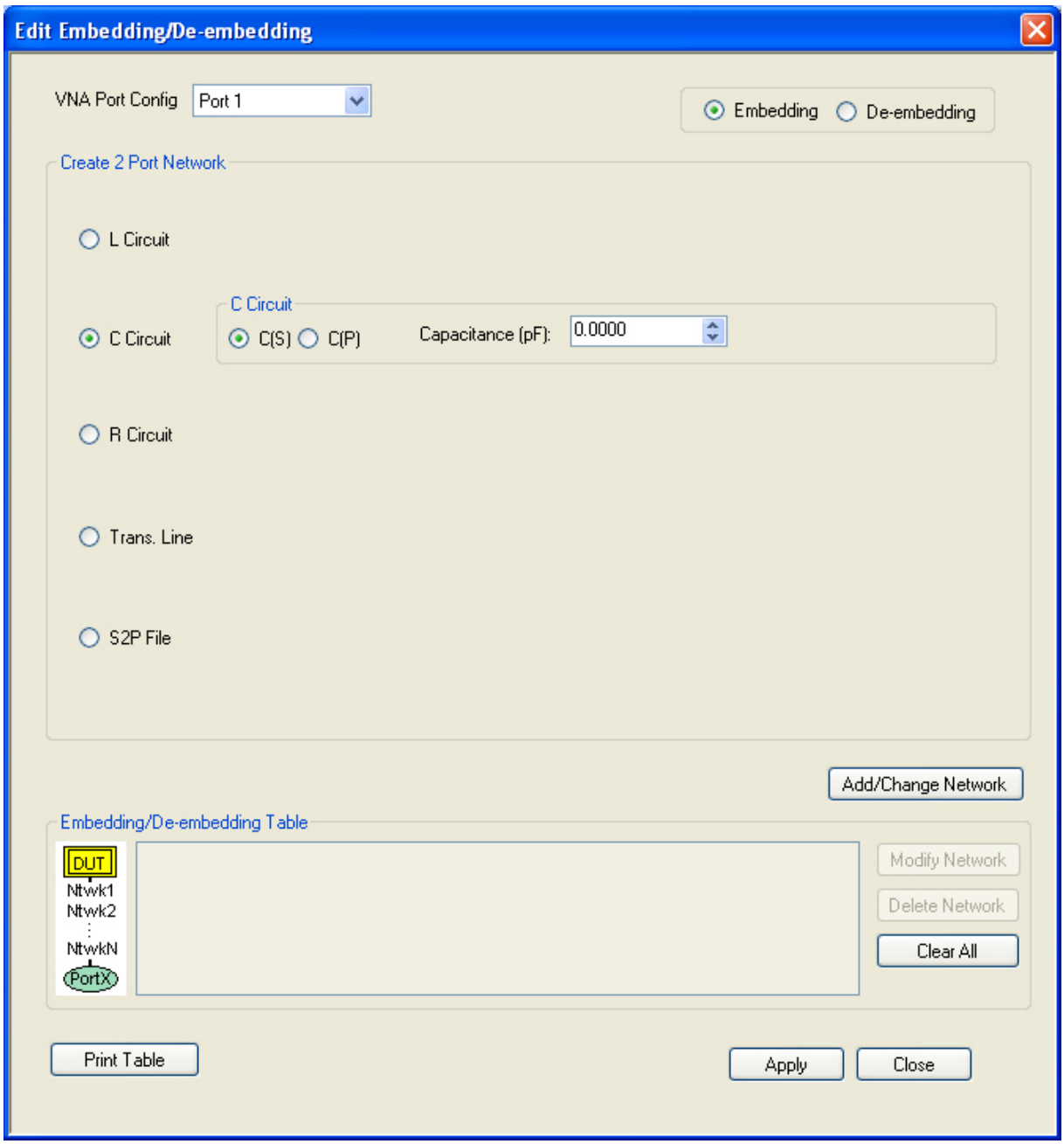


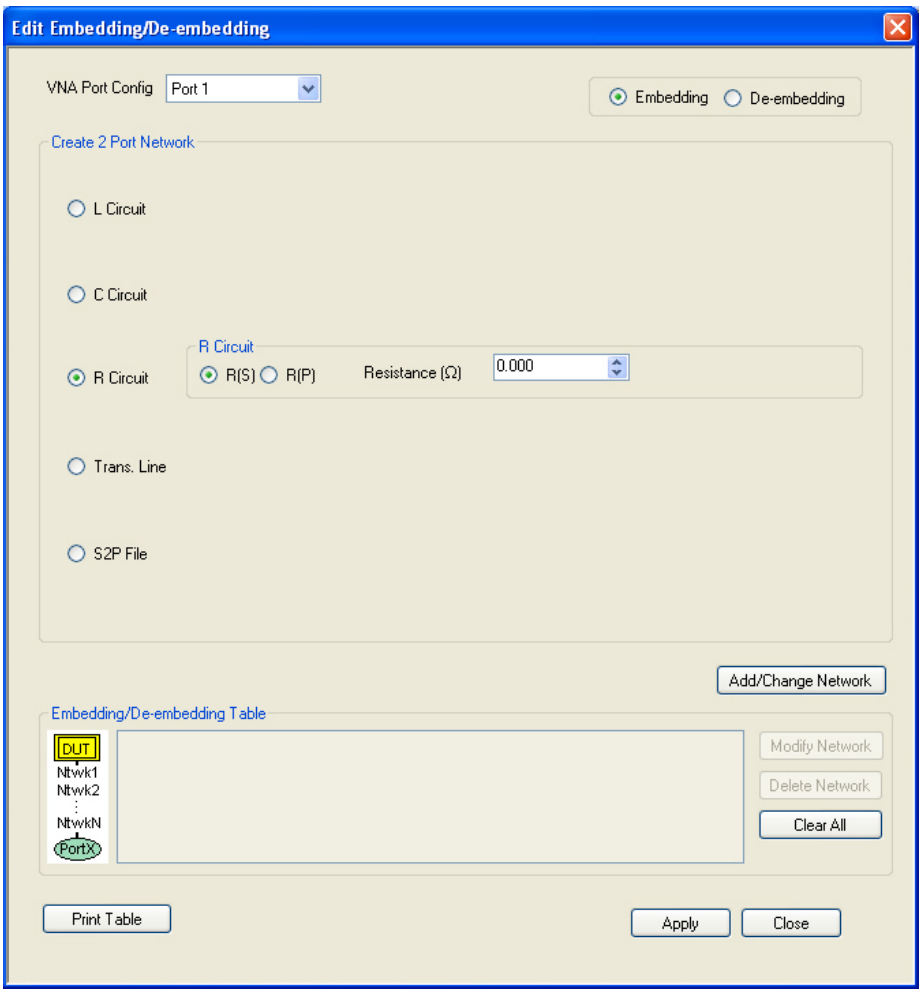
Figure 11-5. EDIT EMBEDDING/DE-EMBEDDING—C Circuit

Entry Mode for Resistive Elements

The entry mode for these resistive elements is shown in the E/DE dialog box below (Figure 11-6). Both series (denoted by an (S)) and shunt (to ground) elements (denoted by a (P) for parallel) are allowed and selectable with the radio buttons. Since this element is symmetric, no orientation knowledge (with respect to DUT port and VNA port) is needed. The default units are:

- Inductance: nH
- Capacitance: pF
- Resistance: ohms

It should be emphasized that the shunt or (P) elements are always shunting to ground (not to the other port). If cross-port elements are desired, then the multiport version of the instrument should be used with an appropriate calibration.



Resistive Element Entry

Figure 11-6. EDIT EMBEDDING/DE-EMBEDDING Dialog Box—R Circuit Setup Selected

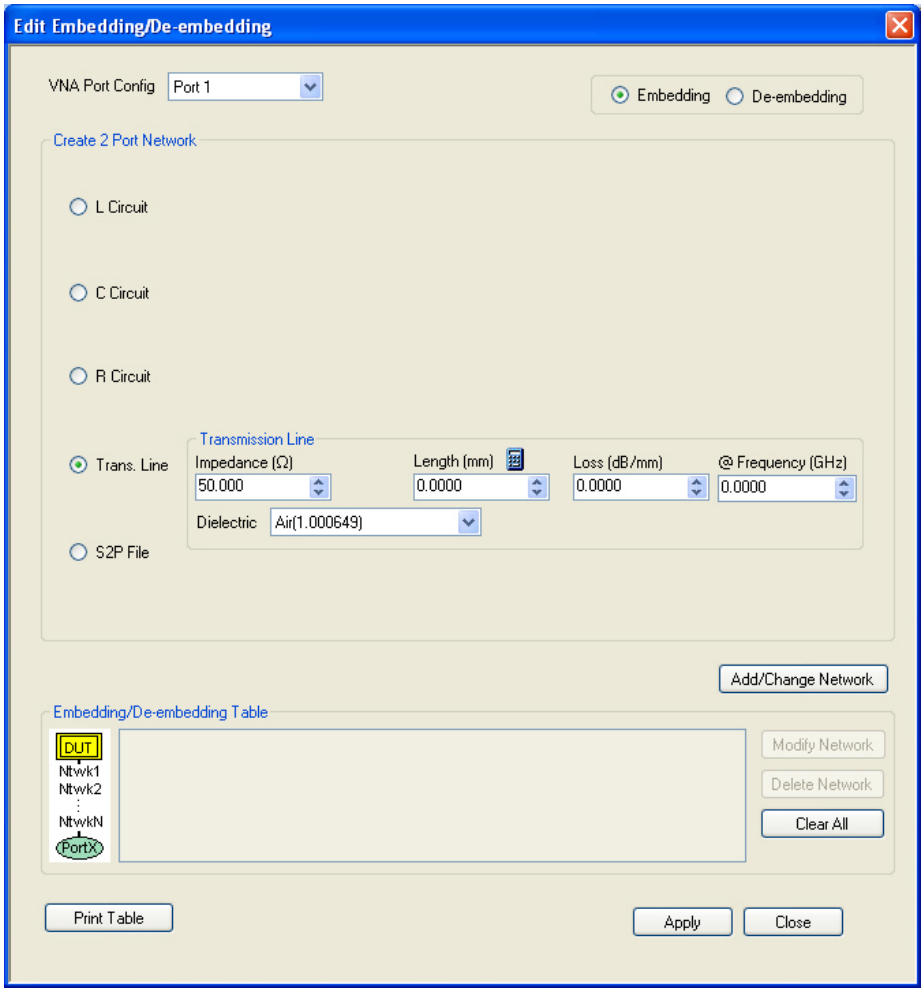
Entry Mode for Transmission Lines

Transmission line entry is illustrated in the E/DE dialog box below (Figure 11-7). As with transmission line entry in other parts of the system, loss can be entered along with a reference frequency. The loss at other frequencies will be computed using:

$$Loss\ (f) = Loss\ (f_0) \times \sqrt{\frac{f}{f_0}}$$

Equation 11-1.

As elsewhere in the system, if a 0 (zero) frequency reference is entered, the loss value entered will be used as a constant at all frequencies.



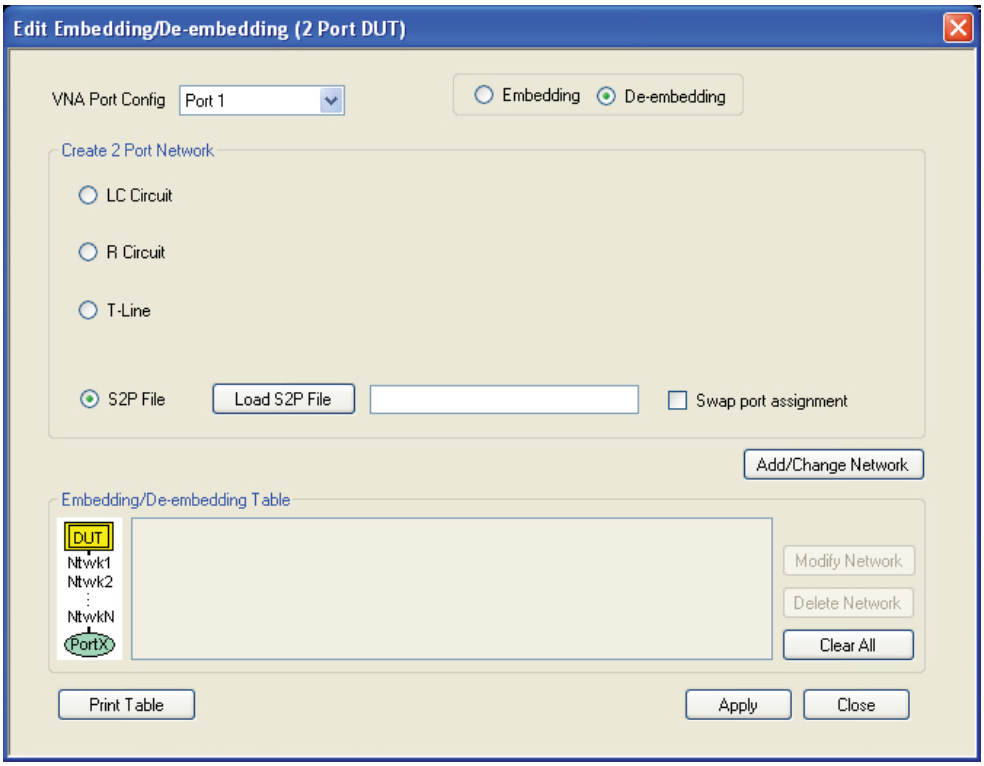
T-Line Selected—Transmission Line Element

Figure 11-7. EDIT EMBEDDING/DE-EMBEDDING Dialog Box

Physical line length is normally entered here along with a dielectric constant but the calculator icon shown in above in Figure 11-7 can be used (which links to the AIR EQUIVALENT LENGTH CALCULATOR dialog box) if only a time delay is known. Again, since this element is symmetric, no orientation knowledge is needed.

Entry Mode for S2P Defined File-Based Networks

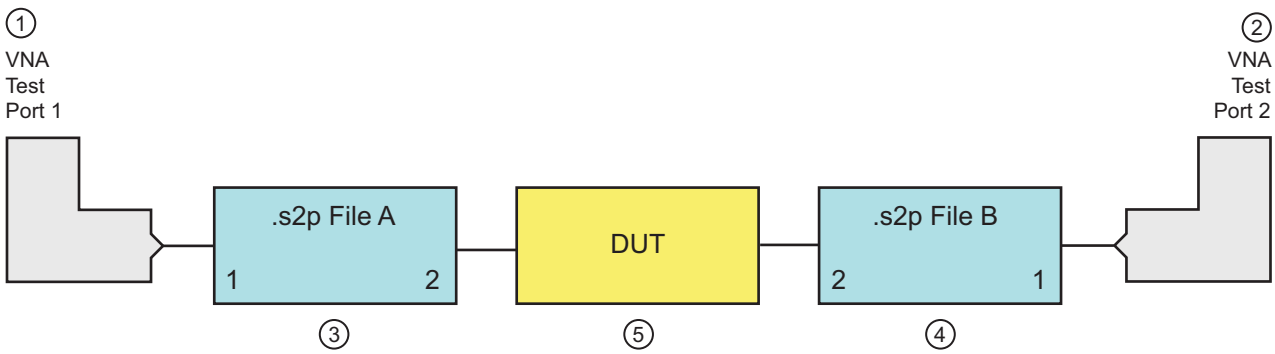
Finally, direct file entry of network S-parameters is shown in the E/DE below (Figure 11-8). A standard S2P file format is assumed and the headers will be interpreted. The system will attempt to interpolate the provided data the best it can in the context of the current channel sweep range. If there is no overlap between the sweep range and the file frequency range, an error will be generated.



S2P File-Based Element Entry

Figure 11-8. EDIT EMBEDDING/DE-EMBEDDING Dialog Box

Note	The network Port 2 is always assumed to be closer to the DUT regardless of which VNA port is involved. See Figure 11-9. If “Swap port assignment” is checked, the relationship is reversed.



1. Test Port 1
2. Test Port 2
3. .s2p File A—Port 1 to left, Port 2 to right.
4. .s2p File B—Port 2 to left, Port 1 to right.
5. DUT

Figure 11-9. Orientation of Loaded S2P Files

Saving and Recalling Embedding Network Configuration

Once a set of networks (consisting of one or more individual networks) is defined, the E/DE configuration information can be saved to a file using the Save Setting button on the EMBEDDING menu (Figure 11-2, "EMBEDDING Menu" on page 11-4). Similarly, a stored E/DE setup can be recalled by using the Recall Setting button on the EMBEDDING menu.

The current E/DE setting is also saved as part of the system setup save (under the menu bar File menu) but multiple embedding and de-embedding circuits can be saved in these menus as well.

11-3 Reference Plane Control

A simplified means of performing de-embedding (and embedding in some contexts) can be accomplished using reference plane control. The function of this control is to remove transmission line lengths from the data. By entering a time or distance, this length of line will be removed (negative lengths are allowed to effectively add length). Various dielectrics and the full dispersion choices (see calibration section of the Measurement Guide for more information) are available.

Reference Plane Menu

The REFERENCE PLANE control menu is shown in Figure 11-10. The ports are handled independently, as in E/DE, and the current port being affected is indicated by the toggle at the top of the menu.

Navigation:

MAIN | Measurement | MEASUREMENT | Reference Plane

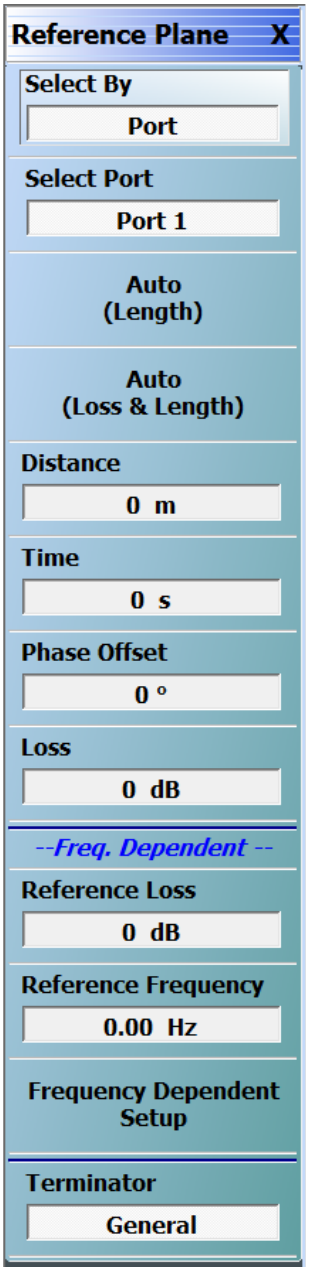
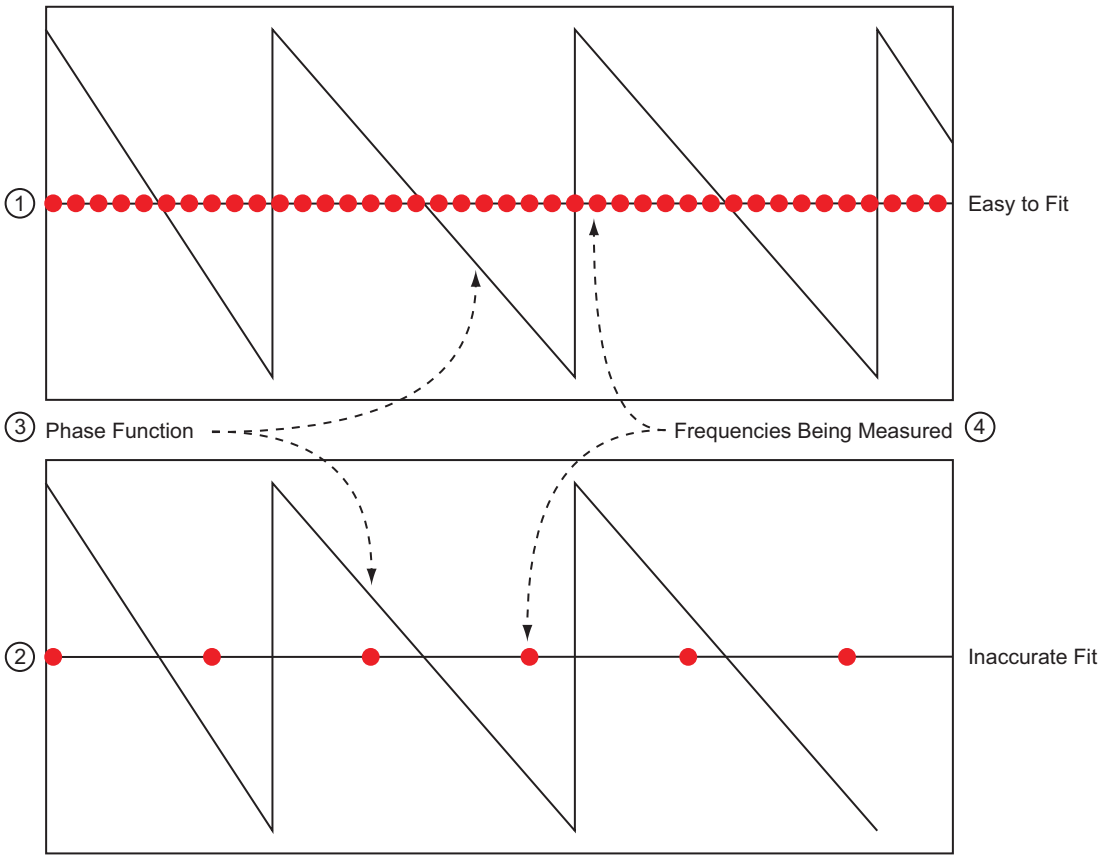


Figure 11-10. REFERENCE PLANE Control Menu

Auto Button Functions

The Auto button performs a best fit operation to the current phase data to estimate the equivalent line length. It will attempt to generate a line length that, when removed, will make the phase flat. This routine will be less accurate if the DUT has very non-linear phase (a dispersion function not matching that selected) or if the DUT is electrically long relative to the current frequency step size. This latter problem, related to aliasing, occurs because not enough information is being collected relative to the true behavior of the DUT phase function (see Figure 11-11 below). Increasing the frequency point density can help this problem.



1. Easier to Fit—Small frequency step size compared to phase function is easier to fit and yields higher accuracy.
2. Inaccurate Fit—Larger frequency step size compared to phase function is can produce inaccurate results.
3. Phase Function
4. Frequencies being measured.

Note: The auto reference plane function can produce inaccurate results if the frequency step (the distance between the red ovals in the figure above) is large relative to the phase function period.

Figure 11-11. Auto Reference Plane Function, Frequency Step Large Relative to Phase Period

Reference Plane Loss

In addition to the frequency-dependent phase shift with length/time, the fixed phase shift and the fixed loss shift discussed so far, there is also a frequency-dependent loss term and an Auto Ref Plane function that applies to it. Unlike the linear-phase-with-frequency model assumed for length/time, the function form for loss/gain is:

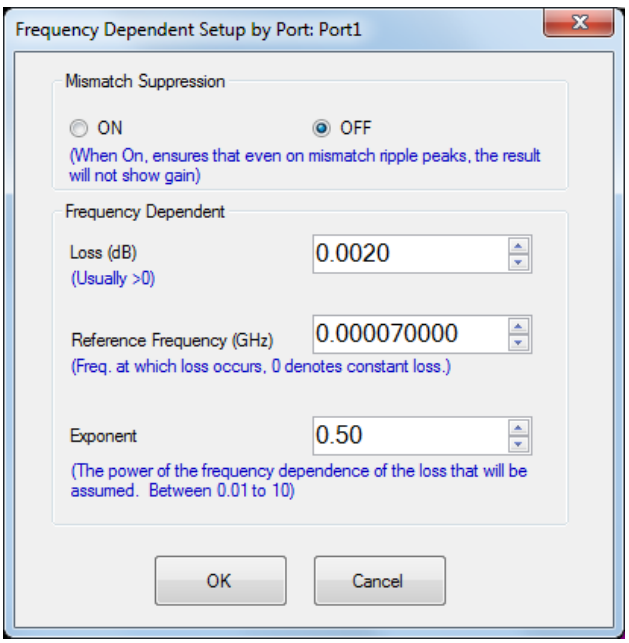
$$A \bullet \left(\frac{f}{f_0} \right)^n \text{ (in dB)}$$

Equation 11-2.

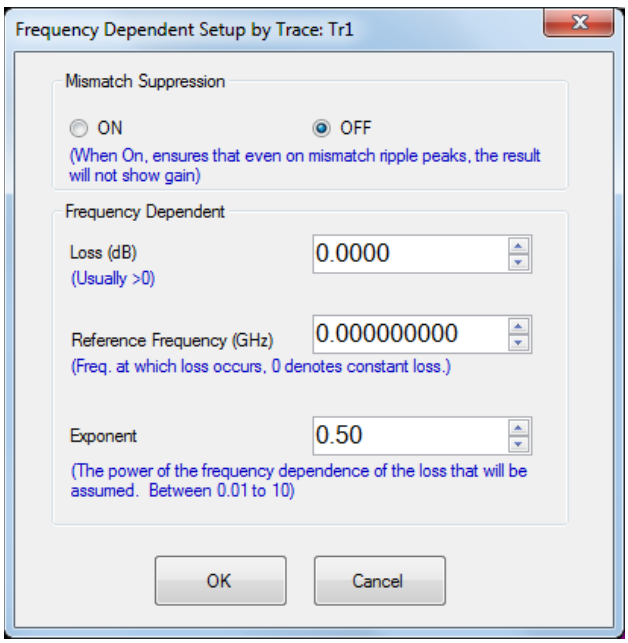
Where:

- A is a nominal loss (or gain if negative) value in dB and f_0 is the frequency at which that loss or gain value occurred. The functional form then describes the value at other frequencies.
- The exponent n is user-selectable with a default value of 0.5 which tends to describe loss in coaxial lines and in coplanar waveguide rather well for many materials. The exponent may be closer to 1 for microstrip structures and other values for other geometries. The allowed range for the exponent is 0.01 to 10 but it is fairly rare to get outside the range of 0.25 to 2. The exponent, along with a *mismatch suppression* option for the auto ref plane process (to be discussed), are entered in a Frequency Dependent Setup configuration dialog (accessed from the Measurement | Reference Plane menu) as shown in Figure 11-12.

Port Dialog



Trace Dialog



The configuration dialog for some frequency-dependent loss reference plane items is shown above.

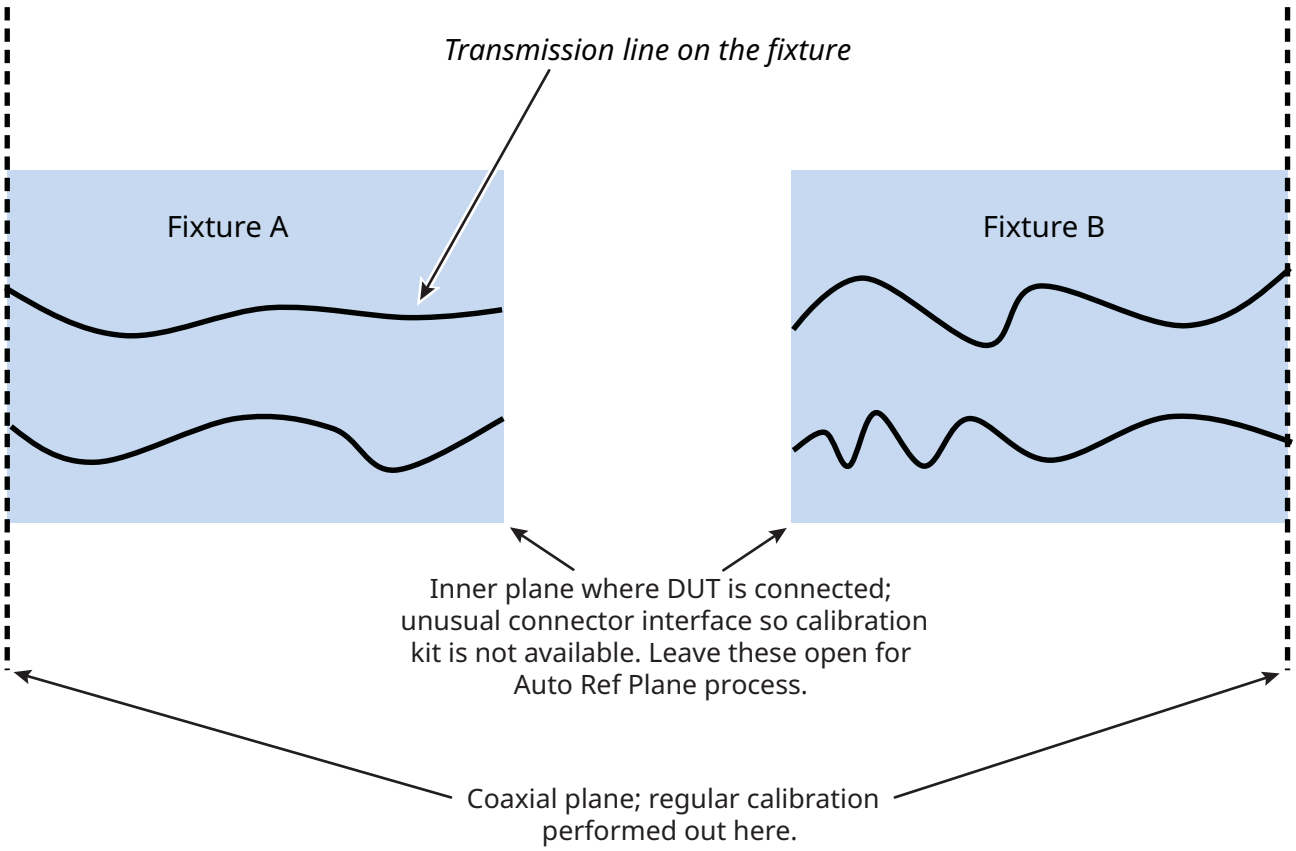
Figure 11-12. Frequency Dependent Setup Dialogs (Per-Port or Per-Trace)

Per-trace and per-port conventions apply for this frequency-dependent loss entry just as they do for the other reference plane parameters.

Reminder: Per-trace reference plane entries do not affect saved .sNp file results but per-port entries do. In a transmission parameter, the reference plane correction will be the product of per-port reference plane adjustment functions from the involved ports (whether that is for regular S-parameters or user-defined parameters).

When the Auto Ref Plane Length/Loss is used, fits are done on both the phase (using the process described previously) and the magnitude (using the exponential functional form just mentioned) independently. Values for the fit parameters are entered in the appropriate menu fields and the adjustments applied to the trace data.

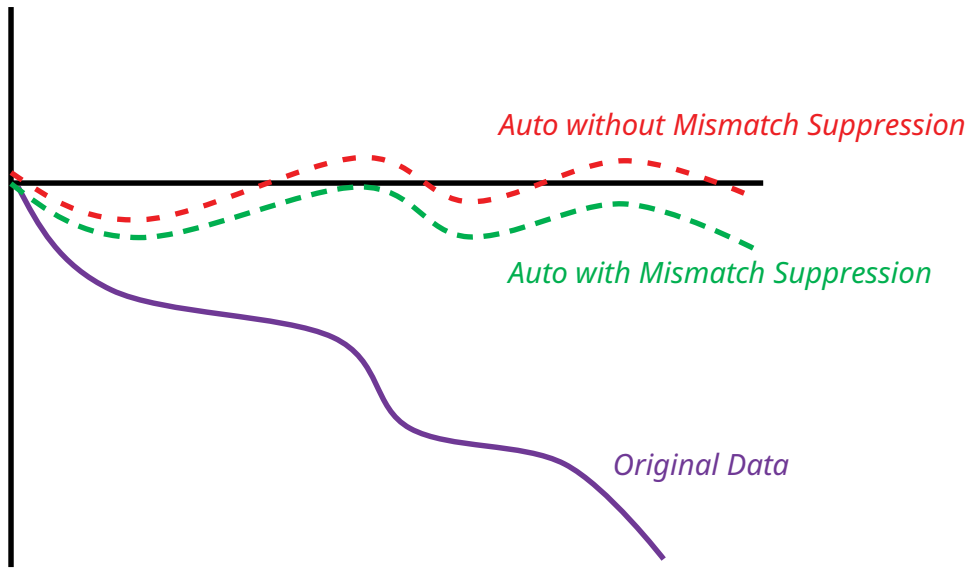
This frequency-dependent loss aspect of reference plane extension can be useful for very simplified de-embedding of fixtures or cabling. In one case, the fixture assembly is left with 'opens' at the DUT interface (a regular calibration was performed at the outer interface shown in Figure 11-13). Using the reflection measurements at each port, the auto ref plane process can fit the loss (as well as length) and remove the effects of the fixture to some extent (other variations of this process use a thru connect between fixture halves and, in this case, a transmission parameter would be used). This is a simplified process in that match of the fixture itself is being neglected entirely and the loss of the fixture is assumed to follow the exponential dependence discussed. More accurate de-embedding methods (using a variety of network extraction tools) are discussed elsewhere in this guide that make use of (sometimes) additional standards or assumptions. The auto reference plane approach is, however, very easy to perform and may be adequate for more well-behaved fixtures and cabling structures.



An example test fixture is sketched here showing how auto ref plane length/loss is used to perform very simplified de-embedding of the fixture response.

Figure 11-13. Example Test Fixture

An additional option for the length/loss auto ref plane function is that of Mismatch Suppression. The concept is to limit that amount of loss correction so that no ripple peaks in the adjusted result exceed the initial (lowest loss) value of the parameter. If the DUT has very low loss at low frequency, not suppressing the effect of mismatch-induced ripple could result in an adjusted parameter value above 0 dB which may be objectionable in some applications. With Mismatch Suppression activated, the fitting process is modified so any ripple peaks will stay below the nominal initial value of the parameter in question. This concept is sketched in Figure 11-14.



Auto ref plane extension (length/loss) effects are shown here with and without mismatch suppression enabled for some example input data with considerable ripple.

Figure 11-14. Auto Ref Plane Extension (Length/Loss) Effects

11-4 Impedance Transformation

Most VNA calibrations are performed referenced to 50 ohms, as this is usually set by the calibration kit. While some calibration kits exist for other impedances (75 ohm N and F connectors for example), they are not common and a custom impedance may be of interest. The impedance transformation function allows performance of a calibration in one impedance and then transformation of the result to appear as if it had been calibrated in a different impedance. As a crude example, if a 75 ohm N calibration kit is not available, but a 50 ohm kit is, neglecting issues with adapters, the 50 ohm calibration could be performed and the utility used to reference the results to 75 ohms. The IMPED. TRANSF. (IMPEDANCE TRANSFORMATION) menu for this function is shown below (Figure 11-15).

Another application of this function is to correct calibrations for the case where it is found that the actual reference impedance (e.g., the characteristic impedances of lines in LRL/TRL calibrations or the load impedance in LRM and SOLT calibrations) is not ideal. Once the actual impedance of the calibration is known, the impedance transformation tool can be used to convert results to those referenced to 50 ohms (or some other value).

The ports can be set independently by using the Port Number toggle button and can be set to any complex value except 0 (zero). Generally, a non-real value will only have meaning in a narrowed frequency range.

A full-term calibration must be active for impedance transformation to take effect and only those ports currently calibrated will be affected.

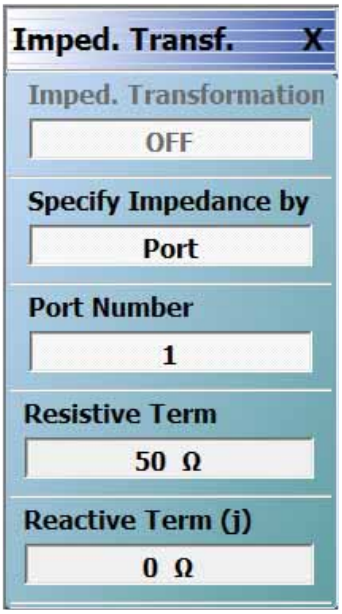


Figure 11-15. IMPED. TRANSF. (IMPEDANCE TRANSFORMATION) Menu

The calculations are based on a pseudo-wave framework (e.g., R. B. Marks and D. F. Williams, "A general waveguide circuit theory," J. Rsrch. of the Nat. Inst. Of Stds. And Tech., vol. 97, pp. 533-561, Sept-Oct 1992). The equations used are matrix-based (scaling with number of ports) and defined by:

$$S' = P^{-1}(S-\gamma)(I-\gamma S)^{-1}P$$

where:

$$P_{ii} = \sqrt{\frac{\mathrm{Re}(A_{ii})}{\mathrm{Re}(B_{ii})}}\left|\frac{B_{ii}}{A_{ii}}\right|\frac{2A_{ii}}{A_{ii}+B_{ii}}$$

$$\gamma_{\mathrm{ii}} = \frac{B_{ii}+A_{ii}}{B_{ii}-A_{ii}}$$

The default is to specify the new reference impedances by port, but it can also be done by port pair. In this case, common-mode and differential impedances (per pair) are specified instead of by single-ended ports. This approach can be useful for certain engineered transmission lines where the single-ended impedances are complex.

11-5 Processing Order

With so many post-processing choices available, it is important to note that the order of operations can matter. A few things are fixed by the way computations are performed and others are changeable to suit user needs. The sequence of computations is as follows for S-parameter measurements:

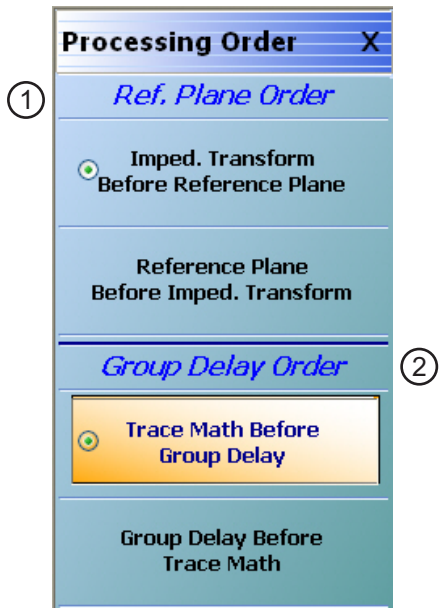
- Acquire raw data and average/filter
- Apply calibration if enabled
- Apply EDE if enabled (impedance transform and reference plane)
- Apply parameter conversions if enabled
- Apply time domain if enabled

Reference Plane Processing Sequence

The selectable item is the order in which impedance transformation and reference plane control are applied. This matters since the current impedance state determines the impedance of the line length that is adjusted using reference plane control (unlike in E/DE where the impedance can be specified). The menu to make this order selection is shown below (Figure 11-16).

Group Delay Processing Sequence

The second selectable item concerns the order of the group delay computation and trace memory. If trace memory precedes group delay (normal), then the complex operation can precede the numerical differentiation that is part of group delay. It could be useful to do a data normalization prior to connecting a DUT. If group delay precedes trace memory, then the trace match will act on the final group delay result. An example might be a group delay comparison using data(-)memory.



1. Reference Plane Order controls (at top).	2. Group Delay Order controls (at bottom).

Figure 11-16. PROCESSING ORDER Menu

11-6 Conversions

While S-parameters (or the UN-ratioed wave parameters) are usually the display variables of interest, conversions to other parameters may be required and are possible with the VectorStar.

1/S is sometimes plotted, particularly for oscillators and other negative resistance devices, where it is desirable to fold the outside of the Smith chart back to the inside. Equivalent impedances and admittances are commonly needed for device modeling and the Z and Y conversions can be used for this (note that these are not, in general, Z and Y parameters).

For each of these parameters, a conversion mode of reflection or transmission must be selected which indicates how the current parameter is to be interpreted. The calculations proceed as follows, where X indicates the current displayed parameter such as S_{11}, S_{21}, b_1/a_1, a_2/a_1, and user-defined parameters:

$$Z_{reflection} = Z_0 \frac{1+X}{1-X}$$

$$Z_{transmission} = Z_0 \frac{2(1-X)}{X}$$

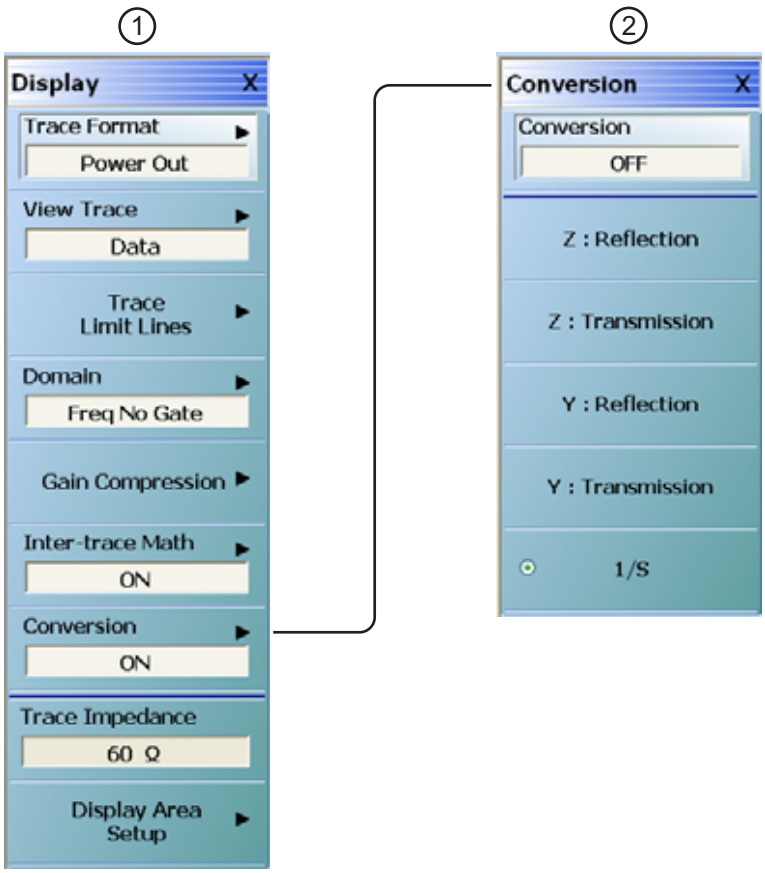
$$Y_{reflection} = \frac{1}{Z_0} \frac{1-X}{1+X}$$

$$Y_{transmission} = \frac{1}{Z_0} \frac{X}{2(1-X)}$$

Equation 11-3.

Note that reflection Z or Y values here represent shunt impedances or admittances while the transmission values represent series impedances or admittances.

All of the choices available are on the CONVERSION menu shown in Figure 11-17 on page 11-20. Note that this function is *PER TRACE* and is located in the DISPLAY menu.



1. DISPLAY menu (on left).	2. CONVERSION menu (on right).

Figure 11-17. Conversion Control Menu

The calculations are a function of the current reference impedance, which defaults to the calibration reference impedance unless impedance transform has been used (see the section above on "Reference Plane Control" on page 11-11) or the trace reference impedance has been changed.

Chapter 12 — Performance Verification Tools

12-1 Introduction

While there are many ways of verifying instrument performance, including the procedures described in the **VectorStar MS4640B Series VNA Operation Manual** and the **Maintenance Manual**, sometimes a simpler procedure can be useful. Verification kits available from Anritsu verify the measurement capabilities of the instrument by analyzing the measurement of artifacts that are traceable to national standards laboratories.

This chapter introduces and describes the verification process. Verification kits and software are available as a separate product and are described in detail in their associated documentation listed below: In Chapter 1, "Calibration and Measurement Overview", see "Calibration Kits, Verification Kits, and Components" on page 1-2 for a detailed list.

- *MS464xB Series VNA Operation Manual* (OM)—10410-00317
- *MS464xB Series VNA Maintenance Manual* (MM)—10410-00320
- *366X-1 Verification Kits (3666-1 3.5mm Connectors, 3668-1 K Connectors, 3669B-1 V Connectors) and 2300-579 (67688) Performance Verification Software (PVS) User Guide*—10410-00270
- 366X-1 Verification Kits and 2300-579 (67688) PVS Quick Start Guide—10410-00285
- *3656B W1 (1 mm) Calibration/Verification Kit and 2300-584-R System PVS User Guide for the VectorStar ME7838A/ME7828A and Lightning ME7808A/B/C BB/mmWave VNA Systems*
- *3656C W1 (1 mm) Calibration/Verification Kit and 2300-616-R System PVS User Guide for the VectorStar ME7838A/ME7828A BB/mmWave VNA Systems*—10410-00784
  The 3656C kits also support 1 mm calibration and verification to 125 GHz, and separate PVS software exists for those kits.
- *3659 Calibration/Verification Kit and 2300-580 Performance Verification Software (PVS) User Guide for BB/mmW ME7838D with 0.8 mm Connectors*—10410-00327

12-2 Concepts and Levels of Verification

There are many levels to the concept of VNA verification which, in one sense or another, is a comparison against expected behaviors.

Hardware Level

On the explicit VNA hardware level are operational checkout items such as port power, receiver signal levels, and noise levels. These items are covered in the Operation Manual. On the calibrated instrument level (which includes the VNA and the calibration kit or AutoCal) are the residual specifications (corrected directivity, source match, load match, and tracking) which are measured using traceable airlines (absolute impedance standards).

Intermediate System Level

An intermediate level which can look at overall system behavior (VNA, calibration kit, cables, environment) in a traceable fashion is through the use of a verification kit. While not intended for day-to-day use, the verification kit can provide a periodic check on system behavior without going through the rigor needed for full residual analysis (which can usually be done less often).

Comparison to Known Devices

The central idea of the verification kit is to have a collection of "known" devices (not calibration components) that have been measured with a calibrated VNA. By comparing the results to the "known" values, some measure of confidence can be gained in the measurement abilities of the VNA-under-test. The values in all cases are vector quantities so that both magnitude and phase responses are analyzed.

The "known" part of this discussion involves a process termed characterization performed on the same devices by Anritsu. Through a traceable process, a VNA at the factory is calibrated and validated against controlled standards before being used to measure the devices that go into the verification kit that is delivered to the user. By carefully controlling this process, measurement uncertainties on the characterization end can be carefully controlled. This allows a useful window to be defined as to what an acceptable measurement result is. At each frequency point, the measurement is compared to the characterization measurement in the context of the uncertainties. If the delta between the two measurements is consistent with the uncertainty window, the measurement is considered acceptable at that point.

Metric of Comparison

The metric of comparison, termed E_n, is a check to see if the measurement differences are consistent with the uncertainty windows of both the characterization and the verification measurements. The quantity is shown below:

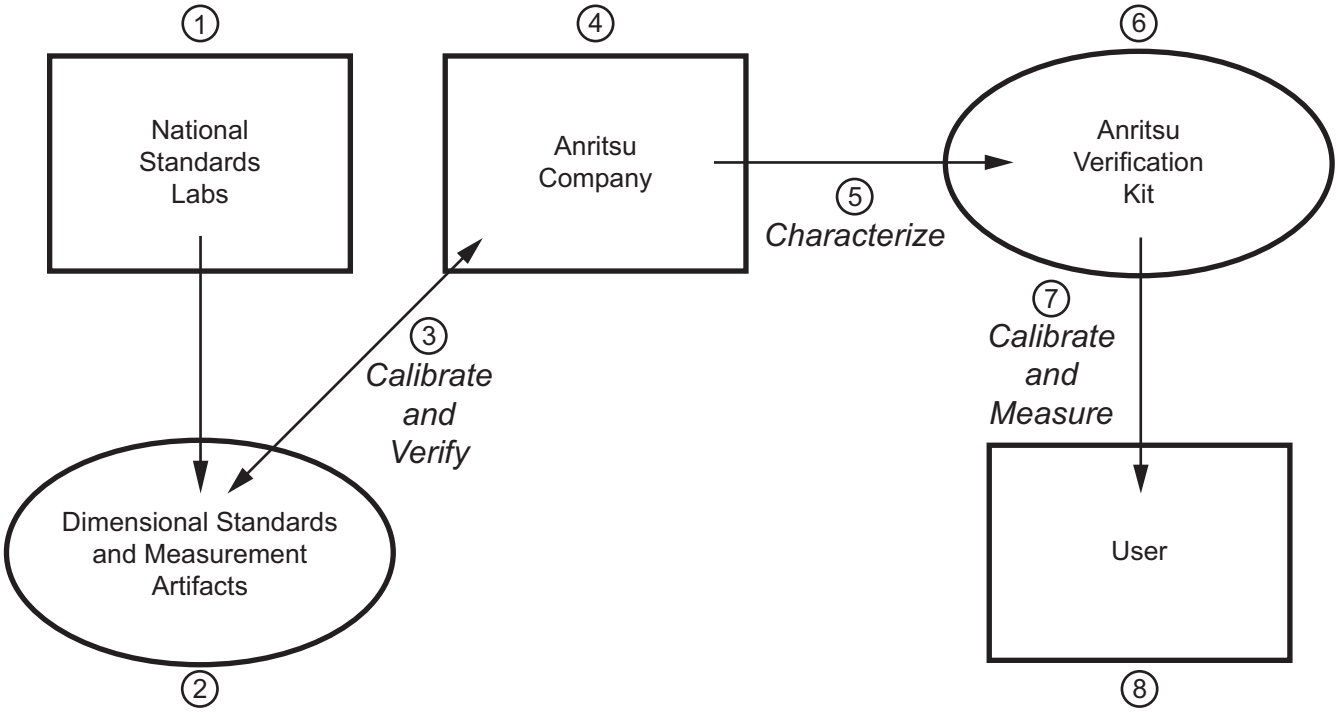
$$E_n = \frac{\left| S_{xy}^{char} - S_{xy}^{ver} \right|}{\sqrt{\left(U_{xy}^{char} \right)^2 + \left(U_{xy}^{ver} \right)^2}}$$

Equation 12-1.

The numerator contains the S-parameters measured during characterization (by Anritsu) and during verification (by the user), and the denominator contains the respective uncertainties. These uncertainties are calculated based on the VNA, the calibration kit, and repeatability. If this quantity E_n is less than 1, then the measurements during the two phases are within the overlap of the uncertainties and can be considered "equivalent" and, in some sense, verified.

Verification Kit Creation Process

The general process of the creation and use of a verification kit is shown in Figure 12-1.



1. National Standards Labs.	5. Characterization processes for verification kits.
2. Dimensional standards and measurement artifacts.	6. Anritsu Verification Kits with characterization data.
3. Calibrate and verify using artifacts.	7. Calibrate instruments and measure DUTs.
4. Anritsu Company processes and standards.	8. Verification kit used to verify instrument calibration.

Figure 12-1. General Process of the Creation and Use of a Verification Kit

A national standards laboratory (through standards and measured artifacts) helps validate the calibration and Anritsu which is then used to characterize the verification kit sent to the user.

It is important to note that the results are influenced by not only the instrument, the calibration kit, and the verification kit, but also the cables, the environment (temperature, humidity and vibration), connector quality, and the care exercised by the user during calibration and measurement. It should also be emphasized that this is not a measurement against absolute standards (which is the case for the residuals measurement process) but is a study of an “equivalent” measurement; the same devices measured with the instrument at the user site versus those devices measured with a controlled instrument at Anritsu (under traceable conditions).

12-3 Verification Kit Components

The three main verification kits for the MS464xB Series VNAs are those corresponding to 3.5 mm, K, and V connectors:

- 3666-1 series for 3.5 mm
- 3668-1 series for K
- 3669B-1 series for V

The verification kits and their use are described in the following documents:

- *VectorStar MS464xB Series VNA 3666-1 3.5 mm, 3668-1 K, and 39669B-1 V Verification Kits and 2300-579 (67688) Performance Verification Software (PVS) User Guide*—10410-00270
- *VectorStar MS464xB Series VNA 366X-1 Verification Kits and 2300-579 (67688) PVS Quick Start Guide*—10410-00285

The devices in these kits are selected based on their ability to stress the envelope of possible measurement parameters while still providing a very stable and repeatable behavior. The key attribute of the devices is that of long-term stability. These kits contain the following devices:

- An airline to represent a low-loss, well-matched device:
  - 7.5 cm for K and 3.5 mm
  - 5 cm for V
- A stepped impedance (Beatty) airline to represent a variable match device with a range of insertion losses:
  - 7.5 cm for K and 3.5 mm
  - 5 cm for V
- A 20 dB attenuator to represent a moderate-loss, well-matched device.
- A higher value attenuator to represent a very lossy device:
  - 50 dB for K and 3.5 mm
  - 40 dB for V

No Verification with Calibration Kits

Calibration kit components cannot be used for verification since the result would be biased (some calibrations force the result to match expectations for the components used during calibration). Higher loss devices could have been used but they become more difficult to characterize accurately and the value of the verification would be reduced. Active devices with gain have not been used due to concerns about stability of the response over time. The envelope of coverage of the existing standards is illustrated in Figure 12-2, “Regions of Parameter Coverage of Verification Kit Component” on page 12-5. Each standard has a regime of coverage in terms of insertion loss and return loss; when combined, the entire kit reasonably exercises a wide variety of parameter values.

The diagram in Figure 12-2 illustrates the regions of parameter coverage of the various verification kit components. The regions are not drawn to scale.

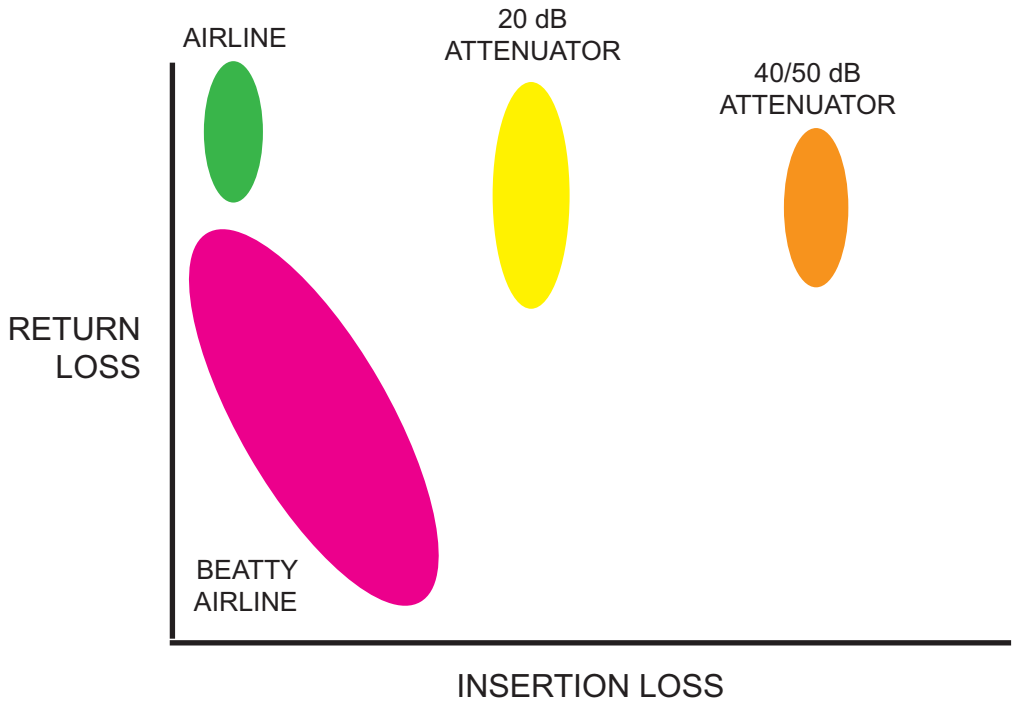


Figure 12-2. Regions of Parameter Coverage of Verification Kit Component

For the MS464xB Series VNAs, the verification comparison is valid for calibrations performed with the calibration kits listed under each verification kit:

- **3668-1 K Verification Kit**

  3652A-1 K Manual Calibration Kit or the 36585K-2MF Precision AutoCal Module

- **3669B-1 V Verification Kit**

  The 3654D-1 V Manual Calibration Kit or the 36585V-2MF Precision AutoCal Module

- **3666-1 3.5 mm Verification Kit**

- The 3650A-1 SMA/3.5 mm Manual Calibration Kit

Other combinations of calibration kits with verification kits are not supported. Verification kits based on other connector types such as GPC-7 or N exist for other Anritsu VNAs, but the MS464xB Series VNA verification software does not support all of these.

12-4 Verification Kit Software

The application provided with the verification devices prompts calibration of the VNA, acquires measurements of the devices, and compares those measurements against the characterization values generated by Anritsu (these values ship with the verification kit). The software also generates reports indicating the outcome of the verification. More information is available in the user guide provided with the verification kit:

- 366X-*1 Verification Kits (3666-1 3.5mm Connectors, 3668-1 K Connectors, 3669B-1 V Connectors) and 2300-579 (67688) Performance Verification Software (PVS) for MS464xB Series VNAs User Guide*—10410-00270
- *366X-1 Verification Kit and 2300-579 (67688) PVS Quick Start Guide*—10410-00285
- *3656B W1 (1 mm) Calibration/Verification Kit and 2300-584-R System PVS User Guide for the VectorStar ME/7838A/ME7828A and Lightning ME7808A/B/C BB/mmWave VNA Systems*—10410-00286
- *3656C W1 (1 mm) Calibration/Verification Kit and 2300-616-R System PVS User Guide for the VectorStar ME/7838A/ME7828A and Lightning ME7808A/B/C BB/mmWave VNA Systems*—10410-00784
  The 3656C kits also support 1 mm calibration and verification to 125 GHz, and separate PVS software exists for those kits.
- *3659 Calibration/Verification Kit and 2300-580 Performance Verification Software (PVS) User Guide for BB/mmWave ME7838D with 0.8 mm Connectors*—10410-00327

Chapter 13 — Measurement Setup Requirements

13-1 Chapter Overview

This chapter provides general measurement setup fundamental concepts, requirements, and options for different types of measurements. Specifically, this chapter describes channels, traces, limit lines, external analog input/output, averaging, and smoothing and organizes their configuration in the same hierarchy. Channel is the highest level configuration with up to 16 channels possible, each with a different frequency range, power level, IF bandwidth, and RF calibration. On a per-channel basis, traces are lower level concepts that represent a data group with up to 16 traces per channel. Limit lines are described with setup tasks and test functionality. External analog input/output is described with setup issues, range, functions, resolution, and accuracy. A description of averaging and smoothing with their available functions and the effects on measurements conclude the chapter.

13-2 Channels and Traces Introduction

Two of the central concepts in the MS464xB Series VNA family that will enable the maximum functionality of the system are channels and traces.

Channel Concept

At a high level, the channel defines the sweep configuration and the calibrations for a measurement. Sixteen channels are possible and each can have a different frequency range, different power levels, different IF bandwidths and different RF calibrations (among other things). In a sense, 16 distinct VNAs within one instrument are possible with each one executing sequentially.

Trace Concept

The trace is a lower level concept that represents a data group. Sixteen traces are allowed per-channel and each can represent a different response parameter, can be on a different graph type, and have certain different levels of post-processing applied to it.

The objective of this section is to explore how the channels and traces can be set up, what possibilities are available, and what configurations are commonly used. Many of the functions of the VNA can be separated into per-channel and per-trace groups; the delineation of these functional groups is also an important part of this section.

13-3 Channels and Traces General Concepts

The hierarchy of setups is illustrated in Figure 13-1. At the highest tier is per-system; these are variables that apply to all measurements on a given physical instrument. There are very few of these variables and they include:

- Step attenuator settings
- Bias tee on/off
- Certain portions of the hold system and certain triggering functionality
- AutoCal characterization files and cal kit files (are accessible from any channel)
- SnP and text file header/format setups
- Blank frequency display (security feature)
- Interface setup items (GPIB addresses, network config, touchscreen setup, …)

Per-System Variables

These variables are often per-system to prevent a setup scenario that could significantly shorten the life span of the hardware. In the case of hold and triggering, it also allows an entire measurement suite to be somewhat more easily controlled externally. Others fall more in the category of utilities that are somewhat per-system in nature. The hierarchy of system, channel, and trace setup information is shown below. Generally, N (the channel count) may be up to 16 (except when more than 25,000 points in a sweep are needed) and M (the trace count) may always be up to 16.

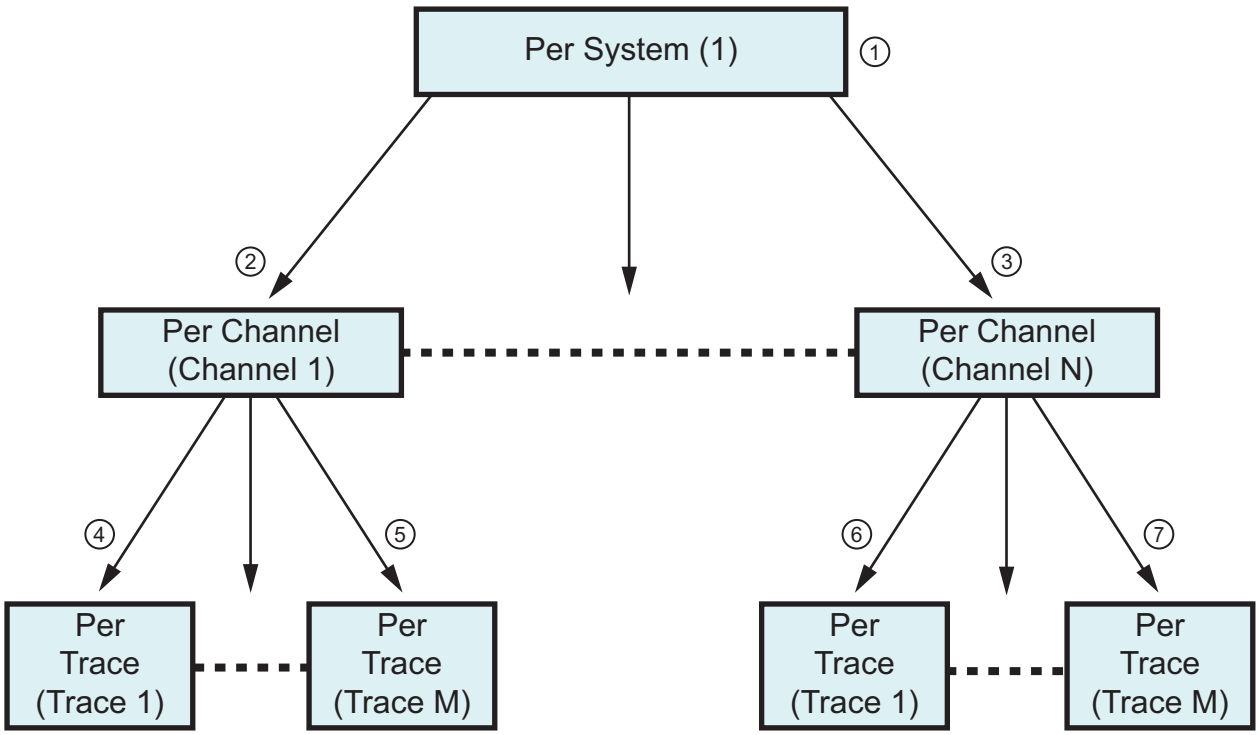


Figure 13-1. Setup Information Hierarchy (1 of 2)

Setup Hierarchy—System, then Channels, and then Traces 1. Per System Settings—These settings affect the entire instrument such as the maximum number of measurement points as either 25,000 or 100,000.	2. Per Channel Settings—If in 25,000 point mode, from 1 (one) to 16 channels can be configured. If in 100,000 point mode, only one channel is available. 3. Per Trace Settings—Each configured channel can have from 1 to 16 traces. Each trace can be configured as a separate measurements, with separate markers, and a different display method. Each trace can have up to 12 total measurement markers and one (1) reference marker.

Figure 13-1. Setup Information Hierarchy (2 of 2)

Per-Channel Variables

The second tier is that of the channel. As mentioned in the overview, the channel can almost be thought of as a separate virtual VNA. Although this term has been used differently in the past with other Anritsu VNAs, in the MS464xB Series VNA Series family, it includes a separate frequency list, separate calibrations and separate sweep control.

Per-Trace Variables

The third tier is that of the trace. As discussed above, this can be thought of as a data element (for example, S_{21} data for a given channel-based sweep setup). The per-trace flexibility mainly comes under the heading of post-processing and display.

Note	For historical reference, the 37xxxX family of Anritsu VNAs was limited to 1 channel and 4 traces. Although it may be somewhat confusing, in those instruments, what we now call traces were, then, called channels. This discrepancy was difficult to avoid when transferring to modern usage of the terminology.

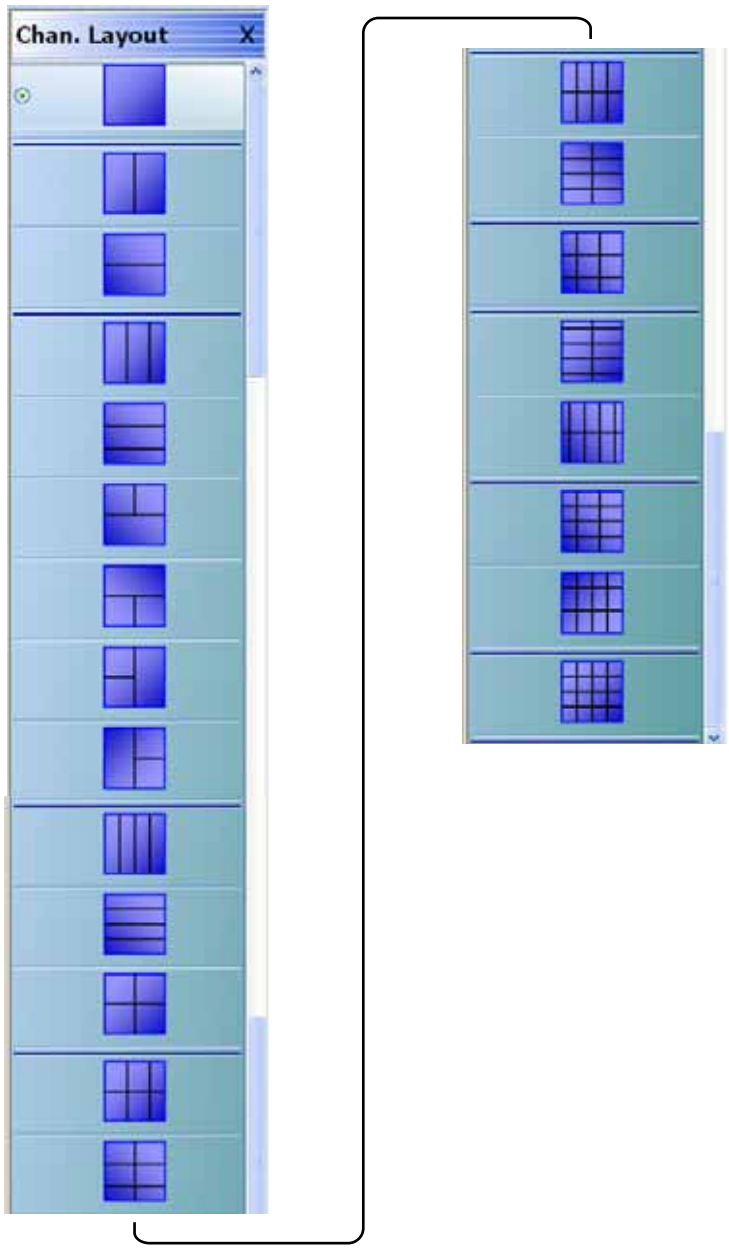
13-4 Channels and the Channel Menu

The channel menu itself is fairly simple. While the system defaults to 1 channel (under a preset command), here a different number of channels may be selected. The active channel (indicated by a thick white border) may be selected by clicking on that channel's window or it can be incremented using this menu (Chan. Next and Chan. Previous). The button Chan. Max can be used to force the active channel to occupy the entire data screen area. Note that the other defined channels will continue to sweep even when this mode is entered. The Chan. Max command is a toggle and can be undone by clicking it again. Chan. Max can also be accomplished by double-clicking the channel area away from a specific graph (that leads to Trace Max discussed later). These commands are illustrated in the top level menu shown in Figure 13-2.



Figure 13-2. CHANNEL Menu

Once a given number of channels is selected, the layout of those channels is selected in a submenu shown in Figure 13-3. Note that selecting a layout with more channels will update the channel count, since gaps in the sweep processing are not allowed. Because there are many combinations of channels possible, this window is quite lengthy (shown in two parts in Figure 13-3).



The instrument menu has a right-side scroll bar to access the entire menu.

Figure 13-3. CHAN. LAYOUT (CHANNEL LAYOUT) Menu

Number of Channels

The number of channels available first depends on the VNA maximum point setting which in turn depends on the VNA model number.

- If the VNA is set to 100,000 point mode, only one channel is available, with up to 16 traces.
- If the VNA is set to 25,000 point mode, up to 16 channels can be defined, each with up to 16 traces.

If multiple channels are available in the VNA, any number of channels can be configured between 1 and 16 channels.

Using the CHAN. LAYOUT menu shown in Figure 13-3 above, layouts are limited to values 1, 2, 3, 4, 6, 8, 9, 10, 12, or 16 channels.

Only those values corresponding to semi-symmetric layouts are allowed, as suggested by the channel layout menu, and other entries will be coerced.

Once the number of channels and the layout has been selected, it then remains to define each of the channels. The sweep control parameters apply to the active channel so one may cycle through the channels, entering values as needed. Alternatively, the setups for a channel may be copied through the setup save/recall mechanism since setups can be saved on a per-channel basis. Note that save/recall can also be applied to all channels.

It has been discussed that most sweep-setup parameters are per-channel in nature. To clearly delineate when these apply, the per-channel functions include:

- Frequency
- Power (although step attenuator settings, for units with Option 61 or 62, are per-system)
- Option 61—Active Measurement Suite with Two (2) Attenuators, Option 62—Active Measurement Suite with Four (4) Attenuators Calibrations (RF, power, receiver, etc.); this includes the calibration reference impedance
- Embedding/de-embedding and reference plane extensions
- Multiple source control settings (to include broadband/mmWave)
- Impedance transformations (note that this is distinct from impedance parameter conversions which are a post-processing calculation on trace data)
- Post-processing order
- Media type (dielectric constants, coaxial vs. waveguide, etc.)
- Hold functionality (portions can be per-system)
- Trigger functionality (portions can be per-system or per-port as well as per-channel)
- Analog out (rear panel)

Maximum Number of Points Setup

One additional behavior associated with channel count relates to the maximum number of points that may be defined in a sweep. Normally this limit is 25,000 pointsMS464xB Series VNAand this value can be used with any number of channels.

The VNA mode can be switched to a maximum of 100,000 points, but in this case, only one (1) channel is allowed. This mode is selected with the MAX. POINTS menu (shown below in Figure 13-4 on page 13-7) which is available at:

- MAIN | System | SYSTEM | Setup | SETUP | Max Points Setup | MAX. POINTS

This selection itself is, of course, per-system since the selection determines the number of channels that will be available.

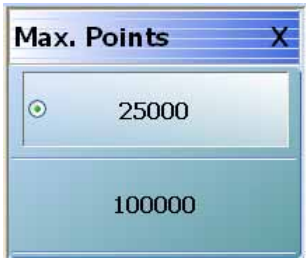
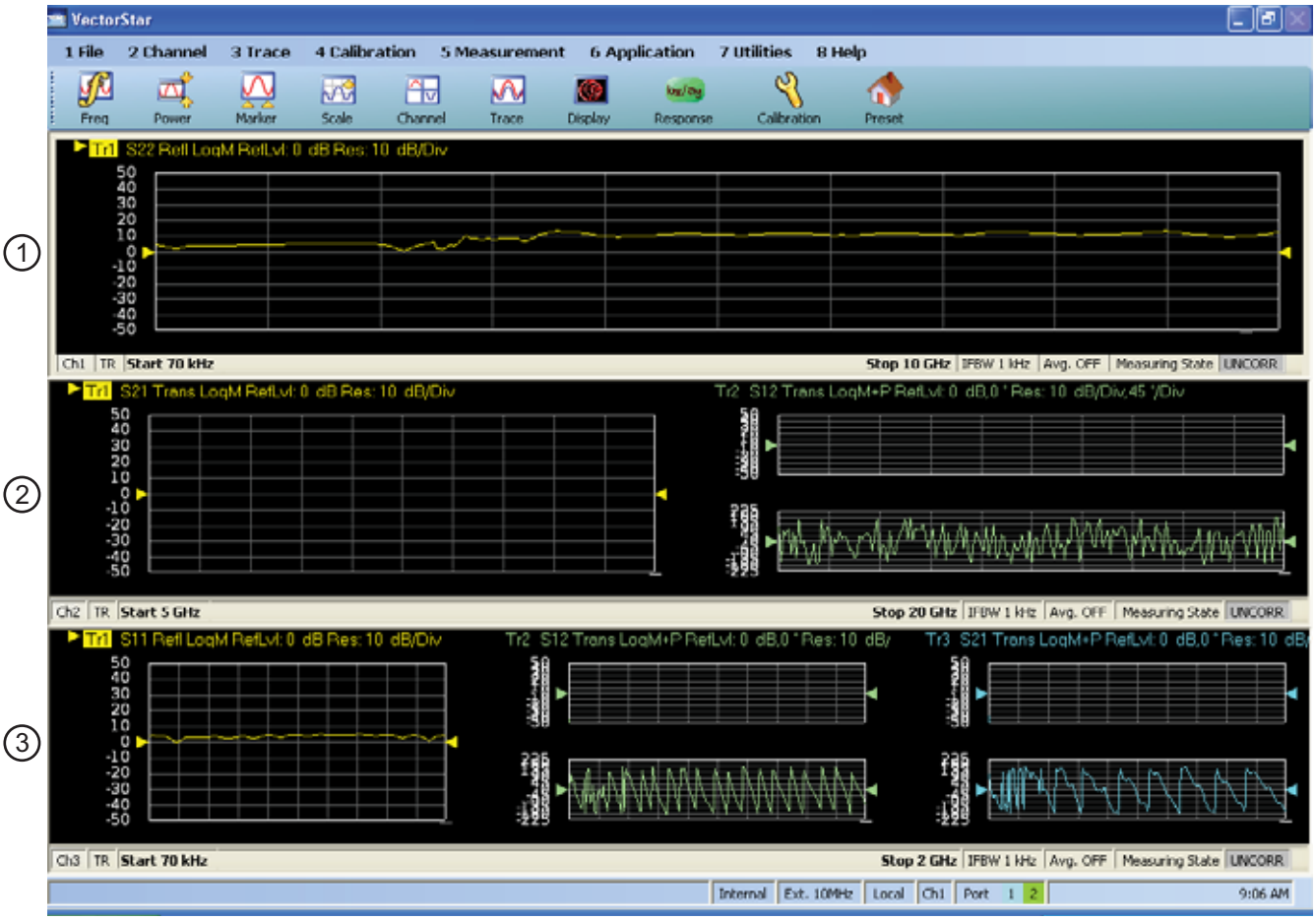


Figure 13-4. MAX. POINTS (MAXIMUM POINTS) Menu

Example Three-Channel Setup

An example of a three-channel setup is shown in Figure 13-5. In this case, all three channels have different frequency ranges, the same IF bandwidth was used in each, and the first channel (at Figure 13-5—#1 on the top or Ch1) is active as indicated by the thick white border.



1. Channel 1 (Ch1 at top) is the active channel is active (note white border) and measuring S_{22}.	2. Channel 2 (at middle) is measuring S_{21}. 3. Channel 3 (at bottom) is measuring S_{11}.

Figure 13-5. Typical Three-Channel Display

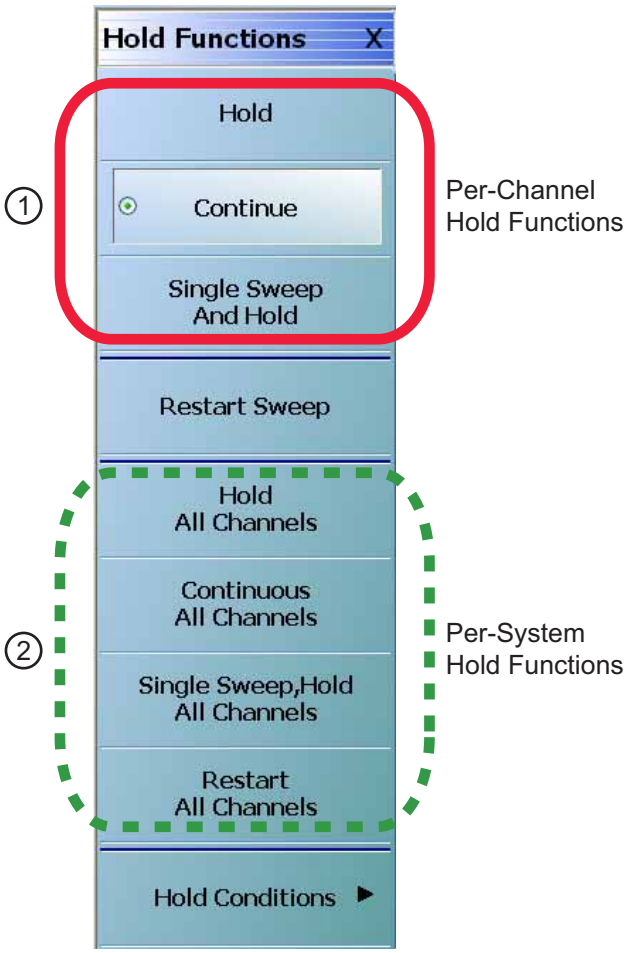
Hold Functions and Manual Trigger for Channels

Generally, the sweeps for each channel are executed sequentially. That is, all of the sweeps for Channel 1 are performed first, followed by all of those for Channel 2, and so on. Hold events and triggering events can be per-channel or per-system (and some subsets of the channel as well.) These menus are shown in Figure 13-6 for hold events and Figure 13-7 for triggering events. The HOLD FUNCTIONS menu is available from the SWEEP SETUP menu.

Navigation:

- MAIN | Sweep Setup | SWEEP SETUP | Hold Functions | HOLD FUNCTIONS

Note that while triggering can be done per-point and per-port, it still refers to the setup of the channel and it is never reduced to measurements related to a trace. As these hold and trigger functions are used, the actual measurement sequence can, of course, change. The HOLD FUNCTIONS menu is shown below to illustrate some per-system aspects (green dashed circle) of the hold function as opposed to the usual per-channel functions at the top of the menu (red circle).



1. Per-channel hold functions (at top).	2. Per-system hold functions (at bottom).

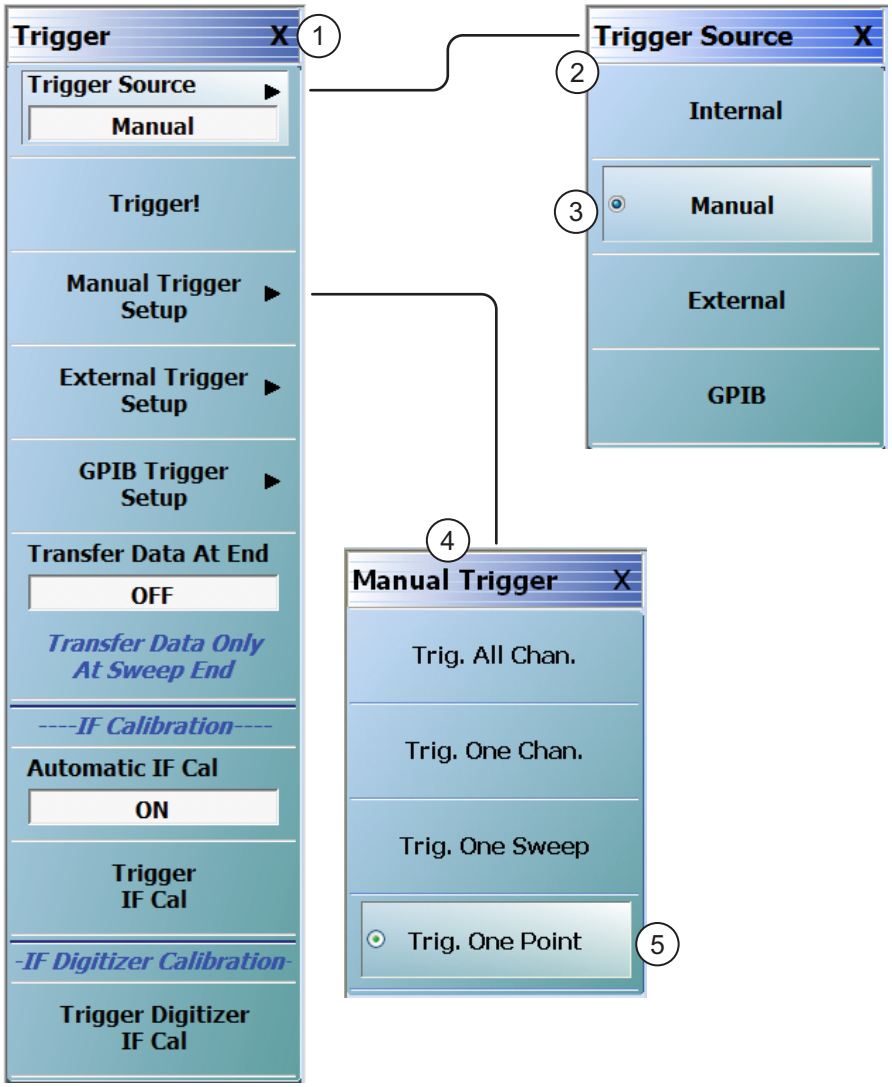
Figure 13-6. HOLD FUNCTIONS Menu

Trigger Source—Manual Trigger Menu

The Manual Trigger menu is shown below to illustrate per-channel behavior (second button), per-portion-of-channel behavior (third and fourth buttons), and per-system behavior (first button). Other trigger modes have analogous behaviors.

Navigation:

- MAIN | Sweep Setup | SWEEP SETUP | Trigger | TRIGGER | Manual Trigger | MANUAL TRIGGER
- For additional information on applying triggering functions via the Trigger Menu, see:
  - Chapter 25, "2PulseView™ (Option 35 and Option 42)"
  - Chapter 29, "Fast CW (Option 46)"



1. Trigger Menu	4. Manual Trigger Menu
2. Trigger Source Menu	5. Trigger One Point selected
3. Manual Trigger Selected	

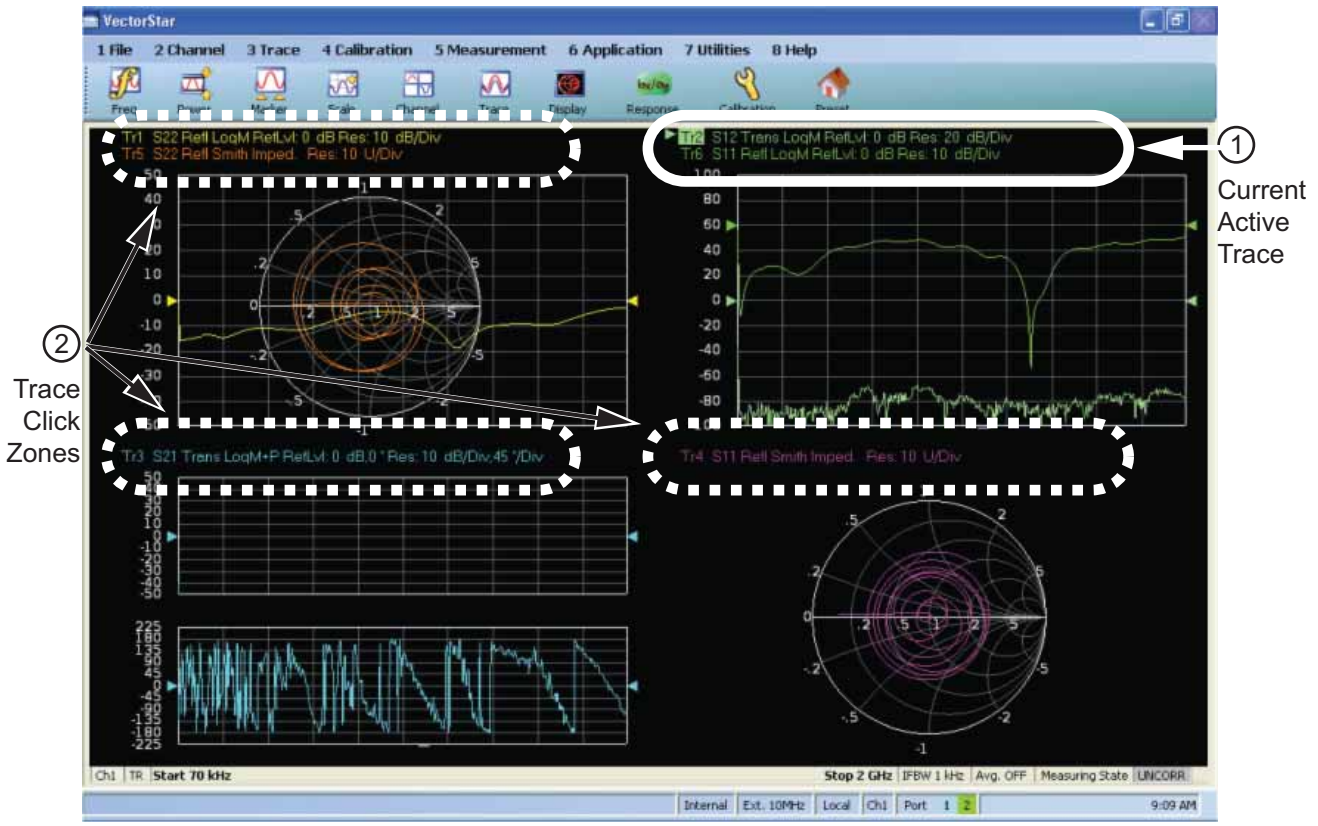
Figure 13-7. MANUAL TRIGGER Menu

13-5 Traces

In some sense, the main trace menu is a parallel of the channel menu. Again, the number of traces in the active channel may be specified and one can rotate through the traces using Trace Next and Trace Previous. One can also make a trace active by clicking on its annotation line as suggested by Figure 13-8. The active trace is denoted by an inverse color field around the trace number (Tr2 in Figure 13-8).

As with channels, the Max function causes the trace (in this case) to occupy the full graph region of the channel. Also as with the Chan. Max, this button is a toggle. Double-clicking on the graph area of the active trace will have the same function as clicking on the Trace Max button. The main Trace menu showing these functions is shown in Figure 13-9.

The area around the annotation line where one can make a trace active with a direct click is shown below. A double click near the active trace graph area maximizes that trace. Trace 2 (upper right) is currently active in this setup, as evidenced by the inverse video trace number.



1. Current active trace (at top).	2. Trace activate click zones (at left and bottom right).

Figure 13-8. Trace Activate Click Zones

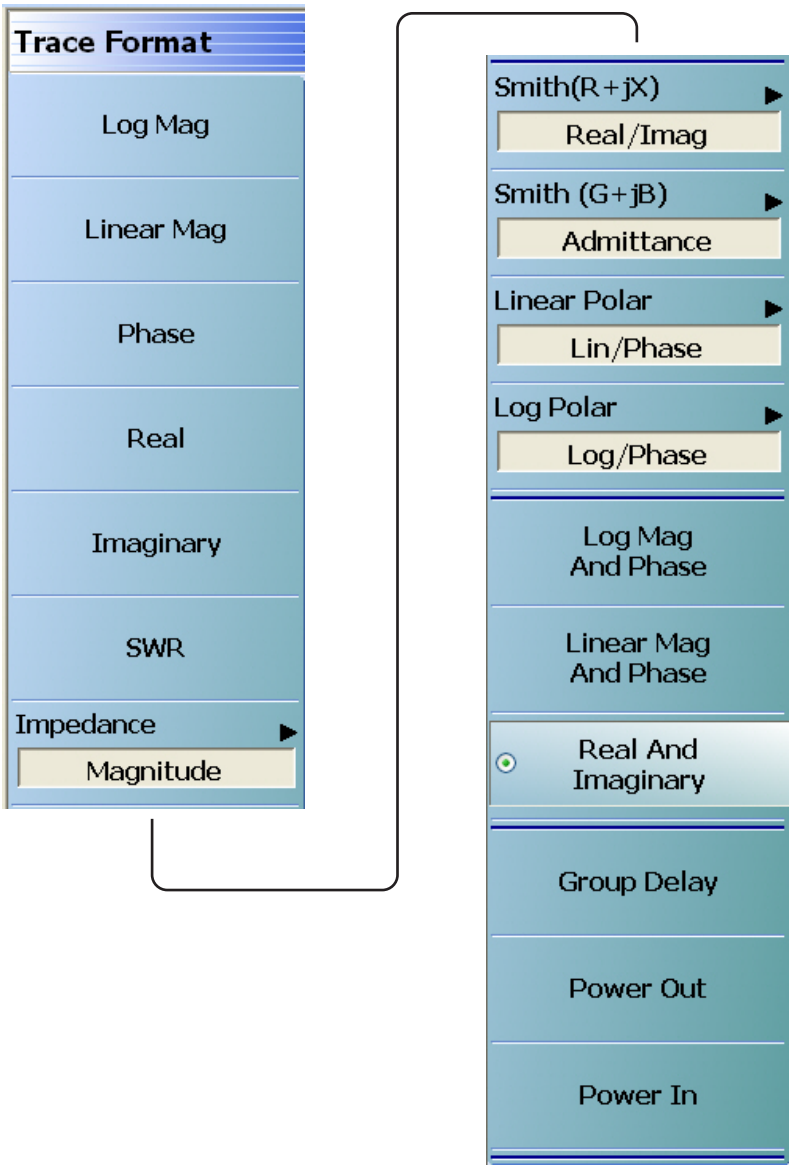
Trace Menu, Format, and Layout



TRACE menu with four traces selected

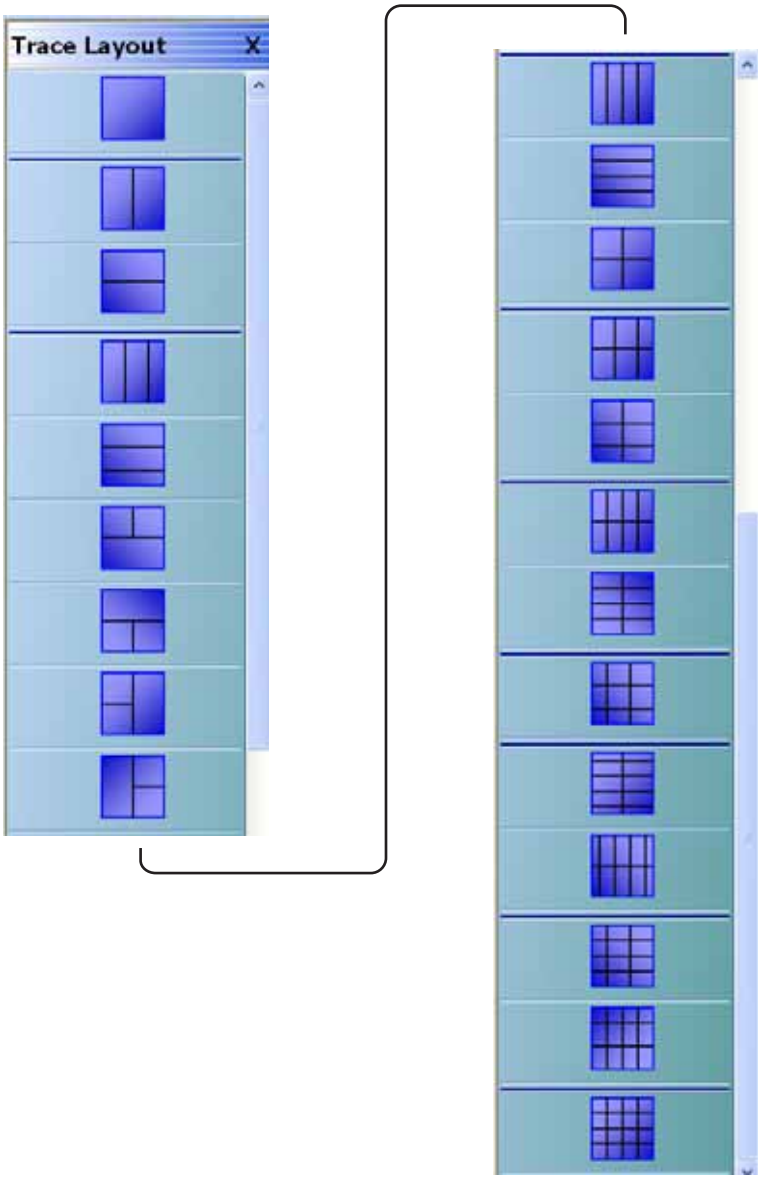
Figure 13-9. TRACE Menu

As with the channel system, there is also a layout submenu for traces (see Figure 13-10 for this long menu). Here, however, one may have any number of traces on any graph configuration and the number of traces will not be coerced based on the layout choice. If the number of traces exceeds the number of graph areas selected in the layout, the traces will be overlaid in a sequential fashion that will minimize the number of overlays. As with channels, only certain numbers of graph areas (in the case of traces) are allowed that are semi-symmetric (1, 2, 3, 4, 6, 8, 9, 10, 12, 16) but, unlike channels, blank areas in the graph area grid are allowed if the trace count is less than the graph count.



The actual TRACE FORMAT menu is continuous, with a right-side scroll bar.

Figure 13-10. TRACE FORMAT Menu



The actual TRACE LAYOUT menu is continuous, with a right-side scroll bar.

Figure 13-11. TRACE LAYOUT Menu

For example, if five traces are selected but the trace layout is chosen to be three graph areas (vertically oriented), then the traces will be assigned as follows:

- Top graph area: traces 1 and 4
- Middle graph area: traces 2 and 5
- Bottom graph area: trace 3

Of course, a 16-way overlay can be arranged by activating 16 traces but selecting the single graph area (top of the trace layout menu).

Once the number and layout of traces have been defined, usually each trace must be individually configured. In this sense, trace configuration generally applies to post-processing tasks, such as graph manipulation, and some analysis tasks, such as time domain, that are not directly tied to a sweep configuration.

Per-Trace Variables

Per-trace variables include:

- Trace format (graph type)
- Trace memory and math functions (to include inter-trace math which is sort of a hybrid but is defined on a per-trace basis)
- Scale (although autoscale can also be per-channel or per-system)
- Trace impedance (which is different from reference impedance set during a calibration)
- Domain (time domain and frequency w/ time gate)
- Smoothing
- Conversions
- Limit lines
- Markers (although markers can be optionally coupled between traces within a channel)
- Response (S-parameter, un-ratioed parameter, ext. analog in)

Complex Trace Setup Example

A fairly complex trace setup example is shown in Figure 13-12 below. This example covers multiple graph types and scaling options as well as different transformations applied to the data in certain traces. Here Trace 3 is in time domain, while the others are frequency domain. Most of the traces represent S-parameters, but Trace 6 is in an impedance conversion mode (reflection impedance, covered in another section of this measurement guide). The symbols in the annotation line provide information about these trace definitions. Some of the abbreviations that may appear in brackets in the annotation line are shown in Table 13-1 below. Details on these functions can be found in other sections of this measurement guide and the VectorStar Operations Manual.

Table 13-1. Trace Labels and Annotations

Abbreviation	Definition
FGT	Frequency domain with time gating
TLP	Time domain low pass
TBP	Time domain bandpass
D&M	Data and memory
D/M	Data memory math
D+M	
D-M	
D*M	
M	Memory
Tr1/Tr2	Inter-trace math (other trace numbers and other operators possible)
Zt	Conversions: impedance for transmission
Zr	Conversions: impedance for reflection
Yr	Conversions: admittance for reflection
Yt	Conversions: admittance for transmission
1/S	Conversions: reciprocal S-parameters
EQN	Equation Editor

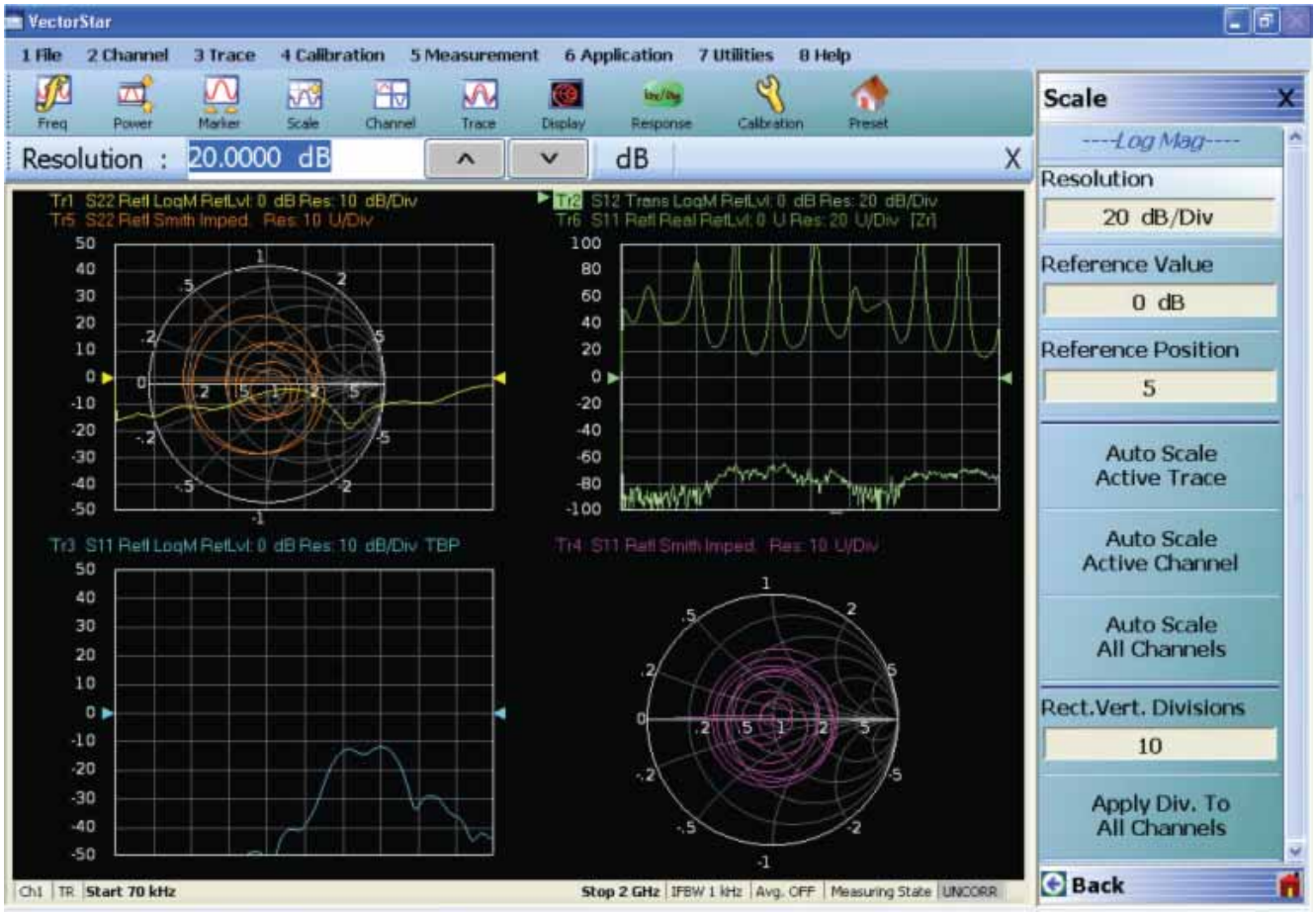


Figure 13-12. Multi-Trace Display with Right-side SCALE menu

Figure 13-12 above also illustrates an example of a function, auto scale in this case, available on per-trace, per-channel, and per-system levels.

13-6 Trace Math, Inter-trace Math, and Equation Editor

Although some trace manipulations were alluded to in the previous section, there are considerable trace computation capabilities that warrant further discussion. The simplest of these is TraceMemMath where some operation (addition, subtraction, multiplication or division) is applied between current trace data and that stored in trace memory for that trace. Some points of interest:

- The math is always applied on the basic, linear complex numbers….not on the displayed, formatted numbers.
  - Example: The trace is formatted as log mag + phase, the first data point is plotted at 20 dB, 0 degrees and the first memory point is plotted as 40 dB, 90 degrees. Data(/)Memory is selected. This will be computed as `(10+j0)/(0+j100)=-j0.1`. The result will be displayed as –20 dB, –90 degrees.
  - It does not matter if the entire complex value is being presently displayed (e.g., a real-only graph type), the math will be applied to the entire complex number.
  - If the user is commonly using log mag graph types, the math may appear counter-intuitive. If one is measuring S21 with a thru line connected, stores data to memory and selects Data(/)Memory, the immediate result is a flat line near 0 dB since one is dividing current data by something very similar (hence close to 1 in linear terms). If one selects Data(–)Memory, the result will be bouncing near the noise floor since now one is plotting the `20*log10(|X–Y|)` where X and Y are nearly identical.
- The math is applied after smoothing.
- The math may be applied before or after group delay, depending on the user selection for order of operations (this matters since group delay is computed as a derivative of phase, so if trace match comes first, it acts on phase).

The math is applied before self-normalization (as used in gain compression).

Now that the math is defined, the trace memory behavior needs to be discussed. Historically, Anritsu VNAs had one memory associated with each trace and they were not accessible from other traces or channels. If the number of points is changed, a stored memory will not be available (but will again be available if the number of points is changed back). Trace memory data (formatted or not) can be saved or recalled into that same trace memory location.

Now there is a more flexible definition where there are 20 trace memories per channel and they can be shared amongst the traces. The point count rules are unchanged. The default behavior is the same as the historic model in that a store operation from trace X will be to memory location X and, by default, `TraceMemMath` will use memory location X. The control dialog is shown in Figure 13-13.

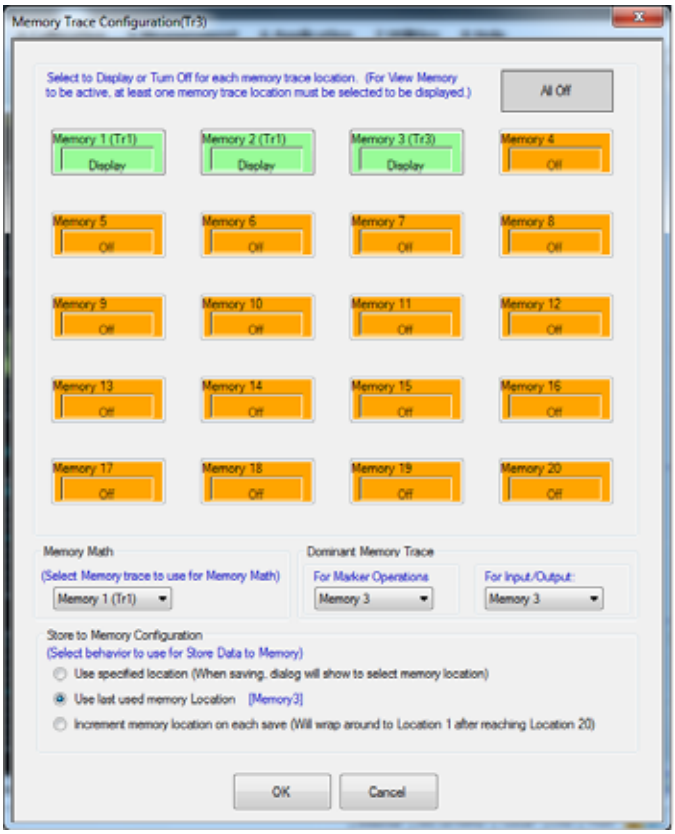


Figure 13-13. Memory Trace Configuration Dialog Box

The matrix in the upper portion of the dialog controls which memories are visible in the current trace (current trace denoted in the title of the dialog) when the display selection is either Memory or Data and Memory. Each box is a toggle button to display or not display the memory and a given button will only be active if something has been stored in that location. The trace where that stored data originally came from will be shown in parentheses (the first two memory locations have data originally from trace 1 and the third location contains data originally from trace 3 in the above example).

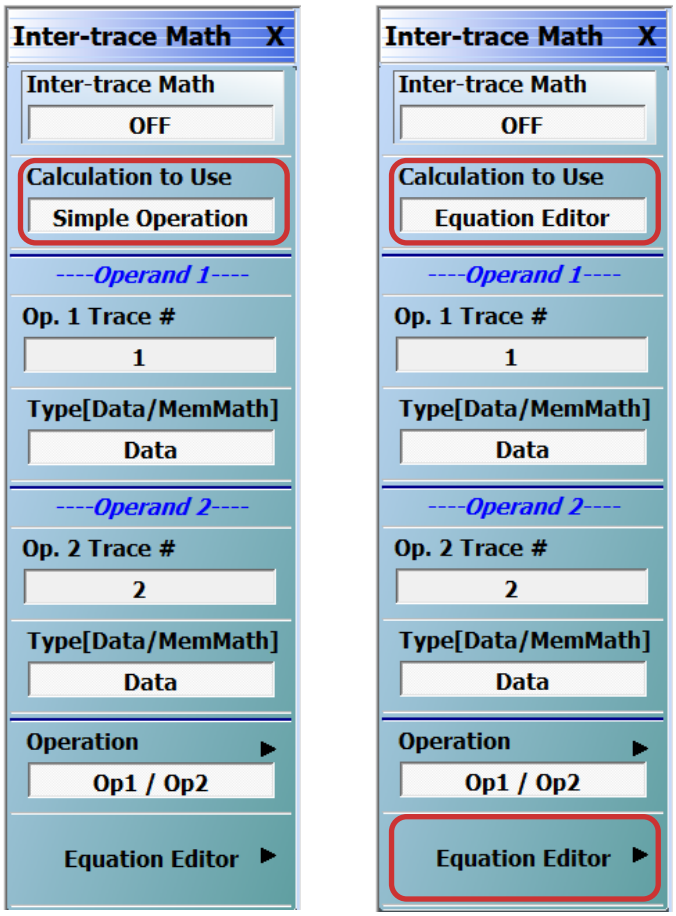
The Memory Math pull-down selects which memory will be used for a math operation of the form DataMemMath (discussed previously). The various locations will only be available if something has been stored in that location.

The Store To Memory Configuration section controls what happens when Store Data To Memory is selected. The default is to use the last used location. Thus if this dialog has not been used, it will default to using the location with the same number as the active trace. Another choice is to force a location selection dialog to appear every time data is stored. The final choice is to sequence through all of the memories beginning with the last used location for that trace. This selection is handy when one wants to capture evolutionary data from a DUT (with time, bias conditions, etc.) and it is inconvenient to save data to a discrete file each time. Note that when the 20 locations have been used up with this approach, the oldest stored information will be overwritten and the saving will continue cyclically.

Finally, certain tasks can only deal with one memory per trace so a 'dominant' memory can be selected for the active trace. One of these tasks groups are marker operations on memory (searching for targets, etc.). Another task group is saving and recalling of trace memory information (formatted or unformatted). These dominant memory locations default to the same numbered location as the current active trace and can be selected independently for the two task groups.

Inter-trace Math

This function allows one to combine measurements from two traces into a result to be plotted in a third trace position. In the Simple Operation mode, the concept is very similar to trace math just discussed, the available operators are the same and the operations again occur on the base, linear complex value. The difference is that the input variables can come from different traces (with completely different S-parameters or user-defined parameters) and can even be based on trace math in that different trace (hence the Type selection in the menu shown in Figure 13-14 on page 13-18).



Depending on the 'Calculation To Use' selection, either a simple (+ - * /) operation set with two arguments is allowed (Simple Operation) or a much more complete set of operations and arguments are possible (Equation Editor).

Figure 13-14. Inter-trace Math Menu

An example could be someone wanting to measure S21/S12 for a device as a measure of reciprocity (e.g., the device is an isolator). If trace 1 was configured as S21 and trace 2 was configured as S12, then the above settings for the Simple Operation would allow a display of the ratio in real-time. In contrast to basic trace math (where one could measure S21, save it to memory, then change the measurement to S12 and plot DataMemMath) which would be quasi-static, the inter-trace measurement version allows for observing changes during tuning correctly. On a more advanced level, one could normalize each of the input variables using DataMemMath on those traces and then operate on the normalized variables using Inter-trace math (e.g., Tr1/Mem1 + Tr2/Mem2).

Equation Editor

The Equation Editor allows a much more complete set of operations between trace data sets (and S-parameter sets) than does the Simple Operation inter-trace math just described. The main dialog is shown next (see Figure 13-15 on page 13-20 and Figure 13-16 on page 13-21) and consists of a selection of functions, input variables (traces and s-parameters in various formats, and sNp file selection) and scalar entry along with some editing tools.

A central concept is that the entire equation is based on complex vectors (which is how trace data and S-parameters come in and what is desired for plotting) of length equal to the number of points. Scalars (real or complex) can be used throughout but, where necessary, will be automatically vectorized (same value at each position in a vector of length equal to the number of points).

Example:

Trace 1 has three points [1+j1, 2+j2, 3+j3]. The equation is `Tr1+pi.` The result of the calculation will be [(1+pi)+j1, (2+pi)+j2, (3+pi)+j3].

Syntax errors will be flagged if parentheses are not used to resolve precedence problems (e.g., `Tr1 * -T2` will not be accepted but `Tr1 * (-Tr2)` will be).

If the input variable format is selected as Raw or Corrected, the variable will enter the equation as a linear complex number (either with or without calibration applied; note that receiver calibrations are applied to all). If Formatted is selected, the current graph type format will be used so the vector may be purely real. The formatted data can cause considerable confusion (e.g., taking log10 of a variable that is already in dB), but can be useful for some computations (e.g., computing 3rd order intercept in dB from product tone levels in dBm). Note that the graph type used for the equation output will act on the equation result directly (e.g., a logmag graph type will take 20*log10(abs(equation output))).

If the time domain checkbox is selected, all traces and parameters will be processed into time domain in the background if they are not already displayed that way. Lowpass Processing will be used if the current frequency list supports it, but otherwise Bandpass Process will be used. Trace time domain parameters will be used, which may be at default if not already configured. It is recommended to configure desired variables in time domain so the results are predictable. See the "Measurement—Time Domain (Option 2)" chapter of this guide for more information.

Note that trace memory and trace math (discussed earlier in this section) can be used as the incoming variables. Constant ***π*** (PI) is available and the '**j**' button is used for entering complex scalars. The scientific notation exponent marker '**E**' is also available (e.g., 1E9 for 1000000000).

Only single-ended S-parameters are available as direct arguments for the editor. Any mixed-mode S-parameter can, however, be created as a trace variable and used that way as an argument. This is because of the wide variety of underlying port permutations possible with the mixed mode parameters.

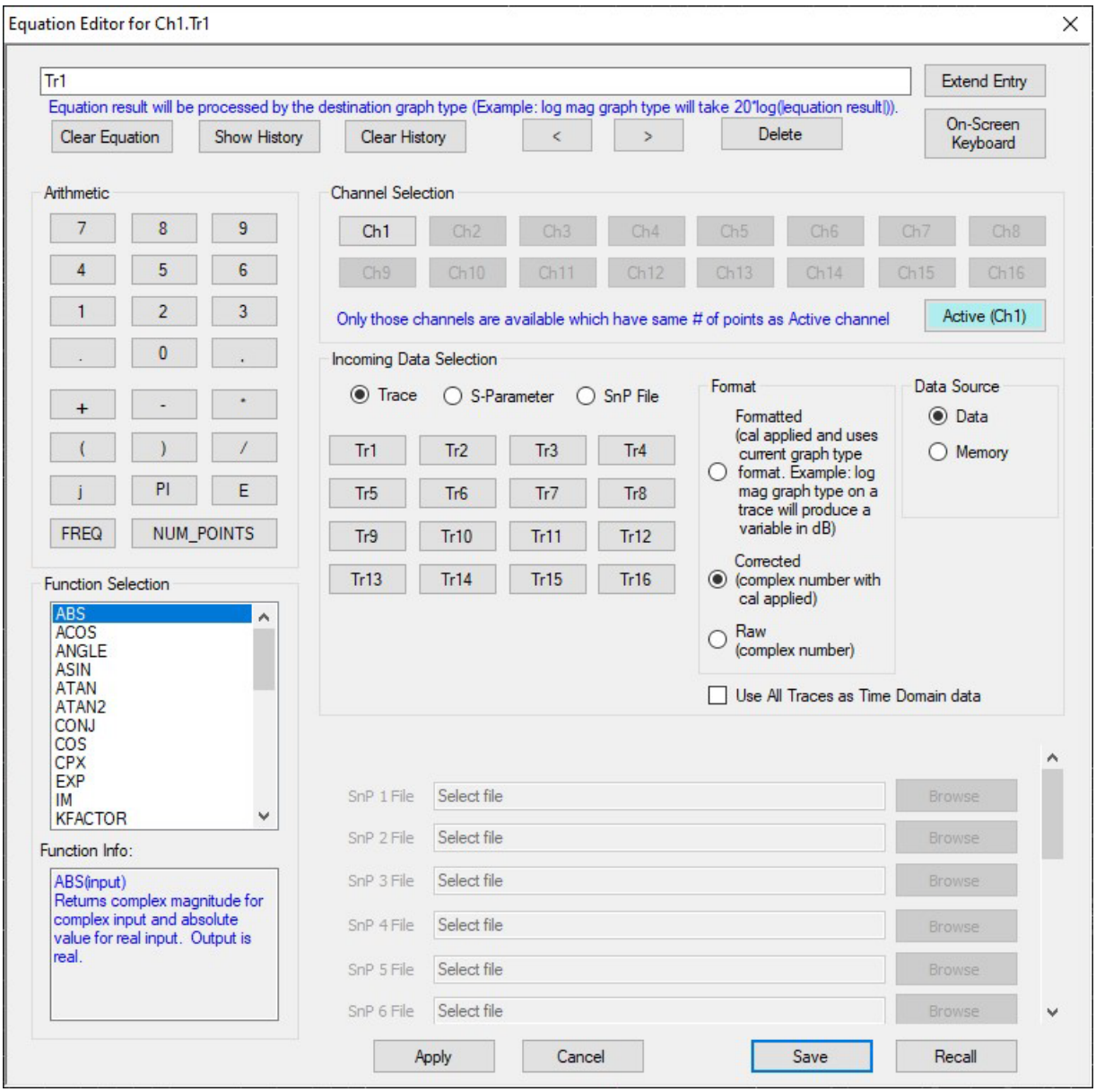


Figure 13-15. Equation Editor Dialog—Trace Mode

Note that S-parameters from .sNp files may also be used in the equations (see Figure 13-16). Up to 16 files may be used and all S-parameters within a given file (for N up to 4) can be used. Interpolation and flat-line extrapolation will apply when the file frequency range does not match the current frequency range. The data from the files is stored as part of the VNA setup files (and during application shutdown) so the original file location does not need to be maintained. The original file name will continue to appear in the dialog as an aid to remember the nature of the data.

Data (and memory and processed results) from other channels may also be used in the calculation for the active channel. This is particularly useful when data over other sweep ranges is needed (e.g., values of various harmonics for computation of total harmonic distortion). It is required, however, that all channels being used in the equation must have the same number of sweep points so that the vector lengths will be commensurate (and a channel will not be available for variable inclusion if it has a different number of sweep points). The various channels do not have to use the same frequency range, although they commonly will be related (e.g., by a harmonic). Also, the data from the various channels being used in an equation will all be converted to time domain if that checkbox has been selected for the current equation.

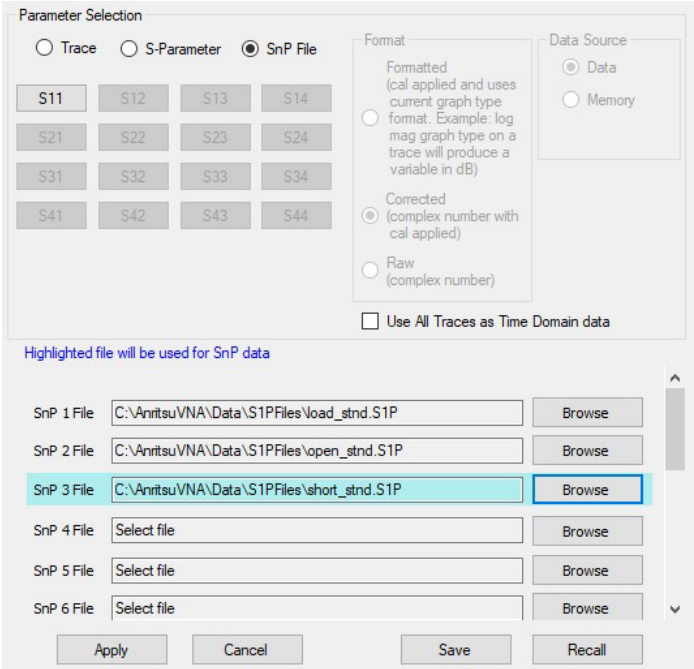


Figure 13-16. Equation Editor Dialog—sNp File Selection Mode

Supported Complex Functions

Following are description of the more complex functions supported (the output of the function is complex unless otherwise noted).

- ABS()—Complex magnitude for complex input and absolute value for real input. Output is real.
- ACOS()—Arccosine; radian output. This will accept complex arguments and uses the standard branch cut.
- ANGLE()—Phase of complex input; radian output. Output is real.
- ASIN()—Arcsine, radian output. This will accept complex arguments and uses the standard branch cut.
- ATAN()—Arctangent, radian output. This will accept complex arguments and uses the standard branch cut.
- ATAN2()—Arctangent with the ability to properly resolve quadrants. The argument is complex and it is internally split into real and imaginary components with sign checking. Radian output
- CONJ()—conjugate
- COS()—Cosine, radian input. Note that this function will accept complex inputs and treat them as such. Commonly one would use this function only with a formatted trace set up for phase and then multiplied by pi/180 to convert to radians.
- CPX(a,b)—Complex equivalent taking 2 real inputs; output is a+jb. If the inputs are complex, the real part of each is taken prior to combination into a new complex variable.
- EXP()—Exponential

- IM()—Imaginary part of a complex input. Output is real.
- KFACTOR()—Stability factor accepting 4 complex inputs (generally representing S11, S12, S21, and S22).

  KFACTOR(Tr1,Tr2,Tr3,Tr4) produces:

$$\frac{1-|Tr1|^2-|Tr4|^2+|Tr1\cdot Tr4-Tr2\cdot Tr3|^2}{2|Tr2\cdot Tr3|}$$

Output is real.

Equation 13-1.

- LN()—Natural log; standard branch cut
- LOG10()—Log base 10; standard branch cut
- MAG()—Magnitude accepting complex input (same as ABS). Output is real.
- MAX()—Maximum value of the MAGNITUDE of the variable selected. (Note that this updates only after a sweep completes so there may be a one sweep delay until the value propagates to a plotted equation). Output is real.
- MAX_HOLD()—Accumulates maximum value of the MAGNITUDE of the argument sweep-to-sweep. The process is reset by clearing the equation or turning inter-trace math off. (Note that this updates only after a sweep completes so there may be a one sweep delay until the value propagates to a plotted equation). Output is real.
- MEAN()—Average value in a complex sense; (note that this updates only after a sweep completes, so there may be a one sweep delay until the value propagates to a plotted equation)
- MEDIAN()—Median value of the MAGNITUDE of the argument; (note that this updates only after a sweep completes, so there may be a one sweep delay until the value propagates to a plotted equation). Output is real.
- MIN()—Minimum value of the MAGNITUDE of the argument …(note that this updates only after a sweep completes, so there may be a one sweep delay until the value propagates to a plotted equation). Output is real.
- MIN_HOLD()—Accumulates maximum value of the MAGNITUDE of the argument sweep-to-sweep. The process is reset by clearing the equation or turning inter-trace math off. (Note that this updates only after a sweep completes, so there may be a one sweep delay until the value propagates to a plotted equation). Output is real.
- MRKX()—Readout of active maker on entered trace, x-value. If no marker is on, a 0 will be returned. If more than one marker is on, the active marker will be used. Output is real. Since this function relies on a trace marker value, the argument can be ONLY a trace and not a function involving a trace.
- MRKY()—Readout of active maker on entered trace, y-value. If no marker is on, a 0 will be returned. If more than one marker is on, the active marker will be used. Since this function relies on a trace marker value, the argument can be ONLY a trace and not a function involving a trace.
- MU()—mu stability factor accepting 4 complex inputs (generally representing S11, S12, S21, and S22).

  MU(Tr1,Tr2,Tr3,Tr4) produces:

$$\frac{1-|Tr1|^2}{|Tr4-Tr1^{*}(Tr1\cdot Tr4-(Tr2\cdot Tr3))|+|Tr2\cdot Tr3|}$$

(where the * denotes conjugate)

Output is real.

Equation 13-2.

- PHASE()—Same as ANGLE but degree output. Output is real.
- POW(z,n)—Raises a complex variable z to the nth power. n is a scalar.
- RE()—Returns real part of a complex input. Output is real.
- REWRAP()—Rewraps phase of a complex variable when range was truncated (often by a power function). The calculation is based on slope of low frequency data.
- SDEV()—Standard deviation of input data. This is evaluated only at sweep completion, so there may be a one sweep delay for values to propagate to a displayed equation. This calculation is based on the equation below where N is the number of points. Note that the output is real.

$$SDEV = \sqrt{\frac{\sum_{k=1}^{N} \left| y_k - MEAN \right|^2}{N-1}}$$

Equation 13-3.

- SIN()—Sine. Note that this function will accept complex inputs and treat them as such. Commonly one would use this function only with a formatted trace set up for phase and then multiplied by pi/180 to convert to radians.
- SQRT()—Square root; standard branch cut
- SUBSET—(scalar, scalar, vector) returns a subset of the trace/parameter variable that is the third argument. The first two arguments are the point count indices (starting from 0) of the start and stop points of interest.
- TAN()—Tangent. Note that this function will accept complex inputs and treat them as such. Commonly one would use this function only with a formatted trace set up for phase and then multiplied by pi/180 to convert to radians.
- XAXISARRAY()—Generates the vector corresponding to the current sweep variable. Output is real.

Example Using the Equation Editor

Power-added efficiency measurements (PAE) are a simple example of how the equation editor can be used. This measurement makes use of the Analog In ports to measure the DC power consumption of a DUT during a sweep and uses it to compute DUT efficiency so it can also be plotted with other sweep data. Power-added efficiency (PAE) and related metrics are important quantities used in the assessment of amplifiers as well as other devices and subsystems/systems. The efficiency of the use of DC power for RF power generation is critical for the overall performance of many systems and increasingly has a considerable influence on product design. It may be useful to directly measure this quantity on a VNA during other measurements and tools exist to make this feasible in the VectorStar family of instruments. The purpose of this note is to show some of the possible hardware configurations of this in-line analysis as well as how the instrument software can be set up to perform the calculations and display/output the results.

The Method

While many definitions exist for PAE, in the end it will be some expression of output RF power relative to input DC power. There may be differences in if gate power dissipation is included or not or at what plane the RF powers are measured, etc. For the purposes of this note, we will use a common definition of:

$$\frac{(P_{RFout} - P_{RFin})}{P_{DC}}$$

Equation 13-4.

Where all powers are expressed in consistent non-logarithmic units (e.g., all W or all mW), the DC power is that fed to the drain or collector terminals (or any single measurable terminal) and the result is expressed as a percentage.

The reader may be familiar with using calibrated wave variables in a VNA to measure RF power: commonly $|b_2|^2$ for the 2-port measurement where the DUT output is at port 2, and $|a_1|^2$ for the input power. A receiver calibration is required to get an absolute power value for these quantities and the units will normally be mW (and this process absorbs reference impedance terms normally seen).

The DC power measurement process using a VNA may be less familiar but there are two analog input ports on the VectorStar that can be used to measure low frequency quantities synchronized with the sweep. With an appropriate sense resistor, or other current-to-voltage converter, one of these can be used to measure DC current supplied to the DUT. A fixed voltage can be assumed (to compute DC power as I*V) or the second analog input can be used to measure voltage at some relevant node. Note that the two analog inputs multiplex to a single analog-to-digital converter so, if both are used, the sweep will slow down as the system measures both inputs at each sweep point. Thus, if one of the DC parameters (current or voltage) can be assumed constant, it will save some measurement time if that parameter is not actively measured.

Some other important notes about the analog inputs:

- Input impedance: 60K ohms
- Range: –10V to +10V
- Accuracy: a slope and offset calibration is available resulting in a typical accuracy of 2 mV + 2%. Example: for a 1V input, the expected accuracy would be +/- 22mV. The user can refresh this Analog In calibration at any time and is recommended when there are significant temperature changes and low-level signals are being measured.
- Connectors: BNC(F); outer shield is chassis ground (see notes below on sense resistor measurements).

The Analog In measurement is activated by assigning one of the analog in response variables to a trace (instead of, for example, assigning an S-parameter or a wave variable to that trace). These selections are available at the bottom of the response menu as shown in Figure 13-17 on page 13-25 where the response menu in shown where external Analog In signals can be selected for analysis. Note that a driving port must be linked to the analog in variable which ensures that the variable is measured during the desired RF sweep.

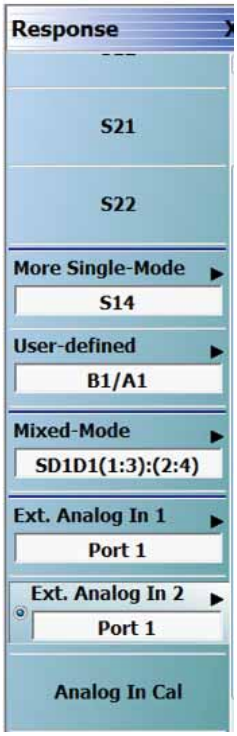


Figure 13-17. Response Menu

The resulting Analog In value generated for plotting is always real so a Real graph type is recommended for viewing. A Lin Mag graph type is also possible if polarity is not of interest. Imaginary graph types will always yield zeros and phase graph types will produce 0 degrees or 180 degrees. The real (or linear magnitude) graph type marker readouts are labeled in Units (U) which should be interpreted as Volts (V) for Analog In.

One note on measuring current: since the analog in inputs have a ground reference, one cannot usually measure the voltage across a sense resistor directly using one input. Instead, one can measure voltages at both ends of the sense resistor, using the two inputs, and get both voltage and current.

For certain measurement setups with high bias network impedances, the 60kohm input impedance may load the circuit. In these cases, a buffer amplifier or other transforming structure may be needed.

The Calculations

The user may export data of $|b_2|$, $|a_1|$ and the desired analog in variables and compute PAE themselves (where those variable names assume a 2-port setup where the DUT input is port 1 and the output is port 2). If one assumes current was measured with a current-to-voltage converter with a transimpedance of one ohm (using AnalogIn1) and the voltage was measured directly on AnalogIn2, the PAE would be given by the following if needed receiver calibrations are in place:

$$\frac{(|b_2|^2 - |a_1|^2)}{1000 * AnalogIn1 * AnalogIn2} x100$$

Equation 13-5.

Here the numerator will be reporting in mW (and the wave variables should be exported in a Lin Mag format or converted to that). Since the AnalogIn1 will represent current in Amps by virtue of the sense resistor, a factor of 1000 is needed to also get the denominator into mW. The final factor of one hundred is to convert PAE into a percentage.

If, instead, a sense resistor (of value R ohms) was used, the DUT-side voltage was measured on AnalogIn2, and the supply-side voltage measured on AnalogIn1, then one would calculate:

$$\frac{(|b_2|^2 - |a_1|^2)}{1000 * (AnalogIn1 - AnalogIn2) * AnalogIn2/R} x100$$

Equation 13-6.

The Equation editor in VectorStar can also be used to implement this computation and plot the result in an additional trace. That formulation of the previous equation (assuming Tr1 is a_1, Tr2 is b_2, Tr3 is AnalogIn1, Tr4 is AnalogIn2, and the sense resistor is one ohm) is shown in Figure 13-18 on page 13-27 where the Equation Editor is setup for computing PAE using a sense resistor is shown here.

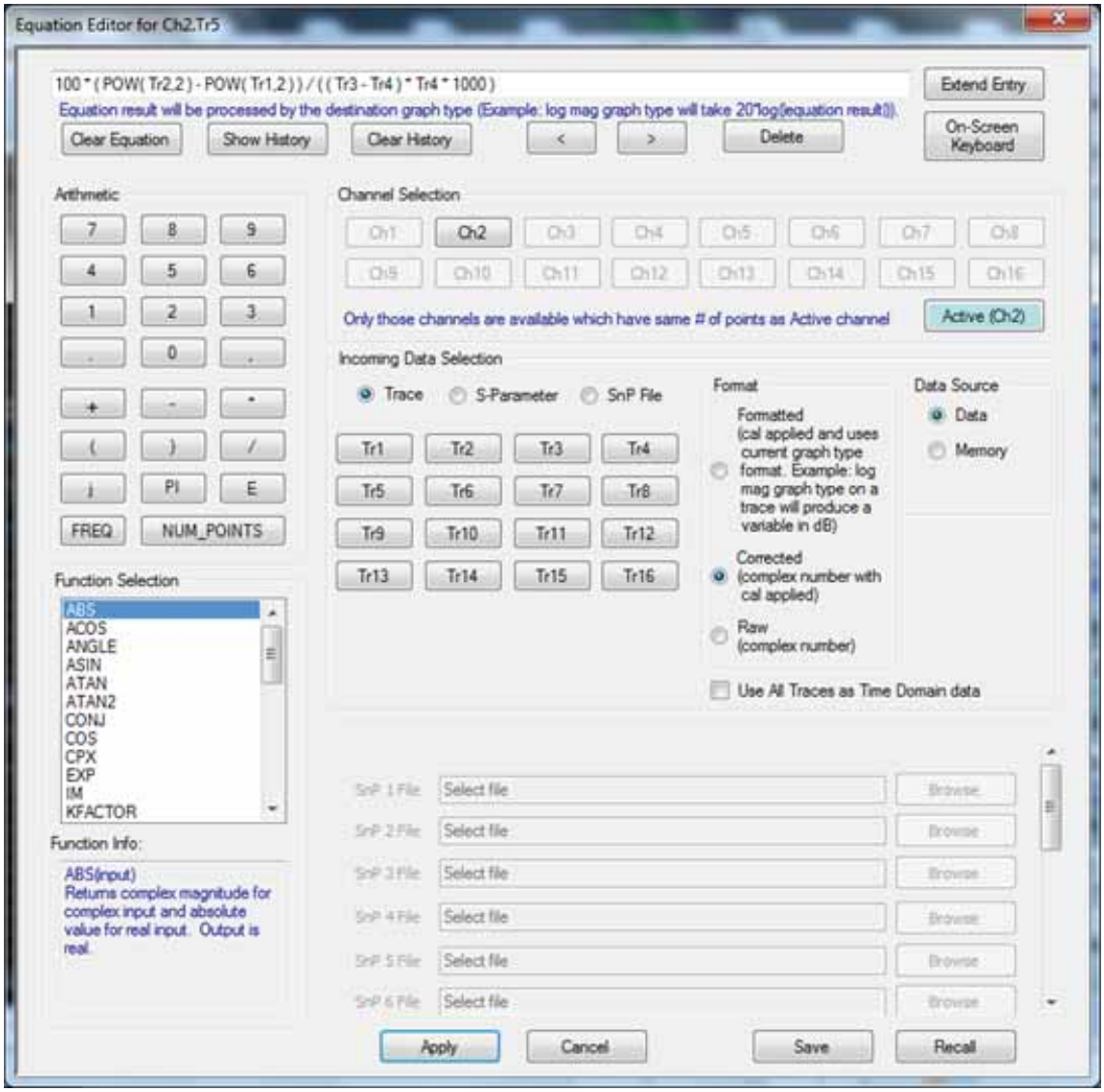


Figure 13-18. Equation Editor

Note that formatted variables were used for the equation which means the graph types for those traces need to be set correctly (Lin Mag for the wave variables and Real or Linear Magnitude for the Analog In variables). More information on using the Equation Editor is available in "Trace Math, Inter-trace Math, and Equation Editor" on page 13-16.

Example

An E-band LNA was the test device (and its PAE was expected to be very low) and it was of interest to see how PAE varied with RF drive power at 89 GHz. This measurement made use an ME7838D broadband system to get the frequency coverage, but the ideas here apply for any frequency range and any VectorStar family model.

Receiver calibrations were performed for a1 and b2 receivers as they will represent input RF power and output RF power, respectively. These measurements were assigned to traces 1 and 2. A 1 ohm sense resistor was used, and the DUT-side voltage was connected to AnalogIn2. The supply-side voltage was connected to AnalogIn1. The RF power was swept from –20 to +5 dBm. The graph types and response variables were assigned as discussed in the previous section and the measurements are shown below. The equation editor output was assigned to the trace at the bottom of Figure 13-19 (changed to have a trace name of PAE). About 0.5V is dropping across the sense resistor so DC power dissipation is about (0.5V/1ohm)*7.5V=3.75W. Given the <10mW max power output of the device, PAE will be low and indeed it is (at little over 1% at high drive). The result of an example PAE measurement of an E-band amplifier in a power sweep is shown in Figure 13-19 (in the bottom trace, expressed as a percentage)..

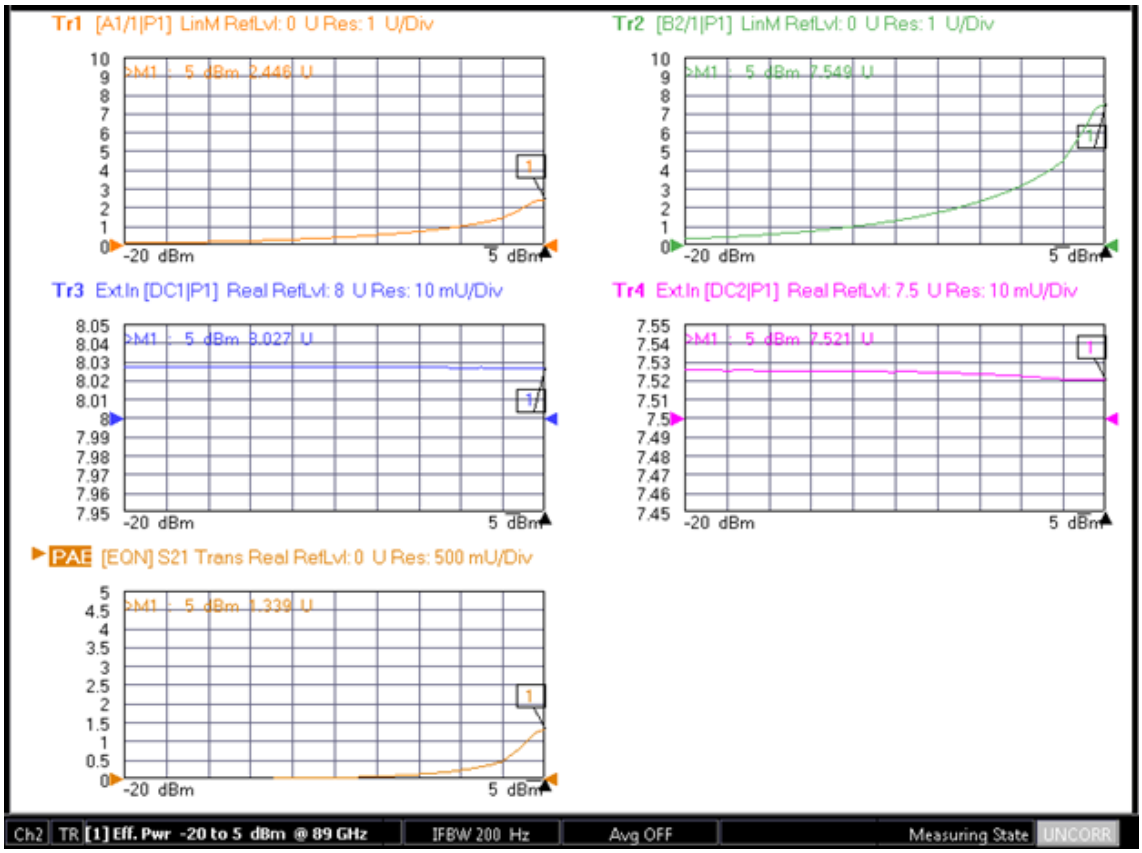


Figure 13-19. PAE Measurement

One can see that the DC power consumption is increasingly slightly at higher drive levels (the DUT-side voltage on the sense resistor in Tr4 is dropping so current draw is increasing) but this is outweighed by the increase in RF output power so the efficiency increases.

A Real graph type was used for the equation editor output which is a desirable choice given the calculation and the value should be read as a percentage because of the 100x factor in the equation.

13-7 Limit Lines

Limit lines are a powerful tool to help quickly compare a set of measured DUT data against specifications or expectations. All limit testing is per trace and, depending on firmware version, limit testing may only be available on rectilinear graph types. Upper and lower limits on any parameter may be set and these may be separated into many frequency bands. There is a limit of a total of 50 segments (upper and lower combined) per trace. The main limit line menu is shown in Figure 13-20.

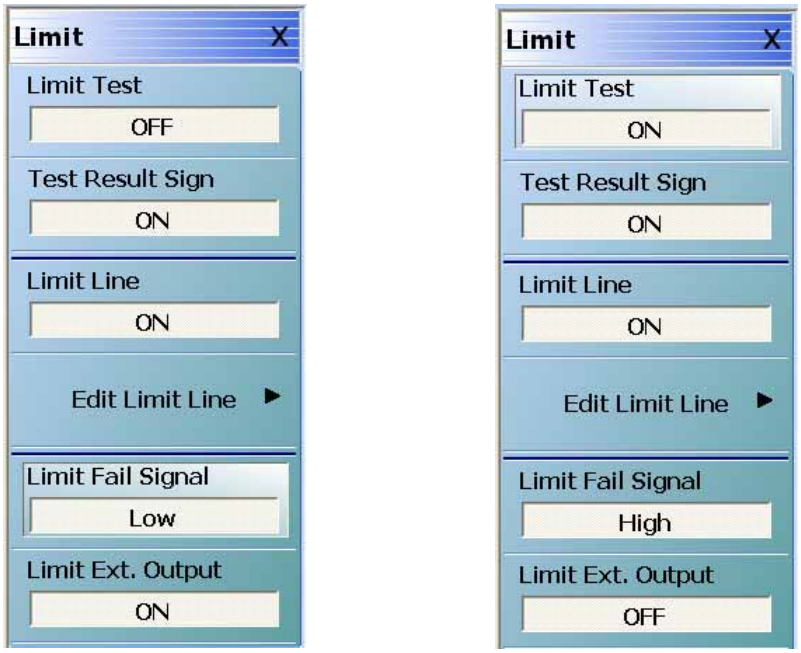


Figure 13-20. LIMIT Menu—Various Functions Toggled ON or OFF

The toggle buttons on the top level of this menu are straightforward:

Limit Test

The Limit Test button enables comparison of the data to the limit lines existing (this is per trace). The results of the test (pass or fail) will appear in the upper right corner (see Figure 13-22) of the graph for that trace.

Test Result Sign

The Test Result Sign button enables a large graphic displaying the pass/fail result. This will be in the middle of the screen (see Figure 13-22 on page 13-30 below) and is visible from a large distance. The Limit Test must be on for this sign to appear. If any limit tests fail, the large fail sign will appear with a notation of which channel has failed.

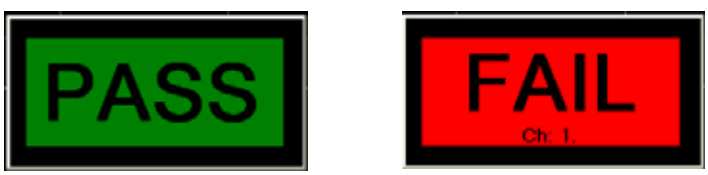


Figure 13-21. Pass and Fail Signs Configured by the LIMIT Menu

Limit Line

Displays the current limit lines on the data graph. The limit lines will appear in red. Failing points are marked with a red dot.

Limit Fail Signal

Determines the state of the external limit status bit for a fail condition (see next item). High or Low (in a 3.3V logic sense).

Limit Ext. Output

The Limit External Output button enables a signal on pin 1 of the rear panel **External I/O** connector that will change state on limit failure. If any activated limit test fails, the bit will go to the fail state. This bit will be active only if Limit Test is ON. In the figure below, the small annunciator in the upper left appears if Limit Test is ON, the large center graphic appears if Test Result Sign is ON, and the red limit lines appear if Limit Line is ON. If a test on a channel fails, the large center graphic includes a notation of which channel has failed.

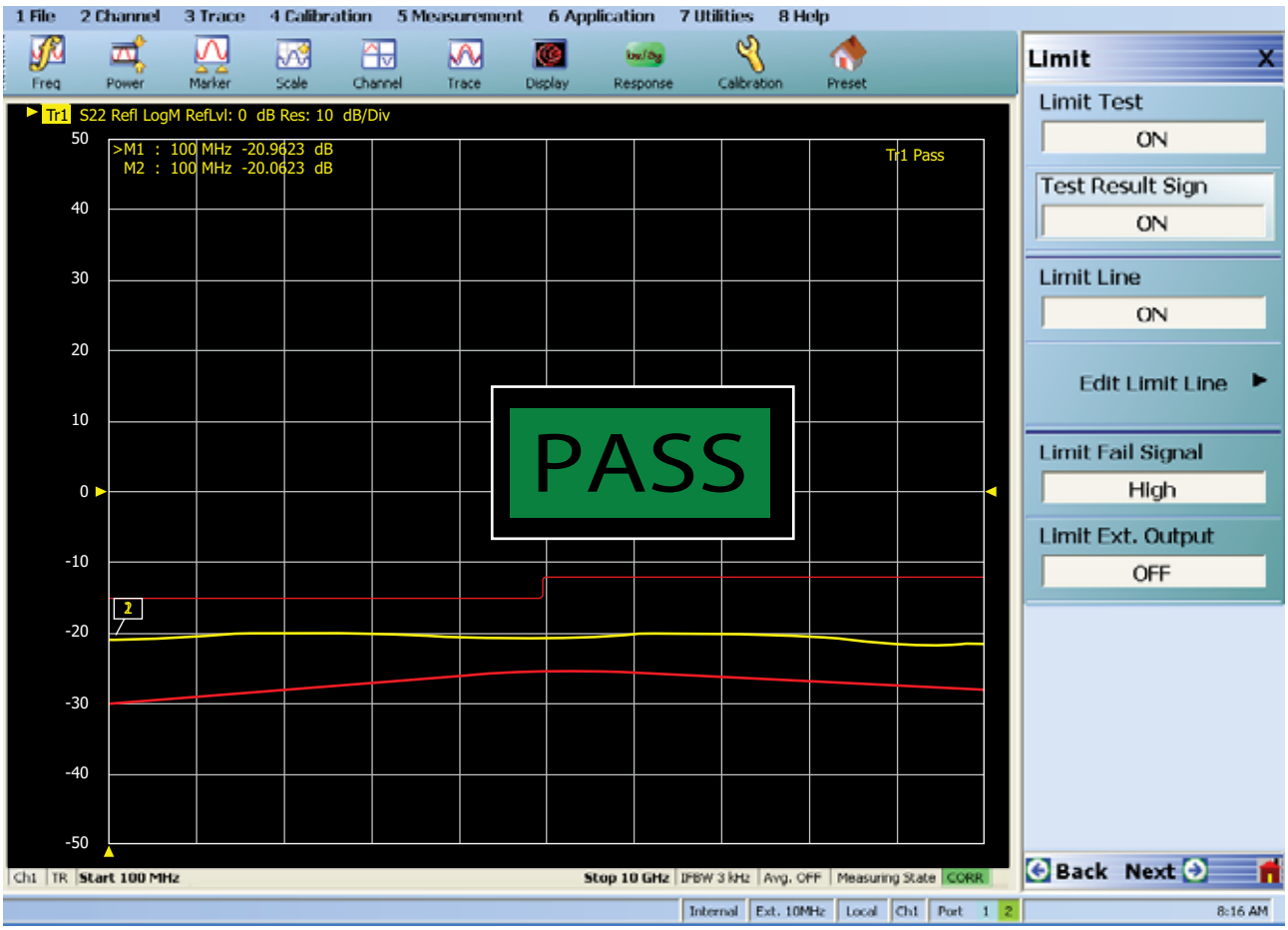


Figure 13-22. LIMIT Menu and Limit Pass Sign Example on Main Display

A more complex, multi-channel example showing some of these concepts is shown in Figure 13-23. Note the dependence of the Test Result Sign on the limit testing going on in multiple channels. Only Trace 1 on Channel 2 has failed, but the overall result will be reported as a failure.

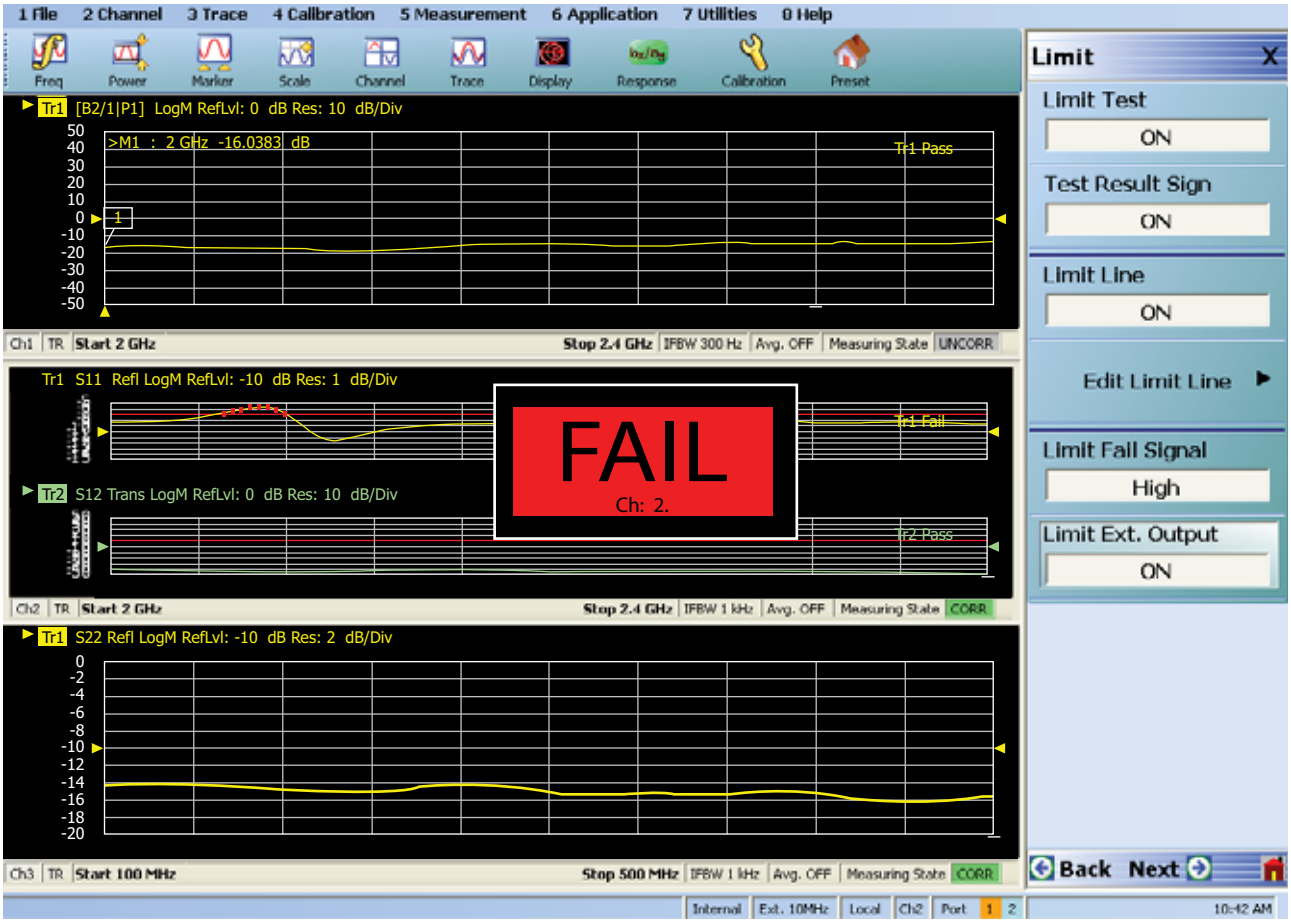


Figure 13-23. LIMIT Menu and Limit Fail Sign Example on Main Display

Editing Limit Lines

The editing of the limit lines is controlled on the one submenu and that is shown in Figure 13-24. When entering this menu, the limit line table will appear at the bottom of the screen (not unlike the multiple source and segmented sweep tables). Initially, the table will often be empty. If a limit line set was created on another trace, those values may appear here but they may be cleared or edited. The limit line tables may be saved and recalled separately using this menu (much like segmented sweep tables) or they may be saved and recalled as part of the global setup (use commands under the File menu to do this).

Edit Limit Line Menu



Figure 13-24. EDIT LIMIT LINE Menu

An example limit line table is shown in Figure 13-25 using two upper limit segments and two lower limit segments. For each segment, a number of things need to be entered: Upper or lower: Use the pull-down to indicate if it is an upper limit or lower limit. Another option on the pull-down is “off” to enable suspension that segment.

X1 and X2

The constraints of the segment in the X-direction. Usually this variable will be frequency (segmented or linear frequency sweeps) but it could be time (time domain) or power (power sweep). If two segments cover the same frequency range (or portions thereof), the first segment will have precedence.

Y1 and Y2

The constraints of the segment in the Y-direction. These will have units of the graph type for the active trace (dB in the examples here).

X1 : 5.000000000 GHz GHz MHz kHz Hz

	Type	X1	X2	Y1	Y2	X Offset	Y Offset
2	Upper	5 GHz	10 GHz	9.8 dB	9.8 dB		
3	Lower	100 MHz	70 GHz	9.8 dB	9.8 dB		
4	Upper	5 GHz	10 GHz	9.8 dB	9.8 dB		
5	Lower	10 MHz	70 GHz	9.8 dB	9.8 dB		

Figure 13-25. Limit Line Table

The X offset and Y offset values allow one to shift both indices in a row by a constant amount. This can be useful in copying multiple rows and, for example, incrementing by a fixed frequency offset.

13-8 External Analog Input and Output

External Analog Input Ports

The VectorStar MS464xB-Series VNA has two external analog inputs (Figure 13-26) where the voltages may be plotted on the same graph types as S-parameters and un-ratioed parameters. Examples include the use of external power detectors or external current probes (to measure amplifier current draw as a function of frequency or power for example). Like everything else on this menu, it is a PER-TRACE selection.

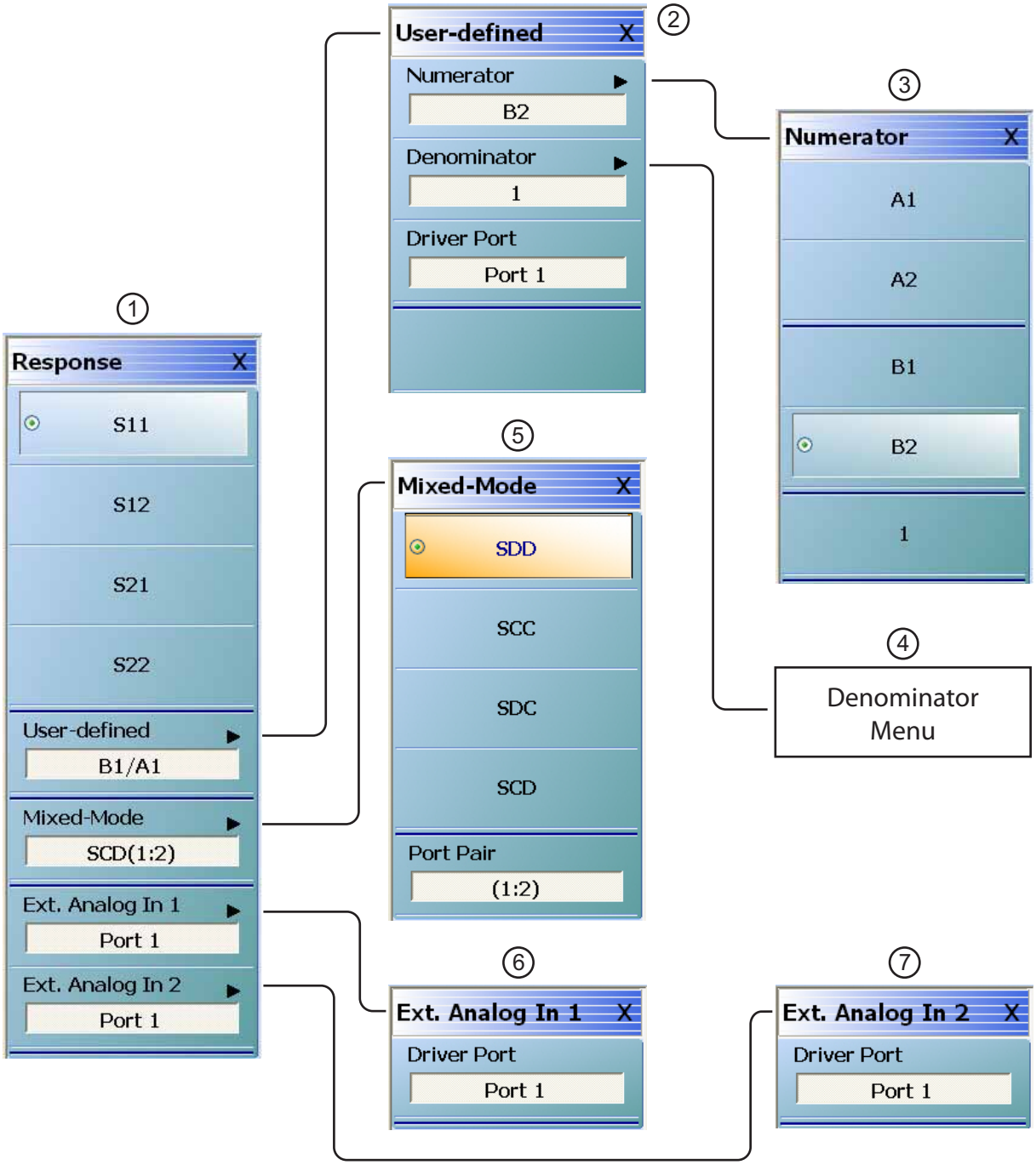


MS4647B VNA Rear Panel Example. 1. External Analog Input Ports (at upper left)	2. External Analog Output Port (at upper center)

Figure 13-26. Rear Panel External Analog I/O Ports

Response Menu

These are selectable on the response menu as shown in Figure 13-27 on page 13-34 below. The submenu allows a choice of which port is driving during that particular analog in measurement. This port selection may be important particularly with the use of external power detectors.



1. RESPONSE menu	5. MIXED-MODE menu
2. USER-DEFINED menu	6. EXT. (EXTERNAL) ANALOG IN 1 menu
3. NUMERATOR menu	7. EXT. ANALOG IN 2 menu
4. DENOMINATOR menu (not shown, the same as NUMERATOR menu)	

Figure 13-27. RESPONSE Menu and Submenus

Rear Panel External Analog Input Ports

These rear panel ports route through a multiplexer to an additional A/D converter. Internal calibration standards are used to correct for offset and channel gain slope, so the accuracies are on the order of 2 mV + 1% for absolute voltages under 5V and 2% for high voltages. The calibration is performed at power on and can be triggered from the diagnostics menus if necessary (see operation manual for details). This calibration is based on a pair of voltage references built into the instrument, so a linear error model of the low frequency analog in path can be generated.

The maximum range for the inputs is ±10V and the nominal input impedance is 60 kΩ. While the input is primarily intended for DC or very-low-frequency measurements, it has a bandwidth of approximately 1 kHz.

External Analog Out Port

There is also one external analog output port on the MS464xB-Series VNA and it is primarily used for sweep synchronization and certain triggering tasks of external instruments or DUTs. The control menu is located under MAIN MENU | SYSTEM | REAR PANEL OUT. and is shown in Figure 13-28.

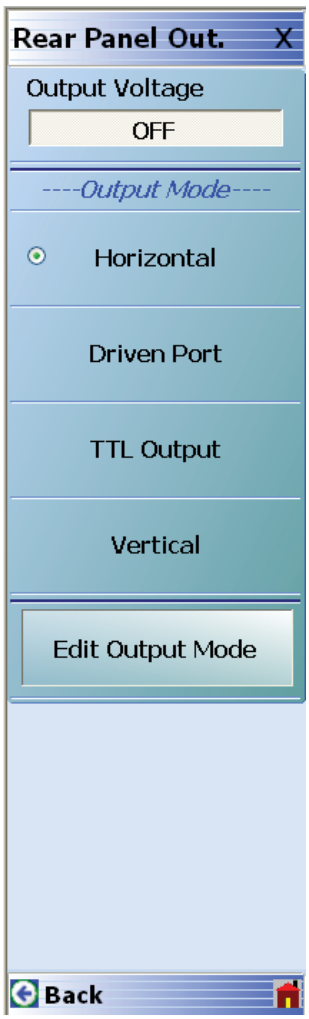
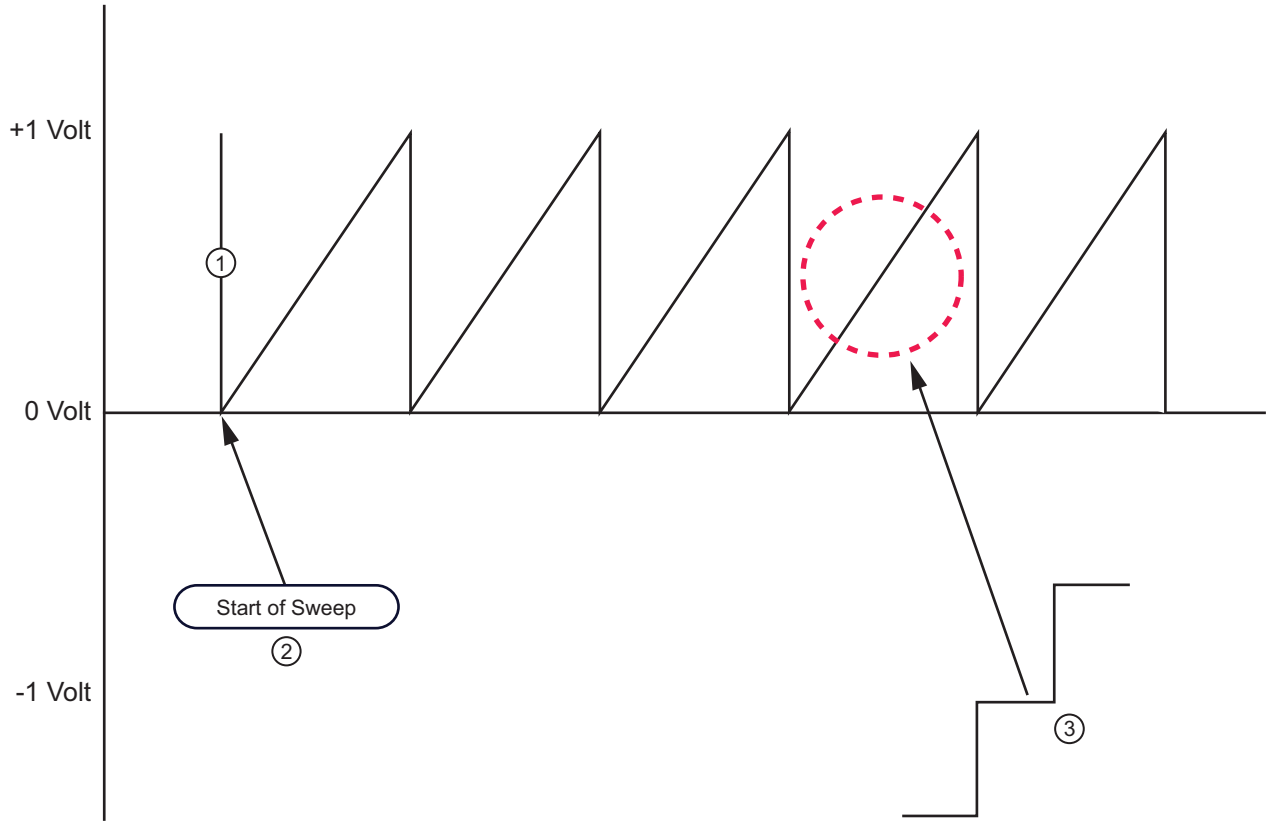


Figure 13-28. REAR PANEL OUT. (REAR PANEL OUTPUT) Menu

The top button is a toggle on/off while the next three buttons select the output mode. Horizontal is a sawtooth wave synchronized with the sweep (resetting to the start voltage at the first point of the sweep). This output really is a sequence of steps (see Figure 13-29) whose width is determined by the per point measurement time.

The horizontal output mode waveform for the settings of Figure 13-30 is shown below. The waveform is actually composed of small steps whose width represents the per-point sweep time.



1. Horizontal sweep waveform. 2. Start of horizontal sweep.	3. Close-up view of waveform showing small steps representing the per-point sweep time.

Figure 13-29. Horizontal Output Waveform

Driven port produces a static voltage level that is different when Port 1 is driving versus when Port 2 is driving. The TTL output is either a static logic value or a pulse (high-going or low-going) at the start of a sweep in a given direction. The details of the output are controlled in a dialog launched by the Edit Output Mode button (see Figure 13-30).

For the first two modes, the allowed voltage range is +/- 10V with an accuracy of about 20 mV + 2%. For TTL output, the selections are high, low, high pulse, and low pulse. The pulse will occur at the start of the sweep for the given driving port and the pulse width is adjustable. Although 0 pulse width may be entered, the actual minimum pulse width is about 0.5 µs. The maximum pulse width can exceed 10 s but note that if the pulse width becomes large relative to the time to complete the sweep, the instrument will wait for the pulse cycle to complete to avoid pulse collisions on subsequent sweeps.

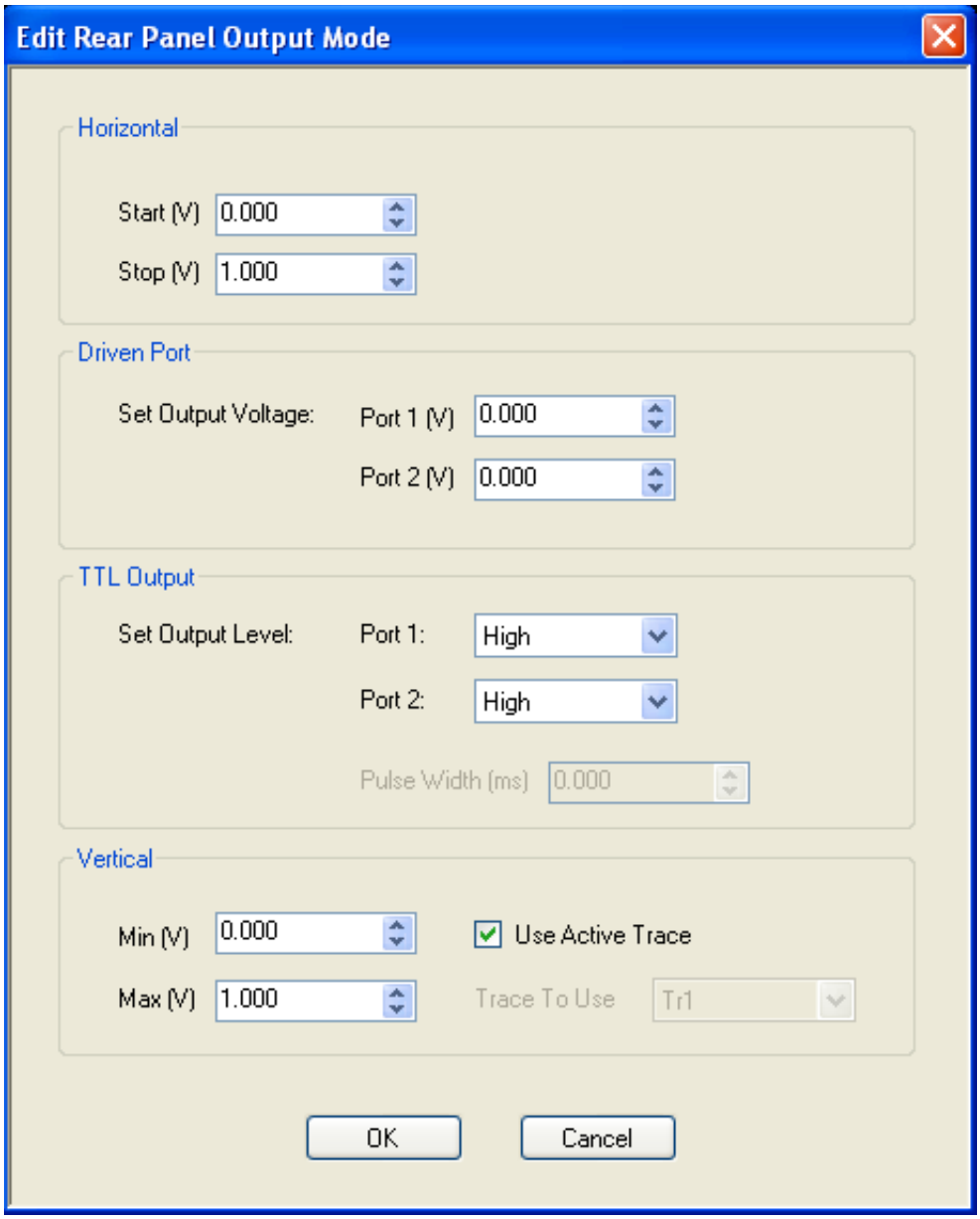


Figure 13-30. EDIT REAR PANEL OUTPUT MODE Dialog Box

The vertical selection will generate an analog output proportional to the plotted value on the active or selected trace data. This can be useful for generating feedback for a position controller or for analog data recording techniques. As shown in the setup dialog, a maximum (corresponding to the top line on a rectilinear graph type) and a minimum (corresponding to the bottom line on a rectilinear graph type) voltage should be entered within the usual +10V to –10V range for this output connector. Some points to observe:

- The voltage will only update when the channel in question is sweeping (the channel on which the rear panel output: vertical was selected). Any channel can have this control active and the rear panel output voltage will update according to the active/selected trace value when that channel is sweeping.

- For all single rectilinear graph types (log mag, linear mag, real, imaginary, phase, group delay, power out, power in, impedance…), the maximum and minimum voltages will correspond to the top line and bottom line of the graph, respectively. If the plotted parameter value goes outside of the graph (with its present scaling), the voltage output will clip at the maximum or minimum value.
- For dual rectilinear graph types (log mag + phase, real + imaginary…), the top graph will control the output.
- For Smith Charts, the maximum voltage corresponds to the outer rim of the chart and the minimum voltage corresponds to the center (and always in the sense of radius of the point being plotted… the parameter reflection/transmission magnitude in linear terms). For log and linear polar graph types, the maximum voltage corresponds to the outer rim of the graph (the reference value on the scale menu) and the minimum voltage corresponds to the outer rim value less five of the resolution values (also on the scale menu). Again, this output will be in the sense of magnitude of the parameter.

  Example: *Linear polar with reference value of 5U and a resolution of 0.2U/div.*

  The maximum voltage will correspond to a magnitude of 5U and the minimum voltage will correspond to a magnitude of 4U (5–5*0.2).
- The output will correspond to the plotted value, which means it is after all post-processing. If the display on the active/selected channel is static (e.g., memory-only display has been selected), then the output voltage will be static.
- The functionality will not be useful in modes where the trace updates all at one time (e.g., time domain—Option 2; pulse modes—Option 42; true mode stimulus—Option 43).

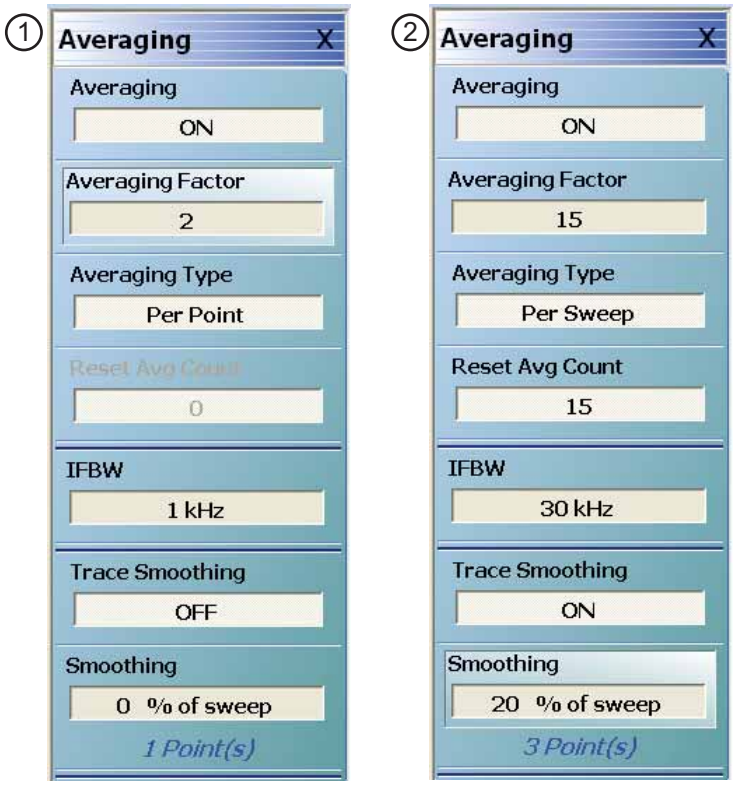
  Example: *Suppose one has used the vertical out feature on the 37XXX VNA in the past and wished to mimic that behavior in the MS464XX VNA for a phase graph type.*

  The 37XXX analog out vertical feature was limited to 1V/div and always worked on the active 'channel' (which is equivalent to the trace for the MS464XX). Thus the voltage range was always –4V to +4V for rectilinear graphs since the 37XXX rectilinear graphs all had 8 divisions. To perform the same setup on a MS464XX, either make the desired trace active or select it from the dialog, select the desired graph type (phase in this case), and enter a voltage range of –4V to +4V on the rear panel out setup dialog.

13-9 Averaging and Smoothing

Averaging and smoothing are covered to a considerable extent in the operation manual but there are some measurement-related impacts that should be discussed in this section. The control menu is shown in Figure 13-31.

- For additional relevant sections on applying Averaging functions, see the Calibration Examples in:
  - Chapter 3, "SOLT/SOLR Calibration"
  - Chapter 4, "Offset Short (SSLT) Calibration"
  - Chapter 5, "Triple Offset Short (SSST) and Multiple Offset Short (mSSS) Calibrations"



1. IFBW is set to 1 kHz and Trace Smoothing set to OFF (at left).	2. IFBW is set to 30 kHz, Trace Smoothing set to ON with Smoothing applied to 20% of the sweep (at right).

Figure 13-31. AVERAGING Menu—Both menus show that averaging has been toggled on.

Averaging

The Averaging button toggles the function OFF and ON.

Avg. Factor

The Avg. Factor (Averaging Factor) represents the number of measurements performed at each frequency point in the case of per-point averaging, and represents the number of sweeps averaged (in a running average sense) for per-sweep averaging.

Averaging Type

The Averaging Type button toggles between per-point and per-sweep averaging.

- **Per-Point Averaging**

Per-point averaging acquires additional samples at each frequency (or power) point and performs the averaging process at that time. In this sense, it is quite similar to an IFBW reduction (adding 10 per point averages is equivalent to a 10x reduction in IFBW). Since the time between sample acquisitions is small in this case, per point averaging works best at removing high-rate noise.

- **Per-Sweep Averaging**

  Per-sweep averaging averages a given frequency (or power) point's behavior on subsequent sweeps. This can be a very long time constant between samples (depending on sweep speed) so this type of averaging does best with low-rate noise.

  Per-sweep averaging is performed on a rolling basis. That is, if 10 per sweep averages are selected, the most recent sweeps are used to compute the result.

  Since per-sweep averaging has a long time constant, setup changes or DUT changes can appear to have an odd effect. Powering down an active DUT, for example, may lead to an S21 display to slowly drift away since it takes some time for the gain change to work its way through the sweep count. If a setup or DUT change is made, it may be desirable to reset the averaging count.

IFBW

The Intermediate Frequency Bandwidth (IFBW) is allowed in the range of 1 Hz to 1 MHz. At lower IFBWs, additional per point averaging will have little effect.

At very low frequencies, where the IFBW may be on the order of the system frequency, there could be measurement issues. By default, the IFBW will be limited at system frequencies below 3 MHz although this can be overridden under the System menu.

Trace Smoothing

Trace smoothing is toggled OFF and ON by this button. Trace smoothing performs a weighted averaging around each frequency point using a window size set by the percentage of smoothing. Since this process combines data at different frequency points, it should be used with care since it can remove valid frequency response information.

Averaging and Smoothing Conclusions

In time domain, averaging and IFBW apply to the basic frequency domain data. Smoothing applies to the time domain data. Everything on this menu is channel-based.

Strictly speaking, the above averaging comments apply to linear/log/CW frequency sweeps and power sweeps. When in segmented sweep mode, the behavior is slightly different. One can assign IFBW and per-point averaging on a segment-by-segment basis (so that, for example, one could use a wide IFBW and little averaging in a low insertion loss part of the DUT's response and a narrow IFBW and more averaging in a very high insertion loss part of the response where one wanted to maximize dynamic range). Sweep-by-sweep averaging, by definition, applies to the sweep rather than a segment so it can be applied only to the aggregate. Unlike in linear frequency sweep, one can then simultaneously apply both point-point averaging (from the segmented sweep table now) and sweep-by-sweep averaging (still from the averaging menu). If sweep-by-sweep averaging is applied in segmented sweep and one goes back to linear sweep, the averaging setting will carry but any use of point-by-point averaging in segmented sweep will not carry back to linear sweep.

Smoothing is also applied on a sweep basis and cannot be dictated by segment when in segmented sweep.

Chapter 14 — Measurement—Time Domain (Option 2)

14-1 Chapter Overview

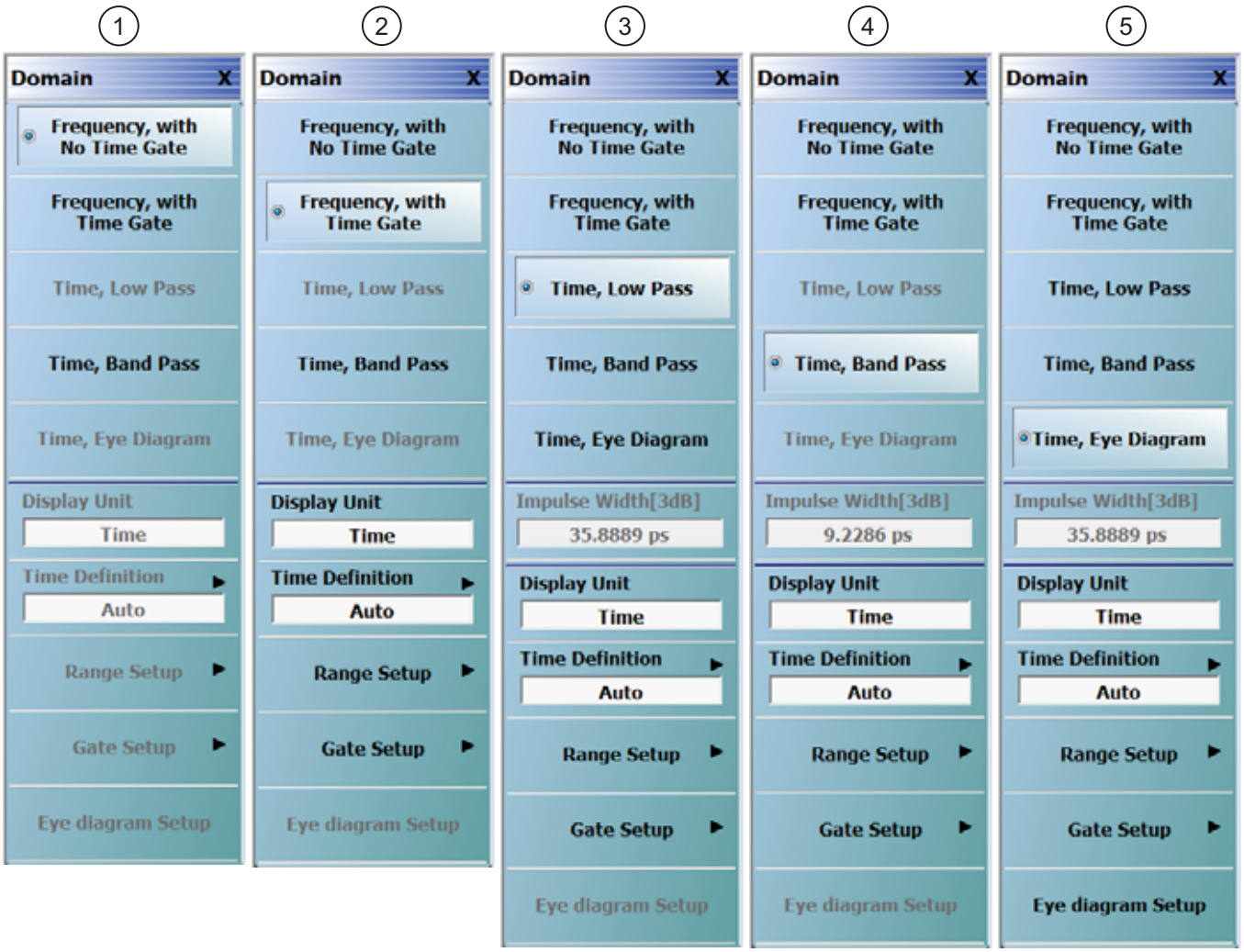
This chapter provides time domain measurement guidelines and procedures. General descriptions, key concepts, and example procedures are presented for time domain measurement modes of low pass, bandpass and gating. This chapter also covers eye diagram calculations (requires Option 47 in addition to Option 2) which present a time domain representation of the result of a digital data stream passing through the DUT being measured.

14-2 Introduction

The *time domain* option offers the ability to transform the native frequency domain data of the MS464xB Series VNA into time domain information for TDR-like displays, distance-to-fault analysis, and general spatial-based circuit and network troubleshooting. Uses for time domain include:

- Identifying the location of significant mismatches/discontinuities in a launch structure in a fixture or PC board
- Finding and quantifying defects in a cable assembly
- Identifying the characteristics of a discontinuity (inductive or capacitive) in a transition within a fixture or on-wafer
- Determining semi-quantitatively the impedance levels in a cascaded series of transmission lines

There are a significant number of choices in how to configure the transformations that this section will cover. To begin, time domain is a per-trace invocation so that, within a given channel, frequency domain and time domain traces can be freely mixed on any response parameter. Note that since there is a single x-axis readout per channel, and it will be in the units of the active trace (either frequency or time range, not both at the same time will be displayed). Time domain and the TIME DOMAIN menu is accessed from the DISPLAY menu (and its top level selections are shown in Figure 14-1 on page 14-2.



The presence and availability of some DOMAIN menu selections depend on the time domain mode selected on the DOMAIN menu, as well as on the SWEEP TYPES and FREQUENCY menus.

Figure 14-1. Domain Menu Variations Based on Time Domain Mode Selection (1 of 2)

1. Frequency, with No Time Gate selected	4. Time, Band Pass selected
2. Frequency, with Time Gate selected (FTG)	5. Time, Eye Diagram (TED) selected
3. Time, Low Pass selected	

Figure 14-1. Domain Menu Variations Based on Time Domain Mode Selection (2 of 2)

14-3 Basic Time Domain Modes

The five basic modes of the Time Domain menu are:

- **Frequency With No Time Gate**

  This is the regular frequency sweep mode.

- **Frequency With Time Gate**

  This is frequency domain data that has passed through time domain where a gate is applied to exclude certain data (such as to remove certain defects) before returning to the frequency domain.

- **Time, Low Pass**

  A time domain mode where frequency content fairly close to DC is available (start frequency no more than about 10 step sizes). Step response (like a TDR) processing is available and resolution is better but this mode may not be available for all frequency lists. The selection will be greyed out if incompatible.

- **Time, Bandpass**

  A time domain mode for any frequency list. Only impulse response can be displayed, defect identification tools are more limited, and resolution is a factor of 2 worse than in lowpass (for the same sweep width) but it can be used for any frequency sweep. This is the only choice in band-limited scenarios such as waveguide.

- **Time, Eye Diagram**

  Eye diagram is a slightly different time domain mode in that it is not a strict transform process but presents the calculated output (expressed as an overlaid time waveform) of the DUT being measured when it is stimulated with a user-specified, synthetic digital data stream. This mode may not be available for all frequency lists. The selection will be greyed out if incompatible.

14-4 General Concepts

Chirp-Z Transform

The time domain functionality is provided by a chirp-Z transform (in most cases) of the available frequency domain data for that parameter. Since the transform simply treats the frequency domain values as input data, any parameter can be transformed (including mixed-mode and differential S-parameters). Unratioed parameters are less useful since they do not contain phase information that the transform relies upon.

The chirp-Z transform is, in a macro sense, very similar to the Fast Fourier Transform with the exception that the output range can be variable. This allows the ability to specify an arbitrary time range to look at while maintaining the desired point count. A different algorithm is used with dispersive media, such as waveguide and microstrip where the time-frequency relationship is more complex, but the functionality remains the same.

Defects as Impulse Functions in the Time Domain

Defects can be considered to be impulse functions in the time domain. This sum of impulses transforms to a sum of complex exponentials in the frequency domain. While these produce the characteristic ripples seen in frequency domain data of mismatched systems, the frequency domain data can be hard to interpret as to the location of the defects causing the ripples. This is the value of a time domain analysis.

One-Way or Round Trip Time

One question that arises is whether the time (or distance) plotted represents a one-way or round trip time, particularly in the case of reflection measurements. The time definition menu, shown in Figure 14-2, controls this behavior. When in Auto, the system will always display one-way times and detects whether the measurement parameter is reflection or transmission to help sort that out. If using user-defined parameters or unusual loop-access configurations, going to manual control may be needed. The transform itself will generate a round-trip time for reflection and a one-way time for transmission without any intervention from the system.



Figure 14-2. TIME DEFINITION Menu

14-5 Low Pass Mode

Low pass mode assumes the existence of data near DC which enables the ability to compute step responses and to create a pure real transform. While any graph type can be used (except Imaginary which would have a flat line), Real is sometimes the most valuable since information about the defect can be determined. An example plot showing a short at the end of a small transmission line length of approximately 100 ps appears in Figure 14-3. Both the impulse response and step response are plotted on real graph types. Many aspects of this plot will be discussed in this section including the impulse and step presentation of the same data.

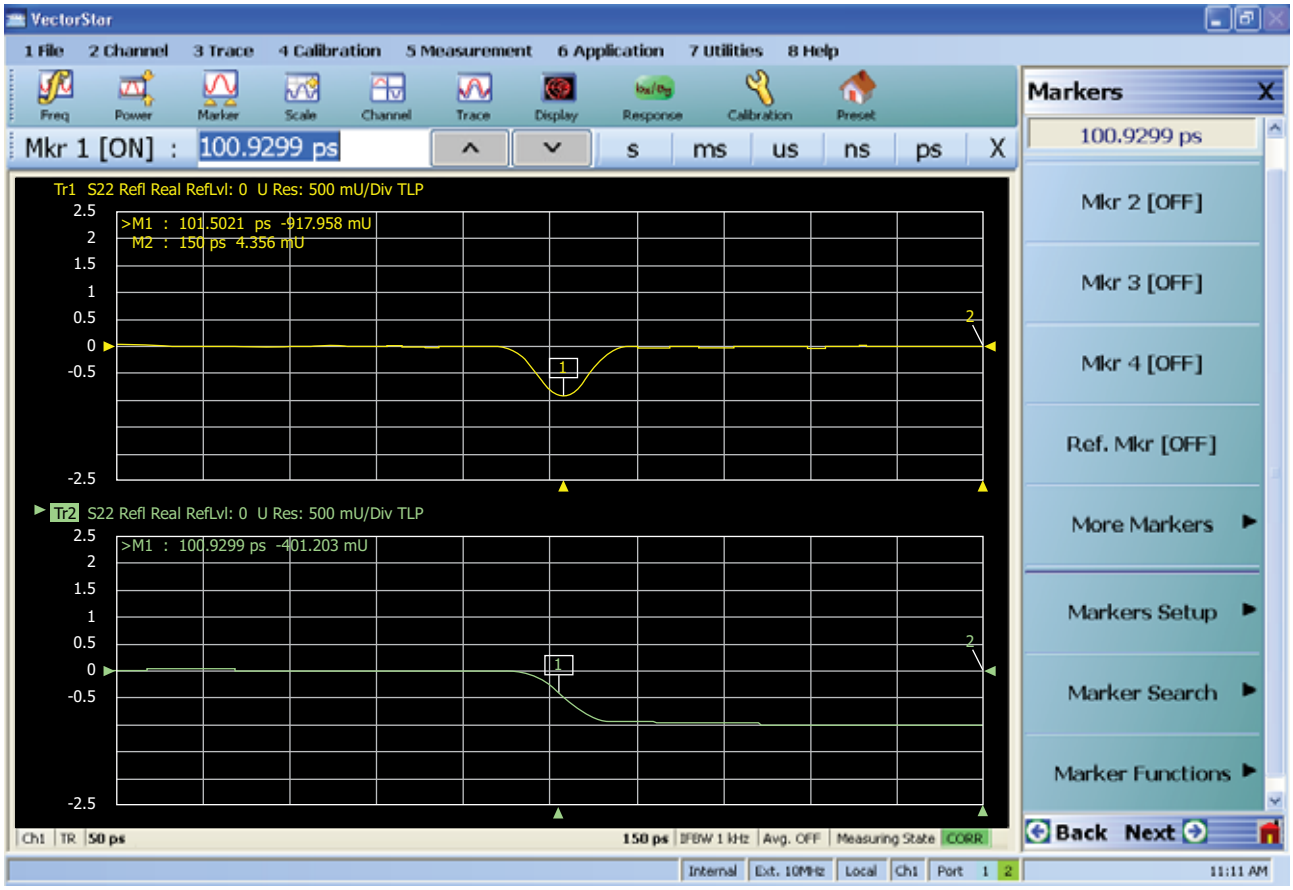


Figure 14-3. Example Low-pass Time Domain Plot

Many of the other submenus change slightly depending on which mode is selected so the remaining subsections will be partitioned according to the mode.

Range Setup Menu Functions

The Range Setup menu for low-pass time domain is shown in Figure 14-4. The top button, Display Unit, toggles between Time and Distance and is a duplicate button to the Display Unit button on the TIME DOMAIN menu (Figure 14-1 on page 14-2).

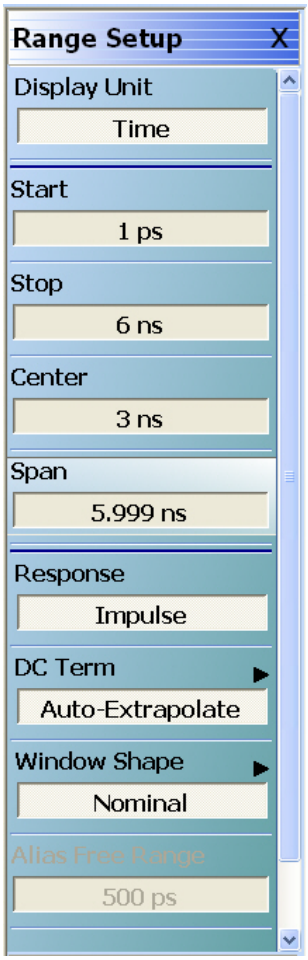


Figure 14-4. RANGE SETUP Menu—Time, Low Pass Domain

The fundamental output of the transform depends on the non-dispersive or dispersive nature of the media. In the case of non-dispersive media (to include coax), time is the fundamental output of the transform and distance is calculated using the media information on the measurement menu. In the case of dispersive media (waveguide or microstrip), distance is the fundamental output and time is calculated from that.

The Start, Stop, and Center buttons all invoke field tool bars that allow user-input for each value (in distance or time); with the Span button displaying the calculated result. There are few limits to what may be entered but extreme entries may not always be useful due to constraints of resolution and alias-free range. These limits are determined by the frequency list used as well as the window selected.

Resolution is interpreted as impulse width (the width of a singular defect) while alias-free range is the maximum time range that can be studied before defects start repeating themselves (due to the cyclical nature of the transform). To help, the resolution (impulse width) is displayed as a read-only variable on the main time domain menu (Figure 14-1 on page 14-2) and the alias-free range is displayed on this range menu as a read-only variable.

The response choice is either Impulse or Step. The step response, which allows a TDR-like display is simply an integration of the impulse response which is the natural output of the transform. Normally, this integration begins at the start time used in the current range (and continues to the stop time). This is advantageous when using results for certain de-embedding activities. It can, however, be confusing if one zooms into a region of impedance different from the reference impedance since the integration would not capture the transition. For this reason, there is the selection available to start the integration at 0 (regardless the range setting). The integrated result will be mapped back into the current start-stop range using interpolation with the linear value set to zero at negative times. If the stop time is also negative, the integrated result will all be near zero.

Note that the integration for step response also requires an initial integration value and this comes from the network's DC response. Since the VectorStar MS464xB VNA cannot get all the way to DC, some additional information is needed to perform this integration. To see this consider:

$$\begin{aligned}
\text{ImpulseResponse} &= \Im^{-1} X(DC) + \{X(\text{sweepRange})\} \\
\text{ImpulseResponse} &\approx A \cdot X(DC) + \Im^{-1}\{X(\text{sweepRange})\} \\
\text{StepResponse} &= \int_0^t [A \cdot X(DC) + \Im^{-1}\{X(\text{sweepRange})\}]dt
\end{aligned}$$

Equation 14-1.

DC Term Menu

Since the DC value ends up being integrated from time 0 (zero), the value used here is quite important and the choices to compute this value are shown in Figure 14-5. The default choice is to allow the system to auto-extrapolate from existing frequency data to estimate the DC value.

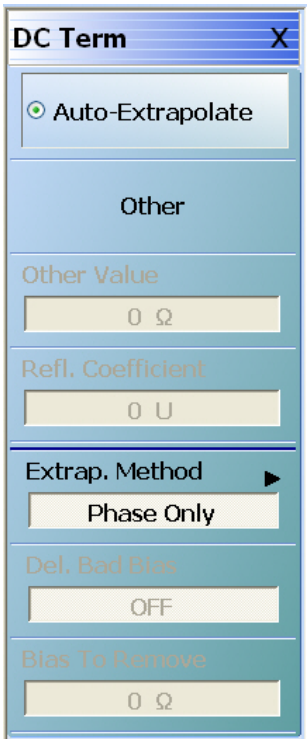


Figure 14-5. DC TERM Selection Menu

There are options on how the extrapolation is done, as shown in Figure 14-6.



Figure 14-6. The DC Term EXTRAPOLATION Menu

The default method, Mag-Phase, extrapolates both portions as would be expected and is energy-conserving. For cases where the start frequency is low and the DUT loss changes slowly over frequency, sometimes the magnitude may be assumed constant and only the phase function need be extrapolated (most common with long cable assemblies). The other option allows a table of low frequency values to be entered (two-column, tab-delimited). If the DUT is well-known, extrapolation can be avoided altogether by entering the DC impedance.

Window Shape Menu

The last item on the Range Setup menu is the Window Shape selection button which displays the Window Shape submenu shown in Figure 14-7.



Figure 14-7. WINDOW SHAPE Menu

Since the frequency range of the VNA is finite, the frequency domain data will have a discontinuity at the stop frequency. This introduces side lobes in the time domain data that can obscure smaller defects and hamper separation of defects. The window provides some pre-processing of the frequency domain data to reduce the severity of the discontinuity and hence the side lobe level. This also reduces resolution but is unavoidable.

The Nominal window is the default and provides about half of the resolution of Rectangular (no window) but with an approximate 30 dB reduction in side lobe levels. The Nominal window is advised for most applications.

Since the window so strongly affects resolution, the Impulse Width display is repeated on this submenu to help determine the impact on the desired measurement. If in low-pass time domain and a step response has been selected, this readout will be in terms of rise time. The rise time definition can be selected on the Range menu to either be 10-90% or 20-80%.

An example of how the window shapes affect the impulse data (main lobe width and side lobe level being traded-off) is shown in Figure 14-8. Here the same data appears with the four different window selections. For this plot, data was saved to TXT files and plotted externally.

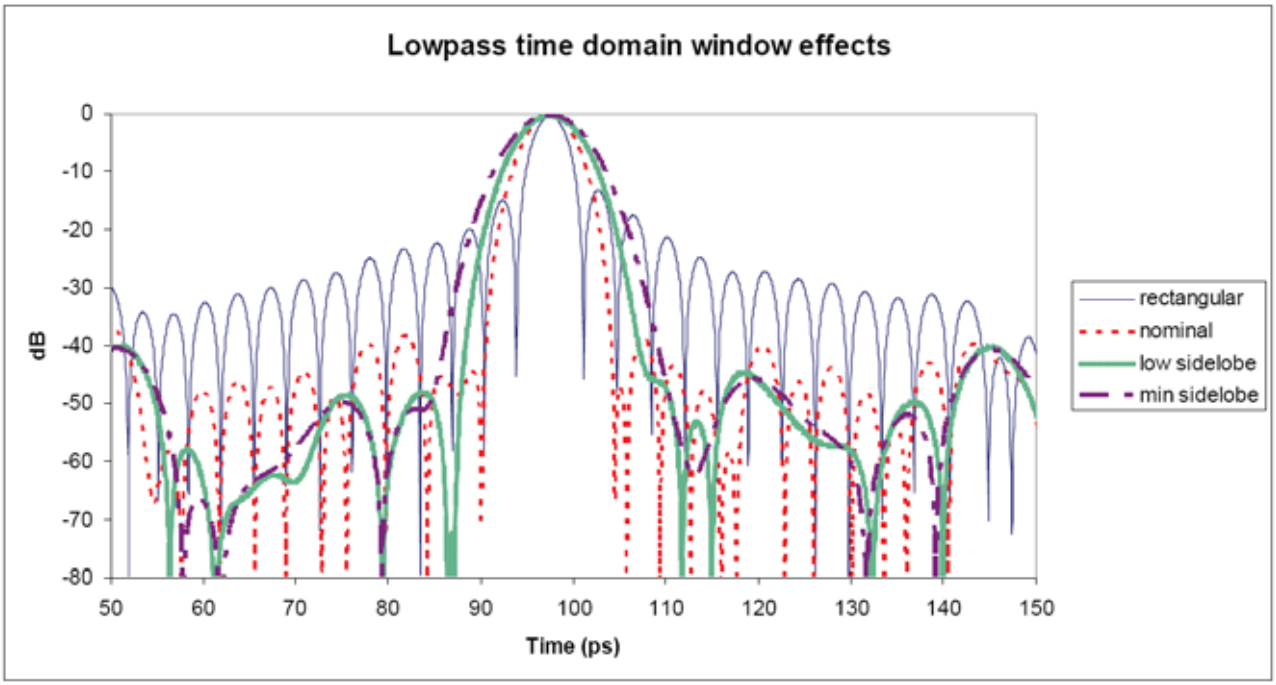


Figure 14-8. Effects of Window Shapes Plot

The Advanced Window selection button brings up the dialog shown in Figure 14-9 that has the previous four choices along with two new parameterized windows, Kaiser-Bessel and Dolph-Chebyshev.

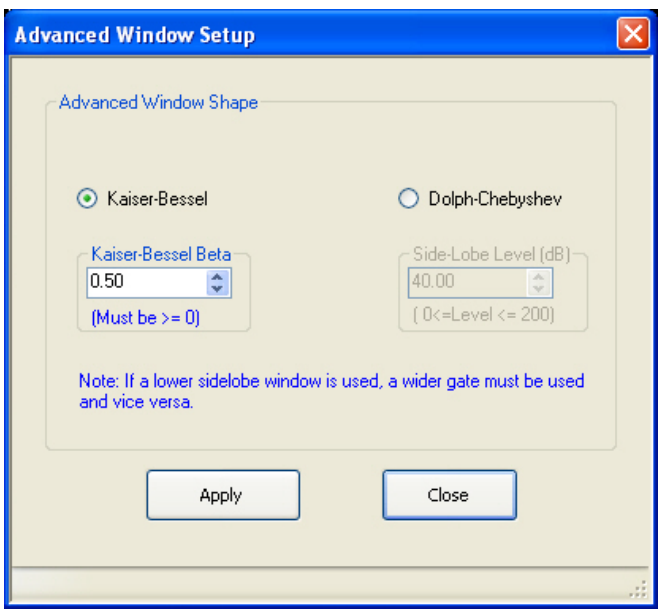


Figure 14-9. Advanced Window Setup Dialog

The dialog for advanced window setup makes two new window choices available (Kaiser-Bessel and Dolph-Chebyshev). The Apply button must be used for a radio-button selection to take effect.

These two new window types allow for a finer selection of the trade-off between side lobe level and resolution. For the Kaiser-Bessel window, a larger Beta value leads to lower side lobes, but a wider main lobe width (and hence poorer resolution). For the Dolph-Chebyshev window, the side lobe level is parameterized explicitly (in absolute dB) and a larger value leads to a wider main lobe width as well. The windows for two parameter values for each of these windows are shown in Figure 14-10 along with the rectangular window for comparison.

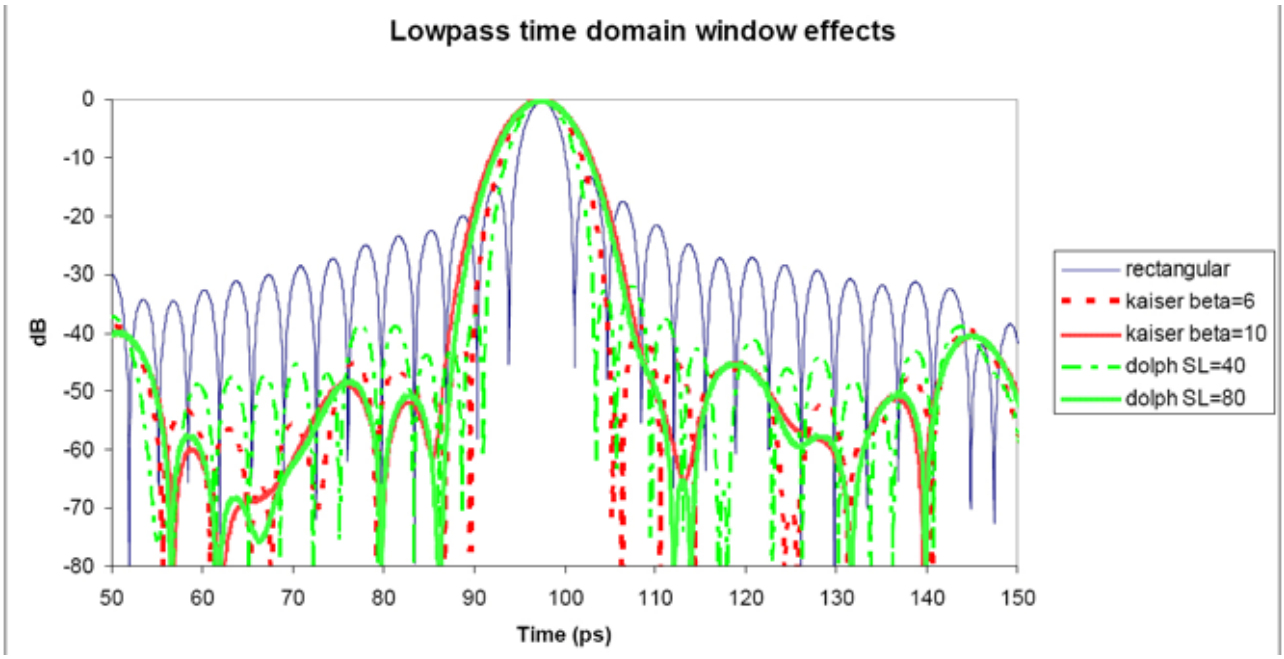


Figure 14-10. Effects of Window Shapes Plot with Advanced Windows Selection

The approximate relationship between these parameters and the main lobe width (null-to-null) is suggested in Figure 14-11. Here, everything is scaled relative to a rectangular window (a nominal window is at 2, a low side-lobe window is at 3, and a minimum side-lobe window is at 4 on this scale) and the y-axis is normalized relative to the lobe width of a rectangular window.

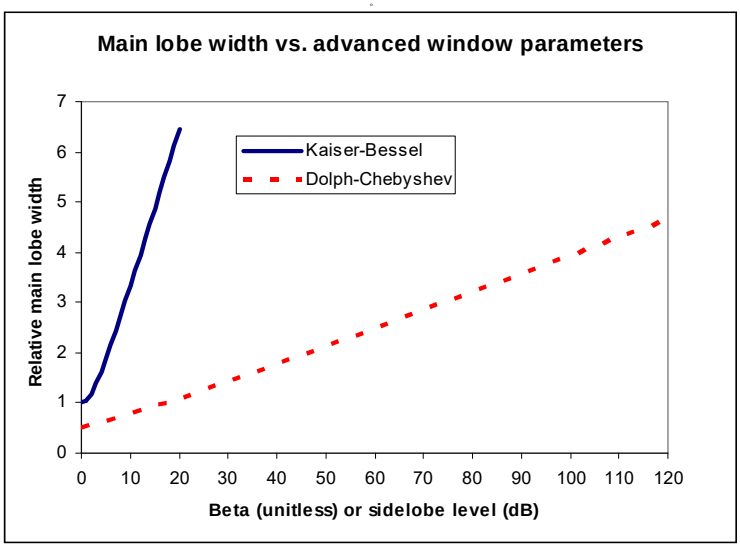


Figure 14-11. Comparison of Lobe Width vs. Window Parameters

Table 14-1. Below are the window widths (expressed in terms of the distance between the first nulls around the central lobe). ‘BW’ refers to the total sweep bandwidth.

Window Type	LP Main Lobe Width (null-null)	BP Main Lobe Width (null-null)
Rectangular	0.5/BW	1/BW
Nominal	1/BW	2/BW
Low sidelobe	1.5/BW	3/BW
Minimum sidelobe	2/BW	4/BW
Kaiser (parameter ß)	$\left(\frac{0.5}{BW}\right)\frac{\sqrt{\pi^2+\beta^2}}{\pi}$	$\left(\frac{1}{BW}\right)\frac{\sqrt{\pi^2+\beta^2}}{\pi}$
Chebyshev (parameter α dB)	$\left(\frac{0.5}{\pi\cdot BW}\right)\sqrt{\left(\frac{\pi}{2}\right)^2+\left(\cosh^{-1}\left(10^{\alpha/20}\right)\right)^2}$	$\left(\frac{1}{\pi\cdot BW}\right)\sqrt{\left(\frac{\pi}{2}\right)^2+\left(\cosh^{-1}\left(10^{\alpha/20}\right)\right)^2}$

Table 14-2. Below are the window widths (expressed in terms of the width between the 3 dB points relative to the central lobe). ‘BW’ refers to the total sweep bandwidth.

Window Type	LP Main lobe width (3 dB)	BP Main lobe width (3 dB)
Rectangular	0.221/BW	0.442/BW
Nominal	0.323/BW	0.646/BW
Low sidelobe	0.412/BW	0.824/BW
Minimum sidelobe	0.472/BW	0.944/BW
Kaiser (parameter ß)	$\left(\frac{0.221+0.019\cdot\beta}{BW}\right)$	$\left(\frac{0.442+0.038\cdot\beta}{BW}\right)$
Chebyshev (parameter α dB)	$\left(\frac{1}{7.68\cdot BW}\right)\sqrt{\cosh^{-1}\left(10^{\alpha/20}\right)}$	$\left(\frac{1}{3.84\cdot BW}\right)\sqrt{\cosh^{-1}\left(10^{\alpha/20}\right)}$

14-6 Bandpass Mode

The Bandpass Time Domain mode is similar to low pass, but a few menu items change. Any graph type can be used with bandpass mode, but log magnitude and linear magnitude are the most common. The top level of the time domain menu is repeated in Figure 14-12 for convenience. This menu level does not change between the time domain modes. An example measurement (of a short on a transmission line like in Figure 14-2) is shown in Figure 14-17. Here, a real and imaginary plot is shown to illustrate the difference from the pure real low pass time domain result, but this graph type is not commonly used in practice.

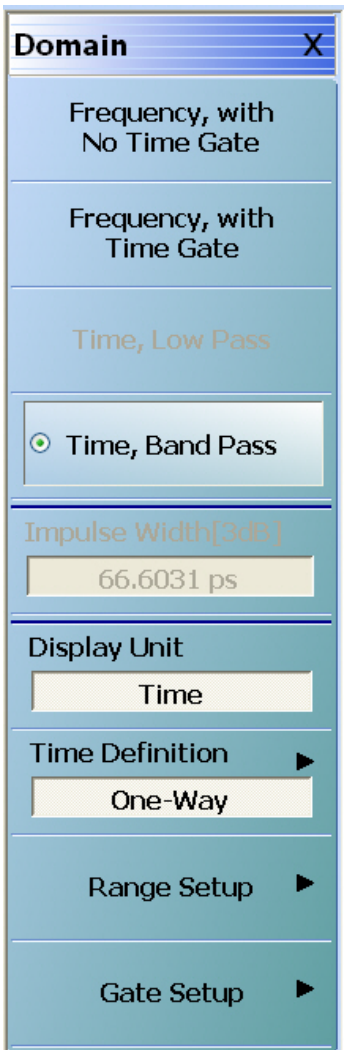


Figure 14-12. Top Level DOMAIN Menu—Time, Band Pass Selected

An example bandpass time domain plot is show below for a short at the end of a transmission line. In a log magnitude display, there is a single impulse of approximately unity amplitude near the 100 ps mark.

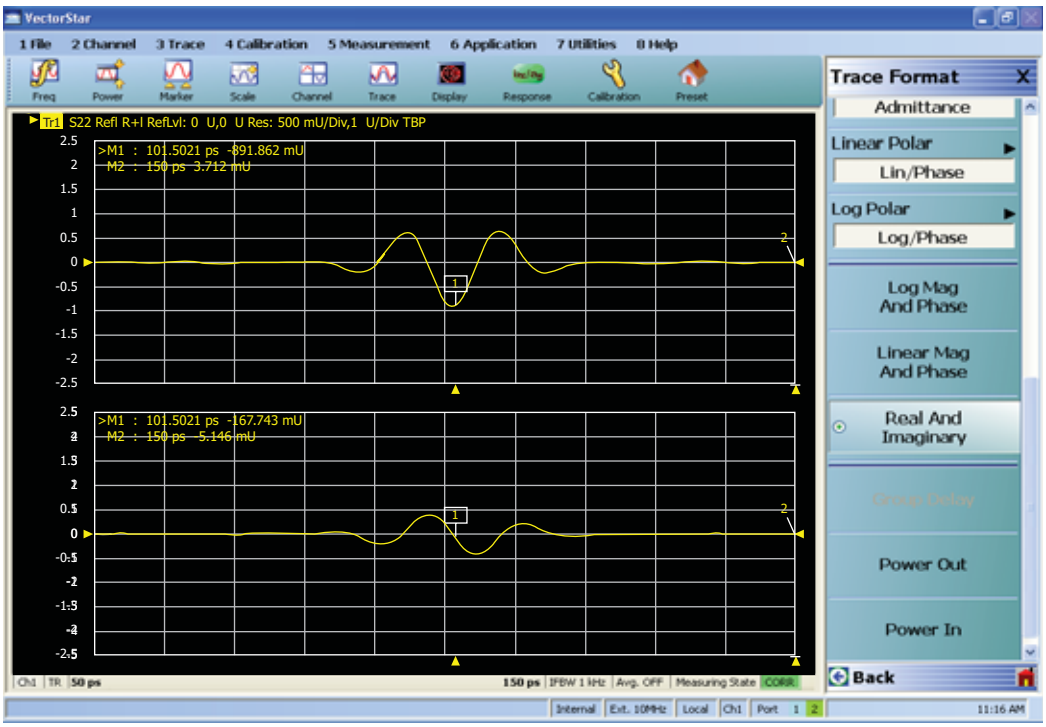


Figure 14-13. Example Bandpass Time Domain Plot

The range menu for bandpass mode is shown in Figure 14-14. The differences here are that the response choice and DC terms are gone since they do not apply to this mode, and a new item appears: Phasor Impulse.

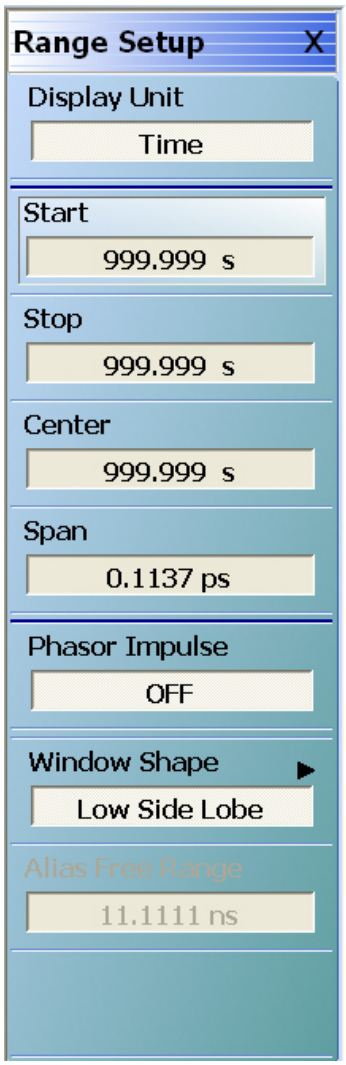


Figure 14-14. RANGE SETUP Menu for Bandpass Time Domain

In low pass mode, the sign of the data can be used to provide some hints as to the nature of the defect (inductive or capacitive). It is less obvious in bandpass mode since the time domain data is complex. A function termed Phasor Impulse Mode is an attempt to simulate the data reduction of low pass mode when operating in bandpass mode. It is only correct if the defect can be described by a single defect (a single complex exponential in the frequency domain). The range should be adjusted to have one peak on screen occupying a reasonable fraction of the span. The Phasor Impulse function processes this single peak to produce a pure real transform carrying sign information much like lowpass mode (positive for inductive, negative for capacitive).

The window shapes have the same effect as in low pass but the starting resolution is only half that of low pass (the window effects are multiplicative). The window effects are illustrated in Figure 14-15 and correspond to the measurement of Figure 14-13 on page 14-14, but expressed in log magnitude. Note the trade-off of side lobe height for main lobe width and that the lobe width is twice that for low pass (Figure 14-8 on page 14-10).

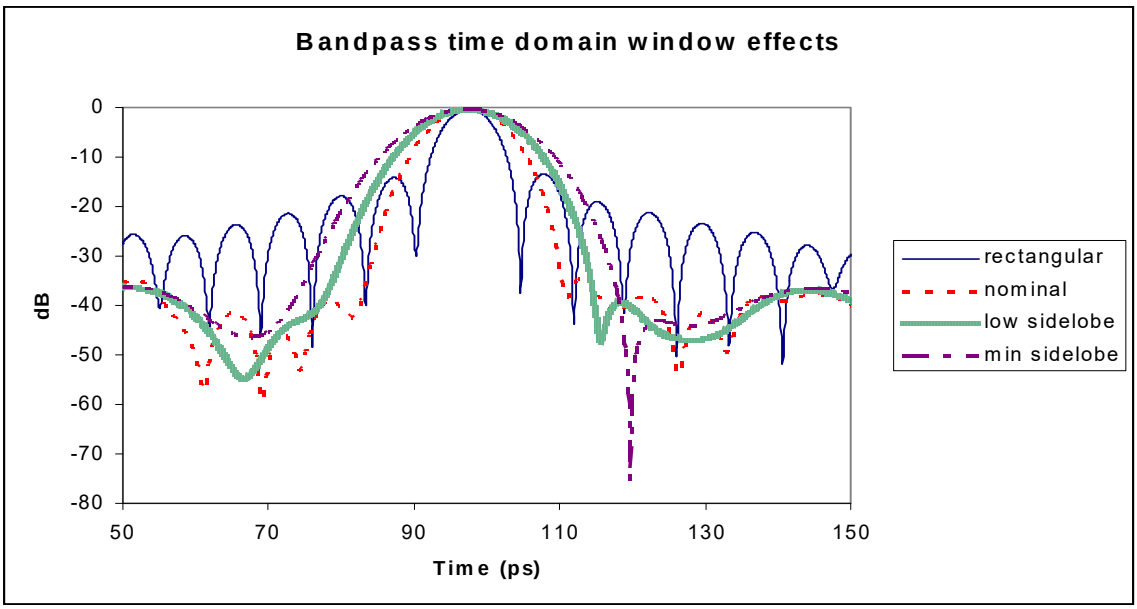


Figure 14-15. Window Effects in Bandpass Time Domain

As with lowpass time domain, the Advanced Windows are also available. Some example results are shown below compared to the rectangular window for a few parameter values.

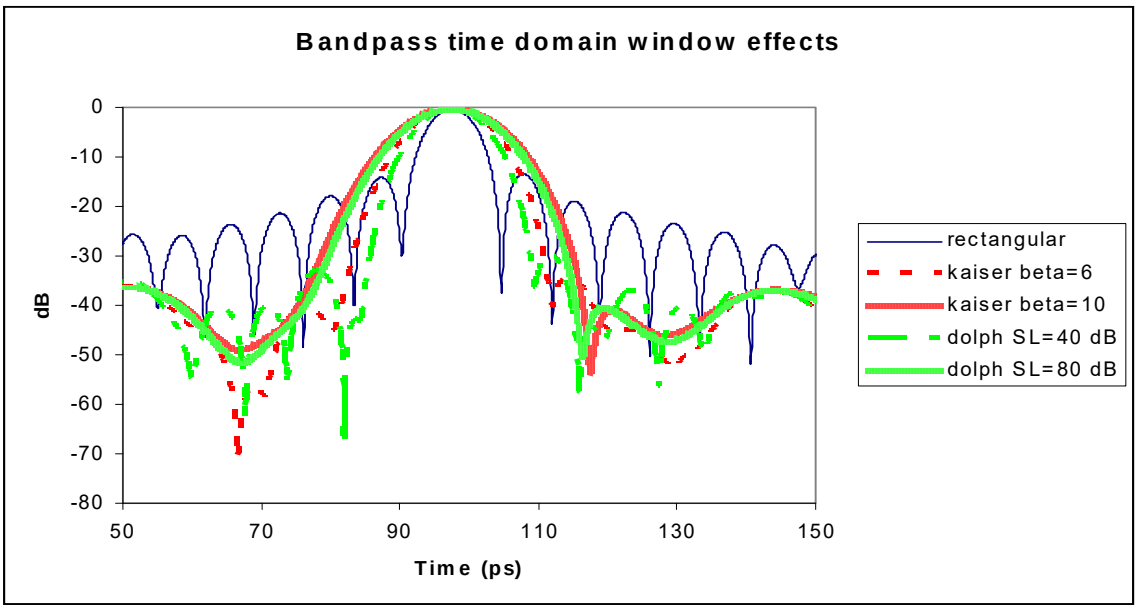


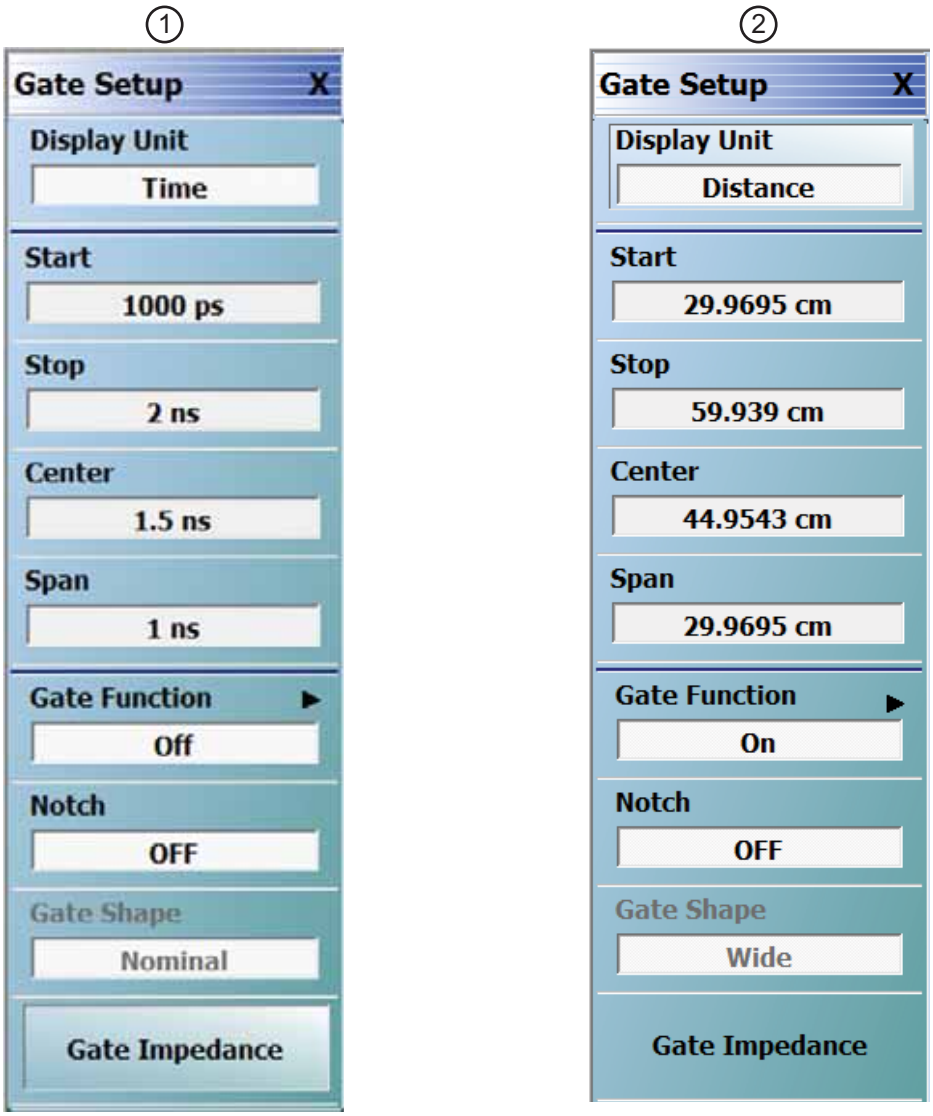
Figure 14-16. Window Effects in Bandpass Time Domain (advanced window types)

14-7 Gating

Both lowpass and bandpass work similarly with regards to gating. Gating is the process of selecting or deleting certain defects to study. This can be left in time domain but, more commonly, the gated results are fed back through the forward transform to get the frequency domain result corresponding to the modified defect scenario just created.

Gate Setup Menu

The Gate menu looks much like the Range Menu. The Display Unit toggle button and Start, Stop, Center, and Span buttons (for the gate this time) control values as described in the sections above.



1. Time units (at left)	3. "Gate Function Menu" on page 14-18
2. Distance units (at right)	4. Gate Impedance Menu—See "Time Gating and Reference Impedances" on page 14-24

Figure 14-17. GATE SETUP Menu

The Notch toggle selects the polarity of the gate. When notch is OFF, the gate will keep everything between start and stop. When notch is ON, the gate will reject everything between start and stop.

Gate Function Menu

The main submenu, Gate Function, is shown in Figure 14-18.



Figure 14-18. Gate Function Submenu

The default gate shape is nominal. By default, the gate is off. Selecting Display will allow the gate function to be drawn on screen (using the current graph type for the active trace). This can be helpful in visualizing what is being included in the gate. Turning gate on will apply the gate to the current time domain data.

The gate shape is analogous to the window selection. If the data was truncated with a sharp gate (minimum, akin to rectangular), maximum resolution is used determining the gate but ripple is introduced in the frequency domain. For more gradual gates, the resolution in separating defects decreases, but the size of the artifacts added to the frequency domain data decreases as well.

The window and gate shapes cannot be selected entirely independently since they interact through the transform. In particular, the use of a very sharp gate with a low side lobe window can lead to large errors. The allowed combinations are shown in Table 14-3 below. If an invalid combination is selected, the variable not being currently modified will be changed to the nearest valid value.

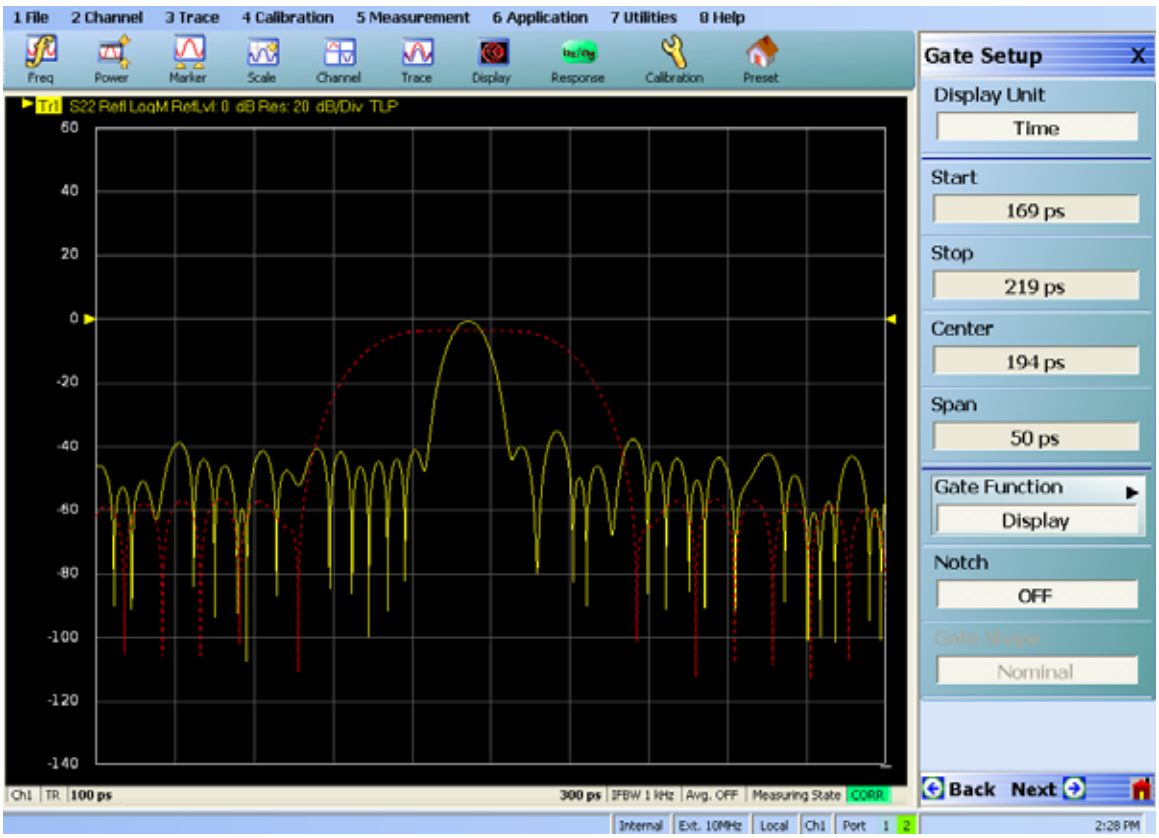
Table 14-3. Window Type and Gate Shape—Allowed Combinations

Window/Gate	Minimum	Nominal	Wide	Maximum
Rectangular	OK	OK	OK	OK
Nominal	OK	OK	OK	OK
Low side lobe	No	OK	OK	OK
Minimum side lobe	No	No	OK	OK

With the advanced gates and windows, selections are not precluded although substantial errors can result if values are chosen without caution. If a more aggressive window is chosen (larger beta or side-lobe level), then the gate must be wider (wide or maximum; larger beta or side-lobe level).

DUT Example—Gate and Window Nominal

To work through an example, a DUT consisting of a short at the end of a slightly mismatched transmission line is used. It is desired to examine the short more closely in frequency domain, excluding the effects of the transmission line. In Figure 14-19, the gate is in display mode surrounding the desired reflection. Both gate and window are set to nominal in this case.



Gate is the red dashed line

Figure 14-19. Gate in Display Mode Example

Next the gate is turned to on. In Figure 14-20, the suppression of the time domain information outside of the gate area is seen.

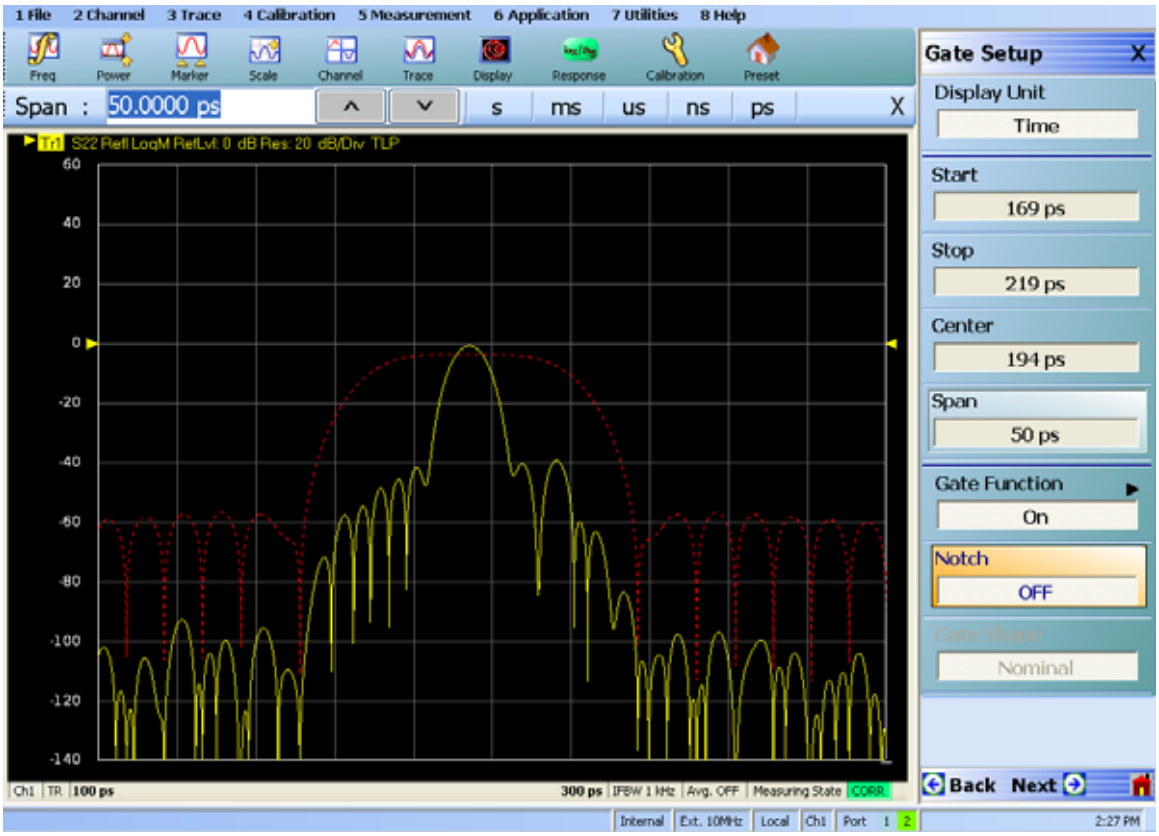


Figure 14-20. Gate Turned On Example

Finally, frequency with time gating is activated and the result is shown in Figure 14-21. The result from frequency without time gating is shown in memory as a darker trace. The time gating has removed much of the ripple due to the mismatched transmission line and residual source match of the instrument.

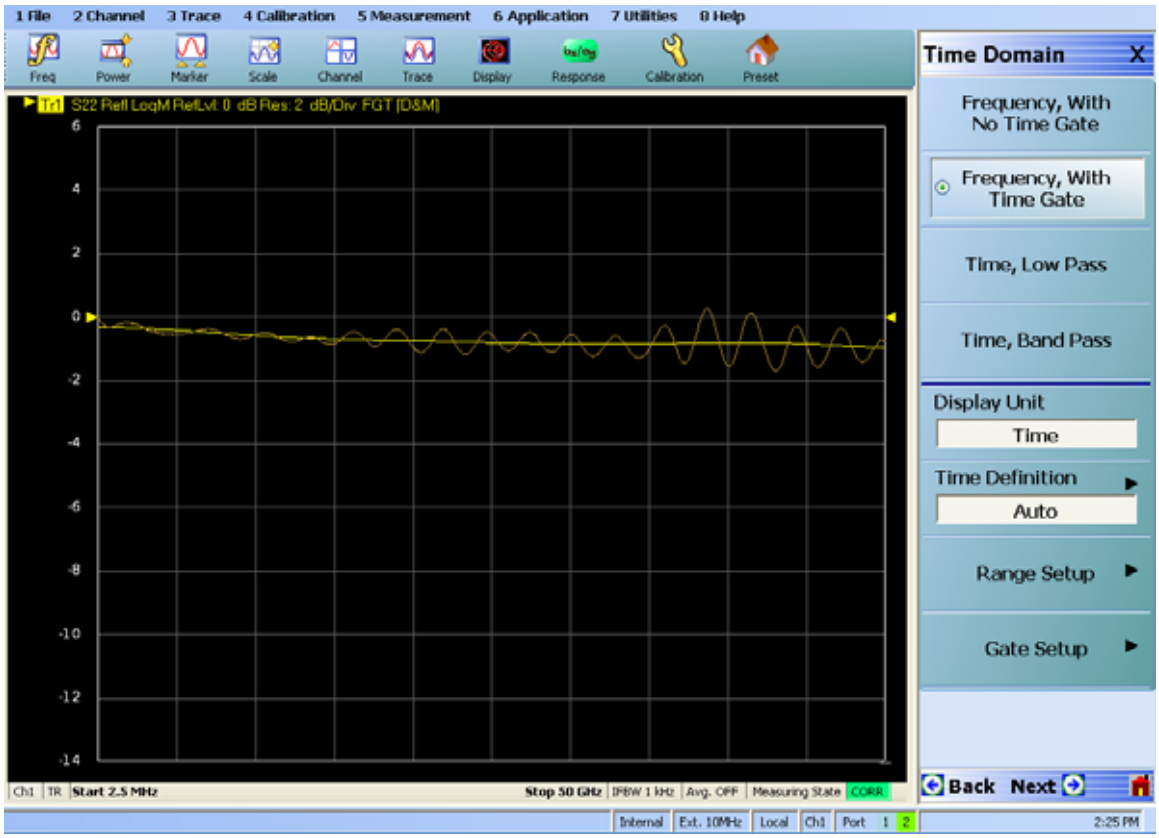


Figure 14-21. Frequency with Time Gating Example

Other Frequency-with-time-gate Calculation Items

Questions are sometimes asked about the details of the gating process and the subject of uncertainty in the final result. The latter topic is addressed in the next section. In terms of the process itself, the basic concept is simple enough: a particular functional form (to exclude or include certain portions) is applied to the time domain data before it is returned to the frequency domain. As the time domain data is theoretically of infinite extent, the limited data roster forces some truncation to happen by default so even with an infinitely wide gate, the process is not conservative.

To get around this problem, a calibration signal (a single, synthetic tone) is applied to the current window/gate setup to generate a set of correction factors. Normally this does not introduce any significant errors. If the gate is very narrow (in the sense of approaching 1/BW), there is an additional issue in that the equivalent frequency domain convolution starts trying to interact more with frequencies outside of the sweep range. In the extreme case, this results in distorted final result, particularly at the extreme frequencies. To improve these results, the gate processing is done on a synthetically larger frequency range (using modeled extrapolation) to minimize out-of-range convolution effects. It is still advisable that the gate not be any narrower than a few resolution intervals.

It is useful to note that the gate shape displayed in time domain is the same as that used during the frequency-with-time-gate process but it does not have the gate calibration term (discussed in the previous paragraph) applied. The gate display in time domain is primarily intended to be informational.

A second type of question that often arises is on how to use and interpret FGT results on transmission parameters. Suppose one had a device in a fixture, one might think that one could de-embed transmission by simply placing a gate around the appropriate place. Unfortunately, it is not quite that simple. Consider the time domain representation of an impulse traveling through our fixture + DUT assembly.

In the time domain sense, an impulse is incident from the left and, at the first interface, some is reflected and some is transmitted. The transmitted impulse then sees the output plane of the DUT and again, some is reflected and some is transmitted. The transmitted impulse (labeled '*A*' in Figure 14-22) goes to the receiver and this is the first response observed in the time domain transmission measurement.

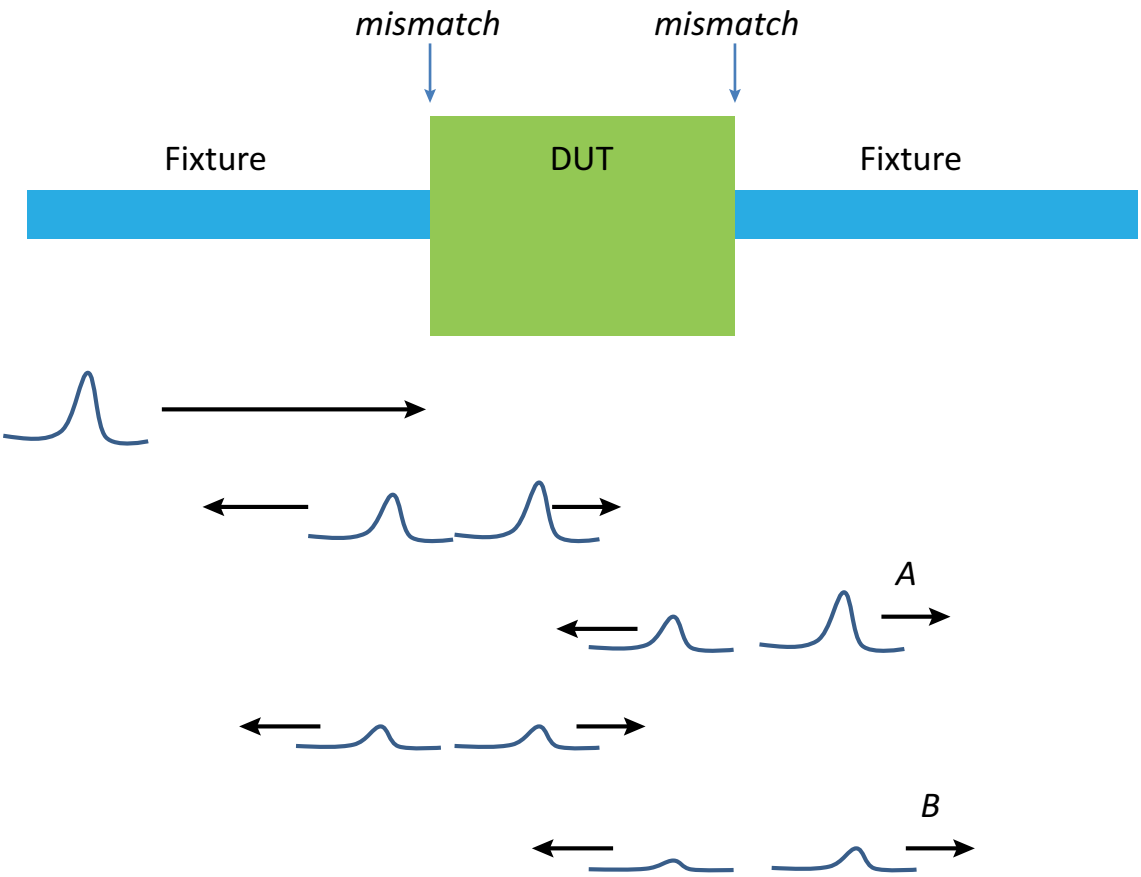


Figure 14-22. Illustration of Gating Effect on Pulse Re-reflections

If one follows the remaining pulse energy, there is an internally reflected impulse in the DUT that again emerges towards the receiver (labeled *B*). There may be additional re-reflections that contribute depending on the loss and reflection levels. Now if one places a gate around '*A*', one will remove the contribution of re-reflections and this may reduce the ripple in the FGT response. This is not complete de-embedding, however, since '*A*' includes loss effects of the fixture as well as any incident mismatch—those effects were not removed by this gating process. For that fuller level of correction, more traditional de-embedding steps usually need to be followed.

Saving Gated Results

As usual, .txt and .csv formats (along with the graphical formats) can be used to save post-processed results. Many users also wish to save gated results in the .sNp file formats but this is not enabled by default to avoid confusion on what the S-parameters represent. This can be turned on using a control at the bottom of the SnP Setup dialog as shown below in Figure 14-23. Note that there is a separate section regarding "Enforced Passivity and Causality" is located in Chapter 9, "Adapter Removal Calibrations and Network Extraction".

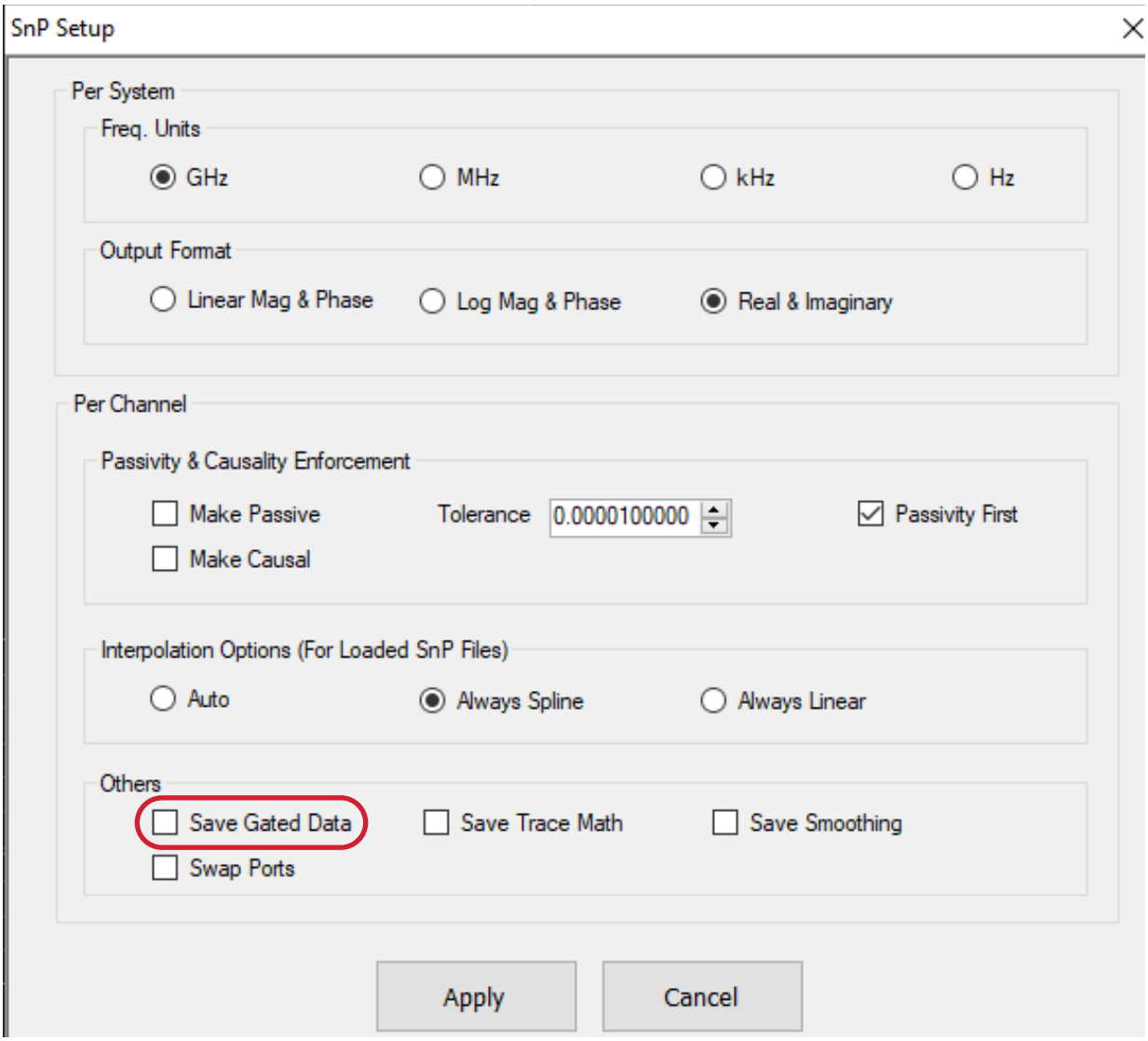


Figure 14-23. Saving of gated data into sNp files is enabled with a check on the SnP Setup dialog

As might be expected, there is some potential for confusion on which gate is applied to which parameter. The following rules are employed:

- If gating is applied on no traces on the current channel, only ungated .sNp data will be saved.
- If gating is applied on at least one trace and the save gated field described above is checked, gated data will be saved for all parameters (that are part of the current .sNp save request). In this case:
  - If all parameters of the .sNp are setup as gated in the current channel, those parameter-specific gate parameters will be used. The data from the last processed run will be used for the save.
  - If not all parameters are setup as gated, the gate parameters of the first gated trace of the same parameter type (transmission/reflection) will be used. If such a trace does not exist, the gate parameters of the first gated trace will be used. If a trace does not exist for the required

parameter, its measurement will be taken from the buffer (if a calibrated parameter) or a measurement re-triggered (if not a calibrated parameter but part of the sNp save definition).

- If gating is applied to the .sNp file, a comment line (! GATING applied) will be added to the header of the file.

Time Gating and Reference Impedances

When a gate starts or stops in a region whose impedance is not the reference impedance (usually 50 ohms), there can be some complications in interpreting the Frequency-with-Time-Gate result. Consider the example of a Beatty line (a 50 ohm transmission line with a length of 25 ohm impedance in the middle) shown in Figure 14-24. It will be used to illustrate a problem with gated reference impedances.

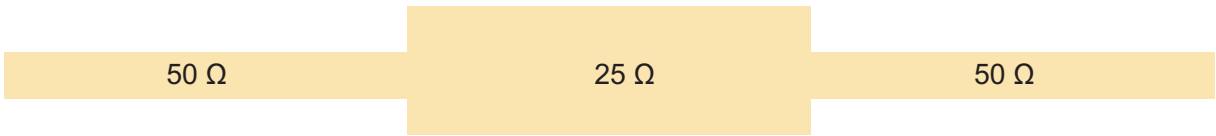


Figure 14-24. Sketch of a Beatty Line Structure

In impulse time domain (low pass in this case) with a gate in Display mode, the result for this example looks like that shown in Figure 14-25. At Marker 1, we reach the initial transition (50 ohms down to 25 ohms) so there is a negative going impulse. Suppose one wanted to gate and study the first half of the Beatty line back in the frequency domain. Since the gate will delete all further reflections, it is as if the line continued at 25 ohms forever and the FGT result like that shown in Figure 14-26 would occur (where there was some other parasitic structure present so the result is not completely flat). If one were to save a .s2p file of this result, this is essentially saying the reference impedance of the far port is 25 ohms. While this may be desirable in some cases, it is normally not what is expected by simulators or de-embedding engines and some additional action would normally have to be taken to get correct results. If referenced to 50 ohms, the result for S11 should have shown large peaks and valleys with spacing dependent on the length of the remaining 25 ohm section.

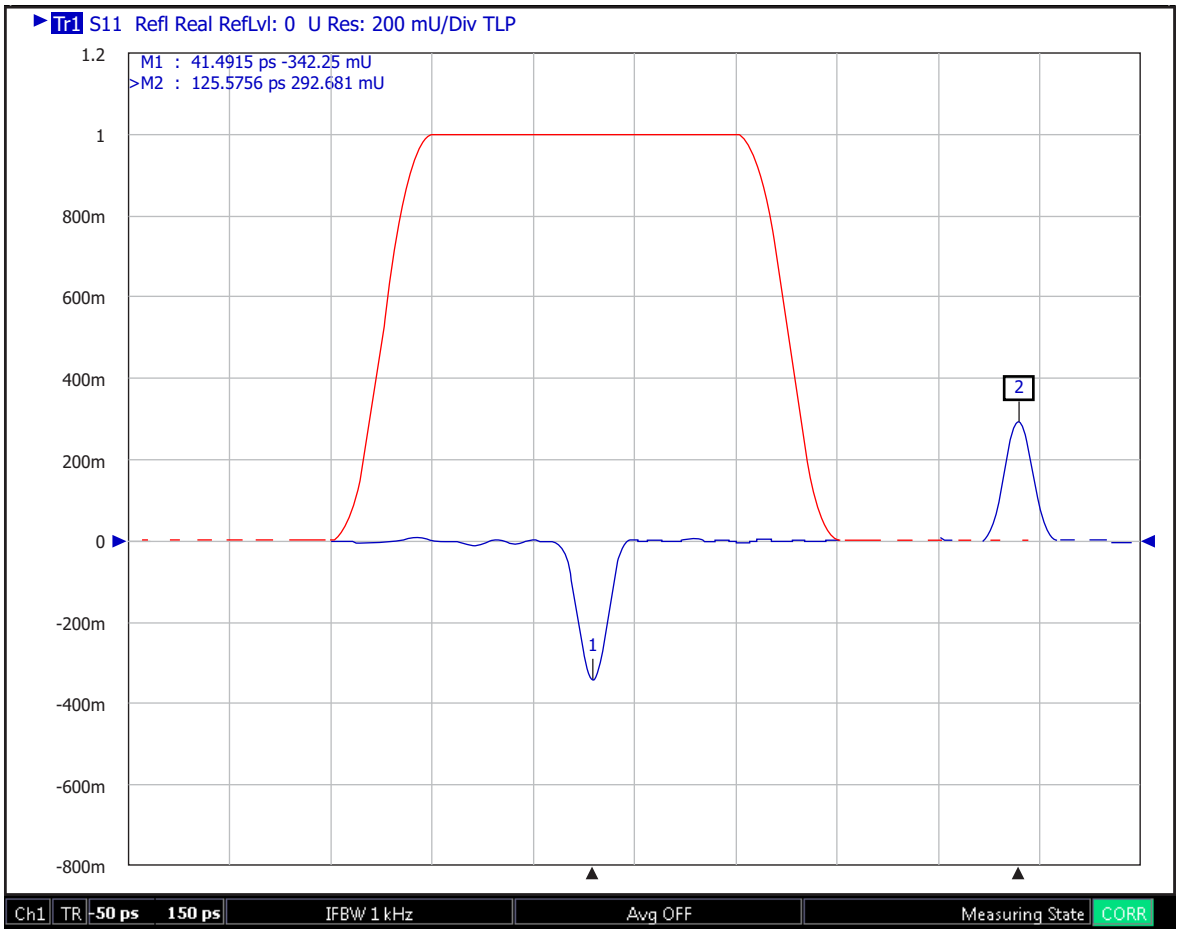


Figure 14-25. Impulse Response of Beatty Line shown in Figure 14-24

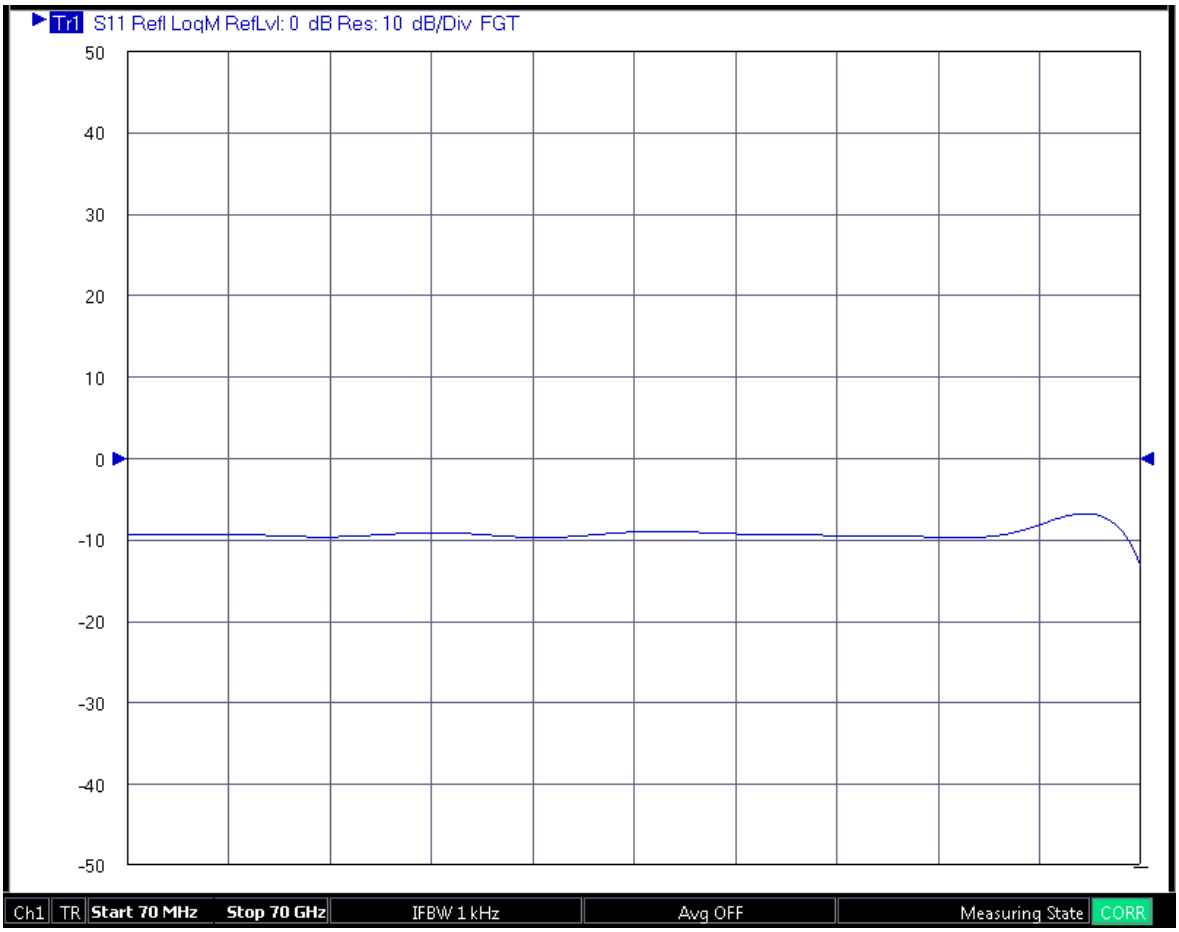


Figure 14-26. FGT Response for Gate Setup shown in Figure 14-25

The Gate Impedance modification tool (accessed via Main | Display | Domain | Gate Setup menu | Gate Impedance | Gate Impedance dialog—see Figure 14-27 on page 14-27) allows one to re-reference the impedance at either end of the gate to something that might be more desirable (often the calibration reference impedance which is usually 50 ohms). This process works by synthetically introducing impedance transitions at Gate Start and/or Gate Stop to get the net result back to the desired impedance planes. If, in the above example, one wanted to use the FGT result to create a de-embedding file for use in a 50 ohm system, then a modification only at Gate Stop would be required where the local impedance is 25 ohms.

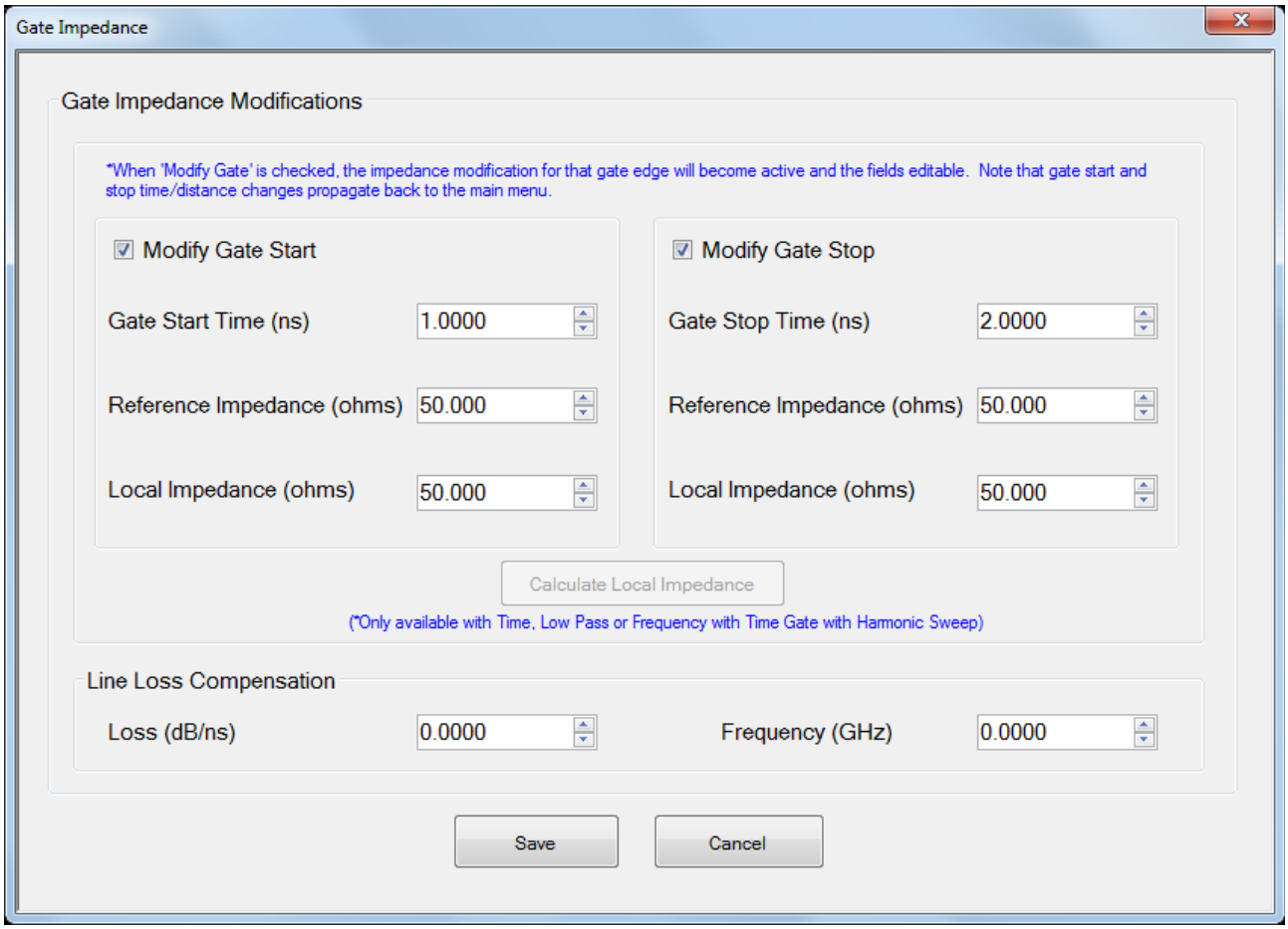


Figure 14-27. Gate Impedance Dialog

Additional Notes:

- The system will calculate the local impedance at Gate Start and Gate Stop locations if requested. This calculation is based on a step response integration of the current data and can be helpful if the DUT impedance levels are not precisely known. Note that if the Gate Start or Gate Stop are placed very close (within a few impulse widths) to physical impedance transitions, the accuracy of this calculation will be reduced.
- If the DUT is lossy, the synthetic modifications should be based on that loss level as well as the impedances involved and the loss-per-unit length at a specific frequency can be entered. A square-root-of-frequency dependence will be assumed and the loss will be treated as being equally distributed along the length of the DUT (up until Gate Stop anyway).
- Local and reference impedances must be positive real.

14-8 Uncertainties in Time Domain Measurements and Recommended Practices

Because native time domain results are an integral of the underlying frequency domain data, understanding the uncertainty in the results is not always intuitive. Consider a measurement of a single point reflect (with little frequency response) so the frequency domain reflection coefficient is:

$$\Gamma = K \cdot e^{-j2\beta L}$$

Equation 14-2.

Where K is the complex amplitude of the reflection (assumed frequency-independent for this example) and L is the distance from the reference plane.

In the time domain (impulse mode), this corresponds to a single impulse at length position L and a magnitude |K|. Now, in the frequency domain, there will be uncertainty including fixed and sloped offsets (from tracking errors, drift, repeatability…) and more complex distortions like ripple (from residual source match and directivity). As these are integrated over frequency, the fixed (in a dB sense) offsets pass-through, sloped offsets have the potential to increase in effect, and ripple-like errors have the potential to reduce in effect due to cancellation (depending on the frequency of the ripple). Immediately, this suggests that uncertainty components that may be of equal weight in the frequency domain may have a very unequal weighting in the time domain representation.

One can look at a simplistic case of a single, relatively large reflection and run Monte Carlo simulations of the effects of various uncertainty terms. Figure 14-28 is an example where source match and directivity terms were elevated (to about the 20 dB residual level) and 1000 iterations were run with random component phases in the measurement. The net scatter was only about 0.1 dB which is noticeably less than the frequency domain uncertainty for a –12 dB reflection. This arises from the correlation between the corrected reflection coefficients at different frequencies.

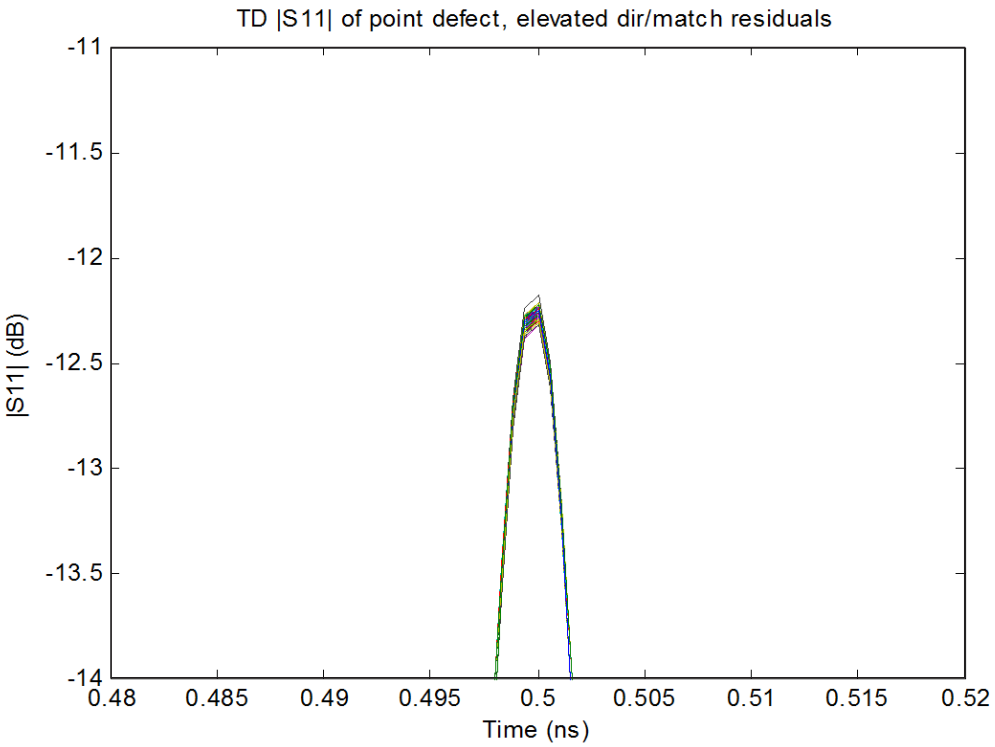
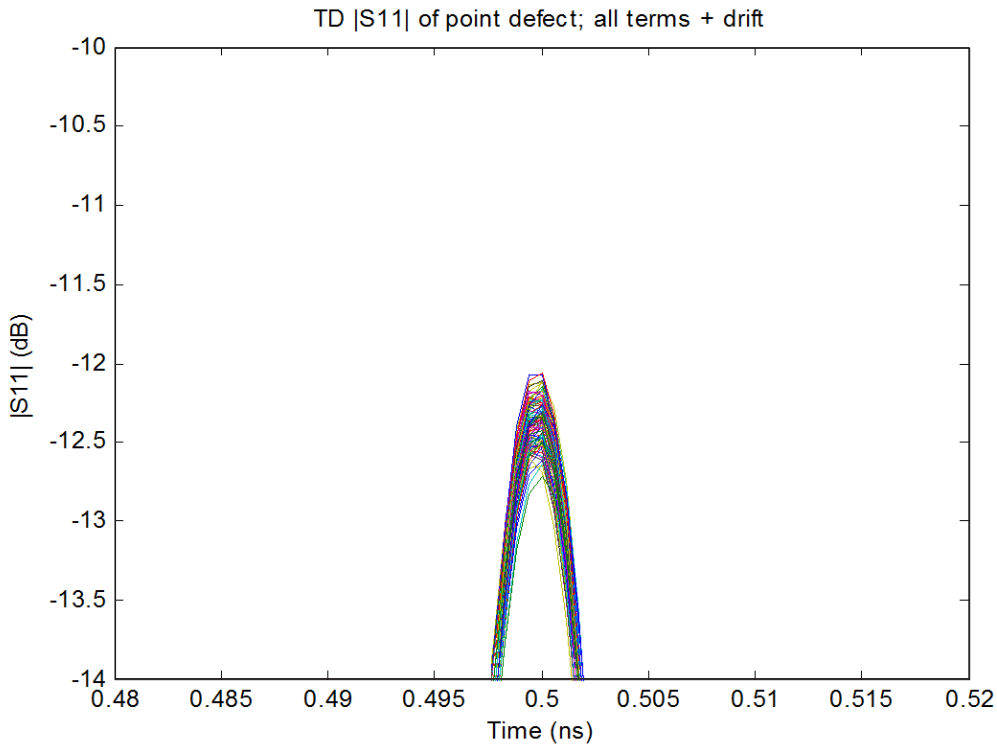


Figure 14-28. Monte Carlo Scatter from a Simulation with Elevated Residual Source Match and Directivity Terms

The Monte Carlo scatter from a simulation with elevated residual source match and directivity terms is shown here. For a single spot defect, these error terms do not have a lot of effect on the time domain result.

Figure 14-28. Monte Carlo Scatter from a Simulation with Elevated Residual Source Match and Directivity Terms

If instead, one added an elevated drift term, the scatter can be much larger as shown in Figure 14-29.

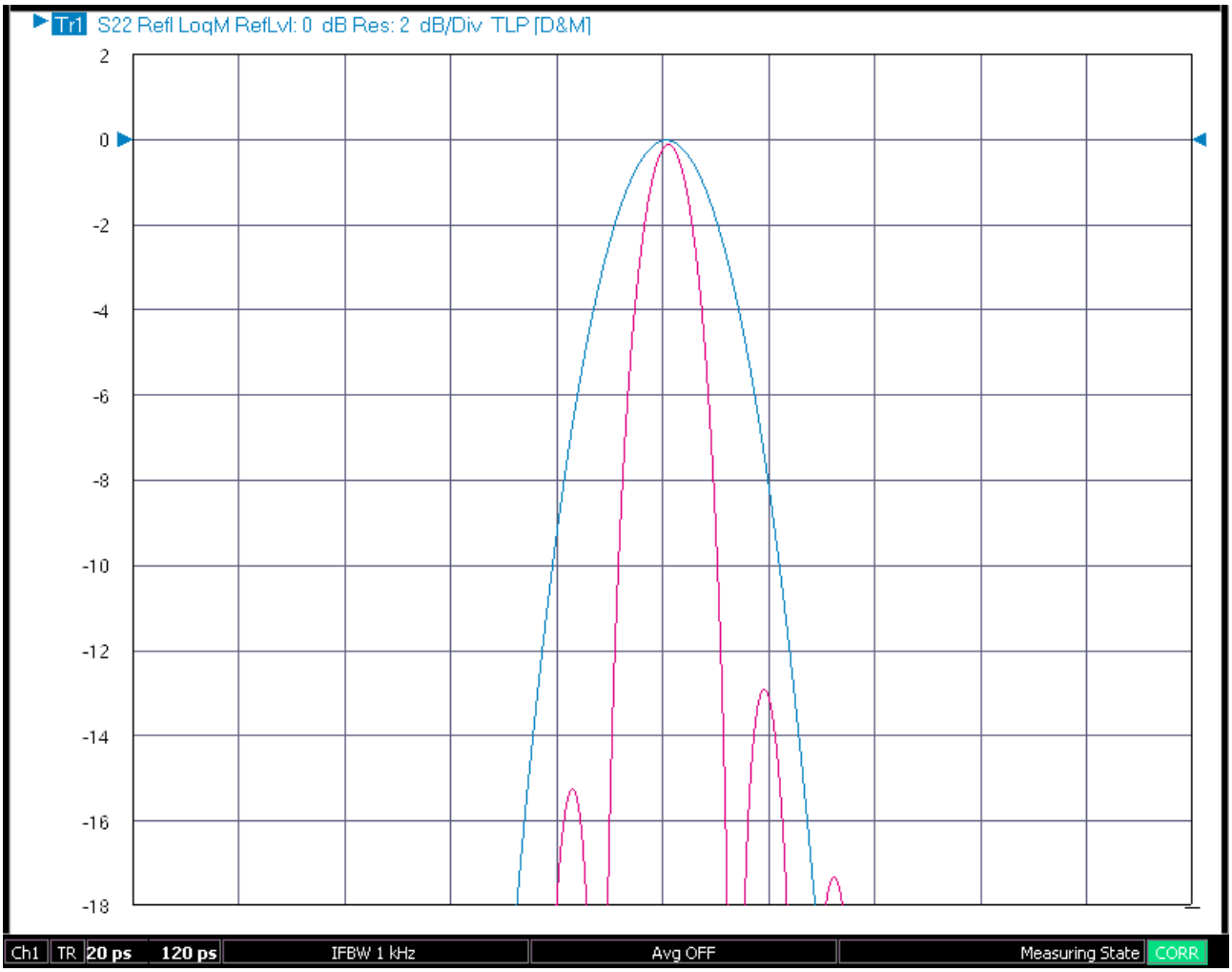


The Monte Carlo scatter from a simulation with both elevated source match/directivity residuals and an elevated drift term is shown here. The effect of the drift error in frequency domain has a major impact on the time domain result.

Figure 14-29. Monte Carlo Scatter from a Simulation with both Elevated Source Match/Directivity Residuals and an Elevated Drift Term

The scatters in Figure 14-28 and Figure 14-29 are particularly simple examples and things become much more complicated when the DUT has to be represented by multiple defects in time. Not only is the correlation between frequencies reduced, but now the windowing function (discussed in an earlier section) can play an outsized role in the uncertainty.

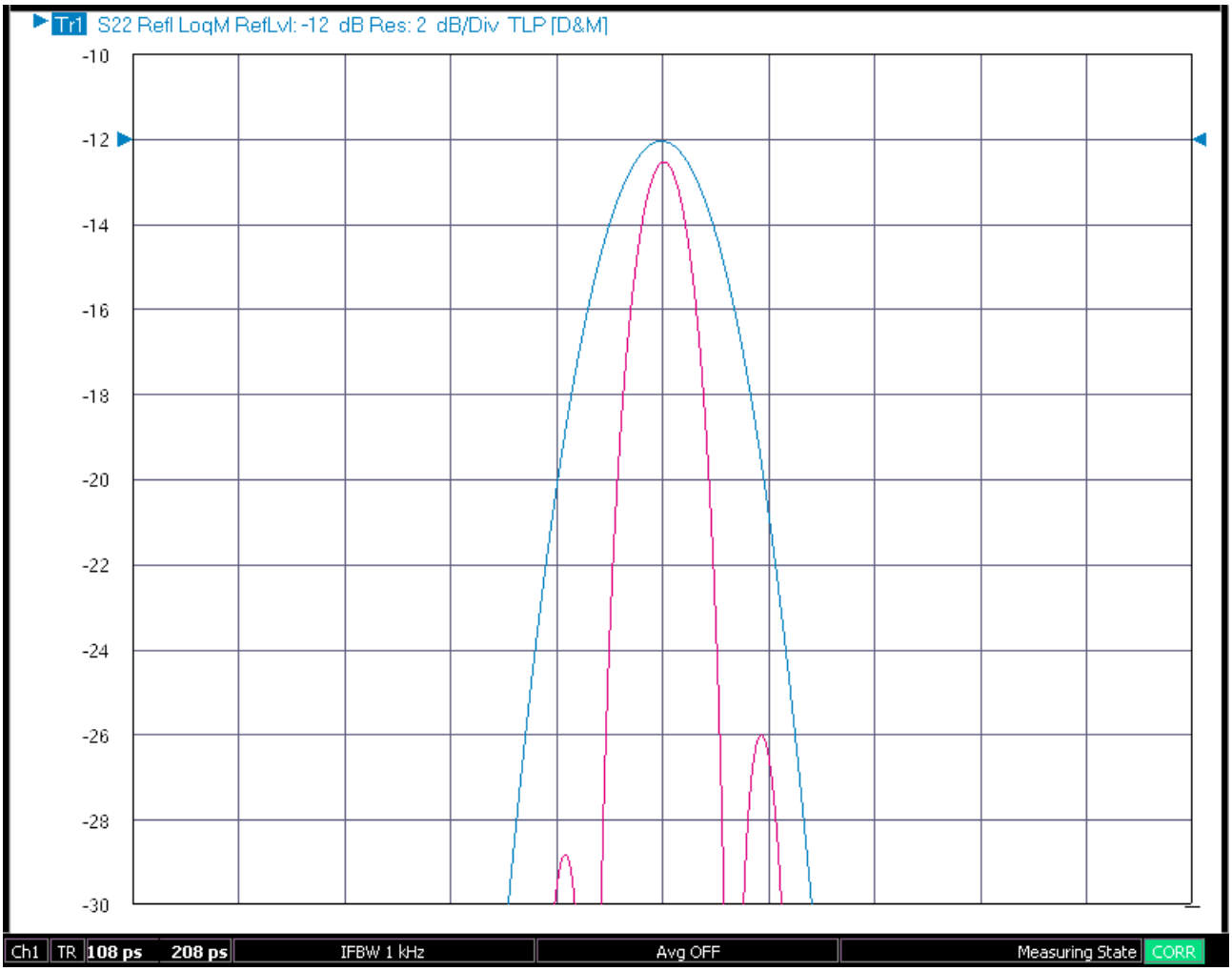
To understand this last concept, recall that the windowing de-emphasizes the extreme frequencies of the data and an aggressive window removes quite a lot of that data. When the structure being tested is complex, the useful information tends to be spread over a wider bandwidth, hence it is more likely that windowing choice may have an impact on the result. In the first example (Figure 14-30), the DUT consisted of a single large reflect and two very different Kaiser-Bessel windows were used (ß=0.5 and ß=12). In terms of the peak amplitude, there is less than 0.03 dB difference.



The difference in results for two different windowing choices is shown here in the case of an isolated time domain high-reflection point. In terms of peak amplitude, the window choice [Kaiser-Bessel with parameter ß= 0.5 (red) or ß = 12 (blue)] had little effect.

Figure 14-30. Results for Two Different Windowing Choices for Isolated Time Domain High-reflection Point

In the next example (Figure 14-31), the portion of interest is a –12 dB reflection and there is another ~ –25 dB reflection 70 ps away. In this case, the window choice makes a much larger difference with the aggressive window (ß= 12) being less correct. The salient difference here is the presence of a non-negligible secondary reflection that is only a few resolution-widths away. This simple example illustrates that time domain uncertainty analysis is much more DUT-dependent than is frequency domain uncertainty analysis and very different effects can dominate. For assistance in evaluating the uncertainties in particular situations, contact Anritsu.



This experiment is similar to that shown in Figure 14-30, except there is a 2nd time domain defect (about 13 dB lower in amplitude) positioned 70 ps away. In this more complex time domain structure, the window choice [Kaiser-Bessel with parameter ß= 0.5 (red) or ß= 12 (blue)] had a much more significant effect.

Figure 14-31. Results for Two Different Windowing Choices with 2nd Time Domain Defect

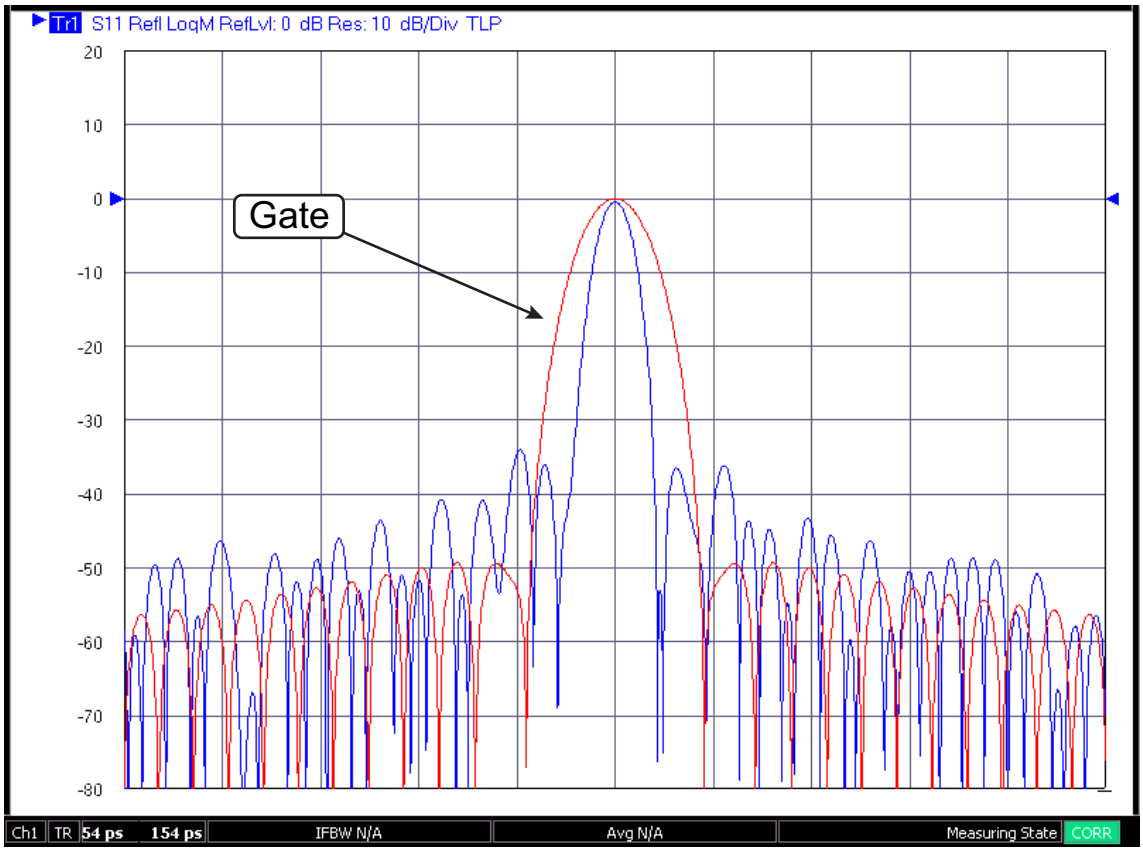
When one looks at step response time domain, there are additional issues in that the result is a time integration of the transform integration. In some sense, this secondary processing compounds the problems already discussed. One dominant aspect of this is the DC term discussed previously. Since the DC term transforms to a constant offset in the impulse response data, the step response will acquire a slope which can dwarf the actual reflection changes if the integration time is long enough. To add another layer, impedance is often plotted in step response and that represents another transformation. Combining these, one gets a reasonably complicated total transformation.

$$Z_{step} = Z_0 \bullet \frac{\int dt\left[K\int df \cdot S_{xy}(f)\cdot e^{-j2\pi ft}\right]+1}{\int dt\left[K\int df \cdot S_{xy}(f)\cdot e^{-j2\pi ft}\right]-1}$$

Equation 14-3.

The main added uncertainty effect in the impedance transformation is one of sensitivity expansion at extreme impedances. Since the impedance changes more for a given S-parameter change (usually reflection) as one gets further away from Z_0, any of the earlier effects will be magnified for very high or very low impedances.

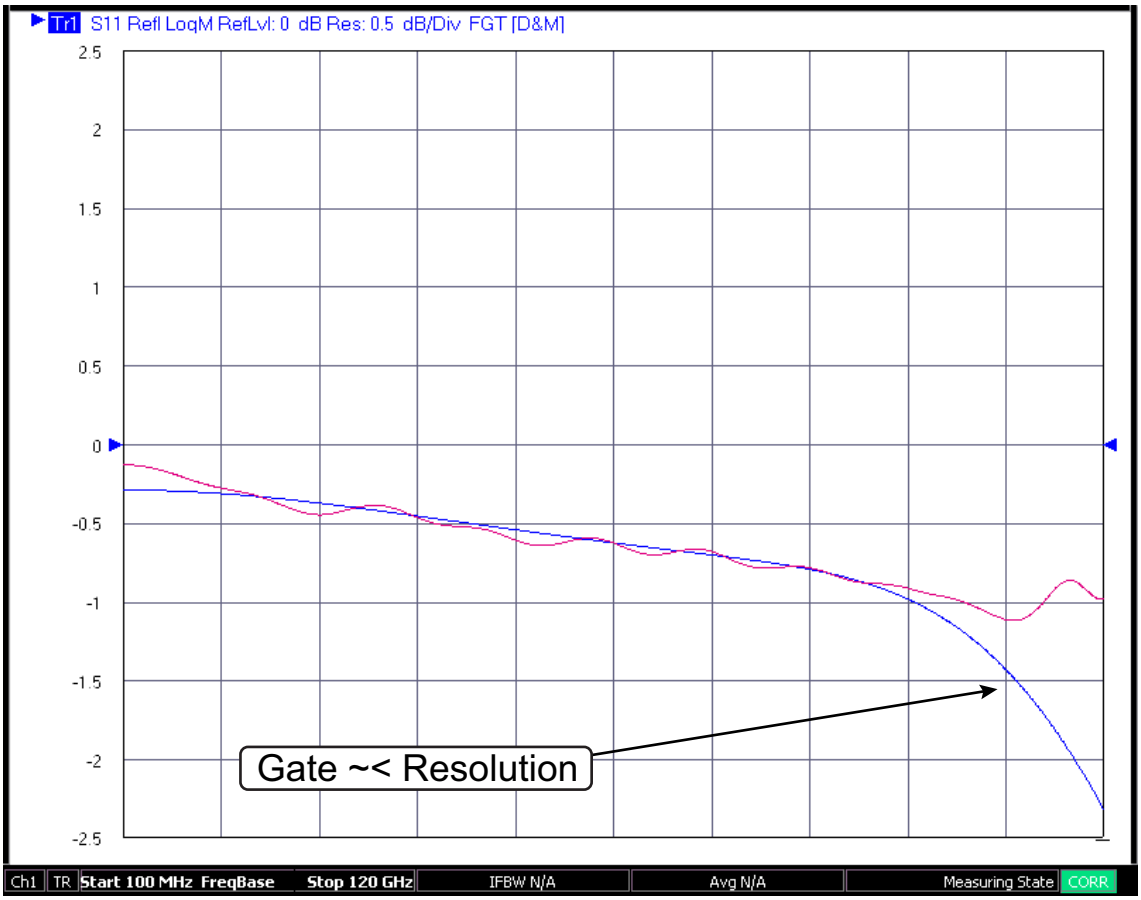
Shifting from time domain representations to the additional transform back to frequency domain (gating), one can also see a complicated uncertainty picture as alluded to earlier. Because of the windowing process, some information is lost in getting to the time domain which cannot be recovered. In addition, if the gate is on the order of resolution in width, then additional distortions occur due to the gate and window shape interactions. It is easy to envision the problem scenario in the plot below (Figure 14-32). The central lobe and its sidelobes represent the information about the defect to be analyzed. Since the gate is cutting off some of that information, the FGT response is now a distorted representation of that defect. While such an analysis may be useful in some cases, it is not a typical de-embedding-like application where one is trying to remove secondary defects.



The selection of a gate that is getting narrow relative to resolution is shown here. This choice will tend to increase errors if frequency-with-time-gate results at extreme frequencies.

Figure 14-32. Selection of a Gate that is Getting Narrow Relative to Resolution

An example is shown below (Figure 14-33) where two different gate widths were used to isolate the same point defect in reflection. In this case, the wider gate was actually a bit too wide in that some secondary reflections remained resulting in the midband ripple. The narrow gate was, however, too narrow and resulted in major distortions at high and low frequencies. Note that this is a common occurrence: as the gate width gets too narrow the errors will increase first at the extreme frequencies.



A frequency-with-time-gate result for two gate width choices: one that is probably too wide for optimum isolation (red) and one that is probably too narrow (blue).

Figure 14-33. Frequency-with-time-gate Result

Optimizing Time Domain-related Uncertainties

Summarizing some of the above analyses, one can arrive at some general recommendations for optimizing time domain-related uncertainties:

- Do not use a more aggressive gate than necessary to get the sidelobe levels where needed. If there is a particular low-level defect near a large defect that needs to be identified, one of the parameterized advanced windows may be helpful.
- Pay particular attention to cable motion and drift if time domain uncertainty is critical.
- If using step response (or in some cases, frequency with time gate), pay attention to the lowest frequency used. A lower start frequency (if the DUT bandwidth permits it) can allow better DC extrapolation and more accurate step integrations. A lower IF bandwidth and/or averaging can also help in this regard.
- When using gating, do not select a gate shape that is more aggressive than needed and do not make the gate narrower than needed to isolate the defect of interest. The frequency extremes are most likely to be affected when the gate is inappropriate.

14-9 Eye Diagram Generation (Option 47 Required)

One feature in the time domain system is the ability to calculate an eye diagram representation of what the currently measured trace data would do to a digital data stream (that can be configured by the user). This is particularly valuable in seeing the data stream signal integrity issues that could occur with a given transmission path and can help with building up subsystem simulation results. Since the eye diagram computation is per-trace, one can configure a single channel having frequency domain, time domain impulse response, TDR-like and eye diagram traces simultaneously and all responding to the same live data.

Eye diagrams have been discussed extensively in the literature, but the basic concept is that they represent a data stream that has passed through an analog channel and it is fundamentally an analog data object where level and transition distortions can be analyzed. In this sense, it is a natural extension of VNA measurements.

A generic eye diagram plot shown in Figure 14-34 illustrates some of the key attributes that will be discussed.

A “heat map” type of plotting is used for eye diagram plotting with the color tending to blue for low occurrences and working up through the rainbow to red for very high occurrences.

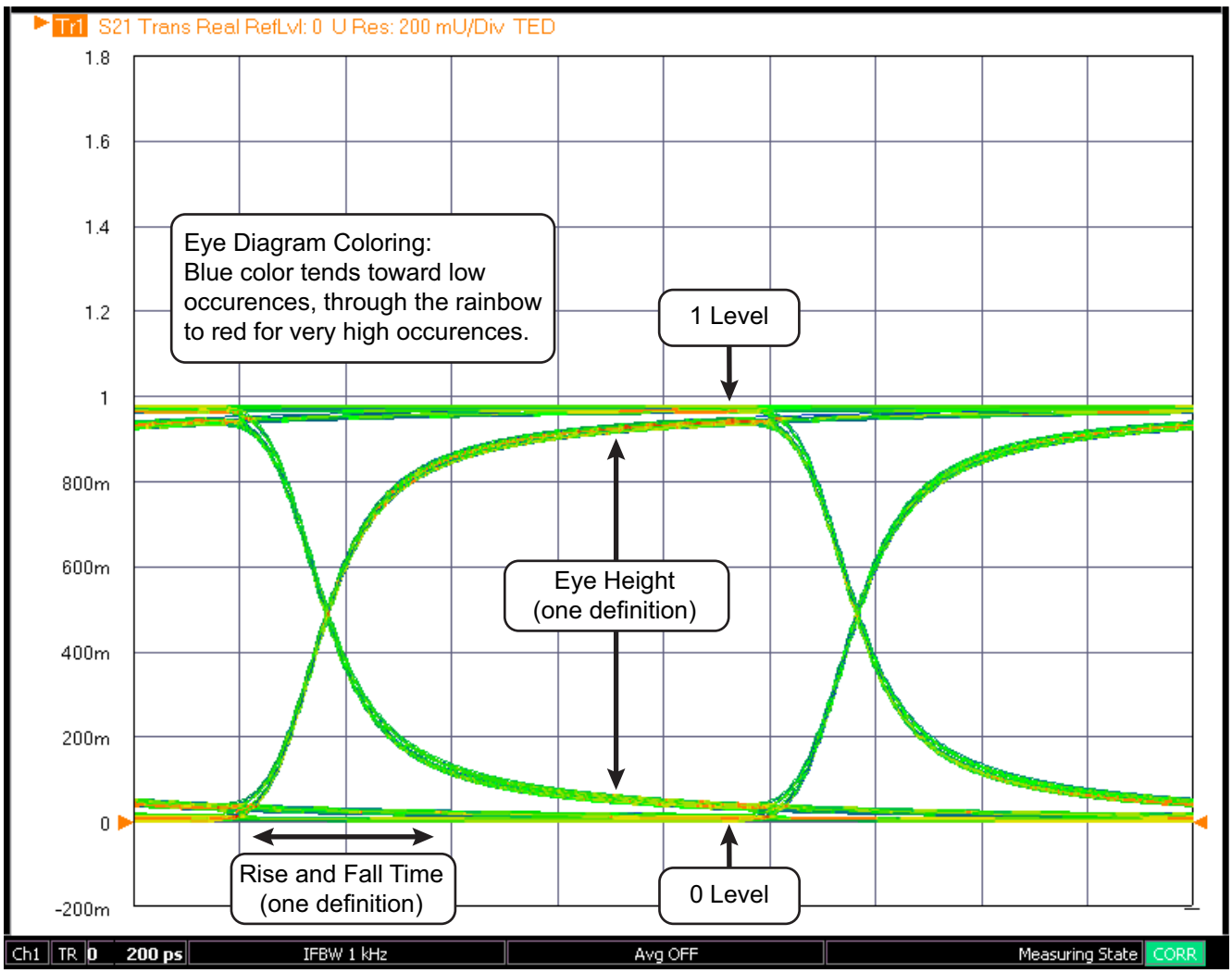


Figure 14-34. Generic Eye Diagram Plot

There are a few general comments that can be made about the eye structure in Figure 14-34. The more the insertion loss increases with frequency, the softer the edges will become and the eye will further close in from the sides. Lower frequency defects may create structure throughout the eye, not just the edges. Jitter will cause a general edge broadening while additive white Gaussian noise (AWGN) will cause a more complete line broadening.

EYE DIAGRAM CONFIGURATION Dialog

The data stream parameters are configured on the eye diagram calculation setup dialog shown in Figure 14-35. The signaling format is the first choice: NRZ (non-return-to-zero, the standard binary waveform) or PAM-4 (the four-logic-level discrete signaling format where each symbol contains 2 bits of information). The type of data stream is the next choice, with a Pseudo-Random Binary Sequence (PRBS) being the default. A user fixed pattern (up to 64 bits that then repeats) and the K28.5 standard (ISI maximizer, only available for NRZ signaling) are also possibilities. The data rate defines the symbol period that will be used for the evaluation. The high and low levels (1 level and 0 level on the input stream for NRZ; four discrete levels are needed for PAM-4) are primarily for scaling and for setup of limit lines later. The pattern length determines the cycle over which the pseudo-random sequence will start to repeat. As the instrument only plots with finite 'persistence' (the number of cycles entered in the dialog, which defaults to 200), the pattern length is only important if the persistence level gets very high. Note also that a higher persistence level will slow the processing time and screen update rate.

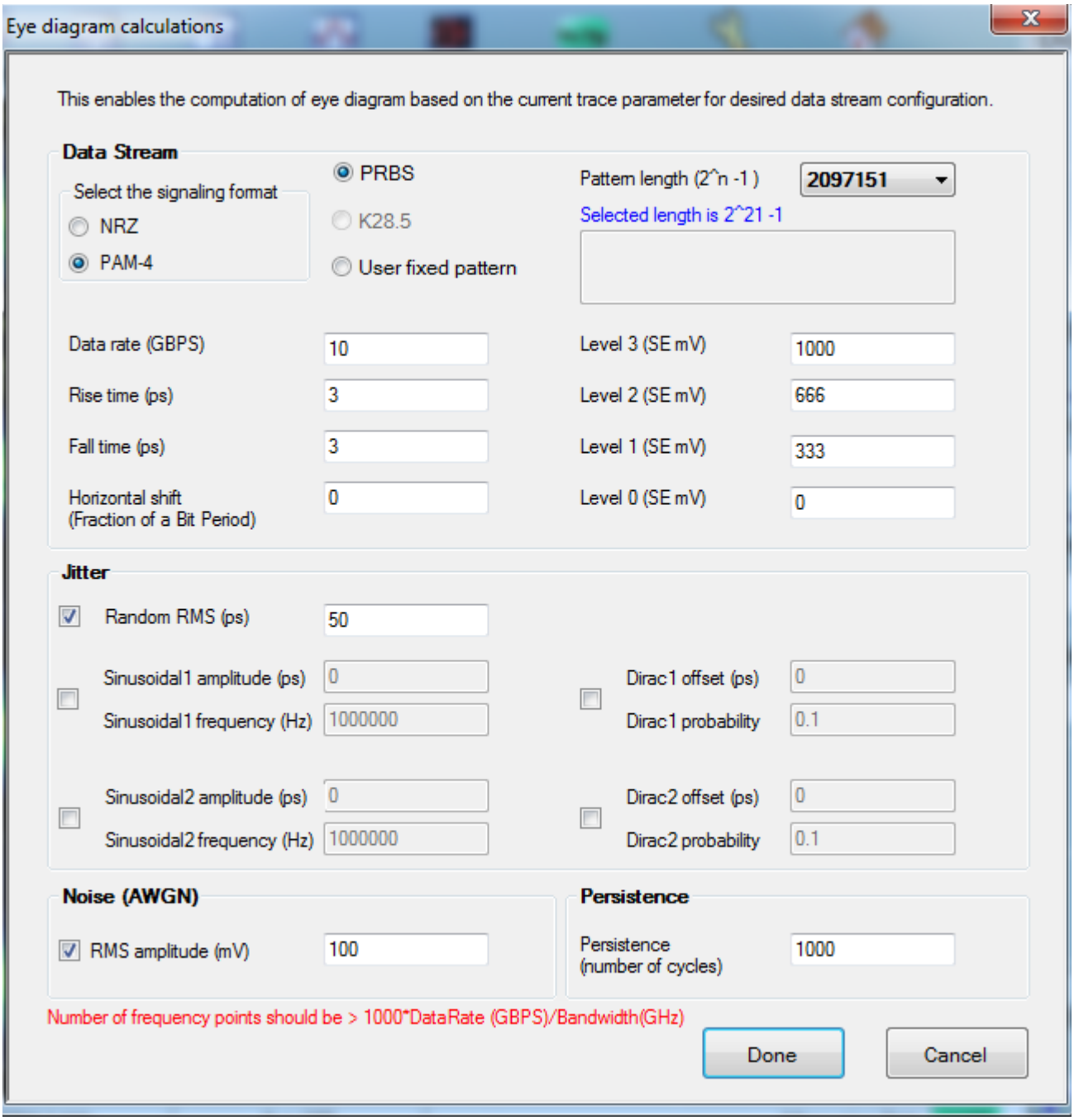


Figure 14-35. EYE DIAGRAM CALCULATION Dialog

Other waveform defects can be entered as shown, including:

Rise and Fall Time:

Note that these values cannot exceed 40% of the symbol period. So for an example data rate of 10 GBPS (100 ps symbol period), the maximum rise and fall times that can be entered are 40 ps.

AWGN Noise:

This is pure amplitude noise that is entered in an RMS mV sense. 5 V is the maximum entry.

Random Jitter:

An RMS time addition to the transitions.

Sinusoidal Jitter:

A deterministic addition to the transitions where the frequency of oscillation of the additive amount can be entered. Two independent sinusoidal jitter mechanisms are allowed. Note that normally the frequency is entered as a substantive portion of the data rate in order for variation to be visible. The amplitude of the jitter cannot exceed 40% of the data rate.

Dirac Jitter:

A fixed time error added to the transitions with an entered probability. Two independent Dirac jitter mechanisms are allowed. The probability is evaluated on a per-transition basis. The amplitude of the jitter cannot exceed 40% of the data rate, and the sum of the probabilities (if both Dirac entries are used) cannot exceed 1.

Eye Diagrams Generation

Here are some points about how the eye diagrams are generated:

- The eye will be based on the data and the parameter in the current trace (eye diagrams are 'per-trace' in the nomenclature used elsewhere in this guide) and will update as the underlying frequency domain data updates. It is possible to have traces in the same channel with frequency domain data, regular time domain results, frequency with time gate results and eye diagrams simultaneously.
- The data waveform is constructed in time domain using the parameters entered and this is convolved with the low-pass time domain representation of the current frequency domain data. Eye diagrams will not be available if the current frequency list does not support low-pass time domain (i.e., the start frequency is not multiple of the step frequency).
- If the bandwidth of the current data is far below the frequency content of the requested data waveform, a very flat, low-amplitude eye diagram will result. If the current data suggests a much larger bandwidth, the eye diagram will approach that of the input data waveform overlaid on itself.
- If the DUT is electrically very long (more than an alias free delay ~ 1/ (frequency step)), the calculation may have difficulties. Using a reference plane shift to reduce that length can help as can reducing the frequency step size in the measurement.
- Since the output of the calculation is a time domain waveform, the result will always be real. Most graph types are allowed due to some unique application requirements but the Real graph type is the most commonly used. The scale will be labeled in Units but the values are in volts with the results driven by the input signal levels specified by the user.
- Any response parameter can be used for the calculation and, particularly in systems with a *loop* option (Option 51/ 61/ 62), any parameter may be practically used. S21 is the most common selection in a 2-port system but all transmission parameters are frequently used. Reflection parameters are normally only studied this way in specialty applications.

Eye Diagram Examples

As an example, consider a long PCB transmission line that has a frequency domain S21 like that shown in Figure 14-36. A 20 MHz to 40 GHz sweep with 2000 points was used to maintain the harmonic sweep requirement for low pass time domain and eye diagram processing. The loss of this DUT is indeed substantial at high frequencies and there is some additional structure above about 30 GHz. The group delay also shows structure at high frequencies.

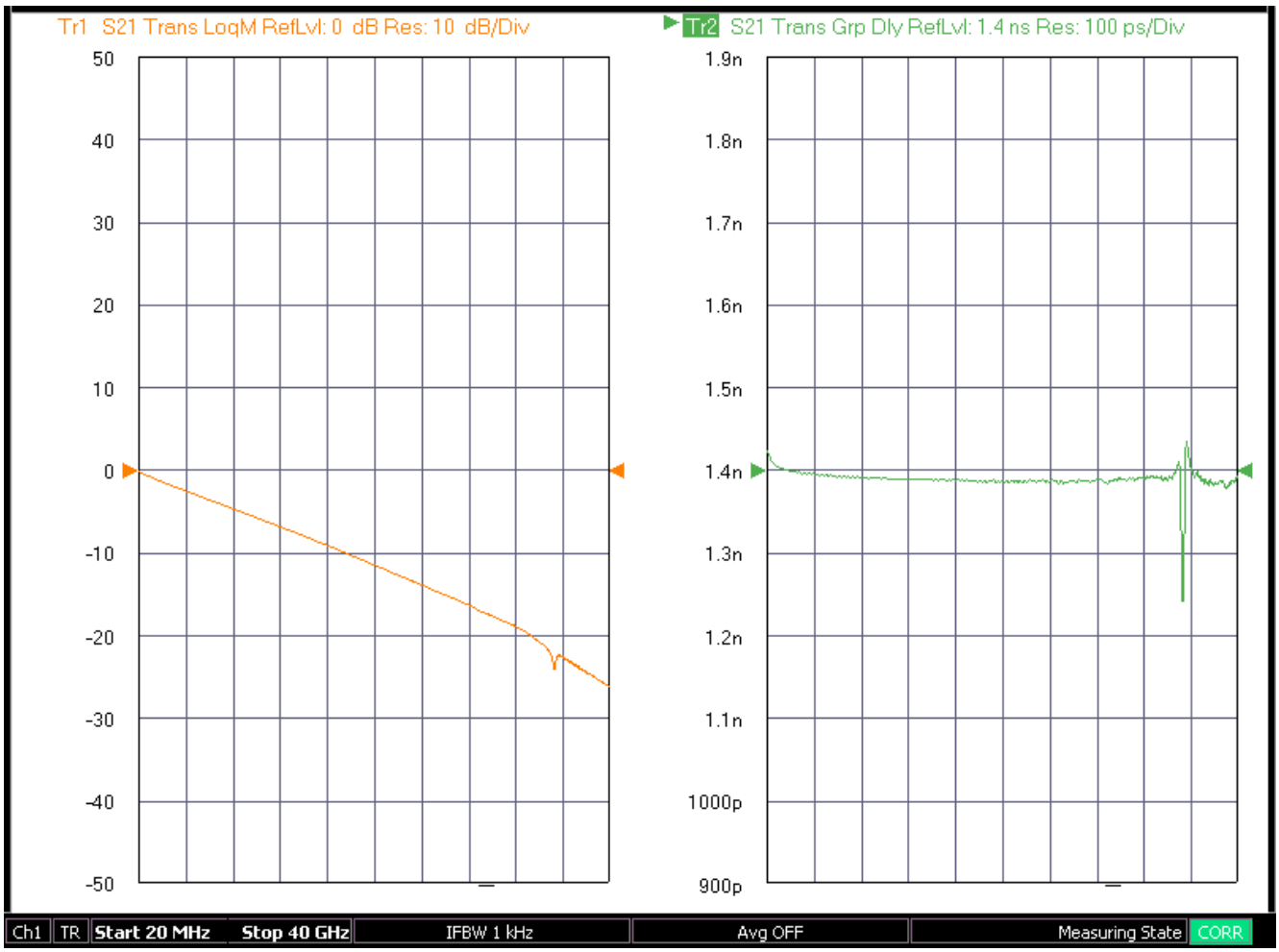
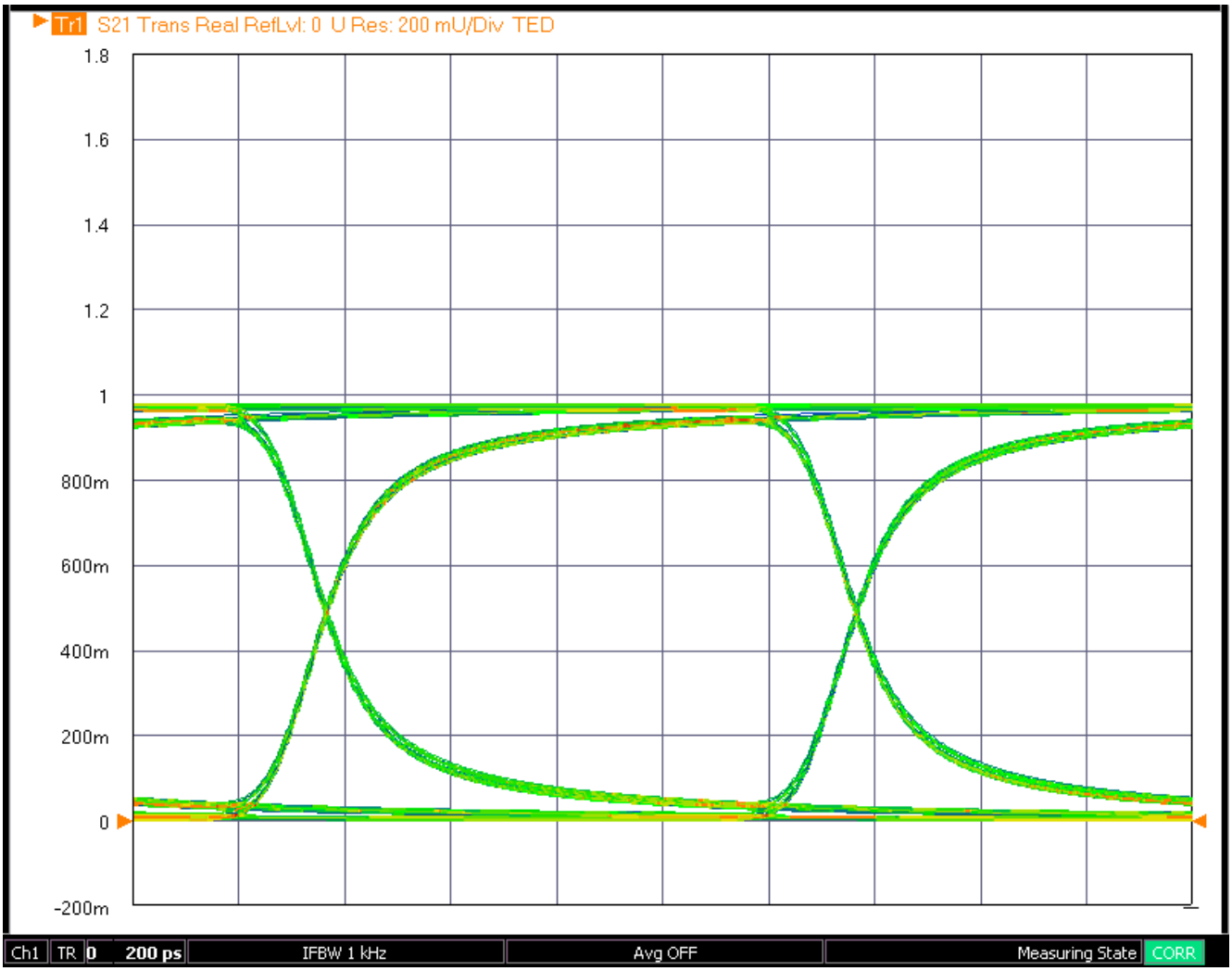


Figure 14-36. Frequency Domain Transmission Response of the Example DUT

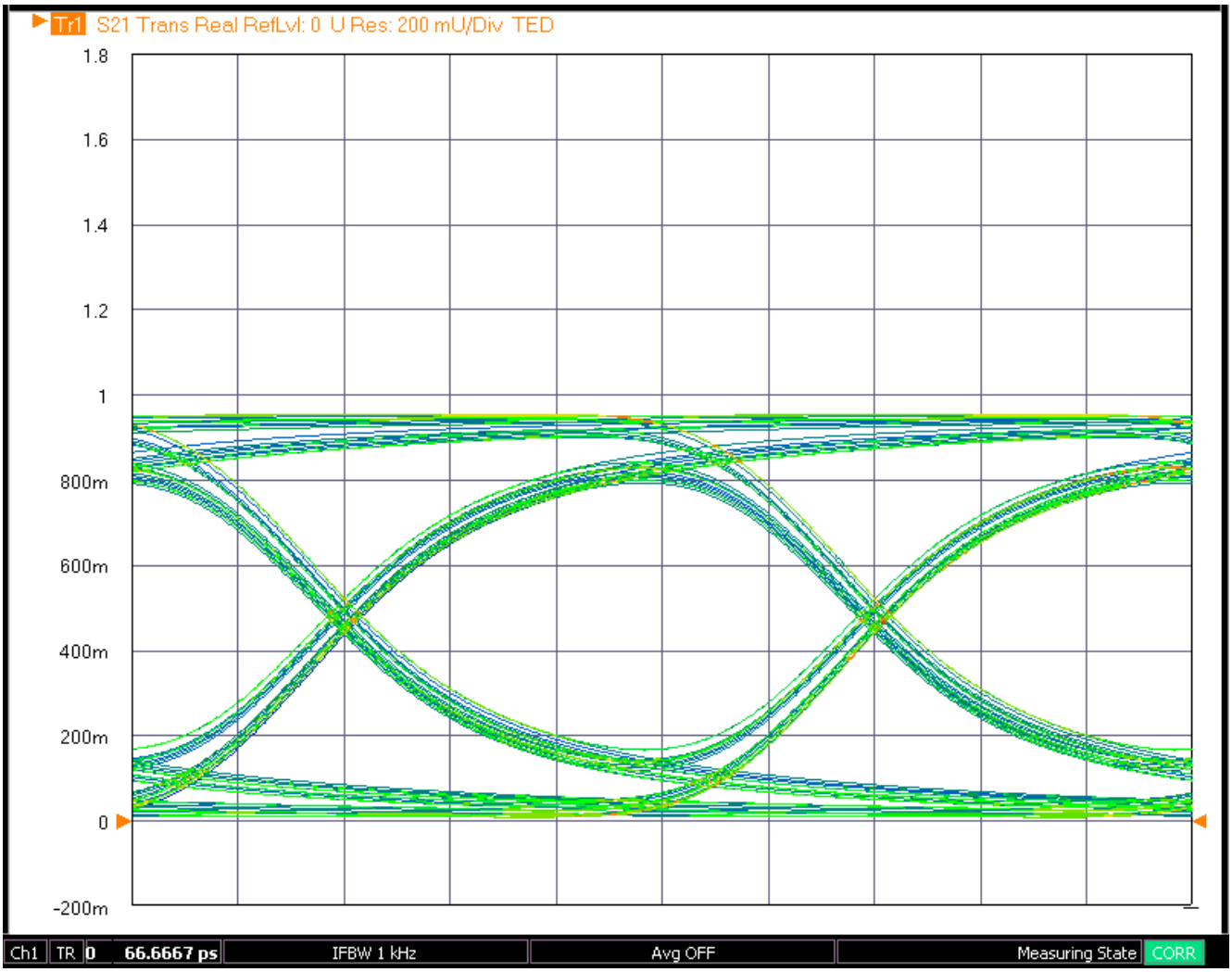
With a 10 GBPS data rate and no assumed jitter, the eye diagram in Figure 14-37 is generated. The eye is still quite open at this data rate.



The eye diagram for the DUT of Figure 14-36 when excited with a 10 GBPS PRBS signal is shown here.

Figure 14-37. Eye Diagram of Example DUT

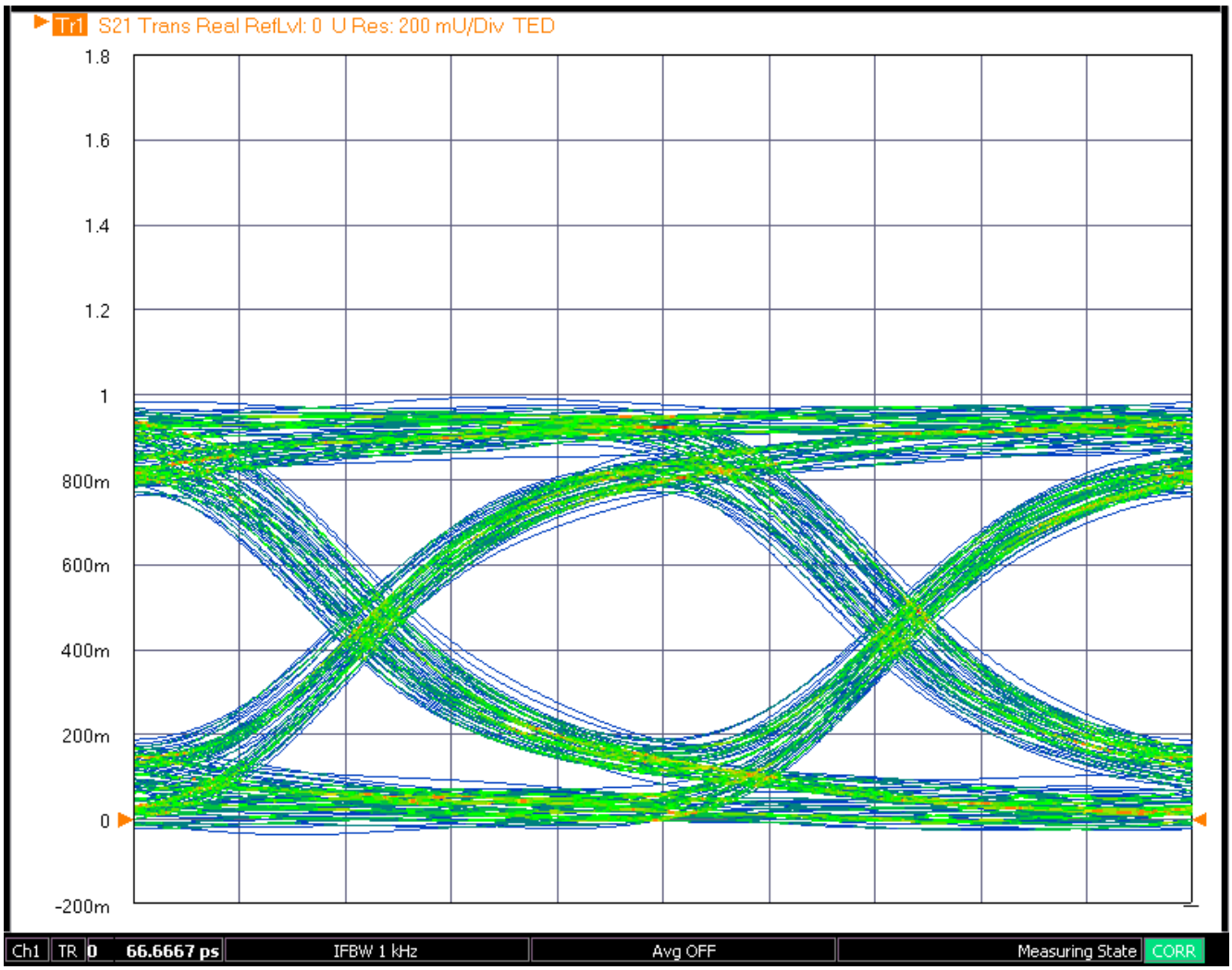
When excited with a 30 GBPS PRBS signal, the eye noticeably degrades further (see Figure 14-38). The eye height has reduced from nearly 800 mV to perhaps 600 mV and the rise and fall times as a fraction of the UI have increased further.



The eye diagram for the DUT of Figure 14-36 is shown here for a 30 GBPS data rate (PRBS sequence).

Figure 14-38. Eye Diagram for Example DUT for 30 GBPS Data Rate

Adding 5 ps RMS random jitter, 100 mV RMS AWGN noise and 5 ps rise/fall times (staying at 30 GBPS) further reduces the eye opening but may come closer to predicting a practical environment (see Figure 14-39). Because the slewing of the DUT is already so severe, the added rise/fall time effects appear more minor than if the DUT had a less dramatic frequency response.



The 30 GBPS example is repeated but with higher AWGN noise and higher incident rise and fall times.

Figure 14-39. Eye Diagram for Example DUT for 30 GBPS Data Rate with Higher AWGN Noise and Incident Rise and Fall Times

A PAM-4 example is shown in Figure 14-40 with relatively high random jitter and AWGN noise. This is a 10 GBPS signal using a 40 MHz–40 GHz sweep range. The default logic levels of 0 V, 333 mV, 667 mV, and 1 V were used. Because of the greater number of possible paths, increased persistence is sometimes used (1000 in this example) to make the display more uniform.

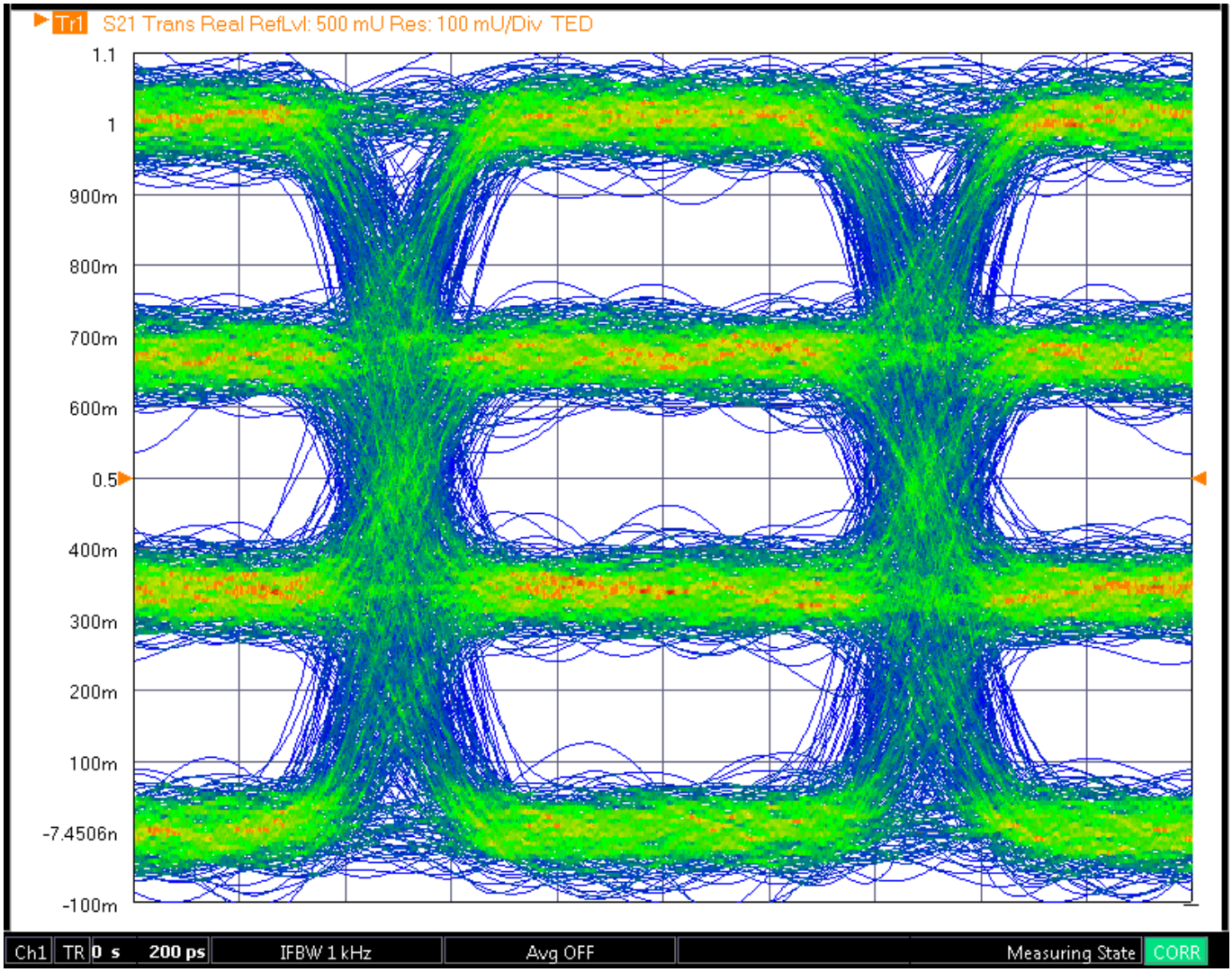


Figure 14-40. PAM-4 Example with Added Jitter and AWGN.

Eye Diagram Readouts

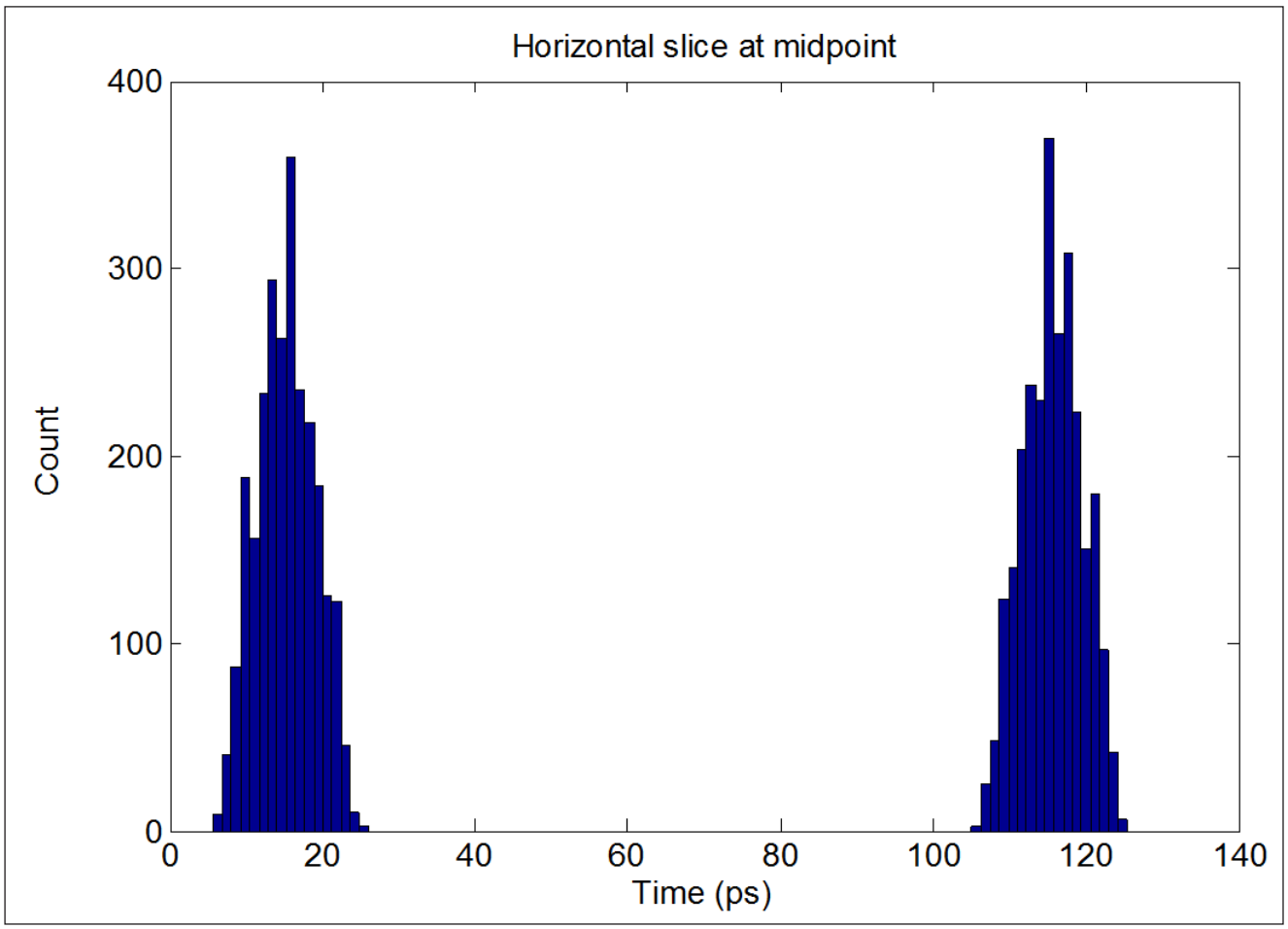
There are a wide variety of readout values (or eye statistics) that are available. These parameters can be enabled in two groups (vertical-based and horizontal-based) and will appear on the trace graticule much like certain marker search parameters (e.g., resonator Q) discussed elsewhere. These parameters are

- Vertical-based (note these are all on a per-eye basis for PAM-4)
  - Logic Levels (0 and 1 for NRZ, four levels for PAM-4)
  - Level mean
  - Amplitude
  - Height
  - Signal to noise
  - Rise time
  - Fall time
  - Duty cycle distortion
- Horizontal-based
  - Width
  - Opening factor
  - Jitter (pk-pk and RMS)

As might be gathered from Figure 14-34, these quantities take on a statistical nature since the data at each point of the eye diagram (in the sense of a slice in the vertical direction or one in the horizontal direction) contains a collection of values. Indeed, the above parameters are evaluated based on histogram-like analyses of the distributions.

Note that this list is separated differently from the amplitude-based and time-based separations in the user interface. The above list is based upon how the parameters are calculated, whereas the user interface split is based on how the parameters are typically used. For reference purposes, the user interface items based on amplitude are Logic Level, Level Mean, Amplitude, Height, Opening Factor, and Signal to Noise; and those based on time are Width, Rise Time, Fall Time, Pk-Pk Jitter, RMS Jitter, and Duty Cycle Distortion. For PAM-4, the vertical quantities are analyzed one eye at a time so there are three sets of numbers in each slice.

As an example, consider a horizontal slice in the middle of the vertical scale. The histogram of this slice is shown in Figure 14-41 and shows two distributions at the transition midpoint; one on each side of the eye. The distance between the mean of the two distributions gives an idea of eye width while the standard deviation of one of the distributions gives a measure of jitter.



The histogram pertaining to a horizontal slice through the middle of an eye diagram is shown here. Although mainly for illustrative purposes, the distributions shown help visualize how the eye parameters are calculated.

Figure 14-41. Histogram Pertaining to Horizontal Slice Through the Middle of Eye Diagram

For the horizontal parameters, these are the definitions used:

- **Width:** (Right distribution mean—3 ∗ standard deviation of right distribution)—(left distribution mean + 3 ∗ standard deviation of left distribution)
- **Pk-Pk Jitter:** Absolute width of the distribution
- **RMS Jitter:** Standard deviation of the distribution
- **Duty Cycle Distortion:** If the rising and falling edge distribution means separate at the mean amplitude point, this difference (in time or in % of the bit period) is the duty cycle distortion.

Similarly, a vertical slice, positioned in the middle of the open eye, is shown in Figure 14-42. Here the difference between the means of the two distributions gives a measure of eye height while the standard deviation could be used to compute a signal-to-noise ratio. The histogram bins were chosen rather randomly for these plots but is the computed statistics that are used for the readout parameters.

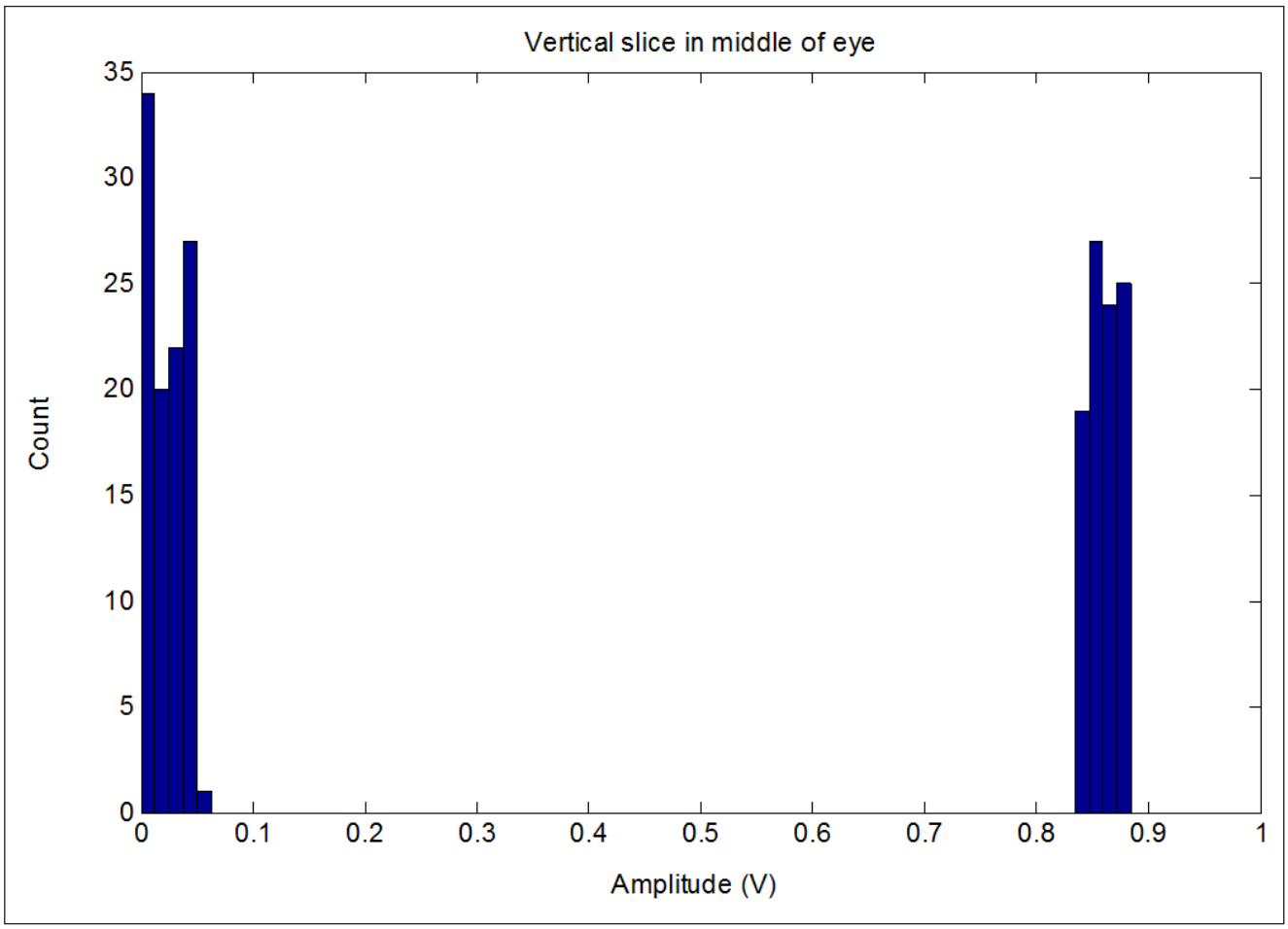


Figure 14-42. Histogram Corresponding to Vertical Slice Through Middle of an Open Eye

Consider first NRZ signaling. For the vertical parameters, these are the definitions used where a slice in the middle of the eye is used:

- **0 level:** Mean value of left distribution
- **1 level:** Mean value of right distribution
- **Level Mean:** Mean of all data
- **Amplitude:** Mean of right distribution—mean of left distribution
- **Height:** (Mean of right distribution—3 * standard deviation of right distribution)—(mean of left distribution + 3 * standard deviation of left distribution)
- **Opening Factor:** Ratio of eye height to eye amplitude
- **Signal to Noise Ratio:** `(Eye level 1 – Eye level 0)/ (sigma of left distribution + sigma of right distribution)`

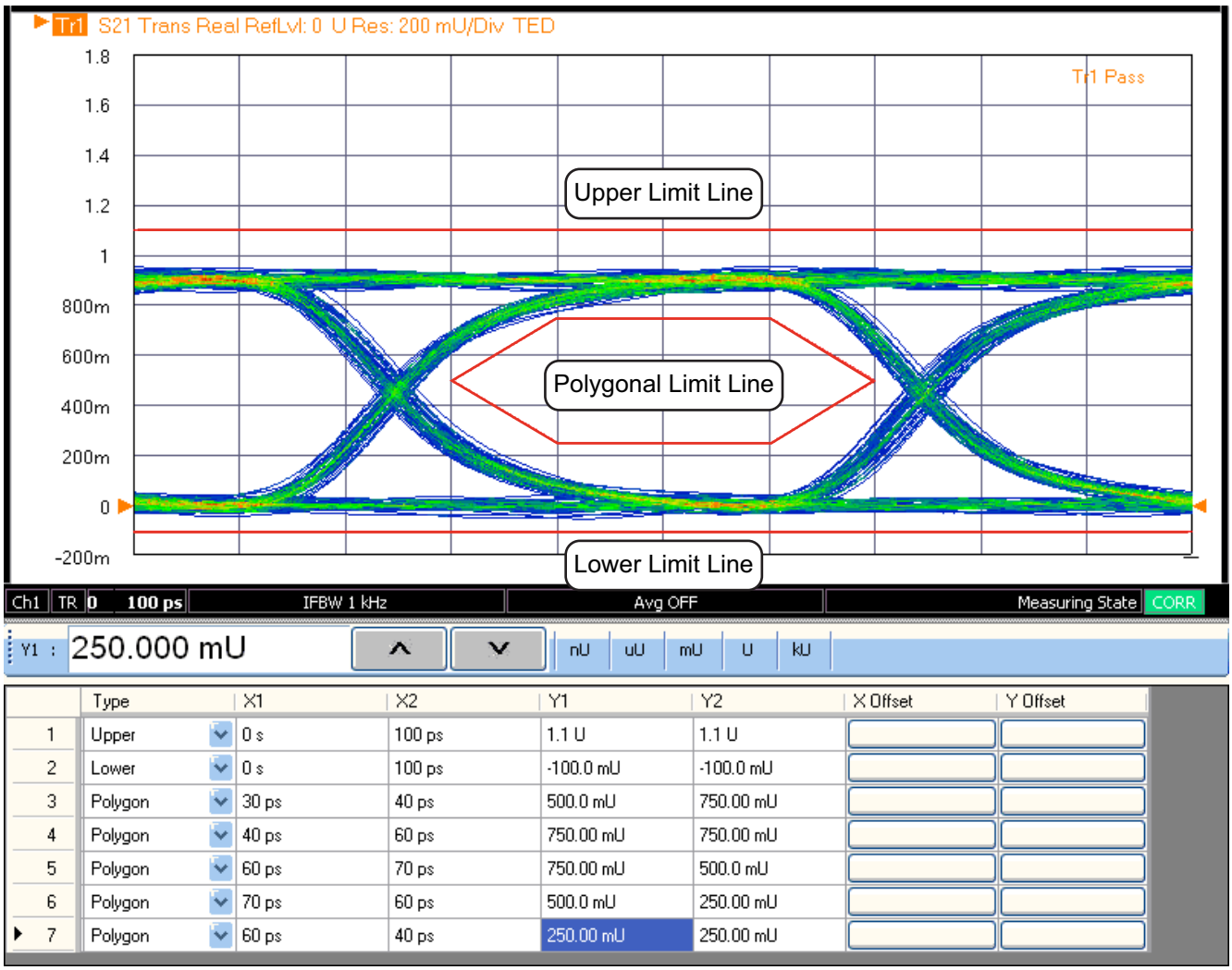
Rise and fall time need a sweep through some vertical slices:

- **Rise Time:** Time from when the mean of the left distribution reaches the set low rise time, e.g. 10%, of the eye amplitude above the eye level 0 (starting from a mean at eye level 0) to the time (after a crossing) when the mean of the right distribution reaches, e.g. 90%, of the eye amplitude above the eye level 0. The low rise time is user-defined, the high fall time is `100% – RiseTime`.
- **Fall Time:** This is the same as above except starting from eye level 1 on the right distribution and working down.

Consider next PAM-4 signaling. The essential quantities are the same as the above, but each eye (thinking vertically) is considered one at a time, so there are three sets of parameters for each slice. The rise time and fall time are bit ambiguous, since different transitions are in play, but the value is computed from the same distributional transfer.

Eye Diagram Limit Lines and Their Configuration

Limit lines are of value in eye diagram measurements just as in other VNA measurements although the topology of those limit lines is somewhat different due to the nature of the data. While traditional 'upper' and 'lower' limits still are useful and serve to demark limits of overshoot and undershoot, there is now an additional concept of a 'polygonal' limit line to constrain the amount of eye closure to be allowed. An example of all three types of allowed limit lines are shown in Figure 14-43.



	Type	X1	X2	Y1	Y2	X Offset	Y Offset
1	Upper	0 s	100 ps	1.1 U	1.1 U		
2	Lower	0 s	100 ps	-100.0 mU	-100.0 mU		
3	Polygon	30 ps	40 ps	500.0 mU	750.00 mU		
4	Polygon	40 ps	60 ps	750.00 mU	750.00 mU		
5	Polygon	60 ps	70 ps	750.00 mU	500.0 mU		
6	Polygon	70 ps	60 ps	500.0 mU	250.00 mU		
7	Polygon	60 ps	40 ps	250.00 mU	250.00 mU		

The display and limit line table in the eye diagram mode is shown here with polygonal line items shown. Normally three or more polygon segments are used to form a closed structure. The polygon will automatically be closed if a complete set of segments have not been entered. The segments of the polygon should be entered sequentially (clockwise or counterclockwise) so the system can figure out which region to enclose.

Figure 14-43. Eye Diagram Mode Display and Limit Line Table

The upper and lower limit lines are covered elsewhere in this guide. The polygonal lines apply only to eye diagrams and that choice in the limit line table only appears in this mode. The polygonal limit line must have at least 3 vertices and must be closed. The system will automatically close the polygon if a complete set of segments have not been entered. The interior of the polygon is always considered the 'Fail' zone.

For PAM-4 signaling, there are 3 polygonal limit lines allowed (scanning each of the three eyes stacked vertically) with their own set of independent vertices. Again, at least 3 vertices are needed for each and the polygon must be closed. A limit failure will be indicated for a violation on any min/max line or any of the polygons.

Eye Diagram Recommendations

As with many time domain-based VNA measurements, using the largest frequency domain sweep range possible where the DUT still has a response and is not radiating severely will lead to the most complete time domain response.

In terms of the calibration quality, cable/connector issues most strongly affect the low frequency and wide-ranging phenomena that have the most time domain impact. Paying more attention to those aspects of the setup and calibration process can pay good dividends. The very low frequency response has less of an effect as the data rate approaches a reasonable fraction of the measurement bandwidth. In these cases, the intrinsic rise/fall time starts to dominate the eye shape.

Depending on the level of mismatch involved, a transmission frequency response calibration (or a simple trace memory normalization) may be an adequate level of calibration. A full port calibration will reduce uncertainty and may be required for reasonable eye diagrams if the mismatch levels are high. Any of the VNA's de-embedding tools may be invoked prior to the eye diagram calculation.

When using jitter and rise/fall time entries, keep in mind that the limits on these scale inversely with the bit rate (as they tend to do in practice).

The eye diagram results are effectively simulations based on frequency domain data of the connected DUT. As such, they may not represent an entire system and they rely on relevant entries for noise, jitter, rise/fall times, and amplitudes.

Chapter 15 — Measurement—Sweep Types

15-1 Chapter Overview

This chapter covers the different sweep types available with the VectorStar MS464xB Series VNA to increase measurement functionality.

15-2 Introduction

A number of different sweep types are available within the MS464xB Series VNAs, including:

- Traditional frequency sweep (defined by a start frequency, a stop frequency and a number of points)
- Log sweep (with a constant ratio or logarithmic step size)
- Power sweep (frequency is constant and power is defined by a start value, a stop value, and a number of power points). Gain compression will be covered as a part of power sweep.
- Frequency-based segmented sweep (frequencies are defined individually or in sub-spans, that are monotonically increasing)
- Index-based segmented sweep (frequencies are defined individually or in sub-spans and can be in any order; all plotting is in terms of index rather than frequency because of the possible direction changes)
- Power sweep (swept frequency) which is a limited multi-dimensional sweep mode that allows the evaluation of gain compression across swept frequency.

The setup complexities in regular frequency sweep and power sweep are minimal and will only be discussed briefly. The segmented sweep possibilities are considerably larger and some explanation will help in setting them up.

The SWEEP SETUP and SWEEP TYPES menus are shown in Figure 15-1 and Figure 15-2

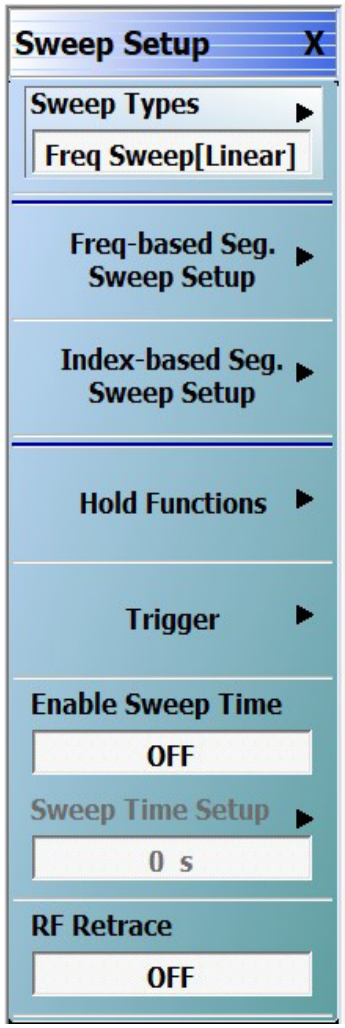


Figure 15-1. SWEEP SETUP Menu

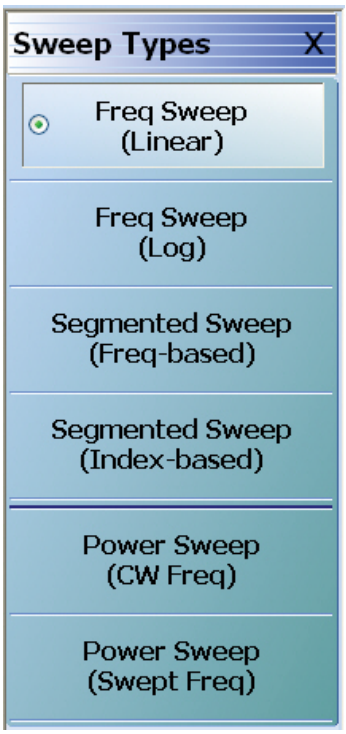


Figure 15-2. SWEEP TYPES Menu

15-3 Setting Up Traditional Frequency Sweeps (Linear and Log)

A traditional frequency sweep is based on a start frequency, a stop frequency, and a number of points (or, alternatively, substitute center/span for start/stop). The number of points is not confined to certain preset values. The minimum number is two (otherwise use CW mode) and the maximum number is usually 25,000.

A mode allowing 100,000 points is also available, but operation is limited to a single channel.

Power Menu in Frequency Sweep Mode

Power entry while in frequency sweep mode is accomplished through the MAIN POWER menu and the POWER CAL menu for frequency sweep, as shown in Figure 15-3 and Figure 15-4.

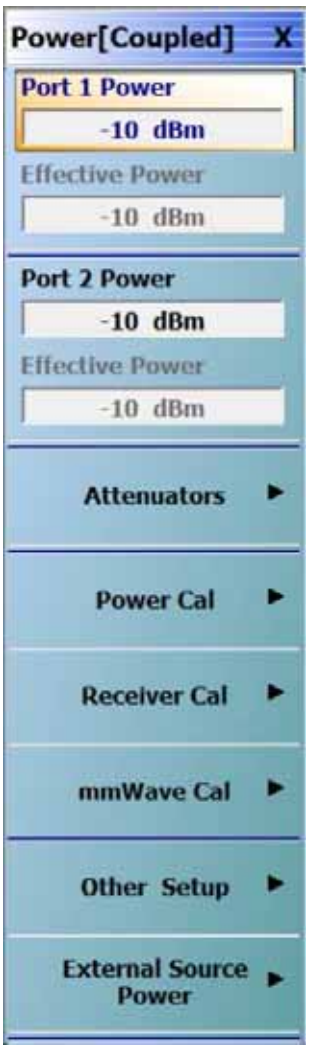


Figure 15-3. Main POWER Menu—Port 1 and Port 2 Coupled—Attenuators Installed

Power Cal Menu

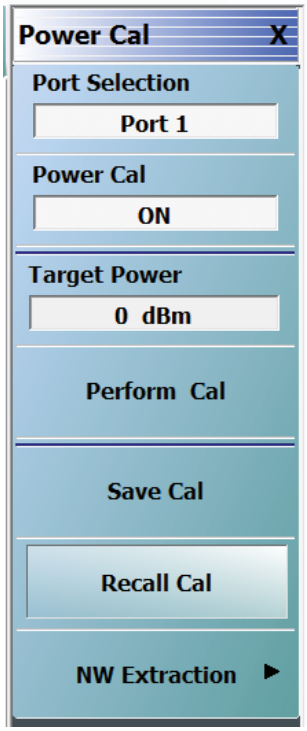


Figure 15-4. POWER CAL Menu

The objective of the power cal is to improve the accuracy of the power delivered to the DUT beyond that provided by the factory ALC calibration (0.1 dB vs. on the order of 1 dB). This is particularly useful if a preamplifier or other network is needed between the test port and the DUT. The exact loss/gain of that network over frequency can be corrected for with reasonable precision. A common setup for executing this calibration is shown in Figure 15-5.

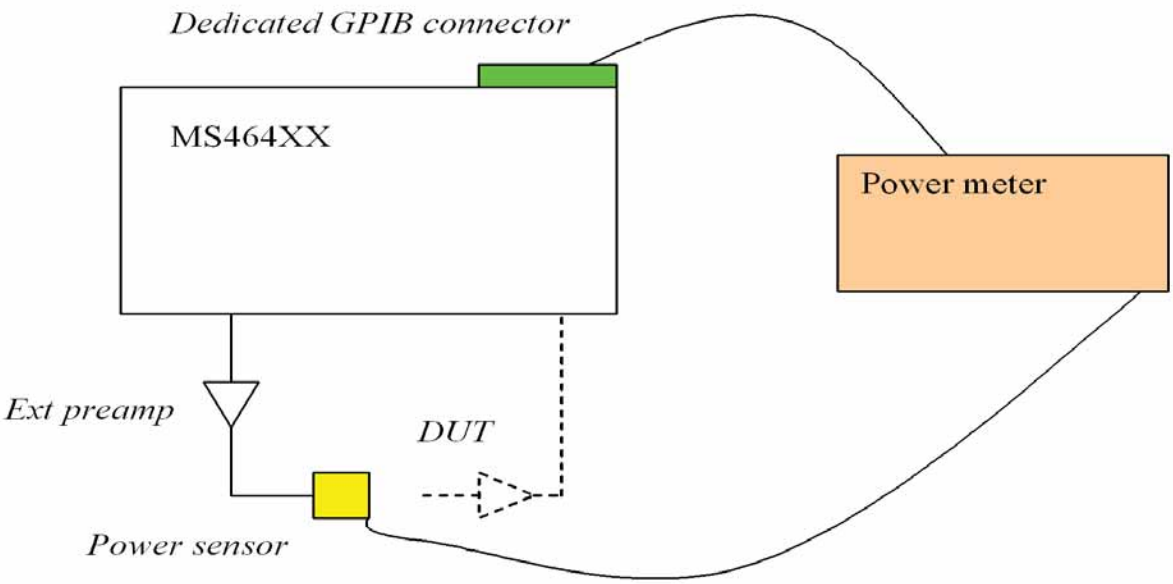


Figure 15-5. Port 1 Power Calibration Example Setup

The relationship between the power entry fields and the Target Power field of the power calibration is important. First a few definitions:

ALC Entry Field (Port 'n' Power on Power Menu):

Where the requested power is entered on the main Power menu for Port 1, Port 2, Src2 out Port 1, etc. (this will also apply to the >54 GHz power fields for the 3739-test-set-based mmWave modes and to the Aux Module power levels for mmWave IMD measurements (when present).

Effective Power Field (on Power Menu):

The read-only fields below the ALC entry fields. These normally reflect the settings of source attenuators (if present) but play an additional role when power calibrations are active.

Target Power Field (on Power Cal Menu):

The entry area on the Power Cal submenu where the objective of the power calibration is entered.

The normal procedure is to enter the desired power at the user reference plane (where the power sensor will be connected for the power calibration). If the gain/loss between the VNA port and the user reference plane is small, the ALC entry field can be set to the same value as the Target Power. If the gain/loss is significant, the ALC field should be offset by an estimate of that gain/loss (ALC entry field should be higher if it is a loss, and lower if it is a gain). This offset need not be extremely accurate; it serves to inform the system of where roughly to set the internal sources in order to achieve the desired user reference plane power. After the power calibration is completed, the Effective power field will match the Target power. If the ALC entry field is later changed (from the value at the time of the calibration), the Effective power field will also change to reflect an estimate of the new user reference plane power. This new value is only an estimate since the calibration was not performed at that power level and the accuracy will decrease as the change from the calibrated level increases.

Example

Target power is –20 dBm and there is about 8 dB of loss from the VNA port to the desired reference plane.

1. Set Target Power to –20 dBm and set the ALC entry field (for that port) to –12 dBm. Perform the power calibration.
2. After the calibration is successfully completed, the Effective power will read –20 dBm and the ALC entry field will remain at –12 dBm.
3. Change the ALC entry field to –15 dBm. The Effective power field will change to –23 dBm (the estimate of the new power at the reference plane).
4. Turn the power calibration off. If the ALC entry field is still at –15 dBm, the Effective power field will change to also be at –15 dBm. With the power calibration no longer applied, the system's best estimate of the power being applied is that based on the internal power calibration.

Note that the source attenuator settings (if Option 61 or 62 is installed) will be reflected in the Effective Power value, even if no power calibration is applied.

Since the power cal performs the calibration at every point, this calibration can be time-consuming (particularly if a slower thermal power sensor is being used). One should exercise some restraint when selecting the number of power points if this time delay will be an issue. Details of power meter connection and setup are covered in the **Operation Manual**, but one should ensure that the power meter GPIB address matches that shown on the MS464xB Series VNA GPIB menu system and that the dedicated GPIB connector on the VNA is used. The dialog that appears when executing this calibration is shown in Figure 15-6.

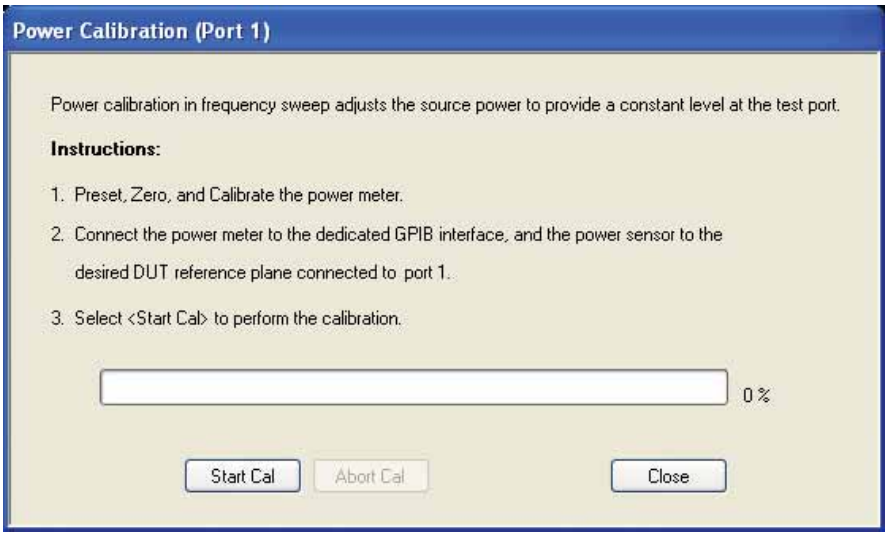


Figure 15-6. POWER CALIBRATION (PORT 1) Dialog Box

Power Meter/Sensor Selection Options

Many possible power meters/sensors can be used with the MS464xB VNA and these choices proliferate with the broadband and millimeter wave versions (see Chapter 17 of this guide) so some guidance may be helpful:

- The ML24XX series of power meters and appropriate sensors can be used if the frequency ranges match (of the sensor and the range over which the calibration is being attempted) and these are usually the default. Control must be via GPIB and the address the VNA will look for is determined by the Ext. Power Meter address entry under System/Remote Interface/Power meters. The highest frequency sensor available in this family reaches 70 GHz and models are available that reach down to 70 kHz.
- Any GPIB meter can be used if it has a HP437B emulation mode. These meters will be expected also at the Ext. Power Meter address. Note that not all emulation modes are equal and the timing may not always behave correctly (particularly at very low power levels where settling time may be longer).
- Broadband (to 110 GHz) sensors are supported (also GPIB controlled, the R&S NRP2 or NRX meter with NRP-Z58 sensor as one example). This will be searched for under the Broadband sensor address.
- Certain waveguide-based sensors are specifically supported for millimeter wave applications.
    - An HP437B-equivalent meter using a W8486A W-band sensor. This meter will be searched for under the W-band address.
    - A D-band sensor (ELVA DPM-06 sensor with appropriate meter with GPIB) can be used above 110 GHz (for ME7838D/G systems mainly) and this is searched for under the D-band address. Normally the W8486A sensor will be used up to 125 GHz if it is available and the D-band sensor above that. The D-band sensor will be used above 110 GHz if the W8486A sensor is not available. Particularly for ME7838G systems, make sure an appropriate waveguide-to-0.6mm adapter is being used in that case (A WR6G adapter is available that covers the full 110-170 GHz range. The WR5G adapter can be used as well over most of this range as its cutoff frequency is 115 GHz).
    - A G-band sensor (ELVA DPM-05 sensor with appropriate meter with GPIB) can be used for ME7838G systems mainly. This sensor is normally used only above 170 GHz but will be used above 140 GHz if a D-band sensor is not available and the calibration requires that frequency range. A 35WR5G adapter is normally used when using this sensor with ME7838G systems.

- Keysight power meters are also supported. The Keysight N191XX EPM series of meters in their native mode are supported under two different address classes: Ext. Power Meter (in lieu of the ML243XX power meter) and the W-band Power Meter (using the W8486A sensor). Since this meter supports sensors that cannot return their frequency limits, it is incumbent upon the user to only connect this meter when the sensor range is valid for the calibration of interest. The VNA will assume the sensor can cover 70 kHz–70 GHz if the meter is connected on the Ext. Power Meter address, and will assume it can cover the extended W-band if connected under the W-band Power Meter address.
- USB sensors from Anritsu are also supported where the sensor's specific frequency range relative to the intended range for calibration is relevant. Multiple USB sensors can be selected and the user will be asked to choose at calibration time.
    - **Note:** Usage of the MA24500A series sensor requires a dual USB Type A male to single USB Type A female cable to supply needed current draw.
- The MA25400A sensors are different from the sensors discussed above in that they use a heterodyne detection scheme rather than a broadband approach. The way this sensor is used in Vector Star calibrations, the integration bandwidth is 7 MHz, and this has some implications. First, the minimum detectable power is in the range of -35 dBm (it can be lower when the sensor is used with a separate PC when the integration bandwidth can be smaller). Second, the accuracy is slightly degraded at very low frequencies (principally below 500 kHz) when the front end corrections start deviating due to the bandwidth.

When not in broadband mode, the user will have the option to select a USB sensor if one is detected. If not, the system will first scan for the Ext. Power Meter address and will use that meter if found and the reported frequency ranges match (note that 437B emulating meters do not report a frequency range so the system will use them regardless). If such a meter is not found, it will next search for the Broadband meter. If that is also not found, the calibration will terminate with an error.

In broadband modes, there are many more choices and the system will switch between sensors/meters to cover the desired range. This is perhaps illustrated best with a series of examples and those are in the chart shown in Table 15-1. The system will start from the left of the chart and use the first option (or combination) that is available. Many of the sweep ranges in the chart are only possible with the ME7838x broadband systems (see Chapter 17 of this guide for more information).

Example:

Setup: Sweep 10 MHz–125 GHz and broadband sensor, an Anritsu 70 GHz GPIB sensor and a W-band sensor are connected.

1. Since no USB sensor is available, it skips the first device column.
2. A broadband and a W-band sensor are available so it will use the 2nd device column

The general principles are, in this order:

- If a USB sensor is connected and selected, use it as far as it can be used.
- Use the connected sensor that can cover as much of the frequency range as possible.
- Use the connected sensor with the lowest uncertainty.

One can always override this decision tree by only leaving connected the meters/sensors that one wants to use.

Table 15-1. Power Sensor Selection Guide (1 of 2)

When Broadband is Selected				
Power Cal Frequency Range	**If USB Sensor Present and Selected**	**Next Choice**	**Next Choice After That**	**Next Choice After That**
70 GHz CW	USB	Broadband	Anritsu GPIB or 437B emulation	W-Band
110 GHz CW	USB	Broadband	W-Band	D-Band
125 GHz CW	USB	W-Band	D-Band	

Table 15-1. Power Sensor Selection Guide (2 of 2)

When Broadband is Selected				
Power Cal Frequency Range	**If USB Sensor Present and Selected**	**Next Choice**	**Next Choice After That**	**Next Choice After That**
10 MHz to 70 GHz	USB	Broadband	Anritsu GPIB or 437B emulation	
10 MHz to 110 GHz	USB	Broadband	Anritsu GPIB or 437B emulation & W-Band	
10 MHz to 125 GHz	USB	Broadband & W-Band	Anritsu GPIB or 437B emulation & W-Band	Broadband & D-Band
10 MHz to 150 GHz	USB	Broadband & W-Band & D-Band	Anritsu GPIB or 437B emulation & W-Band & D-Band	Broadband & D-Band
70 GHz to 110 GHz	USB	Broadband	W-Band	
70 GHz to 125 GHz	USB	W-Band	Broadband & D-Band	
70 GHz to 150 GHz	USB	W-Band & D-Band	Broadband & D-Band	
110 GHz to 125 GHz	USB	W-Band	D-Band	
110 GHz to 150 GHz	USB	W-Band & D-Band	D-Band	
125 GHz to 150 GHz	USB	D-Band		

Whatever the basic power meter accuracy is will be transferred to the VNA within about 0.1 dB (convergence criteria is normally 0.07 dB, this limit loosens for frequencies above 70 GHz) for short periods of time. That correlation can drift as temperatures change, connection repeatability, etc. The basic accuracy of the power meter is outside the scope of this guide (consult the documentation for the meter/sensor in question) but includes terms such as linearity, mismatch, basic power factor correction, offsets, and other items.

Embedding/De-embedding

An additional function possible with power calibrations is embedding or de-embedding where an existing power calibration is modified by the S-parameters in a .s2p file to reflect a network that was added after the power calibration (embedding) or removed after calibration (de-embedding). These functions are placed on a submenu labeled NW extraction (short for network extraction, this refers to the calibration options submenu where the .s2p file is often generated). They are accessed by selecting Power | Power Cal | NW Extraction. Figure 15-7 shows the NW Extraction menu, which is enabled only if a power cal exists.

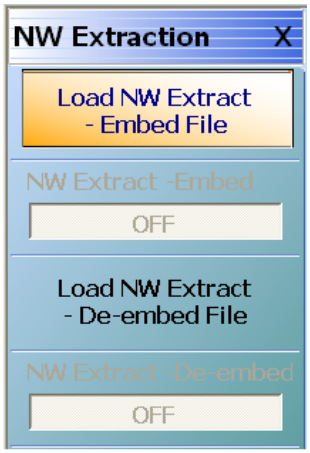


Figure 15-7. NW EXTRACTION Menu

This menu enables one to load files separately for embedding and de-embedding. Note that the files are separate for port 1 and port 2 power calibrations. Also note that only the S21 parameter is used in the adjustment of the power calibration (and, of course, only the magnitude). Interpolation and extrapolation are used if the frequency lists of the .s2p file and that of the power calibration do not match. A warning dialog will appear if extrapolation will be used.

If the network loaded is lossy and is being embedded (e.g., the power calibration is performed coaxially and then the reference plane is connected to a wafer probe and the desired reference plane is at the probe tip), the power delivered by the source will increase by the amount of the loss. Similarly, if such a lossy network is de-embedded (e.g., an adapter was used to facilitate the calibration but is then removed for the measurement), the power delivered by the source will decrease by the amount of the loss.

All of the above discussion applies to both linear and log versions of swept frequency. In log sweep, the frequency list is calculated using a constant ratio step as opposed to a constant step size in linear frequency sweep. Also, the graticule for rectilinear plots while in log sweep will change to reflect that modified frequency list. An example is shown in Figure 15-8. Integral numbers of decades of lines will always be used.

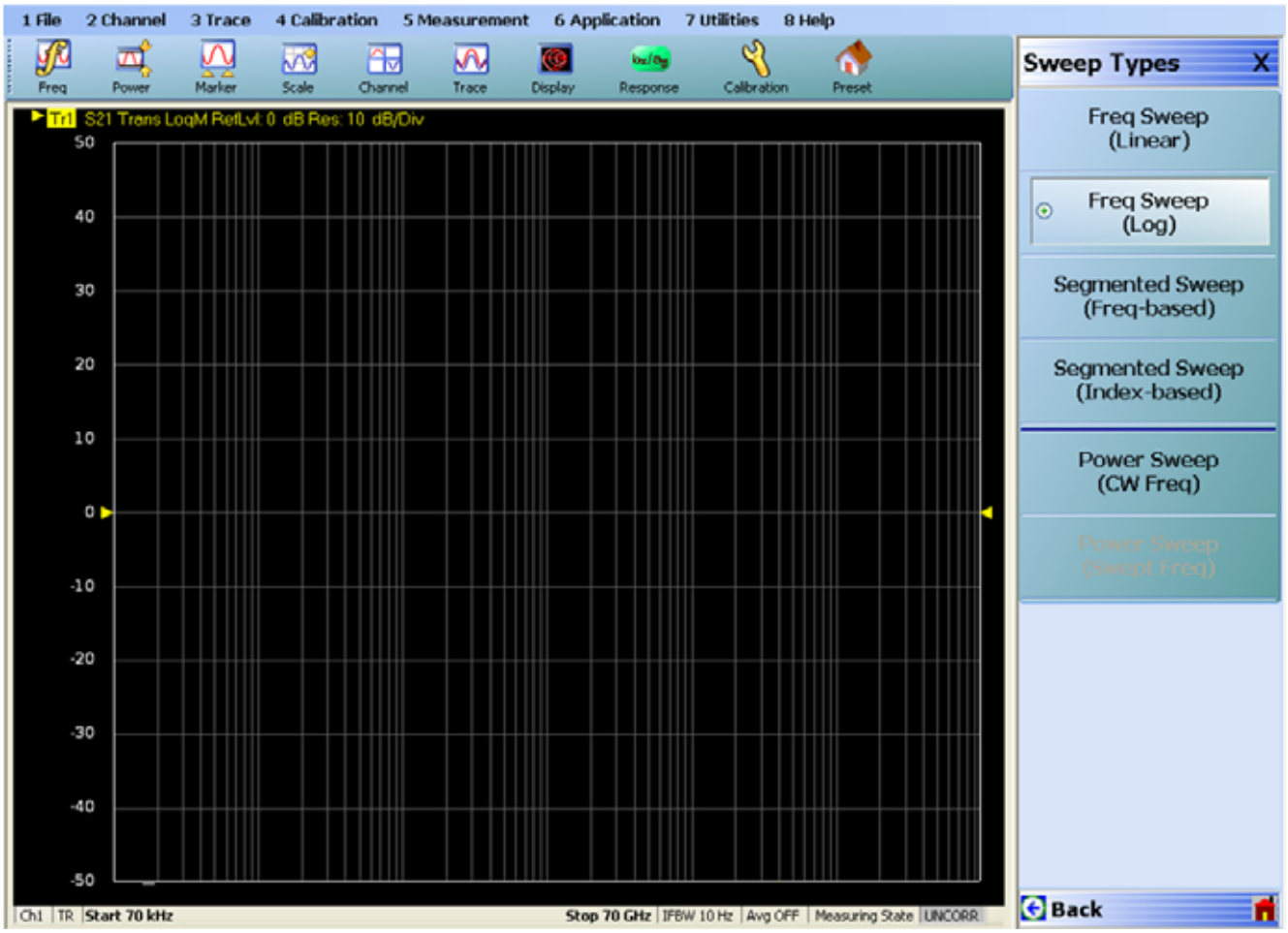
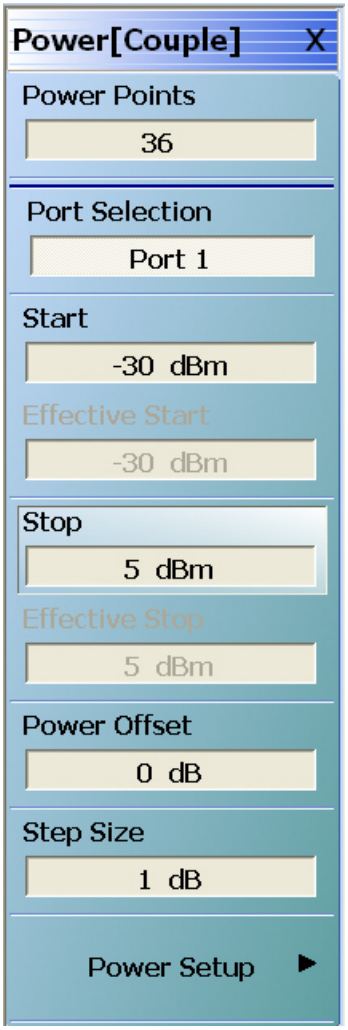


Figure 15-8. Example Graticule for the Log Sweep Type

15-4 Setting Up Power Sweeps

A power sweep is valuable for making gain compression and other power-dependent semi-linear measurements. The frequency is set at a single CW value and the power range is specified in a way analogous to frequency sweeps as suggested by Figure 15-9.



Sweep Type set to Power Sweep (CW Freq)

Figure 15-9. Main Power Sweep Menu—POWER [COUPLE] Menu

As with the regular power control in frequency sweep, the power at the two ports may be coupled or uncoupled. This feature takes on new importance in power sweep in that the two ports may drive with completely different power ramps. The number of points in these two power sweeps must, however, be the same. As with all sweep types, different attenuator settings in different channels is not permitted to avoid potential attenuator damage in fast sweeping scenarios.

Power offset is an important entry for cases when an external preamplifier, large pad, or other network may be in use between the port and the DUT plane. By entering a value here to approximate the net gain of the external networks, the effective power fields will be updated accordingly and any power calibrations performed will be more efficient. Note that source-side attenuator settings are also reflected in the effective power fields.

Some of the more detailed controls are on the power setup level of the power menu located on the POWER SETUP menu (see Figure 15-10). The coupling control and port selection (when uncoupled, duplicate of first level) is located here as is the single power selection items. This entry allows one to put the system in constant power (and constant frequency) and is often used for making DUT adjustments prior to full power sweep measurements. The receiver cal and external source power submenus are covered in other sections of this measurement guide (Receiver Calibration and Multiple Source control respectively).

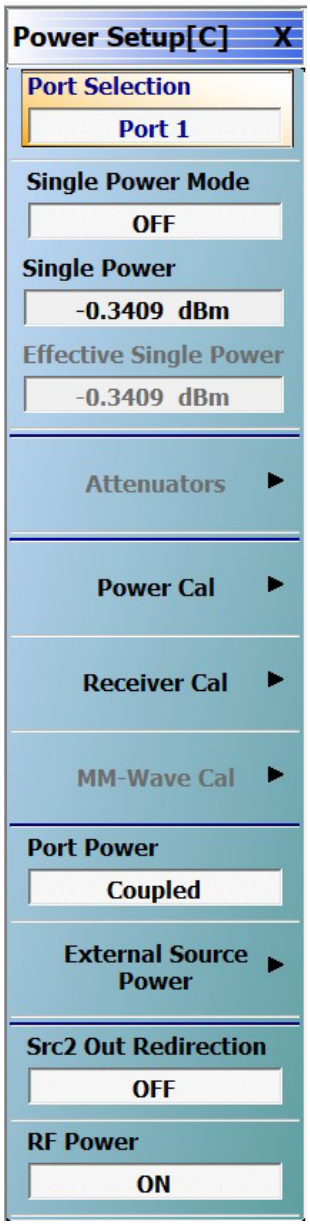


Figure 15-10. Second Level Power Sweep Setup Menu—POWER SETUP [C] Menu

The Src2 Out Redirection pertains to Option 32 (a combiner option available in dual source systems) where the source 2 signal can be switched to combine into the source 1/port 1 signal path. This switch can be activated here when using manual multiple source measurements but is overridden in the IMD measurement (see Chapter 28, "IMDView™ (Option 44)").

The power calibration while in power sweep mode allows an accuracy enhancement of the power levels delivered to the DUT. A power sensor will be connected to the port selected and then at each power point in the power sweep, the system will adjust the control loops to deliver that power to within 0.1 dB. The Power Calibration menu is shown in Figure 15-11 and the dialog executing the calibration is shown in Figure 15-12. Note that the Target power range will want to maintain the span of the power menu entries (e.g., if the main start power is –10 dBm and the main stop power is 5 dBm, the target power spread should normally be 15 dB). This comes from an assumption that the target difference would be due to a loss or gain offset rather than some heavily compressing medium.



Figure 15-11. POWER CAL Menu

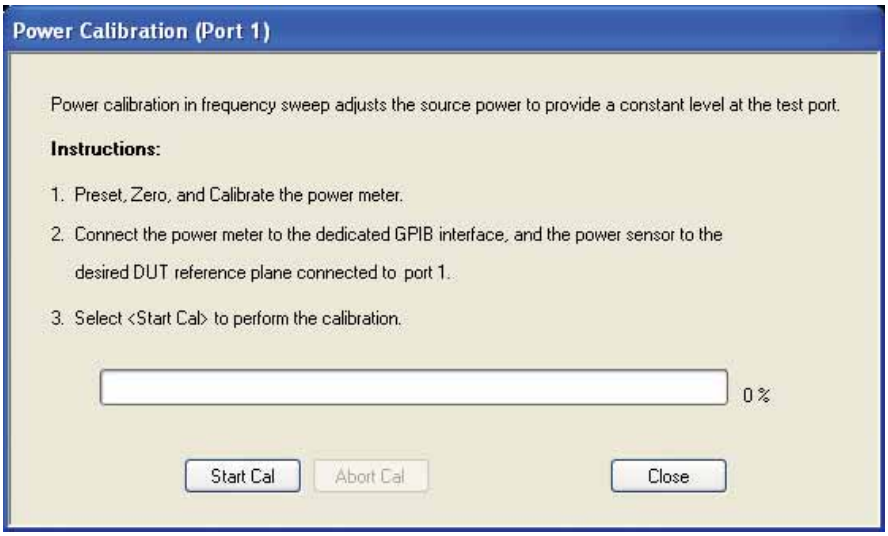


Figure 15-12. POWER CALIBRATION Dialog Box

Gain Compression

Although not a sweep type by itself, gain compression is often thought of as the main application of power sweep. As such, a trace-based utility is provided to simplify the gain compression measurement.

The objective is to provide a marker-like function for quickly finding the X-dB compression point of an amplifier or other DUT. While this is typically performed on S21, it is allowed on any ratioed parameter so that, for example, match deviations can be quantified during compression. The gain compression function is trace-based, unlike the sweep type, so it can be invoked on some traces (S21 for example) and not on others (b2/1 to monitor output power for example).

The main gain compression menu (located under Display while in power sweep) is shown in Figure 15-13. The second button toggles on and off the indicator for the active trace (the indicator is denoted by a "**C**" on screen and will be located at the desired compression point. The top button activates continuous normalization of the trace to the value of the first point in the power sweep. This allows for easier visualization of the compression process but is not necessarily needed for the measurement.

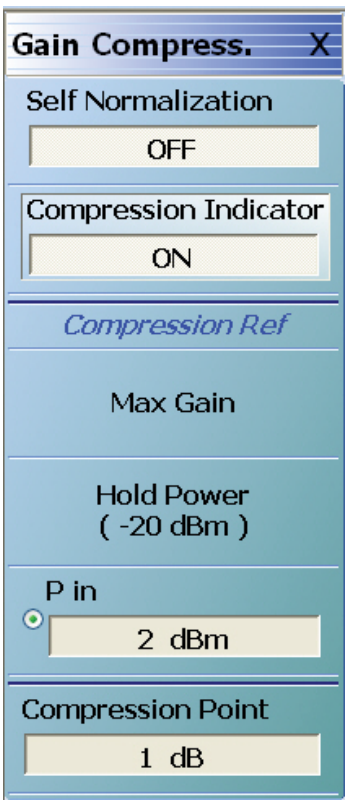


Figure 15-13. GAIN COMPRESS. (GAIN COMPRESSION) Menu—Power Sweep CW Set

Note	The Hold Power value is read-only and comes from the HOLD menu.

The middle three buttons determine how the compression point is to be measured. That is, the X-dB compression marks a decay of X dB from what value? Commonly, this is done from a low power level which is often the hold-state power (see the hold functions menu under Sweep Setup). The reference can also be set to the maximum output power point and this may be useful in cases of gain expansion. Finally, the reference may be set to the point corresponding to an arbitrary, entered input power value.

The compression point can be entered separately and common values are 1 dB and 0.1 dB. Note that this value is allowed to be negative so certain levels of gain expansion can be sensed.

An example measurement showing these features is illustrated in Figure 15-14. The input power corresponding to the compression point is readout in the upper right corner of the screen. The compression indicator "**C**" picks out the 1 dB compression point at an input power of ~3.5 dBm. Note that if Max Gain is selected as the gain reference, the location of the 'R' reference marker will move from sweep to sweep as a new maximum point is identified each time. If the data is noisy (e.g., very wide IFBW, very low input power or output power, etc.), the gain reference can change significantly from sweep to sweep and hence the compression point determined may be more variable.

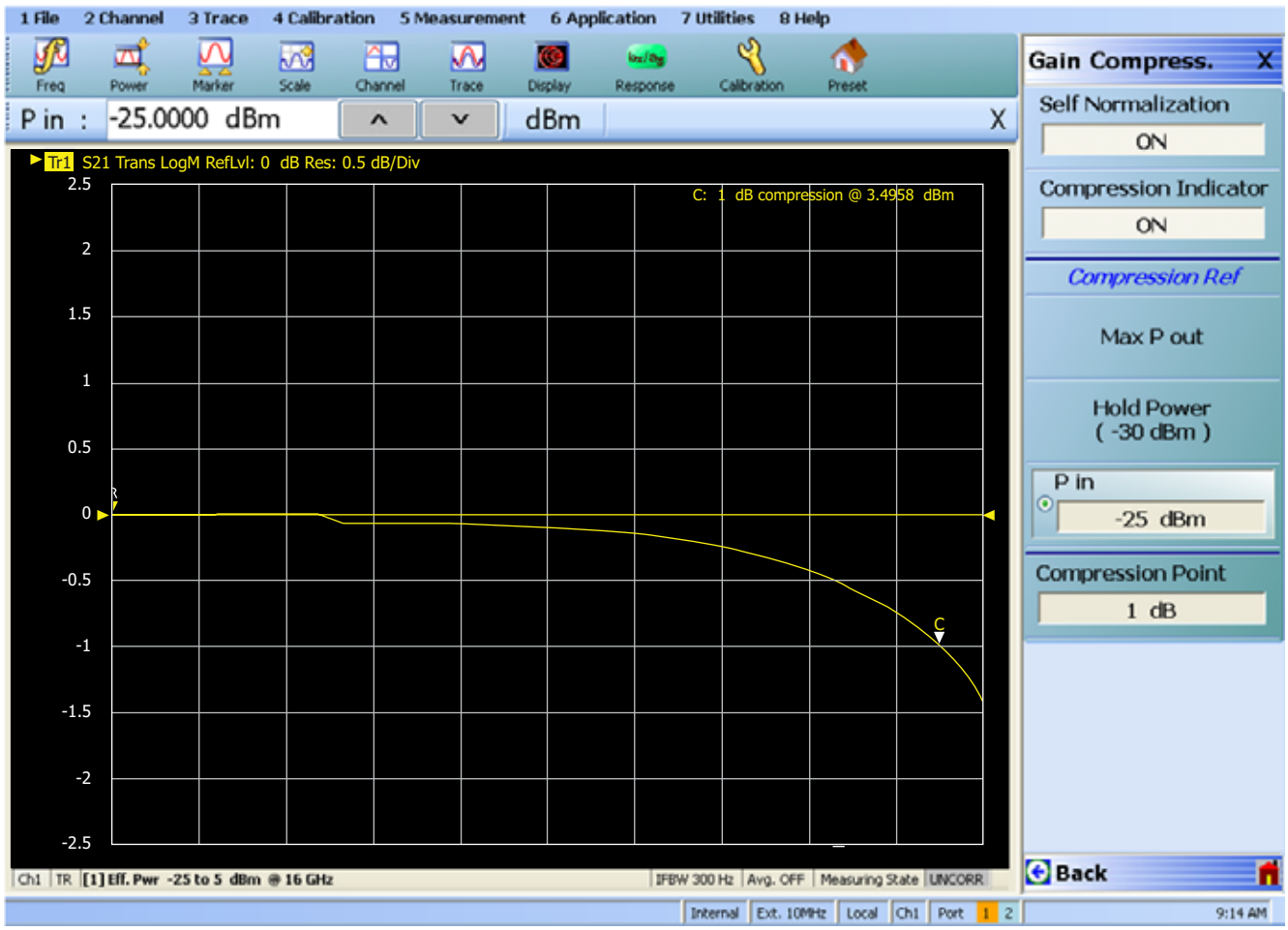


Figure 15-14. Gain Compression Example

The steps that might be taken in executing the above measurement:

1. Set the CW frequency (16 GHz in this case) and the power sweep parameters (–25 to +5 dBm with 61 power points), A higher density of power points will help with accuracy on the compression point (so the system has less interpolation to do) at the expense of longer measurement time and longer power calibration time (if performed). Select the IFBW (300 Hz in this case) for the desired trace noise level.
2. Perform a power calibration if desired. The factory ALC cal accuracy may be adequate in some applications (generally better than 1 dB at microwave frequencies) but the user power calibrations can improve on this. See earlier comments in this chapter on user power calibrations for more information. Note that with thermal sensors, the calibration time per power point could be 5–10 seconds.
3. Perform an RF calibration if desired. This was not done in this example since the input compression point only was of interest and not the absolute gain. A one-direction transmission frequency response calibration may be all that is needed if absolute gain data is required (does not include match correction but is often adequate).

4. Set up the gain compression parameters. In this case, self-normalization was used to center the result near 0 dB (bases on the 1st point and renormalizes every sweep so it may appear to be jumping if the data is noisy). The 1 dB compression point is of interest (that is the default value) and the compression point was to be referenced to the gain at –25 dBm input so 'Pin' was selected as the reference.
5. Turn on the compression indicator and observe the result.

15-5 Frequency-Based Segmented Sweep

In many applications, having a simple list of frequencies where the step size between points is uniform is not adequate. The DUT specifications may have specifications in certain bands and certain specific frequencies that must be tested, there may be certain communications bands that must be tested, or there may be certain spot frequencies that are of interest for troubleshooting or analysis.

For these cases and others, segmented sweep allows one to put together a very arbitrary list of frequencies to sweep as well as having some control of instrument behavior that is distinct at these different points and/or segments. The entire sweep is broken into segments (a segment may contain one or many points) and in each segment, one can independently control:

- IF bandwidth
- Averaging
- Port 1 Power and Port 2 Power (although the step attenuators, if installed, must be in the same position for all segments to prevent premature wear)

There is a distinction made between frequency-based and index-based segmented sweep that should be explained:

- **Frequency-Based**

  Frequency is always monotonically increasing (within a segment and between segments). Plotting may be based on the frequency or the index of the particular point (more on this later).

- **Index-Based**

  The segments do not have to be in any order with regards to frequency. Plotting is always based on the point index.

Frequency-based is most commonly used and will be discussed in this section. Index-based sweeps, which is used when reverse sweeps and particular frequency order is important, is covered in the next section.

The main menu and an example entry table are shown in Figure 15-15 and Figure 15-16. The main purpose of this menu is to aid entering data into the table and to help save and recall that data. Note that segmented sweep tables can be saved/recalled separately from this menu or they can be saved/recalled as part of the global setup using the entries under the File menu.

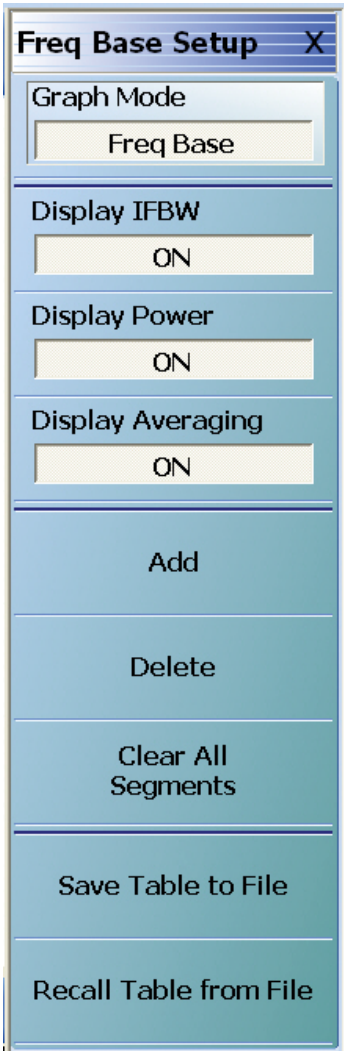
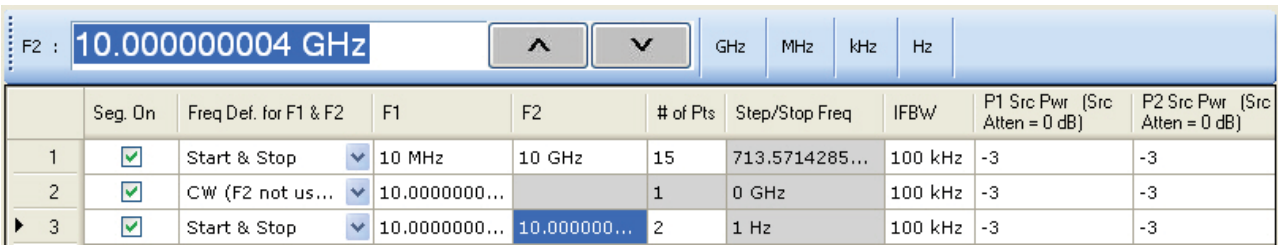


Figure 15-15. FREQ BASE SETUP (FREQUENCY-BASED SETUP) Menu



	Seg. On	Freq Def. for F1 & F2	F1	F2	# of Pts	Step/Stop Freq	IFBW	P1 Src Pwr (Src Atten = 0 dB)	P2 Src Pwr (Src Atten = 0 dB)
1	☑	Start & Stop	10 MHz	10 GHz	15	713.5714285...	100 kHz	-3	-3
2	☑	CW (F2 not us...	10.0000000...		1	0 GHz	100 kHz	-3	-3
▶ 3	☑	Start & Stop	10.0000000...	10.000000...	2	1 Hz	100 kHz	-3	-3

Figure 15-16. Tableau Entry Table for Frequency-based Segmented Sweep

The table will start with one segment and the start, stop and number of points must be defined. The pull-down item in Column 3 allows an alternatively specified start and step or a CW frequency. The step or stop frequency (which depends on the pull-down selection) will appear as a read-only field in Column 7. The IFBW, power and averaging columns can be enabled on the setup menu and entered separately by segment. The current source attenuator setting will appear in the column header and may not be changed by segment (will read 0 dB if the attenuators are not installed). If the display of these fields is not enabled, the values for those variables set in regular frequency sweep mode will prevail for all segments.

The Add, Delete, and Clear All functions are obvious. The delete function applies to the current row as indicated by the caret in column 1.

As with the multiple source tables, there are two ways to enter numbers

- Click on the cell and the text entry box above the table becomes active.
- Click twice on the cell and type directly into the cell. Frequency units must be entered and must have a space between the number and the units.

If an invalid number is entered in any field, the system will change the value to the nearest valid entry.

The one remaining item on the setup menu for frequency-based segmented sweep is graph mode, which controls how the x-axis is setup for all plotting activities but does not affect the sweep itself. In Frequency-based graph mode, the x-axis will be in frequency and all segments will be plotted where those frequencies lie. While correct, this can lead to an odd-looking display if the segments are disjointed as shown in Figure 15-17.

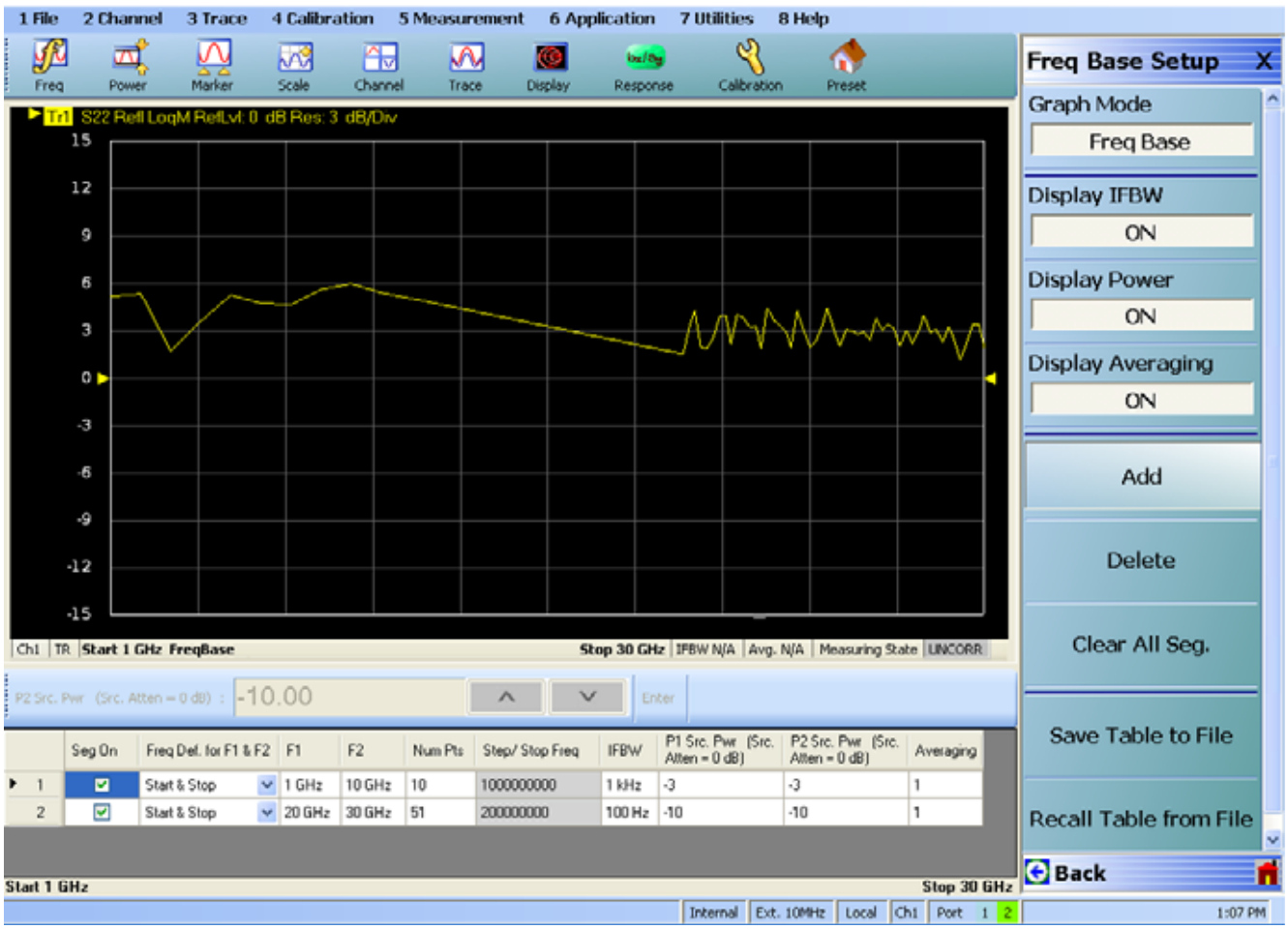


Figure 15-17. Frequency-Based Graph Mode

Since Segment 1 covers 1–10 GHz and Segment 2 covers 20–30 GHz, there is a gap in frequency where no measurements are made. For the purposes of plotting in this graph mode, the two areas are connected by a single line segment. Note that the point spacing in the plot precisely matches the frequency spacing.

When all of the data points plotted without regard to proportional frequency separation are required. For these occasions, the Index-based graph mode is available and an example is in Figure 15-18 for the same setup as Figure 15-17. Here, the x-axis is point index so all plotted points are equally spaced in the x-direction and the frequency based segmented sweep is disjointed.

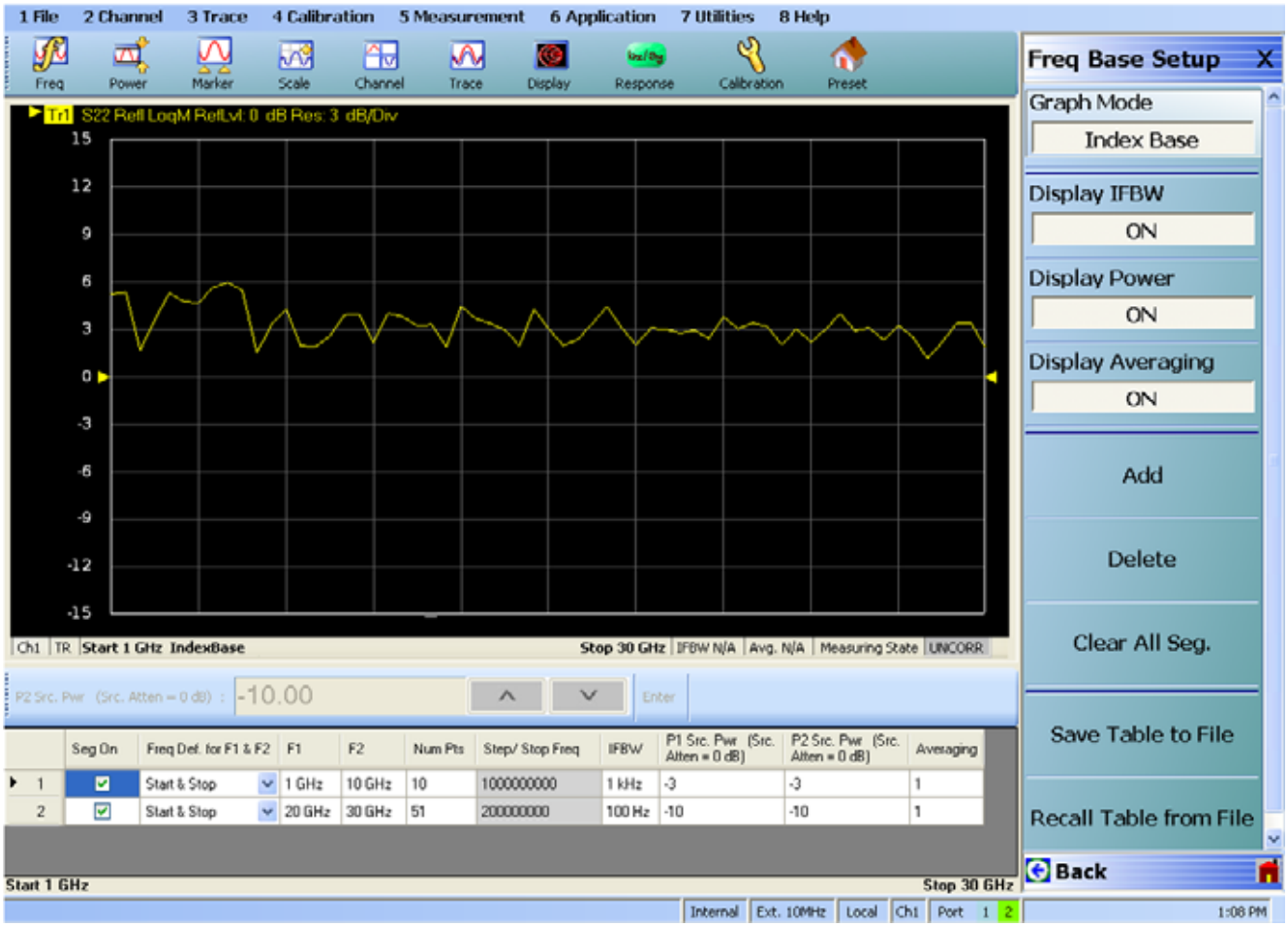


Figure 15-18. Index-Base Graph Mode

It is important to keep separate the concepts of frequency- vs. index-based for the graph mode (which only controls how things are plotted) and frequency- vs. index-based segmented sweep type (which determines how the points are swept by the instrument hardware).

15-6 Index-Based Segmented Sweep

In index-based segmented sweep the frequency segments may be in any order. This may be useful for particular test patterns where reverse sweeps are needed or particular frequencies must be measured before others due to DUT hysteresis. The setup menu and an example table are in Figure 15-19 and Figure 15-20.

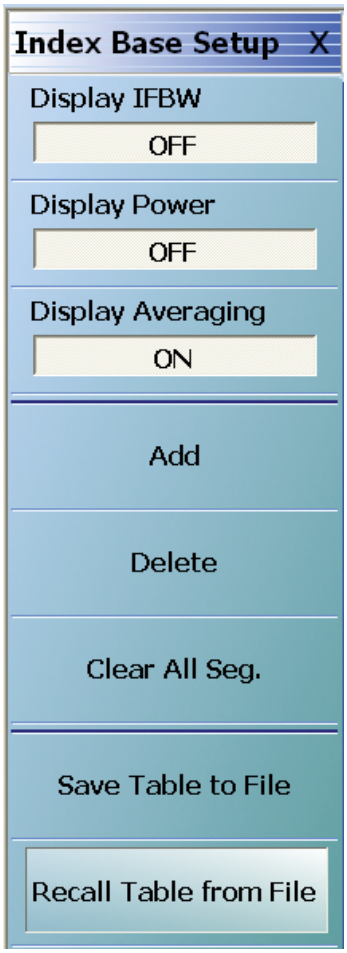


Figure 15-19. Main Menu for Index-based Segmented Sweep—INDEX BASE SETUP Menu

All plotting in this sweep type is based on the point index, which is listed in the last column of the table below.

F1 : 10.000000 MHz GHz MHz kHz Hz

	Seg. On	Freq Def. for F1 & F2	F1	F2	# of Pts	Step/Stop Freq	IFBW	P1 Src Pwr (Src Atten = 0 dB)	P2 Src Pwr (Src Atten = 0 dB)
1	☑	Start & Stop	10 MHz	10 GHz	15	713.5714285...	100 kHz	-3	-3
2	☑	Start & Stop	25 GHz	26 GHz	2	1 GHz	100 kHz	-3	-3
3	☑	Start & Stop	10 MHz	15 MHz	2	5 MHz	100 kHz	-3	-3

Figure 15-20. Tableau Entry Table for Index-based Segmented Sweep

Aside from the increased flexibility in ordering, there are few other differences relative to frequency-based segmented sweep. One exception is that there is now no choice of graph mode; it will always be index-based. In this case, it is to avoid confusing and unreadable displays where one could have many reverse tracing operations.

It should be noted that there is a special form of segmented sweep available over the remote interface only (see the *VectorStar Programming Manual*—10410-00322 for more details). In this mode, a discrete list of frequencies (up to 25000 in the usual VNA operating mode) can be sent, in any order, and without segment definition. This gives considerably more flexibility in defining the frequency sweep without having to enter lengthy data through a graphical user interface. When this entry method is used, a single segment will appear in the segmented sweep table with a notation that it is a special list. The full list being used is available over the remote interface or with any of the text file save processes. This method sometimes is labeled 'discrete fill'.

15-7 Power Sweep (Swept Frequency)

This specialty sweep type is a hybrid of the power sweep and linear frequency sweep types discussed earlier. The application is used to evaluate compression points across multiple frequencies without having to setup power sweeps individually at each frequency. The nomenclature and the meaning of the displays require some interpretation different from other measurements. In some contexts, this sweep mode is called Multiple Frequency Gain Compression (MFGC).

When in this sweep type, the gain compression menu will be available as it is with CW power sweep, but compression determination will always be on and there is no self-normalization. Most of the items are the same as in the single frequency gain compression (CW power sweep), except for the first menu item, as shown in Figure 15-21. As with CW power sweep, these selection are all per-trace. This may require some attention as parameters within the channel may be being evaluated at wildly different compression points depending on the setup.

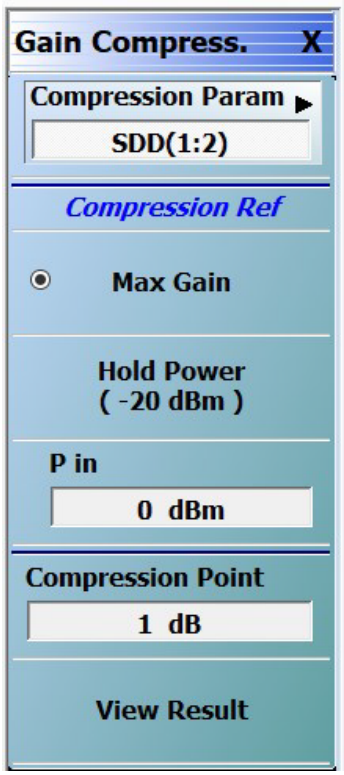
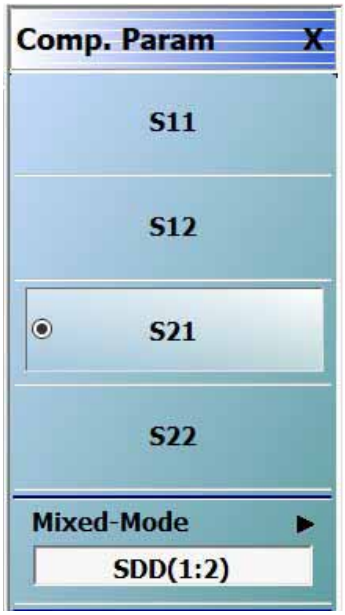


Figure 15-21. GAIN COMPRESSION Setup Menu for Power Sweep (Swept Frequency)

The compression parameter (Figure 15-22), which must be a ratioed S-parameter, is the variable that is used to find the compression point (S21 is usually used for this role).



2-Port VNA



4-Port VNA

Figure 15-22. COMP. PARAM (COMPRESSION PARAMETER) Menu

Note that the compression will be based on an S-parameter and cannot be user-defined. Mixed-mode parameters can be selected as the compression variable and this is useful, for example, with differential amplifiers. If an incomplete calibration is in place, the system will make the required sweeps to compute the selected parameter but, of course, any displayed parameters may have reduced/destroyed accuracy if the calibration does not match the port configuration.

The compression parameter is a per-trace selection; that is, every displayed response parameter can be referenced to compression of a different S-parameter. Commonly S21 will be used for the compression parameter in all cases (on two port systems) but the flexibility is available.

The desired plot variables, which are set up in the usual way with trace and response menus, are then evaluated at the power producing the indicated compression level in the compression parameter. The "view result" button will display (vs. frequency) the active trace parameter at which the desired compression occurred. This display will be in tabular form in a separate dialog (Figure 15-23).

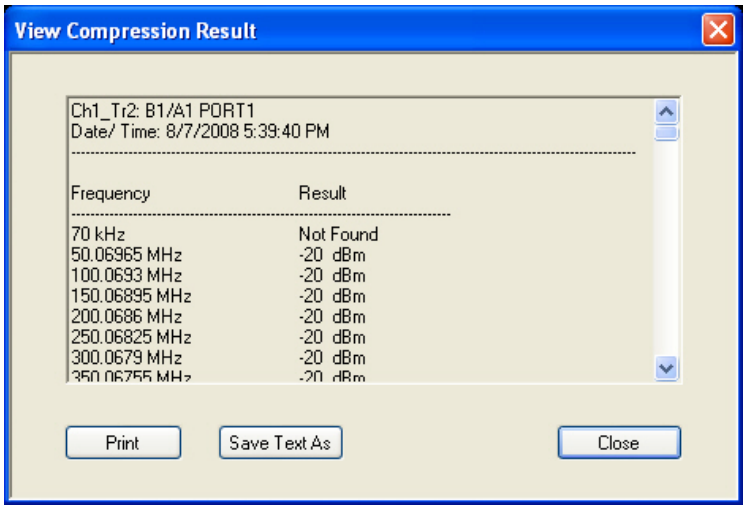


Figure 15-23. VIEW COMPRESSION RESULT Dialog Box

The dialog shows the value of the active trace parameter at the desired compression point. If the active trace has a two graph display (e.g., log magnitude and phase), there will be a third column added.

Power-Sweep Swept Frequency Example

To help understand these concepts, consider an example. We have an amplifier connected in the usual forward direction (S_{21} > 0 dB) and we wish to know the output power and the input match when at the 1 dB gain compression point at a few frequencies. Assume that sufficient attenuation is employed to prevent VNA compression and that all needed calibrations have been performed.

- At 1 GHz:
  - S_{21}
    - = 10 dB at –10 dBm input (low power) and
    - = 9 dB at +1 dBm input
  - $b_2/1$ (output power)
    - = 0 dBm for –10 dBm input and
    - = 10 dBm at +1 dBm input
  - S_{11}
    - = –15 dB at –10 dBm input and
    - = –17 dB at +1 dBm input
- At 2 GHz:
  - S_{21}
    - = 12 dB at –10 dBm input (low power) and
    - = 11 dB at +3 dBm input
  - $b_2/1$ (output power)
    - = 2 dBm for –10 dBm input and
    - = 14 dBm at +3 dBm input

    - S_{11}
        - = –17 dB at –10 dBm input and
        - = –18 dB at +3 dBm input
- At 3 GHz:
    - S_{21}
        - = 7 dB at –10 dBm input (low power) and
        - = 6 dB at –1 dBm input
    - $b_2/1$ (output power)
        - = –3 dBm for –10 dBm input and
        - = 5 dBm at –1 dBm input
    - S_{11}
        - = –10 dB at –10 dBm input and
        - = –8 dB at –1 dBm input

The compression points based on the compression parameter of S_{21} (input power referred) are as follows:

- 1 GHz: +1 dBm
- 2 GHz: +3 dBm
- 3 GHz: –1 dBm

Since output power and input match at 1 dB compression are of interest, one would select $b_2/1$ and S_{11} as the display parameters. In this sweep type, those parameters will be evaluated at the above power levels established by the compression measurement. A plot of $|b_2/1|$ would then show:

- 1 GHz: 10 dBm
- 2 GHz: 14 dBm
- 3 GHz: 5 dBm
  (assuming a power out graph type is being used and a receiver cal is in place)

A plot of |S11| would show:

- 1 GHz: –17 dB
- 2 GHz: –18 dB
- 3 GHz: –8 dB
  (assuming a log magnitude graph type)

Any of the normal S-parameters or user-defined parameters can be selected as the display variables (just as in any other sweep type). Those most commonly used are the S-parameters (to represent amplifier quasi-small-signal behavior at the point of compression), an un-ratioed parameter to represent output power at compression (often $|b_2/1|$), and an un-ratioed parameter to represent input power at compression (often $|a_1/1|$). Again, a receiver calibration is required for the representation of absolute power (input or output) and this is discussed in detail in another chapter of this measurement guide.

The plot resulting from an example measurement is shown in Figure 15-24. Here the frequency sweep range was 1–2 GHz while the power sweep range (at each frequency point) was –10 to +5 dBm. The display parameter is output power ($|b_2/1|$) while the compression parameter is S21.



Figure 15-24. Example Plot from Power Sweep with Swept Frequency

The detailed series of steps that might be used to arrive at the above measurement could follow this script:

1. Select the frequency and power ranges and number of points for each range. (1 GHz to 2 GHz with 51 points, –10 to 5 dBm with 16 points). The power sweep range should, of course, include the anticipated compression point. It may be tempting to use more points, but power calibration time and measurement time will vary as (frequency points)*(power points). As stated earlier, the maximum number of frequency points is 401. Select an appropriate IFBW based on trace noise needs (1 kHz in this case).
2. Perform a power calibration if desired. The factory ALC calibration accuracy may be adequate (<1 dB generally at microwave frequencies) but if not, or if there is a lossy network between the test port and the DUT plane, a user power calibration may be useful. See earlier comments in this chapter on power calibrations. Note that with thermal power sensors or low power levels, the power calibration can take several seconds per point and there are (frequency points)*(power points) points to consider. Interpolation and extrapolation will be used if the ranges are changed after calibration and generally these are smoothly varying functions. Note that a MFGC power calibration can be used in other sweep modes.
3. Perform a receiver calibration if needed. This is important for the current example as we wish to measure output power at 1 dB compression (via $b_2/1$). A thru connection is all that is needed and the calibration is automatically performed at the 'single power' value (not across power; the receiver linearity is relied upon for other levels and is better than 0.05 dB except near receiver compression or the noise floor). Interpolation and extrapolation will be used if the frequency range is changed after cal.

4. Perform an S-parameter calibration if needed. This may be important if gain or mismatch at compression is needed and any of the calibration choices are available. If only gain is needed, a one-direction transmission frequency response calibration may be adequate.
5. Set up the gain compression configuration for each trace. In this case the compression parameter is S21 and the reference is max gain. Note that this reference definition is not universally used and can result in surprising values if the DUT has significant gain expansion or other mid-range distortions.
6. Observe the results and view tabular values with the View Result button if desired.

Some points of note:

- Frequency range is setup using the normal frequency menu. The only constraint is that a maximum of 401 frequency points can be used.
- The power sweep range is setup as with CW power sweep. The usual limits on power points apply.
- If one wants to look at the actual gain compression curve ($|S_{21}|$ vs. power for example), set up a 2nd channel, since power sweep (CW) and power sweep (swept) cannot be combined in the same channel.
- S-parameter calibrations are basically handled as in linear frequency sweep. If performed while in this sweep type, they will be conducted at the start power. The same error coefficients will be applied at all power levels. If the calibration was performed in linear frequency sweep, the point count will be coerced in this sweep type to get under the 401 point limit (interpolation will be used if it is active).
- Receiver calibrations are normally required for accurate plots of output power and can be performed in this sweep type or in other sweep types as long as the appropriate frequency range is covered.
- As discussed previously, receiver calibrations fully use interpolation and extrapolation. Distortions can occur at bandswitch frequencies (2.5, 5, 10, 20, 38, and 40 GHz for base VNAs plus 30, 54, 80, 110, 120, 130, 153 and 170 GHz for some ME7838X systems) if interpolation is used and point density is not high. At very low frequencies (<30 MHz), additional transition frequencies occur. When in doubt, one can redo the receiver calibration over the desired frequency list.
- When in this mode, receiver calibrations are performed at the 'single power' power level. Normally, this should be relatively high so as to minimize trace noise but can be any value the system can achieve.
- Power calibrations are also interpolated and extrapolated. Similar bandswitch frequencies play a role but the effects are usually less dramatic than for receiver calibrations (except at mm-wave frequencies >100 GHz).
- Mixer MFGC can be performed using the same steps if a reference mixer is employed (see Chapter 21 — Mixer Setup and Measurement of this guide for more information; normally such a mixer is placed in the reference loop of the VNA and shares the LO with the DUT). Generally, the reference mixer input is well-padded or is otherwise more linear than the DUT so allow for more unambiguous measurements.

Chapter 16 — Reciprocal Measurements

16-1 Chapter Overview

This chapter describes the reciprocal concept (SOLR or unknown thru approach) measurements.

Related Chapters

The following chapters contain content related to reciprocal measurements:

- Chapter 3, “SOLT/SOLR Calibration”
- Chapter 4, “Offset Short (SSLT) Calibration”
- Chapter 5, “Triple Offset Short (SSST) and Multiple Offset Short (mSSS) Calibrations”
- Chapter 17, “Broadband/mmWave Measurements (Option 7, Option 8x)”

16-2 Introduction

The previously discussed SOLT and offset short calibration techniques all require a known “thru (or through)” as part of the full 2-port (or 1p–2p) calibration. The “thru” is really a defined transmission line having known length, known loss, and assumed perfect match (under most conditions). There are certain cases when this is not possible when there is an undefined or unknown thru

- Coaxial cal when the two ports are different connector types
- On-wafer when the “thru” is a meandering transmission line of imperfect match
- A calibration that must take place through a test set (coax or waveguide) with unknown (and highly frequency dependent) loss and match

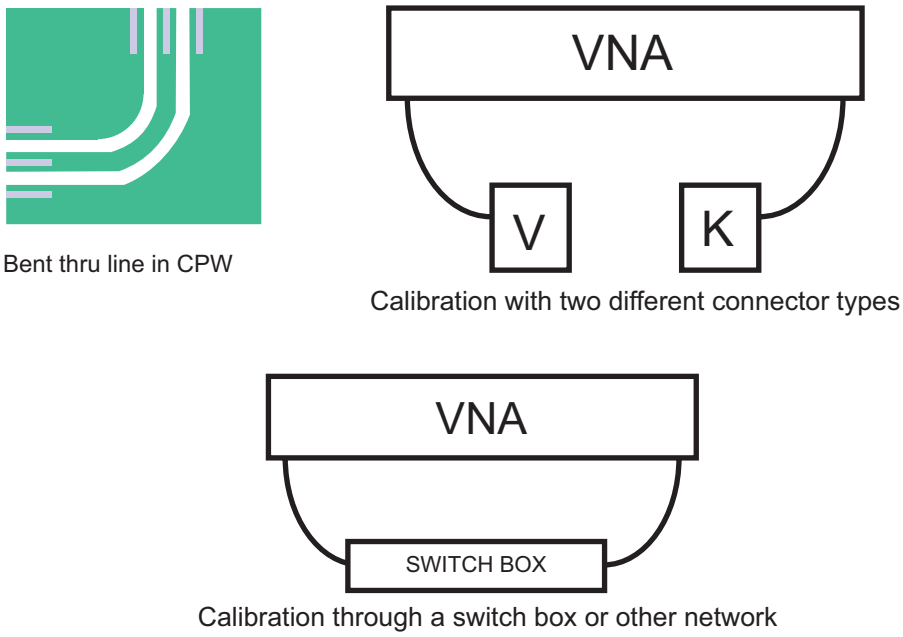


Figure 16-1. Illustration of SOLR or “Unknown Thru” Calibration

For these cases, and others when the “thru” cannot be very well-known, there is the *reciprocal* option (also known as the unknown thru). In this case, the same reflect standards are used, but no assumption is made about the “thru” except that it be reciprocal (i.e., S21=S12; no assumption made about S11 and S22). In practice, there are some limits to this.

The technique borrows from the LRL family and uses some of the redundancy available with the fully-defined families to reduce knowledge needed about something (the thru in this case). The resulting cal will generally not be quite as accurate as the regular thru version if the thru met the conditions described above. It will, however, be better than using the regular thru version when the thru has unknown loss or match.

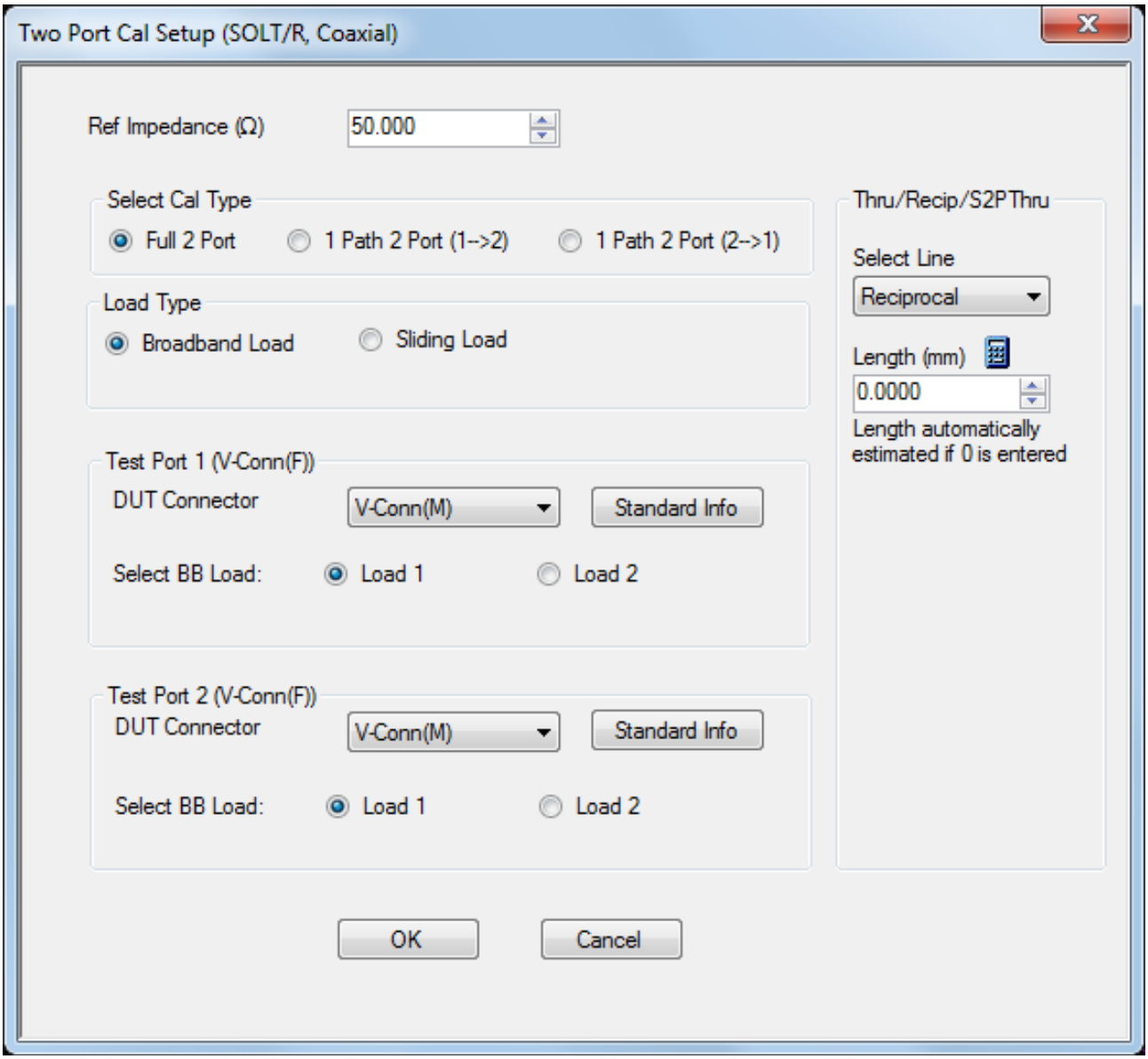
16-3 Line Length Estimate

A line length estimate (electrical delay or free space equivalent length) can be used to help with root choice, but this is not a critical parameter. Typically, one needs only to be within a half-wavelength of the correct length at the maximum desired calibration frequency. If one enters 0 for the length, the software will automatically estimate the length based on fitting-to-the-phase function at the lower part of the frequency range. Details are discussed in Chapter 9 related to automatic length estimation for adapter removal procedures and the same principles apply here. The only danger in auto length estimation is if the frequency step size is large relative to the electrical size of the setup. As discussed in Chapter 9, one can look at the phase of an uncalibrated transmission parameter in the setup to see how fast the phase is changing.

If the match of the reciprocal network is worse than –8 dB or the loss exceeds ~20 dB, the reciprocal treatment will start to degrade, but a calibration will still be possible. Since such a network is at the limits of de-embedding capability, there are few choices except to consider 1p–2p processing with scalar de-embedding.

16-4 SOLR Calibration

For SOLR, the “select line” field would be chosen as “reciprocal” instead of “through” and the length field would be the estimate of length for root choice that was discussed above (see Figure 16-2). Note that the same variants are possible for offset short and triple offset short calibrations.



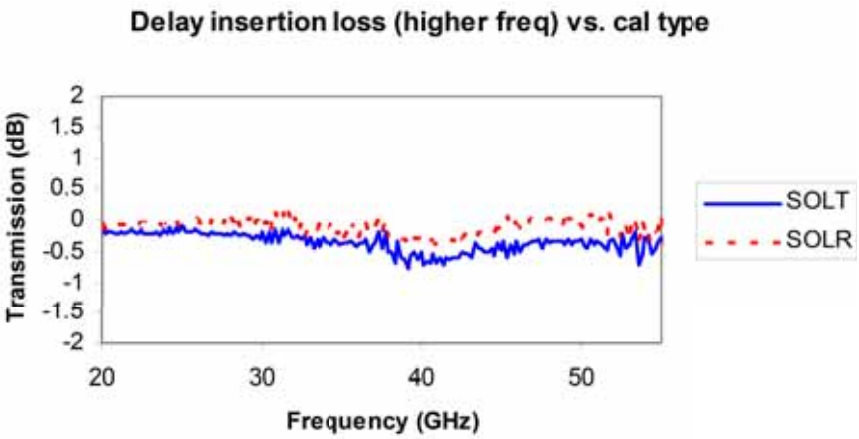
Typical calibration setup dialog for SOLR (Short-Open-Load-Reflection)

Figure 16-2. TWO PORT CAL SETUP (SOLT/R, COAX) Dialog Box

16-5 Uncertainty and Sensitivity Considerations

With reciprocal methods, the usual determining question is *'how bad are the characteristics of a thru?'* in the environment in question. As suggested earlier, if the thru is good (RL~>25 dB and insertion loss known to within ~0.05 dB), then for most purposes, the defined thru methods will do better since there is a more explicit determination of load match. As the reciprocal device worsens in terms of insertion loss and match, the sensitivity of the calibration to small standards problems will increase because less and less information is available to the calibration. In this sense, sensitivity may be a more practical metric than best-obtainable uncertainty, which will be fairly similar in all cases.

When the available thru is poor, however, then the absolute defined thru uncertainty will be high. In one example, delay insertion loss deviation was plotted for two calibrations (SOLT and SOLR) where the only interconnect available had 8 to 15 dB return loss across the band. The SOLT measurement shows almost 0.7 dB deviation (on a low loss device) while the SOLR deviation was kept to < 0.2 dB as shown in Figure 16-3.

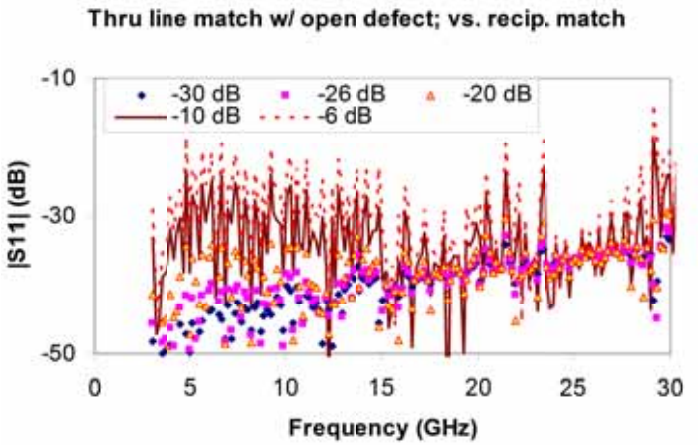


An example comparison (insertion loss deviation) is shown here between SOLT and SOLR when no good thru connection was available.

Figure 16-3. Example Comparison (Insertion Loss Deviation) Between SOLT and SOLR

There are limits on the reciprocal network, as discussed above, and those will become most apparent when measuring reflection of well-matched, low insertion loss devices. The one mitigating factor is that if the media is complicated enough to require a reciprocal technique, there will be fewer well-matched, low insertion loss devices to be concerned about. The issue becomes one where the resolution on load match during calibration becomes limited by the reciprocal itself, so the correction for that term will become more sensitive.

As an example, consider the match measurement of a delay line where a problem was introduced on the open standard during the calibration. A series of different calibrations were done with different reciprocal devices (with different match levels). The resultant measurements are shown in Figure 16-4. For reciprocal elements with reasonable match (~< –20 dB), the measurements did not become distorted significantly by the open calibration error. At the –10 dB reciprocal match level, measurements below ~–20 dB were impacted. At the –6 dB reciprocal match level, even higher reflection measurements would have been distorted. A somewhat obvious guiding principle is that the reciprocal used for the calibration should not be worse (in terms of return loss or insertion loss) than the devices that will be measured with that resulting calibration.



A plot of sensitivity of return loss measurements to a problem with one of the reflection standards is shown here for a variety of reciprocal devices used during the calibration.

Figure 16-4. Plot of Sensitivity of Return Loss Measurements

Somewhat independent of the nature of the reciprocal element, the uncertainty behavior of the SOLR calibration is arguably more nuanced than for SOLT. Part of the reason for this is that the high level reflect measurements are leveraged to a great degree to solve for the transmission tracking terms. This basic transmission uncertainty is dependent to a higher degree on residual source match (essentially the characterization accuracy of the large reflect standards) than with SOLT. As a result, SOLR can benefit from the use of .s1p-characterized reflection components even more so than can SOLT. More details of the underlying behaviors can be found in the literature (e.g., U. Stumper, "Influence of Nonideal Calibration Items on S-Parameter Uncertainties Applying the SOLR Calibration Method," *IEEE Trans. Instr. and Meas.*, vol. 58, Apr. 2009, pp. 1158–1163).

Chapter 17 — Broadband/mmWave Measurements (Option 7, Option 8x)

17-1 Overview

This chapter describes the hardware control and user options that are available for the VectorStar MS464xB Series VNAs employing Option 7 (and for some features, Option 8x), allowing work above the basic 70 GHz limit of the VectorStar family. Calibration, measurement, and troubleshooting approaches and tips are covered.

17-2 Introduction to Broadband and Millimeter Wave

Broadband and millimeter wave measurements using external modules are important applications that allow work above the 70 GHz basic limit of the MS464xB Series VNA family. Part of the APPLICATIONS menu is devoted to making these measurements simple and straightforward to do.

Some key points:

- **3738 Broadband/Millimeter-Wave (BB/mmWave) Mode**

  In classical 3738-based broadband/mmWave mode, two external Anritsu synthesizers are used to form the base RF and LO to drive the external modules. These synthesizers are controlled to provide the correct frequency and powers to these modules. The application also controls an external test set (3738X) that handles transfer switching and IF processing. This mode applies to OEM mmWave modules from OML, Inc. and Virginia Diodes, Inc. along with any others that meet the frequency plan and signal level assumptions (see appendix)

- **3739 Broadband/Millimeter-Wave (BB/mmWave) Mode**

  In modular broadband/mmWave mode, external synthesizers are not used. The MS4647B VNA in conjunction with the 3739x Test Set handles the RF and LO requirements to drive the modules. This mode applies to the Anritsu 374xx and MA25x00A modular mmWave components, as well as to OEM mmWave modules from OML, Inc. and Virginia Diodes, Inc. along with any others that meet the frequency plan and signal level assumptions. Option 8x is required to use this mode.

- **Continuous Sweep**

  In broadband mode, one continuous sweep is possible from 70 kHz to 110 GHz (OEM versions and 3743E/EX modules), 125 GHz (with the 3743x modules) or 145 GHz (with the MA25300A modules, operational to 150 GHz) or 220 GHz (with the MA25400A modules, operational to 226 GHz). In mmWave mode, just the external modules are used.

- **RF Calibrations**

  All RF calibrations are supported in both of these modes although certain algorithms may be more optimal for the frequencies involved. Cal merge is particularly useful in broadband mode where one will often want to mix algorithms

The main APPLICATION menu is shown here with the broadband and mmWave selections highlighted.

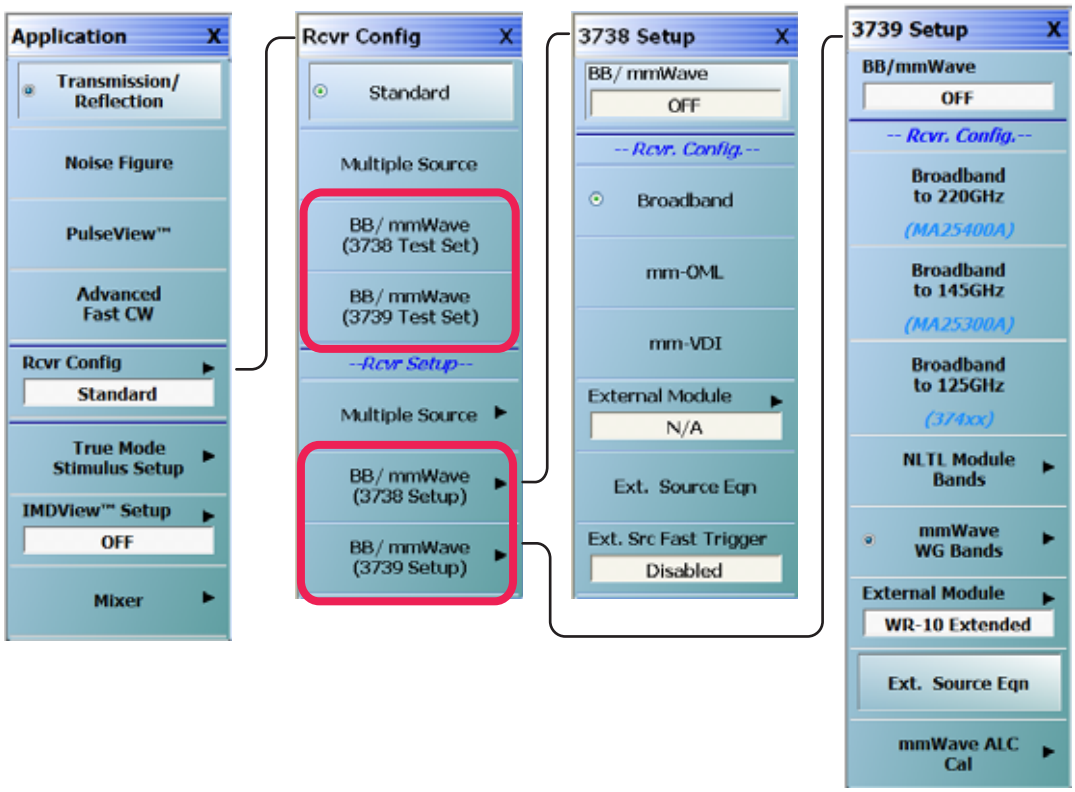
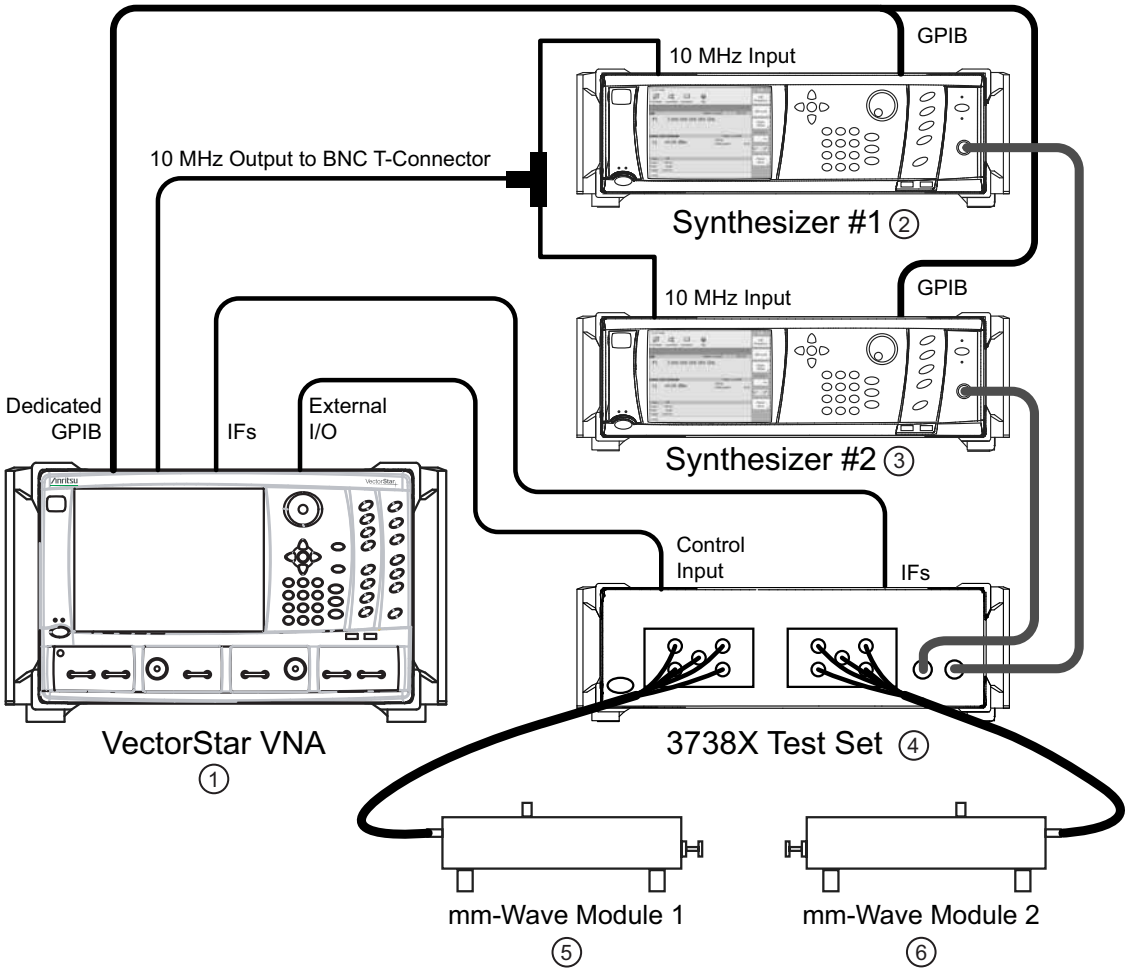


Figure 17-1. APPLICATION Menu—Multiple Source Selections

17-3 Physical Setup for 3738-Based Systems

Broadband and mmWave modes are remote-control intensive, so much of the setup work revolves around setting up communication to the external synthesizers and the external test set (3738X). Of course, some care must be taken with the IF, RF and LO connections but the issues here are typically the usual ones of connector and cable care. The control and reference connections are illustrated in Figure 17-2.



1. VectorStar Series VNA	4. 3738X Test Set
2. Synthesizer #1	5. mmWave Module 1
3. Synthesizer #2	6. mmWave Module 2

Figure 17-2. Broadband and mmWave Setup

Some things to point out:

- The IF cables from the 3738X test set to the MS464xB (both rear panel) are not shown here.
- The External I/O connector is used to supply simple control bits to the 3738X test set. These bits are level-sensitive only, asynchronous, and low speed and there is no addressing so this connection is relatively simple to administer.
- The synthesizers are synchronized to the 10 MHz reference of the MS464xB so frequency accuracy is maintained. External 10 MHz sources can also be used if they have at least the spectral purity of the VNA internal clock. For more information, see:
  - *MS464xB Series VNA Technical Data Sheet and Configuration Guide*—11410-00611
- The cable bundles between the test set and the mmWave modules supply analog signals, bias and simple control information. Anritsu-supplied bundles should be used here in most cases.
- The GPIB connections are used for controlling the synthesizers and usual GPIB cable length practices apply. Note that the "dedicated" connector on the VNA is used. The synthesizer GPIB addresses must match those on the VNA external source address menu (see Figure 17-3 below, the menu is located under in the SYSTEM menu group:
  - MAIN | System | SYSTEM | Remote Interface | REMOTE INTER. | Ext. Sources | EXT. SRC ADDR.

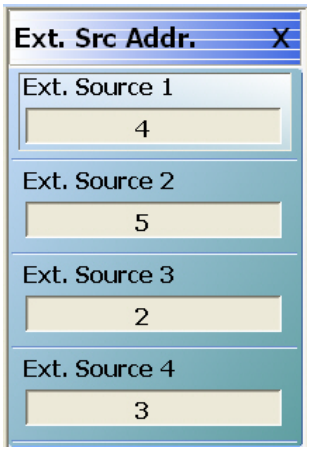


Figure 17-3. EXT. SRC. ADDR. (EXTERNAL SOURCE ADDRESS) Menu

- Note that in broadband/mmWave mode, External Source 1 is defined as the LO and External Source 2 is defined as the RF. Ensure that the RF/LO connections between the synthesizers and the 3738X test set match this designation. If it is desired that other external sources be used, multiple source mode should be used instead of broadband/mmWave.
- For most mmWave modules, the synthesizers must have a maximum frequency of at least 20 GHz and a minimum frequency of no more than 2 GHz. For the control software to operate, the synthesizers must be from Anritsu.
- Normally the IF signals from the mmWave modules are routed through the 3738X test set (amplifiers and some switching) on their way to the rear panel of the MS464xB Series VNA. Depending on the setup, the IF signals can sometimes be routed directly to the VNA.

The system will poll the GPIB bus for the external synthesizers periodically including when the broadband/mmWave mode is activated and when its menus are accessed. If unsure if a connection has been established, check the EXTERNAL SOURCE CONTROL menu (under multiple source setup) to see if they are ACTIVE as in Figure 17-4. In broadband and mmWave modes, the External Sources 1 and External Sources 2 should be automatically activated if they are detected.



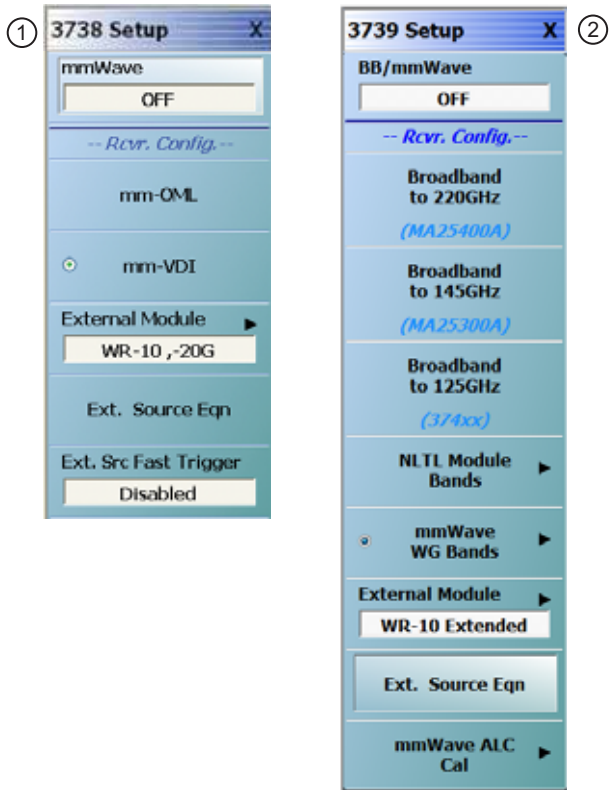
Figure 17-4. EXT. SRC. CONTROL (EXTERNAL SOURCE CONTROL) Menu

If one or both synthesizers are turned off or disconnected while in broadband or mmWave modes, GPIB errors or sluggish system response may be noted. It may be necessary to return to standard mode, reconnect/restart the synthesizers and then re-enter the broadband/mmWave mode to continue. In extreme circumstances, one may want to exit the VNA application, restart the synthesizers, and restart the VNA to establish communication.

17-4 Main Instrument Setup—Classical BB Mode

The BB/mmWave Setup selection brings up the menu shown in Figure 17-5. The On/Off toggle at the top duplicates, in some sense, the Rcvr Config selection in the middle of the APPLICATION menu (turning BB/mmWave OFF will change the selection to standard).

The Rcvr Config section on the SETUP menu provides the user broadband and mmWave selections. As stated above, the distinction involves whether the instrument will be allowed to couple sweeps of the internal VNA hardware with sweeps of the external modules.

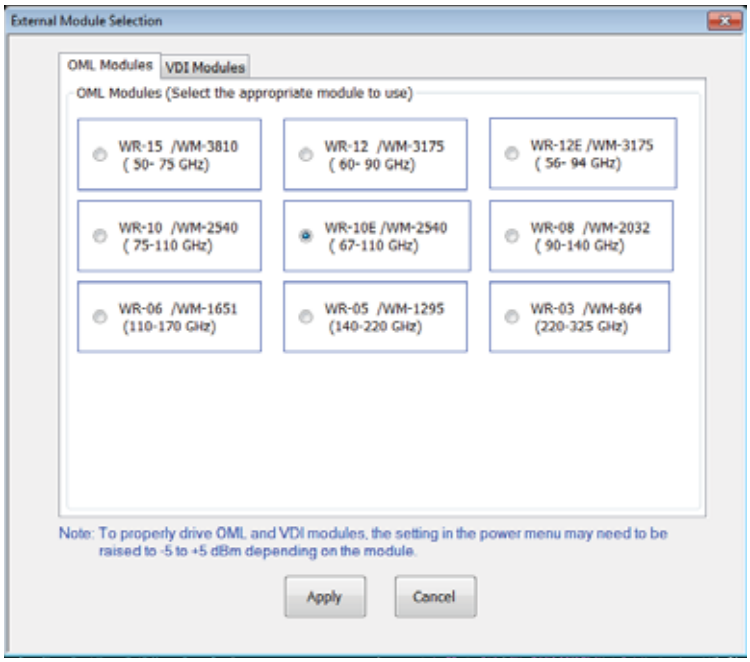


1. 3738 SETUP Menu—For non-MS4647B VNAs	2. 3739 SETUP Menu—For MS4647B, Option 80, 81, 84, or 85 VNAs

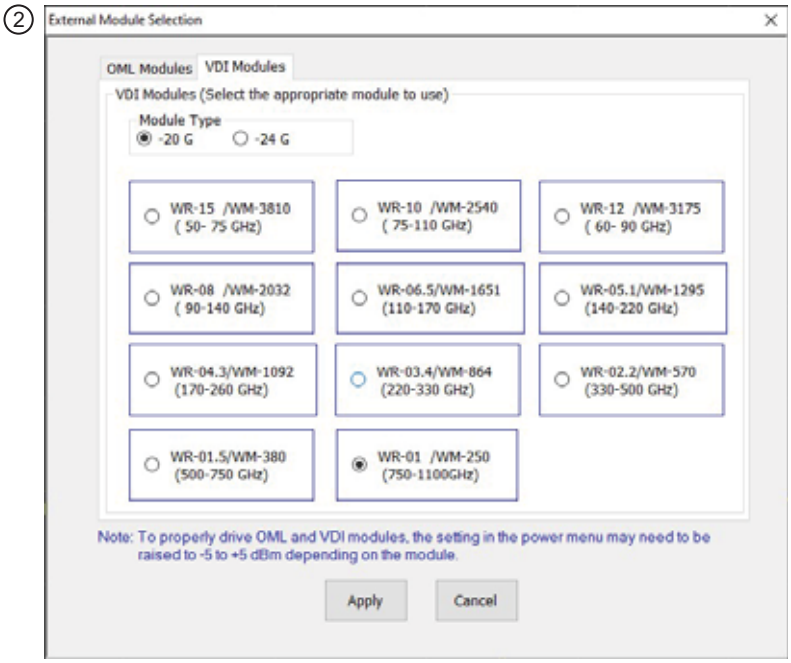
Figure 17-5. BB/mmWAVE Setup Menus

The external module selection, available only when the mode is mmWave (unavailable when in broadband mode), allows the selection of different module frequency ranges. This selection populates the frequency range and synthesizer control fields (harmonic number used for calculating synthesizer frequencies, etc.). These values can be overridden in the EXT. SOURCE EQUATION dialog or one can go to full manual control using multiple source. Selection of the External Module button displays tabbed dialog box with current waveguide selections as shown in Figure 17-6 below.

①



②

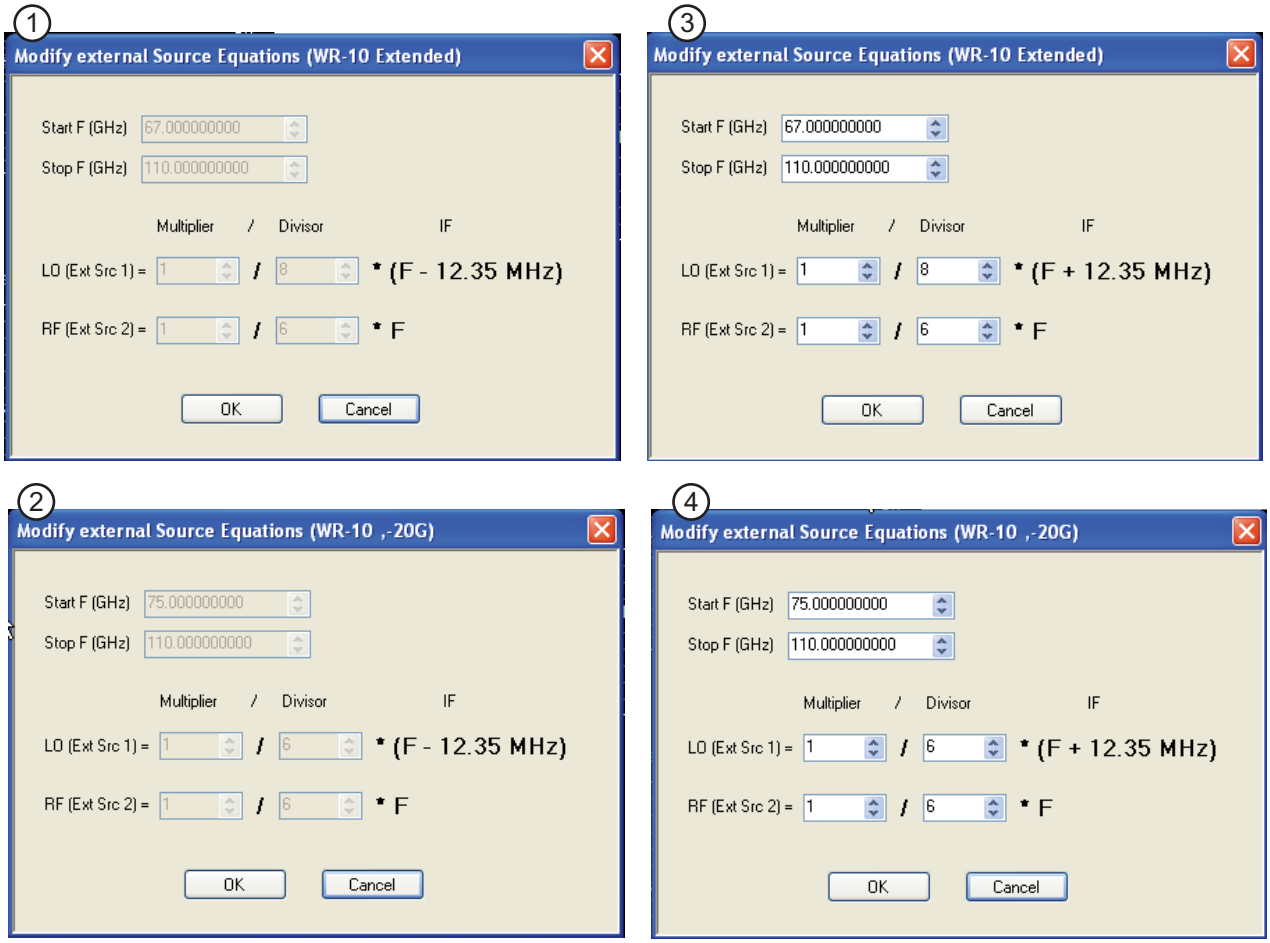


1. At top, the OML tab is selected and the available OML Modules are shown. Select a module radio button to configure that module.	2. At bottom, the VDI tab is selected and the available VDI Modules are shown. Select a module radio button to configure that module. Note that VDI modules have two basic types at –20 G and –24 G. The module type selected here must match the module being used, as the two types have slightly different frequency plans.

Figure 17-6. EXTERNAL MODULE SELECTION Menu

The EXTERNAL SOURCE EQUATION dialog displays the frequency ranges used by the module and the equations used to program the synthesizers. These values can be edited by the advanced user who wants to change the behavior of the modules or if new modules become available (not commonly used).

There are different versions of this dialog for broadband and for mmWave mode, shown in Figure 17-7 below.



1. 3738 Modify External Source Equations (OML Modules)
2. 3738 Modify External Source Equations (VDI Modules)
3. 3739 Modify External Source Equations (OML Modules)
4. 3739 Modify External Source Equations (VDI Modules)

Figure 17-7. MODIFY EXTERNAL SOURCE EQUATIONS Dialog Box

Note that the IF is fixed and un-editable in these dialogs (12.35 MHz). If more advanced adjustments of the frequency plan are desired, the user should go to multiple source mode.

To enable or disable fast trigger mode when using MG37xxxX external synthesizers, click on the Ext. Src. Fast Trigger button which will bring up the dialog. Select the desired mode (ensuring needed cables are connected) and click ok to effect the mode change.

Power control is possible by using the EXT SRC POWER menu for the RF source (Ext Src 2 listed above). It should be noted, however, that the multipliers used in the mmWave modules have extremely non-linear power in-power out transfer curves so the use of the input power (via the menu just referenced) as control can result in violent output power changes. In some firmware versions, an open-loop power calibration can be performed to simplify this task.

From a troubleshooting point-of-view, the most typical problems involve synthesizer communications (either GPIB or fast trigger). As discussed elsewhere, resetting or power cycling of the synthesizers may be required if no problems are found with the cabling or address settings. Sometimes RF and LO cables are swapped which can result in odd ripples in unratioed parameter displays (inappropriate power levels resulting from the swap).

17-5 Modular BB—Physical and Measurement Steps

The typical setup for the modular BB system is shown in Figure 17-8 below. In contrast to the classical BB/mmWave setup discussed previously, external synthesizers are not used but conditioned versions of internal VNA source and LO signals are instead routed to the modules. An option on the VNA (-8x) provides these connection points that the test set (3739x) uses for routing to the modules. Much like in the classical setup discussed before, control and power connections are provided by the test set to the modules and IF signals are taken from the modules for routing to the VNA. As with the classical mode, Option 7 (frequency offset) is required for the more elaborate frequency programming.

The 3739X operation with the MS464XX is controlled by the Options 8x mentioned above. These options are correlated with specific system designations and functionality but there is less correlation with which modules will be used. Table 17-1 on page 17-10 may help explain this.

Two additional rear panel connections are used in this system: an external ALC connection is used between the test set and the VNA to enable leveled power control at the heads and the VNA external analog out function is used to assist in LO leveling. Note that when in Modular BB mode, the normal function of the Ext Analog Out connector is suspended.

When in Modular BB mode, the frequency plan calculations are automatically changed to account for the multiplier behavior and the allowed frequency ranges.

- For ME7838A/AX and related systems (using the 3743A/AX modules), the range is 70 kHz to 125 GHz if Option 70 is installed (10 MHz start frequency otherwise).
- For ME7838E/EX and related systems (using the 3743E/EX modules), the maximum frequency is 110 GHz.
- For ME7838D and related systems (using the MA25300A modules), the maximum frequency is 145 GHz.
- For ME7838G and related systems (using the MA25400A modules), the maximum frequency is 220 GHz (operational to 226 GHz).

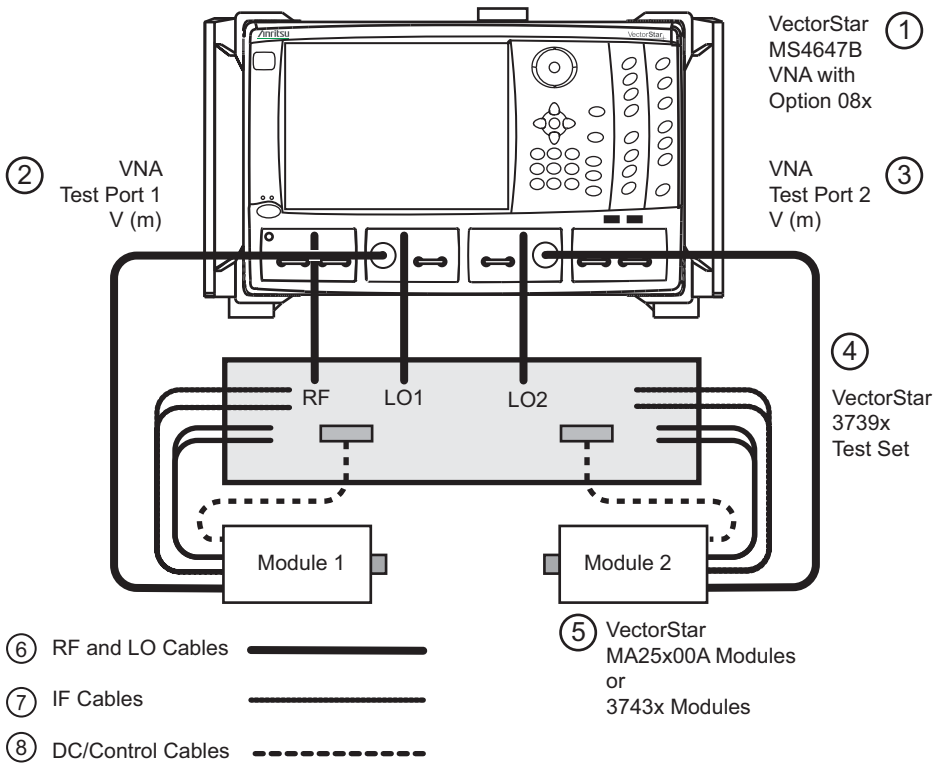


Figure 17-8. System Setup for Modular BB Mode (1 of 2)

The system setup for the modular BB mode is shown here. Rear panel connections are not shown. The shaded boxes on the modules denote the test port connectors. The VNA in this figure has Option 8x which provides the **Src**, **LO1**, and **LO2** connectors shown in the figure. 1. MS4647B VNA with Option 8x 2. VNA Test Port 1 V (m) 3. VNA Test Port 2 V (m)	4. 3739x Test Set 5. MA25300A Modules, 2 each, or MA25400A modules, 2 each, or 3743A/AX/E/EX Modules, 2 each, or 3744A/E/EX-xx Modules, 2 each 6. Heavy lines are RF and LO Cables 7. Light dashed lines are IF Cables 8. Heavy dashed lines are DC/control cables

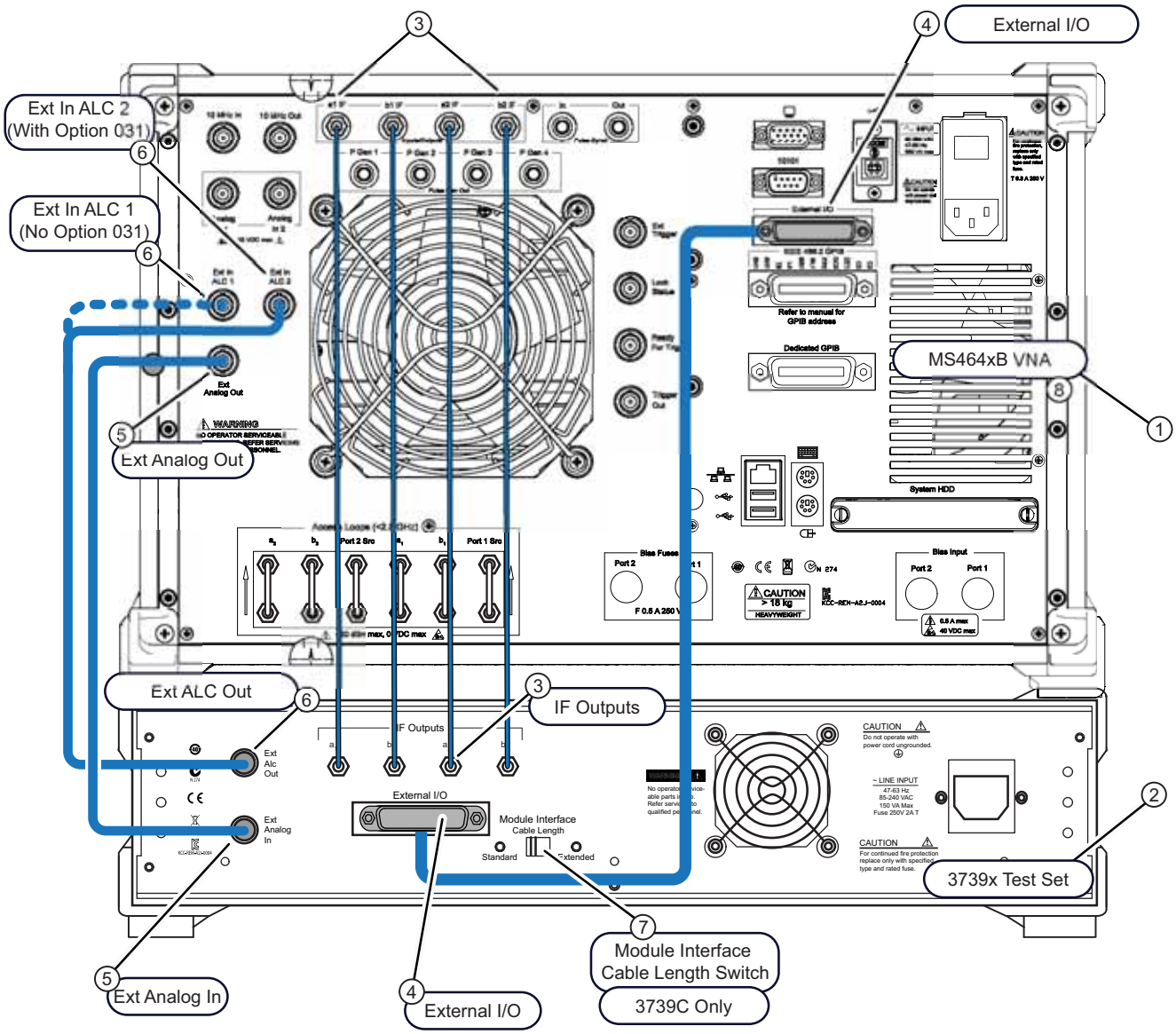
Figure 17-8. System Setup for Modular BB Mode (2 of 2)

Table 17-1. Option 8x Descriptions

Option	Description
080	For ME7838A/AX/ME7838D/ME7838G systems, used for single source VNAs (no Option 31) and ***no*** Option 051, 061, or 062 where broadband operation is desired (to 125 GHz/145 GHz/220 GHz with the appropriate modules).
081	For ME7838A/AX/ME7838D/ME7838G systems, used for single source VNAs (no Option 31) and ***with*** Option 051, 061, or 062 where broadband operation is desired (to 125 GHz/145 GHz/220 GHz with the appropriate modules).
082	For ME7838A/AX/ME7838D/ME7838G systems, used for single source VNAs (no Option 31) and ***no*** Option 051, 061, or 062 where only banded operation broadband operation is desired (E and W band).
083	For ME7838A/AX/ME7838D/ME7838G systems, used for single source VNAs (no Option 31) and ***with*** Option 051, 061, or 062 where only banded operation broadband operation is desired (E and W band).
084	For ME7838A/AX/ME7838D/ME7838G systems, used for dual source VNAs (with Option 31) and ***no*** Option 051, 061, or 062 (no distinction between banded and broadband in this case; operational to 125 GHz/145 GHz/220 GHz with the appropriate modules).
085	For ME7838A/AX/ME7838D/ME7838G systems, used for dual source VNAs (with Option 31) and ***with*** Option 051, 061, or 062 (no distinction between banded and broadband in this case; operational to 125 GHz/145 GHz/220 GHz with the appropriate modules).
086	For ME7838E/EX systems (using 3743E/EX modules), used for single source VNAs (no Option 31) and ***no*** Option 051, 061, or 062 (no distinction between banded and broadband in this case; operational to 110 GHz).
087	For ME7838E/EX systems (using 3743E/EX modules), used for single source VNAs (no Option 31) and ***with*** Option 051, 061, or 062 (no distinction between banded and broadband in this case; operational to 110 GHz).
088	For ME7838E/EX systems (using 3743E/EX modules), used for dual source VNAs (with Option 31) and ***no*** Option 051, 061, or 062 (no distinction between banded and broadband in this case; operational to 110 GHz).
089	For ME7838E/EX systems (using 3743E/EX modules), used for dual source VNAs (with Option 31) and ***with*** Option 051, 061, or 062 (no distinction between banded and broadband in this case; operational to 110 GHz).

In terms of completing the connections, the power/control cable ends at the module side with a snap-in microD connector. The two IF connectors are SSMC and the reference connector is closest to the middle of the module (this is also marked on the module). The V cable coming from the VNA port routes to the module V connector in the middle of the module. The K (LO) and V (Src) cables from the test set route to the appropriate outer connectors on the module. All coax connectors should be torqued properly.

The rear panel connections are shown in Figure 17-9. Two BNC-terminated coax cables are used for routing Ext. ALC (External Automatic Level Control) and analog control signals between the test set and the VNA. A DB-25-terminated cable is used for routing digital control signals, and the IFs are routed using four SMA-terminated coax cables.



1. MS4647B VNA with Option 8x Rear Panel
2. 3739x Test Set Rear Panel
3. IF Input/Output Connections
   —VNA connections at top
   —Test Set connections at bottom.
4. External I/O Connections
   —VNA DB-25P at top
   —Test Set DB-25P at bottom.
5. External Analog Connections
   —VNA BNC at top
   —Test Set BNC at bottom.
6. External Automatic Level Control Connections
   —VNA BNC at top—Test Set BNC at bottom.
   — Ext In ALC 1 (without Option 31)
   — Ext In ALC 2 (with Option 31)
7. Module Interface Cable Length Switch (3739C only)

Figure 17-9. Rear Panel Connections for the Modular BB System

A block diagram of the system is shown in Figure 17-10, “Modular BB Block Diagram” on page 17-13 below. The 3739x test set provides several distinct functions:

- Source power amplification and transfer switching
- LO leveling and distribution
- Generating control signals for the modules from VNA input and supplying power
- Routing and processing IF signals as necessary

The 3739x test set includes higher power amplifiers and further LO condition circuits for use with higher millimeter wave bands. The B model also includes an AUX POWER connector for a third module. This connector supplies power to the module and mimics the control signals sent to the Port 1 module. This makes the port very useful for IMD and mixer measurements. The second tone (for IMD) or the LO (for millimeter wave mixers requiring a high frequency LO) can be supplied this way. The second tone for these applications could be supplied by an external synthesizer (see multiple source in Chapter 18) or the optional second source in the VNA itself (see Chapter 26).

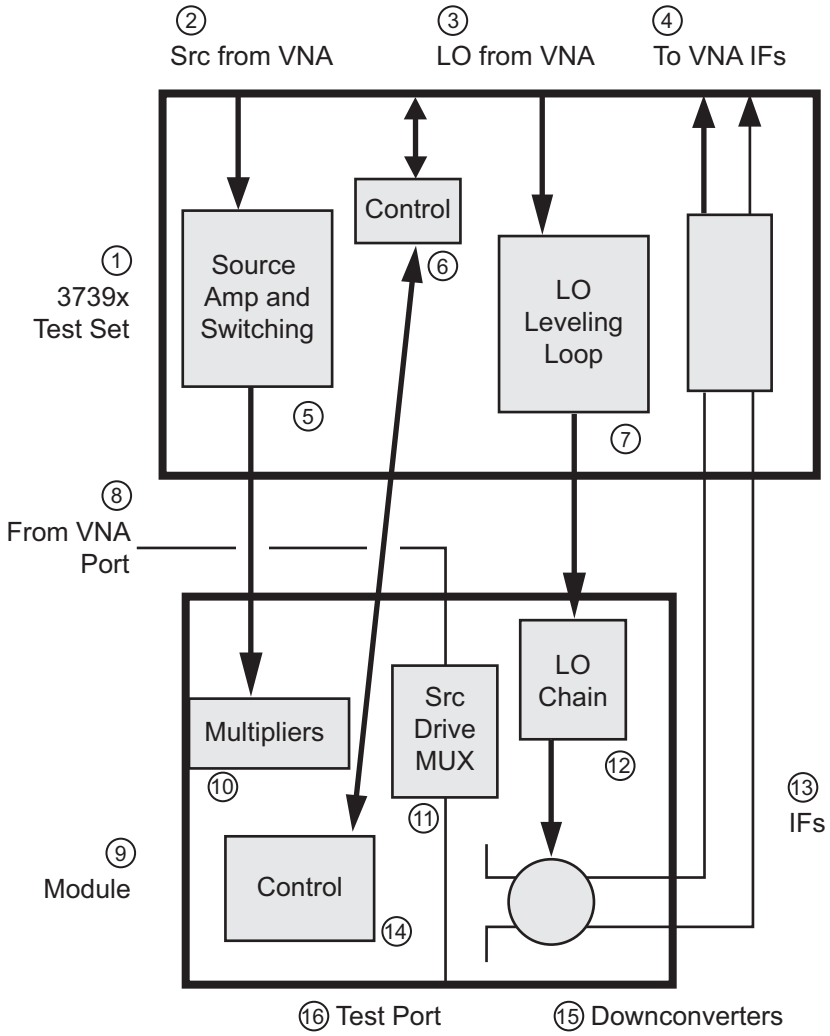
The modules themselves have several unique aspects:

- A bypass path for use at low frequencies. The internal VNA source and couplers/bridges are used for lower frequency measurements.
- At higher frequencies, separate reflectometers in the module handle the measurements.
- Also at higher frequencies, multipliers in the modules provide source power.
- Leveling circuitry (both source and LO) is built into the modules.

There are two main breakpoint frequencies that are helpful in understanding the performance of the system:

- The break where the module receivers take over from the internal VNA receivers. This is nominally at 30 GHz.
- The break where the module source multipliers take over for the main VNA multipliers. This is nominally at 54 GHz.

As with the classical approach discussed previously, this mode can also be invoked from within multiple source control for a more customized setup. Details of this approach are covered in Chapter 18, "Multiple Source Control (Option 7)".



The modular BB block diagram is shown above. Relative to the classical setup, the MUXing is performed in a different location and different multiplier and downconverter technologies are used.

1. 3739x Test Set
2. Src from VNA
3. LO from VNA
4. To VNA IFs
5. Source Amp and Switching
6. Control
7. LO Leveling Loop
8. From VNA Port
9. MA25400A, MA25300A, or 3743x Module
10. Multipliers
11. Src Drive MUX
12. LO Chain
13. IFs to Test Set
14. Control
15. Downconverters
16. Test Port

Figure 17-10. Modular BB Block Diagram

Full power control and power sweeps are available in this mode due to the integrated leveling circuitry in the modules and test set. Flat power calibrations/linear power calibrations are also available and function as described in other chapters. An important point is that the standard Anritsu power meters are used up to 70 GHz while separate waveguide-based power meters are needed above 70 GHz. Normally an Agilent W8486A W-band sensor or equivalent is used for 70 GHz–125 GHz (for ME7838A/AX and ME7838E/EX and related systems, only used to 110 GHz for the E/EX systems) and an Elva-1 power meter with D-band sensor is used for frequencies of 125 to 170 GHz (for ME7838D/ME7838G and related systems). For frequencies above 170 GHz (ME7838G and related systems), an Elva-1 power meter with a G-band sensor is used. For traceability and accuracy information, contact the factory. The calibration routines will prompt when a different meter/sensor needs to be connected. Because of the need for adapters in many cases, the manual power offset feature may be employed on the power meters to correct for adapter loss (e.g., a W1-WR10 coax-to-waveguide adapter typically has 0.5 dB of loss at 110 GHz).

Full ALC calibrations are performed at the diagnostics menu level of the firmware and are normally only employed in service situations. When moving between Standard and Modular BB modes, the ALC calibration files automatically switch to properly handle the hardware being used in those modes.

From a troubleshooting point-of-view, miscabling of the system is the most common problem. A typical symptom is non-ratioed parameters that are very low (< – 90 dB; when at default power, $a_1/1$ and $a_2/1$ will typically be in the –10 to –30 dB range (uncalibrated) depending on frequency). Very low received signals can sometimes also be due to requesting a leveled power too far below specified ranges (the system will allow lower entries than specified since they may be achieved at certain frequencies).

The normal ALC range when above 54 GHz reaches down to –60 dBm, although there may be some significant drop-offs below –55 dBm (and below –50 dBm above 110 GHz for those systems operating to higher frequencies). This is possible because of the highly linear integrated UIF leveling loop based within the modules. A separate RF leveling loop is also available for frequency translating measurements (discussed in Chapter 18). Because no step attenuators are in the mmWave part of the system, this extended ALC range is particularly useful. Since the power structures are fairly different above and below 54 GHz (with or without step attenuator possibilities), two different ALC entries are used and are labeled as such on the Power menu. An example of the wide power range and wide power sweep range is shown in Figure 17-11 (the curve flat lines at the upper end of the sweep as it reaches maximum power of ~+3 dBm for this frequency). To achieve a constant low power level in a broadband sweep, it will be necessary to use step attenuators below 54 GHz and a low base setting above 54 GHz.

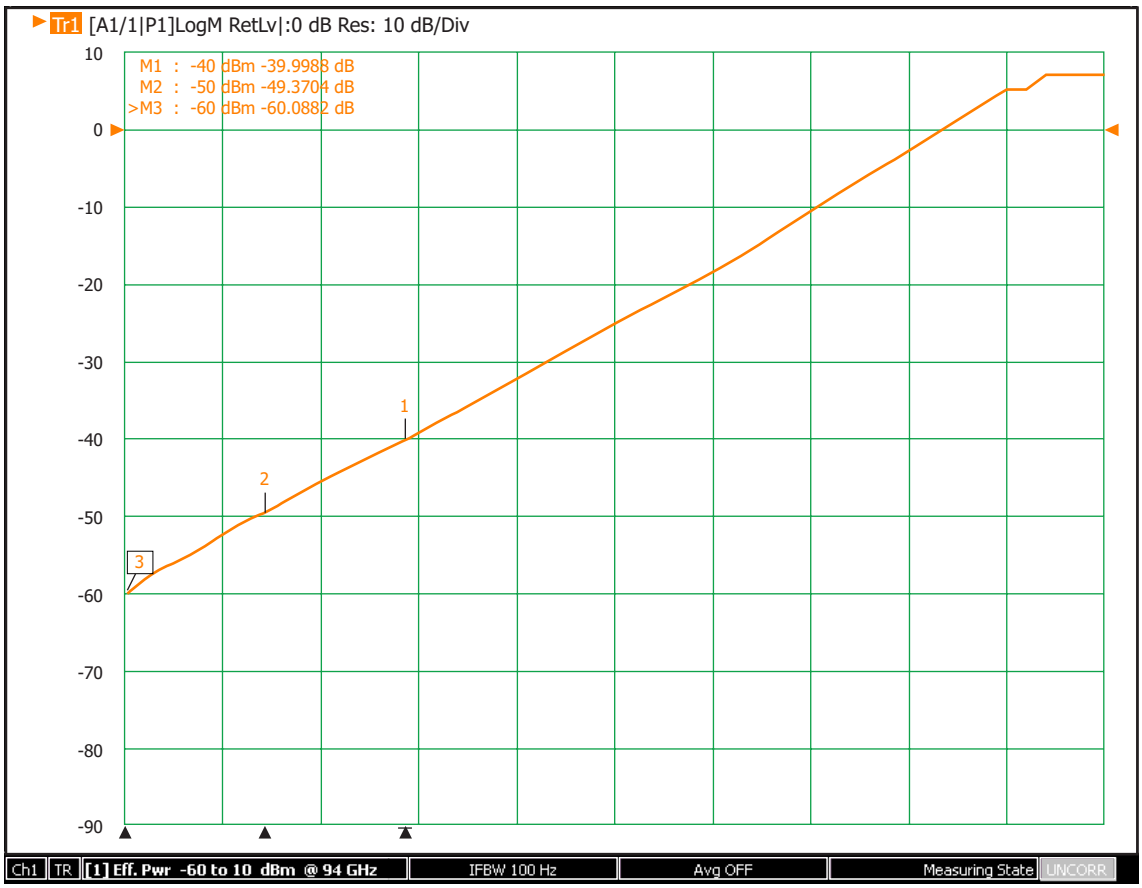


Figure 17-11. Example of Wide Power Sweep Range Available with the Modular BB System.

Example:

- Broadband sweep with port power of –50 dBm
- Set the ALC level (<54 GHz) to –20 dBm and engage 30 dB of source step attenuation (assumes Option 61 or 62)
- Set the ALC level (>54 GHz) to –50 dBm

User power calibrations can be challenging with the broadband systems at lower power levels since most of the available sensors have limited dynamic range (often –25 dBm as a practical lower limit). To get around this issue, the system automatically uses a different approach when the requested power level is below –25 dBm:

- The requested power is set to –20 dBm and a power measurement is taken with the appropriate power meter. At the same setting, a measurement is made by the a_x receiver (the reference signal where x represents the current driving port). This is an internal re-referencing step where the power meter accuracy is transferred to the receiver. Since the instrument receivers are very linear (to within 0.01 dB down to levels only limited by the system noise floor), this is an efficient transfer process to allow for low level calibrations.
- The system is then set to the target power level and an a_x measurement is made and this is converted to port power based on the above re-referencing. The search process then proceeds as usual to get the re-referenced power to the desired target.
- This process is repeated for each frequency and power (depending on sweep mode) for power levels below –25 dBm.

Anritsu PowerMaster frequency-selective power sensors also have extended dynamic range since they use a downconversion process instead of thermal sensing for power measurement. The frequency selective power meters can be used for slightly lower power levels (but generally not less than about –35 dBm).

An added tool to assist in power calibrations is the ability to embed or de-embed .s2p files describing networks added after the power calibration. One example of this is moving the power reference plane from the connector or flange interface to the end of a probe tip.

17-6 3739 mmWave Modes

In addition to the broadband measurements described in the previous section, the 3739x test set supports a number of different banded mmWave measurement setups that are described in this section.

The broadband selections allow operation from the low-frequency limit of the instrument (70 kHz or 10 MHz depending on Option 70 presence/absence) to the upper limit of the mmWave module.

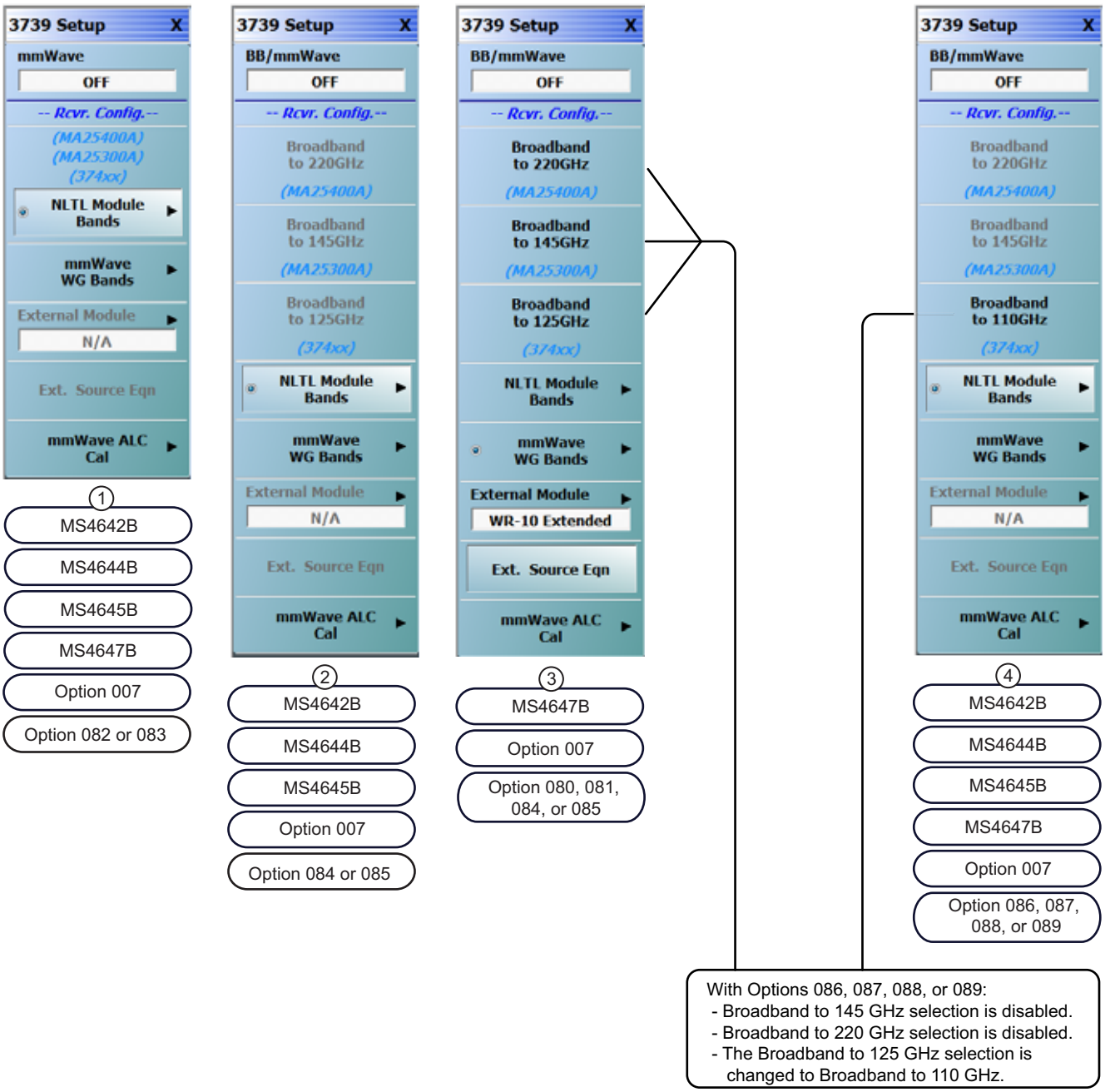
- For ME7838E/EX systems, the maximum is 110 GHz.
- For ME7838A/AX systems (using 3743A/AX modules), the Broadband to 125 GHz selection should be used.
- For ME7838D systems (using MA25300A modules), the Broadband to 145 GHz selection should be used.
- For ME7838G systems (using MA25400A modules), the Broadband to 220 GHz selection should be used.

Note	The system cannot tell which modules are connected so the user MUST select the correct *broadband* option when one is offered.

For use of banded versions of the 3743x mmWave modules (described above and sold as the 3744x), the appropriate frequency band selection should be made from the NLTL MODULES BANDS menu. These modules allow operation in 145 banded GHz range or extended E (54–95 GHz) or extended W (65–110 GHz) bands and are convenient for purely waveguide-based applications. The corresponding options on the VNA are 082 (without step attenuators) and 083 (with step attenuators). All of the 08x options are mutually exclusive.

External OML/VDI mmWave modules can also be selected for operation. Much like the 3738 test set-based systems, OEM mmWave modules are used, except the internal VNA sources are used to drive the modules and additional leveling capabilities are enabled. The mmWave module selection is done from the mmWAVE WG BANDS menu.

The 3739-based setup is enabled from the RCVR CONFIG menu and has additions (see Figure 17-12) relative to the 3738 setup for the E-and W-banded solutions, as well as for a leveling calibration that will be discussed shortly. The EXTERNAL EQUATION dialog has the same structure as for the 3738 setup, and an example for a WR-2.2 (24 GHz) module is shown in Figure 17-13.



1. The 3739 SETUP Menu for MS4642B, 44B, and 45B VNAs with Option 07 and Option 82 or 83.
2. The 3739 SETUP Menu for MS4642B, 44B, and 45B VNAs with Option 07 and Option 84 or 85.
3. The 3739 SETUP Menu for MS4647B VNAs with Option 07 and Option 80, 81, 84, or 85.
4. With Options 86, 87, 88, or 89, the 145 GHz selection is disabled and the 125 GHz selection is changed to 110 GHz.

Figure 17-12. 3739-based SETUP Menu

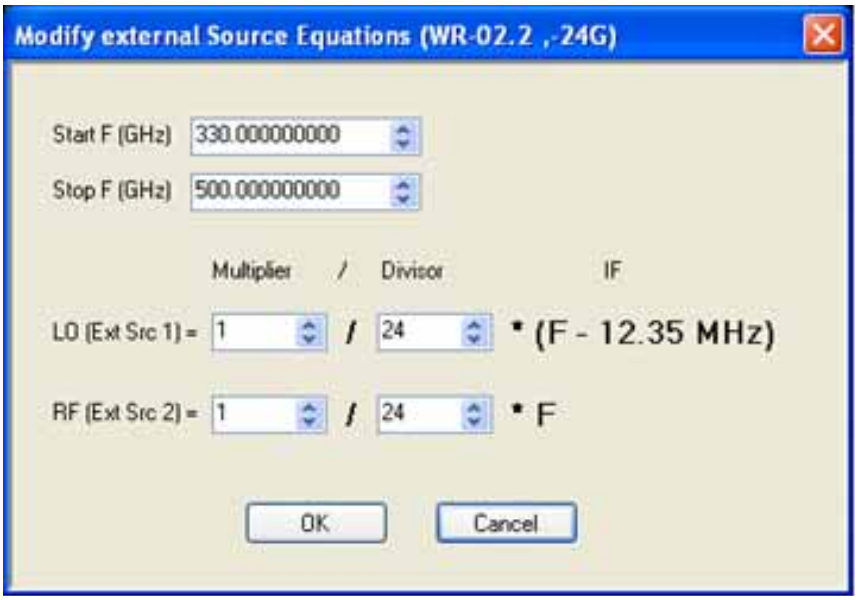


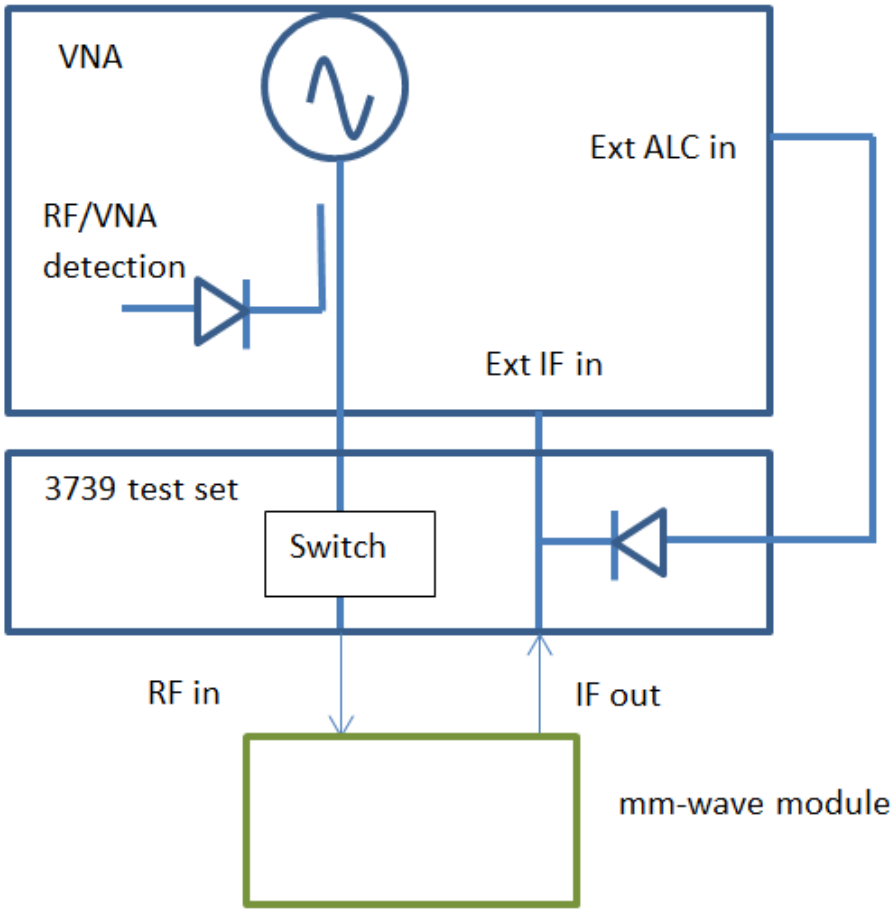
Figure 17-13. Example EXTERNAL SOURCE EQUATION Dialog Box for a WR-2.2 Setup

The 3739-based mmWave systems have some power control options available that are not present on the 3738-based systems and they are illustrated by the mmWAVE ALC Cal menu shown in Figure 17-14.



Figure 17-14. mmWAVE ALC CAL Menu

To understand the leveling/power control options, consider the very simplified block diagram in Figure 17-15. The key point is that there are multiple locations to sample the power, but because of the very high frequencies involved, there is potentially no one ideal location.



A simplified block diagram of a 3739-based mmWave setup is shown here to help illustrate power control and leveling concepts.

Figure 17-15. Simplified Block Diagram of a 3739-based mmWave Setup

In the 'VNA' mode, the standard VNA ALC calibrations are used with the existing detection system in the VNA so that the power menu settings correspond roughly to the actual RF levels being delivered to the 3739x test set. One can then have a general idea of the levels driving the module, and may know the corresponding mmWave output power levels based on data from the manufacturer. The leveling loop is closed, but in this mode the instrument has no knowledge of the relationship between mmWave power levels and the instrument settings.

In the RF leveling mode, the same detection circuitry is used as in the VNA mode, but now a means of relating the actual mmWave power levels is available. Because of the paucity of readily-available power measurement methods in these higher frequency ranges, the method used is somewhat indirect, but relies on the established linearity of the measurement system. The concept is illustrated in Figure 17-16. From the manufacturers' measurements (often based on quasi-optical or other techniques), a value is generally known for the saturated mmWave power. The saturation breakpoint can be found by locally sweeping the VNA RF power until the measured IF power leaves saturation (*Pvna1*). The IF reading can be noted at this point (IF1) and it can be linked to the known mmWave saturated power. Now the VNA plane power can be dropped and the IF magnitude recorded as a function of this setting. Based on the known receiver linearity, we can now link the IF changes to the mmWave power level (note this is only done during the calibration step. The leveling is based on finding the needed VNA setting to get the requested mmWave power level, and then relying on the VNA leveling loop). So if the IF level has decreased to IF1 – 10 dB at some input level *Pvna2*, then we can state the mmWave power is Psat–10 dB with some reasonable level of uncertainty. The concept is to level on the VNA power plane while establishing a link to the mmWave power so that user entries can be in the latter form.

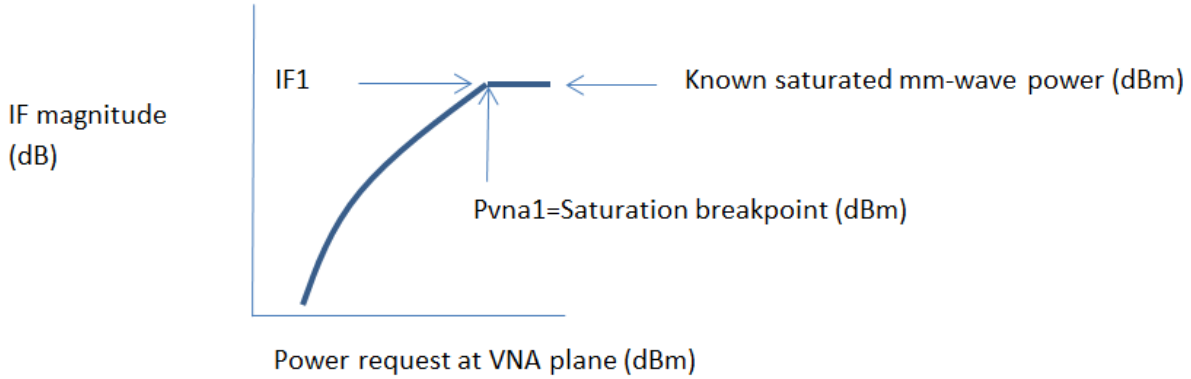


Figure 17-16. RF mmWave Leveling Concept

Since this leveling method only relies on the detection system at the VNA plane, it is useful for all measurement classes including converter and IMD measurements. Because it is leveling prior to the multipliers, it is less stable than the IF leveling to be discussed next and will provide a lower level of long term accuracy.

The calibration is accomplished using the dialog shown in Figure 17-17 (Default mode) or Figure 17-18 (Custom mode). The concept is that the saturated power is known at some frequencies in the band and the user is allowed to enter those values discretely (in Custom mode) or use predefined values for some common modules (in Default mode). In Custom mode, the user specifies exactly which frequencies the calibration will be performed at (maximum of 25). In Default mode, the system will automatically calculate up to 25 frequencies based on the current frequency sweep trying to minimize interpolation issues. If in frequency-based segmented sweep, one frequency in each segment (up to 25) will be used and, if more segments are present, even numbered segments will start to be skipped starting at the end of the list. Default mode is not available for index-based segmented sweep.

When the calibration is executed, the above power sweep and iteration scheme is executed individually at each of the frequencies defined above. Note that Port 1 and Port 2 calibrations are separate, since modules are rarely identical. When the calibration is completed, the calibration may be saved and recalled later using the buttons shown in Figure 17-14. In many modules, the power sweep linearity can change rapidly with frequency so the frequency points of calibration may need to be carefully selected in some cases.

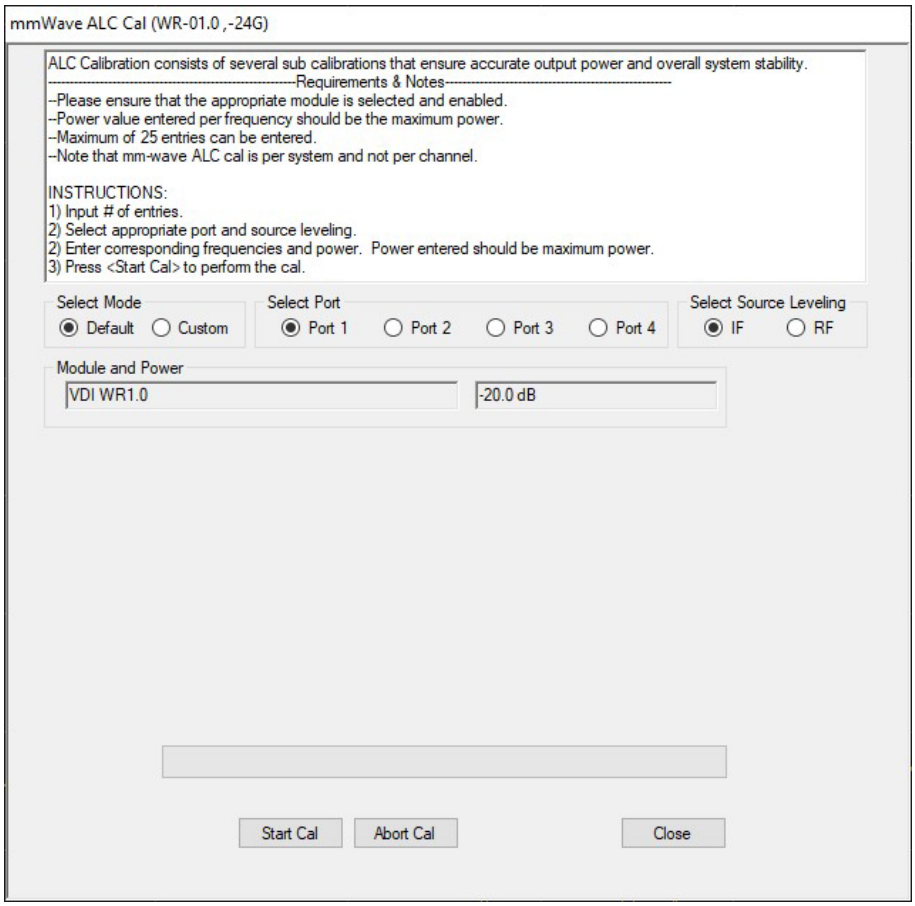


Figure 17-17. mmWAVE ALC CALIBRATION Dialog for a Single Frequency Point of RF Calibration on Port 1 – Default Mode

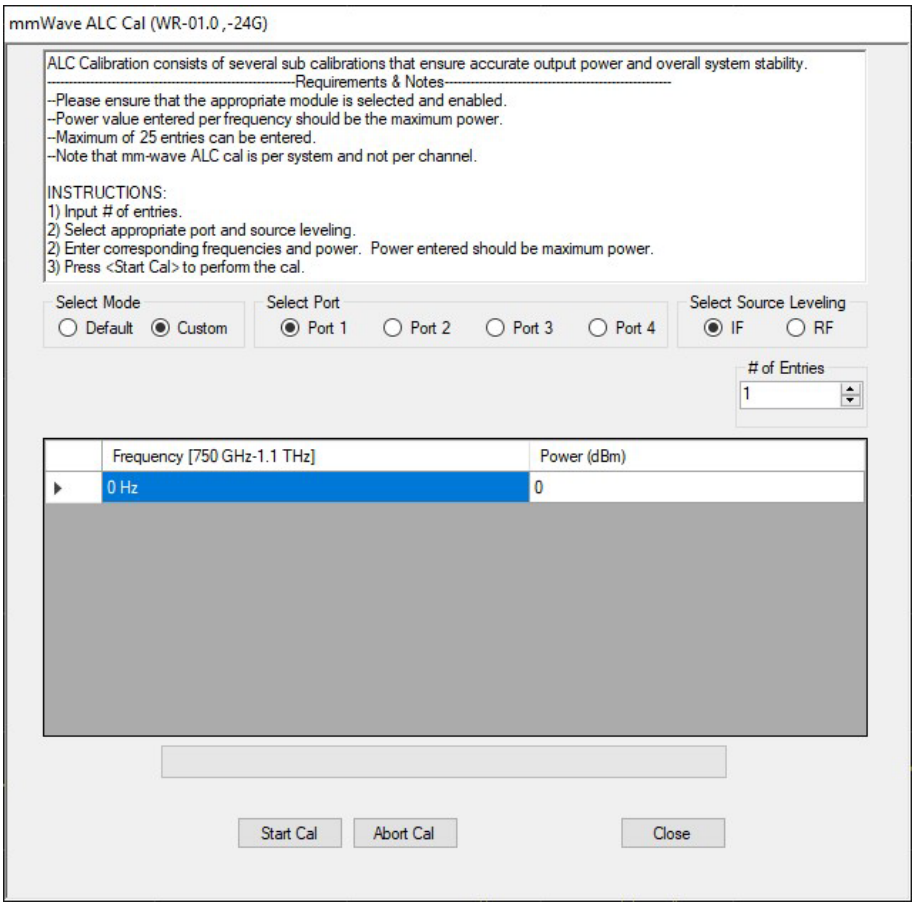


Figure 17-18. mmWAVE ALC CALIBRATION Dialog for a Single Frequency Point of RF Calibration on Port 1 – Custom Mode

The IF leveling scheme relies on sensing the IF amplitude coming out of the module directly and using that IF signal as the leveling signal. This requires that an IF be present on the reference channels so it is not appropriate for mixer or IMD measurements. Since the IF is derived post-multiplication, however, it is much more stable over time and temperature than is the RF leveling scheme discussed before. The block diagram of Figure 17-15 still applies except now the Ext ALC path shown is used for leveling.

The calibration proceeds in the same way as discussed for the RF leveling scheme and the same dialog is used (with different button selections). In both cases, since the leveling is analog, it is very fast so that the DUT rarely sees transient power spikes that can occur under purely software leveling schemes. The detected transient response time for one measurement is shown in Figure 17-19 and illustrates that a ~5 dB power request step is resolved in a few microseconds in an overdamped manner.

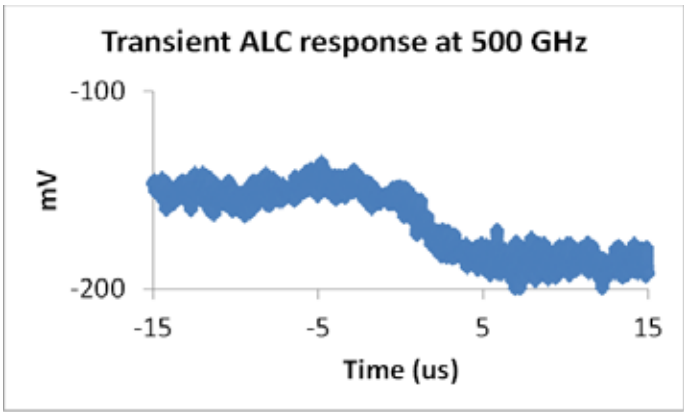


Figure 17-19. Example Transient Response of the mmWave ALC System

Both leveling schemes enable power sweeps as well as finer power control when sweeping frequency. Both modes of control can be useful for active device measurement or for other RF-sensitive or potentially non-linear situations. An example of a power sweep at 400 GHz is shown in Figure 17-20 to illustrate the range that may be possible.

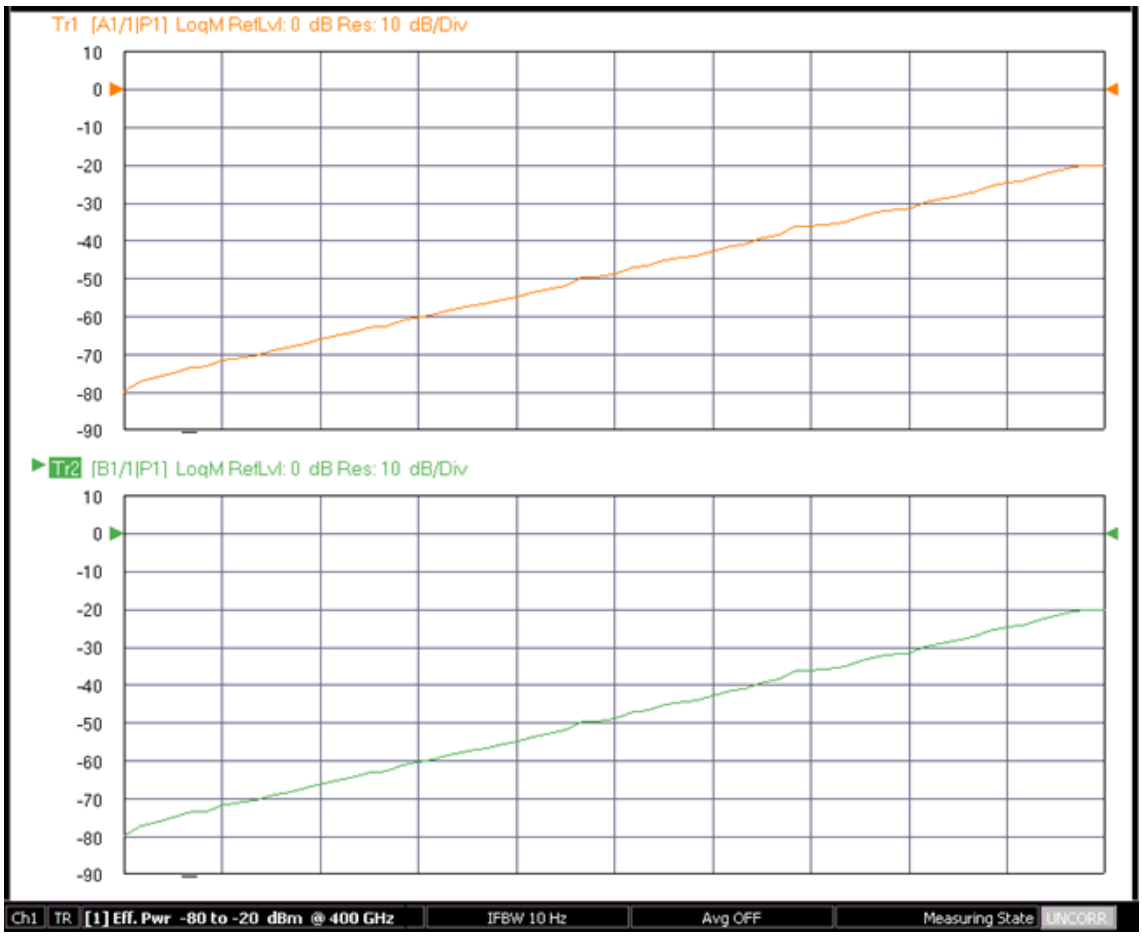


Figure 17-20. Example Power Sweep Measurement

17-7 Measurement Tips

While calibrations are covered in detail in another section of the measurement guide, there are some peculiarities relative to broadband and mmWave operation that require some comment.

- The mmWave bands are often executed in waveguide thus requiring that media type selection (to take care of dispersion correction).
- SOLT/SOLR are often not recommended due to the difficulty of fabricating a reasonable open standard. An open waveguide flange radiates quite effectively and, as a result, is both unstable and has a relatively high return loss.
- SSLT/SSLR are commonly used, particularly at lower frequencies and do require a good load standard. The offset short lengths must be known with some precision.
- SSST/SSSR is quite popular, particularly at higher frequencies since a load standard is not required. Accurate knowledge of the short offset lengths is critical.
- LRL is also a popular technique and is quite effective although it can be sensitive to the condition of the waveguide flanges.

BB Mode

In classical broadband mode (3738-based), a hybrid coupler combining the bands into a single W1 coaxial connector is typically used with a default breakpoint of 67 GHz (i.e., at 67 GHz and below, the base VNA is used for both source and receiver; at 67.000000001 GHz and above, the external modules are used for source and receiver). This structure is key to allowing the combined sweep. To calibrate this connector type, typically two calibrations are recommended: SOLT from the low frequency limit to 67 GHz and SSST from 67 GHz to 110 GHz. The two calibrations are performed separately and then combined using calibration merge (see calibration section for details). The only constraints are the two calibrations must be of the same type (e.g., both 1-port S11 or both full 2-port) and the total point count must not exceed the current maximum for the instrument (25000 or 100000). Note that other calibration types (e.g., LRL) can be successfully combined with this connector type.

Note	It is important that the system be in the appropriate 3739 receiver configuration even when performing the lower frequency calibration. This ensures that the correct receiver mode is used in the merged calibration.

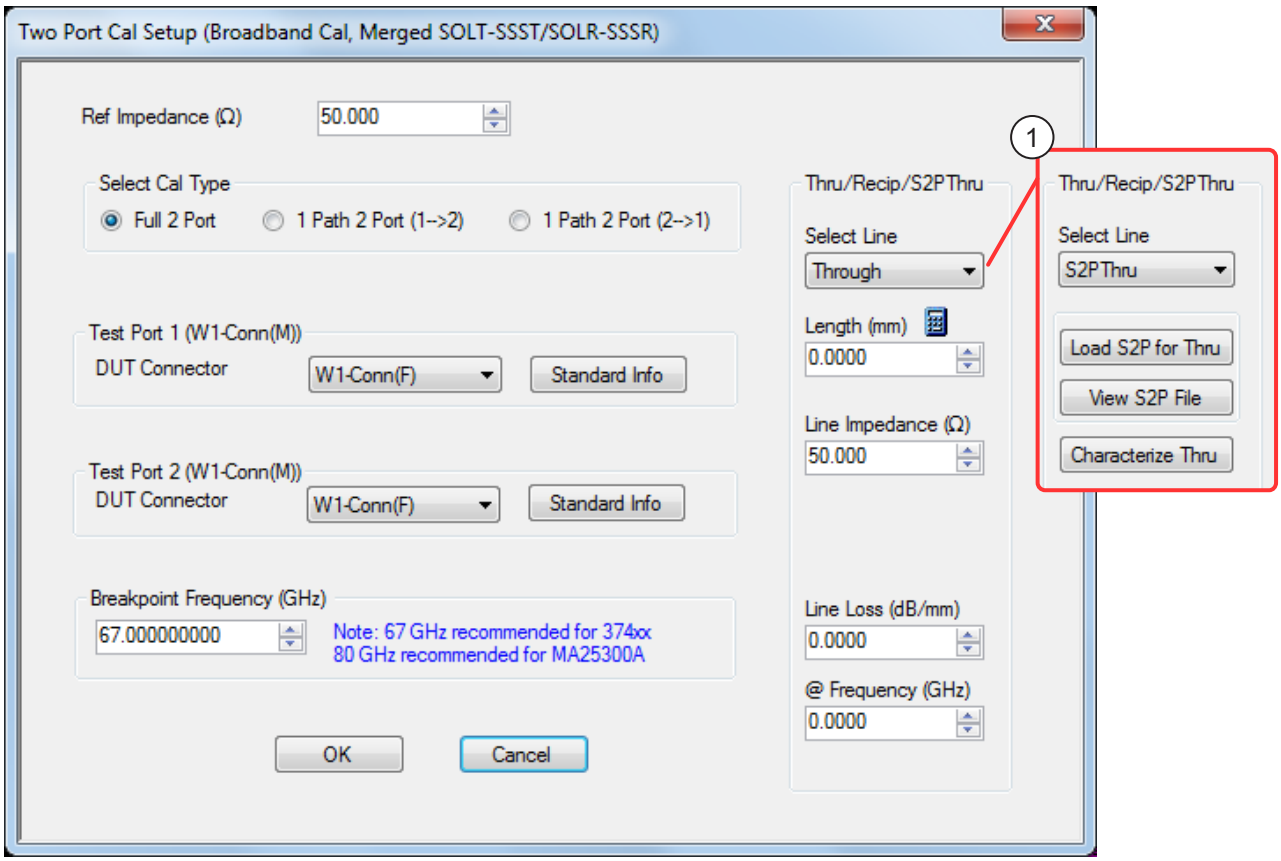
In the 3739-based systems, a similar protocol is used with a 67 GHz breakpoint also typically used for ME7838A/AX and ME7838E/EX (and related) systems (W1 connector-based). For the ME7838D and related systems (using the MA25300A modules) which are 0.8 mm-connector based, a breakpoint for the calibrations of 80 GHz is typically used. In all cases, it is generally a SOLT calibration at the lower frequencies and an SSST calibration at higher frequencies. For the ME7838G and related systems, a coaxial-mode flange-based interface is used for direct connection to wafer probes so the same merged calibrations are used less often (and an on-wafer LRM calibration, for example, is broadband inherently). Adapters from the MA25400A module's native interface to a 0.8 connector are available and calibrations as discussed above using the 0.8 calibration kit are possible. Adapters to WR-5 and WR-6 waveguides are also available. Traditional waveguide calibrations covering the respective bands are then possible.

The 1mm (W1) and 0.8 mm calibration kits both contain 3 shorts, an open, and a load for both connector genders (along with many adapters, tools and verification components). For the SOLT calibration, the coefficients are setup assuming that SHORT 1 is used for that calibration.

Broadband Cal

In some software versions (including and after V2017.06.01), a calibration method termed 'broadband cal' is available for ME7838X systems that allows one to perform the SOLT/R and SSST/R calibrations in one session without using the merge. This broadband calibration saves some connections and a number of keystrokes. Also, since the thru/reciprocal is only connected once, some steps in error coefficients at the breakpoint frequency due to connector repeatability can be reduced. Note that since different algorithms are used above and below the breakpoint, some small steps in data are to be expected (within the uncertainties of each calibration).

As shown in the modify setup dialog below, the concept of the breakpoint frequency is still relevant and one can use the 67 GHz or the 80 GHz (for the 0.8 mm calibration kit and ME7838D) default values or one can enter different numbers. If the current frequency range is either entirely above or below the entered breakpoint, only one of the calibration types will be used but all standards will still need to be connected. Note that the published uncertainty curves only apply if the default breakpoint frequencies are used. Note that options are available on the 3656B/C calibration kits to have .s1p definitions of the standards (in addition to the traditional .ccf files which define the components by means of frequency-dependent circuit models). These .s1p-based calibrations can offer improved uncertainties, particularly for reflection measurements (and all measurements for reciprocal-based calibrations). Lower frequency cal kits (e.g., 3652 and 3654 series) also offer this option.



1. Thru Line selected allows user entries for length, line impedance, line loss and frequency.
   S2P Thru selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 17-21. Edit Cal Params—TWO PORT CAL SETUP Broadband Cal, Merged (SOLT\R-SSST\R) Dialog Box

Note that this calibration method choice will only be available if the system is in an appropriate 3739 mode. The reciprocal choice is available as it is for the base open-short-load and triple-offset-short calibrations but the same reciprocal (or thru) will be used for the entire sweep. If it is desired to use different thrus/reciprocal elements, then the traditional merge method described above should be employed. During the calibration, the open, the load, three shorts and the thru/reciprocal must be measured. Short 1 and the thru/reciprocal data will be used for both upper and lower calibrations. As with the regular versions of these calibrations, losses in the opens and shorts can be defined. See Chapter 3 — SOLT/SOLR Calibration for more information on these entries, but the loss is entered in terms of dB/mm at a specific reference frequency (f_{ref}). Losses at other frequencies are found by scaling as the square root (sqrt(f/f_{ref})). If 0 is entered for the reference frequency, the loss is assumed to be constant with frequency.

As discussed above, the ME7838G and related systems (using the MA25400A module) do not general use this same type of broadband calibration since those systems are almost exclusively used on-wafer with inherently broadband calibrations. Adapters to 0.8 mm can be used with the MA25400A modules and a broadband calibration in 0.8 mm can be performed to 145 GHz.

Modular BB Approach

In the modular BB approach, the multiplexing is instead done behind the high frequency reflectometers and happens at two disjoint frequencies instead of at a common breakpoint. This typically does not affect the calibration break frequency as the latter is more determined by the characteristics of the calibration kit. Thus when using Anritsu's 3656x W1 calibration kit, the same 67 GHz breakpoint is commonly used for both the classical and modular broadband systems.

Some additional comments related to use of the modular BB system:

- Since the thru path of the modules is DC-connected, the internal VNA bias tees can be used (the bias routes from the VNA rear panel through the VNA front panel ports to the modules, and then to the W1 plane). For Kelvin-based measurements, Anritsu Kelvin bias tees can be mounted on the rear of the modules directly (see Figure 17-23). A variety of bias tees are available with different video bandwidths and current handling capabilities are available; contact Anritsu for more information. The internal DC resistance of the modules is very low so the drop from the bias tee to the DUT can be minimized. Contact Anritsu for a detailed report.

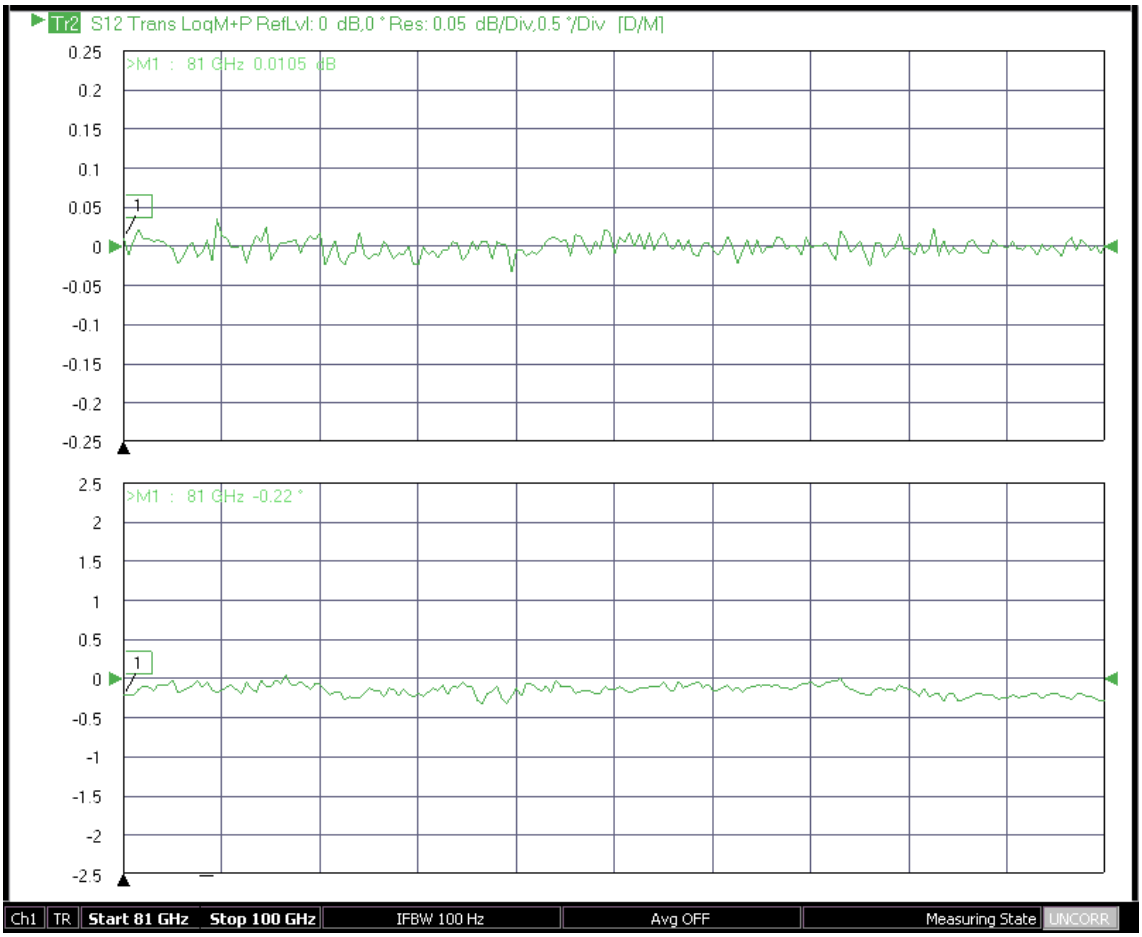
- Because the bias tees often have 70 GHz bandwidth and are very well matched, they have little impact on system raw directivity, in part because they are positioned behind the high frequency couplers.



Figure 17-22. Connection of a Kelvin Bias Tee to a mmWave Module

- Isothermal measurements are often of interest and the position of the module very close to the reference plane sometimes raises the question of thermal effects. When normally mounted, the temperature rise at a wafer probe tip is less than 1 degree C. If even low temperature rises are required for on-wafer or other applications, mounting variations are available to reduce this rise to under 0.3 degree C (about the limit of the thermometry being used). Contact Anritsu for more information.

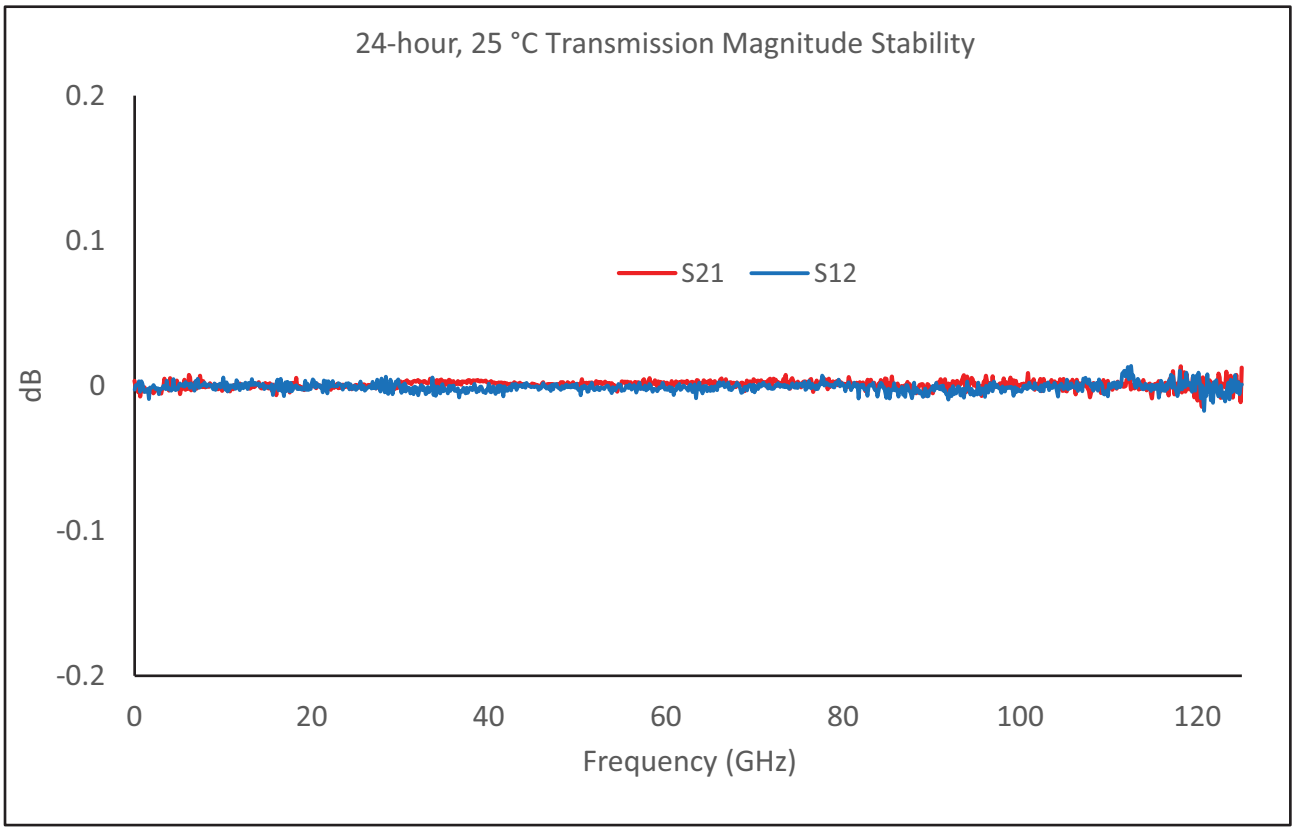
- It is sometimes desirable to position the modules remotely (or further than the usual 1 m cables would allow) because of a complex test setup or for antenna measurement and other applications. Up to 5 m cable runs can be used with reasonable performance (power and dynamic range below 54 GHz will be affected by insertion loss changes, but mmWave performance will not be because of saturating mixers and multipliers) and 15 m cable runs can be used in some circumstances (covering 30–125 GHz for the 3743A/AX module). For information on configuring longer cable sets or for special modifications for even more remote applications, contact Anritsu. Trace noise and short term stability from measurements with a 15 m system are shown in Figure 17-24. This measurement was done with conventional horn antennas in a laboratory environment and mainly serves to show reasonable close-in trace performance with long cable runs.



Trace noise and stability over about 1 hour are shown here for a 15 m antenna-based measurement.

Figure 17-23. Trace Noise and Stability Over About 1 Hour

- In broadband and mmWave measurements, stability is often of interest because of the possibility of long measurement sessions or longer calibrations. It would be desirable if a calibration could last for a longer period of time to reduce the chance of invalid measurements and to reduce the overall time spent calibrating. The modular BB approach offer some advantages in having the high frequency couplers (covering 30–145 GHz) being right at the 1 mm (or 0.8 mm) connector to minimize raw directivity degradation, a small integrated package to avoid thermal gradients, and a tightly integrated control system. The result is good stability in both reflection and transmission over time as suggested in Figure 17-24. In order to optimize stability, minimizing environmental temperature variations certainly helps as does protecting the cable runs from overstress or unnecessary exposure to thermal variation.



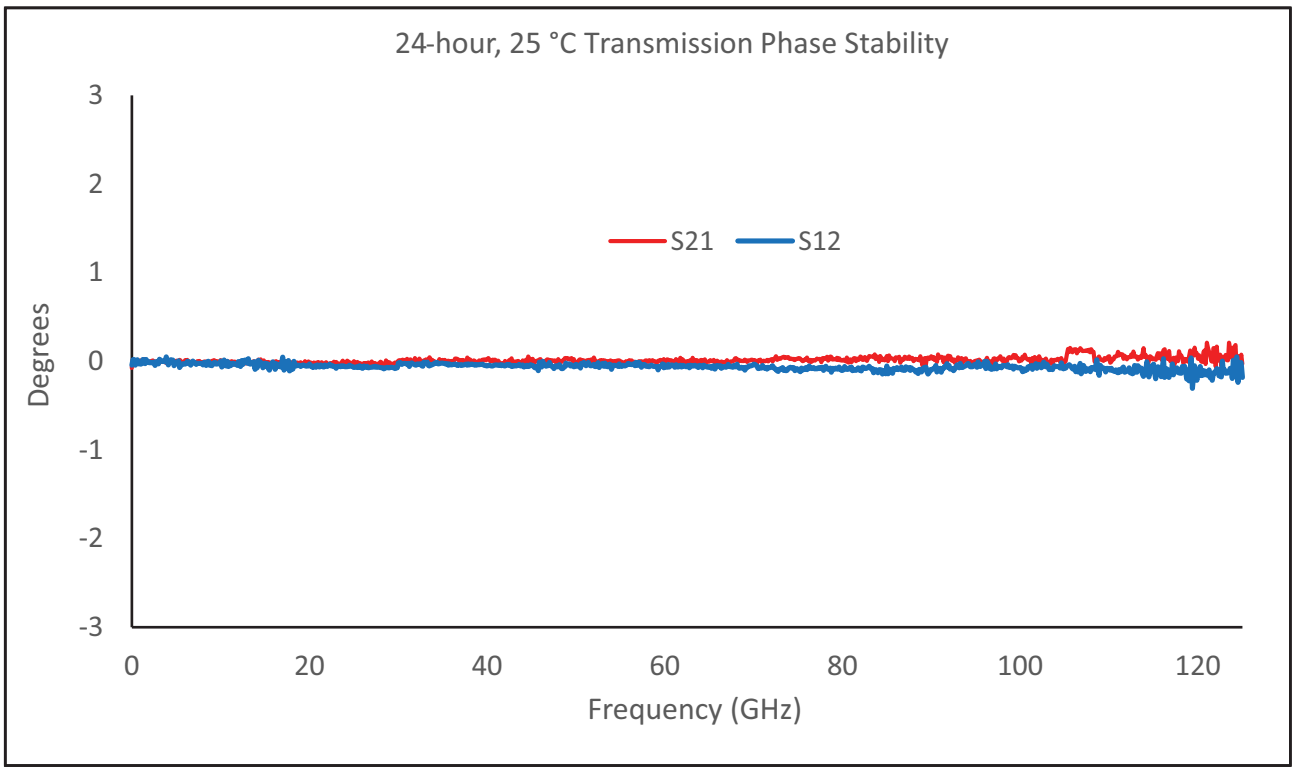


Figure 17-24. Stability Data for Modular BB System

- Waveguide-based measurements are sometimes needed. The classical approach can simply demount the mux coupler and use the exposed waveguide flange (WR-10, 75–110 GHz in that case). A variety of adapters are available for the modular BB approach in WR-15, WR-12, and WR-10 sizes (50–75, 60–90, and 75–110 GHz nominal, respectively). A bracket comes with these adapters to prevent stress from being applied to the 1 mm connector and to improve measurement repeatability. Waveguide calibration kits (3655 series) are available to support SSLT and LRL calibrations referenced to these waveguide planes. The uncertainties do not differ markedly between coaxial and waveguide-based calibrations in these bands as suggested in Figure 17-25, but the waveguide measurements can be more subject to repeatability issues depending on the physical setup and flange quality on the DUT or on any extensions (some high quality extensions are included in the calibration kits to help with the latter).

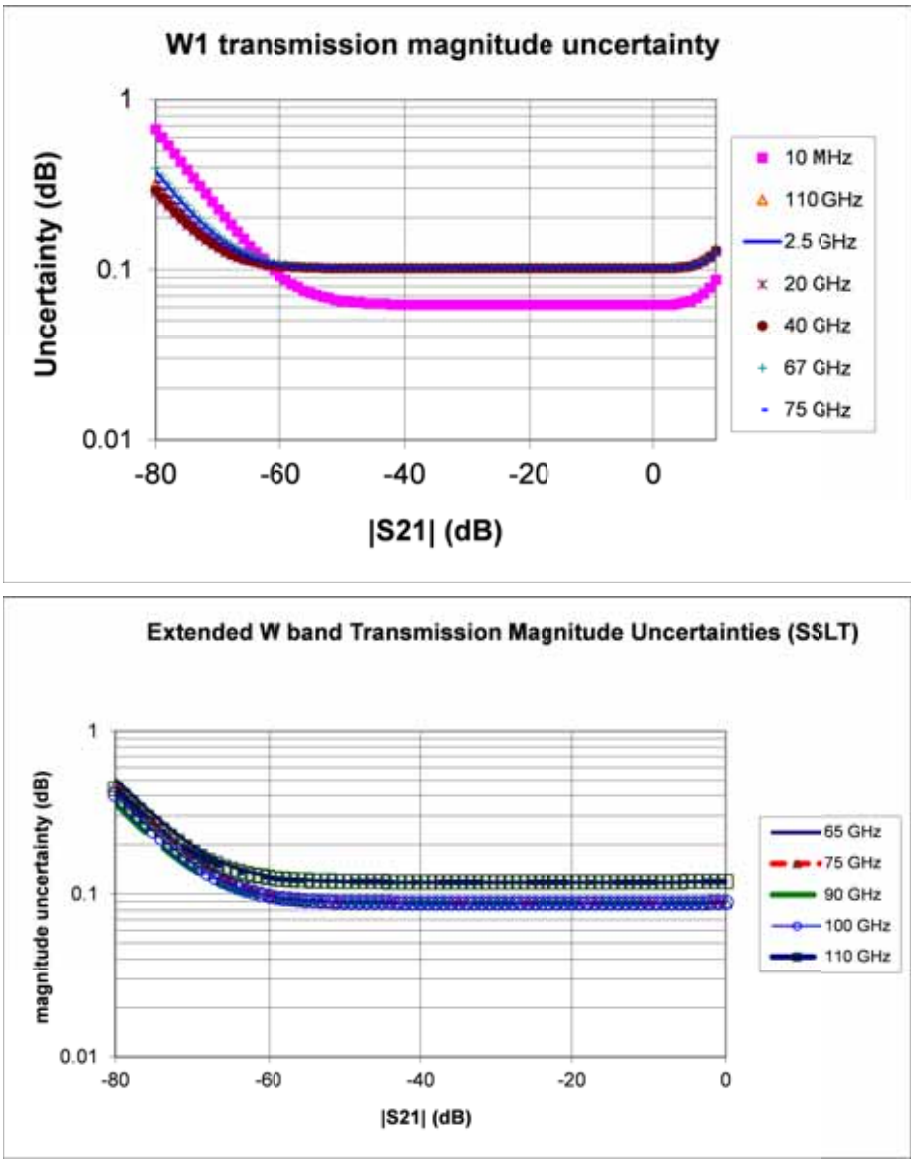


Figure 17-25. Example Coaxial and Waveguide Uncertainty Curves

- The frequency plan for the modular BB and mmWave modes was selected to optimize noise floor, spur and trace noise performance. In the True Mode Stimulus mode (available in 4-port, dual source systems; see Chapter 26 of this guide), a different frequency plan can provide improved trace noise performance at higher frequencies (at the expense of increase spurs which normally impact only low-level measurements which are less common in true mode stimulus applications). This frequency plan is available with the TMS Coherence switch (available via System | Setup | Misc. Setup menu) and should be turned ON for both the 4-port calibration phase and for the TMS measurement. If one later decides to turn this off, the 4-port calibration must be redone.

- The ME7838G system is designed primarily for on-wafer use and has a 0.6 mm coaxial flange-based interconnect for attaching probes. Power calibrations can be done with adapters (to 1 mm, 0.8mm, and WR-05 waveguide) and the power reference plane moved to the probe tip using embedding/de-embedding tools and provided generic .s2p adapter files. Other on-wafer power calibration approaches can also be used. RF user calibrations (TRL, etc.) are generally performed on-wafer using substrate-based standards.

Chapter 18 — Multiple Source Control (Option 7)

18-1 Overview

Multiple source control is an application to independently control the internal source and receiver as well as up to four external synthesizers. Since there are no constraints on frequency linkage (other than the ranges the hardware is capable of), a wide array of mixer, multiplier, converter and other specialized measurements can be performed. Some examples include:

- Mixers (up and down conversion, many conversion stages)
- Frequency multipliers
- Dividers
- Harmonic measurements (including the ability to look at fractional harmonics)
- IMD measurements
- Very high frequency measurements where the source and LO are generated externally

Since the interface is extremely flexible, this procedure also works for broadband/mmWave measurements for applications where the broadband/mmWave interface is too limiting.

There are two modes of the multiple source interface: Complete and Simple, where the default is Simple. Most of this chapter addresses all of the functionality contained in the Complete mode, and the Simple mode is a subset of that. Differences will be discussed along the way. The intent of the Simple mode is to reduce the interface to its most commonly-used elements to help simplify data entry and interaction.

At its core, Multiple Source Control is simply about pointing the sources and receivers to the frequencies of interest in as flexible way as possible. There are a number of extra choices to configure special parts of the hardware but, in the end, the basic part is defining source and receiver frequencies with a simple set of equations.

18-2 Introduction

This section discusses the interface and how to configure the instrument and the hardware for generic measurements. The MULTIPLE SOURCE menu selections are available from the primary APPLICATION menu as shown in Figure 18-1 below. The MULTIPLE SOURCE menu will vary, depending on the Control Format: Simple or Complete. In single-source systems (no Option 31), the Int. Src. Control button changes to Int. Source State, and the Int. Src. Control menu isn't there.

Navigation

Main | Application | APPLICATION | Rcvr Config | RCVR CONFIG | Multiple Source.

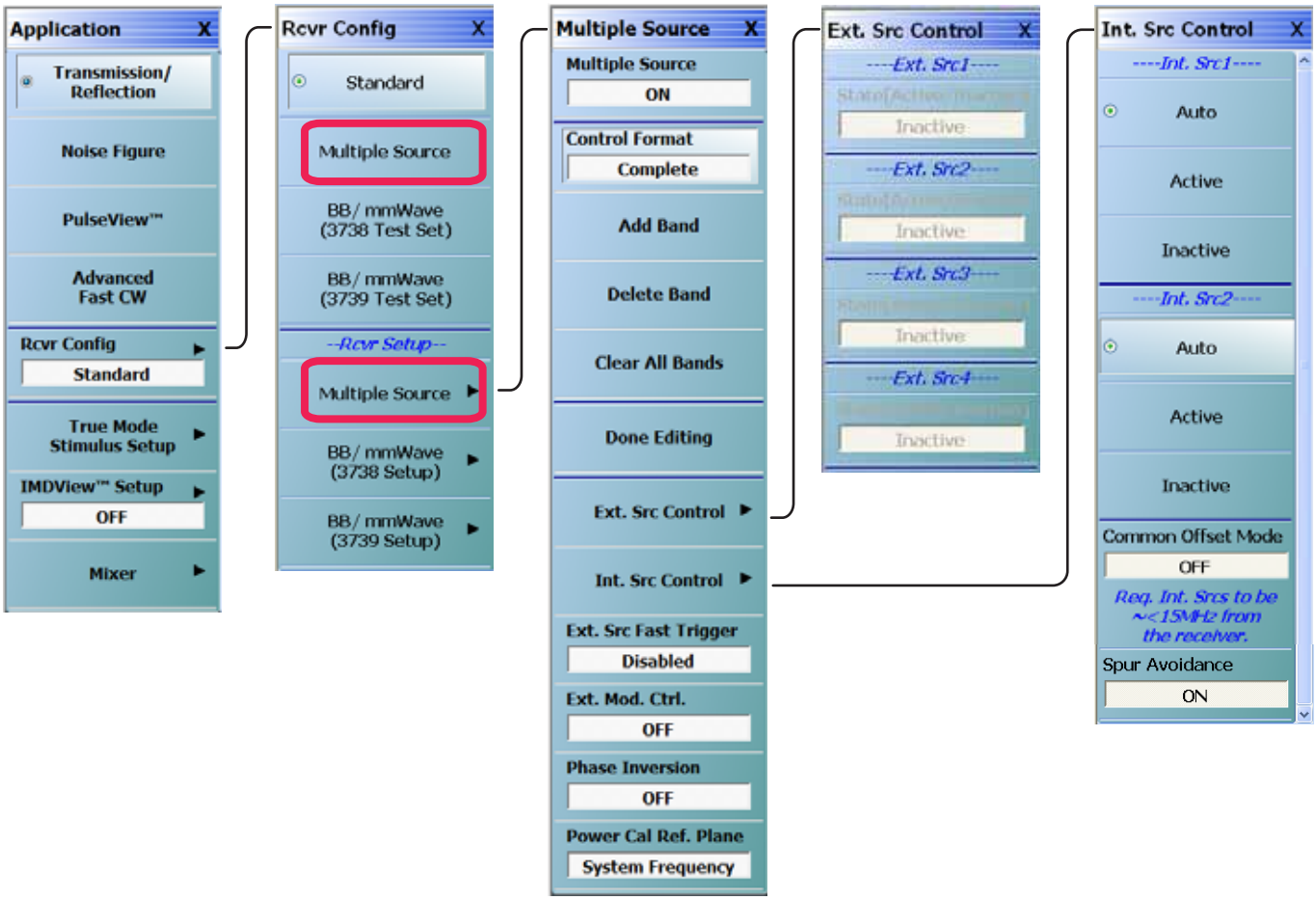


Figure 18-1. APPLICATION Menu—Multiple Source Selections (shown here for a dual source system)

18-3 Multiple Source Control Set Up

The key concepts to setting up multiple source control are:

Four External Sources for MS464xB VNAs

For MS464xB Series VNAs, up to four (4) external Anritsu sources can be configured.

Frequency Plan Bands

The frequency plan is separated into “bands”. There may be as many as 50 or as few as one. One needs multiple bands ONLY if the relationship between sources and the receiver change in an unusual way at some point in the sweep. Examples:

- A mixer measurement is being set up where the RF and LO are offset by a fixed amount and the IF (which will be sent to the VNA receiver) is constant. Only one band is needed.
- A harmonic converter is being tested where it operates on the 2^{nd} harmonic of the LO up to 10 GHz and the 3^{rd} harmonic beyond that. Two bands are needed here.
- The 'simple' mode is restricted to one band.

Linearly Linked Source Frequencies

All source frequencies (internal or external) and the receiver frequency must be linearly related. All are expressed as linear equations as a function of a runner variable “*f*”. This variable *f* is always the one displayed on the X-Axis although it need not represent an actual frequency (although for convenience it usually does).

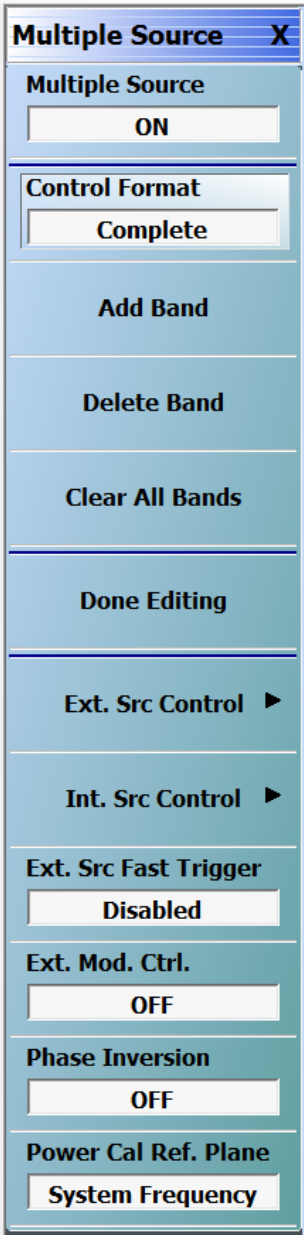
These linear relationships can change in different “bands” that the user defines. The band edges are always in terms of the runner variable *f*.

Band Start Frequency, Band Stop Frequency, and Runner Variable *f*

The band start and stop frequencies are in terms of the runner variable *f*. Thus they may or may not be physical. To reiterate, the choice is usually based on what one wants the X-Axis of the plots to be labeled in terms of, for example, the RF frequency or the IF frequency for a mixer measurement.

MULTIPLE SOURCE Menu

The main multiple source setup menu is shown below in Figure 18-2. The top half of the menu, pertaining to equation editing, will be dealt with first. The special functions of the lower half of the menu will be discussed later.



The MULTIPLE SOURCE menu appearance differs slightly depending on the instrument model number, installed options, and band edit state. In this case: **Multiple Source** is set to OFF. Toggling **Multiple Source** here also changes the **Rcvr Config** selection on the **APPLICATION** menu. **Mixer Mode** is set to ON.	**External Source Fast Trigger** set to Enabled. **External Module Ctrl.** set to OFF. **Phase Inversion** set to OFF. **Power Cal Ref. Plane** is set to SYSTEM FREQUENCY. Select toggles the Power Cal Reference Plane between System Frequency and Source Equation. For more information, see “Power Calibration Reference Plane” on page 18-31

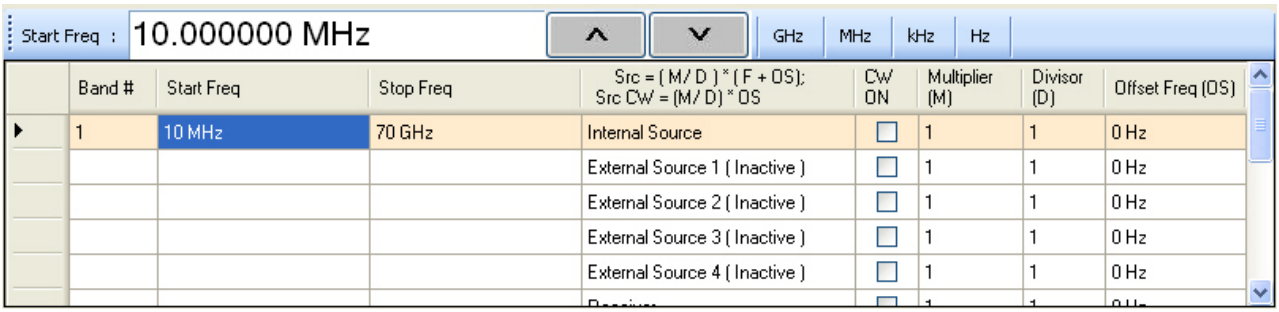
The ‘Simple’ form does not have the band elements, or some of the other features.

Figure 18-2. Main MULTIPLE SOURCE Setup Menu in its ‘Complete’ form.

The top button on the menu toggles Multiple Source mode on or off, similar to the mode selection buttons on the APPLICATION menu (turning Multiple Source mode off here will change the Receiver Config mode on the APPLICATION menu to Standard). When the MULTIPLE SOURCE menu is entered, the MULTIPLE SOURCE tableau will appear in the lower part of the screen (Figure 18-3 below).

Note	The **Done Editing** button MUST be clicked for new values to take effect. Band frequency ranges DO NOT define a required start and stop frequency. These merely set the min and max possible for start and stop. Readjust the desired range for actual measurements on the frequency menu.

When the MULTIPLE SOURCE tableau first appears, the first band is in the table. The Add Band, Delete Band, and Clear Band buttons will have the obvious effects.



The MULTIPLE SOURCE tableau dialog appearance differs slightly depending on the instrument model number and installed options.

Figure 18-3. MULTIPLE SOURCE Tableau

Defining a Multiple Source Band

A red exclamation point (**!**) in the first column of the table indicates that an error is detected.

For each band, the following must be defined:

- A start frequency for the band.
- A stop frequency for the band.
- Equations for each source, the receiver, and receiver source (an index used to work with receiver calibrations). If a source is inactive, its equation may be left at anything. If active, the result of the equation must be a valid frequency for that source (or receiver).

Each equation is of the following form:

If CW OFF is selected, then:

$$Src_X = \frac{M}{D} \cdot (f + OS)$$

If CW ON is selected, then:

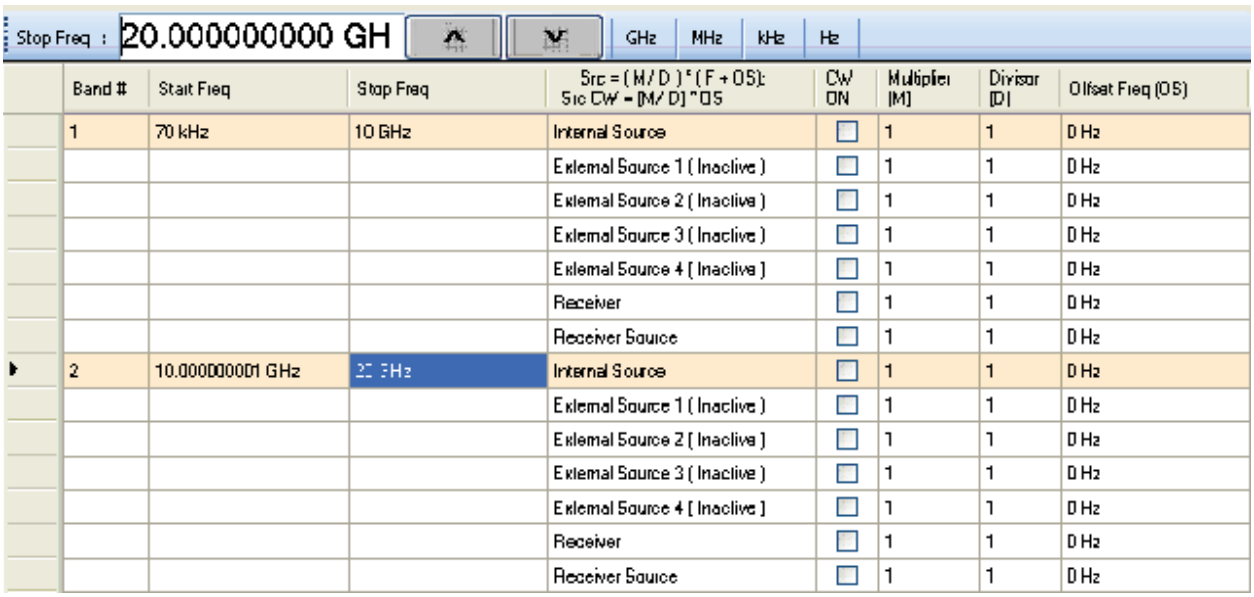
$$Src_X = \frac{M}{D} \cdot (OS)$$

Equation 18-1.

Using the multiplier (M) and divisor (D), a rational relationship can be created between the desired frequencies. The offset (OS) completes the remainder of the linear relationship and is the CW frequency when the source is set to CW ON. Any of these parameters may be negative as long as the result of the equation is a valid frequency for that source (or receiver). This can be used for a reverse sweep (in certain mixer measurements for example).

When a cell is highlighted in the table (with the mouse or touch screen), the text entry box becomes active. Text can be directly entered into the table by double-clicking on the cell. The entry must be typed with a space between the number and the frequency units.

When a second band is added, the table adds another block as shown in Figure 18-4. Note that the bands must be contiguous although the sweep range need not be (by using segmented sweep for example).



The MULTIPLE SOURCE tableau dialog appearance differs slightly depending on the instrument model number and installed options.

Figure 18-4. MULTIPLE SOURCE Tableau—Two Band Multiple Source Tableau

As soon as values are changed in the table, the Done Editing button shown in Figure 18-2 becomes active. Selecting the Done Editing button will start an error check of all entered parameters. An error dialog may appear if an error is found. During the error checking, the system applies the band limit frequencies to each equation and checks that the results are valid for a given source or the receiver. If an external source is inactive, the error checking will not be performed for that line.

Note

The **Done Editing** button in the **Multiple Source** menu MUST be clicked for new values to take effect.

Band frequency ranges DO NOT define a required start and stop frequency. These merely set the min and max possible for start and stop. Readjust the desired range for actual measurements on the frequency menu.

'Internal Source State'

The state of the internal source(s) can be altered between Active and Inactive (and Auto in the case of dual source systems). If an internal source is Inactive, the RF power will be removed (to at least the –90 dBm level) and tuned off in frequency to prevent any interference. This will be the case regardless the trace configuration (e.g., port 1 driving). Note that Inactive is essentially the same as selecting RF Power = Off on the power menu (under other setup) and either of these commands will remove RF power.

In a dual source system, Active means the source will be driving all the time (regardless the trace configuration) at the frequency indicated by the equations. When in Auto, that particular source will be driving only if the sweep pattern (as dictated by the trace configuration or RF calibrations). In a single-source system, Auto and Active concepts are the same because of the use of a transfer switch so only the Active choice is provided.

External Module Control

The EXTERNAL MODULE dialog box is accessed off the MULTIPLE SOURCE menu.

- MAIN | Application | APPLICATION | Rcvr Setup | RCVR SETUP | Multiple Source | MULTIPLE SOURCE | Ext. Mod. Ctrl | External Module Ctrl dialog box

This dialog box allows the use of a 3738x or 3739x Test Set and external broadband or millimeter-wave modules. The activation is by band and when turned on, the test set will be informed to activate its modules and use current knowledge of the driving port. This would commonly be used for broadband or mmWave measurements when the broadband/mmWave mode is too restrictive for the desired measurements. This may happen if different IFs are desired, additional frequency converters are involved in the measurement, or other unusual scenarios. The external IF ports are automatically activated in any of the non-Off modes, as are leveling and power control schemes associated with the given test set (see Chapter 17, “Broadband/mmWave Measurements (Option 7, Option 8x)” for more test set information). Only one global mode can be selected per channel but that external module selection can be activated or de-activated on a band-by-band basis.

External Module Ctrl = OFF

External mmWave or broadband test sets are not being used. The dialog in this state is shown in Figure 18-5. The 3739x test set does not require external synthesizers, so these non-OFF choices will always be available assuming Option 8x is present. The 3738 test set does not require additional options, but does require the presence of external synthesizers 1 and 2 on the GPIB bus.

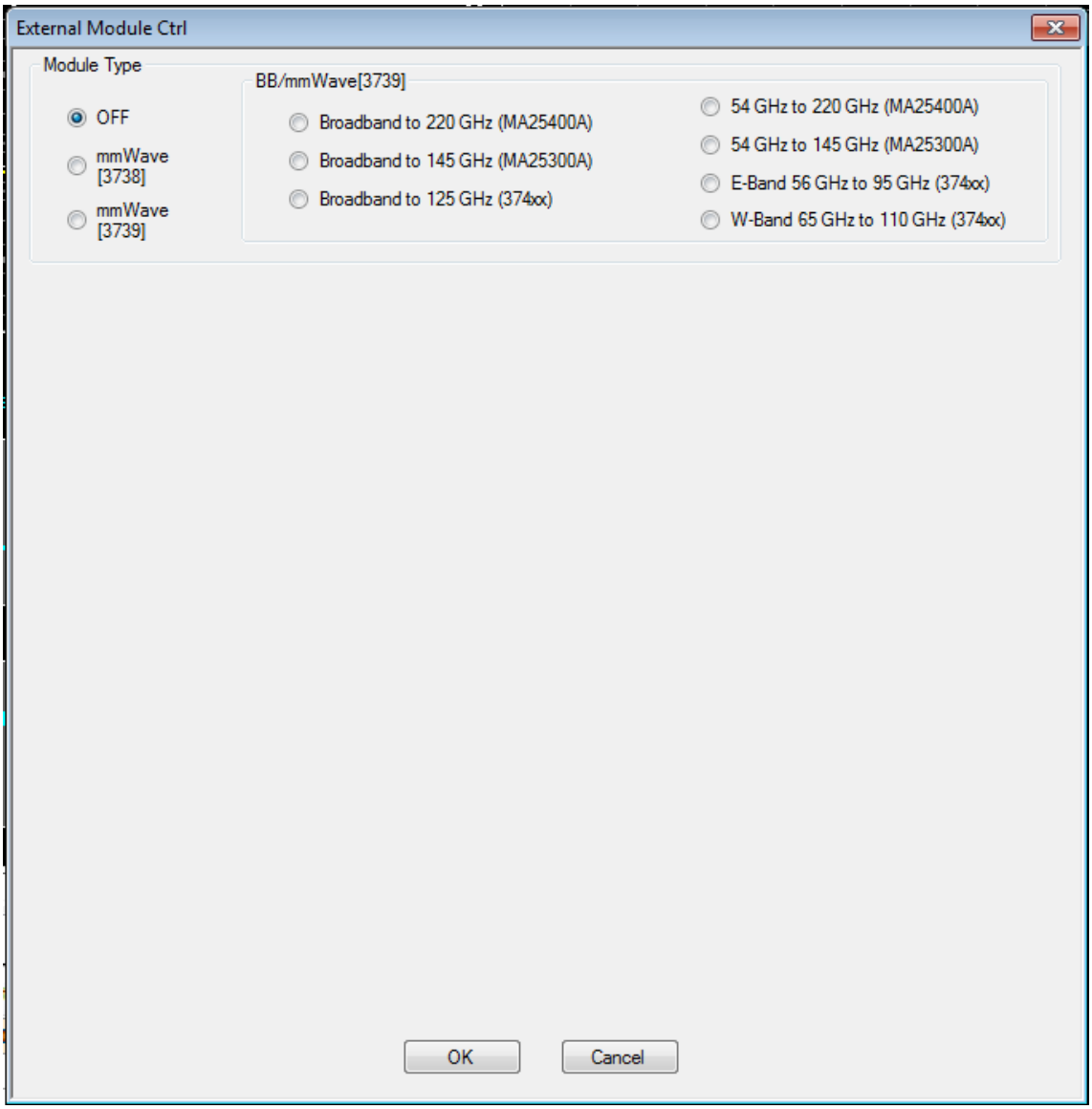


Figure 18-5. External Module Control Dialog Box with OFF Selected

External Module Ctrl = mmWave (3738)

The 3738x test set is being used with conventional OEM mmWave modules (the ME7828 family of systems). In this case, a check box will be available per-defined band to allow activation of the test set in that band. This option is only available if External Sources 1 and 2 are connected and active. There are no other parameters to select other than to activate or de-activate on a band-by-band basis.

As an example to show how this selection might be used, consider a DUT that is a downconverter taking 250 to 300 GHz down to a fixed IF of 1 GHz. The LO is sub-harmonically driven with an effective multiplier of 18 and will be driven from External Source 3 (EXT. SRC 3). The RF is supplied by an external mmWave module (WR-03 waveguide in this case), which also has an effective source multiplier of 18, and this will be driven from External Source 1 (EXT. SRC. 1). This external module also has an LO input (x20 effective multiplier) that will be driven from External Source 2 (EXT SRC 2). It is desired to measure both return loss (at the RF) of the DUT as well as its conversion loss. This will require two different multiple source setups since the conversion paths in the VNA are different.

Return Loss Measurement

Connect all external synthesizers and make sure the GPIB addresses match, and that 10 MHz clocks are synchronized (or higher frequency references if those are being used). It is desired that the DUT LO port be driven since that can affect return loss. Since the mmWave module IF outputs will be used (and routed to the VNA rear panel), Ext Module Ctrl for BB/mmWave must be enabled.

- Band 1: 250–300 GHz
- Int Src (Internal Source) = 1/1(CW 3 GHz)
- We are not using the internal source so it is just parked.
- Ext Src1 (External Source) = 1/18(f +0)
- Ext Src2 = 1/20(f+12.35 MHz)
- The system IF (rear panel) is 12.35 MHz.
- Ext Src3 = 1/18(f+ 1 GHz)
- Rcvr (Receiver) = 1/1(CW 2 GHz)
- We are not using the internal LO so it is just parked.

Conversion Loss Measurement

In this case, Ext Module Ctrl must be OFF since we will not be using the rear panel IFs. Note that this also disables the test set so some care is required with RF signal cabling.

- Band 1: 250 to 300 GHz
- Int Src = 1/1(CW 3 GHz)
- We are not using the internal source so it is just parked.
- Ext Src1 = 1/18(f +0)
- Ext Src2 = 1/1 (CW 2 GHz)
- The external mmWave module LO is not being used.
- Ext Src3 = 1/18(f+ 1 GHz)
- Rcvr = 1/1(CW 1 GHz)
- The DUT IF is routed to a port or receiver loop so it can be converted.

External Module Ctrl = mmWave (3739)

The mmWave (3739) selection also assumes the use of OEM mmWave modules, but with the 3739x test set. Details on some applicable modules, from a hardware perspective, are discussed in Chapter 17. Any multiplied transceivers can be used here as long as the frequency and power plans are consistent with the VectorStar system with the 3739x test set. The test set offers additional power control options and assumes the use of the internal VNA sources so there are some additional selection options as suggested by Figure 18-6. Note that as additional bands are added to the multiple source definition, the band check boxes must be manually checked for each in this EXTERNAL MODULE CTRL dialog.

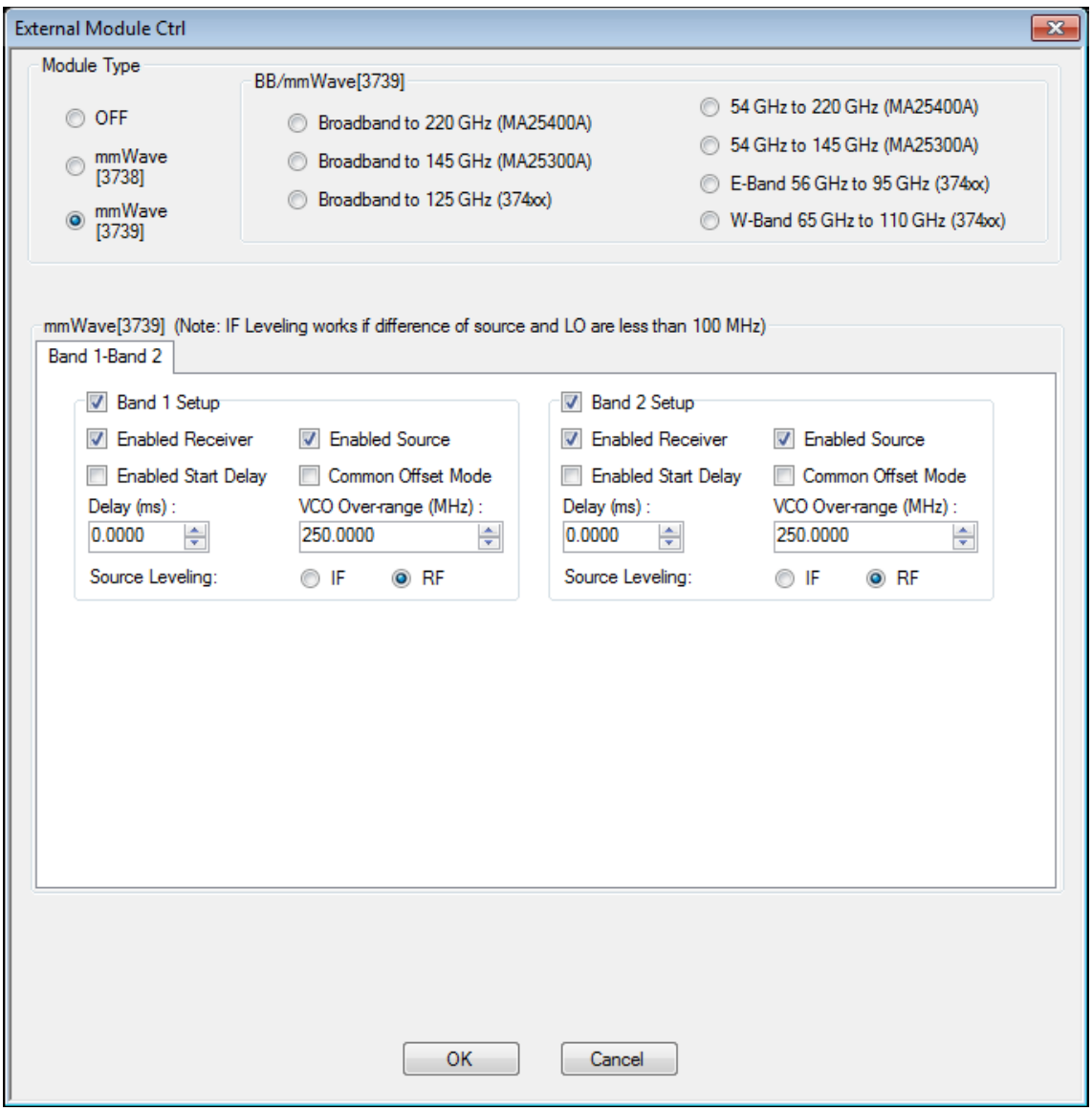


Figure 18-6. External Module Control Dialog as Configured for a Single Active Band of mmWave (3739)

The selection choices per band are described below:

- **Enabled Receiver:** Use the receiver in the remote modules. Above this breakpoint, the VNA system LO will be set appropriately in terms of frequency and power, the test set configured, and the VNA's rear panel IFs will be activated.
- **Enabled Source:** Use the source multipliers in the remote modules. When enabled the VNA synthesizers will be set appropriately, the test set configured as needed, and ALC leveling prepared (see below).

- **Enabled Start Delay:** This enables a fixed delay at the beginning of the band. This could be useful for certain slow settling DUT measurements or for very low power levels.
- **Common Offset Mode:** At higher multiples (generally for modules running over 300 GHz), it may be desired to improve source correlation to reduce trace noise. This can be done with the Common Offset mode bit (and it is automatically done in the 3739 mmWave modes discussed in Chapter 17). If the source and receiver frequencies (when reduced to the 2.5–5 GHz range by division) differ by more than about 50 MHz with this bit selected, phase lock errors may occur so this selection is not appropriate for mmWave mixer measurements but is useful for IMD and other related measurements.
- **VCO Over range:** Normally, the internal VNA multipliers switch at 5, 10, 20, and 40 GHz. In some mmWave modules operating with high multipliers, it may be desirable to push those switch points out further (up to 5.5, 11, 22, and 44 GHz typical, but not guaranteed). This overrange (expressed in MHz relative to the 5 GHz breakpoint) sets the new breakpoints.
- **Source Leveling:** The leveling choice determines which detection path is used (RF implies the VNA detection on the RF drive path, IF uses the test set detection on a reference IF). Additional cals using the mmWave ALC subsystem may be required. Generally, IF leveling is only valid if the IF coming from the mmWave modules is under 100 MHz.

External Module Ctrl= BB/mmWave (3739)

For use with 374xx and MA25300A/MA25400A modular components. This right side of the radio button array is for the broadband operation and for banded operation. The selection for BB/mmWave is straightforward. When checked in a given band, the test set will be activated, the VNA's internal transfer switch will be shut down, and the rear panel IF ports on the VNA will be activated. There are choices for broadband and banded operation. The distinction between these selections is based on the imposed frequency range limits when activated. Broadband allows from the lower instrument limit up to the upper limit of the system (110 GHz, 125 GHz, 145 GHz (150 GHz operational), or 220 GHz (226 GHz operational) for ME7838E/EX, A/AX, D, and G systems respectively), E band allows 56–95 GHz (under-range to 54.000000001 GHz allowed) using 3743x modules, and W band allows 65 GHz to 110 GHz using 3743x modules. For systems loaded with Option 80/081/084/085, three broadband buttons will be available for 125 GHz, 145 GHz (operational to 150 GHz), or 220 GHz (operational to 226 GHz) situations. Not all of the selections make sense depending on which modules are in use (e.g., if 3743A/AX modules are being used, only the first selection is useful). For ME7838E/EX systems (using Options 86/87) only operation to 110 GHz is allowed. Otherwise, the selections and instrument operation are the same. The dialog in the case of broadband set for two bands is shown in Figure 18-7. Note that as additional bands are added to the multiple source definition, the band checkboxes in this External Module Ctrl dialog must be manually checked.

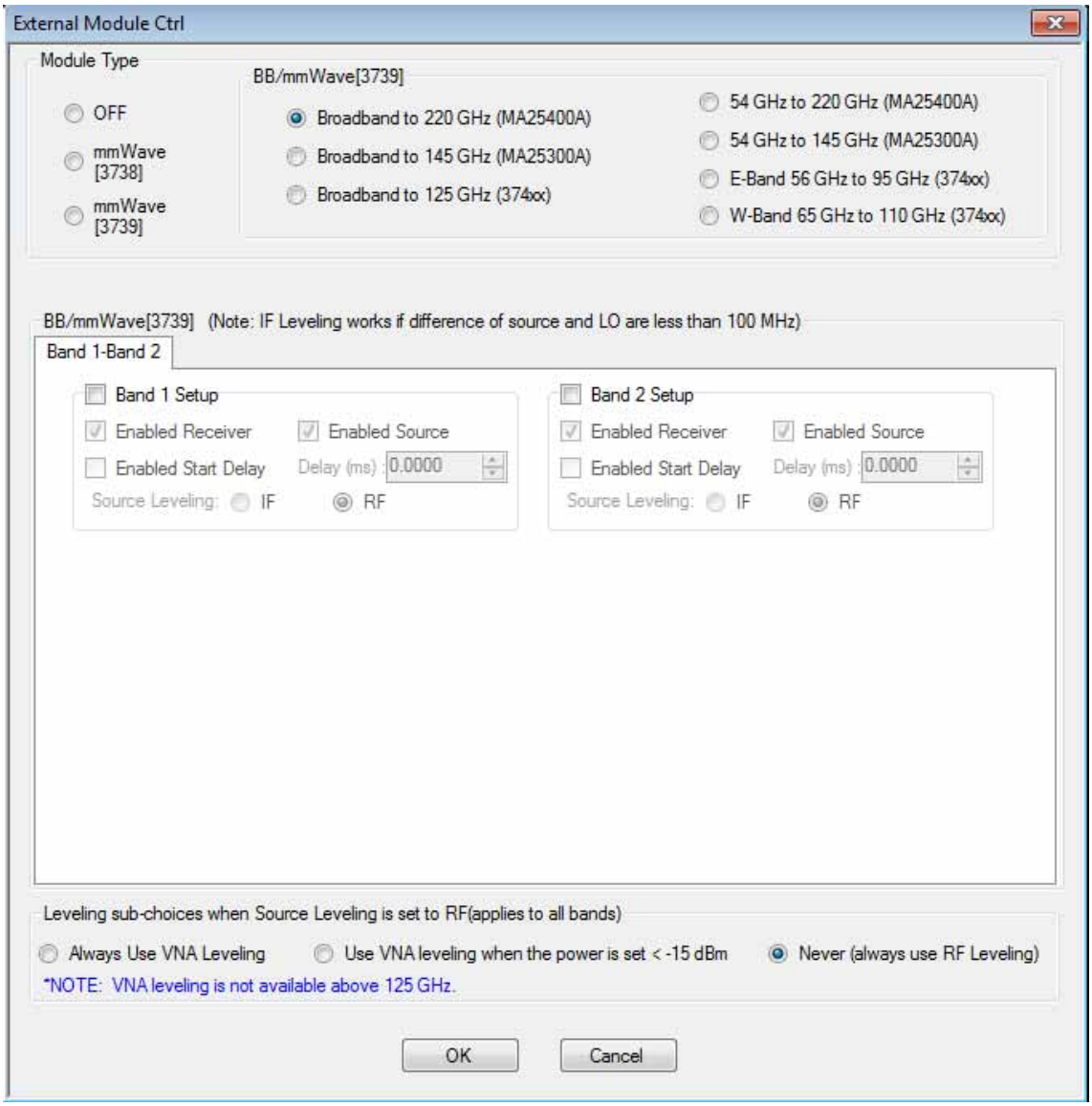
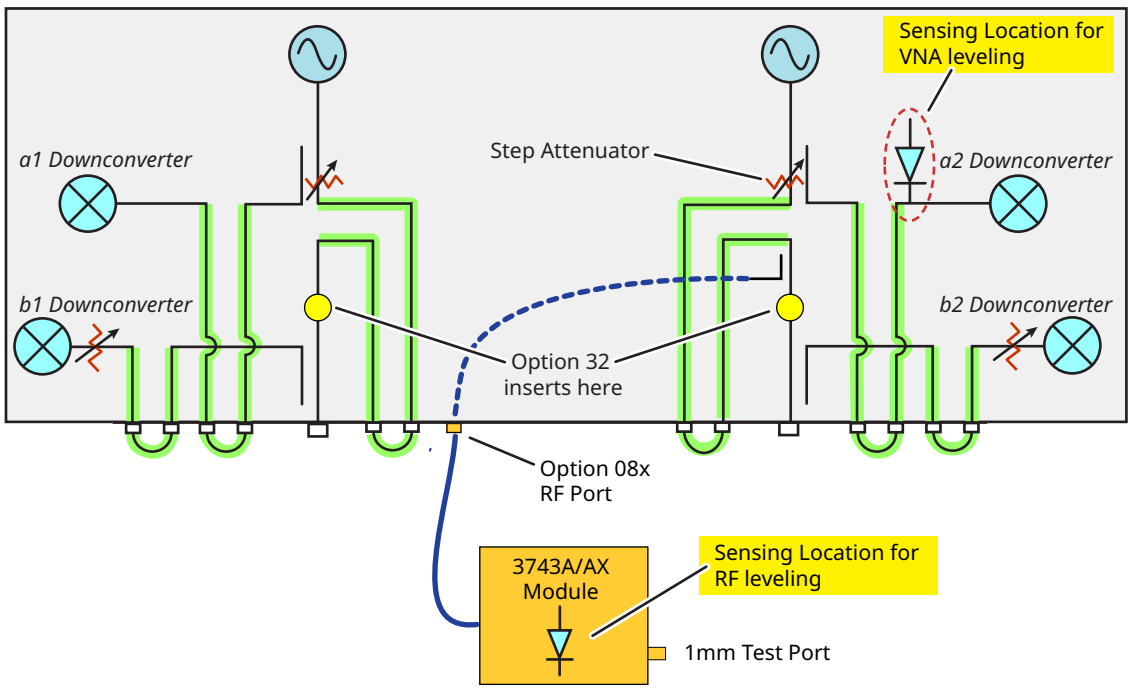


Figure 18-7. External Module Ctrl Broadband Mode (3739x-based) with Two Bands Configured

These MA25x00A, 3743x, and 3744x modules have independent source and receiver paths that can be selected from the above dialog. The selections can be interpreted as follows:

- **Enabled Receiver:** Use the receiver in the remote modules above the receiver breakpoint (30 GHz). Above this breakpoint, the VNA system LO will be set appropriately in terms of frequency and power, the test set configured, and the VNA's rear panel IFs will be activated.
- **Enabled Source:** Use the source multipliers in the remote modules above the source breakpoints (54, 80, and 120 GHz for x2, x3, and x4 respectively for 3743A/AX modules; 54 and 80 GHz for x2 and x3 for 3743E/EX modules; 54, 80, and 110 GHz for x2, x3, x6 for MA25300A modules); 54, 80, 110, and 150 GHz for x2, x3, x6 and x12 for MA25400A modules. Above the first breakpoint, the VNA synthesizers will be set appropriately, the test set configured as needed, and ALC leveling prepared (see below). Note that the above and below 54 GHz power control settings apply when this source feature is enabled.
- **Enabled Start Delay:** This enables a fixed delay at the beginning of the band. This could be useful for certain slow settling DUT measurements or for very low power levels.
- **Source Leveling:** The default operation of the modular BB system (when not in multiple source) is to use IF leveling in order to get wide power ranges, but if the source and receiver frequencies are not close enough (or the receiver is not enabled), then an IF is not available for leveling. The leveling defaults to RF in multiple source modular BB and this is recommended unless one knows the IF will be available (up to about 100 MHz) in the module for leveling. Separate ALC calibrations are available for RF and IF leveling and the system will automatically index the correct calibration table.

  For RF leveling, there is an additional global choice (for dual-source, Option 31 instruments) for all bands < 125 GHz and with RF leveling selected: RF or VNA plane selection. The RF selection uses the detector in the mmWave module which has stability advantages since it is later in the chain but has limited dynamic range (–25 dBm to –20 dBm lower limit commonly). The VNA selection uses a detector in the host VNA which has the advantage of improved dynamic range (can usually reach –35 dBm to –40 dBm) but slightly reduced stability since the detection plane is behind some gain stages. Another selection is to automatically use VNA leveling for requested power levels below –15 dBm and RF leveling for higher levels. The default is RF leveling. The location of the leveling positions is sketched in Figure 18-8. Note that the VNA leveling sub-choice is not available for single-source instruments.



The RF and VNA leveling locations when in the broadband/mmWave (3739) mode are shown here for the case of a dual-source (Option 31) instrument. VNA leveling is not available for this mode in single-source instruments.

Figure 18-8. RF and VNA Leveling Locations

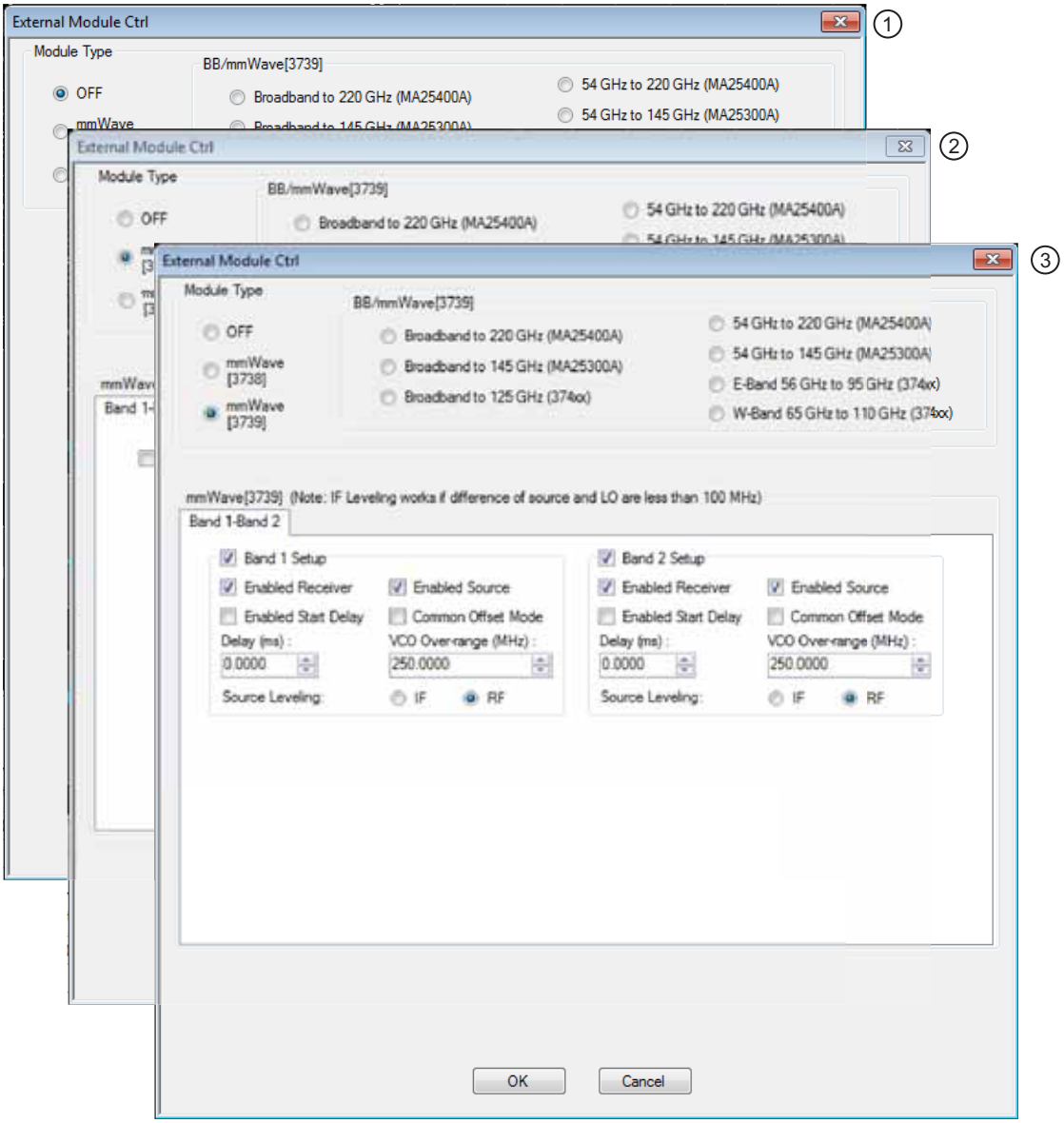
Figure 18-9 on page 18-15 shows the EXTERNAL MODULE CTRL dialog box variants for OFF, mmWave 3738, and mmWave 3739. In Figure 18-10 on page 18-16, examples are shown for Broadband, and Figure 18-11 on page 18-17 shows the variants for E-Band 56 GHz to 95 GHz and W-Band 65 GHz to 110 GHz.

In the BB/mmWave receiver configuration of multiple source when the instrument has two sources, there are a number of additional measurement combinations possible including manual IMD (exclusive of Option 44), mixer measurements, de-sense measurements, and others. There are, however, some intricacies that warrant further discussion.

Key Point: In dual source systems operating as 2-port instruments, source 2 always drives the mmWave test set (true for pure mmWave 3739 modes as well).

As such, source 1 is the free source to act as drive for a 3rd module, to act as a subharmonic LO, etc. The source 2 equation state value will determine the test set state (assuming the corresponding box has been checked) and >54 GHz power control values will affect the 3739 test set drive for BB/mmWave when in the appropriate frequency range. The port 1 <54 GHz power control field will always affect source 1 drive (available from the VNA port 1 or the access loops if so equipped) and the label for that power field will change (Src1 to Aux Module). Note also that both sources must manually be made active (under Int. Src. Control) if it is desired that they drive simultaneously.

When both sources of Option 31 are active in a mmWave configuration (using this auxiliary module for source 1), power calibrations will be more complex. When in this state (multiple source with both sources active, one of the **mmWave Ext. Mod. Ctrl.** buttons clicked and in a mmWave frequency range), the power menu will have 'Src1 driving aux module' for the first field and the power calibration selector will have three choices (Port 1, Port 2 and Aux Module) in a 2-port system. The Target Power and other concepts are similar in all cases except the Aux Module calibration does not have a factory calibration as a starting point. At this stage the power associated with the Aux module is referencing the Port 1 test port of the baseband VNA. Because the output power of the module has a large offset from the VNA test port power it may take longer to converge and one may have to adjust the main power menu setting to get closer to the desired target. When one of these power calibrations is performed, the unindicated source will be deactivated so that unindicated source will not interfere with the calibration (if it, for example, was connected into the power measurement path via a combiner that was used for an intermodulation distortion measurement). After the calibration is complete, the sources are returned to their desired driving state.



1. Dialog box with Module Type OFF selected.
2. Dialog box with Module Type mmWave (3738) selected. Selection boxes enable Band 1 and/or 2.
3. Dialog box with Module Type mmWave (3739) selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2.

Figure 18-9. EXTERNAL MODULE CTRL Dialog Box Variants—OFF, mmWave 3738, and mmWave 3739

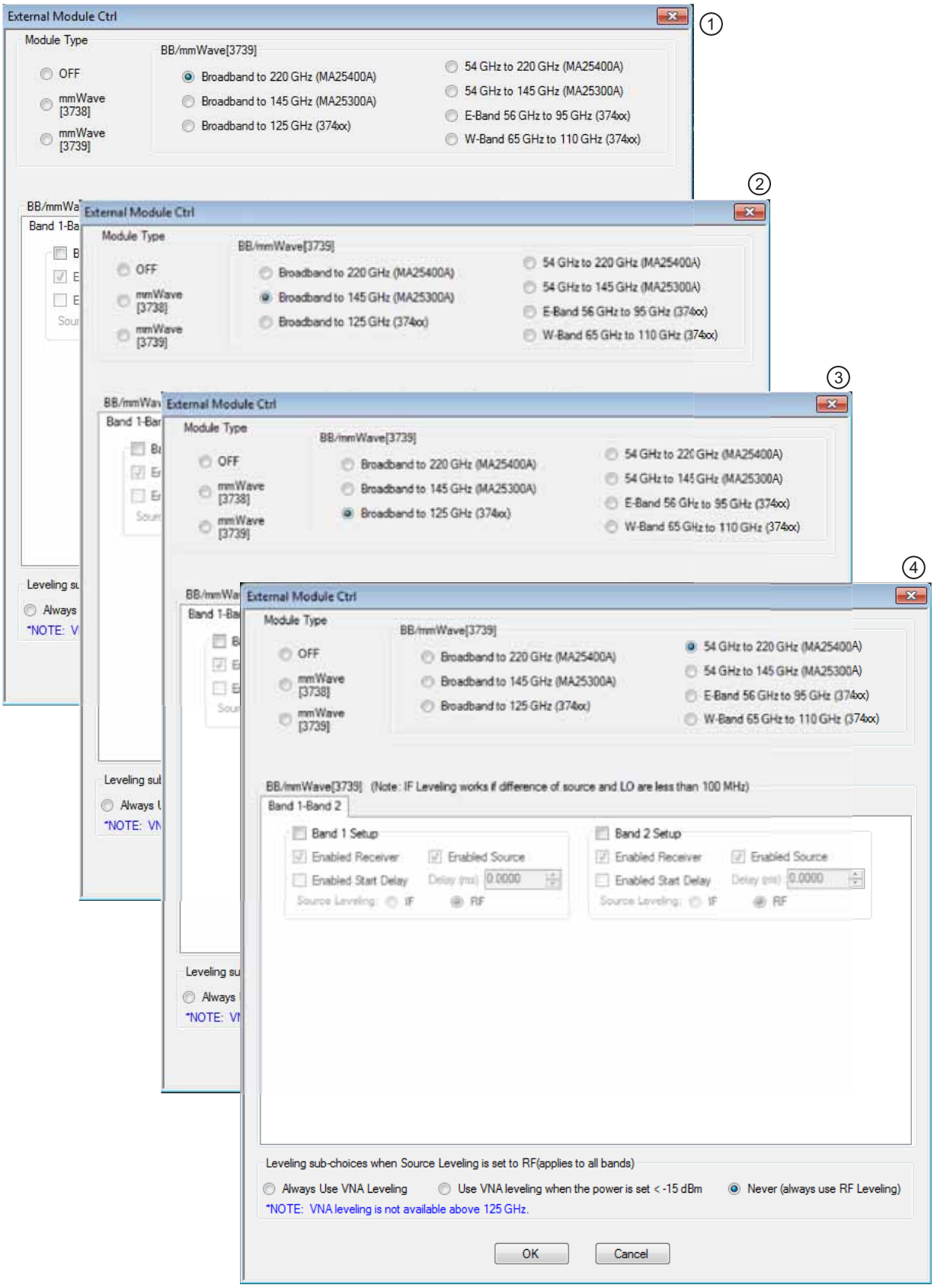


Figure 18-10. EXTERNAL MODULE CTRL Dialog Box—Broadband (1 of 2)

1. Dialog box with Module Type Broadband to 220 GHz selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2. This module type and mode only available with MS4647B VNAs equipped with Option 8x.
2. Dialog box with Module Type Broadband to 145 GHz selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2. This module type and mode only available with MS4647B VNAs equipped with Option 80/81.
3. Dialog box with Module Type Broadband to 125 GHz selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2. This module type and mode only available with MS4647B VNAs equipped with Option 80/81.
4. Dialog box with Module Type 54 GHz to 220 GHz selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2. This module type and mode is available with MS464xB VNAs equipped with Option 8x.

Figure 18-10. EXTERNAL MODULE CTRL Dialog Box—Broadband (2 of 2)

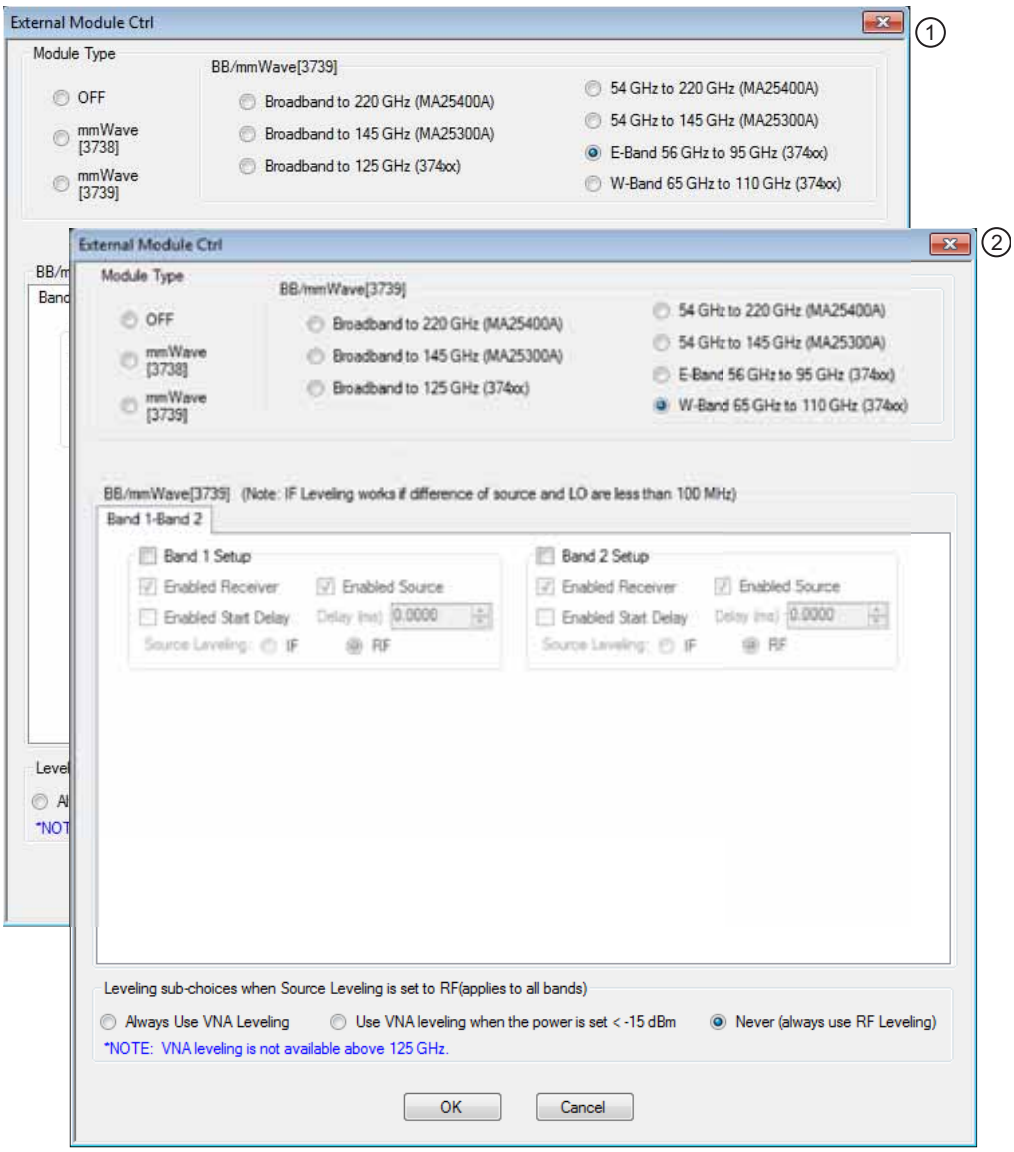
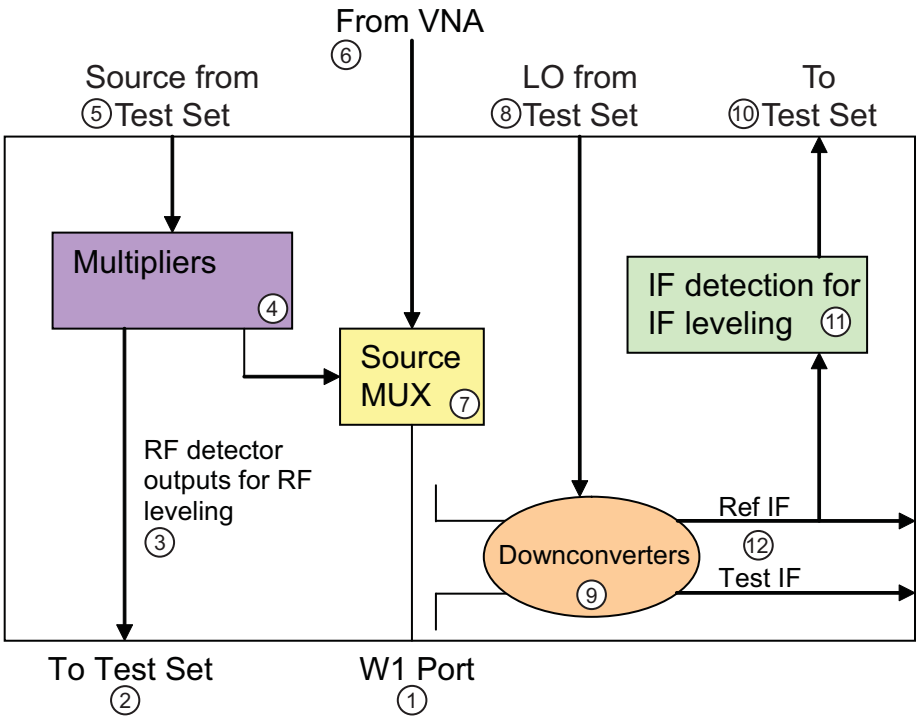


Figure 18-11. EXTERNAL MODULE CTRL Dialog Box—E-Band and W-Band

1. Dialog box with Module Type E-Band 56 GHz to 95 GHz selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2. This module type and mode only available with VNAs equipped with Options 80/81 or 82/83.	2. Dialog box with Module Type W-Band 65 GHz to 110 GHz selected. Selection boxes and control fields enable configuration control of Band 1 and/or Band 2. This module type and mode only available with VNAs equipped with Option 80/81 or 82/83.

Figure 18-11. EXTERNAL MODULE CTRL Dialog Box—E-Band and W-Band

The module block diagram below in Figure 18-12 illustrates where different leveling detection points are located.



1. W1 Test Port	7. Source MUX
2. To Test Set	8. LO from Test Set
3. RF detector outputs for RF Leveling	9. Down converters
4. Multipliers	10. To Test Set
5. Source from Test Set	11. IF Detection for IF Leveling
6. From VNA	12. Reference IF and Test IF

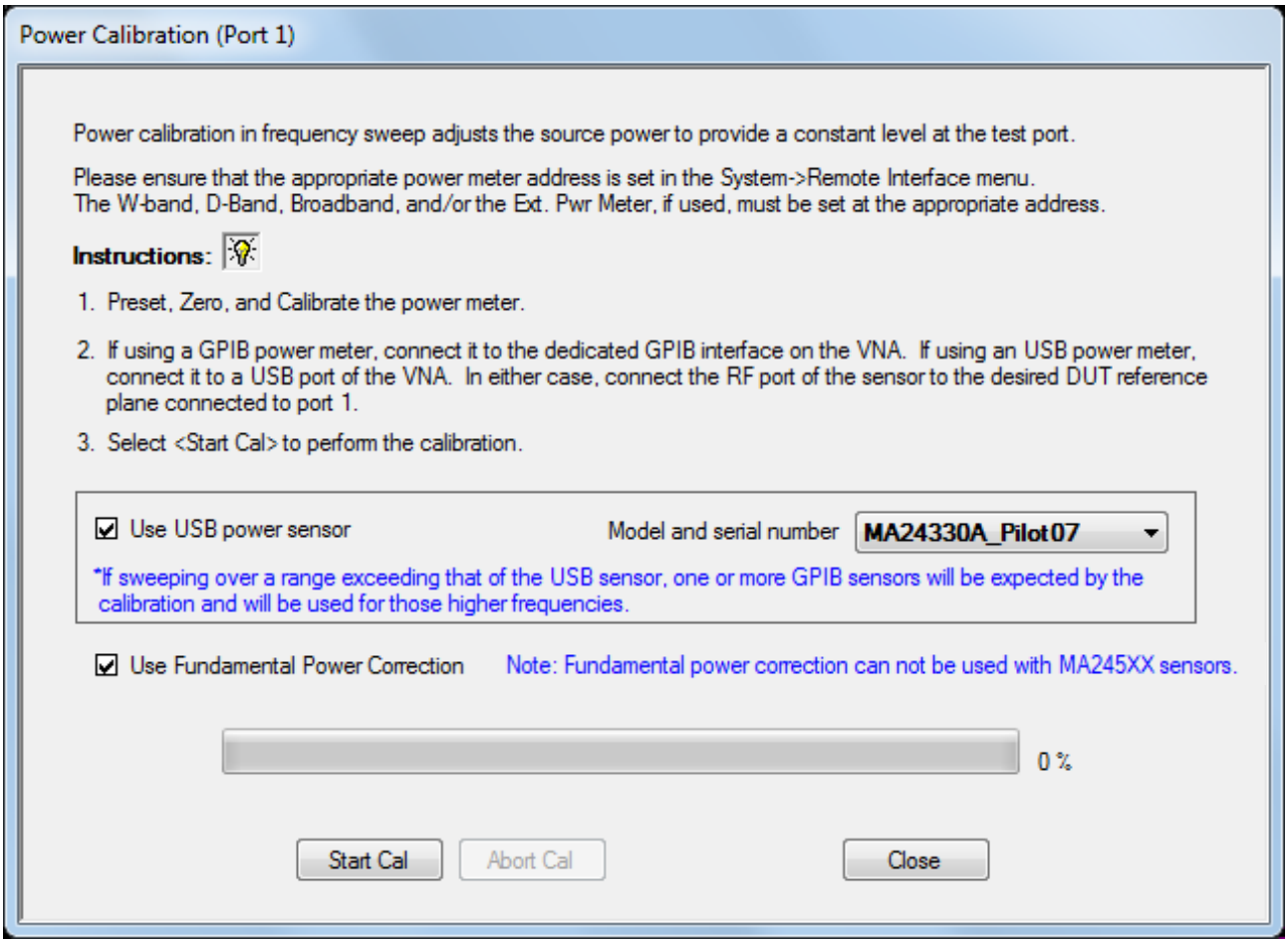
Figure 18-12. MA25x00A or 3743x Millimeter-Wave Module—Block Diagram of Leveling Options

Some of these choices can be made clearer with the following measurement examples in Section 18-4 and Section 18-5.

Fundamental Power Correction

When in a broadband or mmWave version of modular 3739 modes, power calibration can be an important topic as these modes are frequently used for pseudo-absolute power measurements such a multiplier or mixer conversion, harmonics or IMD. Since the mmWave module can only have limited internal filtering of its source chain, the internal source harmonics can sometimes be relatively high (on the order of –10 dBc at some frequencies). Since the power meter used for the power calibrations can be sensitive to this harmonic energy but the tuned response of the VNA is not, there can be a disconnect between integrated and fundamental power that can impact measurement accuracy.

A function, known as fundamental power correction, is available as part of the user power calibrations in these modes that adjusts the power calibration to take into account these differences related to spectral composition of the source signal. When enabled (by the checkbox as shown in Figure 18-13; this checkbox is only visible when in modular mmWave/Broadband modes), the user power calibration appears to proceed the same as always but, behind the scenes, the system receiver is used to estimate source harmonic levels (at each calibration frequency and/or power) and adjust the target power to take those contributions into account. When the calibration is finished, if one goes back with the power meter to check readings, they may appear high (e.g., the target value entered was –10 dBm but the power meter now reads –9 dBm at a calibration frequency) but this is because the calibration is now trying to get fundamental power to the target rather than integrated power.



Appears only in modular broadband or banded mmWave mode.

Figure 18-13. Fundamental Power Correction Checkbox in Power Calibration Dialog

Because of the extensive bookkeeping of receiver frequency plans needed for this process, the fundamental power correction is only available if the multiple source definition includes only one band. The correction only applies when the module is sourcing power (above the sourcing breakpoint frequency which is 54 GHz by default but can be changed with the configuration file as described elsewhere).

As an example, consider a power calibration over the range 81 GHz to 86 GHz and with a target power of –15 dBm using 3739 modular broadband mode. If one performs a standard power calibration, the power meter, after calibration, will generally measure quite close to –15 dBm (subject to drift, connector repeatability, etc.). In this case, however, source harmonics ranged from about –20 dBc at the edges of the range to about –15 dBc in the middle of the range. This leads the fundamental power (using the basic power calibration) being about 0.9 dB to 1.3 dB lower than the target value. By applying the fundamental power correction, the fundamental power is closer to target (which will make the VNA measurements more accurate) but the power meter will read a slightly higher value since it is measuring the fundamental plus the harmonics. The plot in Figure 18-13 shows the difference in power meter readings after the calibration for the two different calibration choices. The fundamental power correction adjusts the integrated power target higher to get the fundamental power closer to the desired target value. In this case example, the target was –15 dBm and the harmonics were in the –15 dBc to –20 dBc range.

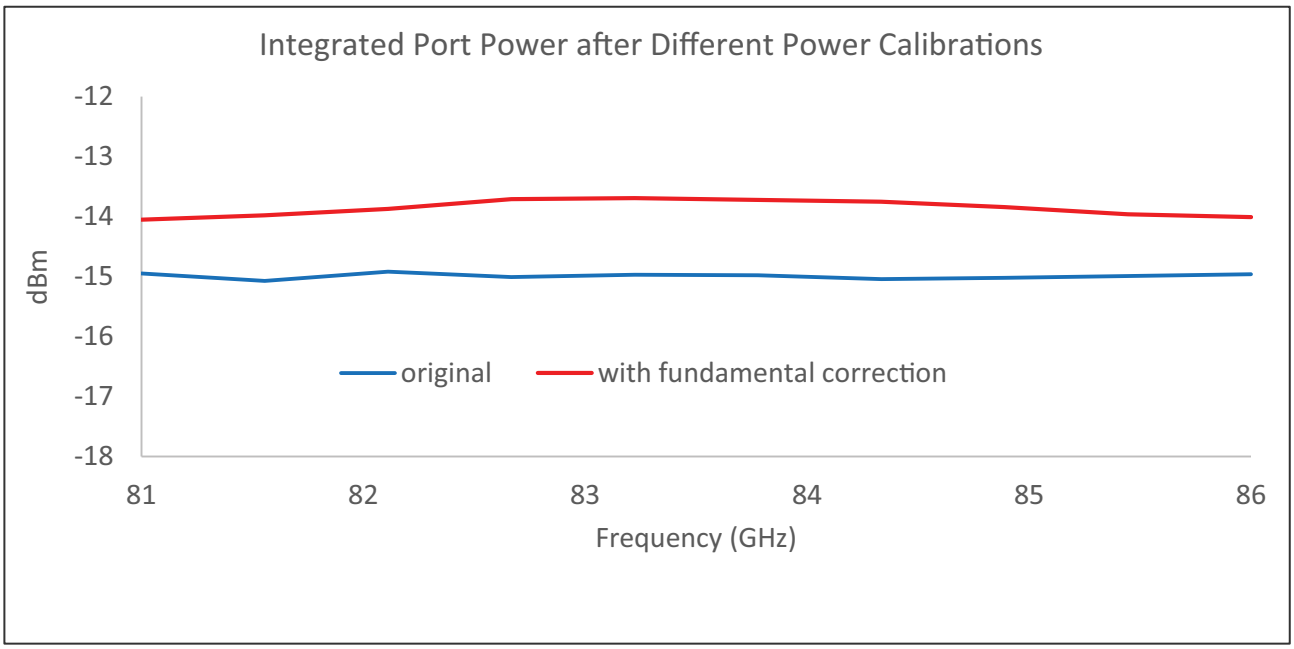
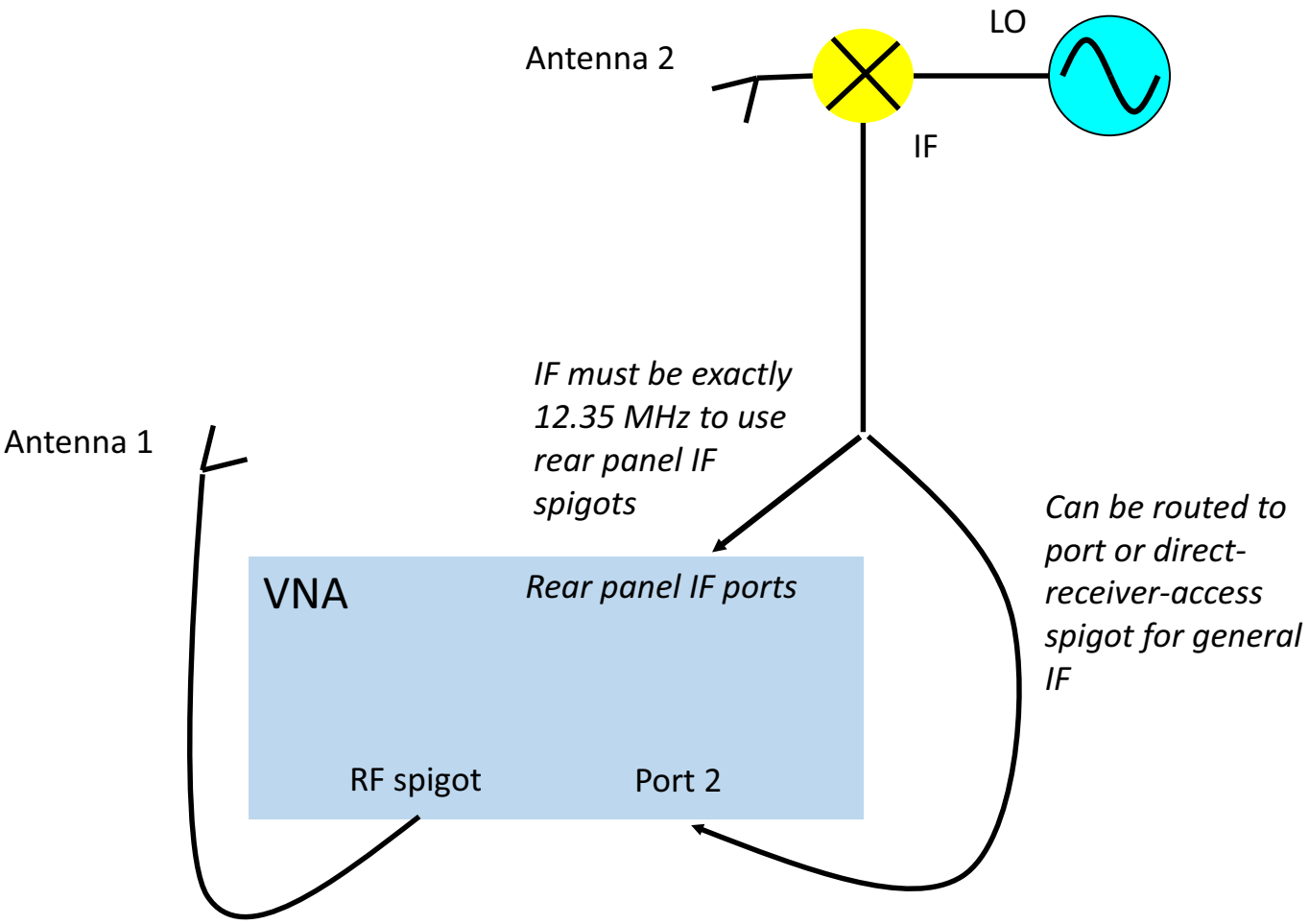


Figure 18-14. Integrated Port Power Calibration Comparison—with and without Fundamental Power Correction

External Module Control and Antenna Applications

Although the labeling may not be obvious, External module control= mmWave (3739) can also be used for a variety of remote antenna measurement applications even if in the microwave band. In these applications, an external downconverter (often at a measuring antenna mount) is used to allow the measured signal to be sent long distances back to the VNA at a lower frequency. This structure is pictured in Figure 18-15. The IF of this remote downconversion process can be at any VNA-accessible frequency and it can be routed to a VNA port (or a receiver direct access port on instruments with Option 51, 61, or 62) for final downconversion by the VNA. In this case, External Module Control is NOT used. The LO for the external conversion can be from an external synthesizer or routed from the VNA (if Option 31 is installed-the second source) and can be configured using standard multiple source control. Also the receiver equation can be set to look at the external IF frequency generated by the remote downconverter. If the internal source(s) is(are) used, they are available from the standard ports (or direct loop access, if so equipped) and Option 8x is not required.



Downconversion is done at the receiving antenna (which may be far away) to reduce the effect of cable losses. The IF from this downconversion can be routed to a VNA port or direct-access-receiver port in the general case (where the VNA will again downconvert to reach its internal IF) or could be routed to the rear panel IF spigots if already at the internal VNA IF of 12.35 MHz. For the latter case, external module control (and Option 8x) is required.

Figure 18-15. Example: Remote Antenna Measurement

Alternatively, if the remote downconverter can be set to produce an IF at 12.35 MHz (the VNA internal IF in most cases) and the instrument has Option 8x, the remote IF can be routed directly to the rear panel IF ports that are normally used for the mmWave applications discussed in previous sections. Such a direct IF injection can provide noise advantages since an internal downconversion is omitted.

The mmWave (3739) External Module Control selection is important in that it enables the rear panel IF ports while imposing minimal other constraints on the frequency plan (unlike the broadband and E/W-band selections on that dialog). The internal sources can be used up to 40 GHz (for a MS4644B or MS4647B, 20 GHz for a MS4642B) while external sources can be controlled over their valid frequency ranges. The VNA source power is routed to the RF spigot on the front panel associated with Option 08x (RF1 and RF2 for dual-source systems (those with Option 031)). For the dual-source systems, note that the RF spigot being driven will be controlled by which source is active and both can be made active within multiple source control as discussed elsewhere in this chapter. It is also important to note that these RF spigots are coupled off of the main source drive paths so the power levels are 10–15 dB below that indicated on the power menu.

The Receiver equation can be set in this direct IF mode, but because the rear panel IF spigots are being used, the internal receivers are not active. The LO spigots on the front panel are active and the signal present there can be used, but note that the internal LOs are capped at 10 GHz so the receiver equation must be set appropriately. As an example, suppose one wanted to use the internal LO to act as sub-harmonic (1/3) drive for a remote converter and suppose the RF to be converted is at 28 GHz. We wish a 12.35 MHz IF, so the final LO (internal to the remote converter) should be 28.01235 GHz assuming a high-side LO. The sub-harmonic drive should be one-third of that or 9.33745 GHz. The multiple source system adds 12.35 MHz to the equation value so the equation should be reduced by the same amount to compensate. Thus, the receiver equation entry should yield 9.3251 GHz. An example setup is shown in Figure 18-16 that accomplishes this measurement over a finite range. Note the –24.7 MHz used in the receiver equation. For a simple harmonic conversion using a factor of N, this number will generally be –(N–1)*12.35 MHz. If multiple conversions are used or there is a more complex IF plan, the equation will differ. The universal point is that the LO will be programmed to 12.35 MHz above the output value of the equation.

Note

When using a mm-wave module with an LO multiplier of N:

- Set receiver (and receiver source) multiplier to 1
- Set receiver (and receiver source) divisor to N
- Set receiver (and receiver source) offset to –(N-1)*12.35MHz (i.e., for an N of 24, the offset would be -284.05 MHz)

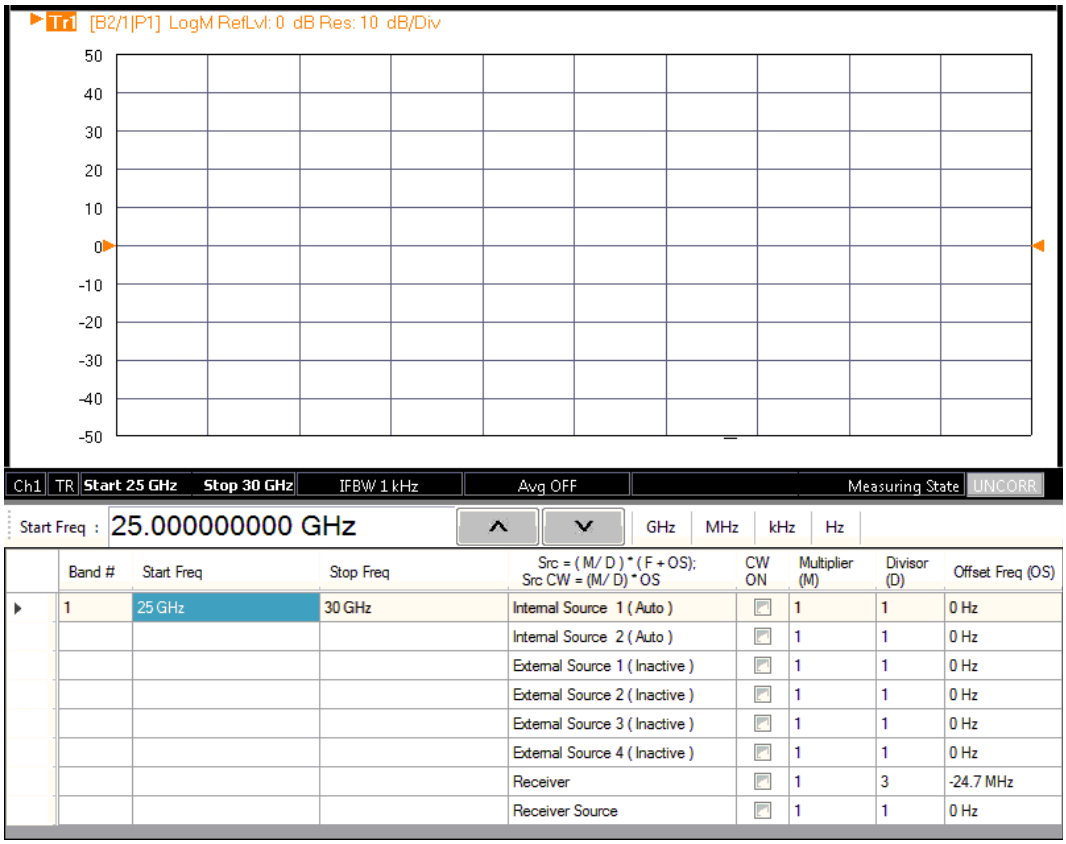


Figure 18-16. Multiple Source Harmonic Conversion Example

Some of the other selections that are part of the external module control dialog are not particularly relevant to this measurement application. Leveling should be left at RF (leveling within the VNA source chain). Common offset should only be selected if the source(s) and receiver are within about 50 MHz of each other (when multiplication factors are removed)… this is generally not the case. Start delay and VCO overrange can be used (behavior detailed in the previous section) but they are not normally changed for this type of measurement application. The band and receiver checkboxes must, of course, be enabled for the IF switching to happen.

18-4 DUT Measurement Example #1—Mixer

The DUT is a mixer with:

- a 90 to 100 GHz RF input
- a 44.5 to 49.5 GHz LO input (with the DUT operating on the second harmonic of this LO input)
- a 1 GHz fixed IF being sent to Port 2 of the VNA

Setup

A 3743A/AX module will be used to drive the RF port, an external synthesizer (Ext Synth 1) will be used to drive the LO and the DUT output will be sent to Port 2 of the VNA.

- Define band 1: 90–100 GHz
- Int Src = 1/1(f+0)
- Ext Src 1 = 1/2(f–1 GHz)
- Rcvr (Receiver) = CW 1 GHz

Once multiple source is activated, the FREQUENCY menu will be limited to covering 90–100 GHz (the FREQUENCY menu values have to be within the Band range, but do not need to be equal). One could change the FREQUENCY menu to sweep, for example, 92–98 GHz, if that was really of interest, and the multiple source setup does not have to change.

Result

External module control activated in 3739-based modes. The source is enabled but the receiver is not. RF leveling should be used. Phase inversion should be activated (see next subsection, "Phase Inversion" on page 18-25 below).

18-5 DUT Measurement Example #2—Up Converter

The DUT is an up converter with:

- a 1–2 GHz input
- a 76.5–7.5 GHz output
- The LO is fixed at 37.75 GHz and a 2^{nd} harmonic is used

Setup

Port 1 of the VNA will drive the input of the DUT and the output will be fed to a 3743A/AX module connected to Port 2 of the VNA.

- An external synthesizer (Ext. Synth 2) will provide the LO
- Define band 1–2 GHz
- Int Src = 1/1(f+0)
- Ext Src 2 = CW 37.75 GHz
- Rcvr = 1/1(f+75.5 GHz)

Again, the FREQUENCY menu values will be limited to the Band range (1–2 GHz in this case), but the start and stop on the FREQUENCY menu can be placed anywhere within that range.

Result

External module control is activated in 3739-based modes. The receiver is enabled but the source is not. The leveling choice does not matter since the base VNA is providing source drive.

18-6 Phase Inversion

Particularly in mixers, it is important to know the relationship between RF and LO frequencies. Consider a down converter:

- **LO > RF or** High-Side Mixing

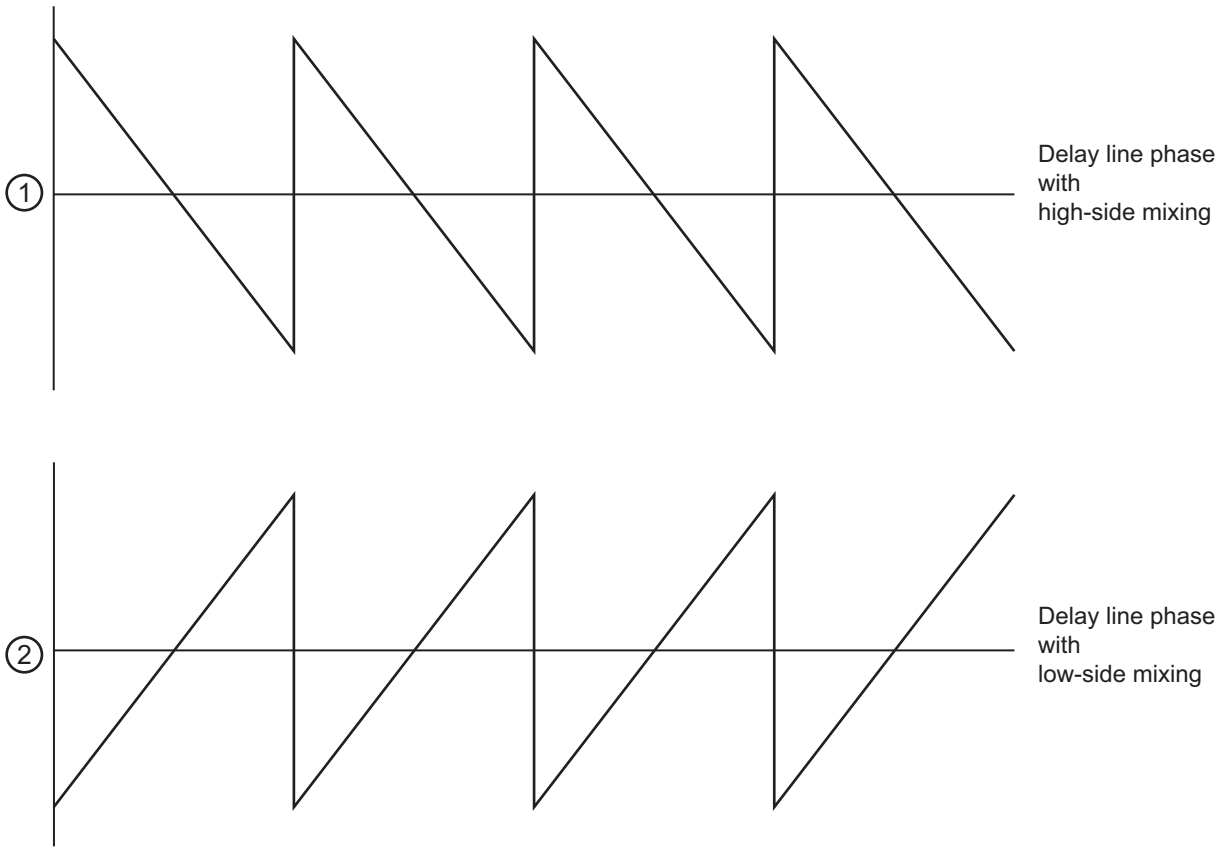
  Here the IF will have the expected phase behavior for the system and phase inversion is not required.

- **LO < RF or Low-Side Mixing**

  Here the IF will be conjugated by the DUT and the phase behavior will be the inverse of that expected. Phase inversion should be used.

Since the system has no way of knowing which source is the RF and which is the LO, or if additional fixed conversions are occurring, the information must be input manually through the Phase Inversion button on the MULTIPLE SOURCE menu, as shown in Figure 18-2 on page 18-4 above

To understand how this affects measurements, consider the phase of a delay line (after normalization with a thru line for example) when high side vs. low side mixing occurred as shown in the 3743x Module Block Diagram in Figure 18-12 on page 18-18 above.



1. Delay line phase with high-side mixing (at top).	2. Delay line phase with low-side mixing (at bottom).

Figure 18-17. Phase Inversion Behavior

The selection of the phase inversion button reverses the effect seen in low side mixing. The most severe impact can occur in calibrations where line and offset lengths are presumed to impact phase according to equations like (where L is the length and β is the propagation constant):

$$\varphi = -n \cdot \beta \cdot L$$

Equation 18-2.

If low side mixing is present, the sign of this equation is incorrect and the calibrations will proceed with the wrong phase values. Proper selection of the phase inversion button will avoid those problems.

The phase inversion feature is not available in 'simple' mode, but the state will be retained for when one returns to 'complete' mode.

Receiver Source and Receiver Calibrations

Although receiver calibrations are covered in more detail in Chapter 8, “Receiver Calibrations”, the special use of a multiple source equation will be discussed here.

The receiver calibration is sufficiently general that it may help perform power normalizations even at the plane of an external converter (acting like a pre-receiver). This receiver may be well-characterized at its frequency range (different from the frequency range of the MS464xB Series VNA receiver) but there must be some means of looking up the correct value in that characterization. This is the purpose of the receiver source equation: to act as an index to a receiver calibration that may refer to an unusual reference plane. **In most common cases, the receiver source equation is just set equal to the receiver equation.** If receiver calibrations are not in use, this equation can be set to any valid frequency.

18-7 Setting Up External Synthesizers

The MS464xB Series VNAs support up to four (4) external Anritsu sources.

External Synthesizer Connections

External synthesizers can be used when in multiple source mode. The control connections for a maximally configured multiple source scenario are shown here.

It is important to note that the synthesizers are synchronized to the 10 MHz reference of the MS464xB Series VNA so frequency accuracy is maintained. External 10 MHz sources can also be used if they have at least the spectral purity of the VNA internal clock. Note that not all external source 10 MHz outputs have enough amplitude to drive the MS464X (and vice-versa). One should check that the 'Ext Ref' annunciators are 'on' on the respective instruments. Choosing a different primary reference is an option, as is reducing cable lengths, using fewer 'star' distribution schemes or adding amplifiers.

The GPIB connections are used for controlling the synthesizers and usual GPIB cable length practices apply. Note that the dedicated connector on the VNA is used. The synthesizer GPIB addresses must match those on the VNA external source address menu located at:

- MAIN | System | SYSTEM | Remote Interface | REMOTE INTER. | Ext. Sources | EXT. SRC ADDR.

The EXT. SRC ADDR. menu is shown below in Figure 18-18.

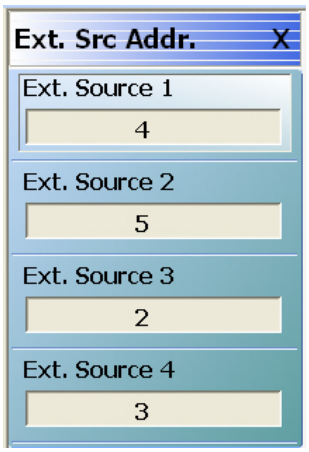


Figure 18-18. EXT. SRC ADDR. (EXTERNAL SOURCE ADDRESS) Menu

The system will poll the GPIB bus for the external synthesizers periodically including when Multiple Source mode is activated and when the MULTIPLE SOURCE menus are accessed. To verify if a connection has been established, check the EXTERNAL SOURCE CONTROL menu to see if the State is Active.

- MAIN | Application | APPLICATION | Rcvr Setup | RCVR SETUP | Multiple Source | MULTIPLE SOURCE | Ext. Src. Control | EXT SRC. COTROL

A connected source can be manually activated prior to entering data in the table to avoid having to re-enter the data, since data is only saved when Done Editing is selected. Flatness calibrations of external synthesizers may be available in certain versions of instrument firmware, in which case those buttons in will enabled.

If one or both synthesizers are turned off or disconnected while in broadband or mmWave modes, GPIB errors or sluggish system response may be experienced. It may be necessary to return to standard mode, reconnect/restart the synthesizers and then re-enter multiple source mode to continue. In extreme circumstances, it may be necessary to exit the VNA application, restart the synthesizers, and restart the VNA to establish communication.

Note	The receiver source equation should be set to the same values as the receiver equation unless an external converter is being used as part of the measuring receiver.

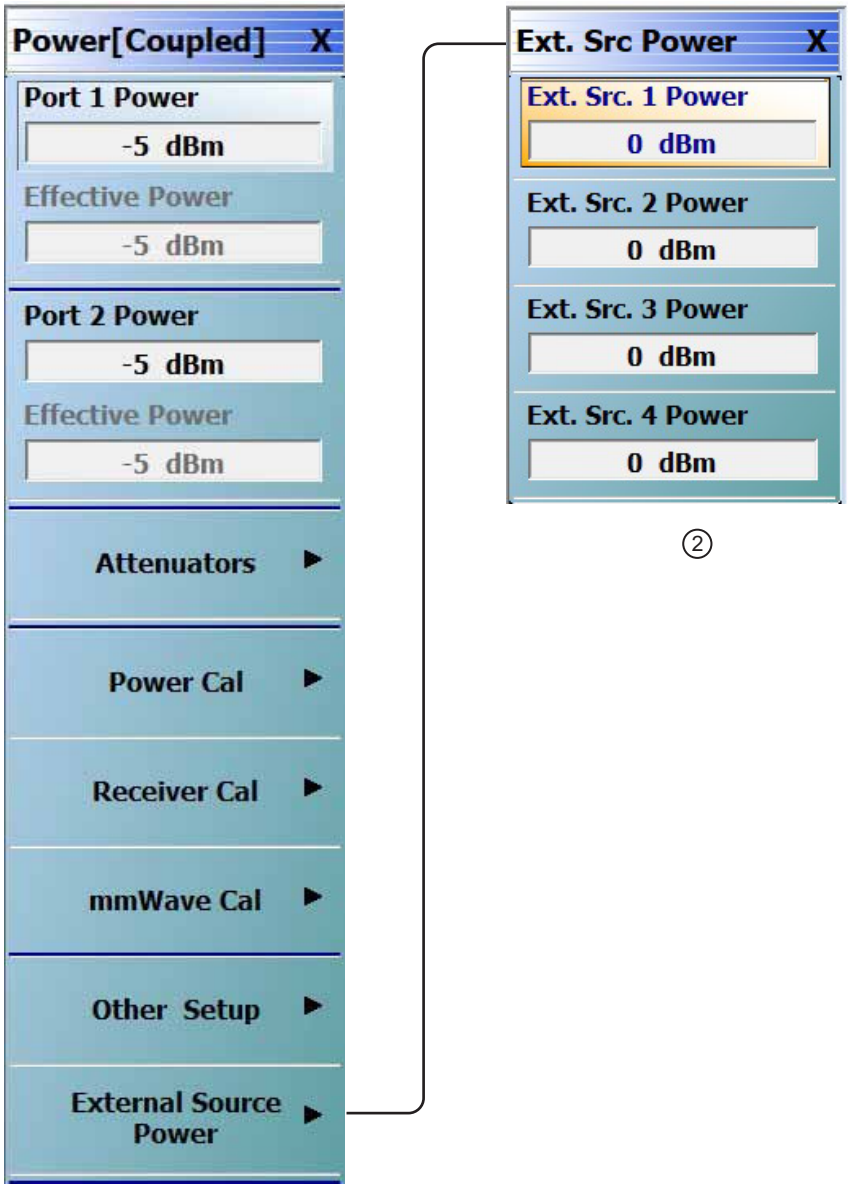
External Sources Power Level

The power level of external sources is controlled by the External Source Power button on the main POWER menu as shown in the first part (left) of Figure 18-19.

Navigation

- Main | Power | POWER[COUPLED] | External Source Power | EXTERNAL SOURCE POWER

This leads to the power entry fields as shown in the second part (right) of Figure 18-19. Again, flatness calibrations may be available with certain firmware versions.



1. POWER [COUPLED] Menu and External Source Power button.	2. EXT. SRC POWER (EXTERNAL SOURCE POWER) Menu with settings for External Source Number and Source Power Level

Figure 18-19. POWER and EXT. SRC POWER Menus

External Source Fast Trigger Mode

To enable or disable fast trigger mode when using MG3702xA external synthesizers, on the MULTIPLE SOURCE menu, click on the Ext. Src. Fast Trigger button, which will bring up the dialog shown in Figure 18-20.

Navigation

- Main | MAIN | Application | APPLICATION | Rcvr Setup | RCVR SETUP | Multiple Source | MULTIPLE SOURCE | Ex. Src Fast Trigger

Select the desired mode (ensuring needed cables are connected) and click OK to effect the mode change.

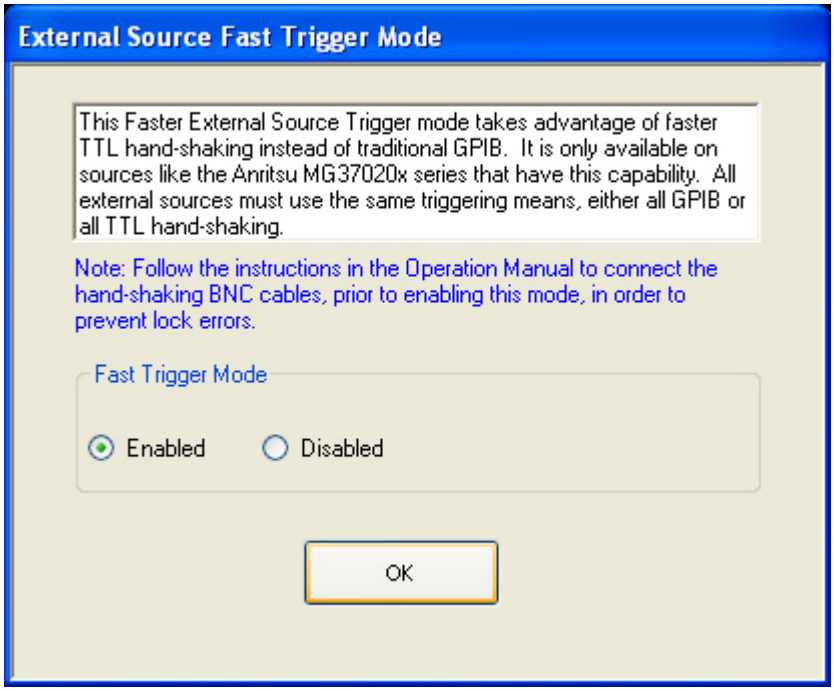


Figure 18-20. EXTERNAL SOURCE FAST TRIGGER MODE Dialog Box

18-8 Multiple Source Hints

So many measurements are possible with the multiple source system that it is hard to provide detailed direction for all of them, but some typical watch-points can be discussed:

Multiple Source Table

- The appearance of the table and the meaning of the band start and stop sometimes cause confusion. The band start and stop do not define the frequencies that you must sweep over, but rather define the range where the given set of equations will be applied. Thus multiple bands can be created with a different equation set for each. When multiple source is ON, the system frequency (i.e., that set on the frequency menu or the segmented sweep screen) must be between the band 1 start frequency (or equal to it) and the last band entered stop frequency (or equal to it). If the Frequency menu range was outside the multiple source range when editing started, when DONE EDITING is hit (or multiple source turned on if it was off), the start and stop frequencies will be coerced to the multiple source band range. One can always sweep any range within the span that has been defined.

External Synthesizers

- Activation of external synthesizers can sometimes be mysterious and part of this is related to the limitations in constantly polling the GPIB bus. To avoid a slowdown in sweep time, the bus is only polled when entering multiple source or the ext. source submenu (and certain other times related to mmWave modes). Thus the sources must be connected at their correct addresses (and of course turned on) when the menu is entered in order for it to be visible. If multiple synthesizers were connected and were accidentally set to the same address, it may be necessary to set the addresses correctly and then cycle power on the synthesizers for the bus communication to recover. Also do not forget to connect the 10 MHz references (often the VNA acts as the 10 MHz source for other instruments, but this is not required).
- When an external synthesizer is under remote control by the VNA, sometimes the display of the synthesizer will still show an old frequency (this depends on firmware revisions and the model of the synthesizer). This may not correspond to where the synthesizer actually has been programmed. That can best be verified with the VNA measurement itself or with a spectrum analyzer. The power readout, if present, will update real time and this is another way to check if control has been established (by changing the entry on the external source submenu of the Power menu).
- Complete control has only been established for the MG3690B (and beyond), MG362x1A, and MG37022A synthesizers. Certain older Anritsu and Wiltron synthesizers will work under some frequency plans, but may not always function correctly due to frequency resolution differences.

  VectorStar control of the MG362x1A assumes the SCPI language selection on the synthesizer and not a native emulation mode of the 3690 series. The SCPI language choice is the default for that synthesizer family.
- The discussion in this chapter generally assumes the synthesizer was in its default state when control began. As the VNA only accesses the frequency and power lists, other aspects of the synthesizer may be manually configured and the VNA will not interfere. This can be useful for certain applications (e.g., performing flat power calibrations on the synthesizer itself using its software and then using the resulting improved power accuracy for VNA measurements), but may lead to confusing results in other cases (e.g., low rate modulation turned on).

External Module Control Dialog

- On the Ext. Module Control dialog, the maximum VCO overrange defaults to 250 MHz (this is the amount of overrange allowed on the base 2.5–5 GHz VCO, thus at the 20 GHz plane, it translates into 2 GHz), but this amount may not always be available depending on temperature and other variations. If one is using a lot of overrange and lock errors are visible on the message bar, the frequency plan of the setup should be re-arranged to require less overrange.

- When using RF leveling, less control range is available than when using IF leveling (only applies to the 374x modules) due to lower dynamic range detection. The IF leveling only works well when the net IF is less than a few hundred MHz so it can work well for IMD measurements, but not so well for mixer measurements. RF leveling should be used for the latter. Within RF leveling, there is a subset of using RF leveling detection and VNA leveling detection when the base instrument has the *dual source* option (default is RF). VNA leveling allows for more control range in exchange for a small degradation in level stability. For all leveling types when doing a user power calibration, the system automatically uses the very linear module receivers to improve accuracies at requested power levels below –25 dBm.
- The common offset selection (Ext. Module Control dialog) is valuable for high order multiplied measurements as has been discussed, but requires the base receiver and source (referenced to a 5 GHz plane) to be within about 80 MHz of each other. Thus at a 20 GHz max frequency plane, they must be within about 320 MHz. If lock errors are observed, consider unchecking this box.

Power Calibration Reference Plane

- If the user needs to perform a user power calibration with a multiple source control setup, there can sometimes be ambiguity at which frequency the power calibration is being performed if the source equation(s) is(are) not at the default value(s). This is important since that frequency is sent to the power meter to use the appropriate correction factor and is used to determine if an appropriate power meter is connected. This is more likely to be an issue at mmWave frequencies when multiplication of the internal sources is more common.

  The Power Cal Ref Plane button on the Multiple Source menu (found via Main | Application | Receiver Config. |(Rcvr Setup) Multiple Source) gives one some flexibility in how the power calibration is assigned. The default value is System Frequency which refers to the multiple source runner variable and the x-axis plotting variable. Source Equation can also be selected in which case the output of the relevant internal source equation will be used.

  As an example, suppose a band is defined as 60–90 GHz (and the sweep is set for 60–90 GHz), Source 1 is defined as (1/2)*(f + 0 MHz) (a x2 multiplier is to be connected to the VNA port which could be the DUT or could be part of the setup) and the power calibration is to be performed on port 1. If System Frequency was selected for the reference plane, the power meter would be sent the frequency list 60–90 GHz and would be appropriate if the power calibration was being performed at the multiplier output. If Source Equation was selected, the power would be sent the frequency list 30–45 GHz and would be appropriate if the power calibration was being performed before the multiplier input.

  Power calibration reference plane selection is not available in the 'simple' mode.

Chapter 19 — Noise Figure (Option 41)

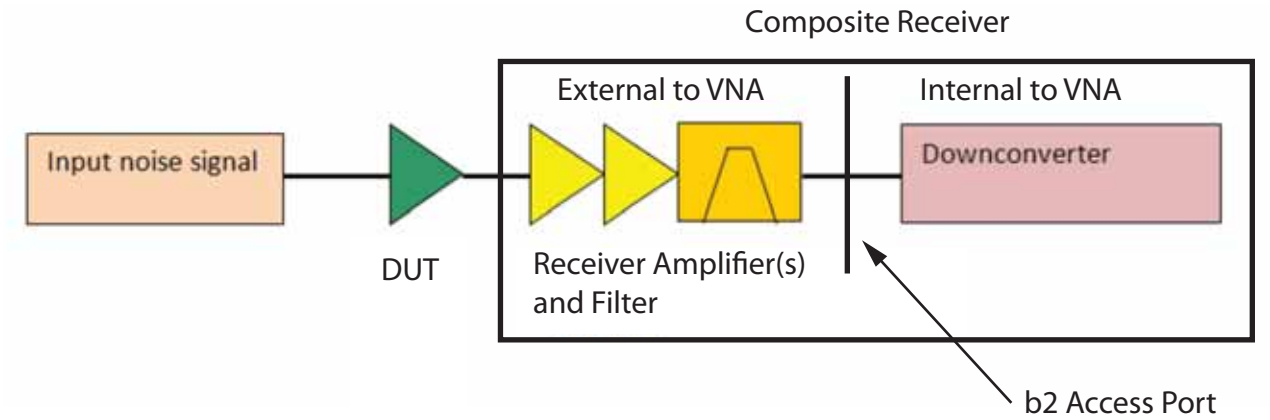
19-1 Chapter Overview

Option 41 provides noise figure measurement capabilities in the MS464xB series VNA family. The intent of this chapter is to explore the physics of the measurement as implemented, offer operational guidance on how to use the option, and to provide hints and tips on how to optimally setup a noise figure measurement.

19-2 Background

In recent years, numerous improvements have been made in noise figure measurements through better algorithmic understanding of the measurements (e.g., [1]-[6]), more sensitive receivers, and simpler methods of processing noise power measurements. This chapter describes one specific set of improvements, based on the cold source measurement technique and related error correction techniques to enhance the noise measurement accuracy.

One particular avenue of noise figure measurements will be discussed here based on a cold source technique and some corrections applied to it. Before beginning with the algorithmic discussion, it may help to briefly look at the measurement hardware. In all cases, the DUT will feed a composite receiver consisting of the main receiver of the VNA augmented with amplification and filtering. The reasons for the extra gain will become clear later in this chapter. The basic structure of the VNA and DUT setup is shown below.



A basic diagram of the noise figure measurement setup is shown here. The MS464xB VNA is represented by the downconverter block and the processing circuitry (not shown) after it.

Figure 19-1. Noise Figure Measurement Basic Setup Diagram

Hot-Cold or Y-Factor Noise Figure Measurement Method

Historically, a popular method for noise figure measurements was termed a hot-cold or Y-factor technique in which a noise source was employed that could produce a low noise output power (cold = N_c) and an elevated one (hot = N_h). This noise source becomes the 'Input Noise Signal' in Figure 19-1.

The ratio of measured noise powers in these two states was termed the Y-factor ($Y = N_h/N_c$) (e.g. [2]) and could lead to a quick calculation of noise figure (F) via the equation (Eq. 19-1) below.

$$F = \frac{\frac{T_h}{T_0} - 1}{Y - 1}$$

Equation 19-1.

Where T_h is the equivalent hot temperature of the noise source and the cold temperature was assumed equal to a temperature of 290 K = T_0 (per IEEE definition).

By making a noise figure measurement of the receiver itself (F_{rcvr}) and of the system ($F_{sys} = F_{DUT}$ + Receiver), it is possible to deconvolve the DUT noise figure with the familiar Friis' equation (Eq. 19-2) below: [1].

$$F_{DUT} = F_{sys} - \frac{F_{rcvr} - 1}{G}$$

Equation 19-2.

Here, the DUT gain G could be measured separately (via S-parameters) or it could be determined from the change in measured noise powers during calibration and measurement. One advantage of the hot-cold noise figure measurement method is that no absolute power calibrations are needed (all based on ratios). This was particularly important in the past when wide dynamic range power measurements over large bandwidths were more difficult. One challenge of this measurement method was the ability to provide a well calibrated noise source.

Because of the signal levels and bandwidths involved, accurate noise source calibration (for T_h) is challenging and normally left to only a few metrology laboratories. In addition to issues in calibrating the noise source, a larger challenge occurred when the match of the noise source changed between the hot and cold states [4]. This could lead to large errors, particularly as the DUT input match worsened. These errors could be partially corrected using a source correction method (e.g., [3]).

Cold Source Noise Figure Measurement Method

The cold source noise figure measurement method was developed to eliminate the requirement for a multi-state noise source, which would allow the use of a simpler, better controlled noise source (nominally a termination at room temperature). In this case, the noise figure is found from a simpler equation (Eq. 19-3), but it has some subtleties:

$$F = \frac{kT_0B + N}{kT_0BG}$$

Equation 19-3.

Where:

- k is Boltzmann's constant
- N is the noise power added
- G is the gain
- B is the bandwidth

To calculate the noise figure, there are a few key steps to accomplish.

First, an absolute noise power is now required (numerator *N*), so a means of a receiver power calibration is needed.

- The receiver can be calibrated with a variety of calibrated signal sources including a power-calibrated sinusoidal source or a calibrated noise source. A highly linear receiver allows this use of different signal sources.
- Other methods are possible to determine added noise power including the use of a calibrated hot noise source. Again, the calibrated noise sources would only be utilized for calibration. In both cases, an absolute power reference is being created.

Second, an effective measurement bandwidth (*B*) is needed. Since measurement bandwidth is largely determined by the digital IF system, *B* can be pre-calculated. This bandwidth value may also be determined with the absolute power calibration step if a noise signal is used, otherwise it is determined separately.

Third, to isolate the noise figure of the DUT, the noise contributions of the receiver must be taken into account. As with the hot-cold method, a measurement of receiver noise is required with the cold source attached to the receiver input. Taking the receiver noise into account, Eq. 19-3 above can be re-expressed in the form below (Eq. 19-4):

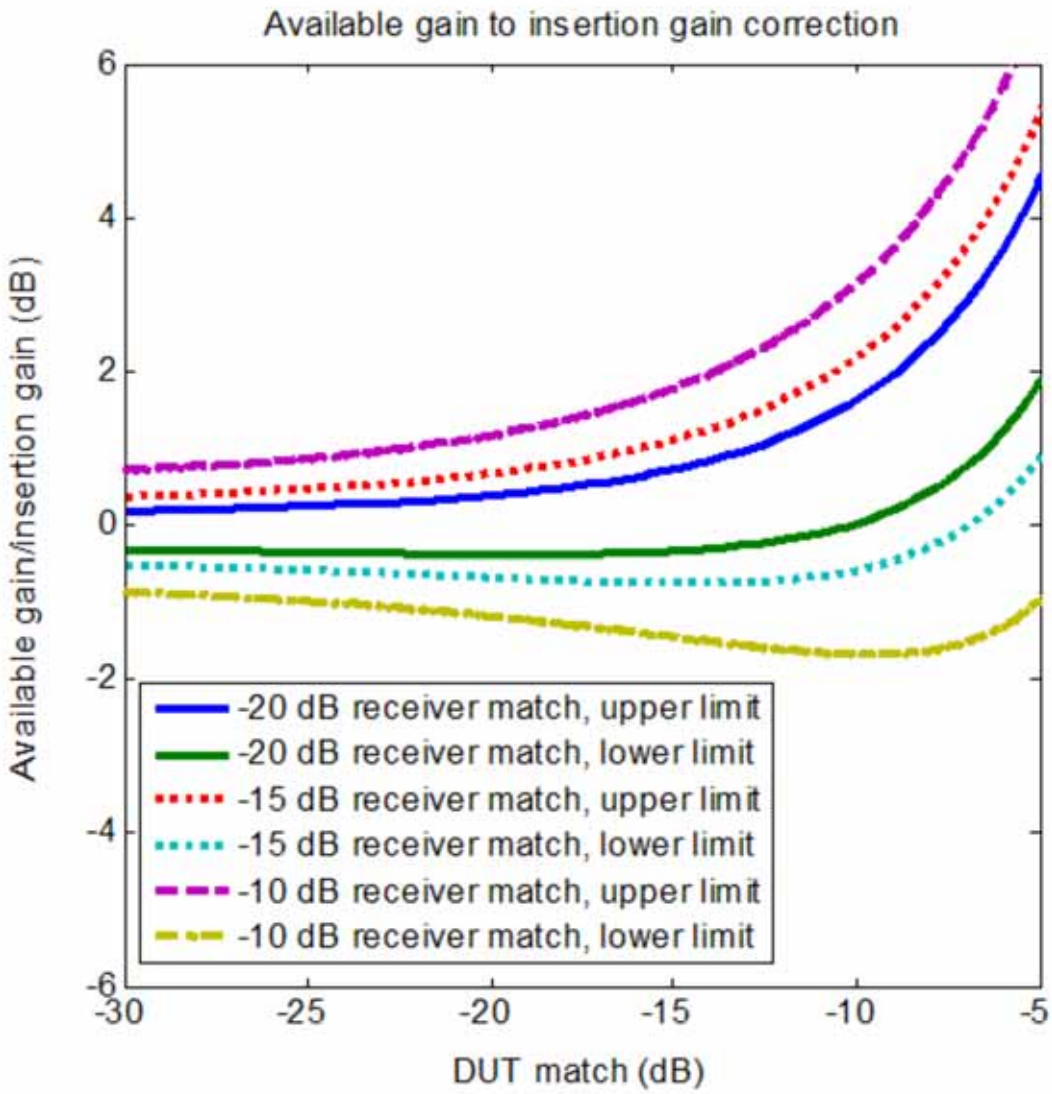
$$F = \frac{1}{G} + \frac{N_{DUT+rcvr} - N_{rcvr}}{kT_0BG}$$

Equation 19-4.

A few things are immediately obvious: errors in gain or the noise power measurement will propagate to noise figure on roughly a dB-for-dB basis (if the gain is sufficiently high at least). These accuracies will be discussed throughout this chapter, but it is worth remembering the approximate dependence.

Measurement Considerations

Of all of these terms, and it is true with hot-cold approaches as well, the term most often misvalued is the DUT gain. As has been discussed many times before (e.g., [3]), the noise figure definition is based on input noise power available to the DUT and output noise power available from the DUT. As a result, it is available gain that is the most appropriate value to use as opposed to insertion gain (which is what was commonly used with Y-factor approaches) or even $|S21|^2$. As the DUT match worsens, the difference between these gain definitions increases. An example plot is shown in Figure 19-2 on page 19-4, where the DUT input and output matches (assumed symmetric for simplicity) form the x-axis and the receiver match is parameterized. The cold source match is assumed to be –20 dB and the DUT is assumed to have decent isolation ($|S21 \times S12| = 0.1$). Multiple dB differences in gain, and hence in calculated DUT noise figure are possible as the match degrades beyond –10 dB.



The difference between insertion and available gain as a function of match levels is shown here. Since gain strongly affects the noise figure calculation, understanding this difference is important.

Figure 19-2. Differences Between Insertion and Available Gain as Function of Match Levels

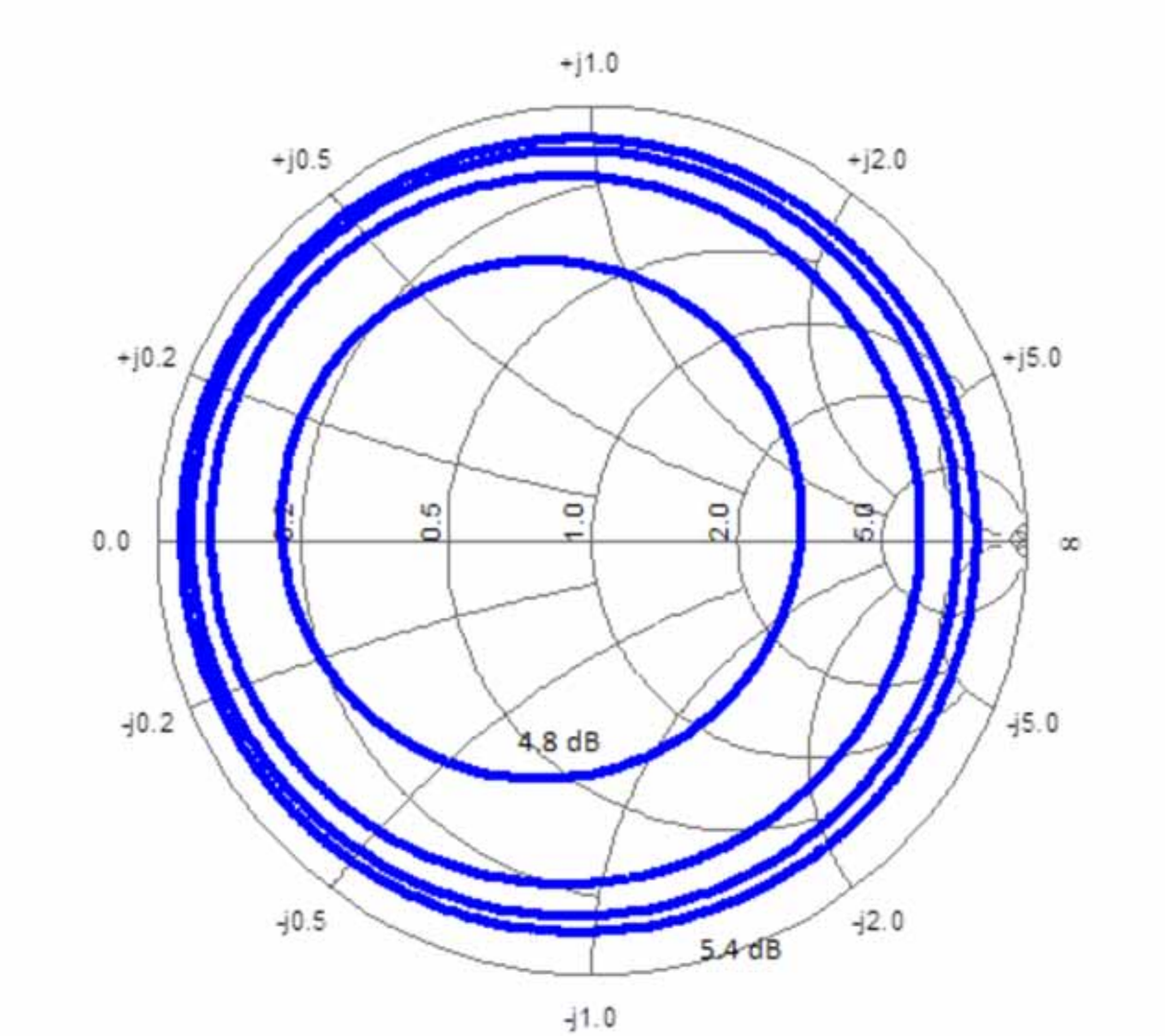
In the case of the cold method (and assuming a quality termination is used on the input), it is primarily the DUT-receiver interaction that causes the difference in gain. From Eq. 19-4, since gain errors map essentially linearly to noise figure errors, using the correct gain term is important even for commonly observed return losses.

Another potential source of errors in the noise figure measurement is the interaction of the receiver with DUT output. If the receiver noise power is strongly affected by the source impedance (i.e., a large R_n in the parlance of noise parameters), then significant errors could be encountered by not accounting for this interaction. This is accomplished by error correcting with the receiver noise parameters. To understand the relationship (e.g., [7]), recall that the noise response of a device (receiver in this case) may usually be characterized by four parameters sometimes represented by F_{min}, R_n, and Γ_{opt} (which is complex, hence counting as two parameters). F_{min} is the minimum noise figure of the device when presented with an optimum source match (given by Γ_{opt}) and R_n is a sensitivity parameter describing the reaction of noise figure to changes in source match. The net noise figure of the device is given by (in this formalism) (Eq. 19-5):

$$F = F_{min} + \frac{4 \cdot R_n \cdot \left|\Gamma_S - \Gamma_{opt}\right|^2}{Z_0 \cdot (1 - \left|\Gamma_S\right|^2) \cdot \left|1 + \Gamma_{opt}\right|^2}$$

Equation 19-5.

If R_n is small relative to the system impedance Z_0, then the net noise figure is relatively insensitive to changes in the source reflection coefficient Γ_S. Commonly, the effective receiver R_n is small and the dependency is minimal, since the receiver is normally a matched/feedback amplifier with some loss in front of it from cables, switching, and similar parameters. The noise circles of an example, typical receiver structure are shown below at 50 GHz. Even if the DUT output is very poorly matched ($|S22| > -1$ dB), the effective receiver noise figure only changes by ≈ 0.65 dB from F_{min}. Assuming the DUT has ≈ 10 dB of gain and ≈ 2 dB noise figure; this would only add 0.1 dB of uncertainty. If the DUT had ≈ 20 dB of gain and ≈ 2 dB noise figure, this would only add 0.01 dB of uncertainty. With many common receivers, even less sensitivity is displayed although the effect is, of course, a function of which pre-amplifiers are used with that receiver.



The effect on noise figure of varying the impedance seen by an example receiver is shown here. As the parameter R_n gets small (as is the case for many receiver structures), this impedance dependence becomes weak.

Figure 19-3. Variable Impedance Effects on Example Receiver Noise Figure

If further reduction in uncertainty is desirable, it is relatively straightforward to add additional correction by including three additional measurements during the receiver noise calibration step (e.g., [4]). They are measuring the noise power with three reflect standards in addition to the cold noise source. Here one is just drawing on the long history of noise parameter measurement to help with an increase in receiver correction.

Linearity

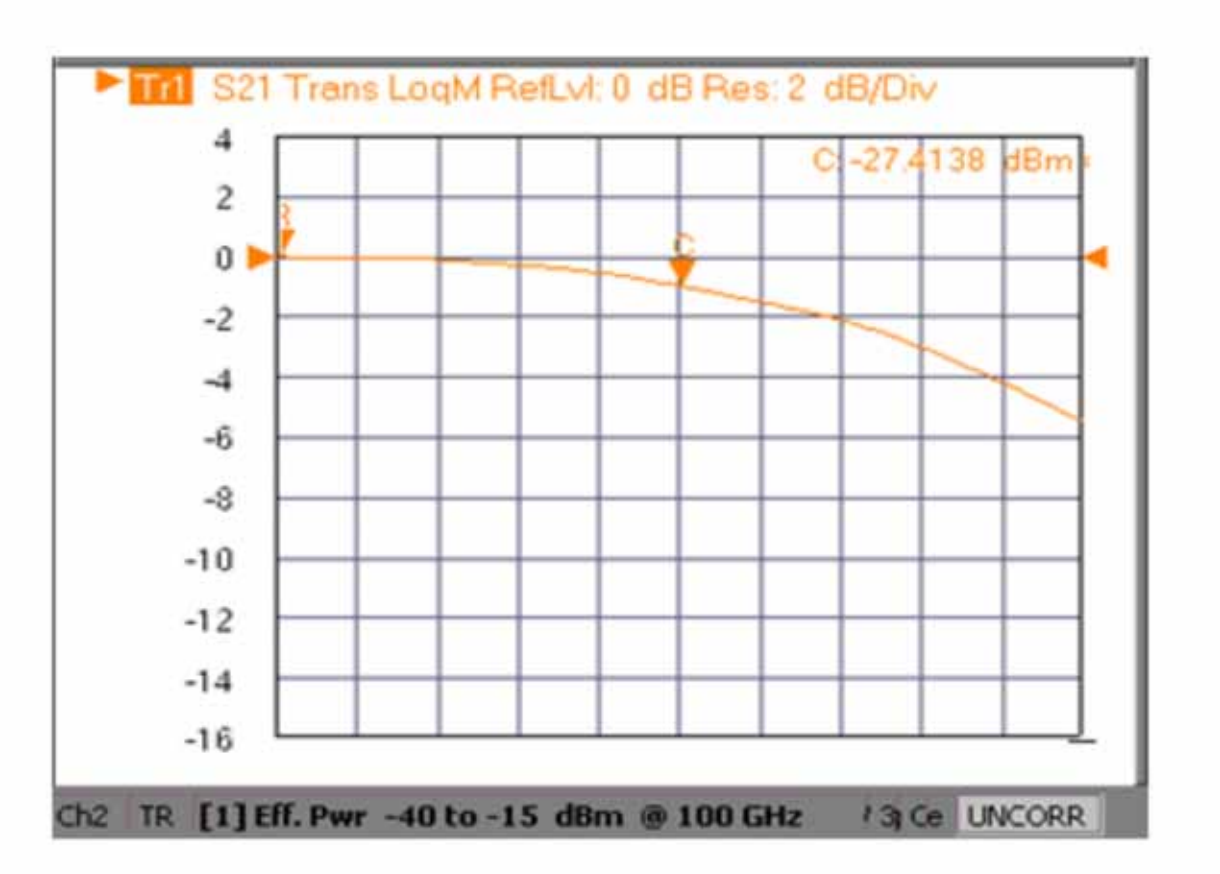
Another aspect of this measurement that sometimes receives less attention is that of linearity. Since it is a measurement of noise, a common perception is that there could never be anything nonlinear about the measurement process. Several issues do come into play

- Since the measurement is over a large bandwidth and the gains of both the receiver and DUT may be large (> 40 dB), the integrated power may be large enough to cause compression either in the amplifiers or in the IF of the instrument.
- Since the DUTs are often small signal LNAs, the power at which the gain was measured must be low enough that the gain result is not compressed. This level may be below –30 dBm for many DUTs.
- The instrument ADCs have low level linearity that must also be respected. This is normally resolved by having enough receiver gain. Receiver gain plays a key role in the uncertainty of the noise figure measurement (see "Uncertainties" on page 19-23 below).

To illustrate the first point, suppose the receiver is composed of a 40 dB gain, 20 GHz broadband amplifier (plus some other components) before the downconverter and suppose the DUT also had about 40 dB gain and about 20 GHz of bandwidth. The input termination (cold noise source) will produce about –174 dBm/Hz noise power at room temperature. The integrated power at the output of the receiver amplifier (integrated over the 20 GHz) could reach +9 dBm (–174 + 80 +103). If the receiver amplifier compressed before this point (or if the downconverter did), then errors would occur. In this particular case, filtering would likely be needed anyway for image rejection (described in sections below) so this may not happen. If compression appears to be a problem, then less gain in the receiver path could be used.

LNA DUT Compression Behavior

The second point is somewhat obvious, but one may not think to look at the DUT compression behavior if it is an LNA. The VNA default power ranges from –10 dBm to +5 dBm depending on model and options. This power range may not be appropriate for the DUT. By looking at |S21| of the DUT as power is varied or by using the gain compression tools in the power sweep modes (see "Power Sweep (Swept Frequency)"of this **Measurement Guide** for more information), one can ensure that the DUT gain data is obtained under linear conditions. As an example consider a W-band LNA whose noise figure was of interest. From the plot in Figure 19-4, the DUT is already 1 dB compressed at an input power of ≈ –27.4 dBm at 100 GHz. The gain measurement in this case should be done at –35 dBm or lower.

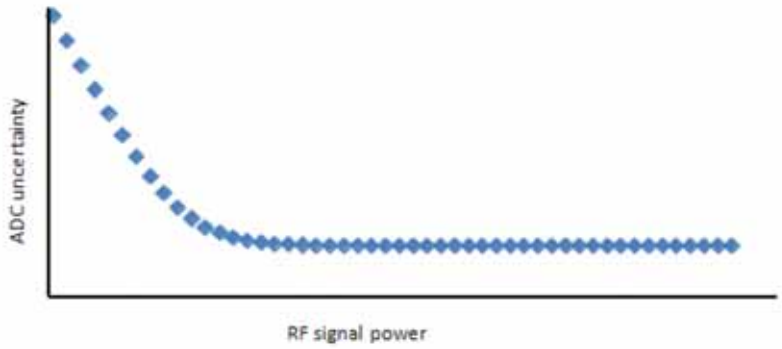


An example LNA power sweep is shown here. Many LNAs compress at very low power levels and it is important that the gain be measured in a linear region for an accurate noise figure result.

Figure 19-4. Example of LNA Power Sweep

Low Level Compression

The third point, related to low level compression, is subtle but true for all analog-to-digital converters (used in every measuring receiver) at some point. In addition to quantization error, integral and differential non-linearities become most obvious at the lower signal levels that are implicit in a noise figure measurement (assuming one avoids the compression conditions described earlier!). In terms of sampling RF power, an ADC uncertainty plot may look something like the figure below. The plot suggests that there is a minimal level of receiver gain necessary to avoid problems from further down the IF chain of the system.



An example analog-to-digital converter uncertainty curve is shown here. At low signal levels, the linearity issues can become more obvious. So, it is important that there be a minimum level of gain earlier in the receiver system to avoid problems in a noise measurement.

Figure 19-5. Example Analog-to-Digital Converter Uncertainty Curve

Noise Power

A key element in Eq. 19-4 above, as yet unaddressed in this document, is the noise power itself. Historically, one way of measuring this quantity is a dedicated noise receiver with a large, variable gain system and appropriate filtering. This can be expensive and cumbersome and the typically wide bandwidths can lead to problems with narrowband devices (if the noise measurement bandwidth is close to the DUT bandwidth, including images, errors can result). Another approach is to use the basic VNA IF system itself, since it is highly linear to extremely low signal levels and has a fairly complete digital filtering system for tailoring the response shape. There may, however, be bias or leakage signal present in the response at these low levels that are not of interest. So in those cases, an RMS-like noise power measurement process is desired. In a non-ratioed channel measurement (usually $b_2/1$), this net RMS noise power can be expressed as (Eq. 19-6):

$$Noise_power = \frac{\sum_{k=1}^{N} |b_{2,k}|^2}{N} - \left(\frac{\sum_{k=1}^{N} |b_{2,k}|}{N} \right)^2$$

Equation 19-6.

The reader may note the similarity to an equation for variance. Here N is a number of measurements (made at each frequency, termed a 'number of RMS points') and is typically a sizable number so that the effects of the mean (the second term in Eq. 19-6) can be removed even if it is varying slightly with time. The default value of N is 3,000. Smaller numbers can be used for greater measurement speed (at the expense of some data jitter, more on this in the uncertainties section) and larger numbers can be used for some reduction in data jitter. Values of N greater than about 7,000 generally produce little benefit in data jitter reduction.

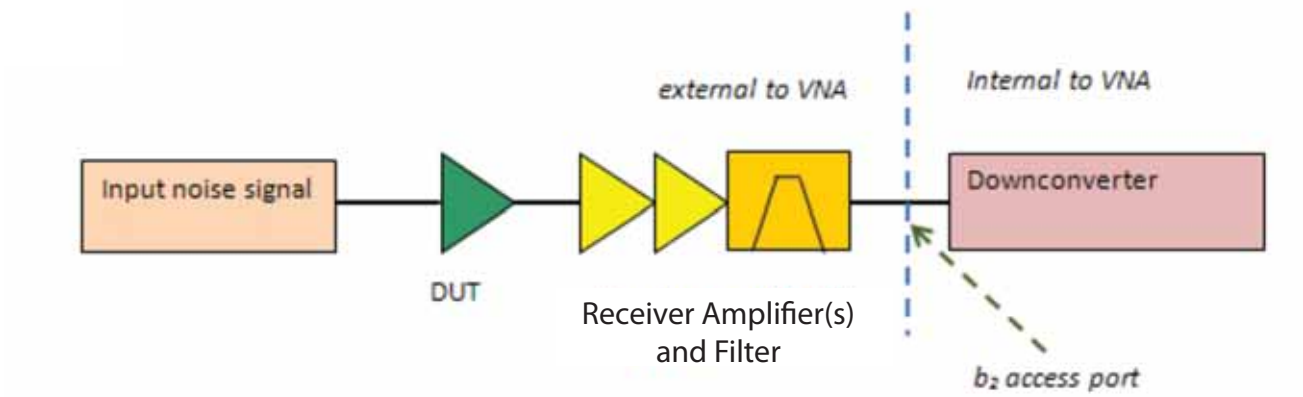
One may also ask how the unratioed parameter, $|b_2|$ in this case, comes to be a proxy for root-power. This comes via the receiver calibration (as is used for mixer, harmonic, IMD, and other measurements), since it transfers the traceable accuracy of the power sensor to the receiver parameter ($|b_2|$). One may rely on the factory ALC calibration to provide the power reference for this calibration. It is more accurate to perform a user power calibration at the frequencies and level of interest prior to the receiver calibration to get the power accuracy roughly to the 0.1 dB level or better. Information on performing the power calibration is elsewhere in the measurement guide and operation manuals. When performing noise figure measurements, they are optimally performed at lower signal levels (e.g., –60 dBm using the internal step attenuators of Options 61/62 or an external attenuator) to stay in roughly the same part of the linearity curve as the noise figure measurement and to improve match. Performing power calibration at low signal levels (e.g. –60 dBm) may not always be practical. A low-level power measurement usually requires a diode-based power sensor, which are not readily available at frequencies above 50 GHz. At higher frequencies, a thermal sensor may be required for power calibration, which typically have a lower power measurement of ≈30 dBm.

19-3 Composite Receiver Component Selection

A number of comments were made in the previous section about the receiver requirements in terms of gain, filtering, linearity and other behaviors. Since these are custom components selected for many measurements, it is important to present a way of assessing a given measurement problem and how to assemble the user provided portions of the composite receiver system.

By slightly modifying the diagram in Figure 19-1, one can see the delineation of system and user components in Figure 19-6. The downconverter is accessed through a port associated with the b_2 loop (this avoids the loss associated with the coupler on Port 2 and helps improve uncertainties). The port itself can be used if the DUT + external gain is high enough (usually over 60 dB).

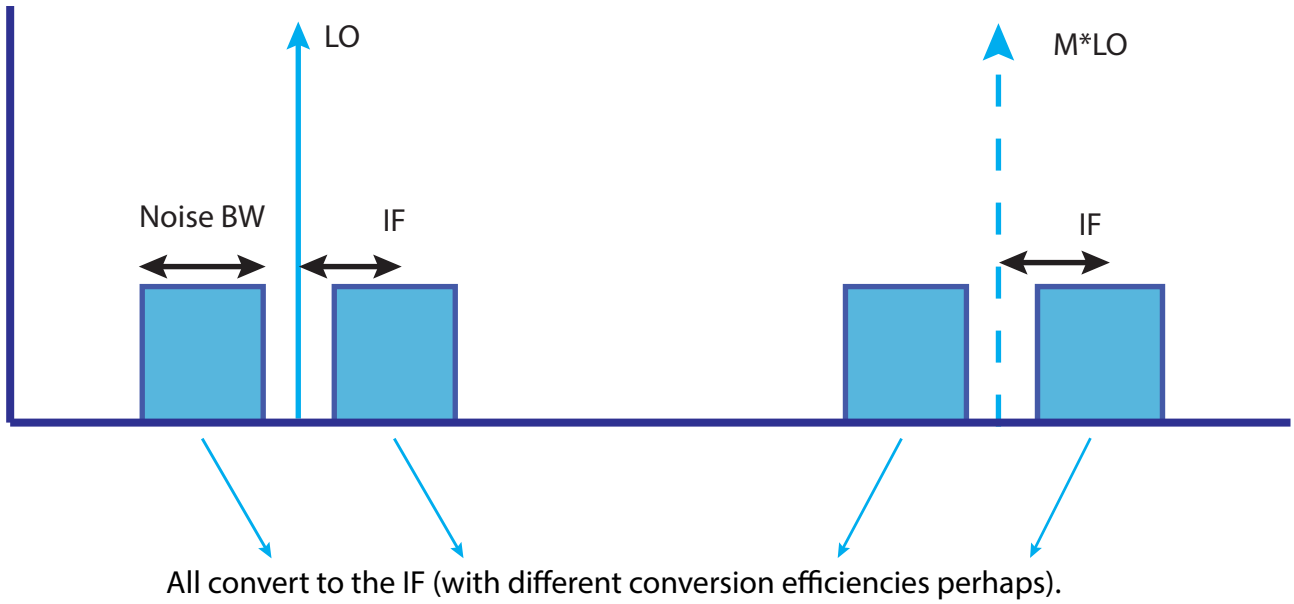
The receiver amplifiers are used to reduce the effective receiver noise figure which can dramatically improve uncertainties as will be shown in a later section. The required gain is somewhat dependent on the DUT but 40 dB of gain is a reasonable starting point. Multiple physical amplifiers can be used. The noise figure itself is generally not as critical but less than 5 dB or so is generally desirable. The lower the DUT noise figure, the more important the receiver amplifier noise figure becomes. This will become clearer in the uncertainties section.



A modified block diagram is shown here to illustrate locations better.

Figure 19-6. Modified Block Diagram for Locations

A filter is shown above in Figure 19-1 and Figure 19-6 and this is for several reasons. As mentioned earlier, if the gain blocks are very broadband, some filtering can help prevent inadvertent receiver compression. Perhaps more importantly, the filter helps prevent image responses of the receiver from complicating the noise figure measurement. To understand this, recall that any downconverter has responses for at least some harmonics of the LO (commonly 3^{rd}, 5^{th}, 7^{th}… for balanced mixers and all harmonics for some structures). Thus noise energy at an IF offset from these LO harmonics will also convert to the system IF and be included in the noise power computation. This effect becomes more pronounced when the desired product is already a harmonic of the basic LO and subharmonic components (which may have more favorable conversion efficiencies) can play a role. The general concept is illustrated below in Figure 19-7.



A spectral diagram is shown here is illustrate how harmonic conversion in the downconverter can add noise sidebands to the result that may be unintentional. Filtering as part of the receiver chain can reduce this issue.

Figure 19-7. Spectral Diagram for Harmonic Conversion in the Downconverter

By having a filter in the receiver chain to isolate the LO harmonic of interest, the other multiples can be effectively removed. As an example, suppose measurements are being run over 5 to 10 GHz which happens to be the fundamental LO component. By having a low pass filter in the chain with a corner of about 10 GHz, higher order terms can be removed.

Depending on the roll-off rate of the filter, it may be necessary to split that octave into two bands (e.g., 5 to 8 GHz and 8 to 10 GHz). This decision would be based on the DUT bandwidth as well as the filter behavior: if the DUT is already rolling off at 10 GHz, then a fairly weak stop band filter would be ok but if the DUT gain is fairly flat to much higher frequencies, then a very sharp stop band filter is suggested or the band should be split as mentioned.

For higher frequency measurements, it is useful to recognize that the fundamental LO range is still 5–10 GHz with the normal harmonics being used of 1, 2, 4, and 8 (adding 9 and 11 for ME7838A/AX operation to 110 GHz). The filtering should be designed to remove the other harmonic responses. Recall that the MS464xB has a system breakpoint at 2.5 GHz (lower frequency loops are on the rear panel, higher frequency loops are on the front panel) but this is a natural filtering breakpoint as well. When operating at frequencies lower than a few hundred MHz, additional filtering may be desired (if the DUT has substantial gain to much higher frequencies) to remove effects of the 3^{rd} harmonic and 5^{th} harmonic responses of the low band converter.

From Figure 19-7, one may question the two noise sidebands centered around the LO signal. Since the VNA input is not preselected, noise energy at an IF offset either above or below the LO will be included. The basic IF frequency is 12.35 MHz so the measurement will effectively be an average over approximately a 25 MHz region. Additional filtering can be employed in narrowband applications where this bandwidth is an issue.

19-4 Measurement Setup and Procedure

Noise Figure Measurement Procedure

The noise figure measurement process consists of a few relatively straightforward steps:

1. Measurement of DUT gain or S-parameters (over an appropriate frequency range and at an appropriate power level as discussed before)
2. Perform an optional user power calibration over the frequencies of interest to optimize receiver calibration accuracy
3. Physical configuration for the noise measurement (discussed in the previous section)
4. Basic setup (frequency range, number of points, type of device)
5. Receiver calibration (transferring the traceable power accuracy to the receiver)
6. Noise calibration (measurement of the receiver noise power so it can be removed from the calculations)
7. DUT measurement

The first three items have already been discussed. A few more details are in order on the gain/S-parameter measurement comment.

Gain/S-Parameter Data

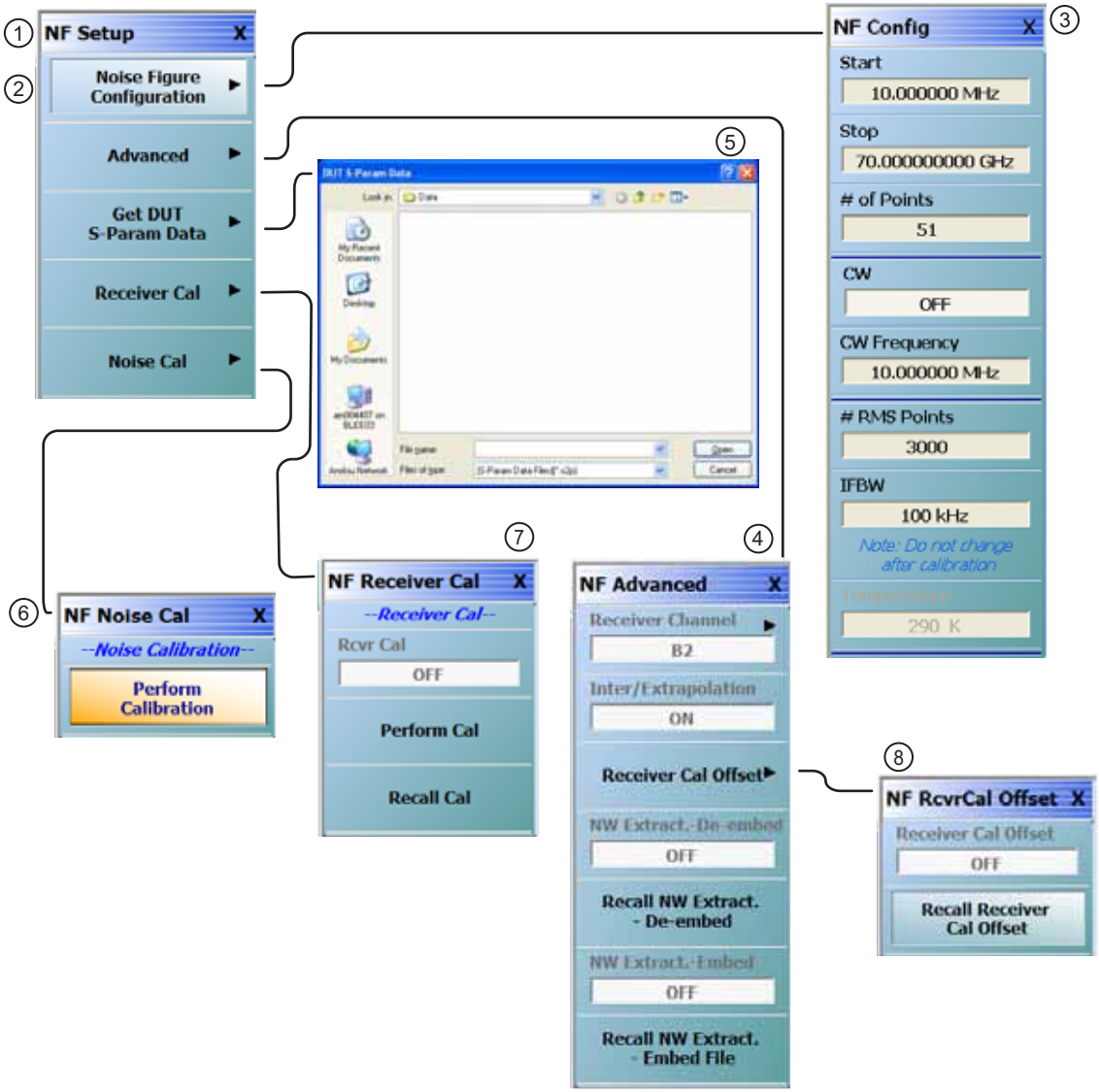
The gain/S-parameter data should be acquired on a frequency range containing the desired noise figure frequency range. The same frequency points need not be used (although that may be desired if the gain changes rapidly with frequency) as interpolation will be invoked. Extrapolation will also be used if needed, but this is not advised for accuracy reasons.

S-parameters should be saved in the S2P file format and they will be loaded in the noise figure application. A full 12-Term calibration is advised for optimal accuracy and to enable the use of the more correct available gain terms, but data from normalization-only calibrations will also be accepted.

The basic setup steps mentioned in Step 4 are illustrated in the menus of Figure 19-8 below:

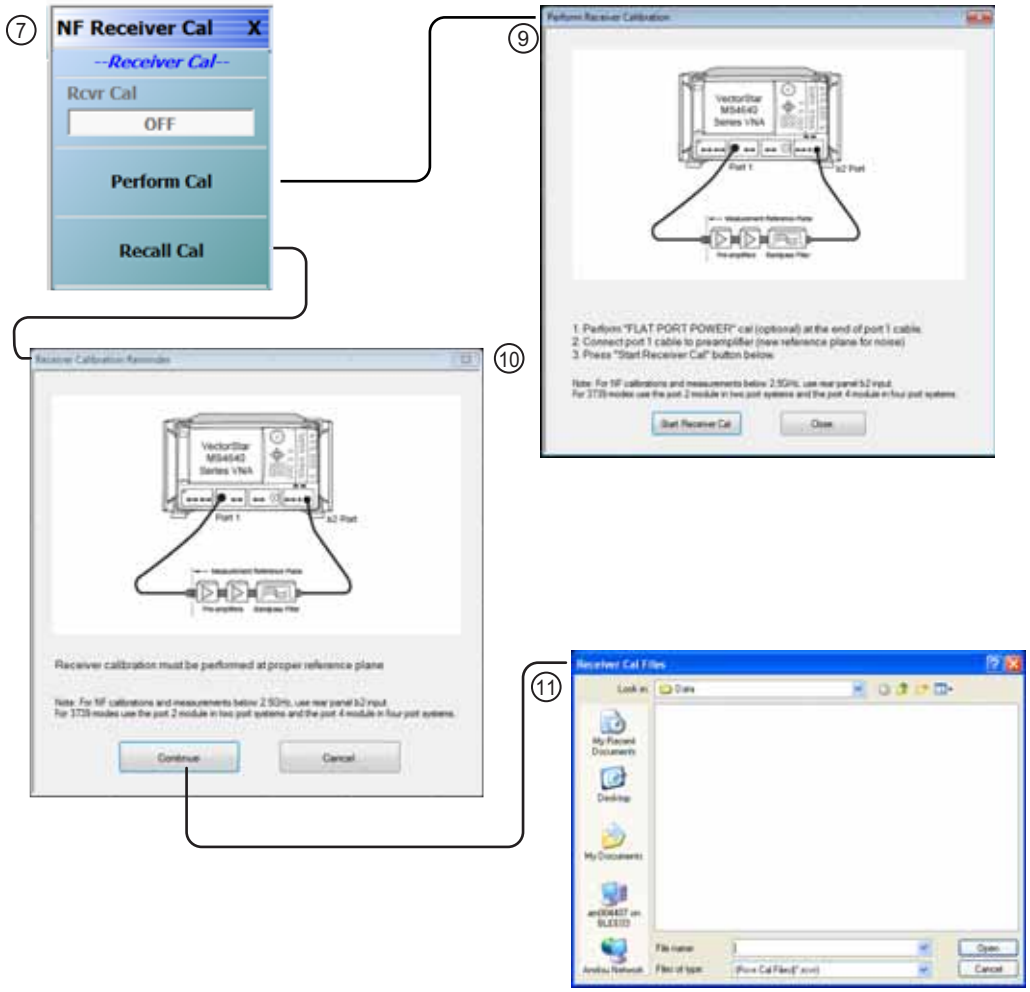
Navigation:

MAIN | Application | Mode selected = Noise Figure | Noise Figure Setup



1. NF SETUP menu.	5. DUT S-PARAM (S-Parameter) dialog box.
2. Noise Figure Configuration button selected.	6. NF NOISE CAL (Calibration) menu
3. NF CONFIG (NF Configuration) menu.	7. NF RECEIVER CAL (Calibration) menu.
4. NF ADVANCED menu.	8. NF RCVRCAL OFFSET menu

Figure 19-8. Noise Figure Menu Set (1 of 2)



9. PERFORM RECEIVER CALIBRATION Dialog Box 10. NF RECEIVER CAL Dialog Box	11. PERFORM RECEIVER CAL dialog box. Provides access to previously saved receiver calibration RCVR files.

Figure 19-8. Noise Figure Menu Set (2 of 2)

	When the instrument changes the VNA application mode from Transmission/Reflection to Noise Figure, the instrument resets to the default settings for that mode.
Note	The default Noise Figure trace condition is a single trace and Noise Figure (LogMag) response. The default Transmission/Reflection condition is four traces, S_{11} (Smith Chart), S_{12} (LogMag + Phase), S_{21} (LogMag + Phase), and S_{22} (Smith Chart) responses.

Most of the setup parameters are self-explanatory. Some of the more noise-specific items

- **Number of RMS Points**

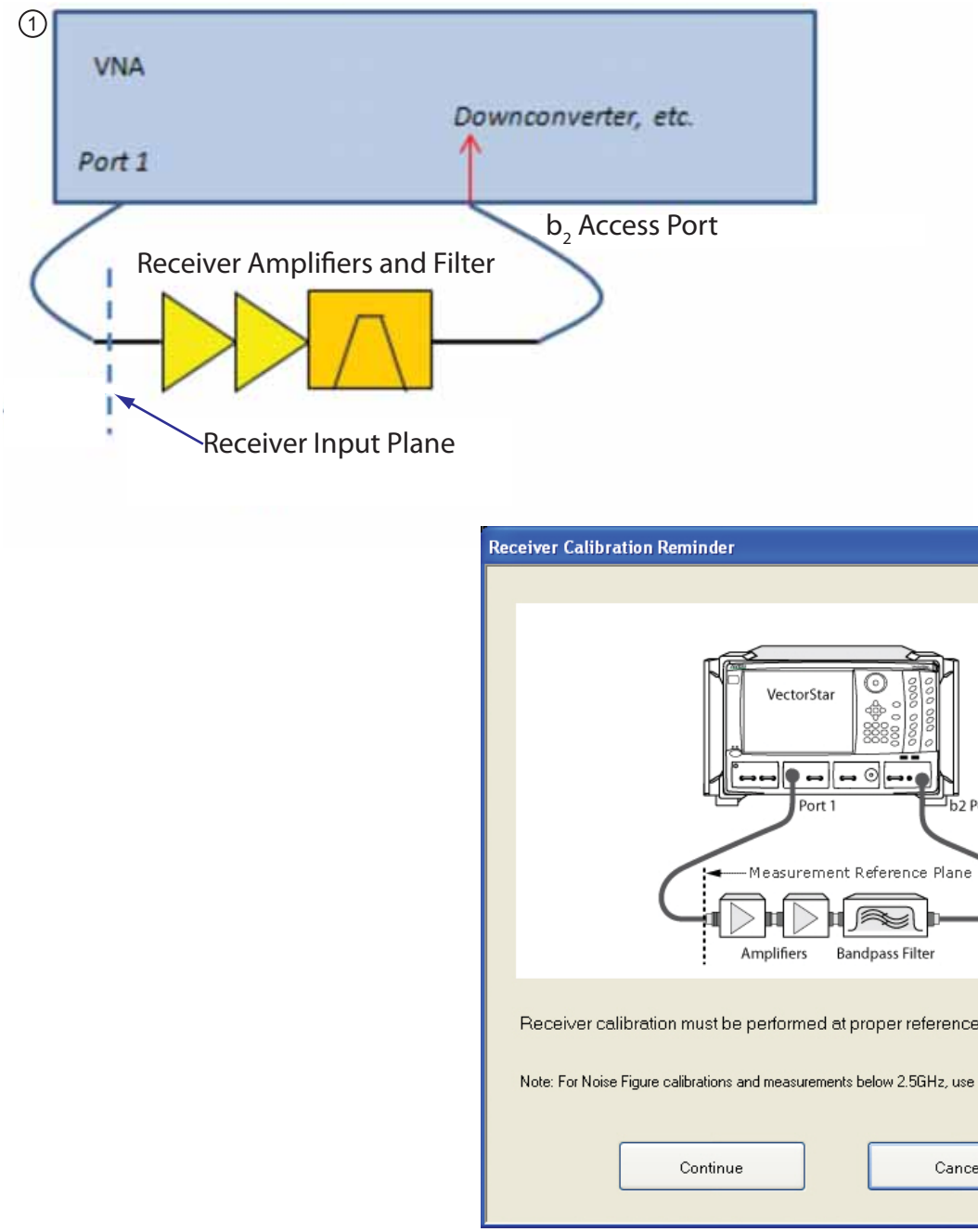
  The number of measurements used in the noise power computation.

- **Temperature**

  This is nominally the temperature of the cold termination used during the measurements and defaults to 290 K. The IEEE definition specifies this temperature, which may be different from the physical termination temperature. For reference:

  - 290 K = 17 ºC (62.6 ºF)
  - 0 K = −273.15 °C (−459.67 °F)

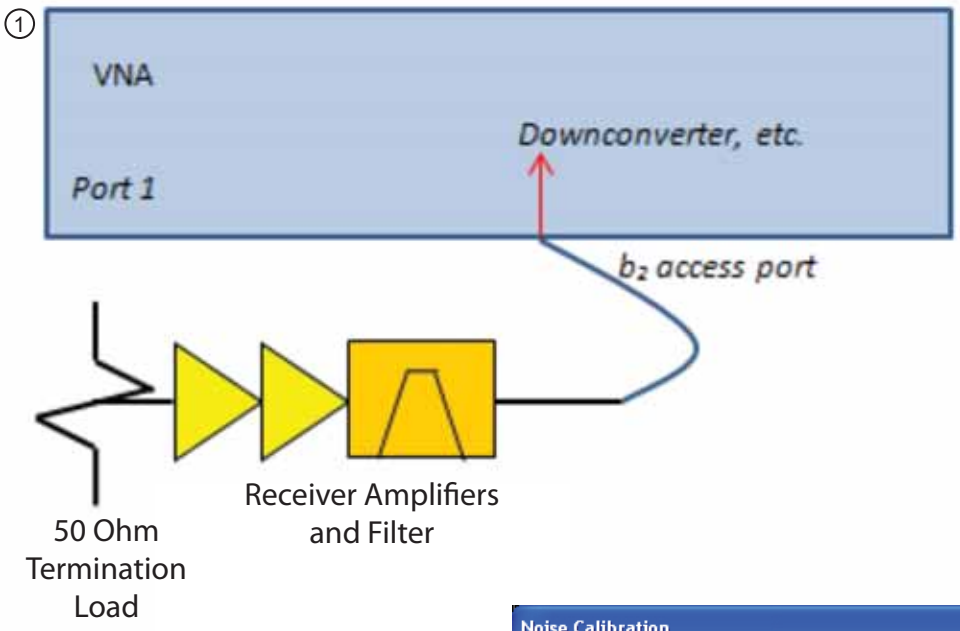
The receiver calibration in Step 5 is performed by simply connecting a cable from the source (usually Port 1) to the receiver input plane as suggested in Figure 19-9. The power calibration discussed earlier and in Step 2 should ideally be performed at the end of this cable. The receiver calibration, as discussed elsewhere in this document, measures the receiver parameter and forces it to match the calibrated power level. Uncertainty addition may come at this point due to mismatch between the source and receiver and due to any residual nonlinearity in the receiver. These points will be discussed in more detail in the uncertainties section.

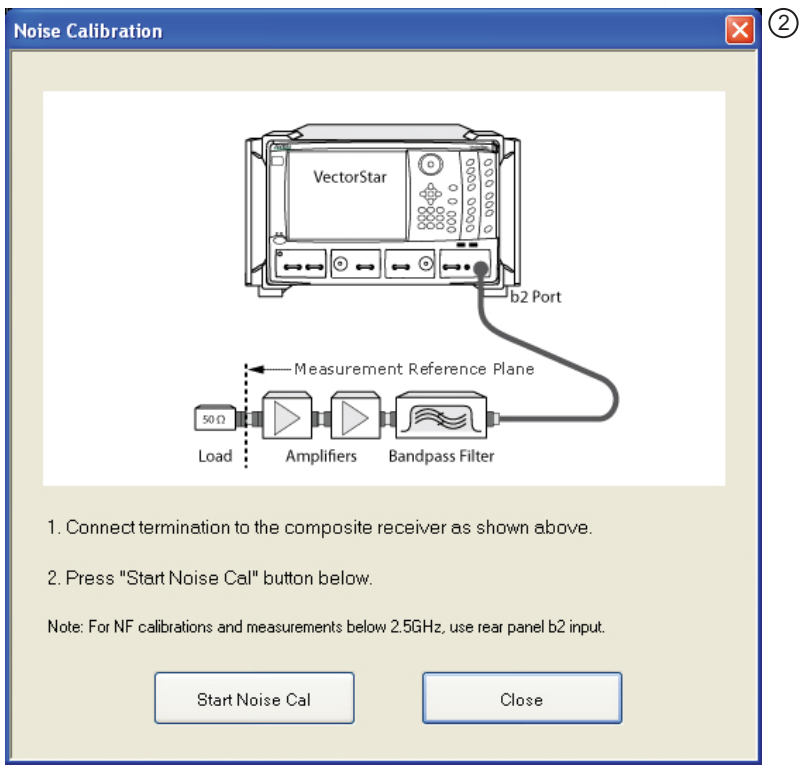


1. The receiver calibration (Step 5) is illustrated here as a schematic view. Note the VNA connections at Port 1 and the b_2 Access Port and also the Receiver Input Plane.
2. The equivalent pictorial view is shows in the RECEIVER CALIBRATION REMINDER dialog box.

Figure 19-9. Noise Calibration Step 5

The noise calibration in Step 6 is even simpler. Here one simply places the 50 ohm cold termination on the input to the receiver plane and allows the system to measure noise power. This is illustrated in Figure 19-10. Depending on how many frequency points are used, this may be the most time-consuming of the setup/calibration steps (allow on the order of 0.2 seconds to 2 seconds per frequency point depending on IFBW, averaging, and number of RMS points). If the frequency range is changed after calibration, interpolation and extrapolation of both the noise calibration data and DUT S-parameter data will be employed. Uncertainties may degrade, particularly if extrapolation is invoked by going outside the original frequency range of either step.

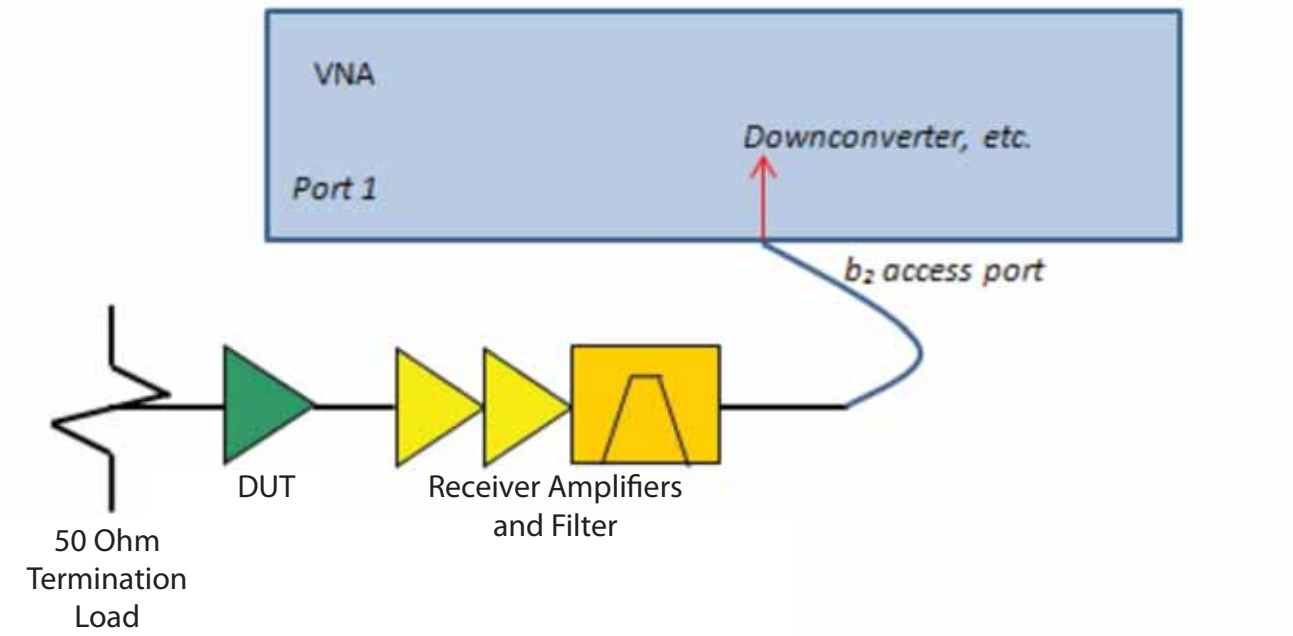




1. The noise calibration (Step 6) is shown here as a schematic view.	2. The equivalent pictorial view is shows in the NOISE CALIBRATION dialog box.

Figure 19-10. Noise Calibration Step 6

Step 7, the final step, is the measurement of the DUT itself and this is equally simple. The output of the DUT is connected to the receiver input plane and the cold noise source with a typical termination load of 50 Ohms is connected to the DUT input. This step is shown in Figure 19-11.



The final step, that of measuring the DUT (here illustrated as an amplifier), is shown here. Note the 50 Ohm Termination Load attached to the DUT.

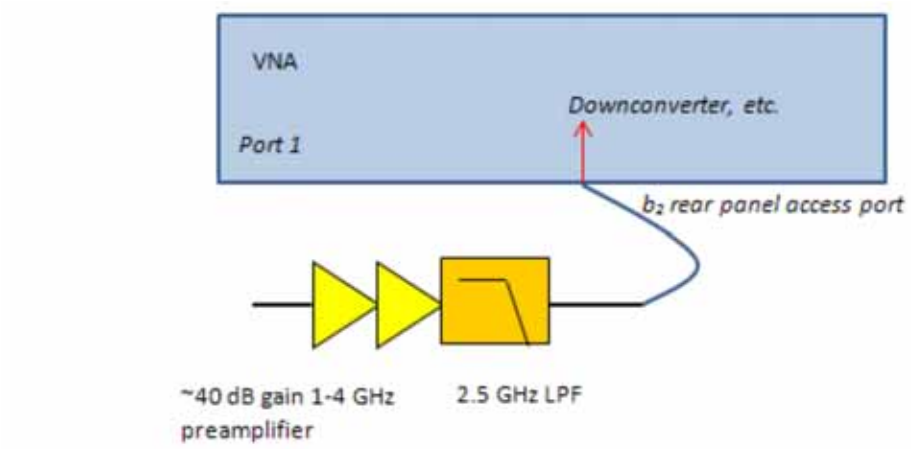
Figure 19-11. Amplifier DUT Final Measurement Step

Measurement Example

To integrate all of these concepts, let us consider an example measurement.

- DUT 1.7 to 2.4 GHz
- With an average gain of 12 dB
- A nominal noise figure of 0.8 dB.

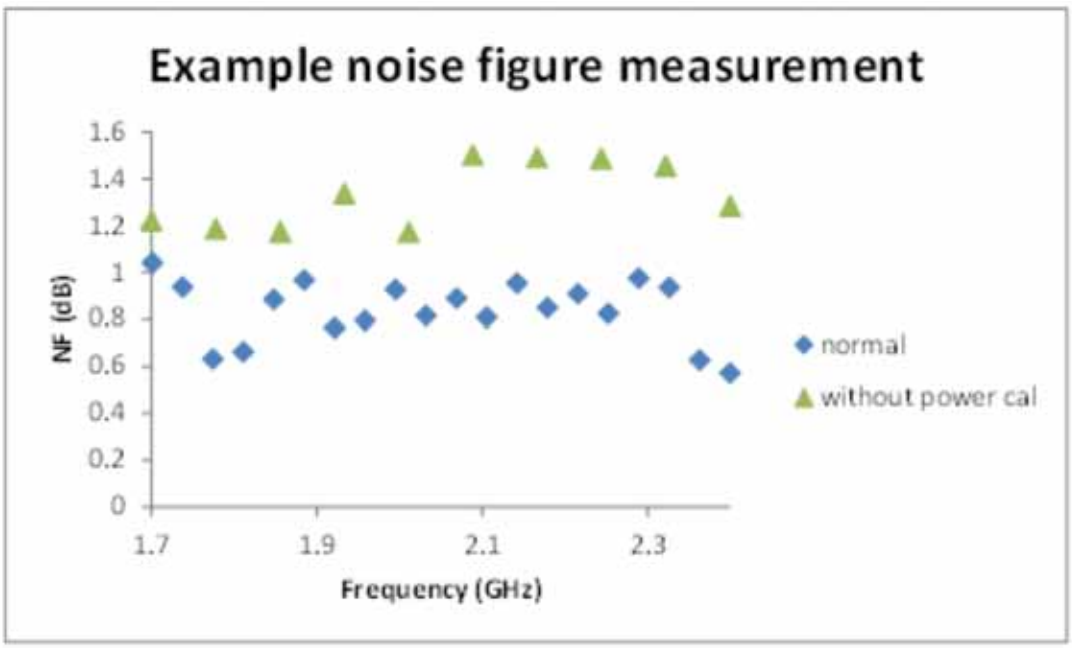
This is a challenging measurement in some sense in that the noise figure is relatively low and the gain is not very high. We will proceed with an IF Bandwidth of 1 kHz and 3000 RMS points and no averaging. The receiver structure is annotated in Figure 19-12. The compression point of this structure is about –10 dBm which should easily avoid linearity issues. The power calibration and receiver calibration will be performed at –60 dBm. The input match of the preamplifier structure was better than –20 dB and the noise parameter R_n was small enough that a 0 dB source match only results in a 0.1 dB receiver noise figure deflection. Since all frequencies are below 2.5 GHz, the rear panel b_2 access port is used.



The example measurement receiver structure is shown here.

Figure 19-12. Example of Measurement Receiver Structure

Because the noise figure is low, accuracy throughout the receiver calibration step is important since those errors will propagate roughly on a dB-for-dB basis. Since the basic ALC factory calibration has a nominal uncertainty of 0.5 dB to 1 dB, using this in lieu of a user power calibration could have a significant impact on this measurement. The difference in results is shown in Figure 19-13.



The example measurement is shown here with and without a power calibration preceding the receiver calibration.

Figure 19-13. Receiver Calibration Example with and without a Power Calibration

For another example, a mixer will be considered. In this case, the DUT has a fixed IF of 70 MHz and an RF range of 2 GHz to 2.5 GHz (and the LO is below the RF). The mixer setup dialogs discussed earlier make this setup straightforward. The conversion gain of the DUT is approximately 10 dB. The basic measurement parameters are the same as in the previous example. A different low pass filter is used to avoid LO leakage corrupting the receiver response (the DUT in this case is already well-filtered but it is reasonable practice). A preamplifier with a 10 MHz low end was used for this setup.

19-5 Receiver Cal Offset and Embedding/De-embedding

The receiver cal offset is a file that can be used to modify the receiver calibration used in noise figure measurements when it is known that there are some power discrepancies within that receiver calibration.

Examples of this situation include:

- The only power calibration that could be performed was at a higher level than that where the receiver calibration was performed and one wishes to correct for the frequency response of a pad/attenuator (or other means of power reduction) used for the receiver cal.
- The source used for the receiver calibration had some harmonic content that could cause inaccurate power readings relative to the receiver calibration readings. Knowledge of these harmonic levels can be used for a correction.
- There are other known inaccuracies with the power calibration (due to fixturing, bandwidth, etc.) that one wishes to correct for.

A common scenario for using this function is in mmWave noise figure measurements when high dynamic range power sensors are not available (bullet one above) and a factory receiver calibration of the mmWave module is available to help. This case will be covered in detail at the end of the section.

The receiver cal offset process simply involves loading a file. The menu button to load the file is shown in Figure 19-14. Once a file is loaded, a check box appears and the correction then can be turned off and on (also shown in Figure 19-13).

The file format is a standard text file (.txt) and is the result of a Save Data operation with a single trace enabled in Log Mag display format.



Figure 19-14. Advanced Noise Figure Menu: Receiver Cal Offset

The response parameter is $b_{2/1}$ in this case but can be anything appropriate (usually will be $b_{2/1}$ or S_{21}) to describe the response. The data in this file should represent the power error incurred when doing the receiver calibration: a value > 0 dB means the actual power was higher than the system detected during the receiver cal and a value < 0 dB means the actual power was lower than the system detected.

A segment of an example text file is shown below. This is the file format used for the receiver cal offset correction. The response parameter can be anything but will most commonly be S_{21} or $b_{2/1}$.

```
MS4647B
!10/23/2013.12:32:31.PM
!CHANNEL.1
!TR.MEASUREMENT
!RF.CORRECTION.OFF
!AVERAGING.OFF
!IF.BANDWIDTH: 10HZ
!NUMBER.OF.TRACES: 1
!TRACE:    TRACE.1
!PARAMETER: B2/1
!PORT: PORT1
!GRAPH: LOGMAG
!SMOOTHING: OFF
!TIMEDOMAIN: OFF
!SWEEPTYPE: FREQ.SWEEP(Linear)
PNT  FREQ1.GHZ  LOGMAG1
1   80.001000000  -8.735106E+000
2   80.150995000  -6.820481E+000
3   80.300990000  -6.195999E+000
4   80.450985000  -6.687141E+000
5   80.600980000  -6.155944E+000
6   80.750975000  -6.188873E+000
7   80.900970000  -6.445963E+000
8   81.050965000  -6.779163E+000
9   81.200960000  -6.733072E+000
10  81.350955000  -6.922758E+000
11  81.500950000  -7.044048E+000
12  81.650945000  -7.252111E+000
13  81.800940000  -6.779829E+000
14  81.950935000  -7.133829E+000
15  82.100930000  -7.282691E+000
16  82.250925000  -7.258348E+000
17  82.400920000  -7.333891E+000
18  82.550915000  -7.438091E+000
```

```
19  82.700910000  -7.777820E+000
20  82.850905000  -7.632741E+000
21  83.000900000  -7.665494E+000
```

Examples

1. An uncompensated adapter with 1 dB of loss was used for the receiver calibration but was not present during the power calibration. The power seen by the receiver was 1 dB lower than what the power calibration is reporting.

   In this case the values in the file will all be around −1 dB. This can be measured with a simple S-parameter measurement of the adapter (using adapter removal or reciprocal techniques as necessary if the connections required are not mateable).

2. Another example is in mmWave systems where power sensors that operate at the desired levels for the receiver calibration (which may be −50 dBm or lower) are rare.

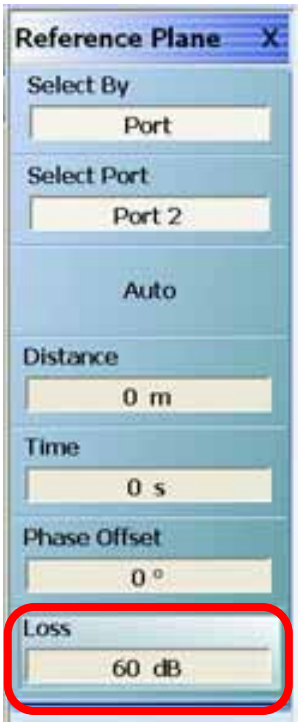
If measurements are performed with the 3744A-Rx direct receiver module, there is a correction path available. This module is usually supplied with a factory receiver calibration file that one can use to back-propagate accuracy to the source power so that when one performs the 'real' receiver calibration for noise figure (with a pre-amplifier/filter assembly attached), the correct receiver calibration values can be obtained indirectly.

Creating the Receiver Cal Offset File

1. Connect the desired source reference plane to the 3744A-Rx module at the target power level (−60 dBm for this example).
2. Load the provided factory receiver calibration file.
3. Set the frequency range to the one desired for the noise figure measurement (with an adequate number of points) and set the display for single trace, $b_{2/1}|P_1$, log mag.
4. Set the reference plane magnitude to correct for the intended power level (60 dBm in this case) as shown in Figure 19-15 on page 19-22. The resulting data then represents an error from the ideal power level and helps to describe the actual source power in a nominal bandwidth of interest.

   Navigation:

   MAIN | Measurement | MEASUREMENT | Reference Plane | REFERENCE PLANE
5. Save the resulting data as a text file. This step creates the receiver cal offset file for the measurement.
6. Connect the rest of the assembly for the noise figure measurement (pre-amplifiers and filters normally) between the 3744A-Rx module and the source reference plane, leaving the power setting the same.
7. Perform the receiver calibration on this setup and save the results as a text file.
8. Load the receiver calibration file from Step 5 and the receiver cal offset file from Step 7 when performing the noise figure calibration and measurement.
9. Load the DUT S-parameters and perform the noise calibration.



The reference plane adjustment for the measurement is shown outlined in red.

Figure 19-15. REFERENCE PLANE Menu with Adjustment

Within the noise figure application, the Network Extraction Embedding and De-embedding recall utilities allow one to load .s2p files describing networks associated with the receiver calibration process that may need to be added or removed from the measurement. Common examples are in on-wafer measurements (for both embedding and de-embedding):

- A power calibration was performed coaxially but the desired receiver reference plane is at a probe tip so a second probe is required to get the calibrated power to that reference plane. The loss of this second probe must be de-embedded from the result.
- A receiver calibration has been performed coaxially but then one wishes to move that receiver calibration reference plane to a probe tip. In this case, that probe tip loss must be embedded.

Both recall utilities read .s2p files and will interpolate and flat-line extrapolate as required for the measurement. If extrapolation is being used, a warning dialog will appear. Both of these files, when turned on, directly affect noise power as well as noise figure so the impact will be observed even when uncalibrated. The two file recalls are independent and they can be used simultaneously. As this is a power correction and the system has no knowledge of the environment where the recalled network is being placed, it is only a scalar correction and only |S21| data is used.

19-6 Uncertainties

As has been alluded to in the previous sections, many factors affect noise figure uncertainty and an understanding of their impacts and interactions can be important in setting up and performing a high accuracy noise figure measurement. The purpose of this subsection is to explore these effects and to produce a more realistic picture of the uncertainties one can expect and how to improve them.

Looking back at the noise figure equation (Eq. 19-4 above), one can recast it to show how component uncertainties affect the result. In Eq. 19-6 above, the Δ terms represent deviations that may occur during the measurement or the calibration. Not all possible deviations are included here but rather only those that are most dominant practically are shown.

$$\tilde{F} = \frac{1}{G + \Delta G} + \frac{(N_{DUT+rcvr} + \Delta N_{DUT+rcvr} - N_{rcvr} - \Delta N_{rcvr}) \cdot (1 + \Delta R)}{kT_0 B(G + \Delta G)}$$

Equation 19-7.

The Gain variance is perhaps the easiest to understand. Gain errors can arise in the usual s-parameter uncertainty context (or the equivalent for a frequency converting device) and this uncertainty is usually what is included. Errors due to linearity faults or other structural measurement problems are generally assumed to be taken care of by the user.

The other potential terms relate to the measurement of noise power and those are split into two groups: those common to both DUT measurement and calibration noise measurement (represented by ΔR) and those unique to the individual noise measurements (the two ΔN terms). All of these have several sub-terms.

- **ΔR**

  Terms related to the user power calibration (power sensor accuracy and mismatch) and to the receiver calibration (match of the receiver, match of the source, linearity and noise floor).

- **ΔN**

  Match interactions (termination with the receiver, termination with the DUT, DUT with the receiver), power measurement linearity, jitter.

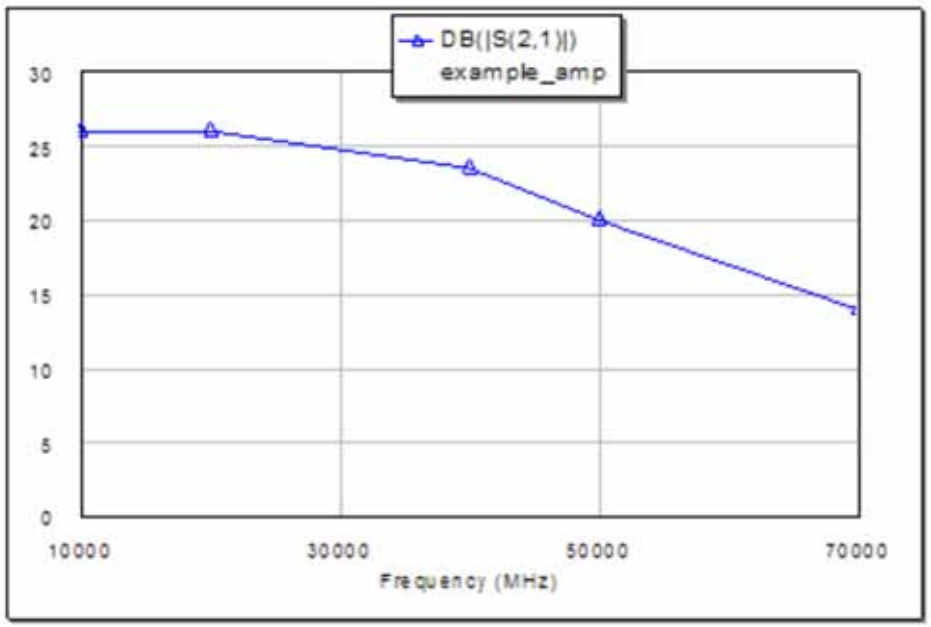
These terms in turn have many dependencies on DUT S-parameters/gain, DUT noise figure, receiver noise floor, receiver linearity, source match, absolute signal levels, power sensor characteristics, etc. Because so many variables are involved, an uncertainty calculator has been created to help in analyzing some of the scenarios. A Monte Carlo approach is used since many of the variables have random distributions (particularly in phase) so a statistical analysis of the uncertainty may be more practically useful and is increasingly accepted as a preferred approach (e.g., [8]).

A few general observations can be made:

- The better matched the DUT and the receiver, all else held constant, the lower the uncertainties
- The higher the DUT gain and receiver gain (up until the point of linearity issues), the better the absolute (in dB) uncertainties
- Lower IFBWs and higher RMS point counts help (in terms of jitter) but only up to a point
- In terms of DUT parameters, the worst-case scenario is a poorly matched, low gain, low noise figure device. The best-case scenario is a well-matched, high gain device.

To understand the parameter space and to illustrate some of the computation options that are available, we will run through some examples.

Consider an example amplifier with the following gain profile:



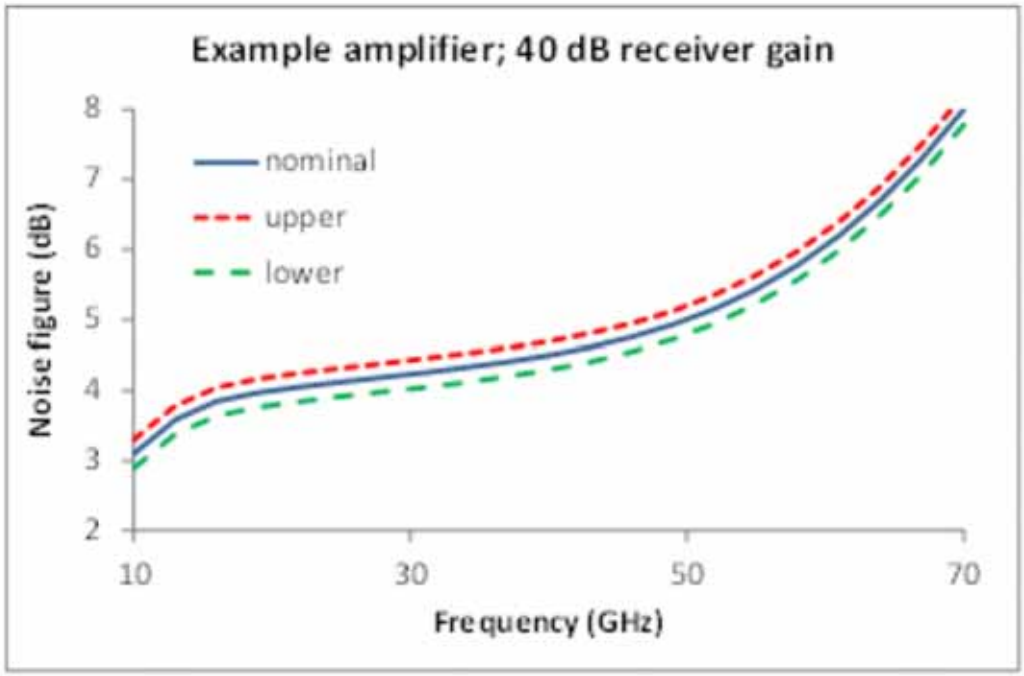
The |S21| behavior with frequency for the example amplifier is shown here.

Figure 19-16. Behavior of |S21| with Example Amplifier

The DUT noise figure ranges from 3 dB at 10 GHz to 8 dB at 70 GHz with a frequency dependence that will be shown shortly. Consider the following example parameters:

- Starting IF Bandwidth: 1 kHz
- Averages: 1
- Number of RMS points: 3,000
- Temperature: 290 K
- Termination match: –30 dB
- Power sensor accuracy: 0.1 dB
- Power sensor match: –20 dB
- Source match: –20 dB
- Receiver match: –20 dB
- Receiver P1dB: 0 dBm
- Receiver calibration power level: –60 dBm
- Starting receiver gain: 40 dB

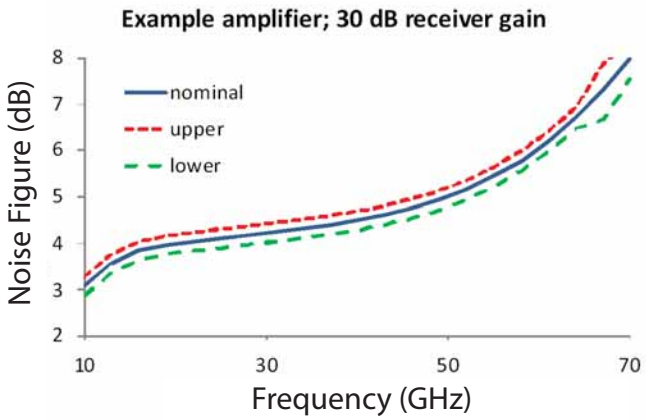
The first plot in Figure 19-17 shows the nominal DUT noise figure (solid line) and the upper and lower 95% uncertainty bounds if the receiver gain is 40 dB.



Example uncertainties are shown for the case of receiver gain = 40dB.

Figure 19-17. Uncertainities for 40 dB Receiver Gain

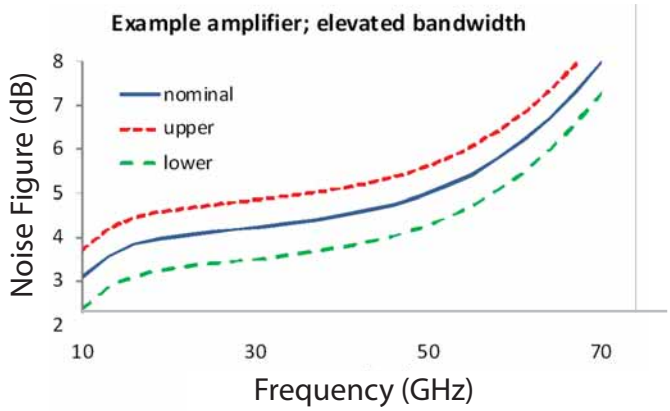
Next the receiver gain is reduced. Note that there is little change except at the higher frequencies where the DUT gain is low. This is a general concept that the DUT gain + receiver gain is a normative parameter.



The uncertainties for the same measurement as in Figure 19-17 above are shown here except the receiver gain has been reduced.

Figure 19-18. Uncertainties with Reduced Receiver Gain

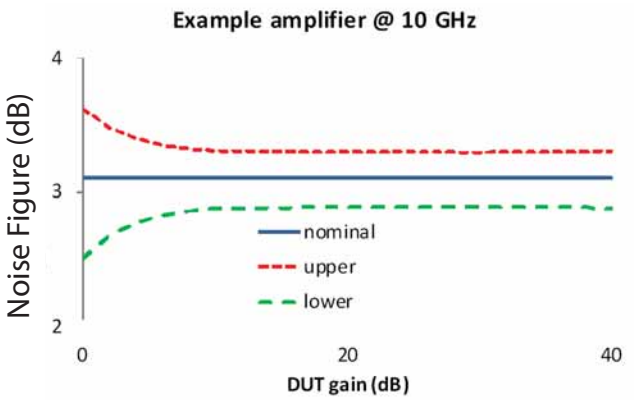
Next, let us return the receiver gain to 40 dB but increase the measurement bandwidth to 30 kHz. This increases the data jitter and the net uncertainty.



The uncertainties are shown here for the case of Figure 19-17 above except the IF bandwidth has been increased.

Figure 19-19. Uncertainties with Increased IF Bandwidth

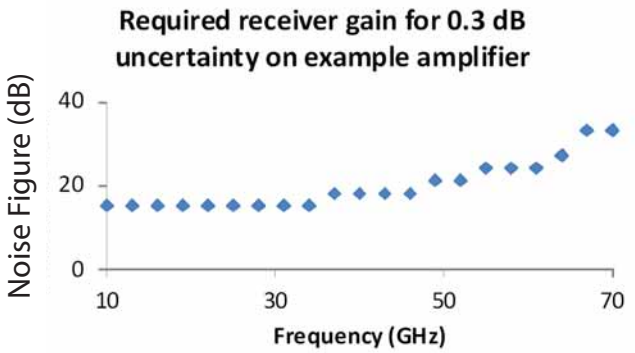
We can also look at a different independent variable. In Figure 19-20, the receiver gain and IF Bandwidth have been returned to their starting values but we will plot versus DUT gain for a single frequency (10 GHz). As might have been surmised from the Figure 19-16 discussion, if the gain gets low enough (in concert with receiver gain), the uncertainty will increase.



A plot of uncertainty as a function of DUT gain is shown here for the parameters discussed in this section at a fixed frequency.

Figure 19-20. Uncertainty as a Function of DUT Gain Plot

Finally, one may want to know the receiver gain for a given level of uncertainty. This is plotted in Figure 19-21 for our example amplifier (using nominal parameters) but with a 0.3 dB desired uncertainty. There is a step-like quality to the plot since the possible receiver gains are parameterized in ~2 dB steps for simplicity. As might be expected, as the DUT gain drops at higher frequencies (and noise floors increase in the VNA slightly), additional receiver gain is needed to hold the uncertainty constant.



A plot of required receiver gain for a given level of uncertainty for the example amplifier is shown here.

Figure 19-21. Required Receiver Gain Plot

19-7 Noise Figure Uncertainty Calculator

The VectorStar Noise Figure Uncertainty Calculator is a standalone program used to provide the user a tool for easily determining measurement uncertainty while using the VectorStar VNA Option 41—Noise Figure. The calculator can be used for non-frequency converting active devices, such as amplifiers. The uncertainty calculator takes into account the five primary parameters of the system configuration:

- Receiver parameters
- DUT parameters
- Power measurement parameters
- Source parameters
- Cold source temperature

In addition, the calculator is helpful in determining the optimum measurement setup for achieving a desired measurement uncertainty target.

Installation

The calculator is installed on the VectorStar instrument desktop and can also operate on an external Windows XP or Windows 7 PC as a standalone executable that is launched from the Program Menu or the desktop Icon shown below.

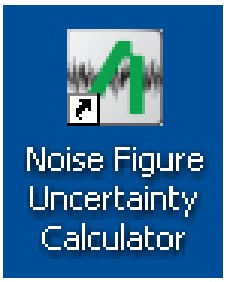


Figure 19-22. Desktop Icon in Windows for Noise Figure Uncertainty Calculator

To start the standalone Noise Figure Uncertainty Calculator installation on a local PC, select the Setup_XXXX.exe file (where the XXXX provides the calculator revision) and follow the installation dialogs.

Best practices recommend pinning the calculator to the Start Menu and to also create a short-cut on the desktop.

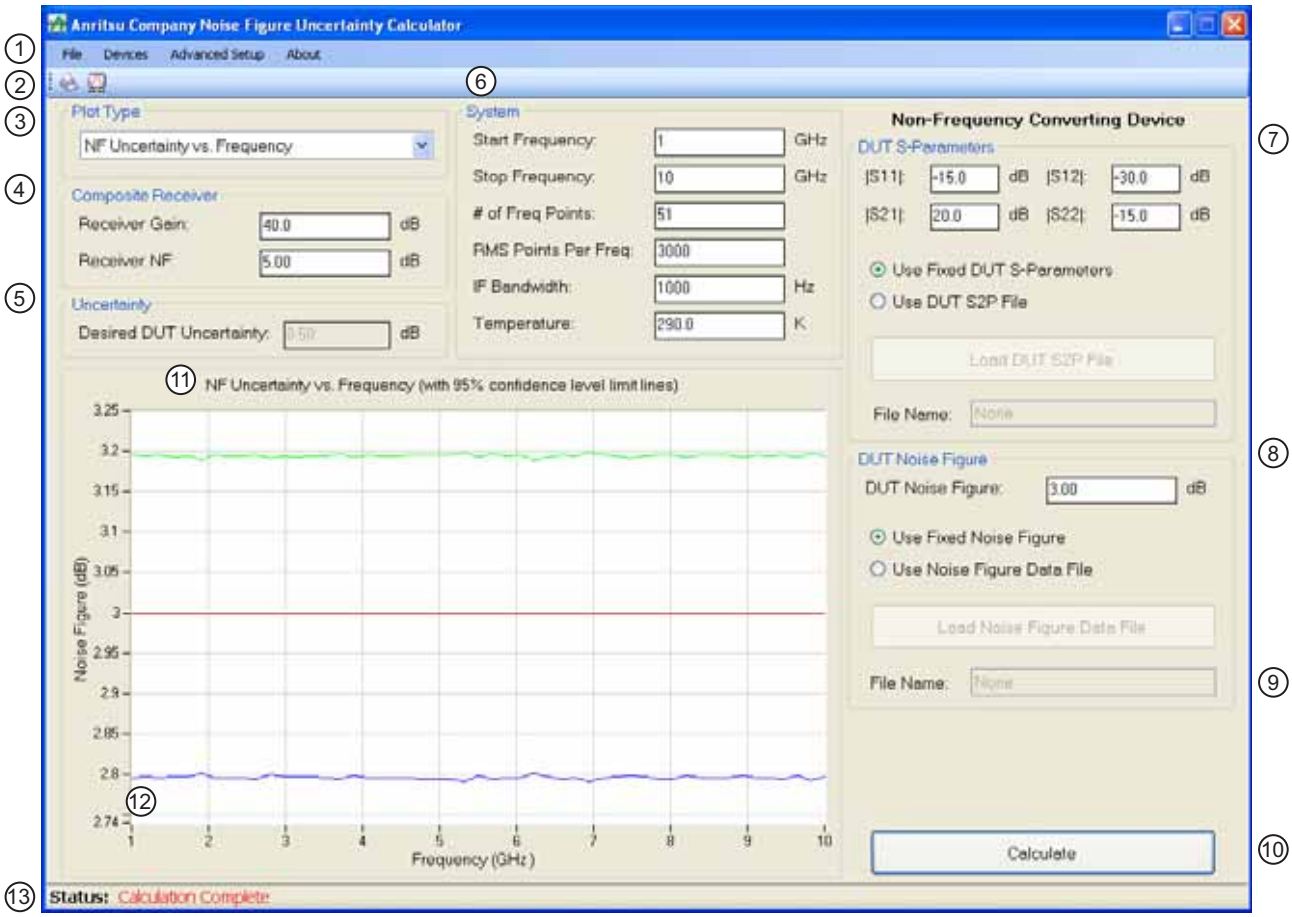
For more information, refer to the Installation Guide contained in the InstallationGuide.txt file, which is provided with the Uncertainty Calculator download package.

User Interface

The program user interface provides editable fields for the most common noise figure measurement test setup variable parameters, such as composite receiver and DUT information. When the Calculate button is pressed, the input is used to determine the overall measurement uncertainty. Conversely, a desired DUT uncertainty value can be input and the calculator will provide a display of required receiver performance (gain and noise figure) for a given DUT performance.

The Advanced Setup menu provides the ability for further modification of the measurement parameters and additional input such as receiver, source, and power sensor match.

The figure below shows the various user configuration and calculation controls. A detailed control description is in the section following.



1. Menu Bar selections for File, Devices, Advanced Setup, and About—Selection displays a drop-down menu or sub-dialog.
2. Icon Bar functions—Selection toggles between:
   - Print Window dialog
   - Markers—Activates user movable plot markers.
3. Plot Type—Select displays pull-down menu for:
   - NF Uncertainty vs. Frequency
   - NF Uncertainty vs. Gain
   - Receiver Gain vs. Frequency
4. Composite Receiver Configuration Area
5. Uncertainty Configuration Area
6. System Configuration Area
7. DUT S-Parameters Area—Start of the Non-Frequency Converting device type controls.
8. DUT Noise Figure Area
9. Noise Figure Data File Name—If Use Noise Figure Data File is selected, file name appears here. A standard VectorStar TXT format file is required.
10. Calculate execute button
11. Graph Display Title—Determined by the selection in Plot Type.
12. Display Graph—Axis labels and range determined by user settings. Graph values change when Calculate button is selected.
13. Status Message—Shows calculator status such as “Calculation Complete” or “Calculation in Process”

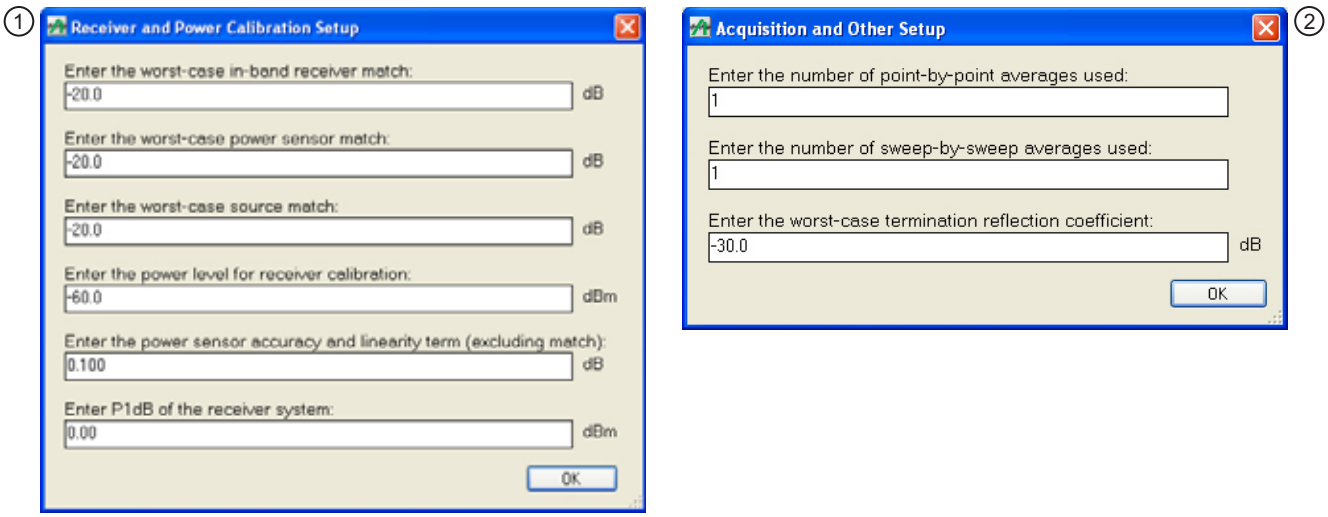
Figure 19-23. UNCERTAINTY CALCULATOR Dialog Box

Uncertainty Calculator Controls

The uncertainty calculator provides user interface (UI) controls in a Menu Bar, an Icon Bar, and in several dialog box areas.

Menu Bar Controls

- **File**—Drop-down menu provides selections for:
  - Save Uncertainties—Saves the calculated uncertainty data as a TXT file.
  - Print Window—Displays the standard print dialog for available printers.
  - Save Window—Saves the current Uncertainty Calculator UI display as a PNG file.
  - Preset—Clears the current user-defined configuration and resets all values to their default state.
  - Exit
- **Devices**—Allows selection of device type as:
  - Non-Frequency Converting—Select also configures the right side controls as Non-Frequency Converting Devices.
- **Advanced Setup**—Allows selection of advanced setup in sub-dialog boxes as:
  - Receiver and Power Calibration Setup—Select displays the same named sub-dialog box as shown in the left side of Figure 19-24 below.
  - Acquisition and Other Setup—Select displays the same named sub-dialog box as shown in the right side of Figure 19-24 below.



1. Receiver and Calibration Setup—Provides input fields for: • Worse-Case In-Band Receiver Match dB • Worse-Case Power Sensor Match dB • Worse-Case Source Match dB • Power Level for Receiver Calibration dBm • Power Sensor Accuracy and Linearity Term dB • Receiver System P1dB dBm	2. Acquisition and Other Setup—Provides input fields for: • Number of Point-By-Point Averages Used • Number of Sweep-by-Sweep Averages Used • Worse-Case Termination Reflection Coefficient dB

Figure 19-24. Advanced Setup Sub-Dialog Boxes

- **About**—Displays a sub-dialog box that provides version number and location of additional sources of information as shown below in Figure 19-25.

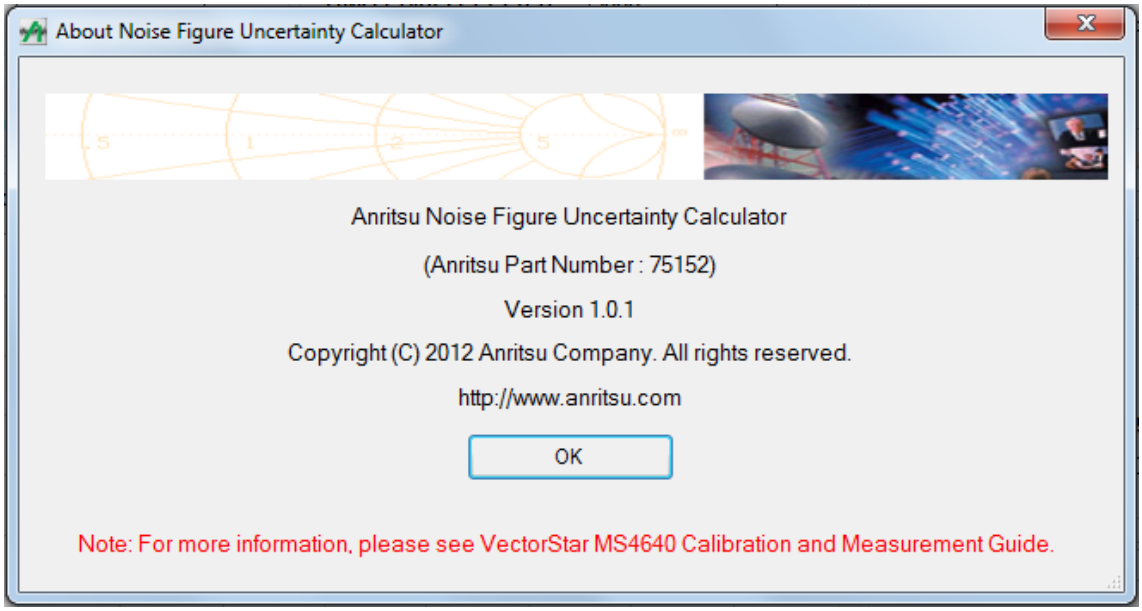


Figure 19-25. About Noise Figure Uncertainty Calculator Dialog Box

Icon Bar Controls

The Icon Bar provides two controls for fast access to print and marker functions:

- Print Icon—Select opens the Print Window dialog box to the default local or network printer. The output is sized to fit on standard A4/8.5 in × 11 in letter-sized paper.
- Marker Icon—Select toggles the plot marker on. The second selection toggles the markers off. If markers are moved and then turned off, turning them back on restores their previous positions.

Dialog Box Controls

These control descriptions are organized by the dialog box area name. Where noted, some controls are only present when the appropriate converting mode is selected.

Plot Type Area

A pull-down menu provides selection between:

- NF Uncertainty vs. Frequency—Display NF uncertainty as a function of frequency.
- NF Uncertainty vs. Gain—Display NF uncertainty as a function of DUT gain.
- Receiver Gain vs. Frequency—For a desired DUT NF uncertainty.

The plot type selection updates the label on the display graph after a calculation is performed.

Composite Receiver Area

Provides input configuration of the VNA plus additional user supplied components:

- Receiver Gain—Provides an input field for gain in dB. Typically used with multiple physical external amplifiers. Receiver Gain is unavailable if Receiver Gain vs. Frequency is selected in Plot Type.
- Receiver NF—Provides an input field for noise figure in dB. Typically dominated by the noise figure of the amplifier closest to DUT.

Uncertainty Area

Provides input of the Desired DUT Uncertainty in dB. Only available if the **Plot Type** above is set to Receiver Gain vs. Frequency.

System Area

Provides input configuration of the relevant VNA parameters:

- Start Frequency GHz
- Stop Frequency GHz—Stop Frequency must be greater than the Start Frequency.
- # of Freq Points—Number of Frequency Points between the Start and Stop Frequencies. Must be 2 points or greater.
- RMS Points Per Freq—Root Mean Square Points Per Frequency Point. Sets the number of measurement points to be used for each frequency points. For example, if the default of 51 frequency points is used with the default of 3,000 RMS Points, there will be 153,000 total measurements (51 × 3,000) used in the noise power computation.
- IF Bandwidth Hz—Should be the same as that used in the receiver calibration.
- Temperature K—Nominal temperature of the cold termination, defaults to 290K (17 ºC, 62.6 ºF), specified by the IEEE definition.

DUT S-Parameters Area

This control area is present if the Menu Bar | Devices parameter above is set to Non-Frequency Converting. Select allows the user to select between entering fixed data or using a previously loaded DUT S2P file.

- Use Fixed DUT S-Parameters—Select allows the user to define and apply fixed S-parameters located in the fields immediately above for: S11, S12, S21, and S22.
- Use DUT S2P File—Select allows the user to load a previously stored S2P DUT file. If selected, the Load DUT S2P File button and the File Name field become available. Select the button and navigate to the previously stored S2P file. When complete, the loaded file name appears in the File Name field.

DUT Noise Figure Area

This control area is present if the Menu Bar | Devices parameter above is set to Non-Frequency Converting. Select allows the user to select between entering fixed data or loading measured data.

- Use Fixed Noise Figure Data—Select allows the user to define and apply a fixed DUT Noise Figure value defined in the field immediately above.
- Use Noise Figure Data File—Select allows the use to load a previously stored noise figure data file in standard VectorStar TXT format. If selected, the Load Noise Figure Data File button and the File Name field become available. Select the button and navigate to the previously stored data file. When complete, the loaded file name appears in the File Name field.

Calculate Button

Select starts the calculation after all parameters are entered. The status message at the bottom changes to Calculation in Progress. The button name changes to Abort where selection stops the calculation and returns the status to Calculation Complete. Once the calculation is complete, the display graphs and axis labels are updated and the status changes to Calculation Complete.

Typical Uncertainty Calculator Use Procedure

1. Select plot type. The plot type selection activates the required fields and controls.
2. Setup system and input device parameters.
3. Plot uncertainties by selecting the Calculate button.
4. If uncertainty improvements are needed, modify the IFBW and # of RMS Points to verify that the current receiver gain and noise figure are adequate for the desired measurement uncertainty.
5. If further improvement is needed, adjust the gain of the receiver to determine if adding additional preamps will reach the objective.
6. Alternatively, enter the desired uncertainty and compute the required receiver parameters.

7. As required, activate and place markers at frequencies of interest.
   - On the Icon Bar, select the Marker icon.
   - A marker is available for each displayed trace.
   - Drag the marker to the position of interest and read off the axis values.
8. As required, print display graph as a PNG file.
9. As required, save uncertainty output as a TXT or CSV file.

19-8 mmWave Noise Figure

mmWave noise figure considerations were briefly covered earlier, in the receiver offset section. As loops are not available in the mmWave modules (usually), the 3744A-Rx module was created that is receive-only and does not have a coupler, so the noise figure is roughly equivalent to that of the loop access ports on the base MS464XX VNA. This Rx module is normally connected in place of the port 2 module for noise figure measurements and the same (conceptually) pre-amp filter combinations in front of this module are required for all but the highest gain DUTs.

One complication relates to the power calibrations done as a precursor to the receiver calibration. The noise receiver (with all of the added gain) will compress at a relatively low power level and mmWave power sensors with range below about –25 or –30 dBm are often not available. For this reason, the receiver offset discussed earlier in this chapter is available. There are cases when the accuracy of this approach is not enough because of the harmonic content of the source leading the power sensor concept of integrated power differing from the heterodyned tuned power measured by the receiver (note that the PowerMaster series of USB sensors are also heterodyne-based and do not suffer from this problem). A more advanced version of the receiver offset process is discussed in the Application Note *Procedure for a Higher Accuracy Receiver Calibration for Use in mmWave Noise Figure Measurements*—11410-00878 that can help correct for harmonic contributions.

19-9 Mixer Noise Figure

Mixer noise figure is also supported with relatively minor changes to procedure. From a physical setup point of view, it is important that if the DUT does not intrinsically filter LO leakage to the IF/RF (output) port sufficiently (meaning that portion that can get through the receiver assembly is below about –120 dBm), additional LO filtering should be added. It does take a large discrete LO signal leakage to dramatically affect a noise measurement. From a measurement execution point-of-view, the only real change is in how DUT gain is loaded since a .s2p file is not generally available. Instead, .txt files can be loaded to bring in conversion gain information. The only requirement is that the gain data be in trace 1.

Alternatively, any .s2p file can be modified by loading conversion gain data into the S21 fields and the system can be told an amplifier is being measured (as long as any required LO programming is handled via multiple source control). The phase of S21 is not used in this case, so it can be left in any form, but the magnitude must be correct.

19-10 Conclusions

The noise figure application has been discussed in this section in terms of its algorithm, a comparison to other approaches, the setup requirements and options, and the associated uncertainties. By considering the setup needs and how they interact with the uncertainties, successful and accurate measurements are considerably more likely. For more information on mmWave noise figure measurements, see the *3744A-Rx Noise Figure User Guide*—10410-00328.

19-11 References

1. H. T. Friis, "Noise figure of radio receivers," *Proc. IRE*, vol. 32, Jul. 1944, pp. 419–422.

2. —, "Noise Figure", *Anritsu Application Note 11410-00210*, Aug. 2000.

3. D. Vondran, "Noise figure measurement: corrections related to match and gain," *Microwave Journal*, Mar. 1999, pp. 22–38.

4. N. Otegi, J. M. Collantes, and M. Sayed, "Cold-source measurements for noise figure calculation in spectrum analyzers," *67th ARFTG Conf. Dig.*, Jun. 2006, pp. 223–228.

5. N. Otegi, J. M. Collantes, and M. Sayed, "Receiver noise calibration for vector network analyzer," 76th *ARFTG Conf. Dig.*, Dec. 2010, pp. 1–5.

6. J. M. Collantes, R. D. Pollard, and M. Sayed, "Effects of DUT mismatch on the noise figure characterization: A comparative analysis of two Y-factor techniques," *IEEE Trans. Instr. And Meas.*, vol. 51, Dec. 2002, pp. 1150–1156.

7. L. Escotte, R. Plana, and J. Graffeuil, "Evaluation of noise parameter extraction methods," *IEEE Trans. Micr. Theory Tech.*, vol. 41, Mar 1993, pp. 382–387.

8. —, "Supplement 1 to the "Guide to the expression of uncertainty in measurement" — Propagation of distributions using a Monte Carlo method", *Joint Committee for Guides in Metrology*, 2008.

Chapter 20 — Differential Noise Figure (Option 48)

20-1 Chapter Overview

Option 48 is an enhancement of, and exclusive of, Option 41 for noise figure discussed in Chapter 19. Option 48 includes the ability to handle 3- and 4-port DUTs in a single-ended sense, differential and common mode noise figure for all DUTs with a variety of processing options, and the inclusion of a level of vector match correction to enhance accuracy. The menu system is augmented to help speed the calibration procedures. The measurement premise, that of cold source analysis, remains the same, as do most of the guiding principles on the receiver preamplifier/filtering discussed in Chapter 19.

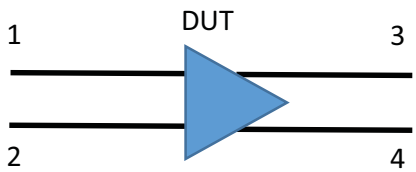
Much work has been published in the area of differential noise figure (e.g., [1]-[7]) although there is still some inconsistency on terminology and on measurement approaches. This chapter does not attempt to resolve these issues, but is instead intended to suggest one way forward that potentially has some practical advantages, is largely self-consistent, and can lead to some reasonable measurement results in a wide range of DUT configurations.

20-2 Multiport DUTs: Gain Aspects

As discussed in Chapter 19, handling of the DUT gain is an important part of noise figure computations and, in the case of 3- and 4-port DUTs, the S-parameters will come in the form of an .s3p or .s4p file. If the DUT outputs are to be treated as single-ended, then the analysis process may just be a simple extraction of the 2-port paths of interest. If it is to be treated as a differential or common-mode measurement, then the gain must be calculated in those terms. To further complicate the practicalities, the port assignment in the .s3p/.s4p file may not match how the DUT is connected to the noise measurement ports.

Port Assignments

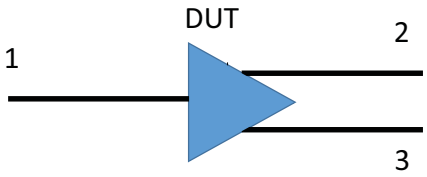
The DUT S-parameter file loader in noise figure ALWAYS uses the following port assignment for .s4p: DUT inputs are ports 1 and 2 connected, respectively, to output ports 3 and 4. When loading the file, a port swap tool is available to let the user align the port configuration used in the actual file to the port assignment (shown below) used by the noise figure application.



If the data was saved using a different port configuration, there is a port reassignment tool available.

Figure 20-1. Port Assignments When Loading .s4p Files for DUT S-parameters

For the .s3p case with an output pair, the following assignment is ALWAYS used: ports 2 and 3 form the output.



If the data was saved using a different port configuration, there is a port reassignment tool available.

Figure 20-2. Port Assignments When Loading .s3p Files for DUT S-parameters.

As with Option 041—Noise Figure, the loaded DUT S-parameters can be plotted on-screen (as static memory data) but now the 3- and 4-port S-parameters can be plotted if loaded (even if the VNA in use is a 2-port instrument). The variables are available under the RESPONSE menu.

Gain Definitions

For differential insertion gain, the mixed mode parameter S_{d2d1} is used. The mode conversion parameter S_{d2c1} is also needed and the net insertion gain is $|S_{d2d1}|^2 + |S_{d2c1}|^2$. Similarly for common-mode, the net insertion gain is $|S_{c2c1}|^2 + |S_{c2d1}|^2$. The concept on these gain configurations relies on the use of uncorrelated terminations at temperature T_0. This will be discussed further in a later section but, from a gain perspective, it means that equal noise inputs at the common and differential modes will exist. Thus, the gain of interest has to take into account the output power in a given mode due to both differential and common-mode noise inputs. The mixed mode parameters are discussed in more detail in Chapter 24, but, for the port assignments listed above, they are defined here as

$$S_{d2d1} = \frac{1}{2}(S_{31} + S_{42} - S_{32} - S_{41})$$

$$S_{d2c1} = \frac{1}{2}(S_{31} - S_{42} + S_{32} - S_{41})$$

$$S_{c2c1} = \frac{1}{2}(S_{31} + S_{42} + S_{32} + S_{41})$$

$$S_{c2d1} = \frac{1}{2}(S_{31} - S_{42} - S_{32} + S_{41})$$

Equation 20-1.

For the 3-port case with a mixed-mode DUT output, there are no mode conversion issues and the insertion gains are $|S_{d1}|^2$ and $|S_{c1}|^2$ for differential and common-mode respectively. The mixed-mode definitions of the underlying parameters are

$$S_{d1} = \frac{1}{\sqrt{2}}(S_{21} - S_{31})$$

$$S_{c1} = \frac{1}{\sqrt{2}}(S_{21} + S_{31})$$

Equation 20-2.

The use of available gain is more correct for noise figure analysis, as has been discussed previously and the multiport equivalent of this added level of vector correction is also available. These definitions are in Eq. 20-3 and, as in the 2-port case, account for the power delivery impact of DUT output mismatch.

$$G_{diff,avail} = \frac{(|S_{d2d1}|^2 + |S_{d2c1}|^2)}{1 - |S_{d2d2}|^2}$$

$$G_{cm,avail} = \frac{(|S_{c2c1}|^2 + |S_{c2d1}|^2)}{1 - |S_{c2c2}|^2}$$

Equation 20-3.

For the SE-in, diff out 3-port DUT case, things simplify:

$$G_{diff,avail,SEin-DiffOut} = \frac{|S_{d1}|^2}{1 - |S_{dd}|^2}$$

$$G_{cm,avail,SEin-CmOut} = \frac{|S_{c1}|^2}{1 - |S_{cc}|^2}$$

Equation 20-4.

The underlying definitions for the reflection mixed mode parameters used here are (again DUT output ports are 3 and 4 for the 4-port case and 2 and 3 for the 3-port case)

$$S_{d2d2} = \frac{1}{2}(S_{33} + S_{44} - S_{34} - S_{43})$$

$$S_{c2c2} = \frac{1}{2}(S_{33} + S_{44} + S_{34} + S_{43})$$

$$S_{dd} = \frac{1}{2}(S_{33} + S_{22} - S_{23} - S_{32})$$

$$S_{cc} = \frac{1}{2}(S_{33} + S_{22} + S_{23} + S_{32})$$

Equation 20-5.

For 2-port DUT analysis, the available gain used is, of course $|S_{21}|^2/(1-|S_{22}|^2)$ where 1–>2 is the path of analysis.

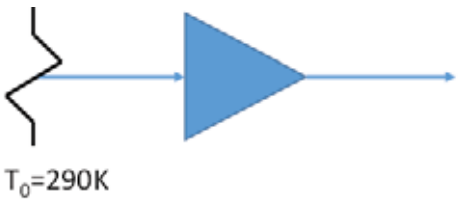
20-3 Multiport DUTs: Network Extraction Aspects

As with 2-port noise figure, it may be desirable to embed or de-embed networks from the measurement (probes, for example). With multiport DUTs, the networks to be added/subtracted can also be multiport, so some logistical complications arise as with gain. .S2p files are always supported and they are embedded/de-embedded in a simple single-ended sense where one file is associated with each port pair. For an .s4p file, the paths will be treated as single-ended as well, since mode-conversion calculations at this stage are prone to error. The user can specify which path in the file should be associated with the b_1 receiver and the disjoint path will automatically be used for the b_2 receiver.

Example: A differential probe is used and the .s4p file has dominant paths 1->2 and 3->4 (i.e., the low insertion loss parameters are S_{21}, S_{12}, S_{43} and S_{34}). The user cables what is nominally port 2 of the probe to the b_1 receiver chain. On the network extraction dialog, the user should select 1->2 as the path for the VNA b_1 receiver. This will result in S_{21} being used for embedding/de-embedding on the b_1 noise data and S_{43} will be used on b_2 noise data.

20-4 Multiport DUTs: Noise Measurement Background

Noise figure definitions present some interesting challenges, particularly in a multiport context. To start, consider the IEEE definition, which is somewhat intrinsically 2-port. It relies on the noise power out of a DUT when the input is terminated, in a reflection-less way, with a body at T_0=290K.



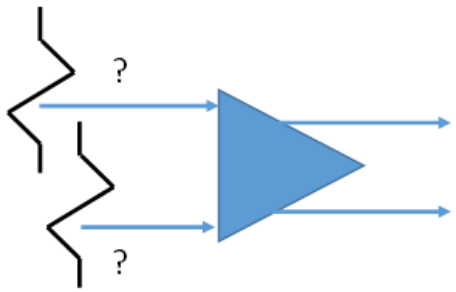
The input termination is assumed to be noisy but reflection-less.

Figure 20-3. Classical Configuration Assumed for 2-port Noise Figure Measurements

Now, as has been discussed in the literature [1], the multiport situation raises some questions:

- Can we segregate ports into ‘input’ and ‘output’ classes?
- How are the input ports terminated? Are all at T_0 or are all but one ‘noiseless’? If all are noisy, are those noise signals correlated? Does it matter if the distribution of input noise is by port or by mode?
- If measuring noise at output port M, how is output port N terminated? Can the output port be terminated indirectly (e.g., through a probe) with minimal accuracy impact?

How are noise parameter inputs to be thought of? A vector of input impedances and an input correlation matrix?



For a differential device (or for a multiport DUT in general), the existing noise figure definitions are somewhat incomplete, but reasonable choices can be made.

Figure 20-4. Choosing the Noise Figure Definition for a Differential Device

The input/output assignment question ends up having little practical impact. For a passive DUT, one can assign these on the fly and for an active DUT, it generally follows logically: those ports that have gain to some other port should be terminated in T0. As we will see later, the passive network case ends up being less confusing than it might seem.

Why should we terminate the inputs in noisy T0 terminations? One could generate a more detailed picture of a DUT by using some noiseless terminations and other combinations, but recall that noise figure is generally intended to represent a simple way (with a single value) of expressing the noise output of a DUT under some representative conditions. The more complete description is left to noise parameters. Practically speaking, it would be hard to apply noiseless terminations to certain ports for measurement and would not represent many practical use cases anyway. Extending the 2-port definition makes sense in this way. The next question is one of correlation of the inputs (and there will be much more on this later): should the input noise sources be uncorrelated or correlated in some manner? Going back to noise figure-as-a-single-number concept, it would seem to make sense to have the inputs be uncorrelated since (a) that is practical from a measurement perspective, (b) it is the most easily described configuration, and (c) the correlated case is best handled in the realm of noise parameters since there is an arbitrarily large number of possible correlated inputs.

The output port terminations again go back to the question of assignment. In the case of a device with gain, the terminations on the output ports not being measured are going to have little practical impact (assuming the DUT doesn't oscillate in this case.). In the case of a passive network, all ports not being measured should be assigned the state of 'input' and be appropriately terminated for the definitions to be consistent.

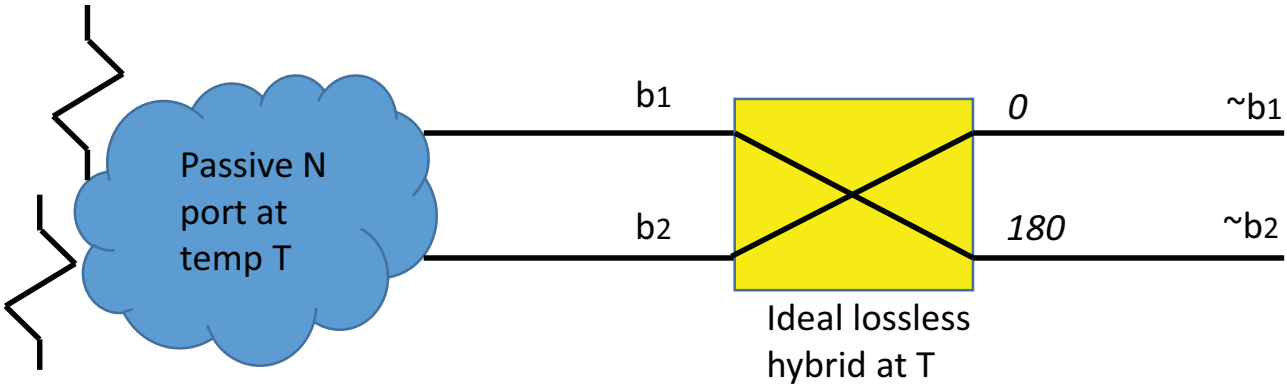
Before going further, it may be helpful to revisit the concept of correlation. Mathematically, the cross-correlation of two signals is simply (where * denotes complex conjugate)

$$\rho_{12} = \sum_{i} b_1(i) b_2^*(N-i)$$

Equation 20-6.

Where N is some positional index that might reveal some information when two signals are related but shifted relative to each other. In general, if b_1 and b_2 are noise signals from independent sources (and have zero mean), one would expect this sum to be very small in the limit of many samples since each individual $b_1 b_2$ product will be randomly phased. If, however, the two noise signals are derived from the same source, there would tend to be some common phase relationship between b_1 and b_2, so the product terms will all tend to align and the sum becomes non-zero. This is the result for two correlated noise signals.

Consider next a passive network (with T-terminated inputs) at temperature T. Ignore all reflections and connect an ideal, lossless hybrid at the output so we have some final output signals ~b_1 and ~b_2. This line of analysis was reported in [2].



A measurement thought experiment (after [2]) is sketched here to illustrate noise correlation behavior in a passive multiport at thermal equilibrium.

Figure 20-5. Noise Correlation Behavior in Passive Multiport at Thermal Equilibrium

Because everything is at temperature T (and is presumably at equilibrium), the noise powers in the various output places must be the same and equal to kTB where k is Boltzmann's constant and B is the measurement bandwidth.

$$\overline{|b_1|^2} = \overline{|b_2|^2} = \overline{|\sim b_1|^2} = \overline{|\sim b_2|^2} = kTB$$

Equation 20-7.

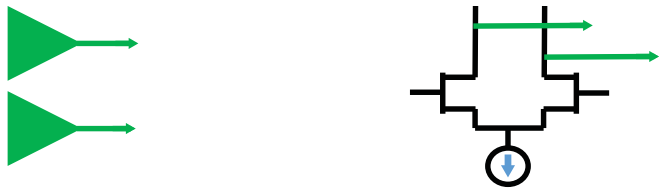
Here the overbars indicate mean summation. From Figure 20-5, ~b_1 and ~b_2 must represent the sum and difference, respectively, of b_1 and b_2. But because of the ideality of the hybrid, those output noise powers must also satisfy

$$\overline{|\sim b_1|^2} = (\overline{|b_1|^2} + \overline{|b_2|^2}) + \mathrm{Re}(\overline{b_1^* \cdot b_2^*})$$
$$\overline{|\sim b_2|^2} = (\overline{|b_1|^2} + \overline{|b_2|^2}) - \mathrm{Re}(\overline{b_1^* \cdot b_2^*})$$

Equation 20-8.

The only way both equations Eq. 20-8 can be true is if $Re(\overline{b_1 \cdot b_2^*})$=0. But this is just the real projection of correlation. If we repeat this thought experiment with a 90 degree hybrid instead of a 180 degree hybrid, we will reach the conclusion that $Im(\overline{b_1 \cdot b_2^*})$=0. This interesting result (called Bosma's theorem) is that noise power emanating from a passive multiport must be uncorrelated. Not only does this lend some further justification to our noise figure definition, but it also makes it practically much easier to analyze the effects of network losses before the DUT (probes, pads, etc.): they do not induce any input correlation.

In terms of an actual active DUT noise figure measurements, what are then the impacts of correlation? On one level are the users' system implications: how will the DUT outputs be used? If they end up feeding single-ended receivers, it may not matter much. If they feed another stage in a fully differential system, it may matter quite a lot. In terms of the DUT outputs themselves, the degree of correlation has much to do with the internal topology of the device. While the analysis cannot be exhaustive here, one can consider two extremes. Suppose a packaged differential amplifier has noise dominant stages that are actually separate single-ended amplifiers (left side of Figure 20-6). Assuming the dominant noise source within those stages is not some shared bias system, it is quite likely that the output noise will not be highly correlated.



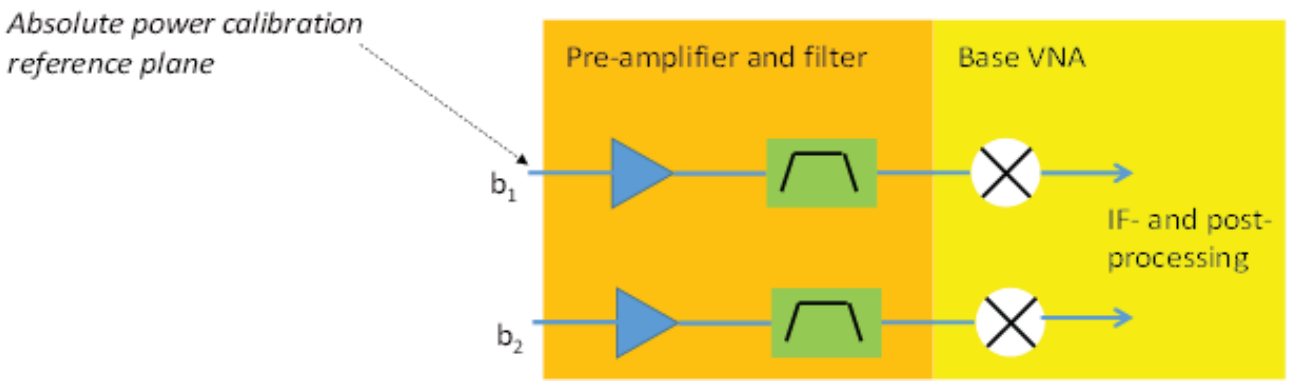
DUT topology has a large influence on the correlation of noise power between output ports. Separate dominant gain stages could make correlation negligible, but a dominant differential pair would do the opposite.

Figure 20-6. DUT Topology Influences Correlation of Noise Power Between Output Ports

If, however, the dominant noise source is an output differential pair with much of the noise derived from a common source bias (right side of Figure 20-6), then the output noise will likely be highly correlated. Depending on one's knowledge of the DUT structure, it may be desirable to choose a measurement method that fits most closely. If not much is known, then doing the additional work to ensure correlation is treated correctly may be useful.

20-5 Uncorrelated and Single-ended Measurements

If one can neglect correlation, the measurement is very simple and is a straightforward extension of the 2-port DUT analysis presented earlier in this guide. Two VNA receivers are used along with pre-amplification and filtering, as discussed. A receiver calibration establishes an absolute power reference plane for the cold-source measurements. Note that for non-mmWave measurements (i.e., those made with the base MS464XX VNA and not on one of the ME7838X systems operating in broadband or mmWave mode), the b_1 and b_2 loops on the VNA proper are used for the noise inputs, EVEN IF A 4-PORT TEST SET IS BEING USED. The instrument can remain in 4-port mode at all times, but the noise connections are not through that test set.



For multiple single-ended or uncorrelated noise measurements, the configuration follows naturally from that for the 2-port DUTs discussed in Chapter 19.

Figure 20-7. Configuration for Multiple Single-ended or Uncorrelated Noise Measurements

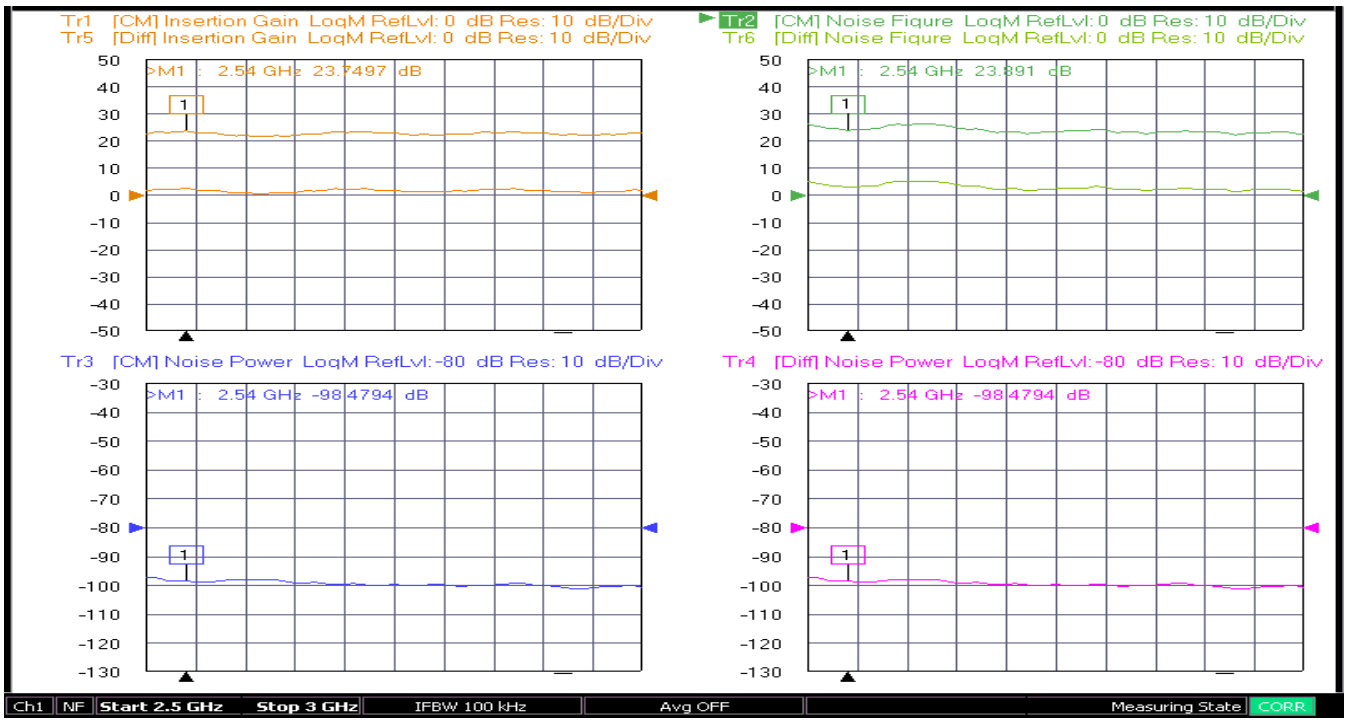
The two paths need not be identical but must obey the same rules, as discussed in Chapter 19 for overall gain, and should have similar frequency responses in the measurement band of interest. Since correlation is neglected, the differential and common-mode noise power become simply

$$\overline{|b_d|^2} = \overline{|b_c|^2} = \frac{1}{2}(\overline{|b_1|^2} + \overline{|b_2|^2})$$

Equation 20-9.

The same concepts of receiver offset and network extraction tools discussed in Chapter 19 also apply here, except independent files can be loaded for each receiver. In the case of network extraction tools, a single .s4p file can be used where paths can be assigned to each receiver.

As a basic measurement example, consider a 3-port differential-output amplifier whose noise generators are known to be relatively uncorrelated. The uncorrelated method was chosen and the receiver and noise calibrations performed. As seen in Figure 20-8 (traces 3 and 4) the common-mode and differential noise powers are the same as is forced by this method. The insertion gains for the two modes (derived from the .s3p file, using Eq. 20-2, for the DUT that was loaded) differ, however, by more than 20 dB (see traces 1 and 5). As a result, the noise figures (traces 2 and 6) also differ by more than 20 dB. The output of this particular device was well-matched so the use of available gain would not have significantly changed the result.



Although this method forces the differential and common-mode noise powers to be equal, the very different gains for the two modes cause the noise figures to also be quite different.

Figure 20-8. Three-port DUT with Uncorrelated Differential Output Measurement Example

General calibration and measurement procedure:

- Set up composite receivers feeding b_1 and b_2.
- Perform a power calibration on port 1 (if desired) at a sufficiently low level that the composite receivers will not compress. Perform a receiver calibration (using this power calibration if desired) on both receiver paths. These calibrations can be done while in the noise figure application or can be recalled.
- Perform a noise calibration on both receiver paths with terminations attached to the receiver inputs.

- Load DUT S-parameter or gain data.
- Connect DUT and measure.

The various steps are illustrated in Figure 20-9. and Figure 20-10.

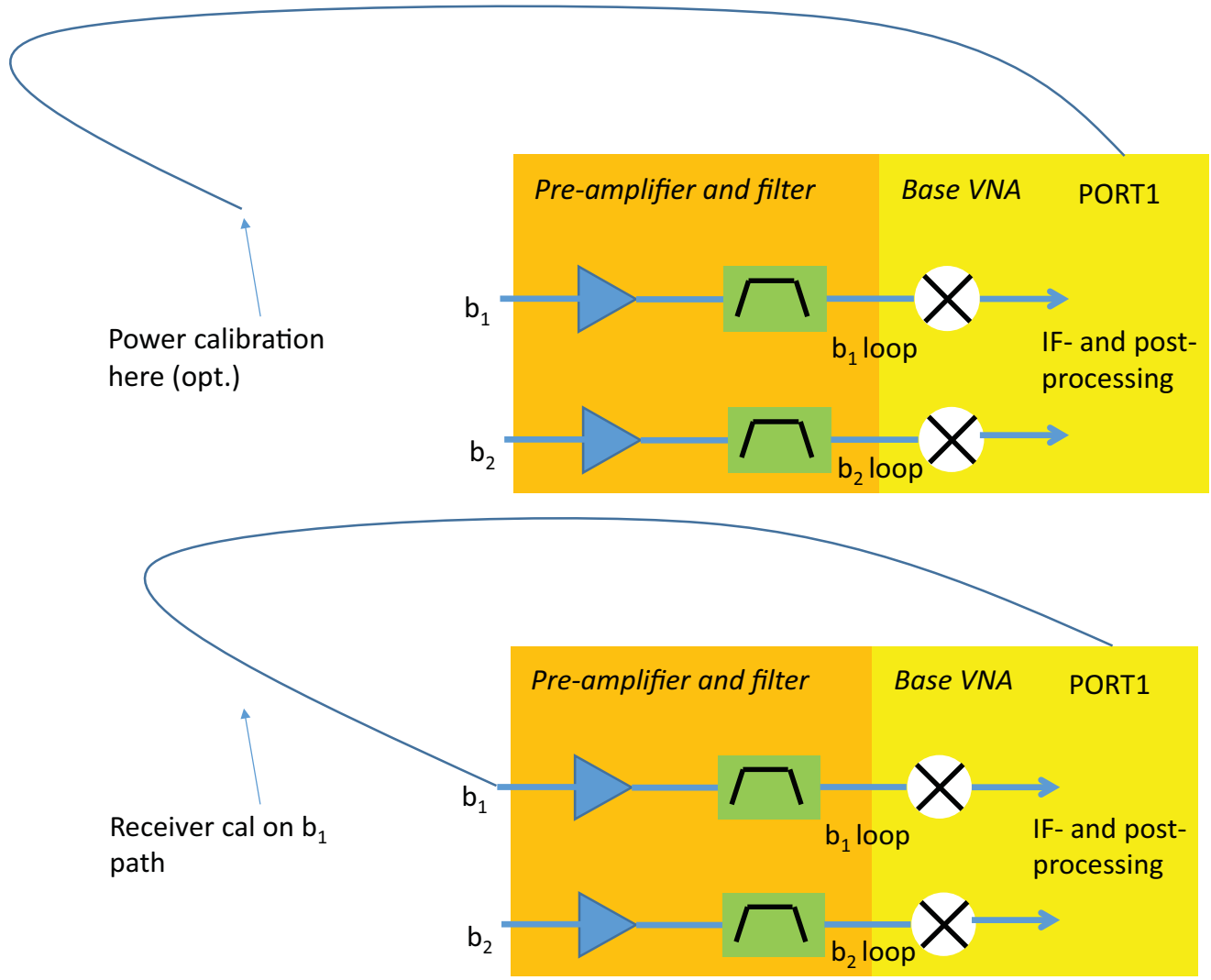


Figure 20-9. General Calibration and Measurement Steps (part 1).

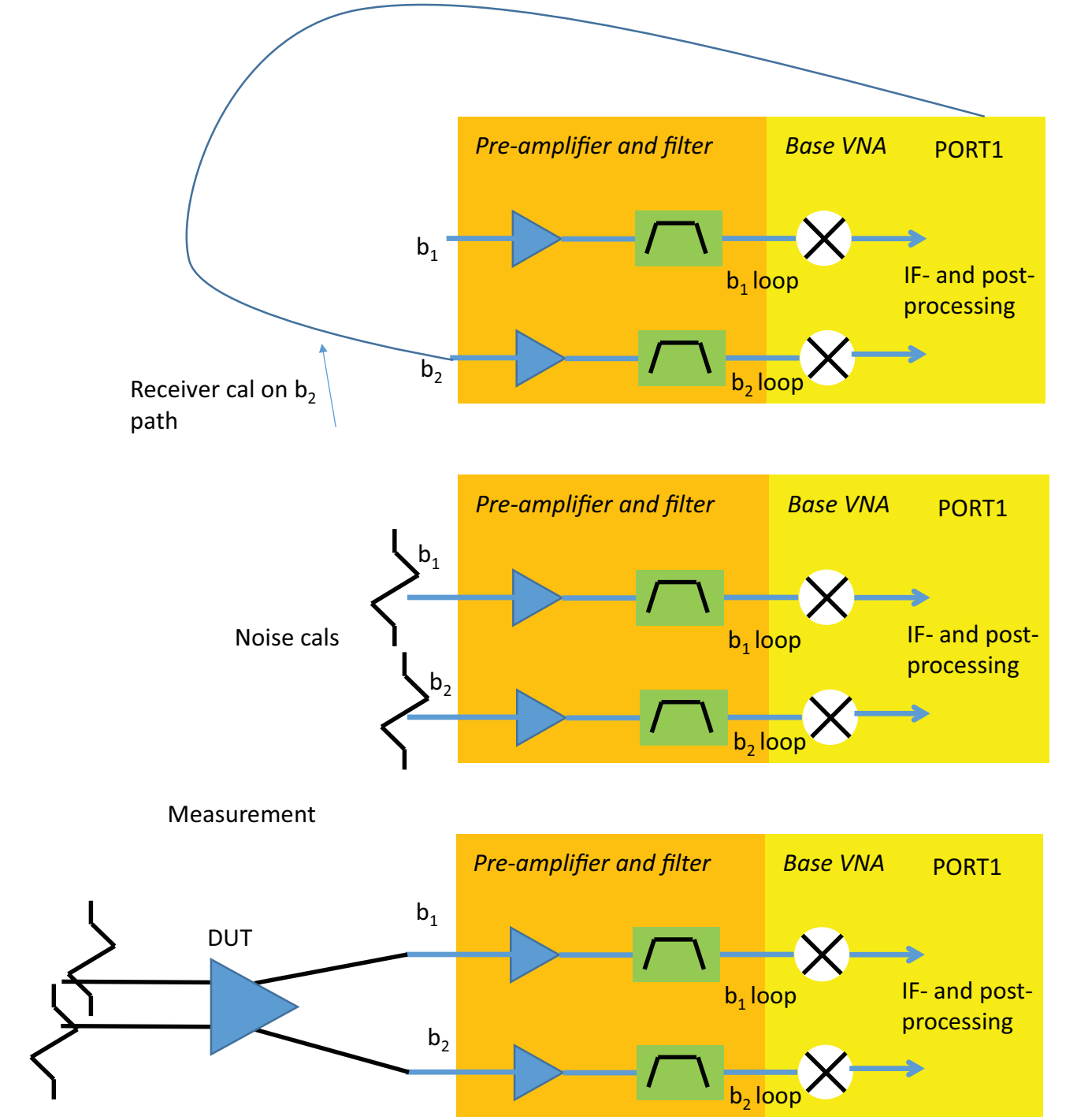
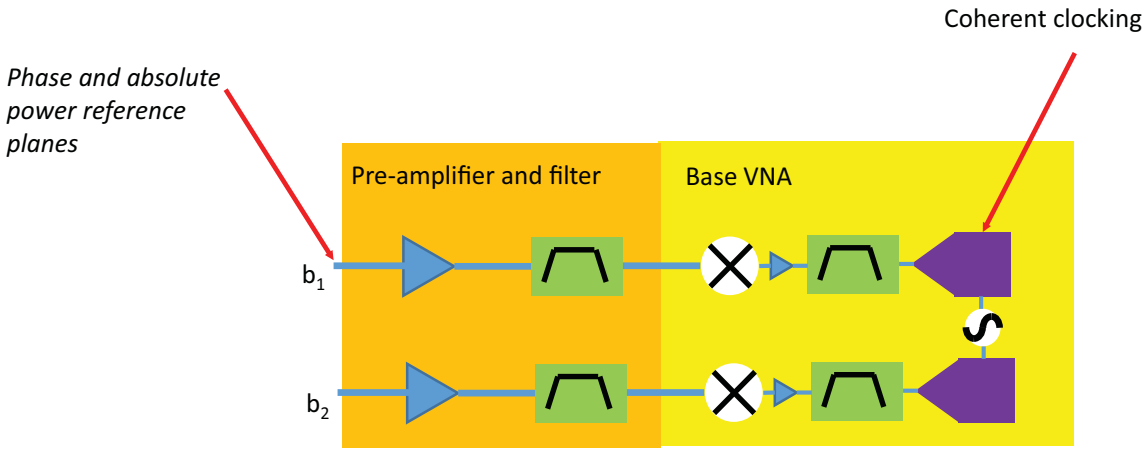


Figure 20-10. General Calibration and Measurement Steps (part 2)

20-6 Correlated Method and Coherent Receivers

Another method can take advantage of having multiple, time-coherent, IF channels in the VNA to get at correlation between DUT output ports directly. Since the noise waveforms are directly digitized after IF processing, correlation between two noise signals can be maintained after certain levels of correction. This is the principle illustrated in Figure 20-11.



In correlated noise measurements, the external hardware configuration is largely the same (see text), but coherent digitizers in the instrument are used to help directly identify the correlation levels.

Figure 20-11. Correlated Noise Measurement Example Showing Coherent Digitizers

If the correlation is directly measurable, then the differential and common-mode noise powers are

$$\overline{|b_d|^2} = \frac{1}{2}(\overline{|b_1|^2} + \overline{|b_2|^2}) - \mathrm{Re}(\overline{b_1 \cdot b_2^*})$$

$$\overline{|b_c|^2} = \frac{1}{2}(\overline{|b_1|^2} + \overline{|b_2|^2}) + \mathrm{Re}(\overline{b_1 \cdot b_2^*})$$

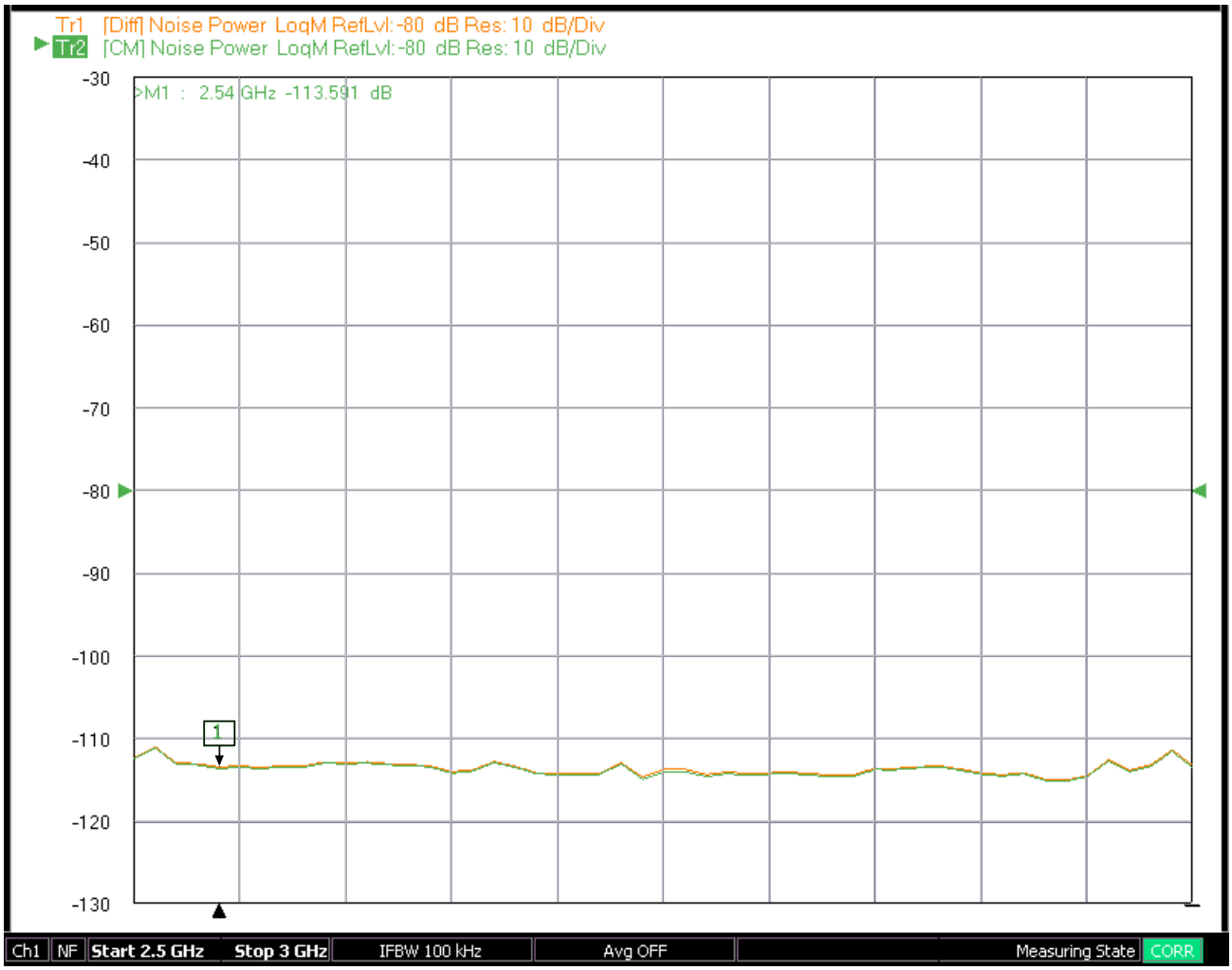
Equation 20-10.

Of course, there are complications:

- The b_1 and b_2 measurements are now complex quantities, so a phase consistent reference plane must be established. The correlation calibration was established to achieve this using ratioed measurements against a deterministic signal from an internal source. A thru line connection to each receiver path is all that is required, and this can be done at the same time as a receiver calibration. As with the receiver calibration, the system will interpolate/ (flat-line) extrapolate if the frequency range is different at measurement time vs. calibration time.
- At some point in a setup, coherence time can come into play. That is, two noise signals will only retain their coherence relative to each other over some finite length of line (even if well-matched). This is not necessarily intuitive, but is demonstrated in a more obvious way with the famous Michelson-Morely experiment (e.g., [8]) when performed with white light. In this experiment, two paths of different electrical lengths are made to interfere and create interference fringes. When the lengths get longer, these fringes get less distinct and eventually vanish when using broadband, white light. The concept is that every source has some inherent coherence time that scales as 1/ (bandwidth) that describes the scale over which this occurs. For a VNA noise measurement, the bandwidth of relevance is the IF bandwidth of the measurement system, which is of order 1 MHz or smaller, so the length scales that could be a concern are longer than 1 microsecond.

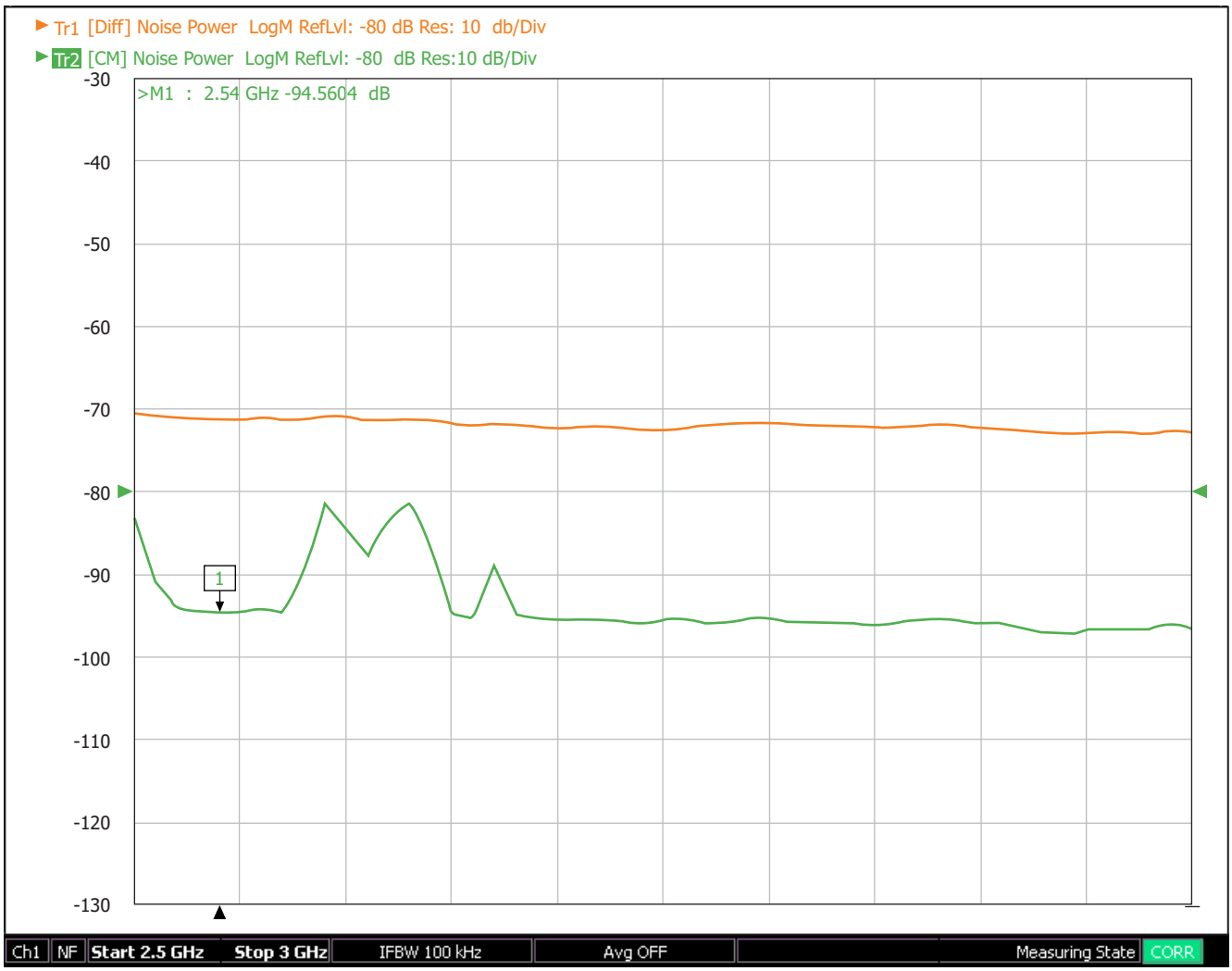
- Even aside from coherence time, the receiver network itself can de-correlate a pair of incoming correlated noise signals. If the gain and noise generation of the receiver system greatly exceeds that of the DUT, the uncorrelated noise of the receiver system (assuming distinct amplifiers are being used) could wash out the signal correlation, and this cannot be recovered. Although good practice, as discussed in Chapter 19, the receiver pre-amplifier chain should have just enough gain to pull the kTB base signal out of the instrument floor. In addition, radically different electrical lengths of the receiver chains can de-correlate the signals in a periodic fashion, but the instrument uses the frequency response of the correlator calibration functions to partially correct for this. This correction relies on the frequency sweep being sufficiently wide relative to the electrical length delta, so adding some asymmetry to the setup is helpful for narrowband devices. If a CW frequency is being measured (or the frequency sweep is extremely narrow relative to the electrical length difference), a different algorithm will automatically activate that de-embeds receiver network correlation effects, but accuracy is reduced, mainly for the non-dominant mode (e.g., common-mode noise power for a differential amplifier will have reduced accuracy).
- Because of the use of frequency relationships, the data only updates at the end of sweep when in this mode.

As an example, consider the noise power from a passive DUT. From the earlier discussion, one would expect these to be uncorrelated and, in this measurement example, the differential and common-mode noise powers do essentially overlay (see the example in Figure 20-12). When a differential amplifier is measured (whose internal topology suggested a fair amount of noise correlation was likely), a very different result is obtained, as shown in Figure 20-13. Here the differential mode noise power exceeds the common-mode noise power by at least 10 dB (and by >20 dB at some frequencies). There is a lower limit to the correlated power measurement, set by the single-ended floor, so soft clipping will result in some measurements. In this particular case, the common mode output impedance was such that even the thermal noise contribution to this mode was small at many frequencies.



Correlated noise measurement examples are shown for a passive DUT in this figure and a differential amplifier in Figure 20-13.

Figure 20-12. Correlated Noise Measurements Example for Passive DUT



Correlated noise measurements examples are shown for a passive DUT in Figure 20-12 and a differential amplifier in this figure.

Figure 20-13. Correlated Noise Measurements for Differential Amplifier

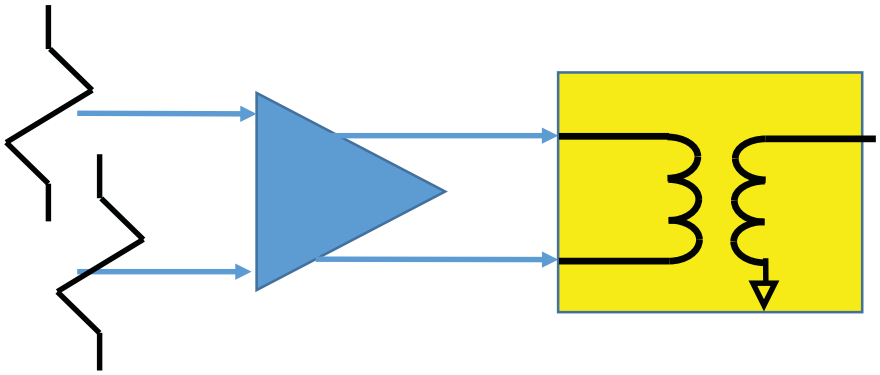
General calibration and measurement procedure:

- Set up composite receivers feeding b_1 and b2.
- Perform a power calibration on port 1 (if desired) at a sufficiently low level that the composite receivers will not compress. Perform a receiver calibration (using this power calibration, if desired) and correlation calibrations on both receiver paths (the receiver and correlation calibrations can be done simultaneously using the checkbox in the CALIBRATION dialog. These calibrations can be done while in the noise figure application or can be recalled.
- Perform a noise calibration on both receiver paths with terminations attached to the receiver inputs.
- Load DUT S-parameter or gain data.
- Connect DUT and measure.

The steps are essentially the same as in Figure 20-9, since the receiver and correlation calibrations can be done with the same connection. A checkbox in the CALIBRATION dialog allows the two calibrations to execute simultaneously.

20-7 Balun-based Methods and Handling of Imperfections

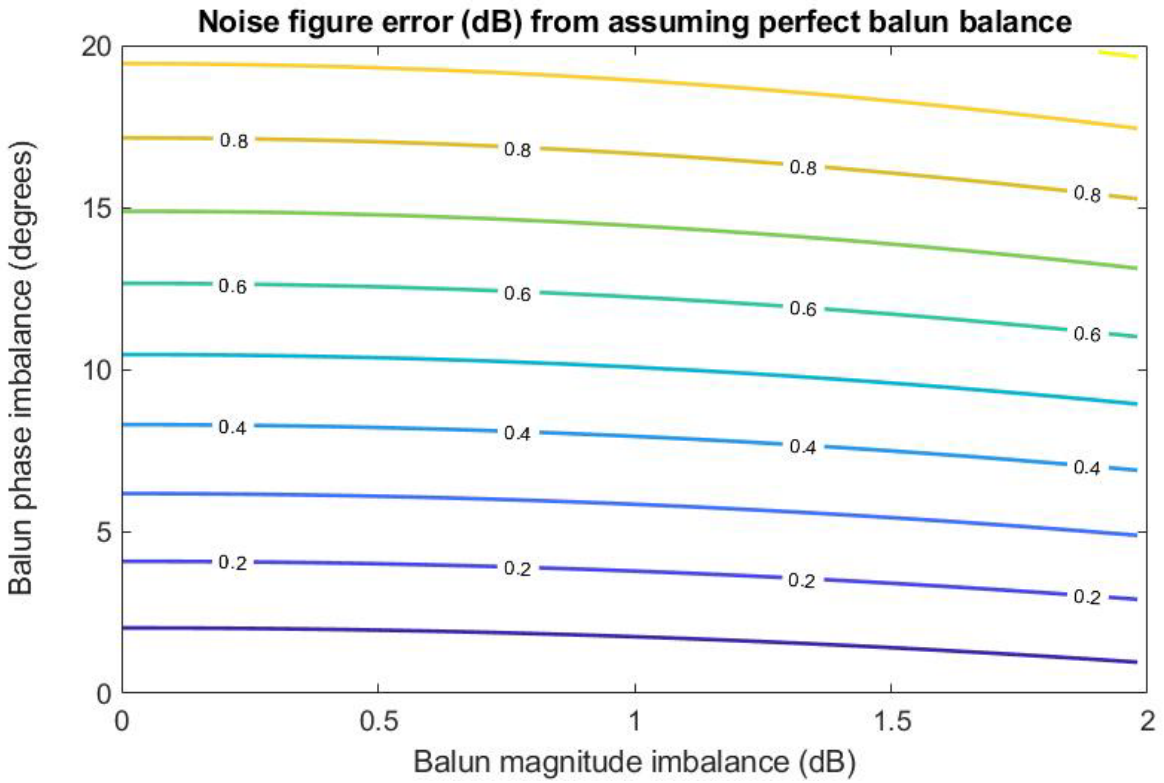
A more classical approach to dealing with correlated DUT noise outputs is to send them first to a balun (or a combiner in more general terms). If the line lengths to the balun are equal and the balun balance is ideal, then a differential signal is generated for conventional 2-port noise analysis and only the insertion loss of the balun must be de-embedded (e.g., [5]–[6]).



The use of a balun (or combiner more generically) is a useful measurement method to get differential (or common-mode) noise figure when coupled with de-embedding.

Figure 20-14. Using a Balun for Differential Noise Figure Measurements

If the balance of the combiner/balun is less than ideal, some errors can be introduced. On a simple level, for noise figure, the gain of the balun is described by Sd1, so asymmetries will be corrected for directly. Noise power, however, is also affected. Similar to the discussion in the previous section, the difference between correlated and uncorrelated noise power is the correlation term. One measure of the effect of balun irregularities may then be the change they introduce into the correlation term. A simulation was run where varying degrees of imbalance were inserted and the fractional change in the correlation term was computed (see results in Figure 20-15). Relatively large amounts (up to 1 dB) of amplitude imbalance did relatively little, but 10 degrees of phase imbalance caused about ~0.5 dB noise figure error. At least for high-frequency broadband baluns, such a level of imbalance is not unusual. Considering also the lines connecting the DUT to the balun (assuming they are not integrated on the same chip, in which case this can indicate the sensitivity to layout): 10 degrees of phase difference at 70 GHz corresponds to about 120 mm length difference with an air dielectric (and less on any substrate). Thus some means of correcting for imbalance might have a detectable effect on overall measurement accuracy.



The calculations assume minimal ohmic balun insertion loss, and neglect mismatch.

Figure 20-15. Noise Figure Error from Neglecting Balun Imbalance

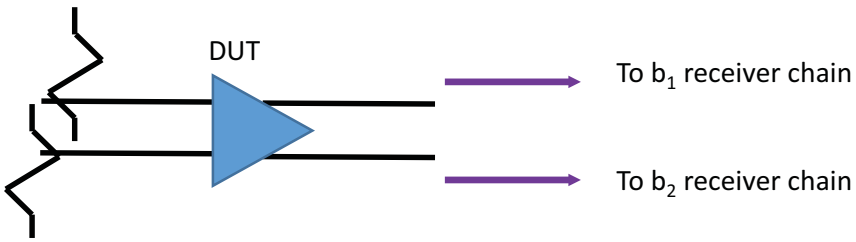
Conceptually, one can arrive at the correlation term by separately measuring the single-ended noise powers from the DUT and removing their effects after propagating them through the S-parameters of the balun (e.g., [7]). This does leave one additional layer of ambiguity on how the noise powers combine in the balun, since the real and imaginary parts of correlation can co-mingle. If one can assume that the imaginary part is negligible, only one balun-based measurement (this updates the display real time) is required. To fully treat the imaginary part, a second swapped-measurement with the balun in place can be done. This swapped measurement is stored, much like a noise calibration step, and should be updated with DUT changes. In both cases (with or without the swapped measurement), the single-ended DUT noise measurements must be stored, also like noise calibration steps, and these should also be refreshed with DUT changes. The sequence of measurements is shown in Figure 20-16.

Some notes on this measurement:

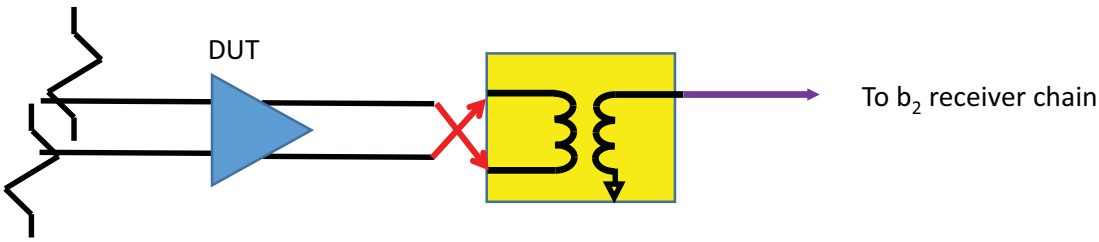
- The balun (combiner) must be pre-characterized and an .s3p file stored with port 3 as the defined common port. A port swap tool is available for files stored with some other port configuration. Interpolation and (flat line) extrapolation will be used for frequency lists that do not match the current list.
- Step 2 (and 3 if the second combiner measurement is used) should be repeated if the DUT or its bias state is changed.
- Only the b_2 receiver chain is used for the combiner measurements. While the b_1 receiver is used for the single-ended measurements, the same pre-amplifier/filter assembly can be used on b_1 and b_2 inputs since these single-ended measurements do not have to be performed simultaneously.

1) Perform receiver and noise calibrations

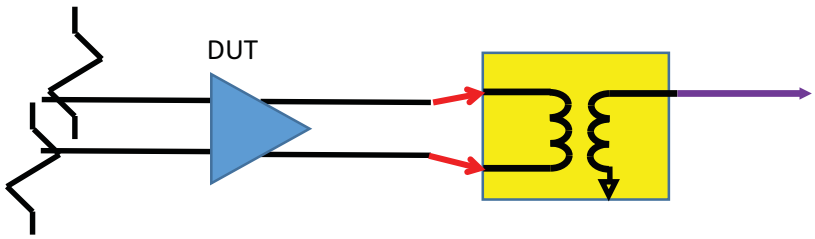
2) Collect single-ended noise data (can be done at the same time or sequentially)



3) (optional) Collect data with balun(combiner) in the swapped configuration



4) Run the real-time measurement with the balun (combiner) in the normal configuration

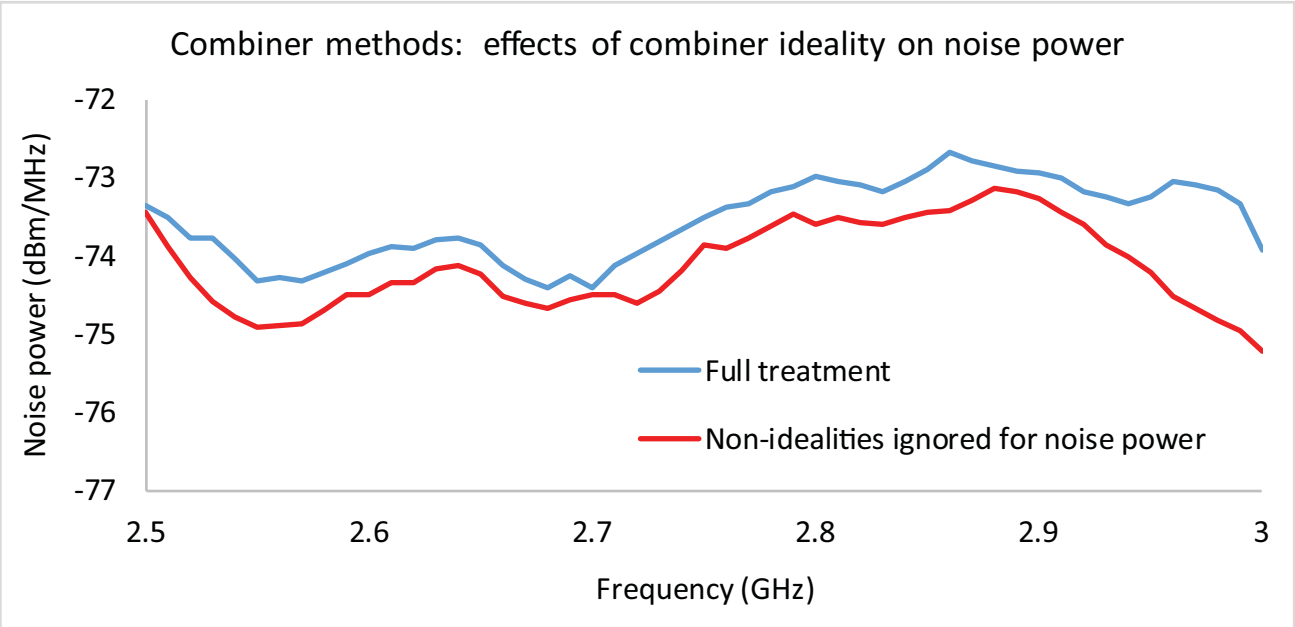


Step 3 is optional when the double-combiner measurement sub-method is selected, but can improve accuracy, particularly for DUTs with partially correlated outputs.

Figure 20-16. Measurement Sequence for the Combiner Method

- There are no particular constraints on the balun parameters, but if the insertion loss is very high, low DUT noise figures (or moderate noise figures on very low gain DUTs) may have higher uncertainties. The algorithm has to work harder the more non-ideal the balun/combiner is, so the characterization accuracy of that balun becomes more important in those cases.

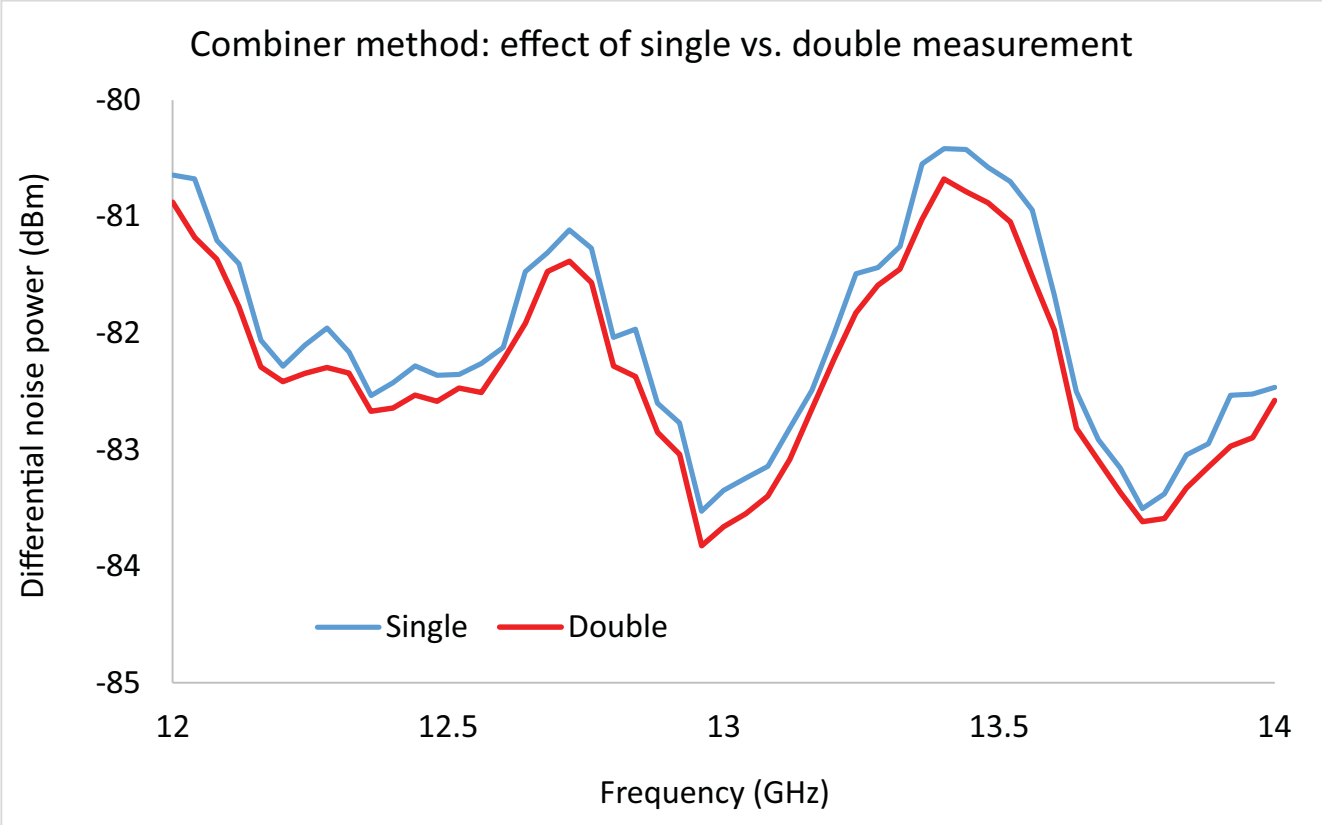
Consider an example measurement that uses the full characterization of the combiner to calculate the correlation, as just discussed. The resulting noise power is plotted in Figure 20-17, where it is contrasted with a measurement where the noise power was just measured directly and the loss (in a differential sense) of the combiner simply de-embedded. Note that gain definitions do not play a role in these plots, as it is simple noise power referenced to a common reference plane (and de-embedding would not be able to treat correlation differences) but the errors do reach 0.5 dB or more in places, which is not entirely inconsistent with Figure 20-15. The net amount of difference is a function of both the non-idealities in the combining structure and the degree of correlation in this particular DUT output. As always, noise power (in excess of calibration values) will propagate to noise figure on a dB-for-dB basis.



An example measurement result of noise power for a differential amplifier using the balun (combiner) method of this section (blue/upper trace), and for a more classical combiner approach (red/lower trace) where the noise power from the combiner is measured directly and the loss of the balun de-embedded.

Figure 20-17. Noise Power Examples using Balun and Classical Combiner Approaches

The effect of the using the single or double combiner measurement can be more subtle depending on the details of the DUT correlation. A relatively representative example measurement of a highly correlated DUT is shown in Figure 20-18 for both approaches to incorporate balun non-idealities. The average difference over the bandwidth was ~0.3 dB, which is higher than the repeatability level for this DUT measurement (~0.1 dB) but less than the absolute uncertainty levels and less than the differences seem in Figure 20-17 (where the balun was treated as ideal for correlation purposes). There may be devices with different correlation structures where the single/double difference could be larger.



An example measurement result of noise power for a differential amplifier using the balun (combiner) method of this section, and a single combiner/balun measurement (blue), or the double measurement (red), which makes fewer assumptions about the nature of the correlation function.

Figure 20-18. Noise Power Examples Using Balun and Single Combiner Methods

To re-iterate, the general procedure can be described by (see Figure 20-16):

- Set up composite receivers feeding b_1 and b_2.
- Perform a power calibration on port 1 (if desired) at a sufficiently low level that the composite receivers will not compress. Perform a receiver calibration (using this power calibration if desired) on both receiver paths (the receiver and correlation calibrations can be done simultaneously using the checkbox in the dialog. These calibrations can be done while in the noise figure application or can be recalled.
- Perform a noise calibration on both receiver paths with terminations attached to the receiver inputs.
- Load DUT S-parameter or gain data.
- Connect DUT (without the combiner/balun present) and collect b_1 and b_2 noise data (can be done at the same time).
- If the 'double' combiner measurement was selected, connect the combiner/balun with the inputs swapped (relative to the desired connection) and collect the data. If 'single' was selected, skip this bullet.
- Connect DUT and combiner/balun and measure.

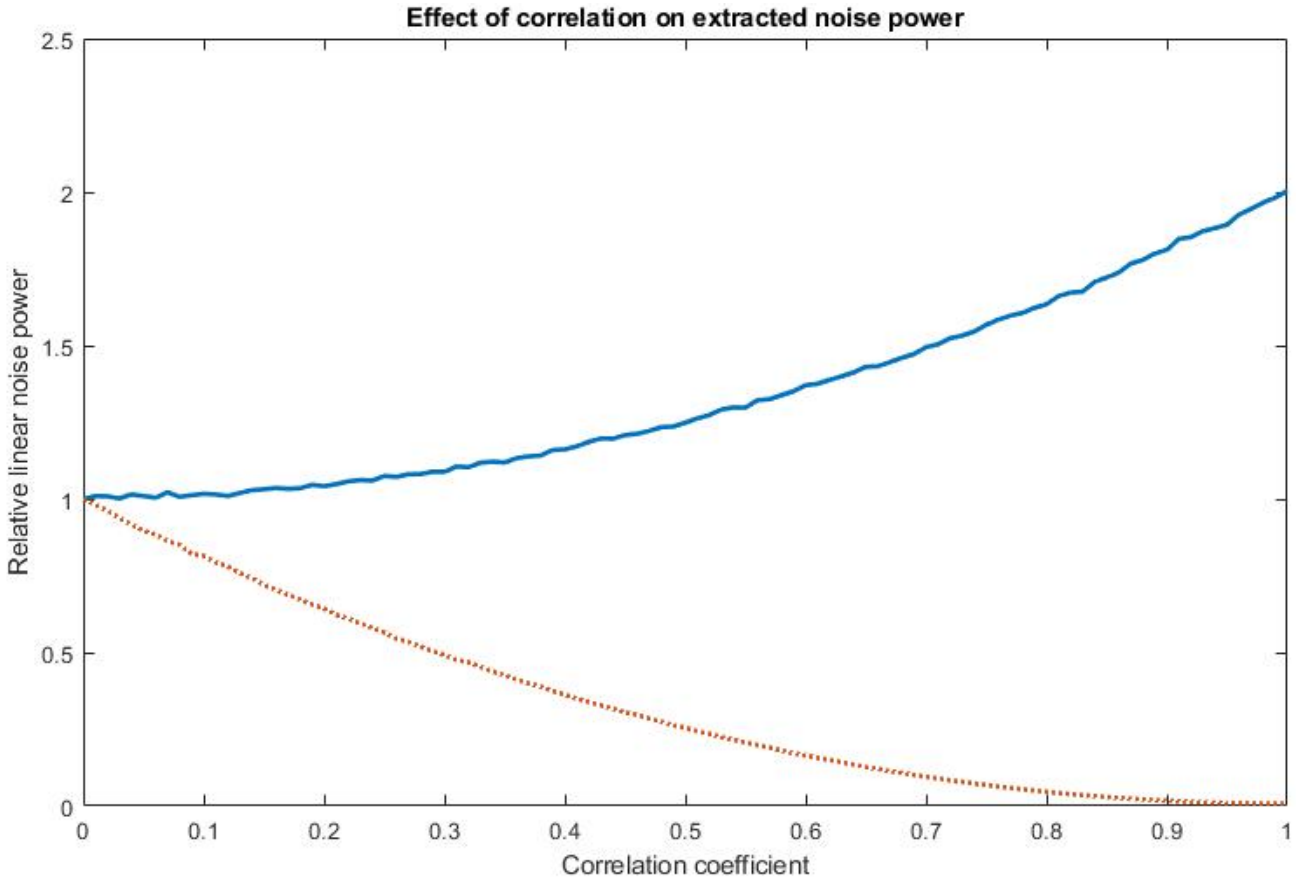
20-8 Measurement Uncertainty

Since at its core, a noise figure measurement discussed in this chapter is a combination of a noise power measurement and analysis of DUT gain, many of the uncertainty components are the same as for single-ended noise figure. As discussed in Chapter 19, some of these elements are:

- Accuracy of DUT gain. This is an S-parameter uncertainty problem, but in this chapter it may involve those derived from 4-port measurements. The contributing factors are the same and the S-parameter uncertainties are discussed in great length elsewhere.
- Power calibration and receiver calibration accuracy. Since a cold source method is used for the differential noise figure measurements as well, how the power reference plane is established matters.
- System noise floor and having adequate gain (and noise figure) in the DUT and composite receiver is essential to getting the noise power in a linear range for the VNA analog-to-digital converters.
- Sufficient RMS points per frequency to keep measurement jitter tolerable.

Specifically for differential noise figure (aside from the slightly different gain computations and different net uncertainties in those terms), there is the context of the measurement that is important: defined with uncorrelated noise inputs at temperature T_0 and all terminations are reflection-less. While not that different from the 2-port assumptions, there are more places for changes to happen in the setup. There is then the topic of correlation between DUT outputs that has been a central topic in this chapter. Section 20-7 showed the scale of errors in neglecting correlation in balun (combiner)-based methods and there is an uncertainty contribution from the S-parameter measurements of the balun. In the case of the direct correlation method, there is an S-parameter-like term embedded in the correlation calibration (how accurately can the phase reference plane be established, which will be affected by port match, etc.) and from the ability to correct for receiver-network-induced de-correlation (which will depend on relative gain levels and electrical length differences).

More generally, the effect can be somewhat self-evident by looking at Eq. 20-10. Using arbitrary scaling, suppose the average of single-ended noise powers was 1 linear unit. If the DUT outputs are completely uncorrelated, the common-mode and differential noise powers will also be 1 linear unit. At the limit of full correlation, they could reach values of 0 and 2, respectively (or the other way around for a common-mode amplifier) so the percentage or dB errors can be extreme, theoretically. In practice, this is usually not the case (and, indeed, often one does not care as much about common-mode noise figure since the gains are so low; there are also additional uncertainty complications with the lower noise power since the subtraction of nearly equal numbers is involved) but the potential errors may be useful to understand.



A Monte-Carlo simulation of extracted noise powers for different levels of correlation for the differential (blue/top)) and common-modes (red/bottom), assuming a differential amplifier. The end points of the curves are easy to understand on an ideal level.

Figure 20-19. Monte-Carlo Simulation of Extracted Noise Powers for Different Levels of Correlation

20-9 Other Comments

Differential noise figure is compatible with the broadband and mmWave systems and, as with Option 41, a receiver module greatly helps this effort by removing coupler losses (as loops are not available in the mmWave bands). Two such modules (like the 3744A-Rx) are required for most of the differential noise figure measurements, and receiver offset corrections can be applied to both receiver paths. In a 4-port setup, these receiver modules should be connected as ports 2 and 4, as these will be the only paths monitored for noise measurements. Port 1 can remain as a standard module, as it will be used for receiver and correlation calibrations. The overall calibration processes follow from what has been discussed.

20-10 References

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Chapter 21 — Mixer Setup and Measurement

21-1 Chapter Overview

In contrast to the fully general multiple source control approach (see Chapter 18, “Multiple Source Control (Option 7)”) and the algorithm-specific NxN approach (see Chapter 22, “NxN Application”), the mixer wizard and (guided) mixer setup approach help to setup common mixer measurements with a simplified, yet still accurate, calibration methodology. The prime objective of the guided mixer setup is to help configure the frequency plan of the measurement using easier to understand diagrams. The mixer wizard goes a step further and helps configure measurement channels to handle any of a suite of possible mixer measurements and to list the required calibration steps. Both of these tools are coupled with a mixer calibration menu system that enables both scalar and vector corrected measurements.

21-2 Mixer Theory

The literature on mixer theory is extensive and many topics will not be covered here. The interested reader is referred to, for example, S. A. Maas, *Microwave Mixers*, Artech House, 1993 and references therein for a complete treatment.

The central purpose is the translation from one frequency (at the input) to another in an additive (or subtractive) fashion (at the output). As an aside, this chapter and the instrument will generally use the terms 'input' and 'output' so the names do not have to keep changing depending on the type of conversion. The software will allow labeling of ports as IF and RF (and allow them to be changed) as will be seen later.

In the simplest sense, the mixer can be viewed as a perfect multiplier so if the input is some sinusoid with frequency ω_{in} and the Local Oscillator (LO) port has some sinusoid with frequency ω_{LO}, then the output signal is given by Eq. 21-1 below, (ignoring any constant phase terms, it is mainly frequencies we are interested in at the moment).

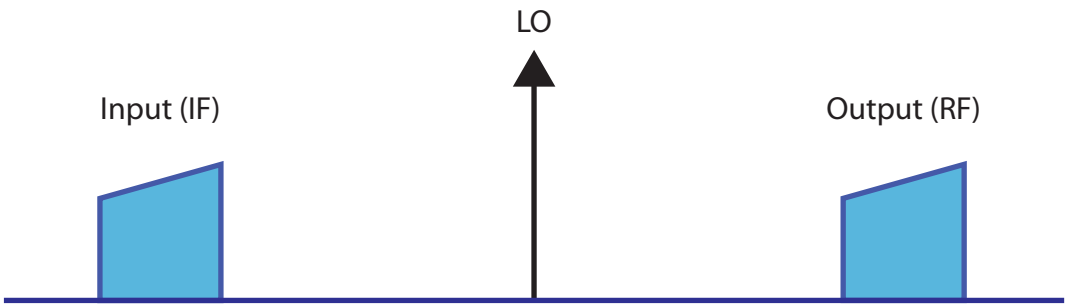
$$Output = A \cdot \cos(\omega_{in} t) \cdot B \cdot \cos(\omega_{LO} t) = \frac{AB}{2}[\sin((\omega_{in} + \omega_{LO})t) + \sin((\omega_{in} - \omega_{LO})t)]$$

Equation 21-1.

Also, the conversion distortions (magnitude and phase) of the mixer itself are not in the above equation. The important point is that sum and difference frequencies are created. Other higher order terms are also created in practice that will be discussed later.

Up Conversion

One case is when the sum term is of interest, this is called *up conversion* and typically the input is labeled as the IF in this case. This is the least complicated case and there is little ambiguity with the signals unless ω_{LO} is very small relative to ω_{in} (not common). The spectra are illustrated in Figure 21-1.



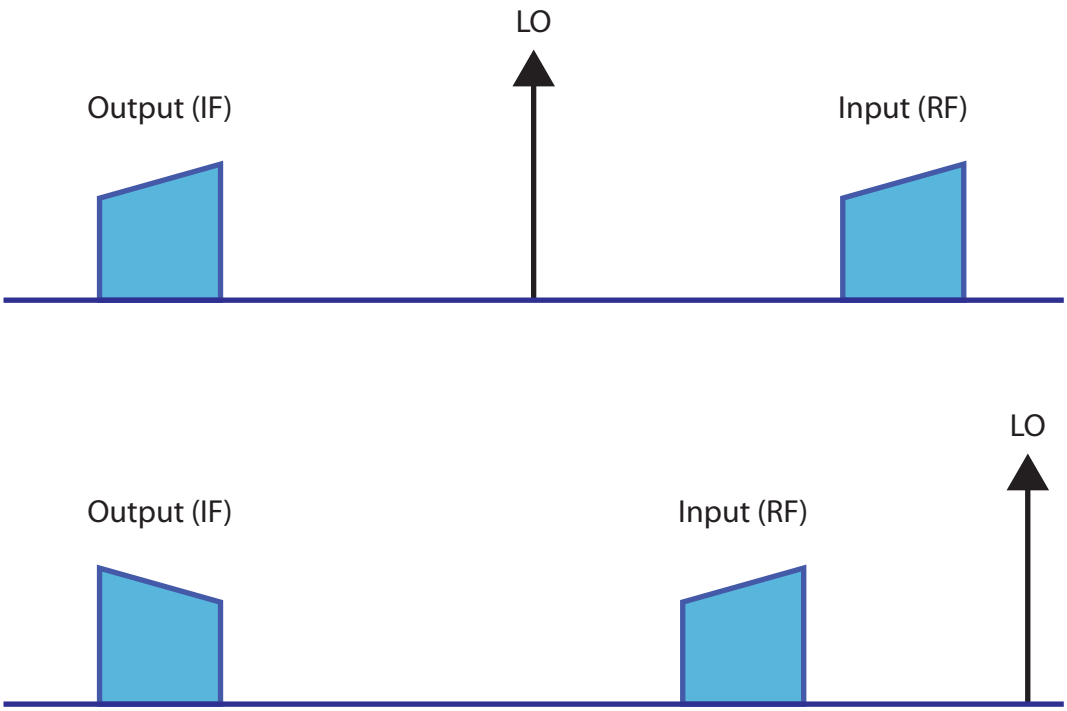
The spectra of signals in an up conversion case are shown here. The LO is normally greater than the input but this is not a requirement.

Figure 21-1. Up Conversion Signal Spectra

Down Conversion

The difference frequency case (down conversion) can be more complicated in that now the relationship of the input frequency range to the LO makes a difference in the behavior (due to a sign flip in Eq. 21-1). If the input frequency range is greater than the LO (Figure 21-2), then there is no inversion. If the LO is greater than the input, there is an inversion. One could argue that the signs in Eq. 21-1 could be re-arranged arbitrarily, but the key point is if there is a minus sign before the input frequency. If there is, then the output sweep will be in an inverse direction from the input sweep and there is a phase conjugation.

Various terms are used to separate these two cases including 'high-side LO vs. low-side LO' and 'USB vs. LSB conversion' (for upper sideband vs. lower sideband). This document and the instrument software will generally use the latter terminology.



The spectra of signals in down conversion cases are shown here. The LO position relative to the input affects the behavior. The spectra picture on top is of a low-side LO or USB case while the spectra picture on the bottom is of a high-side LO or LSB case.

Figure 21-2. Down Conversion Signal Spectra

One may ask about the negative frequency implications in Eq. 21-1 of the LSB case but the sign simply flows out of the sinusoid and inverts the phase. As far as the instrument is concerned, it takes the absolute value of the computed frequency before programming the hardware, which is just what the equation is telling it to do.

Up to this point, we have been treating the conversion as ideal. In practice, the mixer (or converter) will have gain/loss and some phase shift (both typically frequency dependent), some mismatch, and potentially considerable non-linearity. The real embodiment of Eq. 21-1 is more like that shown in Eq. 21-2 below:

$$X = A \cdot \cos(\omega_{in} t) + B \cdot \cos(\omega_{LO} t)$$

$$Output = a_0 + a_1 X + a_2 X^2 + \ldots$$

Equation 21-2.

There are also additional higher order terms. This means that any integer combination of input and LO frequencies can appear at the output (and many do). These other frequency components are often termed spurs with a nomenclature like an 'M by N spur' if the output frequency is of the form $M\ \omega_{in} \pm N\ \omega_{in}$. The measurement of these spurs is also possible with the MS464xB and will be discussed towards the end of this chapter.

As might be suggested by Eq. 21-2, one could conceivably use a harmonic of the LO to do the heavy lifting of conversion. That is, one could intentionally use the 1 by N 'spur' for the conversion. This is indeed done, particularly for higher frequency and millimeter wave applications where creating a sufficiently low noise fundamental LO is either too difficult or too expensive. Harmonic mixers and samplers (or subharmonic mixers which is confusing but equivalent terminology usually) are handled by the multiple source control system described in another chapter.

Another terminology problem is that of mixers vs. converters vs. frequency conversion devices. There is some lack of consistency in the industry but often a converter is a mixer with some additional functionality (gain blocks, switching, etc.). A frequency conversion device can be a mixer but is also sometimes use to denote a multiplier. This chapter is concerned with those devices that translate frequency in an additive or subtractive fashion but with additional gain or linear networks allowed. We will thus use the terms mixer and converter somewhat interchangeably. Multipliers are also covered by the very general multiple source control section in another chapter.

21-3 Mixer Measurement Definitions

While the terms used are common, the below may serve as a way of clarifying the nomenclature to be used in this chapter.

Frequency/Power Plan

This is a description of the frequencies involved in the mixer measurement including the input sweep range (if swept), the LO sweep range (if any), and how the output frequency is related to the input frequency. It also includes the power levels required. In the case of a power sweep, the fixed frequencies are part of the plan along with which variable is sweeping in power (input or LO) and over what range.

Input Match, Output Match, and LO Match

These are the usual return loss terms. The terms 'input' and 'output' are used in the dialog choices in lieu of IF and RF to avoid some confusion. The default has an input of RF and an output of IF but these can be reversed. The arrow directions in the diagram always make clear what the input port is. A standard S-parameter calibration (1-port or more, but with different frequency ranges on the different ports) is used.

Isolation

This is transmission through the mixer (input->output, output->input, LO->output or LO->input) but without frequency conversion. The test frequency range is based on the stimulus port of the selection and the defined frequency plan. This is sometimes also called 'bleed through'. A normalization or a standard s-parameter calibration is normally used.

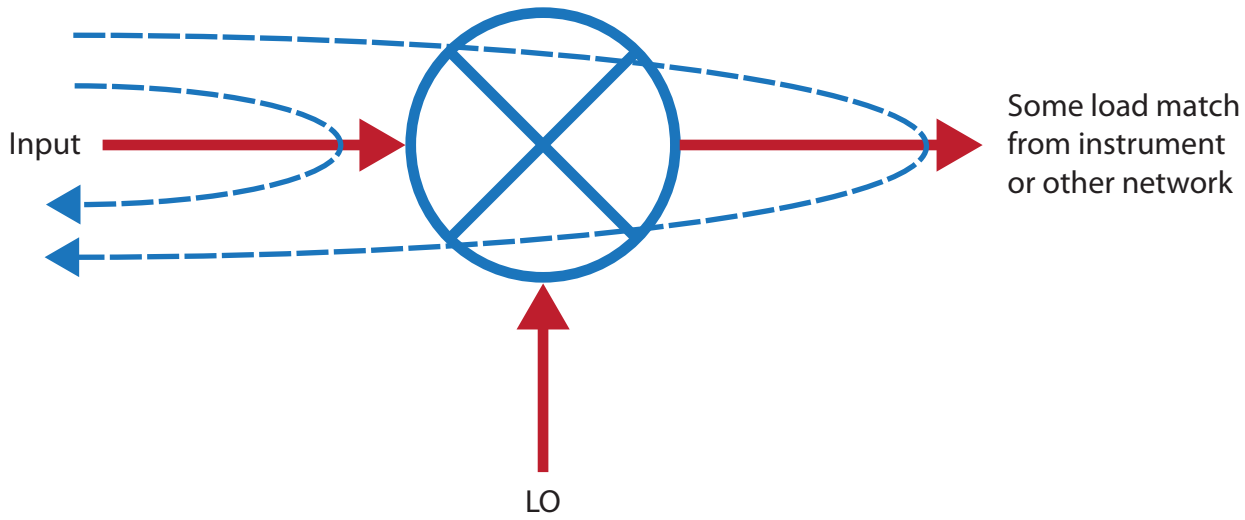
Conversion Gain/Loss

This is a measure of energy transferred from the input frequency to the output frequency. There are a number of different calibrations available ranging from a simple scalar (but narrowband) normalized power measurement to enhanced match correction that vectorially takes into account the DUT's match (input and output) interacting with the VNA. The NxN technique discussed in the next chapter is another method for getting at conversion gain/loss vectorially.

21-4 Mixer Measurements

As was suggested in the previous section, there are many different ways of making mixer measurements and many different calibration approaches. The purpose of this section is to explore some of the options and how they might perform under various circumstances.

Some of the mixer measurements described in the previous section (match and isolation) are essentially simple S-parameter measurements that have been discussed previously in this guide. The match measurements are generally 1-port in the case of mixers (from a correction point of view) since the other ports of the mixer are operating at different frequencies. Depending on the isolation of the mixer, there can lead be some load match dependencies and some confusion in the meaning of the practical match of the mixer (see Figure 21-3).



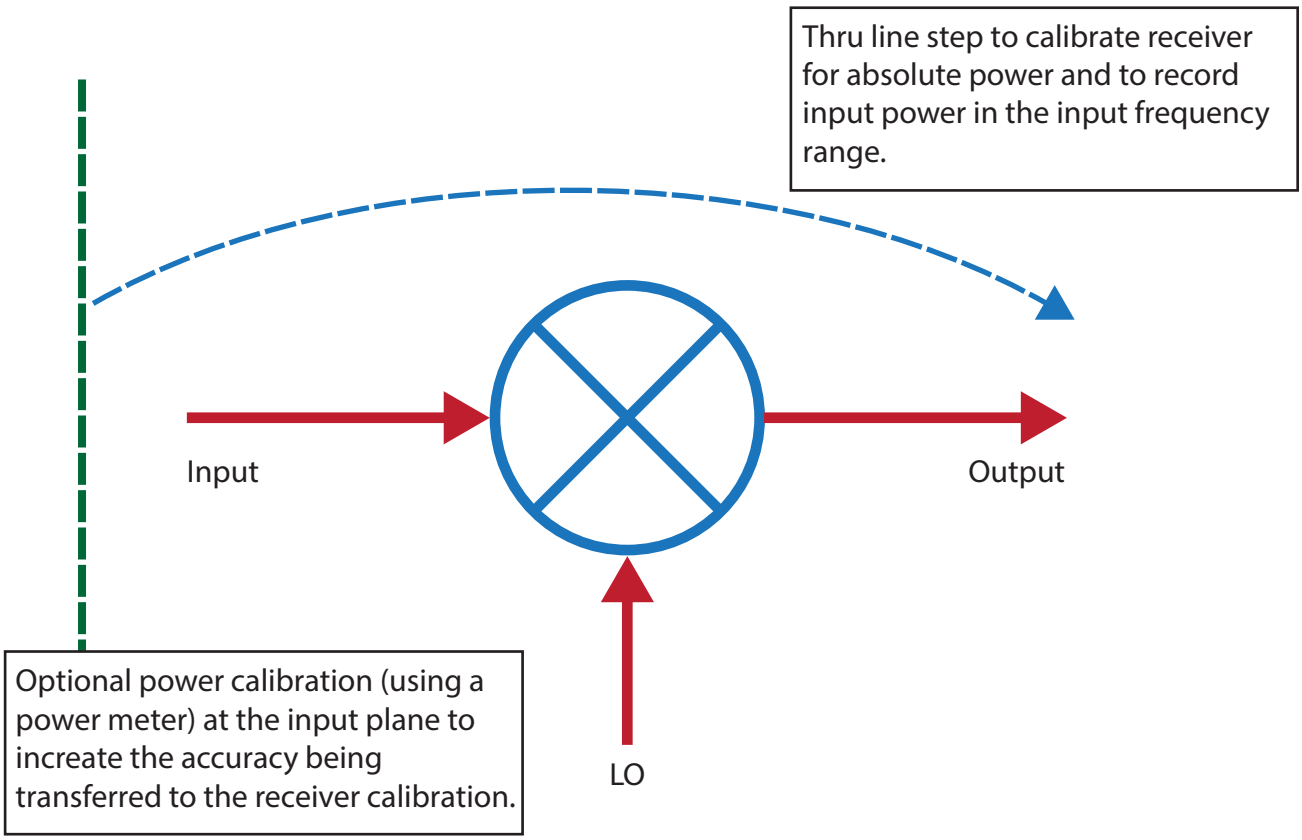
An illustration of the dynamics of the actual match of a mixer is shown here. If the isolation of the mixer is poor and the load match at the input frequencies is poor, multiple reflections can make the actual match of the mixer (in system) worse than one might expect. This can also have implication in conversion loss measurements as will be discussed.

Figure 21-3. Actual Match Dynamics

The newer measurement is really that of conversion gain/loss although it has been touched on in the Multiple Source Control chapter. The concept as discussed above is power transfer from input to output but since the frequencies are different, a classical ratioed S-parameter is not possible (a1 is at the input frequency and b2 is at the output frequency). There are several ways to handle this.

Normalization

This is a scalaresque approach where the receiver is calibrated for absolute power and the input power is calibrated absolutely. This allows one to measure the power received at the output and calculate conversion loss/gain based on knowledge of the input power. This is using the VNAs narrowband receiver so one is not also measuring spurs and leakage that one may be doing with power meter techniques and is hence more accurate than those approaches. Match is, however, neglected and can reduce accuracy. The ports can be heavily padded to reduce any match-related error. Also, group delay of the DUT is unavailable with this method but it is the simplest approach and may be most practical in millimeter wave frequencies or in complex media. The measurement is illustrated in Figure 21-4.

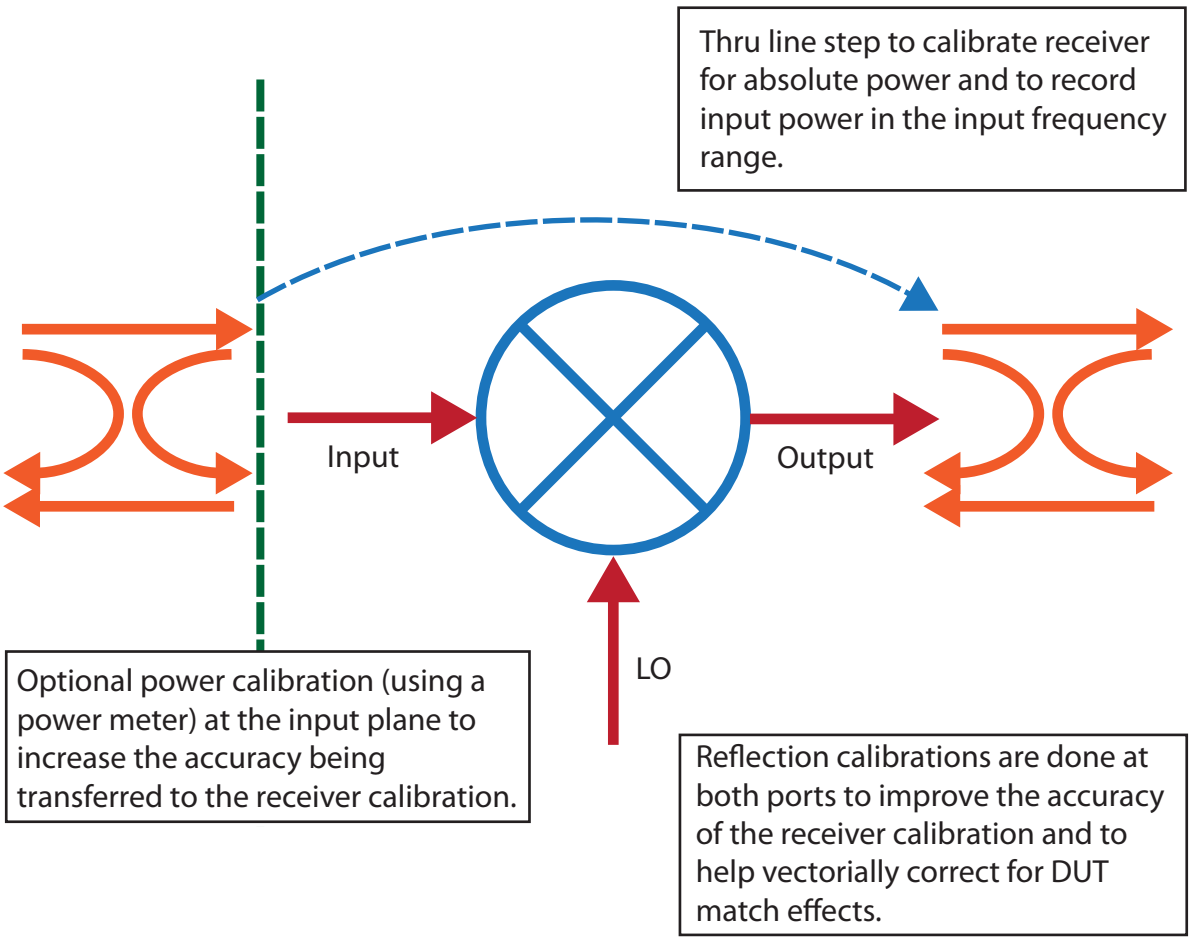


The normalization method of conversion loss/gain measurements is shown here. A thru line is used to establish the input power to the DUT (at the input frequencies) and to calibrate the receiver for absolute power (at the output frequencies). Match is neglected and padding of the ports (prior to calibration) is sometimes done to reduce mismatch effects of a poorly matched DUT.

Figure 21-4. Normalization Method of Conversion Loss/Gain

Enhanced Match

This calibration approach combines power normalization with vector match correction to make a vector method that is still relatively simple to execute and provides better accuracy. The concept is illustrated in Figure 21-5 and relies on multiple one port calibrations in multiple frequency ranges per port (input and output ranges) to correct for DUT match interaction with the ports and for port match interaction during calibration steps. The accuracy is better than that of straight normalization at the expense of some additional calibration steps. Group delay information is also not available with this method.



The enhanced match calibration is shown here. The ports are corrected for mismatch at both input and output frequencies so that match effects during both the DUT measurement and during the calibration steps can be corrected vectorially.

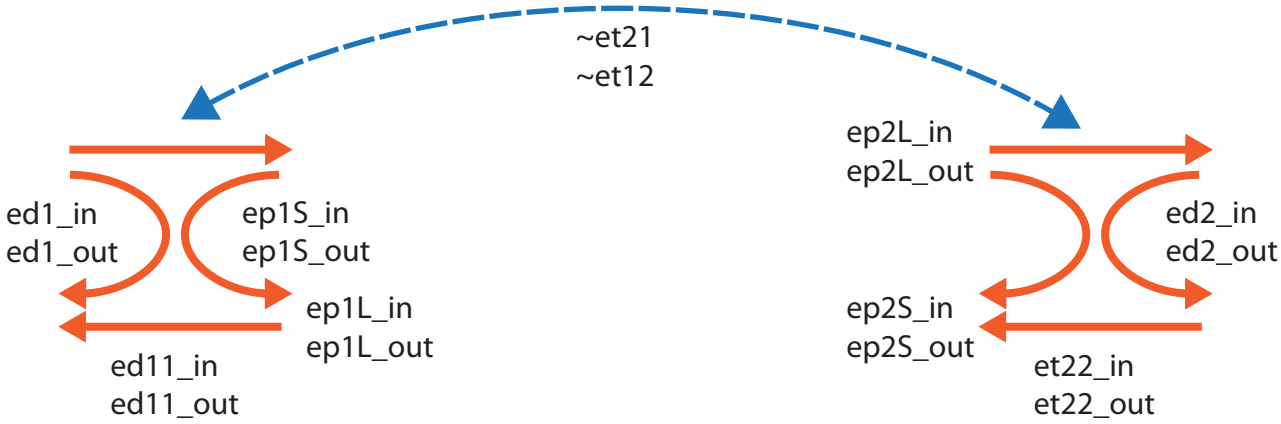
Figure 21-5. Enhanced Match Calibration

The principles of the enhanced match calibration are essentially the same as for the regular S-parameter calibrations covered earlier in this document. There are the obvious disparate frequency ranges involved and some assumptions are implicit that are worth discussing. The error boxes for the present problem are illustrated in Figure 21-6. The reflectometer terms have the frequency range explicit in their names. The transmission terms have the frequency conversion implicit in their definition (with a ~ to denote a variance from the usual S-parameter meaning in that now it applies to unratioed conversion).

The receiver calibration is corrected for the interaction of the input port source match and the receiving port load match (ports 1 and 2 in this description although that is not a requirement). This reflection term is of the form (1-ep1S*ep2L) and is done for both input and output frequencies. For the DUT correction, similar terms apply for the interaction of DUT S22 and ep2L_out and the interaction of DUT S_{11} and ep1S_in. As the isolation terms are included, this correction gets more complex but the form of the terms is similar.

Since only fundamental mixing terms are included, this correction can never be complete so the accuracy will be best when the device is operating more linearly (obvious on many levels), when isolation is high, and when the DUT is more unilateral (i.e., does not convert in both directions well).

A variant on this correction, termed "True Source Match," uses a modified term (instead of ep1S_in) to help describe the input match interaction. This is helpful, particularly in mmWave mixer measurements, where the actual source match starts to diverge more from the ep1S term due to coupler placement. When this mode is enabled, an additional backward-looking measurement is performed during calibration to better characterize the source match that the DUT actually sees.



The error boxes for the enhanced match correction are shown here. Port 1 is used as the input and Port 2 as the output in this figure but that is not a requirement.

Figure 21-6. Error Boxes for Enhanced Match Correction

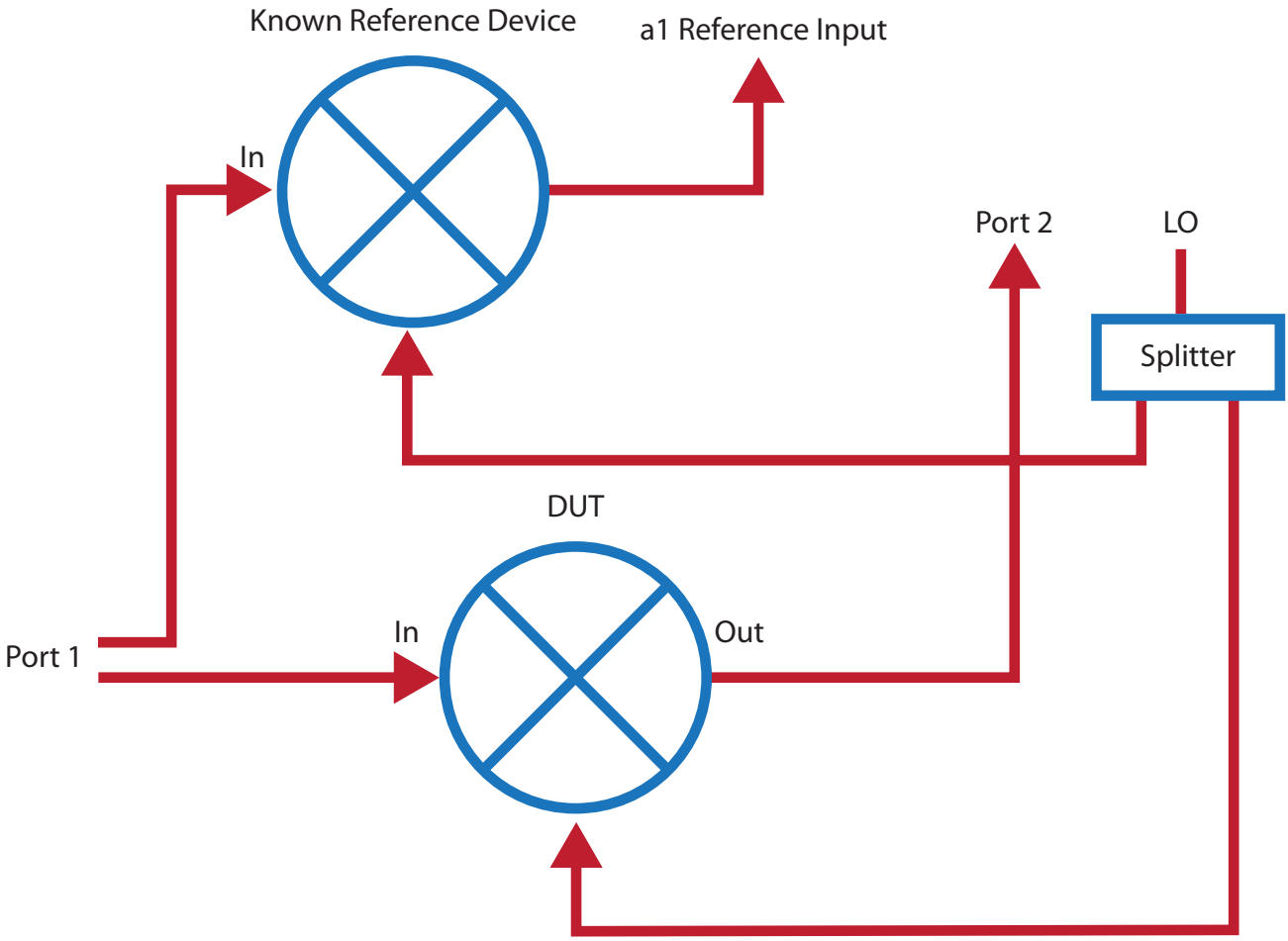
Broadband Enhanced Match Cal

Commonly SOLT, SSLT or SSST calibration cores are used for enhanced match calibrations (or variants using the hybrid calibration processes) but, if 1 mm or 0.8 mm coaxial components are employed, it may be desirable to combine SOLT and SSST calibrations as is done in conventional, broadband S-parameter measurements. In these 'merged' calibrations, the open-short-load or triple-offset-short algorithm is employed depending on if the current frequency is above or below a breakpoint frequency (default is 67 GHz for 1 mm and 80 GHz for 0.8 mm calibration kits). This same merged approach can be used with enhanced match calibrations and the frequency sorting will be performed on both input and output ports.

Reference Mixer

One way to get back to a ratioed measurement is to also place a converter in the reference channel (a1 often). This requires one of the loop options (051/061/062) but allows for a S21 (or similar) measurement to be made on a mixer and hence allow group delay extraction. The simplest approach is to use a very broadband (relative to the DUT) mixer that is well-padded to act as the reference mixer so the correction for its effects is a simple normalization. The concept is shown in Figure 21-7.

The reference mixer sequence is not explicitly supported in the wizard setup systems but is compatible with it. The reference mixer must obviously support at least the frequency plan of the DUT but ideally is very broadband. The LO is shown as split in Figure 21-7 but another synthesizer (as long as it is locked to the same time base as the instrument) can also be used. Indeed, a harmonic mixer with a different multiplier can be used as the reference but multiple source control will generally be needed for this effort (see Chapter 18, "Multiple Source Control (Option 7)").



The reference mixer approach is shown here. The port definitions can be changed as before.

Figure 21-7. Reference Mixer Approach

Embedding/De-embedding

A more complex approach is to characterize the reference mixer and then to de-embed its effects using the embedding/de-embedding engine discussed in a previous chapter. The parameters of this mixer can be found using the NxN method (see the next chapter) or using network extraction (Type B discussed elsewhere in this guide).

NxN

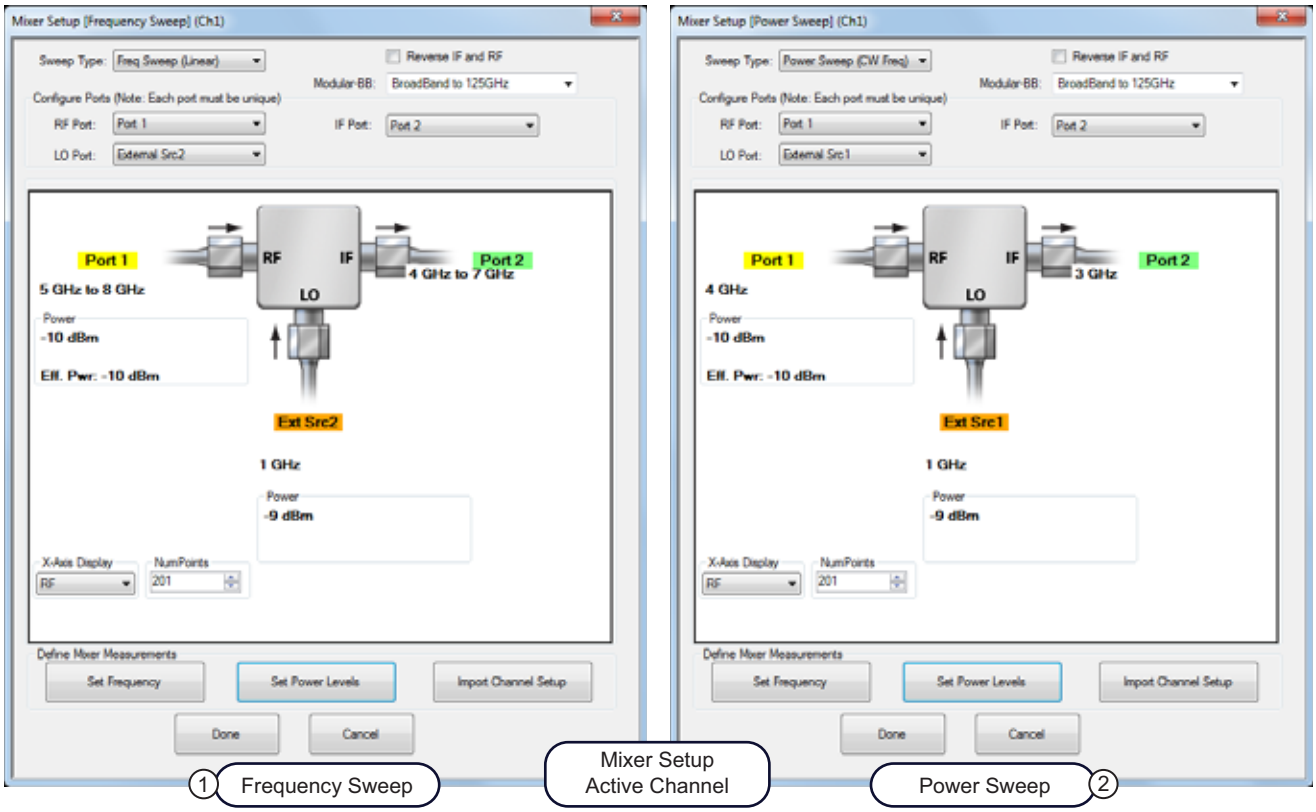
A fourth method, also recreating a ratioed measurement, is termed NxN and is discussed in more detail in the next chapter. The idea here is to use a characterized mixer to reconvert the DUT output frequency back to the input. This does not require a loop option, but does require some care in handling the image and match at the interface between the DUT and the characterized mixer (usually with a simple filter and pads). Accuracy is approximately the same as with the previous two methods.

21-5 Active Channel Mixer Guided Setup Controls

The guided setup and wizard sections of the application menu are split into two parts: one that only sets up the active channel and acts more as a guide (will be discussed now) and a multi-channel wizard that helps one configure multiple measurements in a more organized fashion (to be discussed later in this chapter). Details on the controls themselves are covered in the operations and user interface manuals. The purpose here is to discuss how the controls relate to the mixer measurement concepts discussed previously.

Mixer Setup Dialogs

The basic setup screens for this active channel method are shown in Figure 21-8. The IF and RF ports can be assigned to different VNA physical ports as has been discussed and the IF/RF labeling can be swapped (see the check box near the top). Both frequency sweeps (of input and LO possibly) and power sweeps are supported and are chosen from a pull-down menu. Also part of the input/output duality is how the x-axis of the plots are labeled and this is handled by the pull-down in the lower left part of the dialog. In the case of a fixed output frequency or a fixed input frequency, labeling the x-axis with the fixed variable is usually not helpful.

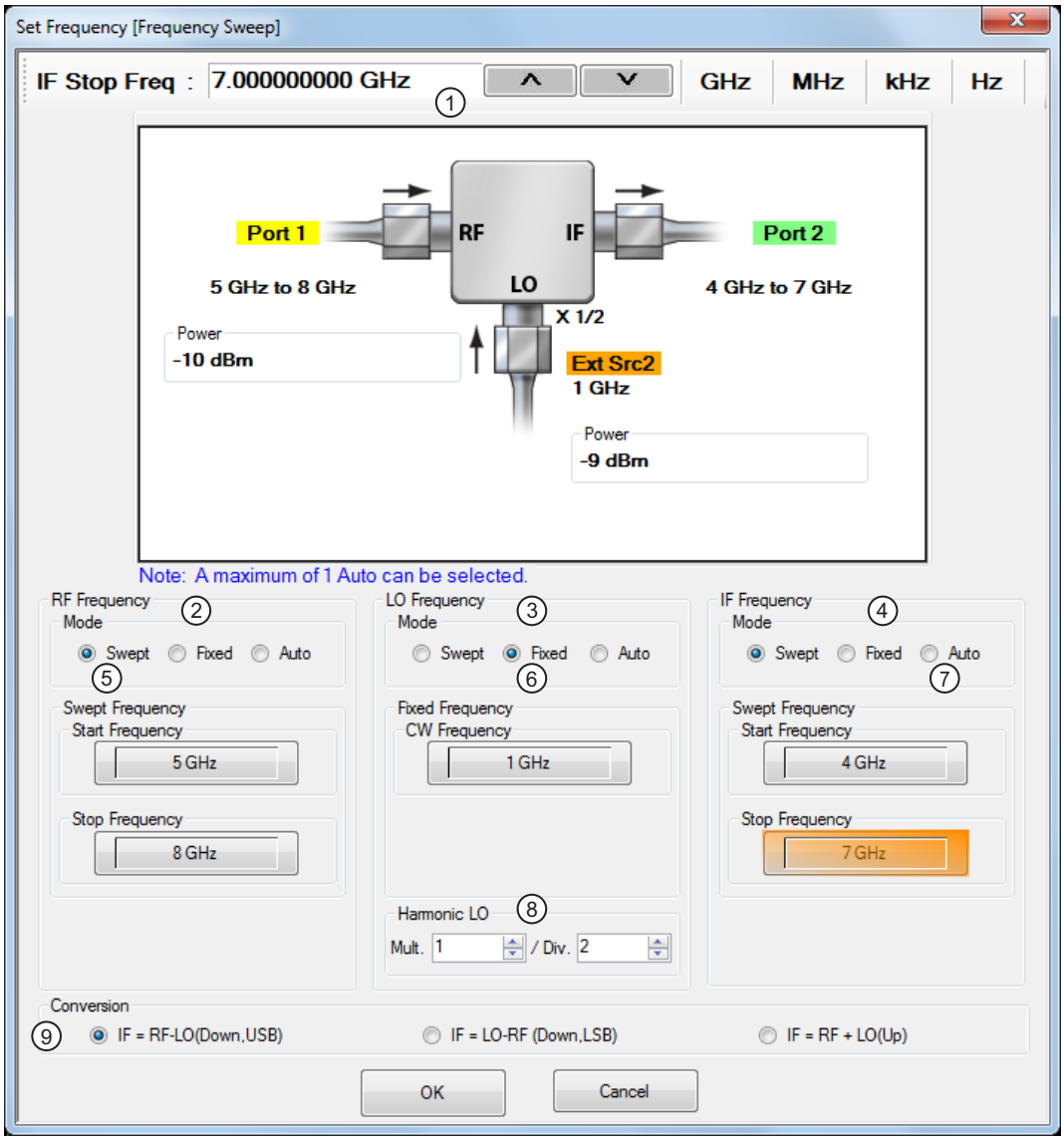


1. On left, MIXER SETUP (FREQUENCY SWEEP).	2. On right, MIXER SETUP (POWER SWEEP).

Figure 21-8. Mixer Setup Dialogs

The frequency ranges (including the possibility of reverse sweeps) are shown in the dialog and are set up using a sub-dialog as shown in Figure 21-9. The conversion direction and sideband choices discussed previously are handled here and there are many permutations possible. The conversion direction is selected from the radio buttons at the bottom and note the two downconversion selections discussed previously (USB and LSB). Input, output, or LO may be fixed (generally only one of them) and an Auto selection may be used to let the system calculate the frequencies for one of the ports. It is, of course, possible for an invalid frequency combination to be entered (out of range of the hardware) but this will be scanned for when the OK button is selected.

The dominant frequency variable (normally RF but will be IF if the Reverse IF and RF checkbox is selected) must have a start frequency less than the stop frequency when in a swept condition.

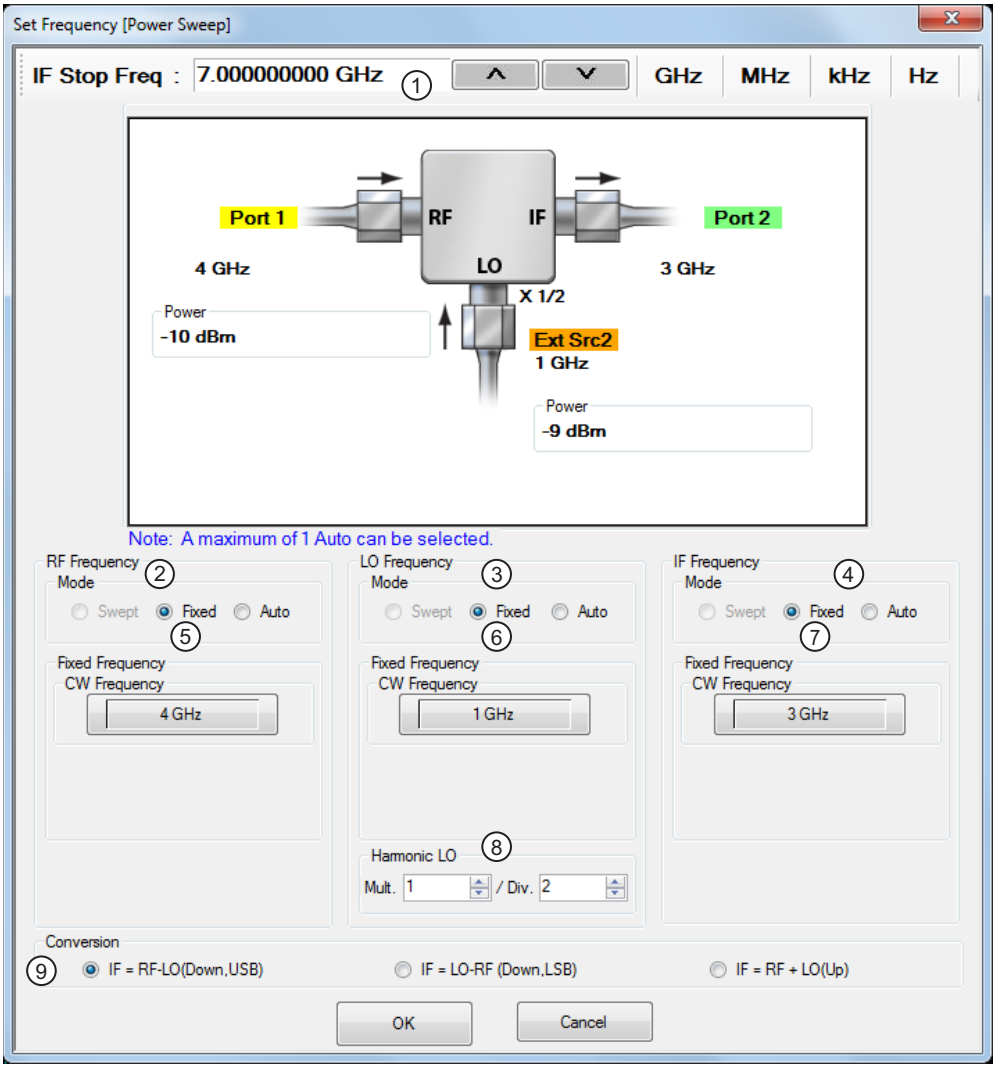


1. Frequency input field toolbar. The label changes depending whether RF, IF, or LO are selected.
2. RF Frequency mode and value area.
3. LO Frequency mode and value area.
4. IF Frequency mode and value area.
5. Mode set as Swept. Any combination of the RF, IF, and LO ports can be set as Swept. When selected, Start and Stop Frequency fields appear.
6. Mode set as Fixed. Any combination of the RF, IF, and LO ports can be set as Fixed. When selected, a CW Frequency field appears.
7. Only one port can be set as Auto. If selected, the instrument determines the frequency settings based on the other two values and the Conversion selection.
8. The Harmonic LO entry allows one to describe any frequency multiplication and/or division that may be present between the synthesizer providing the LO signal and the DUT's fundamental frequency converter. This entry can also be used when the DUT is a harmonic mixer or sampler that has implicit frequency multiplication in the LO path.

   Note: A label showing the Harmonic LO ratio is displayed at the mixer schematic LO port when multiplier/divisor (M/N) ratio is *not* equal to 1.
9. Conversion setting. Changes based on whether RF is input or output.

Figure 21-9. Frequency Setup Dialog for Frequency-Sweep Case

In the case of a power sweep, all of the frequencies must be CW so the frequency selection dialog simplifies as shown in Figure 21-10. The same sideband and conversion choices apply; there is just no frequency sweeping involved in this measurement setup.

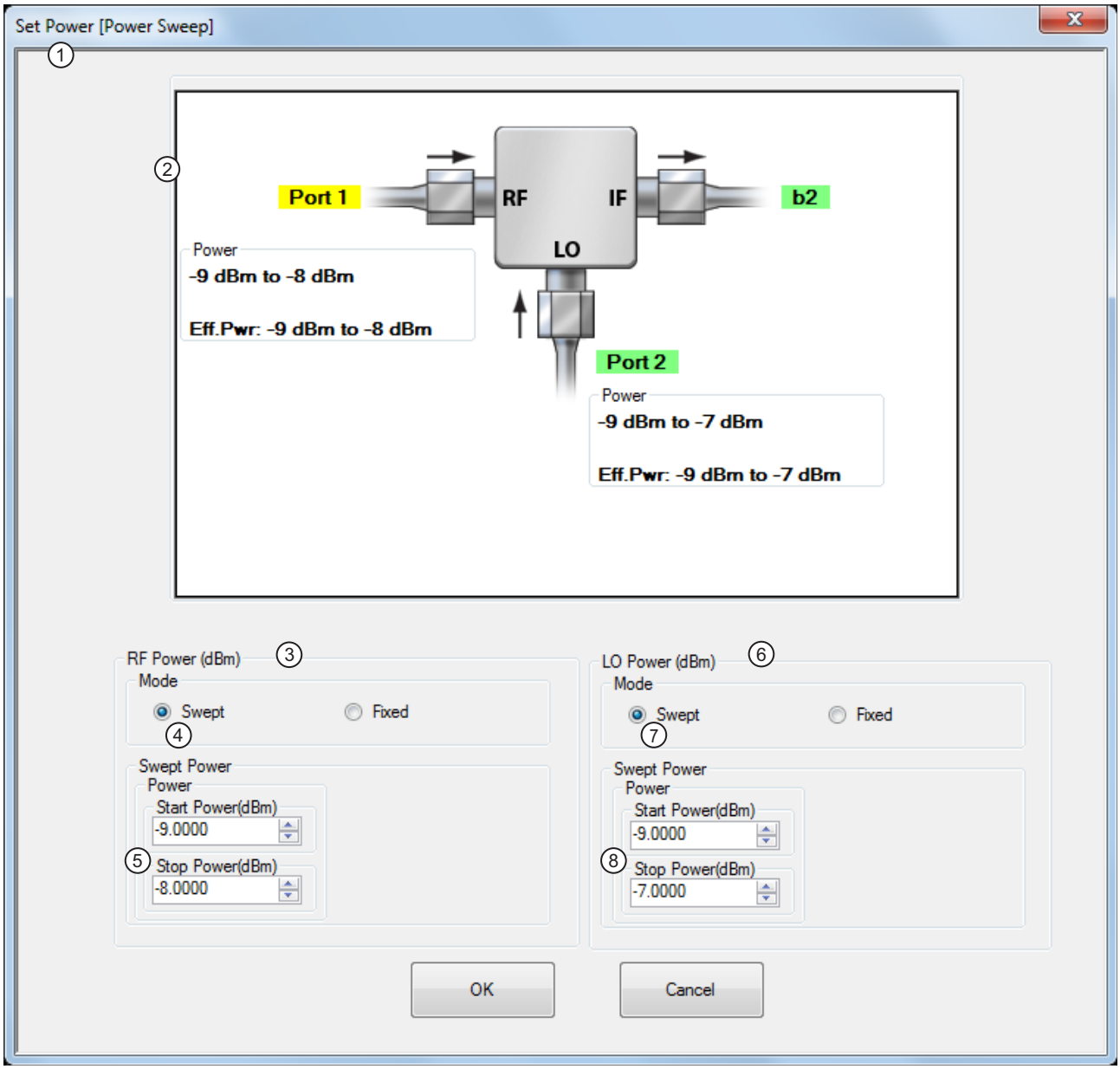


1. Frequency input field toolbar. The label changes depending whether RF, IF, or LO are selected.
2. RF Frequency mode and value area.
3. LO Frequency mode and value area.
4. IF Frequency mode and value area.
5. RF Mode setting can only be Fixed or Auto. Only one mixer port can be Auto. If fixed, RF CW frequency is input in field below.
6. LO Mode setting can only be Fixed or Auto. Only one mixer port can be Auto. LO CW frequency is input in field below.
7. IF Mode setting can only be Fixed or Auto. Only one mixer port can be Auto. IF CW frequency is input in field below.
8. The Harmonic LO entry allows one to describe any frequency multiplication and/or division that may be present between the synthesizer providing the LO signal and the DUT's fundamental frequency converter. This entry can also be used when the DUT is a harmonic mixer or sampler that has implicit frequency multiplication in the LO path.

   Note: A label showing the Harmonic LO ratio is displayed at the mixer schematic LO port when multiplier/divisor (M/N) ratio is *not* equal to 1.
9. Conversion setting. Changes based on whether RF is input or output. If RF is input, equations show conversion calculations for IF. If IF is input, equations show conversion calculations for RF.

Figure 21-10. Frequency Setup Dialog for Power Sweep Case

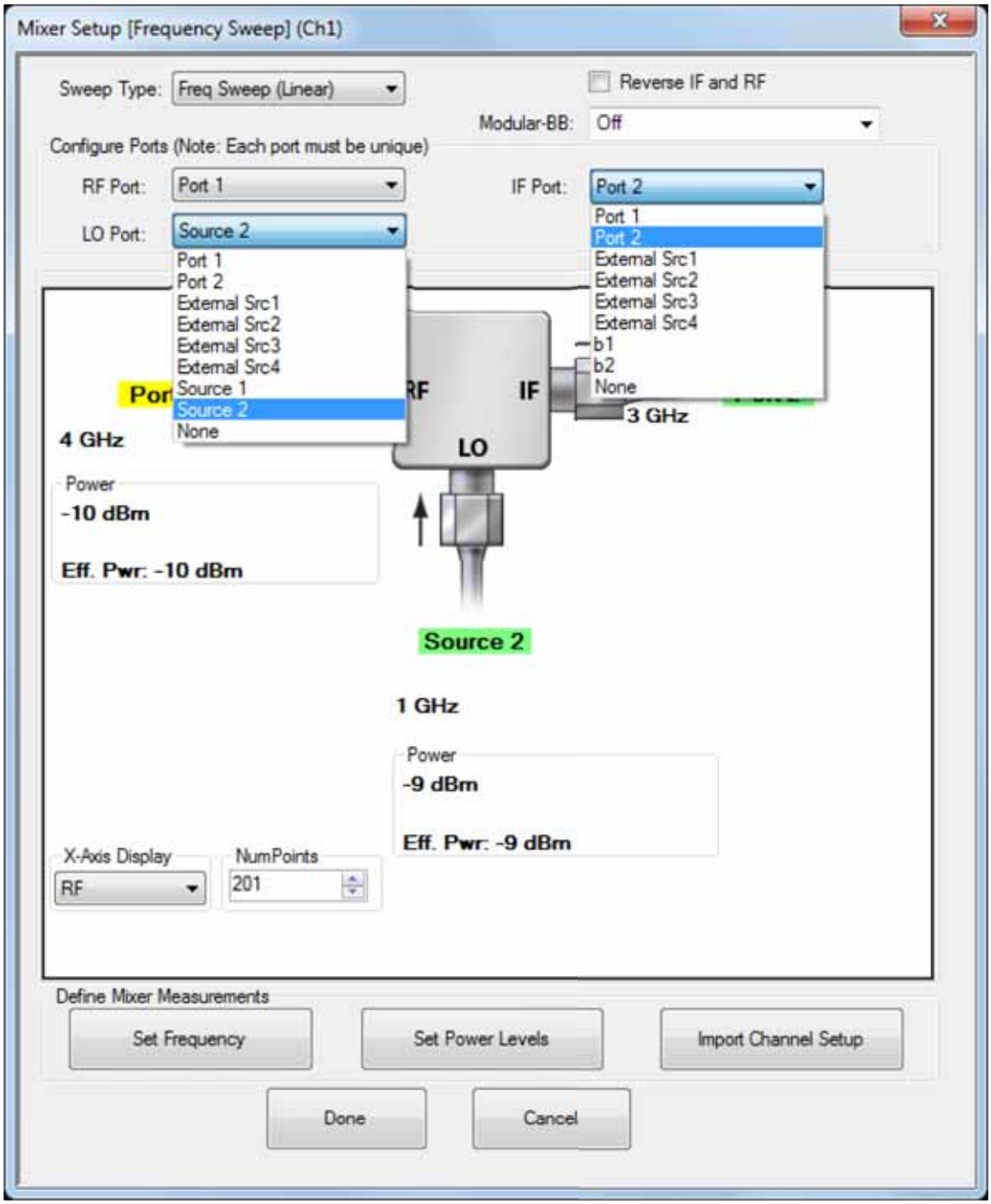
Correspondingly, the power setup is now more detailed as suggested by Figure 21-11. There are no conversion complications, but there could be simultaneous sweeping of power on input and LO (not commonly done, but possible). The swept power input is often done to look at gain compression of the DUT, while LO power sweeps are often employed to look at the saturability of the LO path. The latter can be important since it has many linearity implications depending on the DUT technology.



1. SET POWERS (POWER SWEEP) Wizard Dialog Box
2. Mixer Schematic—Port assignments and power levels are updated automatically.
3. RF Power (dBm) controls.
4. RF Power Mode set as Swept with option of Fixed.
5. Swept start and stop power input fields. The input field is CW Power when Fixed Power Mode is selected.
6. LO Power controls
7. Mode set as Swept with option of Fixed.
8. Swept start and stop power input fields.

Figure 21-11. Power Setup Dialog for the Power Sweep Case

In VNAs equipped with dual sources (Option 31) and one of the loop options (051, 061 or 062), additional options appear in the LO and IF pull-down menus (or LO and RF if the Reverse IF and RF checkbox has been checked) as shown in Figure 21-12.



The modifications to the pull-downs when Option 31 and one of 051/061/062 are available are illustrated here. This allows the second source to be used as the LO in the mixer measurement.

Figure 21-12. Modifications to Pull-downs when Option 31 and One of 051/061/062 are Available

An example mixer setup on a 4-port system is shown in Figure 21-13. The Input (RF) and LO are driven from the two internal sources (not all port combinations are permitted).

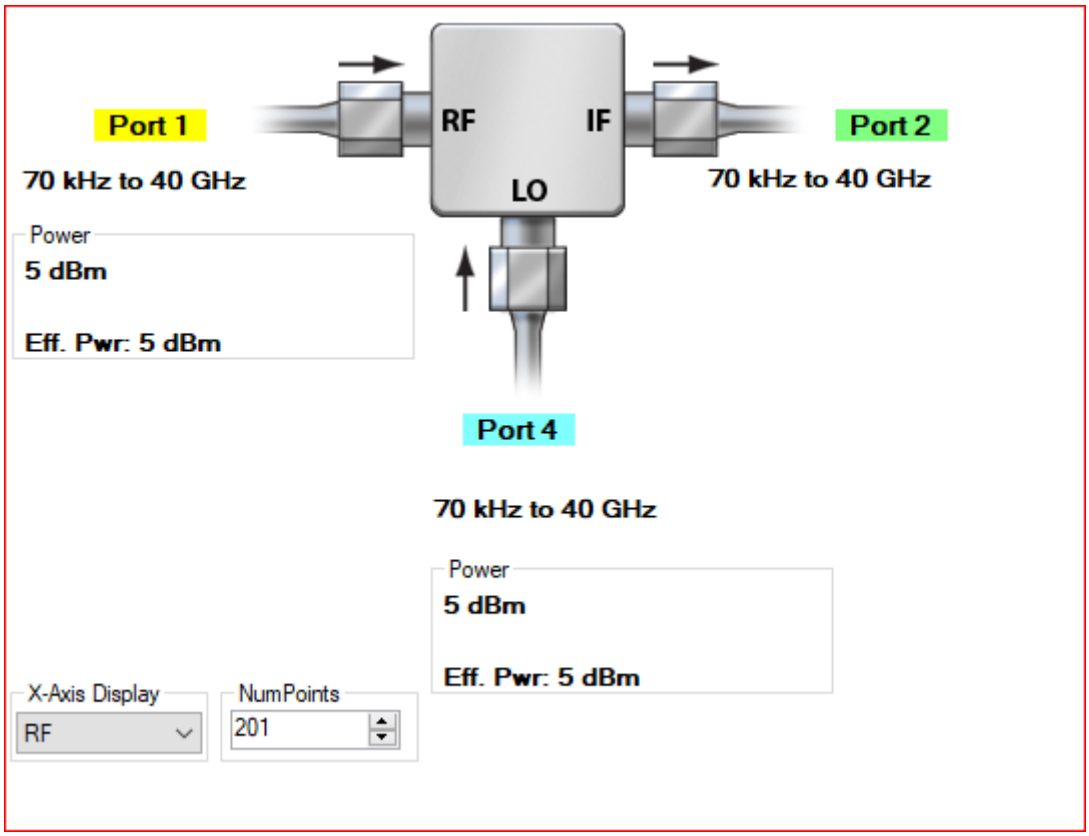


Figure 21-13. (4-Port) Mixer Setup

Most of the concepts discussed above map over to 4-port systems. In the single-source case, the only change is the ability to use any two ports as the input and output (or an LO can be assigned to one and an external synthesizer used for the input). For dual source 4-port systems, there is some additional flexibility in that the two ports can be used for input and LO and an additional port can be used for the output. The only constraint is that the input and LO must be able to access the two sources so allowed port pairings are 1-3, 1-4, 2-3, and 2-4. All mixer calibrations are available. An example setup of this type is shown in Figure 21-14.

As always, external sources can be used for input and LO (with calibration restrictions if used as the input). b1 and b2 can be used for the receiver in the dual source case. Note that these entries still refer to the loop access points on the base VNA (so there are only two).

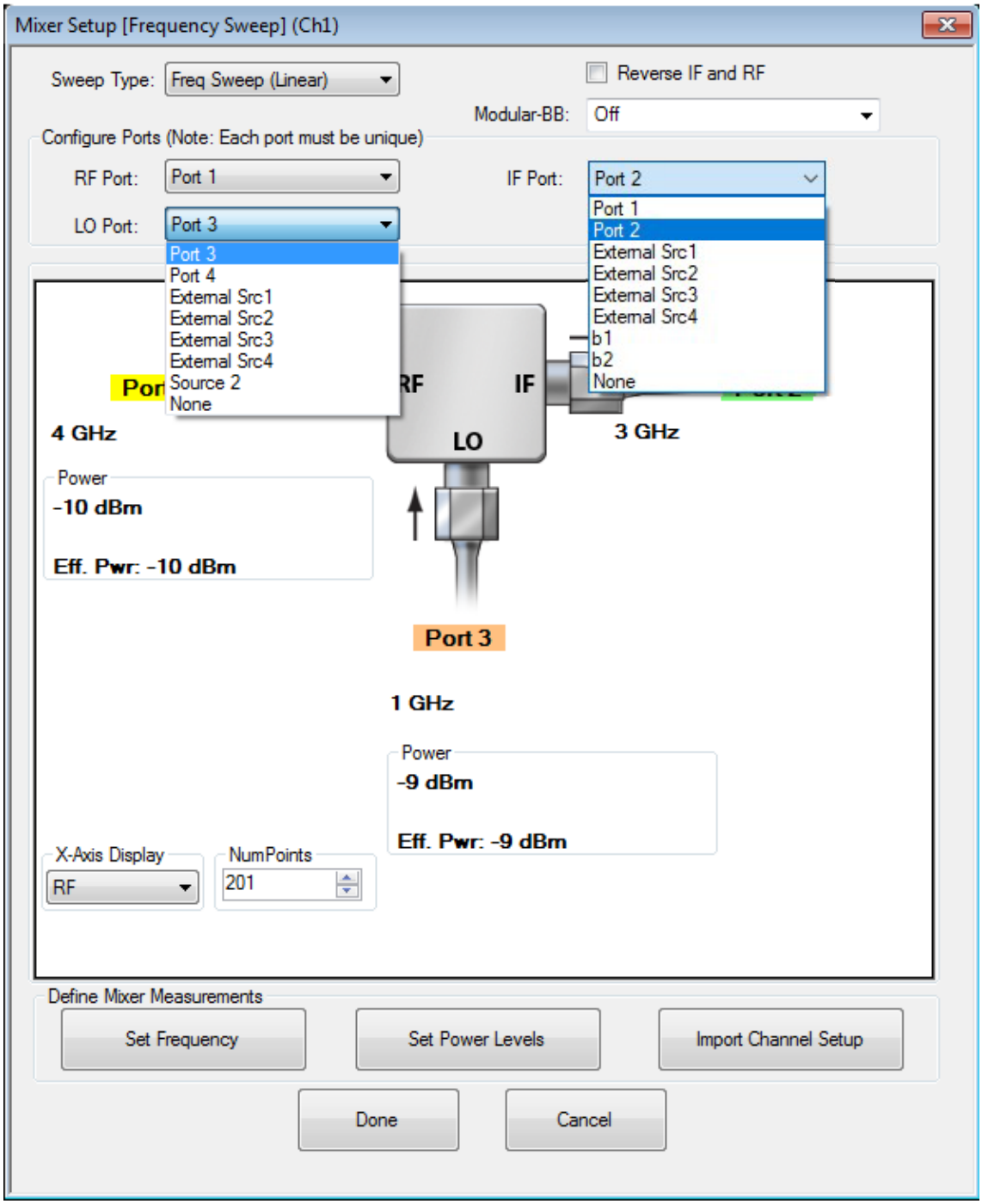


Figure 21-14. (4-Port) Input as Port 1, Output as Port 2 and LO as Port 3

Example Mixer Setups

One can, for example, select source 2 for the LO and then port 2 for the IF or port 2 for the LO and b2 for the IF. These choices all make use of the independent sources available on ports 1 and 2 and the ability to work around the ports to separate LO and IF signals (in this case). To see how these configurations might be arranged, consider the two setups outlined in Figure 21-15.

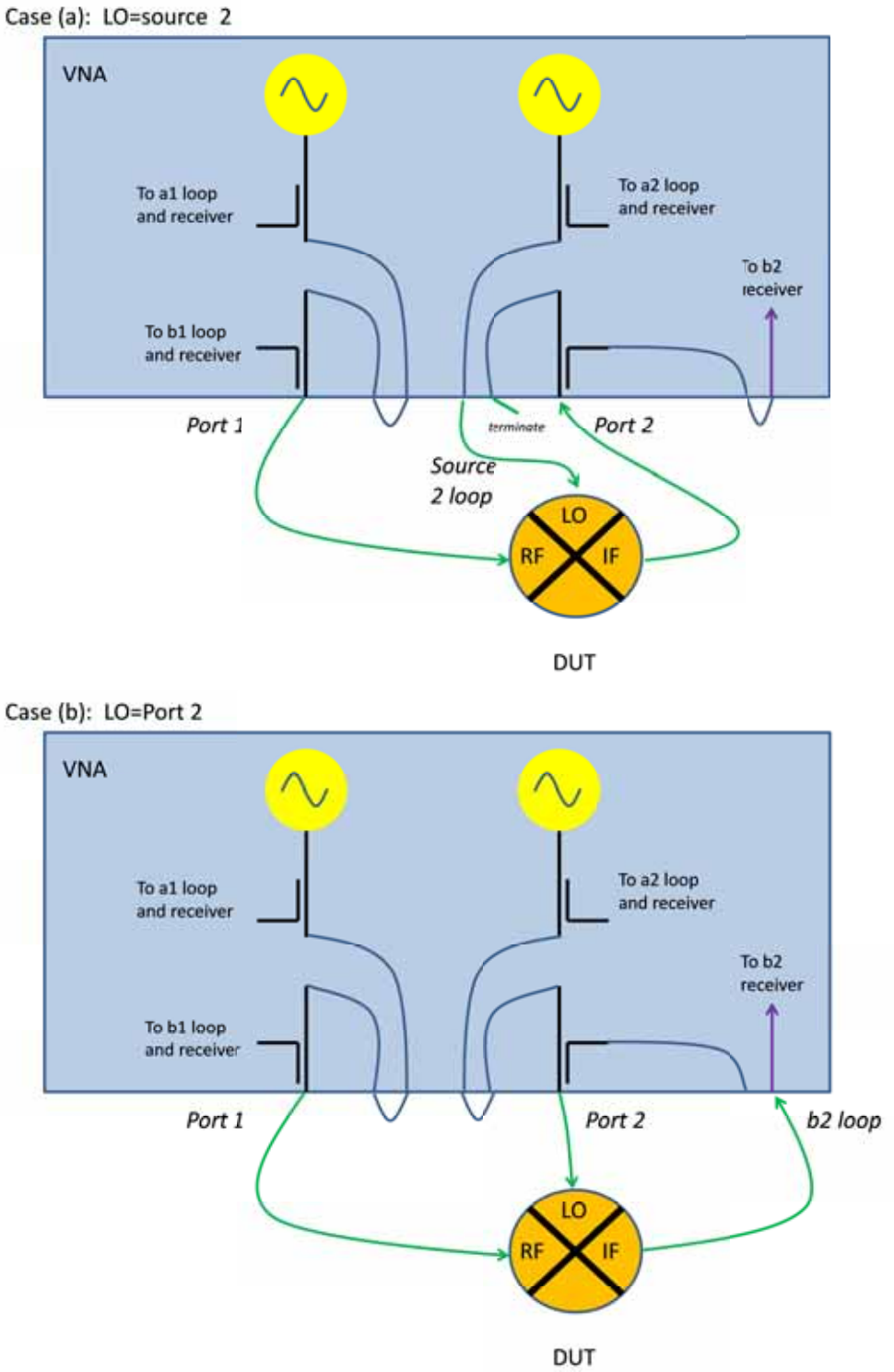


Figure 21-15. Setups for Source 2 and Port 2 LO Drive

- Case (a): LO = Source 2

  In this case, the source loop is used to access the LO drive and the mixer IF can feed the port 2 coupler as usual (one could also select b2 and feed the IF directly into the receiver for more dynamic range). Since the leveling circuitry is behind the source loop, full power control and power calibrations will still be available for the LO. The LO is available from the loop port marked with the tail of the arrow as shown in the figure. It is recommended that the unused loop port (marked with the head of the arrow) be terminated to help improve the raw match that the DUT output will see looking into port 2.

- Case (b): LO = Port 2

  In this case, the LO is fed from the port connector and the IF can be connected directly to the b2 receiver (b2 loop connector marked with the head of the arrow). The noise floor is lower when using the loop receiver access (by about 13 dB) but the compression point is as well (by the same amount) so padding may be necessary for DUTs with output power ~> –3 dBm to avoid system receiver compression. The unused loop port can also be terminated in this case but it is less critical here since it leads to only a coupler arm (isolated from the signal path).

In both of these examples, port 1, source 1 and b1 can be substituted into the discussion if the RF is driven by port 2. Also, all of the terms swap if the Reverse IF and RF checkbox has been checked. Note that if the second source is used as the LO, the enhanced match calibrations will not be available since that calibration relies on full reversing measurements which are not possible if one source is dedicated to the LO drive. When doing a normalization calibration, the relevant connection for the 'thru' step on the output side would be the port 2 connector or the b2 loop connector (head of the arrow) as one might expect.

Note	An additional point for all active channel mixer setups is that the input port frequency range must have a start frequency less than the stop frequency. The output and LO ports are allowed to sweep backwards.

Mixer Power Cal Assistant

At the conclusion of the Active Channel setup (hitting 'Done' and the configuration has been validated), the user will be offered a directed path to a user power calibration process. Such a power calibration is recommended for increased accuracy (particularly at millimeter wave frequencies) and the directed path helps by setting up all of the power calibration variables based on the mixer setup for power calibration at both RF and IF frequencies. An appropriate power meter must still be connected to the system (GPIB or USB) and to the relevant port. If the user skips this directed path, a power calibration can always be orchestrated manually at a later time.

21-6 Using the Active Channel Guided Setup

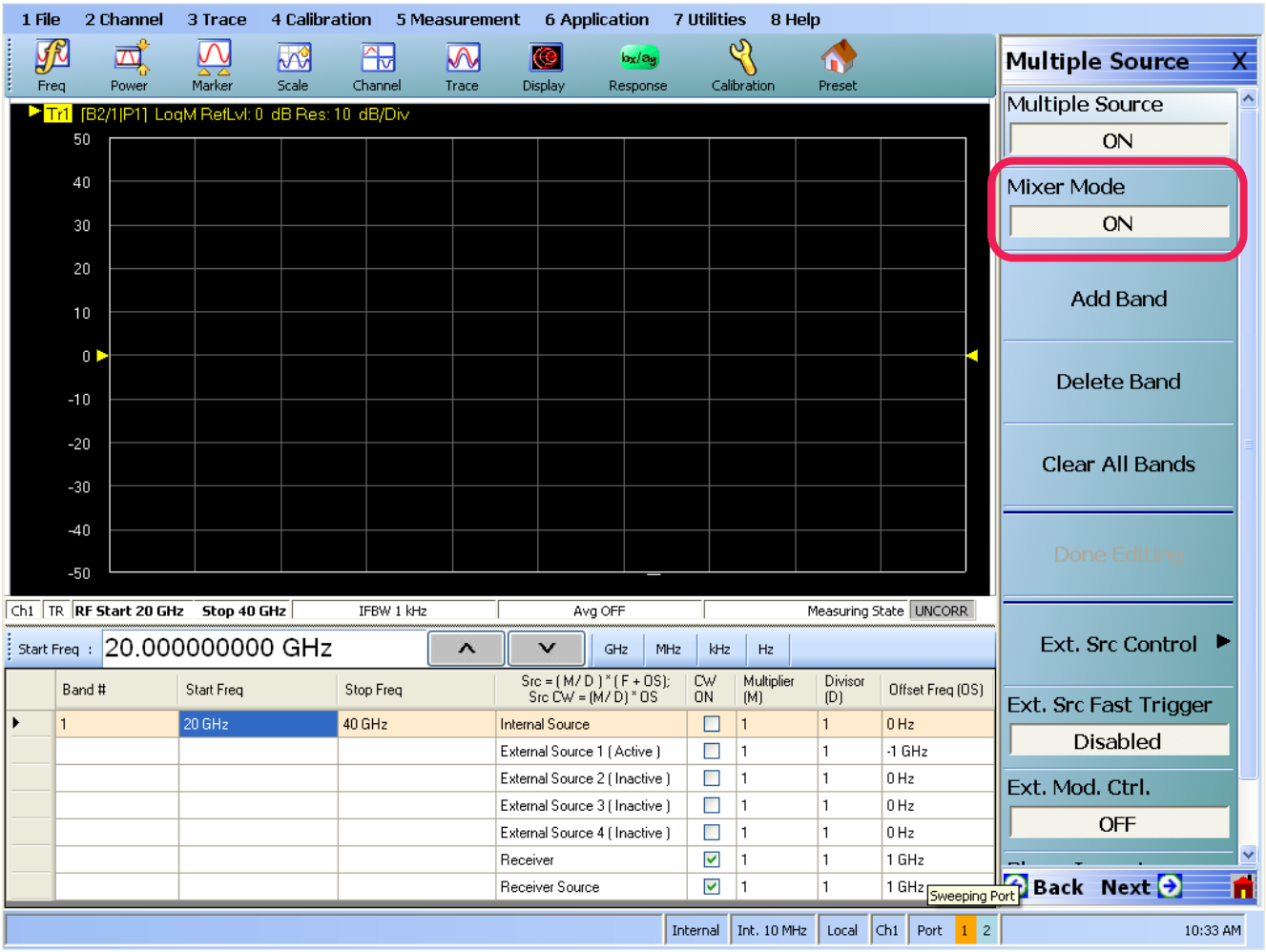
The general procedure for using the guided setup is relatively simple:

1. If a previously used mixer setup exists that is of interest, it may be imported first.
2. Decide on the type of sweep needed (frequency or power)
3. Decide on the frequency plan for the DUT addressing the following parameters as required:
   - What are the Input, LO and Output Frequencies.
   - Which mixer ports are swept and which are fixed.
   - If downconversion is used, is it USB (upper side band) or LSB (lower side band).
   - What are the power levels.
4. Decide on the number of points needed and what X-Axis labeling is desired (input or output frequency in the case of a frequency sweep)
5. Enter the guided setup dialog and input the desired parameters. Auto can be used to help with the frequency entry.
6. Exit the dialog by selecting the OK button. If not successful (i.e., the error checking of your entries indicated a problem), return and fix the errors. When successful, proceed to configure your desired measurements.

Note that when one successfully leaves the dialog, the channel will be setup in converting mode (sourcing at the input frequency and the receiver tuned to the output). If one wishes to perform some non-converting measurements, proceed to multiple source control and put the receiver and source equations where desired. In the dialog of Figure 21-16, if one wanted to measure input match or input-to-output isolation, change the receiver and receiver source equations to 1/1(f+0 GHz).

After using the guided setup, one may find that one wants to change or optimize some settings. One can always return to the dialog and try again. For more advanced modifications, one can also proceed to the multiple source dialog (see Chapter 18, “Multiple Source Control (Option 7)”) and edit all of the equations manually. This multiple source system is automatically populated with the mixer settings when one leaves the guided setup so one can perform more extensive modifications later. This includes the possibility of changing to a harmonic converter or a multiple conversion setup. As shown in Figure 21-16, a button labeled Mixer Mode has been added to the screens shown in Chapter 18, “Multiple Source Control (Option 7)”. This mode enables the extra mixer calibration options available on the calibration menus and the linkage to the guided setup and wizard dialogs.

The reader may have noticed the Enable Modular BB check box in the setup dialog. This check box enables the internal frequency plans for use with the MA25x00A or 3743x mmWave modules for extended frequency range. Generally, this enables the VNA source and receivers to reach 145 GHz (requires opt 08X, the 3739x test set, and the appropriate mmWave modules). The same setup controls apply.



The multiple source screen after a successful guided mixer setup execution is shown here. Note the Mixer Mode button which is not present in traditional multiple source control modes.

Figure 21-16. Multiple Source Screen After Successful Guided Mixer Setup

21-7 Typical Required/Recommended Active Channel Mixer Calibrations

The required calibrations are dependent on what measurements are going to be completed. Since the active channel setup is somewhat less of a guided approach than is the full mixer wizard (to be discussed in a later section), the guidance in this section will be at something of a macro level. All but the power calibrations and traditional receiver calibrations are located in the main Calibration menu system.

For Mixer Port Return Loss

A full 1-port calibration is normally used. The calibration must exist over the runner frequency range for this mode and is often done in standard mode. It can be in the form of a full 2-port calibration to capture isolation terms.

For Isolation

A transmission frequency response calibration can be used or it can be combined with return loss measurements in a full 2-port cal.

For Conversion Loss

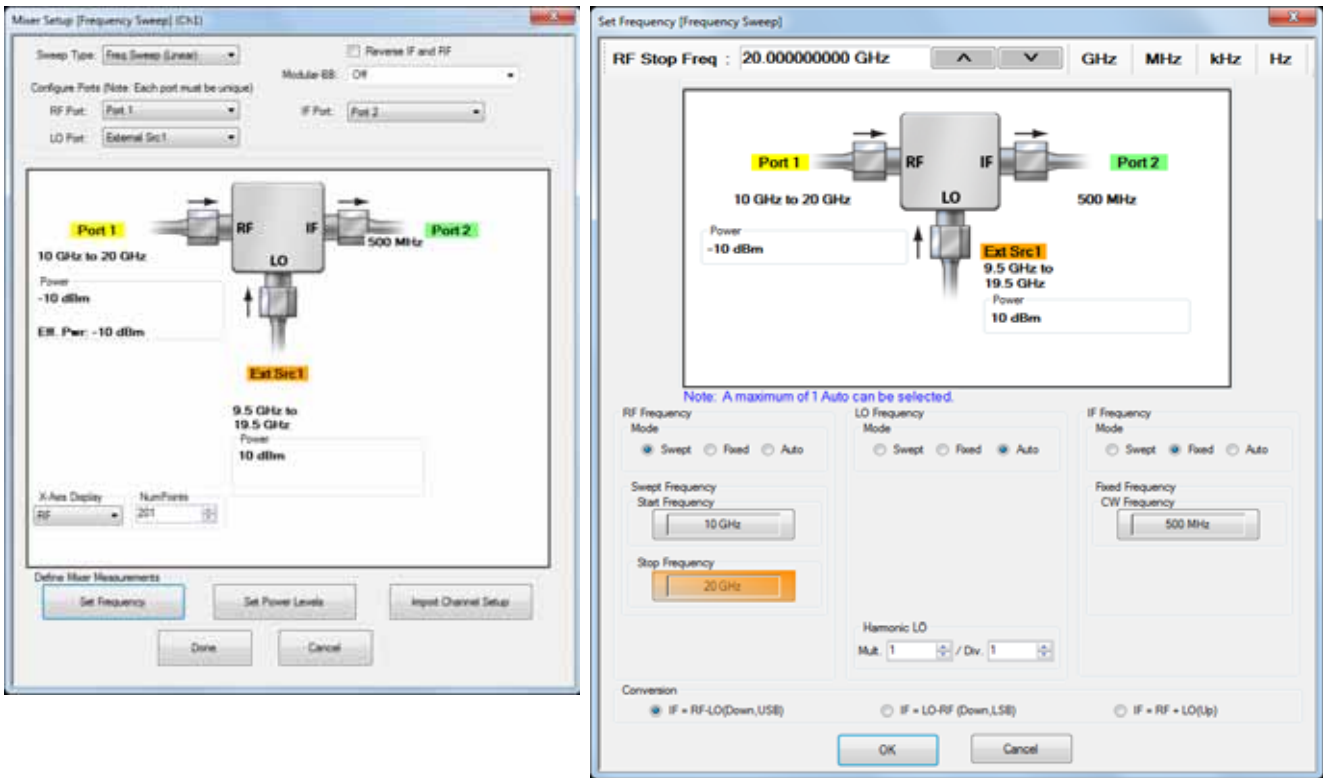
As has been discussed, a normalization-cased calibration or an enhanced match calibration can be used. The latter provides a higher level of correction and provides return loss data as well at the expense of additional calibration steps.

For Power Calibration

A power calibration is also often recommended for conversion loss situations in that it can improve the accuracy of the receiver calibration that is embedded in both types of conversion loss calibrations. The power calibrations are discussed in separate chapters but use a power meter to more accurately establish an absolute power level at a given reference plane (e.g., at the end of a cable where the DUT will be connected).

21-8 Procedure for a Typical Single Channel Mixer Measurement

Consider that one wishes to make a conversion loss measurement of a DUT with a 10 to 20 GHz input range, a fixed 0.5 GHz output and a low-side LO. External synthesizer 1 will be used as the LO (and it is on the GPIB address of 4, which is default for external synth 1) and it has been connected to the GPIB and the 10 MHz reference out of the VNA and turned on. The LO power is to be +10 dBm and the input power is to be –10 dBm. The setup dialogs for this scenario are shown in Figure 21-17. Here we used the 'auto' function on the LO entry to simplify things. Note the selection of the conversion direction on the frequency setup. Once the Done button is selected on the main dialog, the setup will be verified and the system placed into a converting mode where the source at Port 1 is sweeping over 10 to 20 GHz and the receiver is fixed at 0.5 GHz.

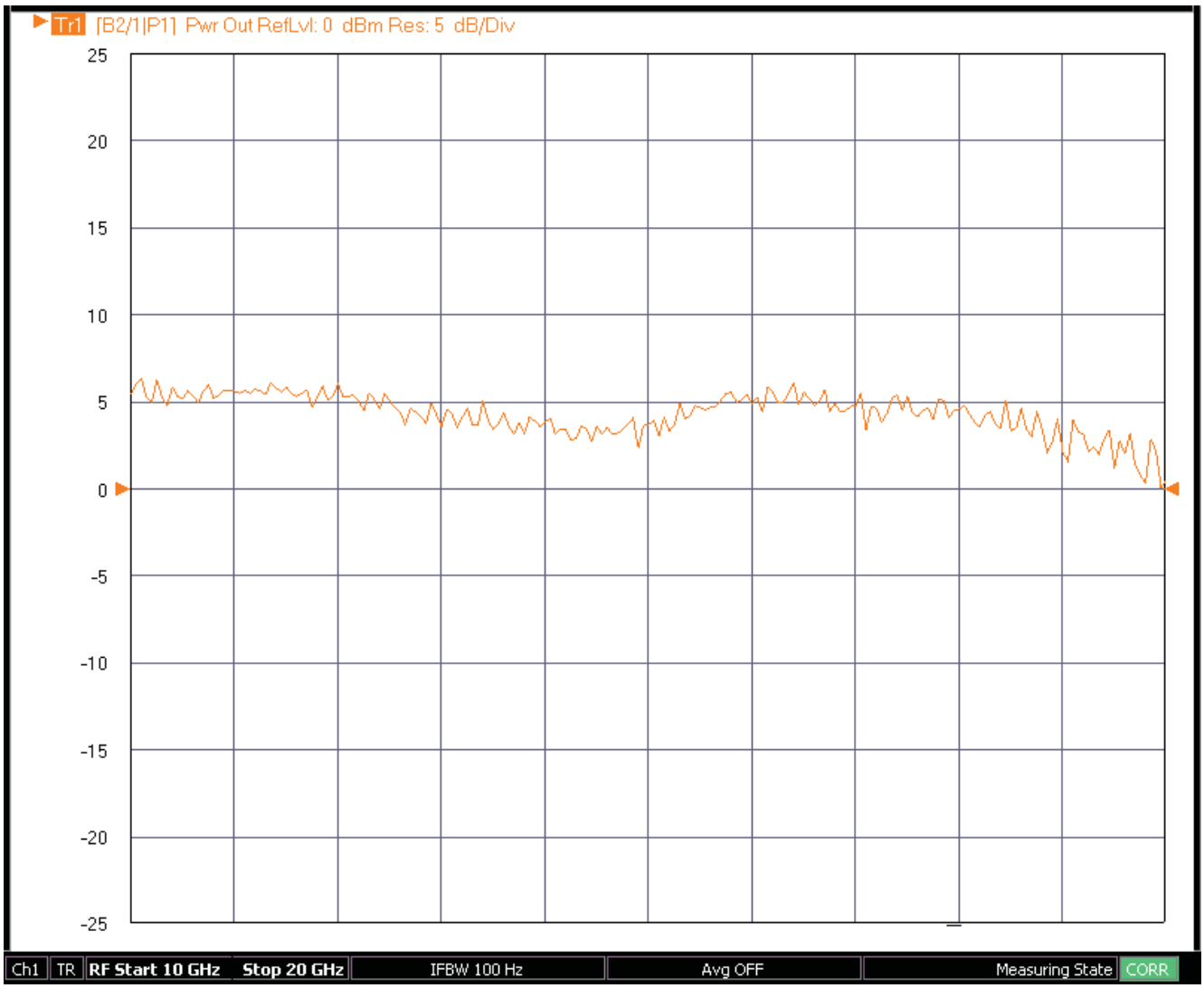


An example setup is shown here for a 10 to 20 GHz downconverter with a fixed IF of 0.5 GHz. The LO range will be calculated to be 9.5 to 19.5 GHz and programmed into an external synthesizer.

Figure 21-17. Example Setup for 10 to 20 GHz Downconverter

An initial step after the setup may be to perform a user power calibration, and a dialog will prompt the user on whether they want to do that. The factory ALC power calibration accuracy (<1 dB generally at microwave frequencies) may be adequate but, if not, the user power calibration can be performed now. Importantly, it is the relative error between the input and output frequency ranges that is of most importance here, since the conversion gain/loss measurement is relative. The user power calibration becomes more useful for very low-IF scenarios and for many mm-wave converter applications. If performed now, this power calibration has the benefit of being automatically performed at both the input and output frequencies of the DUT and merged.

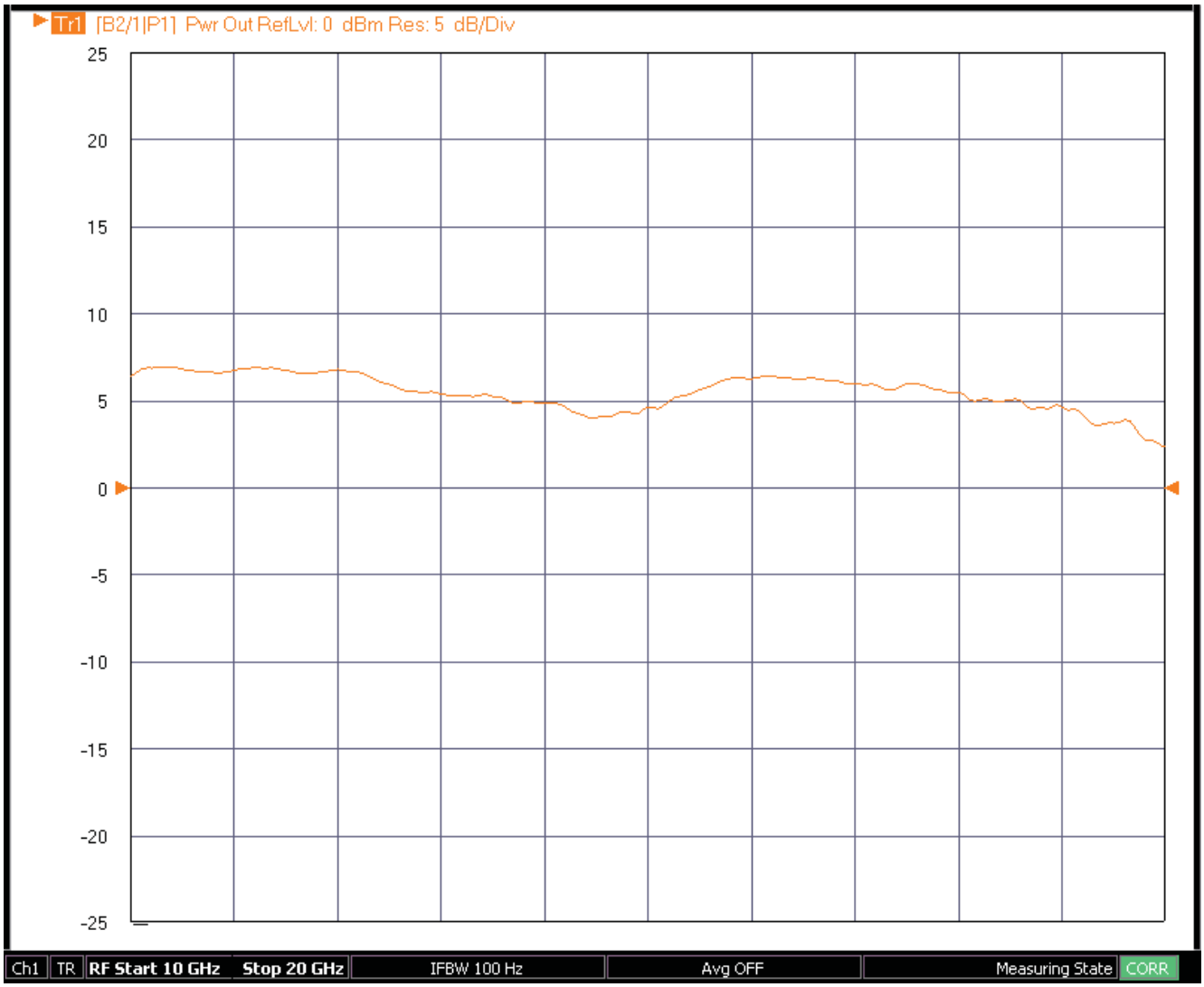
The next step would normally be to go to the calibration menu and select either a normalization or an enhanced match calibration for this conversion loss measurement. During the normalization calibration, only a thru line connection is needed as discussed earlier. The measurement result for this case is shown in Figure 21-18. No port padding was used and the DUT was not particularly well-matched so there is some ripple.



An example conversion gain measurement is shown here using the setup of Figure 21-17 above and a normalization-based calibration.

Figure 21-18. Example Conversion Gain Measurement with Normalization-Based Calibration

An enhanced-match calibration was also performed on this setup and the measurement results are shown in Figure 21-19. A standard manual OSL calibration kit in the K connector (3652X) was used for the reflectometer calibration steps. The measurements were not identical due to some bias changes on the DUT but the reduction in ripple is apparent.



An example conversion gain measurement is shown here using the setup of Figure 21-17 and an enhanced match calibration.

Figure 21-19. Example Conversion Gain Measurement with Enhanced Match Calibration

21-9 Multiple Channel Mixer Wizard Controls

The multiple channel mixer wizard is, in some sense, a natural extension of the single channel setup method introduced previously. The difference here is that multiple measurements are orchestrated for the user in different channels with the corresponding sweeps already setup and the calibrations suggested. The main configuration dialog is shown in Figure 21-20. In addition to some of the controls discussed previously are a list of possible measurements separated into 3 groups: between input and output, between LO and output, and between LO and input. The three groups require different connections since the LO port needs access to receivers in the second two sets. All measurements can be set up at once, but the results from different groups will only be meaningful if the DUT is connected correctly.

All of the measurements shown have been discussed previously, but now they are automatically assigned to different channels that will control the different sweeps. To see this concept, suppose one requests forward conversion gain/loss (input to output), isolation (input to output) and isolation output to input. The first measurement requires the source to sweep over the input frequencies while the receiver sweeps over the output frequencies. The second measurement requires both the source and receiver to sweep over the input frequencies. The third measurement requires both the source and receiver to sweep over the output frequencies. By placing these three measurements in different channels, all the measurements can be completed in one 'super sweep' without user intervention and all of the results displayed at once.

There is a Starting Channel button on this dialog to indicate which channel should have the first measurement and the others will increment from there. While the default is to start with channel 1, the user may have setups in one or more channels that they would like to keep sweeping in addition to those measurement orchestrated by the mixer wizard.

The power and frequency setup dialogs are essentially the same as for the single channel setup system and will not be repeated here. Minor differences can be analyzed in the operating manual or the UI guide. For the power sweep case, the choices are more limited and are shown in Figure 21-21. As had been discussed previously, conversion gain as a function of input power (compression) may be of interest as can conversion gain versus LO power (looking for LO saturation while the input power is constant). Unlike the manual method in the single channel setup, here the swept power on the respective port will be requested.

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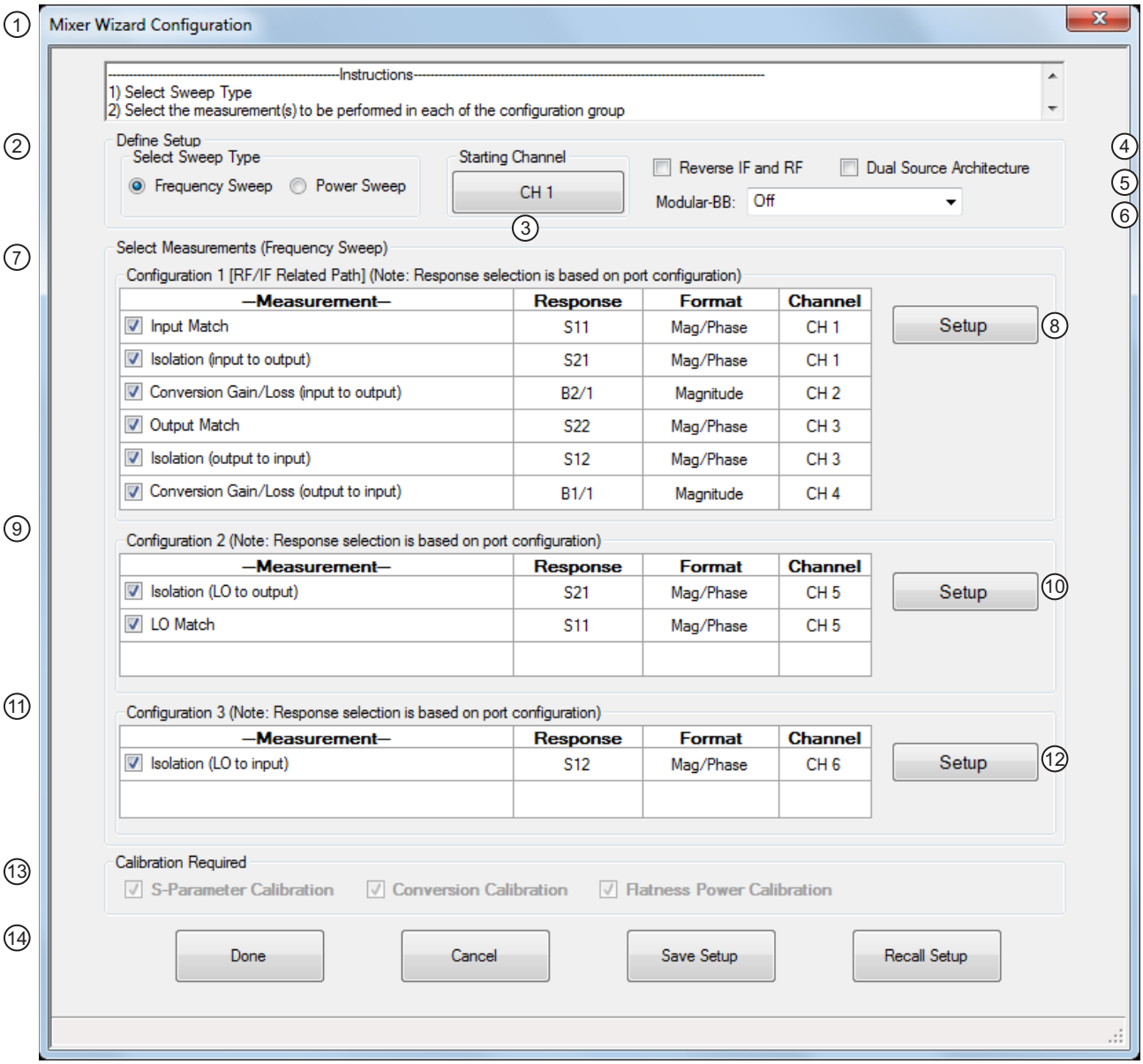
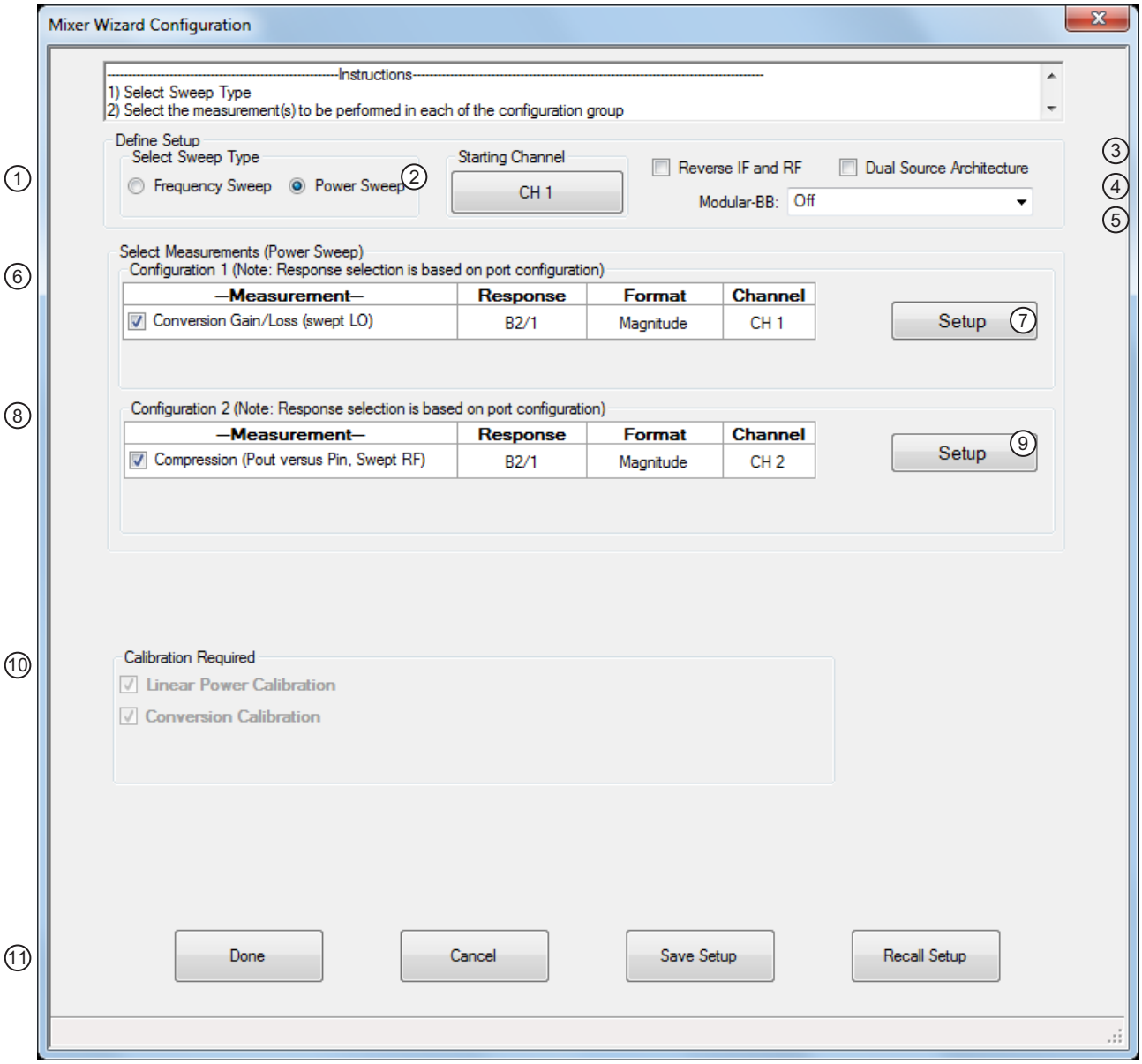


Figure 21-20. Mixer Wizard Configuration (Frequency Sweep) Dialog Box—Control Areas (1 of 2)

1. Dialog Box
2. Select Sweep Type (Frequency or Power Sweep).
3. Channel Start Selection
4. Reverse IF and RF—Default is RF as input and IF as output. Toggles to RF as output and IF as input.
5. Modular Broadband selection dropdown—Only for MS4647B VNAs with Option 8x BB/m-Wave System.
6. Dual Source Architecture (Only for MS464xB VNAs with Option 31) Note that when this feature is selected, the measurement selections change as shown in the figure below.
7. Configuration 1—Conversion Gain/Loss Swept LO—RF/IF Related Path—Select any combination of measurements. Channels are automatically assigned based on Starting Channel.
8. Setup Button—Links to Configuration 1 Dialogs to configure mixer and VNA ports, external sources, frequency, and power levels.
9. Configuration 2—Compression Pout versus Pin Swept RF—Select any combination of measurements. Channels are automatically assigned.
10. Setup Button—Links to Configuration 2 Dialogs to configure mixer and VNA ports, external sources, frequency, and power levels.
11. Configuration 3—Isolation—LO to Input Available Measurement
12. Setup Button—Links to Configuration 3 Dialogs to configure mixer and VNA ports, frequency, and power.
13. Calibration Required Area
14. Dialog Controls

Figure 21-20. Mixer Wizard Configuration (Frequency Sweep) Dialog Box—Control Areas (2 of 2)



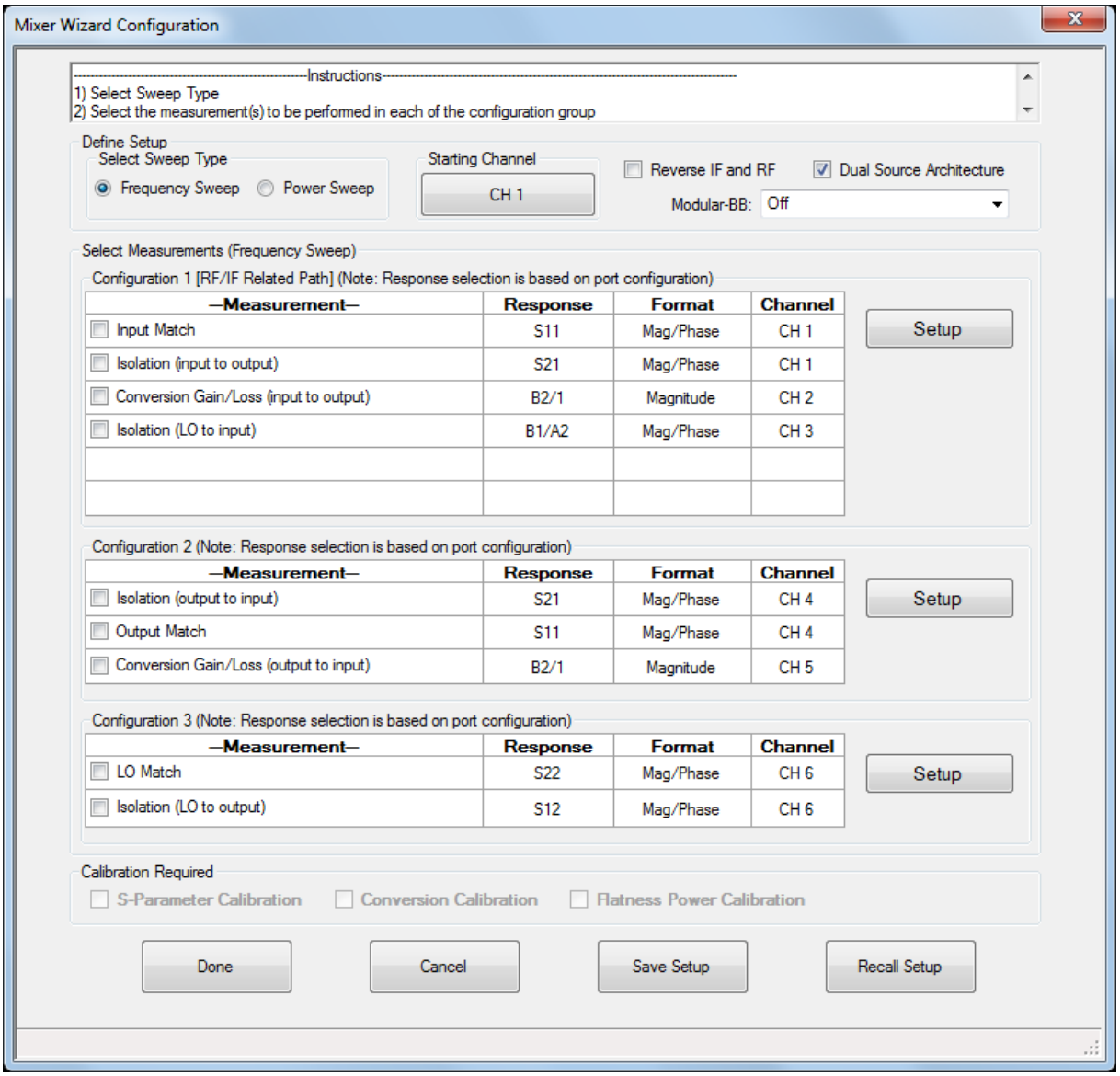
1. Select Sweep Type
2. Select Starting Channel
3. Reverse RF and IF—Standard is RF is input and IF output. Reverse is RF as output and IF as input.
4. Modular Broadband selection dropdown
5. Dual Source Architecture (Only for MS464xB VNAs with Option 31) When this feature is selected in Power Sweep mode, the measurement selections do not change.
6. Configuration 1 Measurements for Conversion Gain/Loss Swept LO.
7. Setup button to Configuration 1 dialogs
8. Configuration 2 Measurements for Compression Pout versus Pin Swept RF.
9. Setup button to Configuration 2 dialogs.
10. Calibration Required Area display.
11. Dialog control buttons.

Figure 21-21. Mixer Wizard Configuration (Power Sweep) Dialog Box—Control Areas

Note	An additional point for all multi-channel mixer setups is that the input port frequency range must have a start frequency less than the stop frequency. The output and LO ports are allowed to sweep backwards.

As with the active channel mixer setups, additional features can become available, and additional limitations incurred, when using the second source and loops (Option 31 and one of Options 51/61/62). In the setup dialog, there is a Dual Source Architecture checkbox that will only be visible when the above options combination exists. By selecting this option, the user is indicating the desire to use the second source as an LO. This selection is mutually exclusive with the Enable Modular-BB selection.

When selected, the measurement choices change somewhat as shown in Figure 21-22 because the use of the second source requires some measurements to move to different setup groups. As with the active channel case, when the second source is used as an LO (the checkbox selected in this case), the enhanced match calibrations will not be available for the conversion gain/loss measurements since full reversing measurements (that would use the second source) required for those calibrations are not possible with this type of setup.



The multi-channel mixer setup dialog with Dual Source Architecture selected is shown here for the case of using the internal second source as a DUT LO.

Figure 21-22. Multi-channel Mixer Setup Dialog with Dual Source Architecture Selected

The port/source selection setups are constrained when the Dual Source Architecture checkbox has been selected to simplify the user interface as suggested by Figure 21-23. As always, more complex setup possibilities are available using multiple source control (refer to Chapter 18, "Multiple Source Control (Option 7)" for more information).

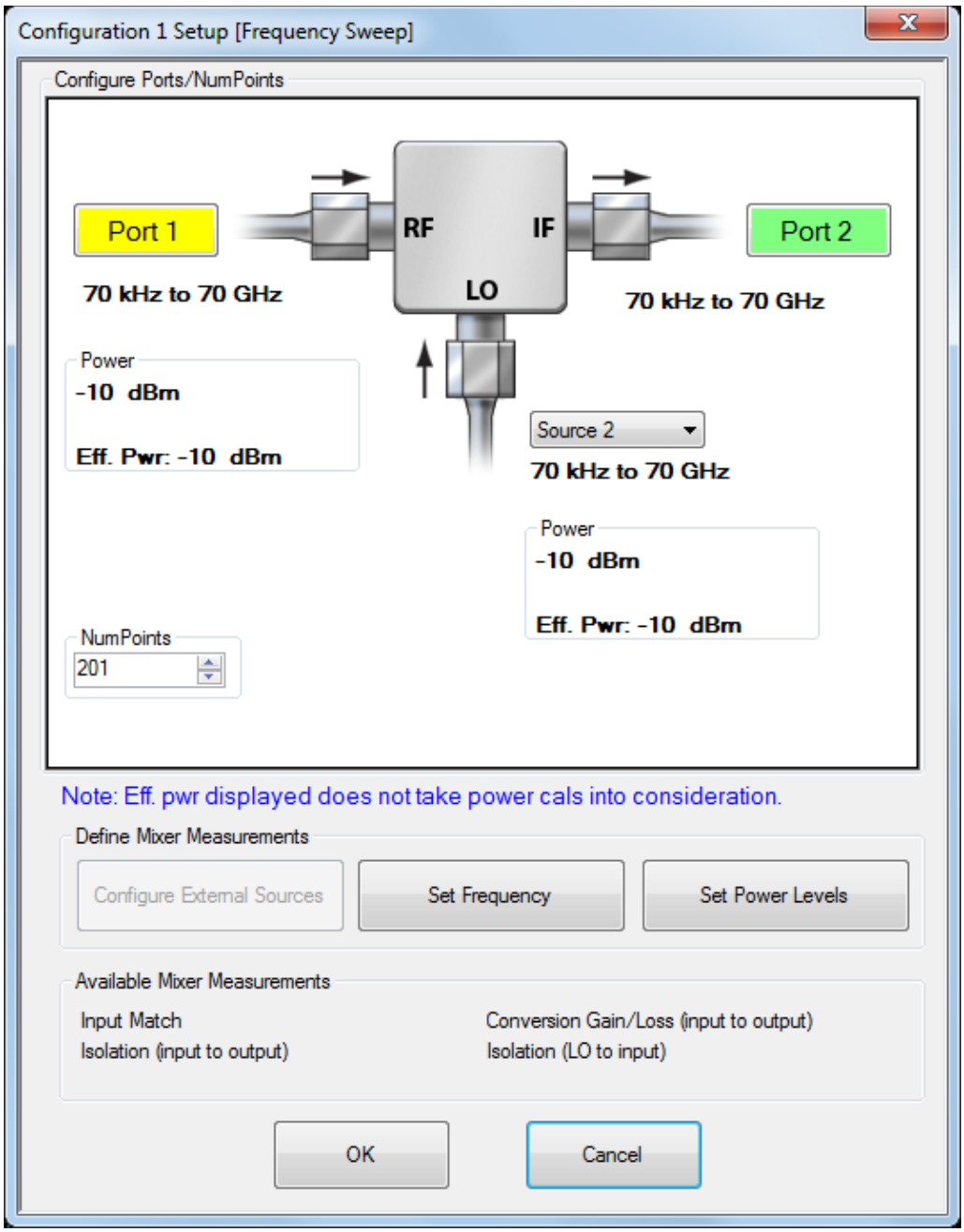


Figure 21-23. Configuration 1 Setup Sub-dialog for Dual-source Selection 2-Port Case

The input and output are constrained to be either port 1 or port 2. The LO is forced to be either source 1 or source 2. The latter case corresponds to the Case (a) setup from Figure 21-15. Again, since the leveling circuitry is behind the source loop (used for source access), full power control and power calibrations will be available for the LO. As in Figure 21-15, it is recommended to terminate the unused loop port to improve the raw match that the DUT output will see looking into port 2. Also, the RF-IF assignments will swap if the Reverse IF and RF checkbox has been selected. If it is desired to use an external source for the LO (or input), then both internal sources are not required and the Dual Source Architecture checkbox should not be selected.

With 4-port single-source systems, the multi-channel setups are basically the same as for 2-port single-source cases except any of the ports can be used as input and output (and, again, an external source can be used as input with some calibration limitations). The dual source architecture selection has slightly different effects.

- For setups 1 and 2, the input, output and LO can be any port, but there are some combination restrictions. If the LO is port 3 or port 4, then the input and output should be selected amongst port 1 and port 2. If the LO is port 1 or port 2, the input and output should be selected amongst port 3 and port 4.

- For setup 3 (dealing with LO paths), if the LO is port 1 or port 2, b1 is not allowed for the output. If the LO is port 3 or port 4, b2 is not allowed for the output.

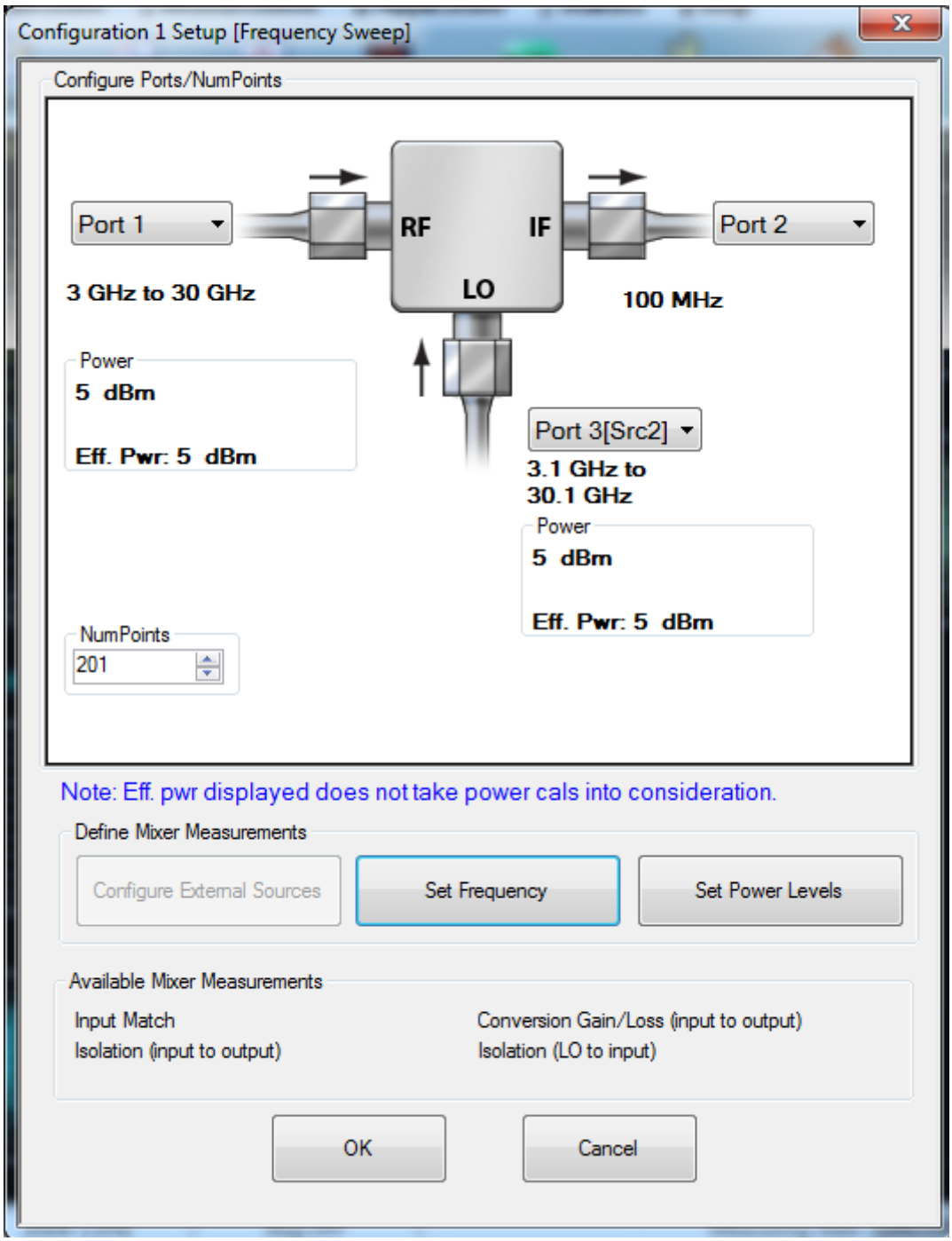


Figure 21-24. Configuration 1 Setup Sub-Dialog for Dual-Source Selection 4-Port Case

21-10 Using the Multiple Channel Mixer Wizard

The procedure for using the wizard is essentially the same as that discussed in "Using the Active Channel Guided Setup" on page 21-20 above for the single channel guided setup except now, one may want to more carefully list the needed measurements so that they can be selected in the dialog. Once the setup is complete, the wizard will issue a list of recommended calibrations for the various channels that can then be executed. Once those are complete, the channels will complete all requested measurements (for a given DUT connection setup).

21-11 Typical Required/Recommended Multi-Channel Mixer Calibrations

The types of calibrations were already discussed in Section 21-7 "Typical Required/Recommended Active Channel Mixer Calibrations" and in earlier sections. The multi-channel wizard will provide a dialog indicating suggested calibrations based on the measurement requested (see Figure 21-25). In this case, input match, input-to-output isolation, input-to-output conversion gain/loss and output match were requested.

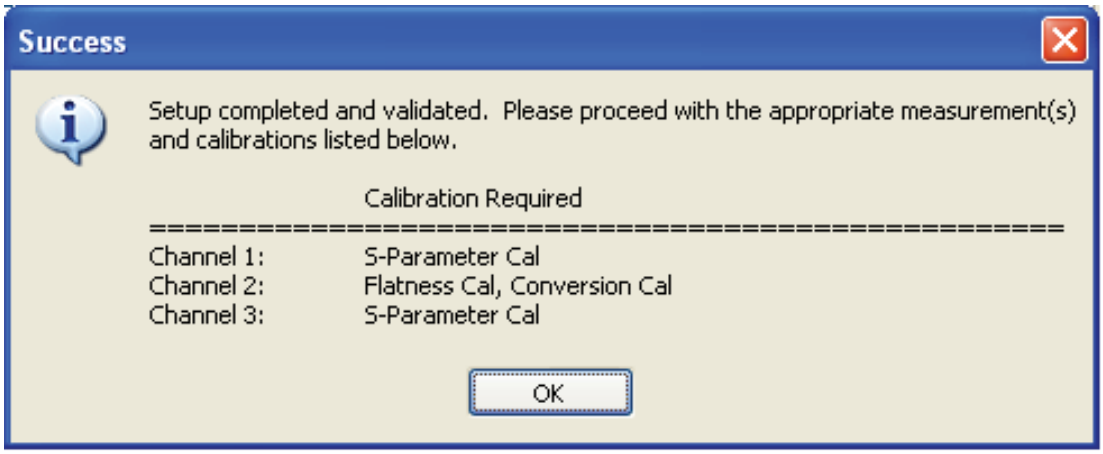


Figure 21-25. Example Listing of Recommended Mixer Calibrations

21-12 Procedure for a Typical Multiple Channel Mixer Measurement

We will extend the example of Section 21-8 "Procedure for a Typical Single Channel Mixer Measurement" to include additional measurements. The setup parameters are the same as before except now we will request input match, output match, forward conversion loss, and input-to-output isolation. The measurement results are shown in Figure 21-26.

- **Channel 1:**

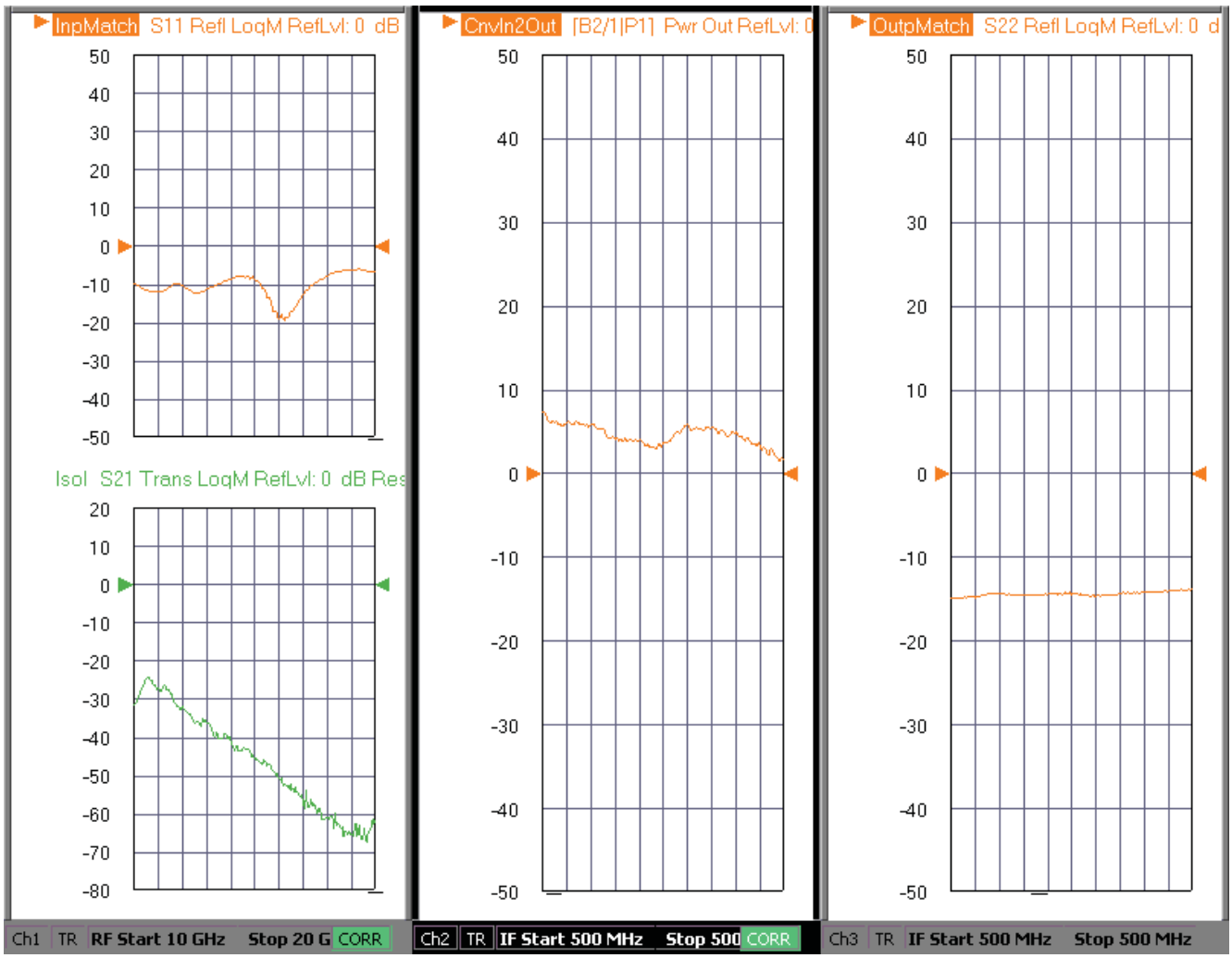
  Sweeping both source and receiver 10 to 20 GHz to measure input match and input-to-output isolation

- **Channel 2:**

  Sweeping the source 10–20 GHz with the receiver fixed at 0.5 GHz to measure conversion loss (actually more is going on here since an enhanced match calibration was used so this channel also measures DUT input and output match behind-the-scenes).

- **Channel 3:**

  CW measurement at 0.5 GHz to get the output match. Note that although the source and receiver are CW during this measurement, the LO is sweeping so one does see a variation over the sweep. Since the LO sweep range is known to be 9.5 to 19.5 GHz (or one can plot with this x-axis), the correlation can be studied.



The three channels of measurement results for our example are shown here.

Figure 21-26. Three Channel Measurement Results

Per the recommendation dialog (Figure 21-25), calibrations were performed independently in the three channels. Channels 1 and 3 received standard S-parameter calibrations using SOLT and a K calibration kit. Channel 2 received an Enhanced Match calibration to get match-corrected conversion gain. A user power calibration was also performed before the Enhanced Match calibration on this channel.

21-13 More on Mixer Calibrations

While many of the mixer calibration topics with regard to conversion gain/loss have already been discussed in passing, it may be helpful to review some of their attributes. When a mixer has been set up using the tools described, the manual calibration menu changes to that shown in Figure 21-27. The two bottom buttons describe two options

- **Rcvr Cal/Normalization:** A receiver cal is performed (as part of the 'thru' step) to establish absolute power accuracy and then a normalization is done to take into account the power incident on the mixer input port. A separate receiver calibration does not need to be performed. A power calibration is recommended for increased accuracy and this must be executed separately. Ideally, the power calibration would be performed over input and output frequency ranges (using segmented sweep can save time). Note that for mmWave applications, the system should be in the proper leveling mode when the calibration is performed (for example, RF leveling for the 3743x modules).

  This calibration is very fast to perform and only one sweep is needed during the measurement, but as will be discussed in the uncertainties section, does become less accurate as the DUT match degrades.
- **Enhanced Match:** This is the process that was described in Section 21-4 "Mixer Measurements" where match components at the input and output of the mixer are taken into account. This calibration is structured very much like a traditional full 2-port calibration, except several different frequency lists are employed: input range→input range, input range→output range, output range→output range.

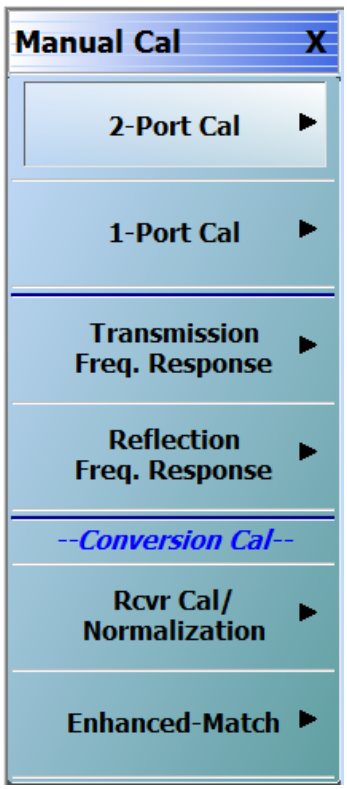


Figure 21-27. The MANUAL CALIBRATION Menu When in Mixer Mode.

The enhanced match calibration can be performed in any media type and with any of the defined-standards calibration algorithms. An example of the setup dialog is shown in Figure 21-28. Note that the direction of the calibration is fixed by the input/output relationship defined in mixer setup and reciprocal connections are not available due to the power transfer computation requirement. The True Source Match checkbox was described in Section 21-4 and refers to a different set of internal calibration steps used, particularly for millimeter wave applications, when a more accurate measurement of the driving match that the DUT sees is needed. This selection will almost never make things worse (one exception being if the driving network is marginally stable when out-of-band due to the addition of an amplifier in the system loops) and can improve conversion loss flatness at higher frequencies.

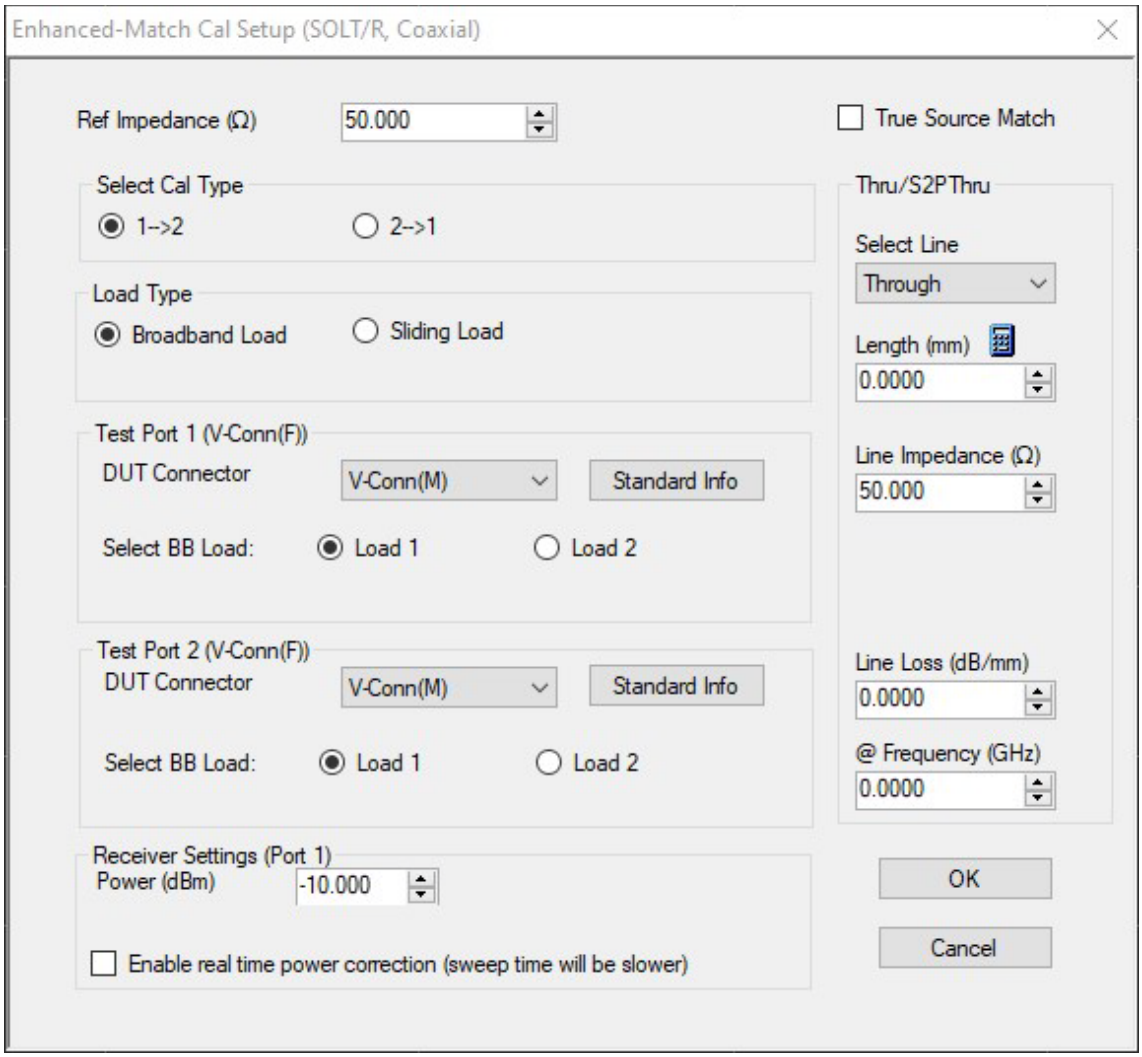
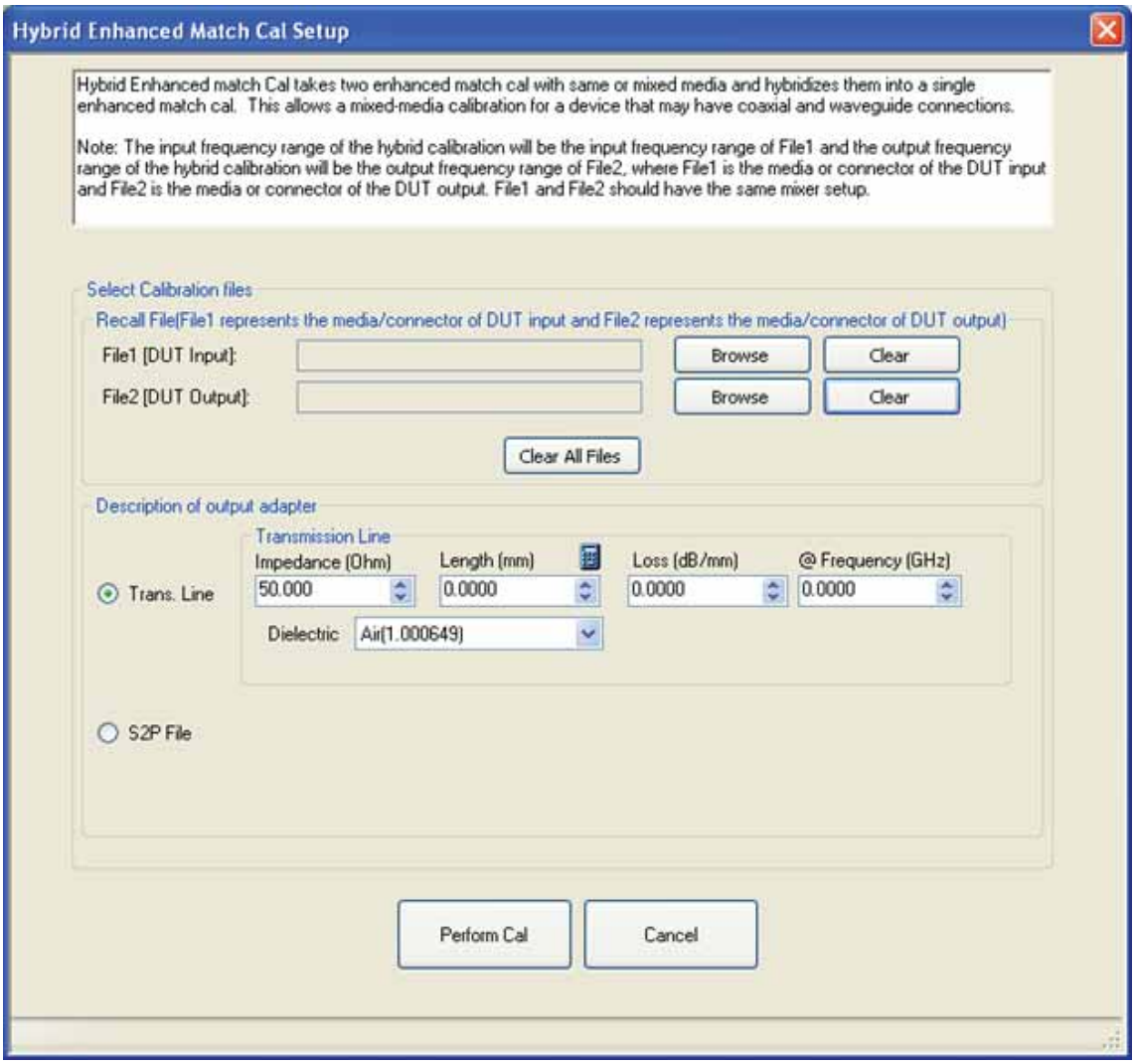


Figure 21-28. Example Enhanced Match Cal Setup Dialog

Systems with options 06x will have an 'attenuation' field visible in the lower part of the dialog. Systems in broadband mode will also have a 'Power > 54 GHz' field.

When an enhanced match calibration is being applied, one may note that three sweeps are required. These are used to get DUT input and output match (non-frequency converting measurements in different ranges) as well as conversion. In certain cases, one wishes to perform an enhanced match calibration but the input and output media of the DUT are different (for example, waveguide and coax) and perhaps different algorithms need to be used on those ports (for example, SSLT and SOLT). Analogous to the hybrid calibrations referenced for non-frequency converting measurements earlier in this guide, there is a hybrid enhanced match calibration to handle this case for mixers. The dialog, accessible from the ALTERNATIVE CALS menu (via the CALIBRATION menu, is shown in Figure 21-29.



The Hybrid Enhanced match calibration dialog is shown here. This approach is helpful for mixers with different connection media at input and output (such as a waveguide input, but a coaxial output).

Figure 21-29. Hybrid Enhanced Match Calibration Setup Dialog

The process, again analogous to regular hybrid calibrations, is that one performs an enhanced match cal in each of the media types (and/or algorithm choices), and this utility combines them. The key point is that each calibration is not frequency converting, so the relevant pieces of the calibration for each media type can be extracted to form the hybrid calibration.

Consider an example using the ME7838A/AX broadband system.

- Input 91–95 GHz in WR-10 (will need to use a SSLT calibration kit at this plane)
- Output 1–5 GHz in coax (will use an SOLT calibration kit at this plane)
- LO (not shown) fixed at 90 GHz

Two complete enhanced match calibrations are completed where the ME7838A/AX system (using 3743A/AX mmWave modules) natively use 1 mm connectors. Coax-to-waveguide adapters are in place for the first enhanced match calibration and are removed for the second. The same mixer frequency plan is used for each calibration. Schematically, the calibration reference planes are as shown in Figure 21-30. We desire the final reference planes to be at A and B.

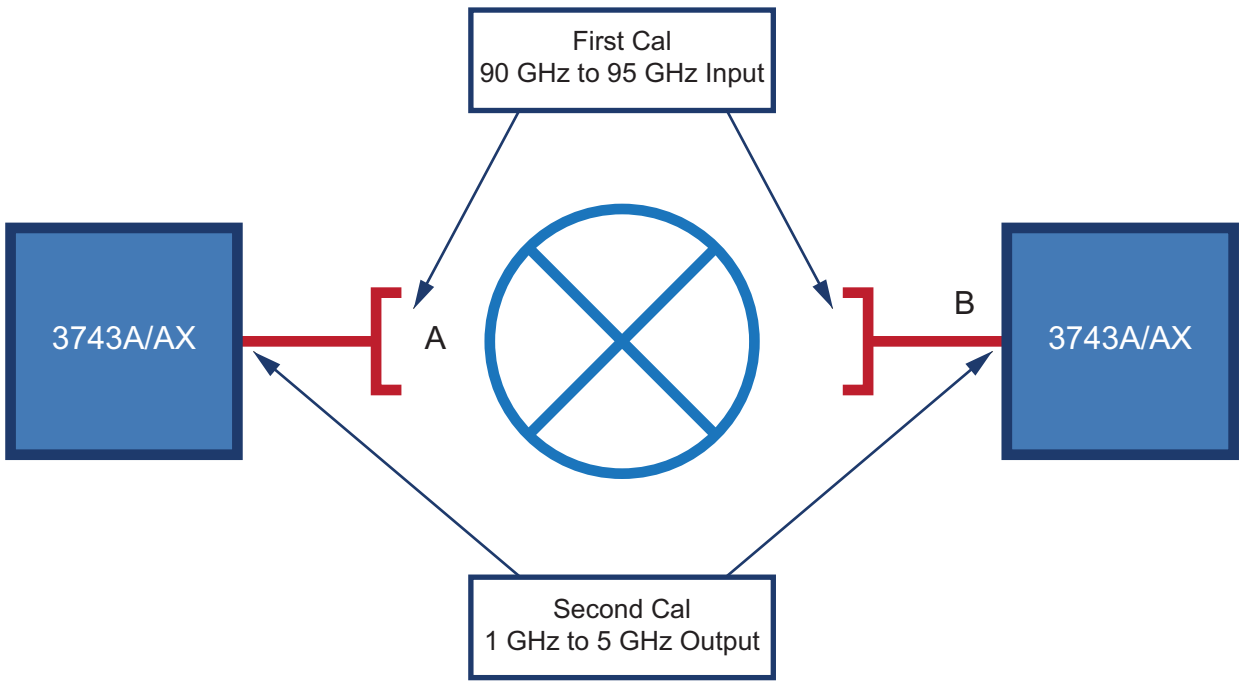


Figure 21-30. Example Hybrid Enhanced Match Calibration Setup

During the first (waveguide) calibration, one may notice some data being acquired that is nonsensical since it is below the waveguide cut-off. This ends up being OK since those measurements are not used in the hybrid computation, but leaving them in the execution sequence allows the instrument to relatively easily offer almost all calibration permutations. When the two calibrations are performed and saved, they can be loaded into the dialog of Figure 21-29.

The final step is a description of the output adapter, which is needed to get the power calibration transfer as accurate as possible. A simple transmission line model can be used or a .s2p file (covering the input frequency range, preferably) can be entered. In the latter case, interpolation and flat-line extrapolation of the file will be used when the frequency lists do not match. Normally this adapter loss is much less than 1 dB and it does not have a great effect, but there are exceptions based on the user setup.

Another important point is what happens to the conversion results (using either receiver/norm or enhanced match calibrations) when the drive power is changed after calibration. Since the input signal to the DUT is measured during calibration and used to compute conversion gain, there is the potential for confusion if that input power changes later.

In the MODIFY CAL dialog is a power field. This field will specify at which level the receiver calibration portion will be performed. Ideally, this would be at the level where a user power cal had been performed or, if none, at default power. This can now be different from the power where the normalization activities occur which are dictated by the main power menu settings. If there are no changes, both receiver calibration and normalization will happen at default power. For option 06x systems, there is also an 'attenuation' field where the drive side source attenuator can be set for the receiver calibration. For broadband systems, there is an additional field for power > 54 GHz so the power during receiver calibration can be configured by band. This allows the receiver cal and normalization to be decoupled, in terms of power level, which can be important when there are large gain differences in the paths.

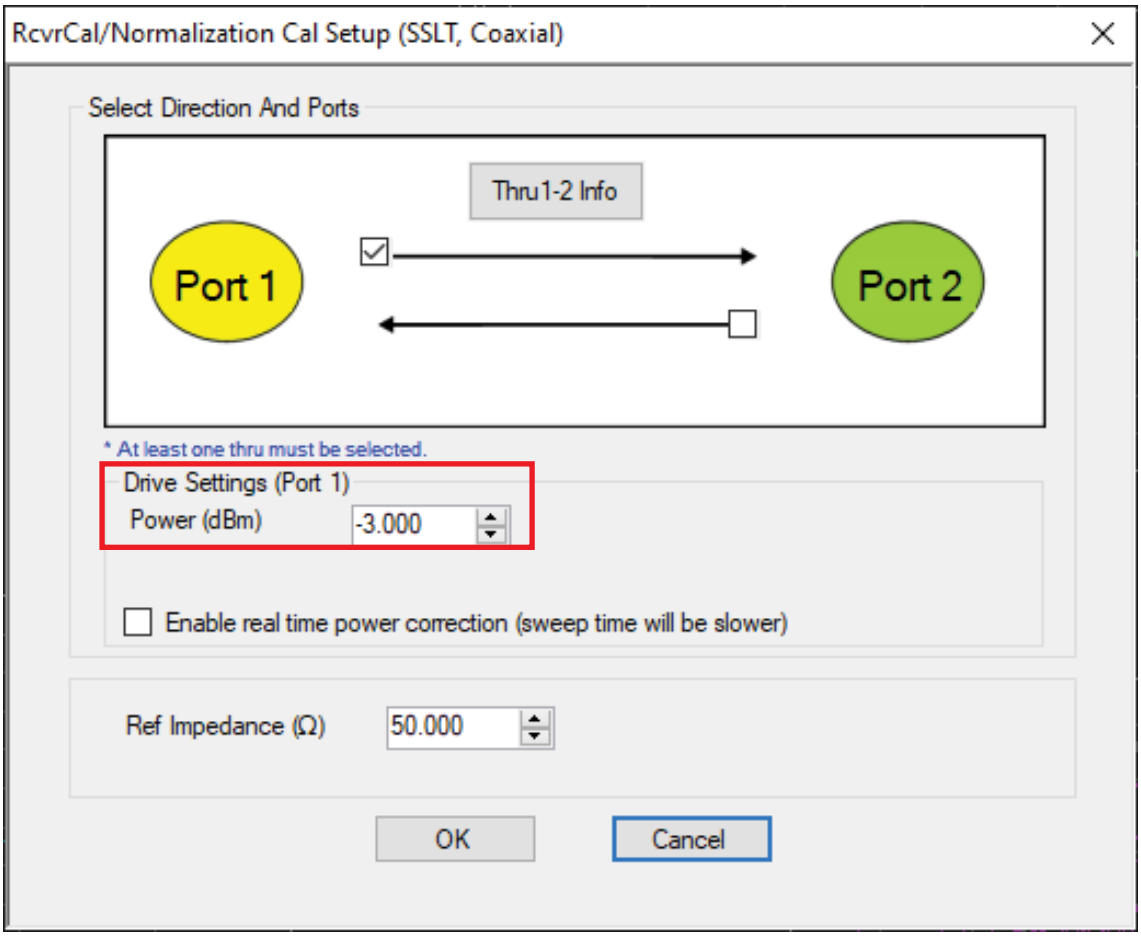


Figure 21-31. A MODIFY CAL dialog with the power-awareness fields highlighted.

Importantly, when the power is now changed after calibration, this system will note the difference relative to the normalization-time power and correct the conversion value by that amount. The correction is only as good as the power calibration (hence a user power calibration can help) since it is simply done on a dB-by-dB basis from the power menu entries.

This correction can be improved upon using the 'real-time power correction' selection. With this mode selected, an extra measurement of the reference receiver power is done at calibration time to establish a baseline. Then, when the calibration is applied, an extra sweep is done on every cycle to measure the reference receiver at the driving frequencies. The difference between the reference receiver measurements, on a point-by-point basis, is then applied to the final result. If the drive power has changed, the reference signal will also change so the compensation should be reasonably accurate. At extreme levels, there may be differences in receiver linearity that could alter the result and if the DUT match is wildly non-linear, there may be some modest coupling changes that could also affect the correction (and would affect the measurement in other ways as well). Also, at extreme levels where the leveling loop goes unlocked, the deviations may become large.

An extra sweep is required on every measurement cycle so there is some measurement time effect to this approach.

Examples

A simple downconverter was measured using both approaches. The input frequency range was 92.5 to 100 GHz and a fixed LO of 102.5 GHz was employed (generated externally so NONE was selected as the LO on the Active Mixer setup interface). The IF thus sweeps backwards from 10 GHz to 2.5 GHz. An ME7838A4X system was employed for the broadband coverage, but this measurement would easily be done with any of the broadband 2 or 4 port systems.

The rcvr/norm calibration will be performed in both cases with the power menu setting at -10 dBm (both frequency ranges) and the receiver calibration settings in the MODIFY CAL dialog will also be set for -10 dBm (no attenuator in play).

For real-time power correction OFF, the results are shown in Figure 21-32,when the measurement was done at -10 dBm (brown curve) and when the power menu was then changed to -20 dBm (orange curve). One can see a midband deviation where the ALC calibration linearity was slightly degraded. This is a mm-wave frequency converting measurement, so RF leveling is used, which has less range and linearity than does IF leveling (see Chapter 17, “Broadband/mmWave Measurements (Option 7, Option 8x)” for more information).

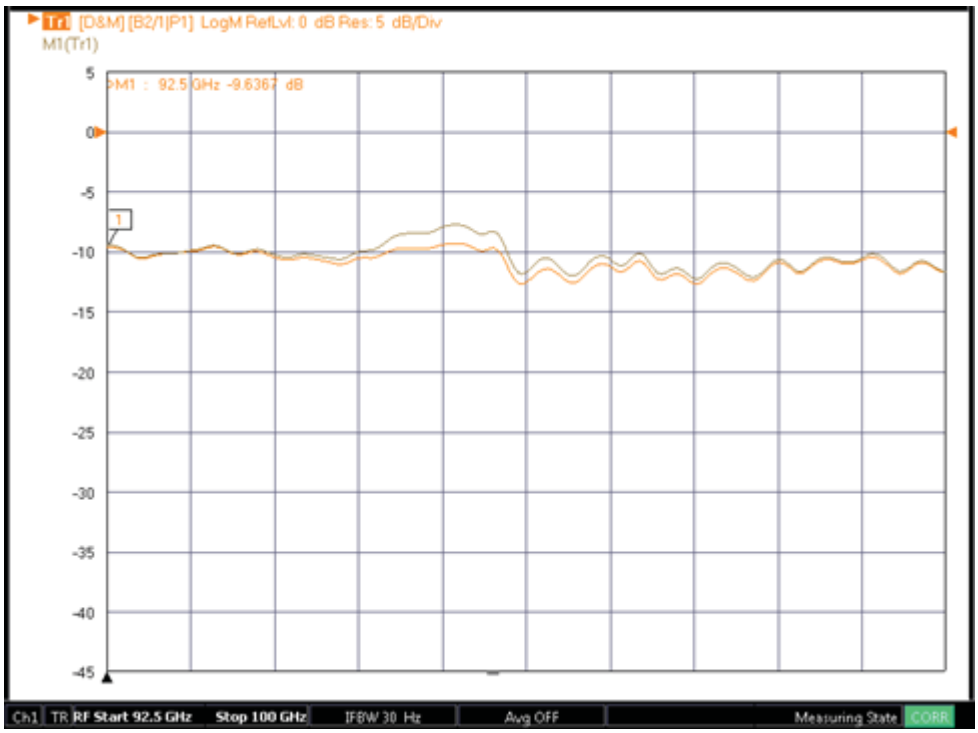


Figure 21-32. Drive Power Changed After Calibration – Real-time Power Correction Off

The experiment was then repeated with real-time power correction ON (see Figure 21-33). This time power levels of -10 dBm (brown) and -30 dBm (orange) were used to better illustrate the benefits of direct measurement of drive power. The deviation in this case was < 0.2 dB. Note that the curves would only be expected to overlay if the DUT linearity was absolutely perfect. In this range, this particular DUT was expected to be relatively linear so only minor deviations were anticipated.

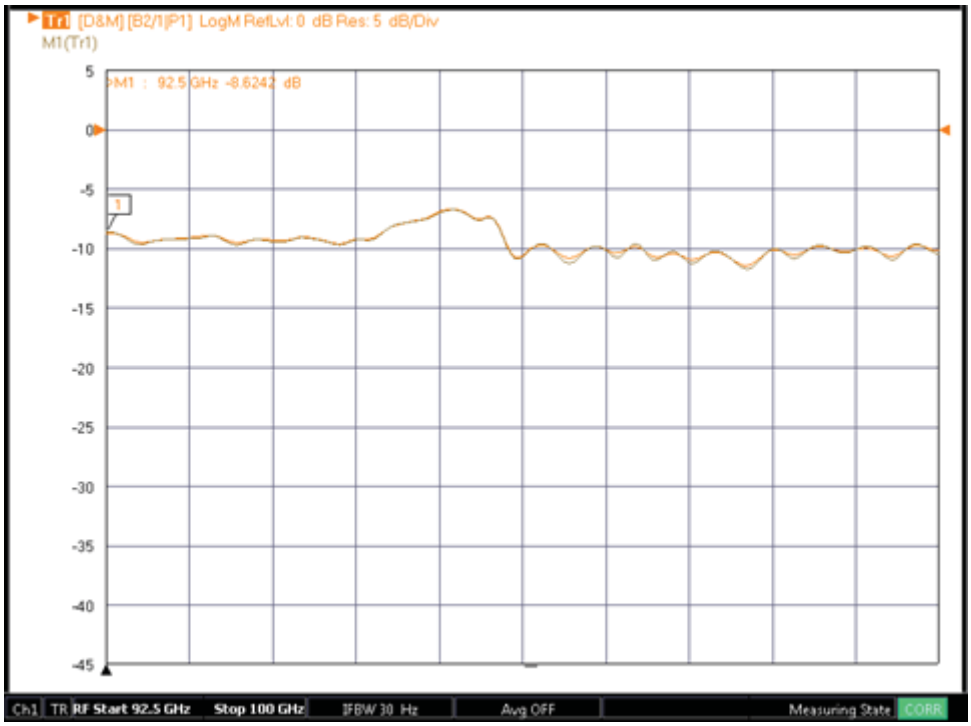


Figure 21-33. Drive Power Changed After Calibration – Real-time Power Correction On

21-14 Tips, Techniques, and Best Practices

This section contains some general hints on successful mixer measurements along with some tips on how to conduct ancillary measurements like coarse spur analysis. The topics are:

- "General Hints and Tips"
- "Dealing with Uncertainty Terms"

General Hints and Tips

- For scalar-like measurements (normalization-based conversion loss measurements for example), mismatch can be critical. If the DUT is poorly matched, consider padding the test ports before calibration.
- Even for enhanced match calibration situations, the match seen by higher order products can matter if the DUT is non-linear enough.
- Know what the compression level of the DUT is. The default power levels of the VNA (+5 dBm for some models) can compress many simple mixers and particularly high gain converters.
- Understand the frequency plan of the DUT, particularly the relationship of the input frequency to the LO in downconversion cases.
- Knowing all of the required measurements in advance can help speed up the setup when using the multi-channel wizard

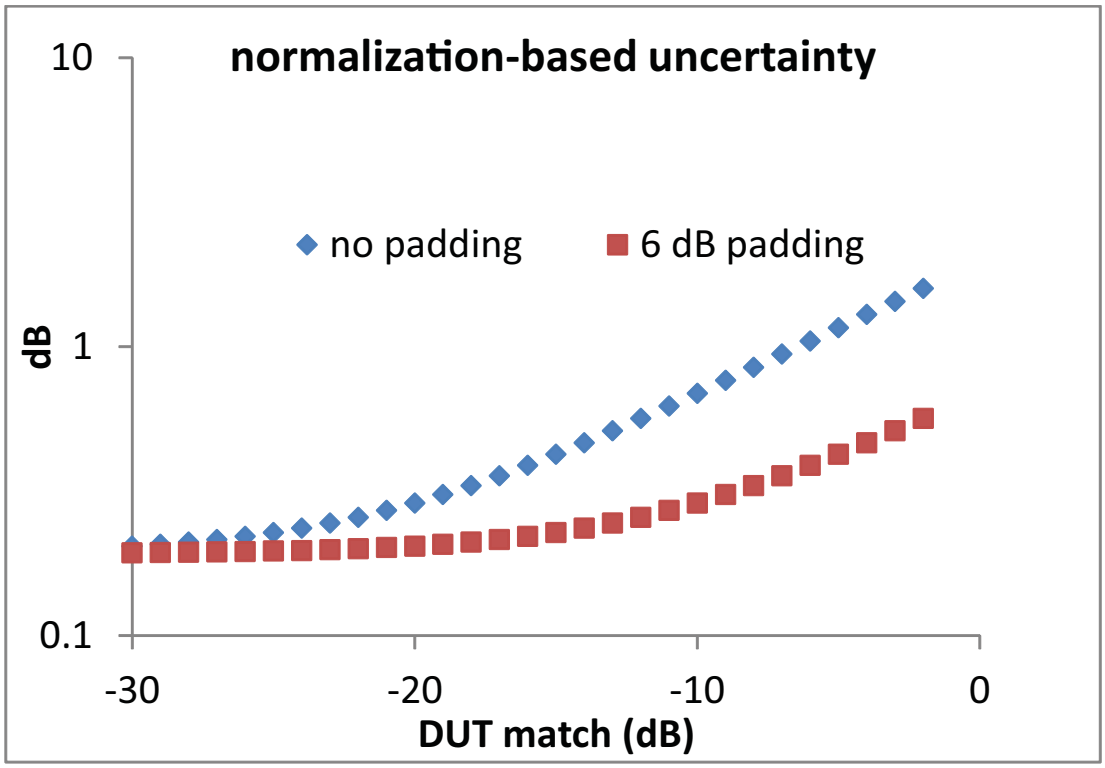
- For complex conversion scenarios (harmonic or subharmonic conversion, multiple LOs and multiple stages, etc.), the multiple source control system see Chapter 18, "Multiple Source Control (Option 7)") gives very flexible control of the internal sources and up to 4 external synthesizers.
- Know the required LO power. Often one might assume a required power based on the mixer architecture (e.g., +7 dBm for a double balanced Si diode mixer), but with so many technologies and topologies available, these assumptions may be less valid than in the past. In certain high gain DUTs, LO leakage out the port can be large enough to induce compression in the instrument as well (filtering may help in this extreme case).
- As with S-parameter measurements, consider the calibration method to be used based on the available calibration kits/standards and their respective uncertainties.
- The receiver/norm and enhanced match mixer calibration measurements help to get conversion gain, but this is all based on an unratioed parameter measurement (e.g., b2/1) and is hence affected by drive power. In earlier software versions (V2021.6.1 and before), this relationship was absolute: if the drive power changed, the displayed conversion gain will be in error by the amount of that power change. Later versions correct for input power shifts, at least so far as the instrument is aware of them, so caution is warranted in complex setups where the converter drive power is not directly related to an internal source setting.

Dealing with Uncertainty Terms

On the subject of uncertainties, the one new measurement (conversion gain/loss) warrants some additional commentary. For a normalization-based calibration, there are several uncertainty terms

- Port power accuracy (1 to 2 dB without a power calibration, 0.1 to 0.2 dB typically with a power calibration)
- Receiver calibration accuracy (typically <0.5 dB based on port mismatch)
- Linearity accuracy (<0.1 dB assuming away from compression)
- Mismatch interaction with DUT

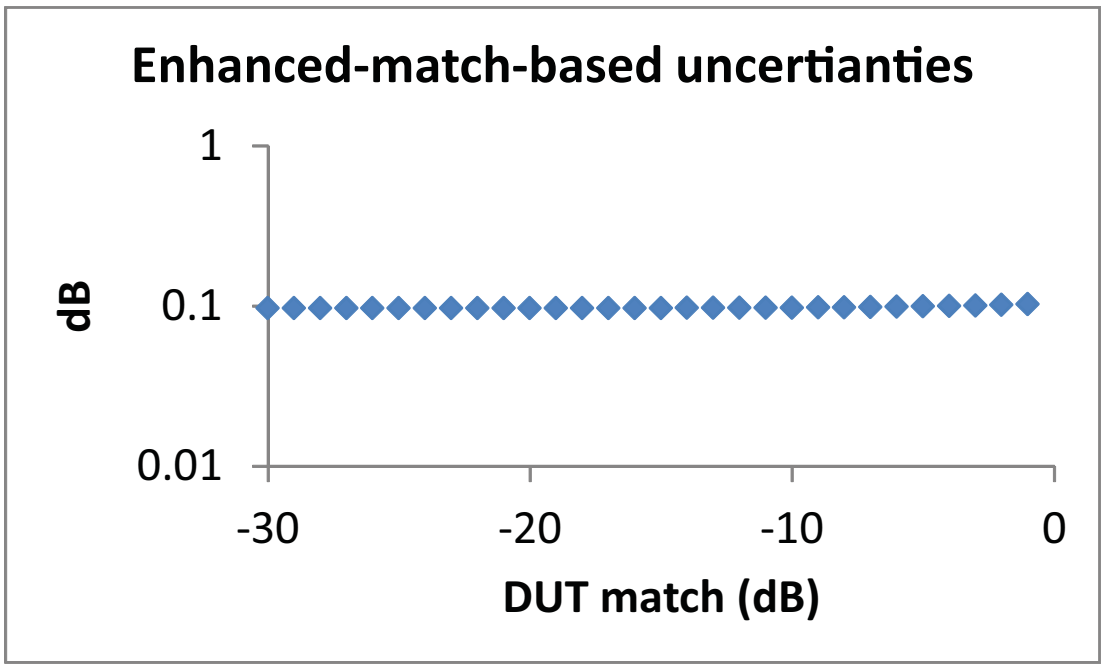
Assuming nearly optimal typical numbers for the other terms, an estimate of total uncertainty on conversion loss vs. DUT match (assumed symmetric and unilateral) is plotted in Figure 21-34 with and without some modest (6 dB) port padding. As might be expected, the worse the DUT match, the greater the value of the port padding.



An estimate of normalization-based conversion gain/loss uncertainty versus DUT match is plotted here

Figure 21-34. Normalization-Based Uncertainty

With the enhanced match calibration, the receiver calibration and DUT interaction uncertainties can reduce dramatically. The residual errors are based on a common SOLT calibration and the same unilateral assumptions as before are employed to simplify the calculation. The results are shown in Figure 21-35. The scales are changed and no padding was assumed. The result is largely match-invariant assuming the surrounding calibration was adequate.



An estimate of enhanced-match based uncertainties vs. DUT match are plotted here under the assumptions stated in the text.

Figure 21-35. Enhanced-Match Based Uncertainties

21-15 Final Topics

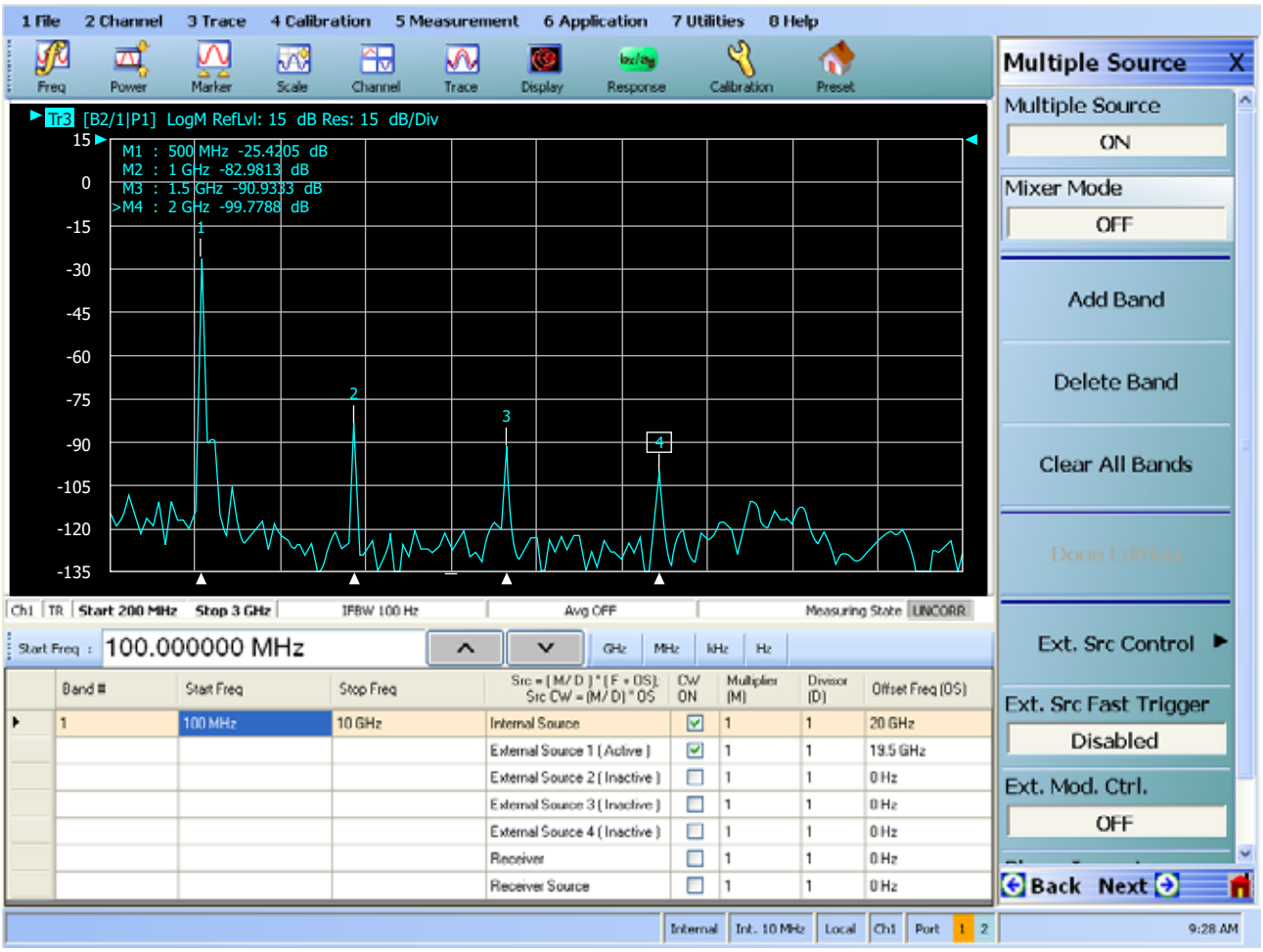
Final topics for this chapter cover some ancillary measurements, including:

- "Spurs"
- "IMD"
- "mmWave Mixer Measurements"

Spurs

The MS464xB can be used as a spectrum monitor to scan for spurious products of the mixer. This is most easily done with multiple source control (see Chapter 18, "Multiple Source Control (Option 7)") with the input source and LO usually set to some CW frequencies and the receiver allowed to sweep over a range of interest to look for spurious products on the LO. The unratioed test channel is often used as a monitor and a receiver calibration can be applied to get an absolute signal level for those spurious products. An example setup is shown in Figure 21-36 where the source is parked at 20 GHz and LO at 19.5GHz. The desired IF is 0.5 GHz but we can sweep over a wider range to look for products and the result is shown in the figure as well.

In this particular case (and with the sweep resolution used here), the dominant spurs are all harmonics of the IF. Using the previous nomenclature, they are of the form `N × (Input – LO)` and they are all at least 55 dB below the desired IF. With a different sweep range and point density, other spurs might have been found so it does help to work out the frequency plan in advance where likely spurs might reside.



A multiple source setup to look for spurious products on an example DUT is shown here. The plot shows the products measured,

Figure 21-36. Multiple Source Setup—Search for Spurious Products

One does have to exercise some caution in interpreting these results since the VNA receiver is not pre-selected and there may be internally generated spurs as well. In addition, the internal converters have an image located (usually) 24.7 MHz above the programmed frequency (at certain frequencies below 50 MHz, the offset is slightly larger). It usually pays to back calculate the product orders to see if they make sense for the DUT's frequency plan.

IMD

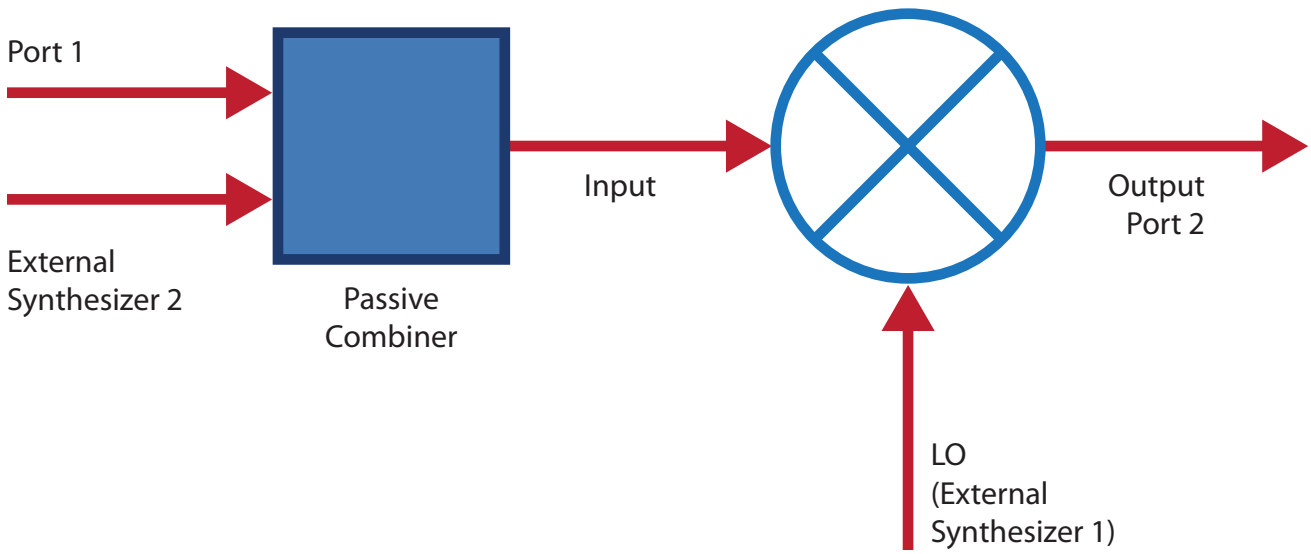
Using multiple source control again and two external synthesizers, it is quite easy to configure an IMD measurement. The measurement concept is covered in more detail elsewhere (the more automated version using IMDView is covered in Chapter 28, "IMDView™ (Option 44)"), but the basic idea is to apply two closely spaced sinusoids to the input of the DUT (at frequency f1 and f2) and examine the third order products at the output as shown in Eq. 21-3 below:

$$\left|\mathrm{LO} \pm (2 \times f_2 - f_1)\right|$$
$$\left|\mathrm{LO} \pm (2 \times f_1 - f_2)\right|$$

Equation 21-3.

The setup is shown in Figure 21-37 and the multiple source configuration is shown in Figure 21-38, "Example of Multiple Source Setup For CW Mixer IMD Measurement" on page 21-48 below.

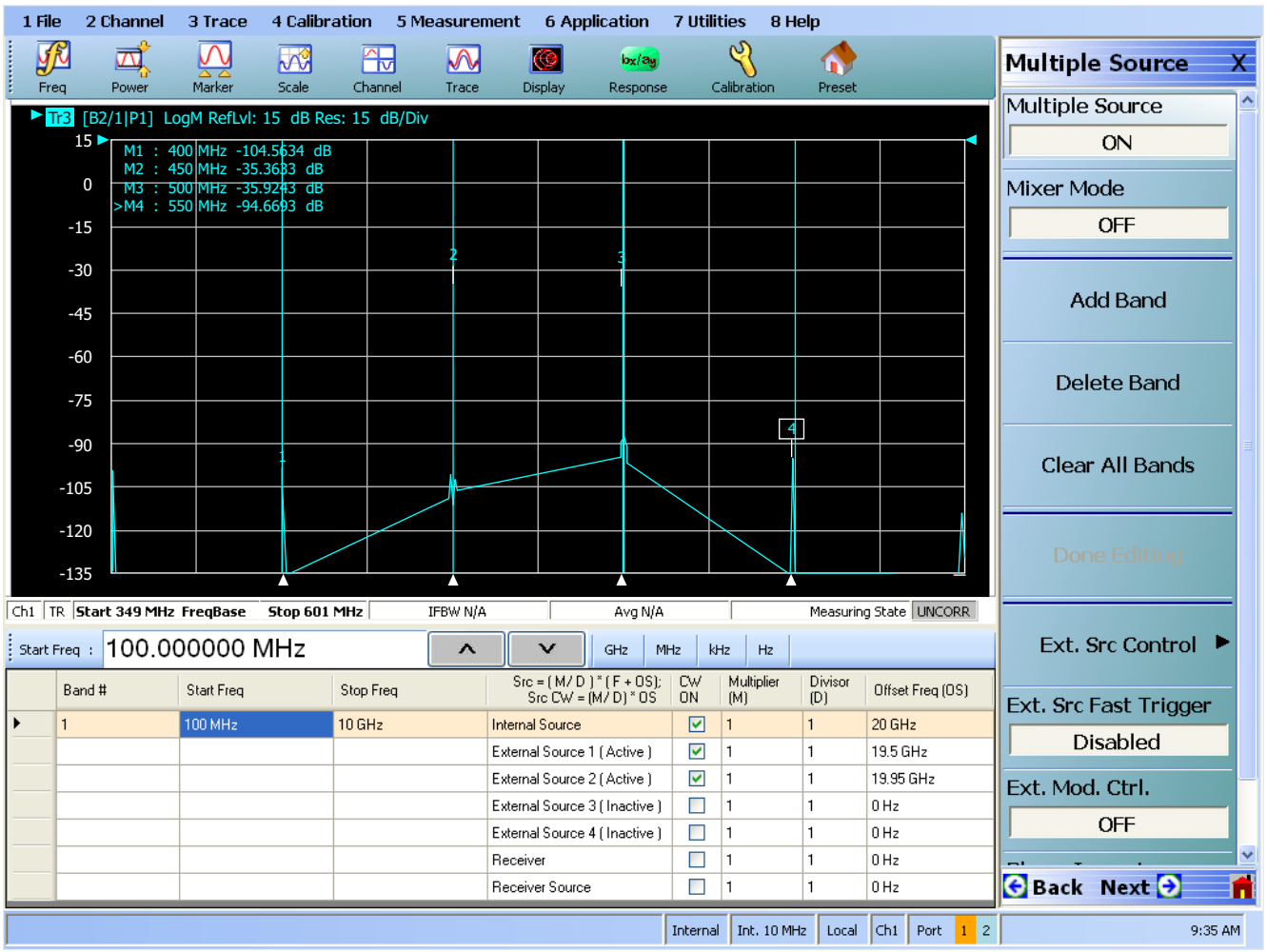
Typically b2/1 is the response variable and the measurement can be constructed to be CW (to look like a spectrum analyzer output) or swept to reveal the IMD product magnitude vs. frequency (in either dBc or dBm terms). The CW-measurement is illustrated in Figure 21-37 below where the sources are kept fixed and the receiver is allowed to move over the range of interest.



A simplified setup for mixer IMD is shown here.

Figure 21-37. Mixer Setup for IMD Measurement

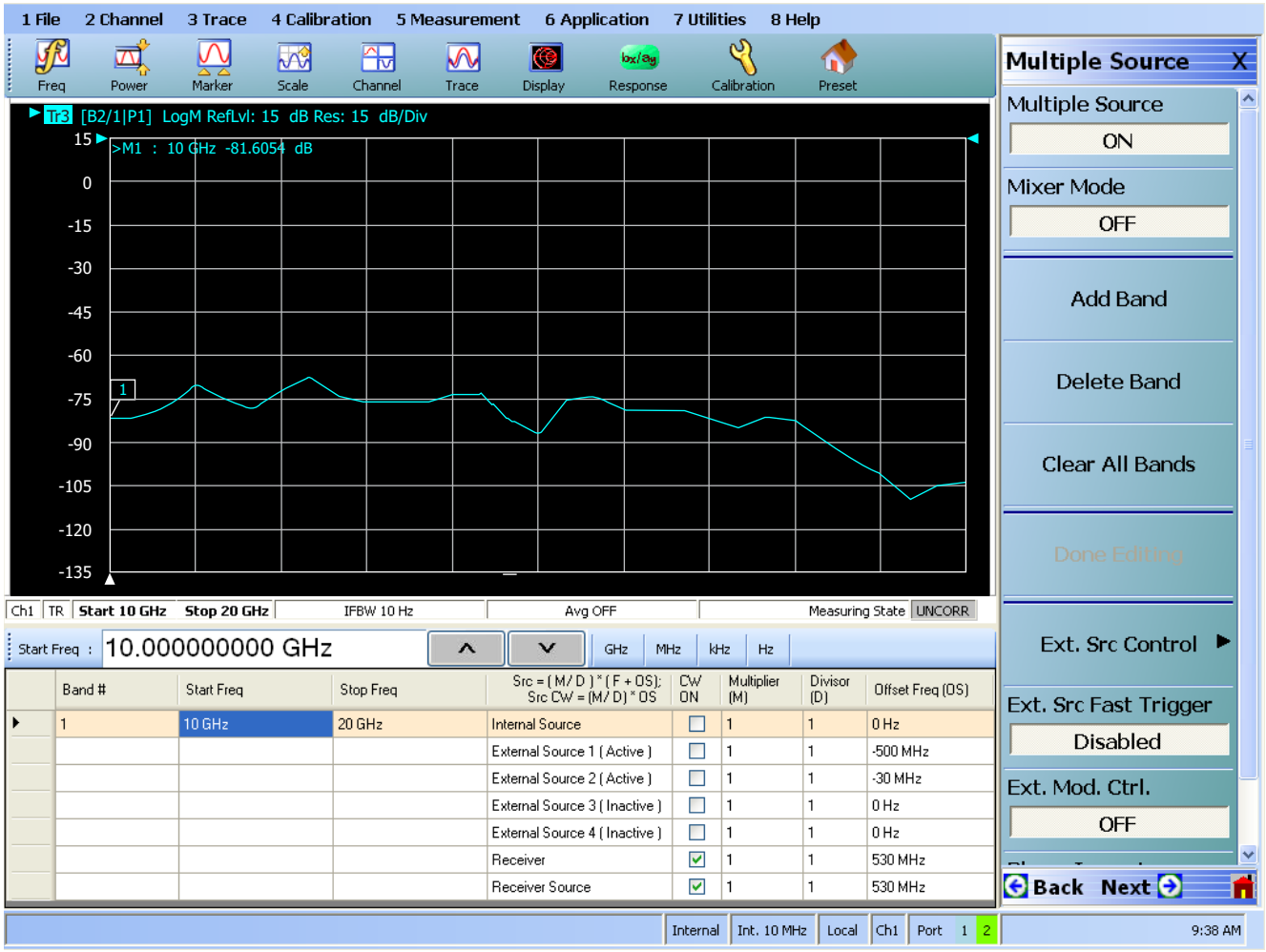
A segmented sweep setup was used in Figure 21-38 with 11 points (spaced by 200 kHz in this case) around the main tones (a 50 MHz tone offset was chosen), the 3rd order products and the 5th order products. This can optimize measurement time by not measuring frequencies that are not of interest and it allows one to measure quickly when at the main tones (large signals) and in a narrower IFBW when measuring the products (since they may be close to the noise floor). External synthesizer 2 was used to create the 2nd tone and its amplitude was adjusted to make the main tone levels approximately equal although other values could be chosen. In this example, the upper and lower products are not symmetric which may be an indication of bias network interaction within the DUT, compressive behavior in parts of the DUT, or other memory effects.



A multiple source setup for an example CW mixer IMD measurement is shown here.

Figure 21-38. Example of Multiple Source Setup For CW Mixer IMD Measurement

The swept measurement is shown in Figure 21-39 where the LO is fixed but the two source-like synthesizers are sweeping along with the receiver. In this case the tone offset was chosen to be 30 MHz. The only practical limit (outside of DUT bandwidth) is that very small offsets may incur a noise penalty and the synthesizer noise skirts start to interact. This normally does not happen until below 1 MHz offset depending on the synthesizers being used. The receiver is kept locked onto the upper IMD product in this case. Here just a receiver calibration was used so the readout is the product in absolute power terms (dBm). By doing a sweep with the receiver at a main tone (470 or 500 MHz in this case) and doing a trace(/)memory normalization, one could then measure the product in relative (dBc) terms.



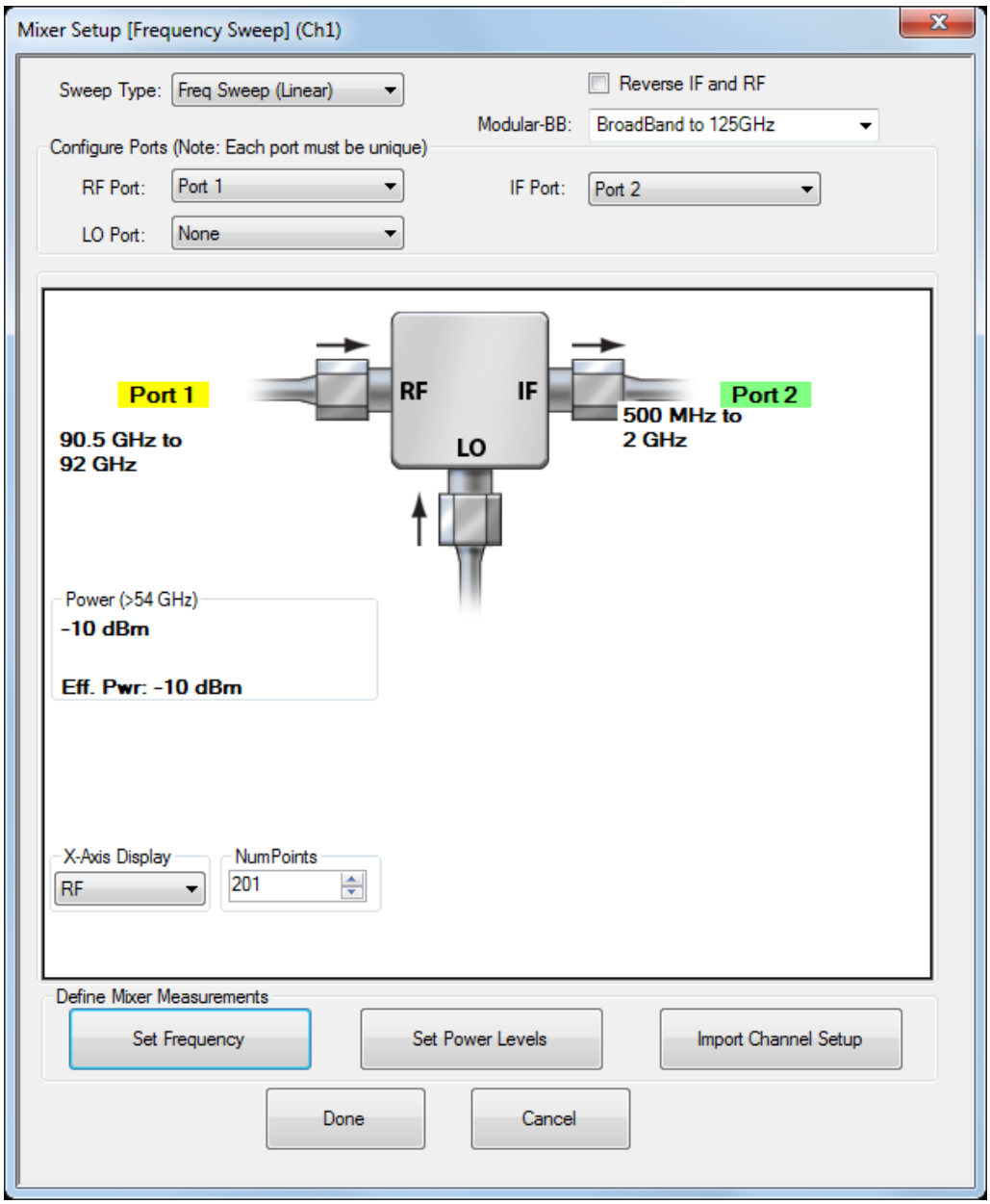
An example swept mixer IMD measurement is shown here.

Figure 21-39. Example of Swept Mixer IMD Measurement

mmWave Mixer Measurements

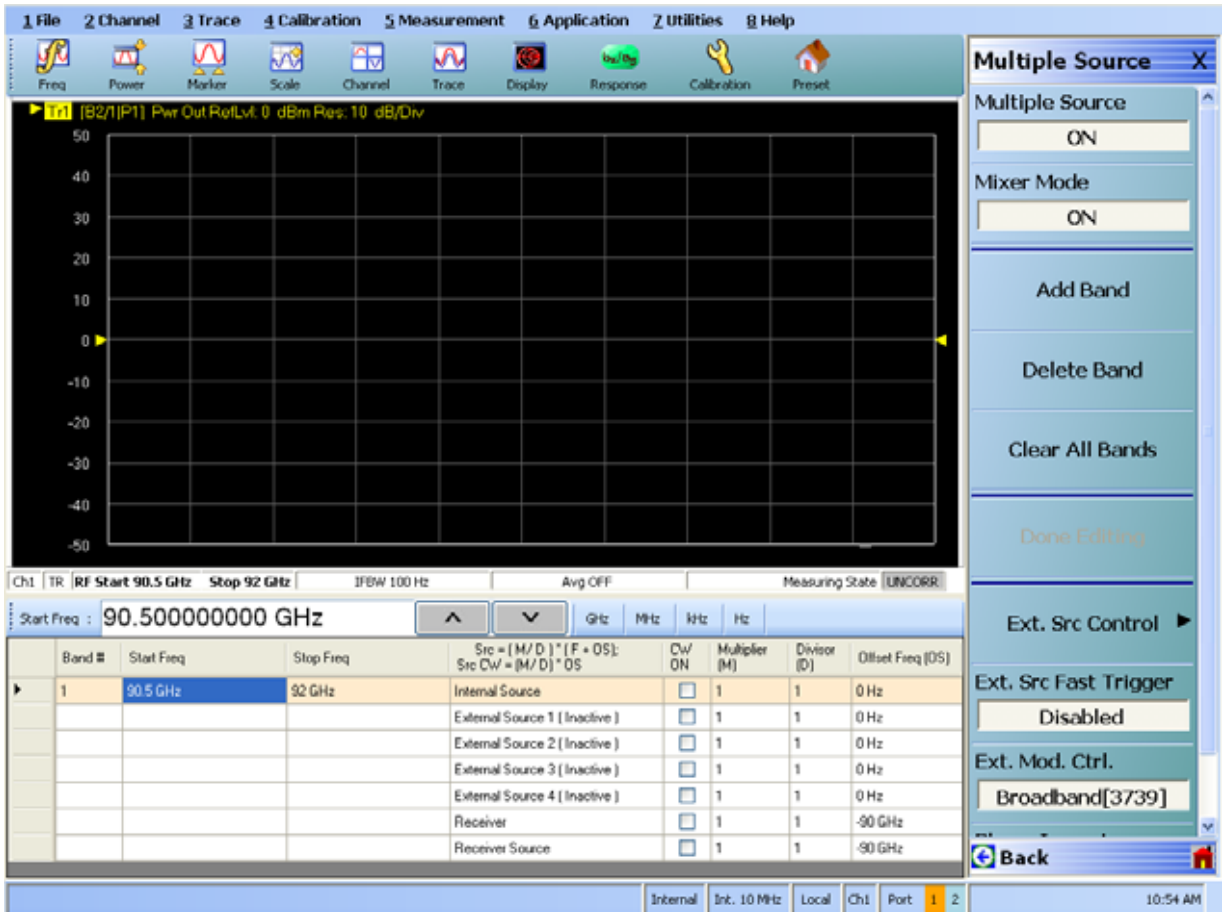
While the ‘modular BB’ check box in the mixer wizard was discussed earlier in this chapter, some additional information on setting up mixer measurements in the higher frequency bands may be helpful. As stated before, the mixer setup assistants are really configured for fundamental mixing while multiple source control is used directly for more complex cases. Many millimeter wave converting DUTs have more elaborate frequency plans, so by nature one will use the multiple source path more often, but we can discuss the use of the assistants.

The checkbox in the wizard/setup aid allow the frequency ranges to enter the millimeter wave band and will properly configure the system for the use of the mmWave modules. As an example, consider a W-band mixer with a fixed external LO at 90 GHz and the desire to run the RF over 90.5–92.0 GHz with a 0.5–2.0 GHz IF. The active channel mixer dialog is set up as shown in Figure 21-40, where no LO was selected since it is external and uncontrolled. One can then look at the multiple source setups (Figure 21-41). Most importantly, the external module control dialog has been activated.



A fixed external LO is used so ‘None’ can be entered here.

Figure 21-40. Active Channel Mixer Setup Dialog for a Millimeter Wave Example



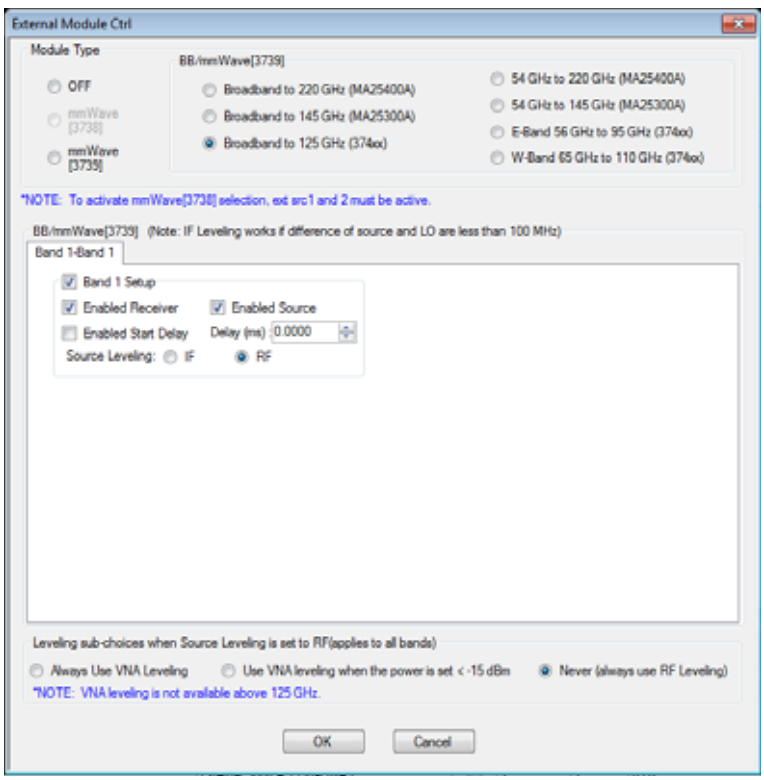
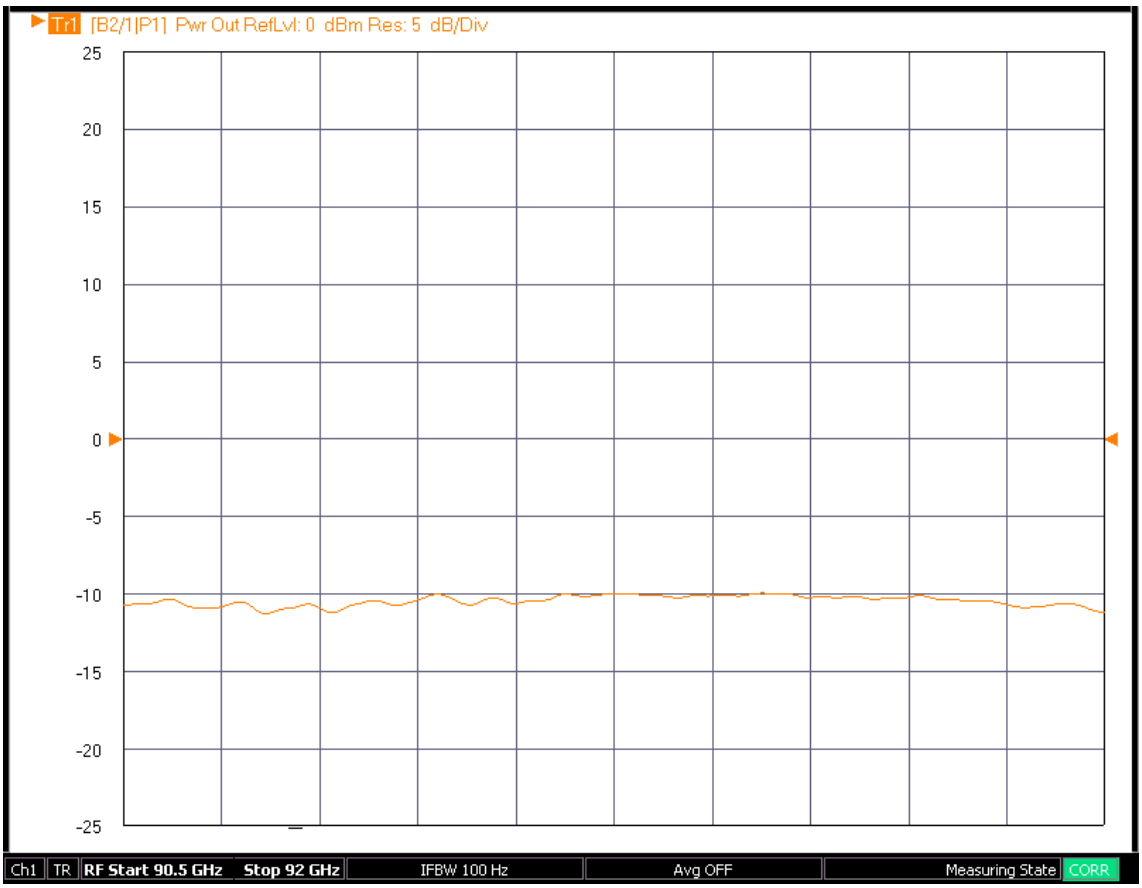


Figure 21-41. Resulting Multiple Source Screens for the mmWave Mixer Setup Example

The same calibration protocols as before are available for the modular mmWave modules. For this particular DUT, a normalization calibration was performed and the resulting conversion loss measurement is shown in Figure 21-42. Prior to entering the mixer mode, a user power calibration was performed on the mmWave module as discussed in Chapter 17. This accuracy transfer is important to the overall uncertainty.



A normalization calibration type was selected.

Figure 21-42. Example Millimeter Wave Conversion Loss Measurement

Even with dual sources, it is not possible to use both internal sources (i.e., both input and LO) for an active mixer broadband setup. It is possible to set this up manually with multiple source control. With a 4-port, dual source broadband/mmWave system, however, it is possible to set this up directly with both active channel and multi-channel tools. To keep the choices manageable, the input and output are restricted to be amongst ports 1 and 2 ('none' and external sources can be still be selected for the input) and the LO can be either port 3 or port 4 ('none' and external sources are still possibilities). An example setup is shown Figure 21-43.

For higher millimeter wave bands, multiple source alone is the setup choice because of the complex frequency plans that are involved.

Note	If the IF is < 30 GHz (on output) or < 54 GHz (on input), the mmWave module itself does not need to be connected as the base VNA will handle the tasks anyway.

Note	The LO power requirements of the DUT will have to be monitored. Below 54 GHz, the available power is < –10 dBm due to test set losses. An external amplifier may be required.

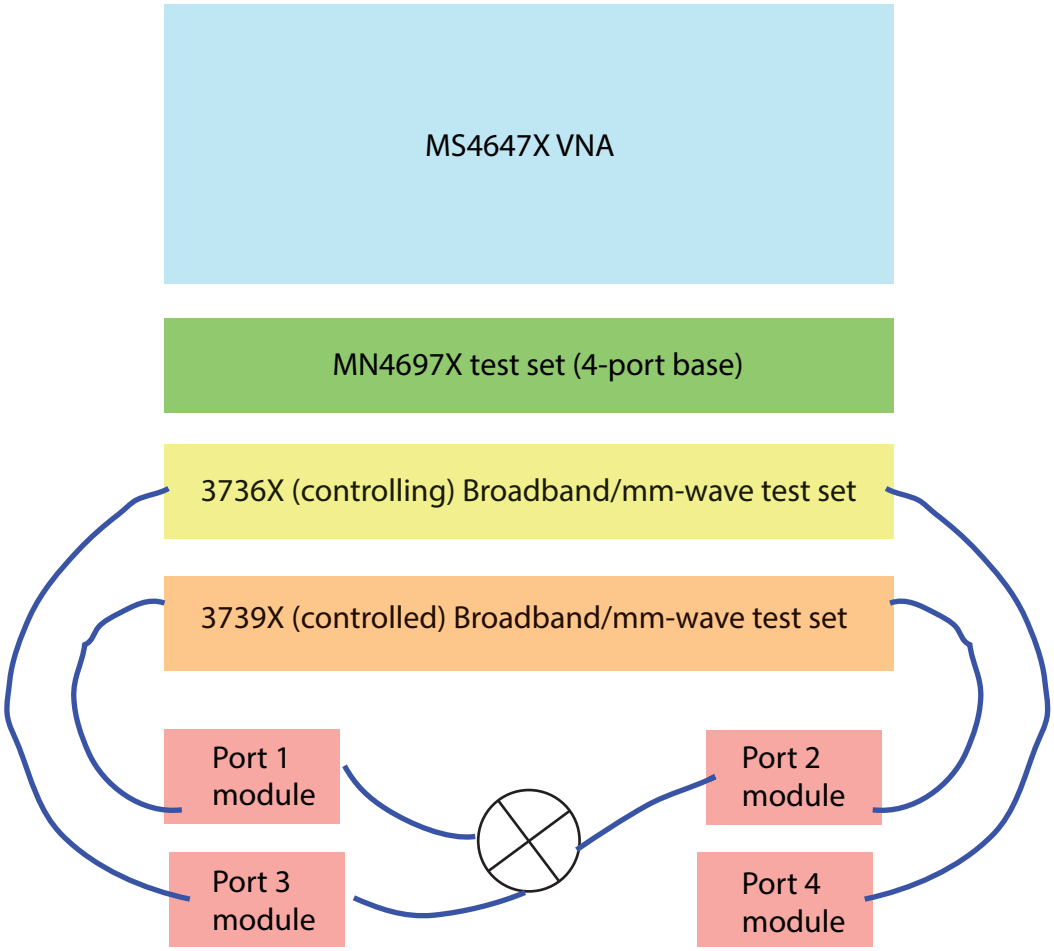


Figure 21-43. Example 4-Port Dual Source Setup

For mixer measurements in general, but particularly mmWave mixer measurements, power calibration is an important accuracy aspect as has been discussed. Since millimeter wave signal generation sometimes does not have as high a level of harmonic and sub-harmonic filtering, there can be differences between integrated power (that a power sensor measures) and fundamental power (that a VNA measures). As the harmonic content levels rise, this can lead to inaccuracies when basing a VNA measurement on a power meter calibration. For the modular broadband/mmWave case, the user power calibrations in VectorStar have an available fundamental power correction (checkbox available on the user power calibration dialog) that helps to reduce this problem by using measurements with the VNA receivers to estimate harmonic content instantaneously during the calibration and correcting the power targets accordingly. More information is available in "Fundamental Power Correction" located in Section 18-3 of this guide.

Harmonic converters can be handled within the mixer setup dialogs (at least for the more common cases) and this measurement type is quite common at mm-wave frequencies. As an example, a downconverter with a 90–105 GHz RF range was to be tested and its LO system operated on the second harmonic. This harmonic behavior could be selected on the set frequency sub-dialog as shown in Figure 21-44. Here the LO range was defined over the base range (this is where the VNA hardware must be programmed) and the diagram notes that a multiplication is being applied in DUT. The main benefit here is that 'auto' frequency plan calculations will take this into account, as would the frequency plan needs on the LO port for multi-channel setups.

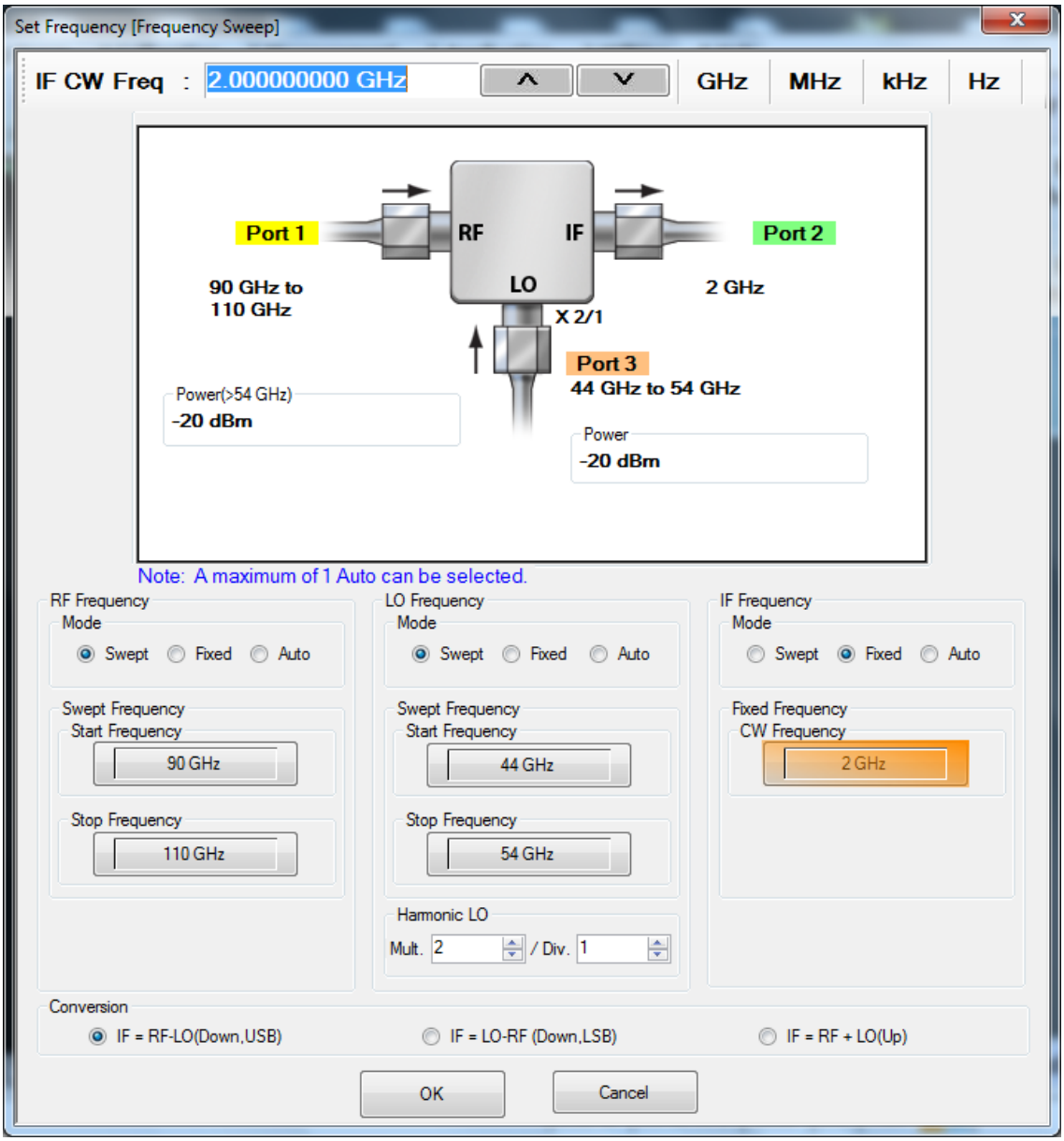


Figure 21-44. Frequency setup for DUT using 2nd harmonic LO conversion on a 4-port system

Chapter 22 — NxN Application

22-1 Chapter Overview

The NxN technique is an application for measuring non-separable devices. NxN is used when devices can only be measured in pairs and information about the individual devices is required. Examples of NxN include:

- Mixers (where one of each pair is an up converter and one is a down converter so that the VNA sees no frequency translation)
- Long cable runs (where two cables are needed to complete a round trip to the VNA)
- Fixture parts (where an individual part cannot be easily measured since one of its interfaces is non-coaxial/non-waveguide).

22-2 Introduction

This section provides a brief overview of the NxN method and how the application within the MS464xB Series VNA is used. The idea is that there are three devices that should or must be measured in pairs, and to measure every pair-wise combination of them. This collection of measurements provides enough information to extract the transmission characteristics of each of the devices. The information on any one of the first three devices can then be used to de-embed from measurements when this reference device is paired with any number of other parts. The measurements pairings to get started are shown in Figure 22-1. Device 2 is required to be reciprocal.

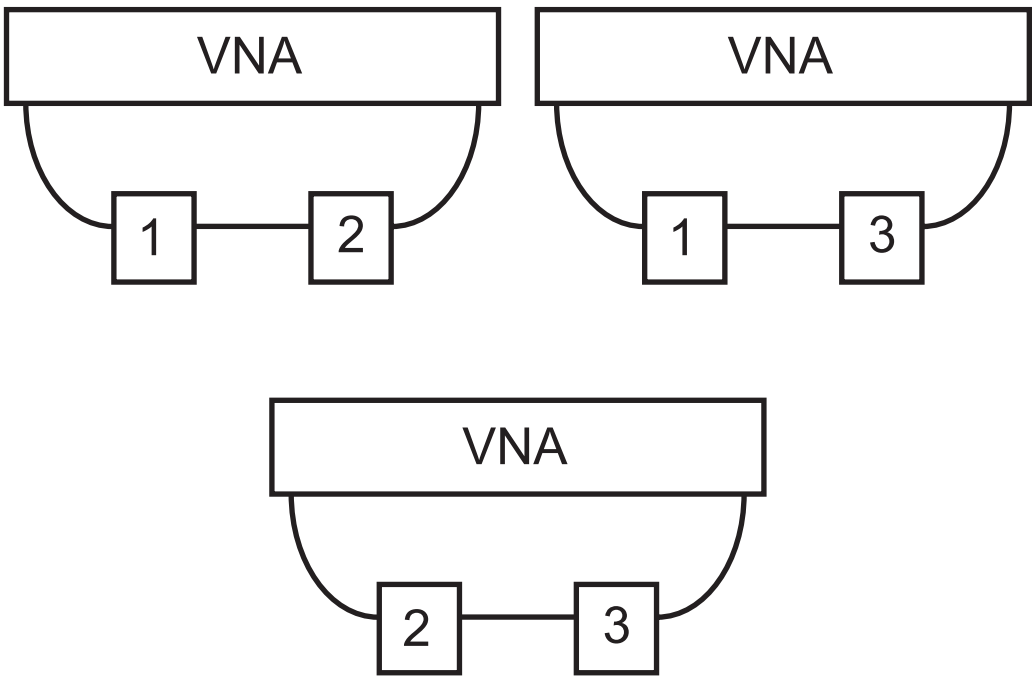


Figure 22-1. NxN Technique Measurements

If the devices are non-translating, few precautions are needed. If the loss of the devices gets high relative to the dynamic range of the VNA (in whatever state it is being used), then accuracy will be degraded. Normally a full 2-port calibration is performed on the VNA (so that outbound matches can be corrected) but simple normalization calibrations can be adequate. Note that the inner matches cannot be corrected (since no access is available) hence match is not an extracted parameter. Full s2p files will be generated for the devices but S_{11} and S_{22} will be set to zero.

If the devices are poorly matched, it may be necessary to place a pad or other network between the devices. As long as this inner network is present for all measurements and its S-parameters are known, its effects can be removed. This technique was originally used in mixer measurements, hence it is termed an IF network. This measurement setup is shown in Figure 22-2. The three measurements required to execute NxN are shown here when using an inner network N (which may be a pad or filter combination). Device 2 is required to be reciprocal.

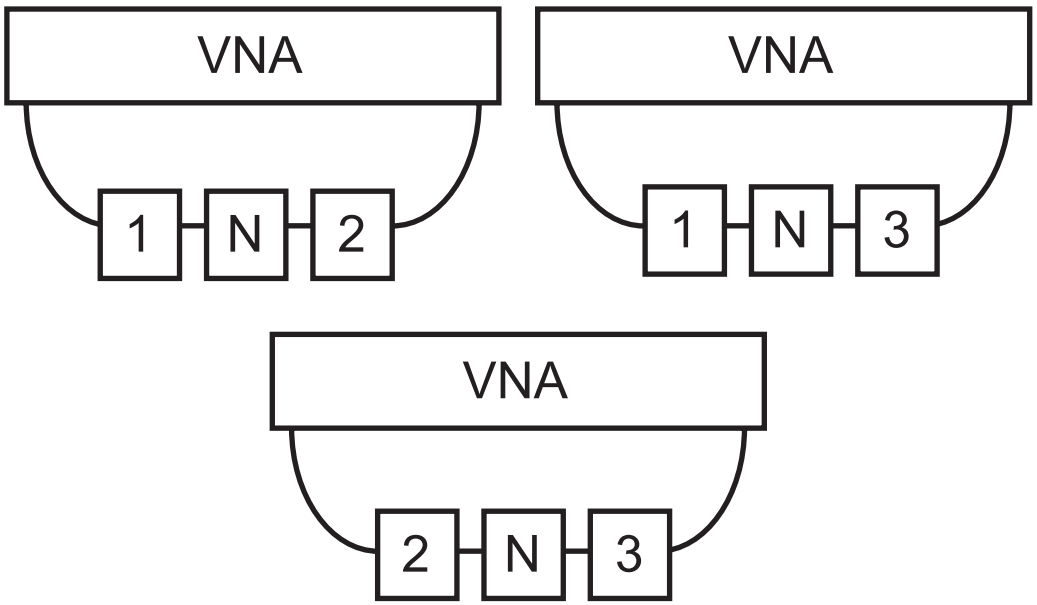


Figure 22-2. NxN Measurement Setup (Using an Inner Network N)

NxN is a popular mixer technique since the VNA sees the same input and output frequencies. Thus normal calibrations can be used, no frequency offsets are needed and there are no reference complications. The first device in each pair will be used as an up converter or down converter and the second device will be the opposite. Device 2 must be able to fulfill both roles so it must be a passive mixer. The same frequency plan must be used in all of the measurements.

Typically a common LO is used for all devices (with some splitter assembly) but multiple LOs can also be used as long as they share a common 10 MHz reference. These sources can often be controlled using multiple source control (see separate section in this Measurement Guide for more information). If one of the devices has multiple conversion stages, it may be desirable for the other devices to have the same conversion pattern for optimal trace noise. If this is not feasible, the only real requirement is that the input and output frequencies for all devices match.

The nature of the frequency plan between the devices must be known if the IF network (N in Figure 22-2) is to be removed. If the first device is not performing an inversion (i.e., if the input signal ramps up in frequency, the output signal ramps up in frequency) then the lookup of the IF network behavior is not inverted. If the first device does perform an inversion, then the lookup of the IF network must be inverted.

22-3 Using the NxN Application

The NxN application is started from the APPLICATION menu as shown in Figure 22-3.

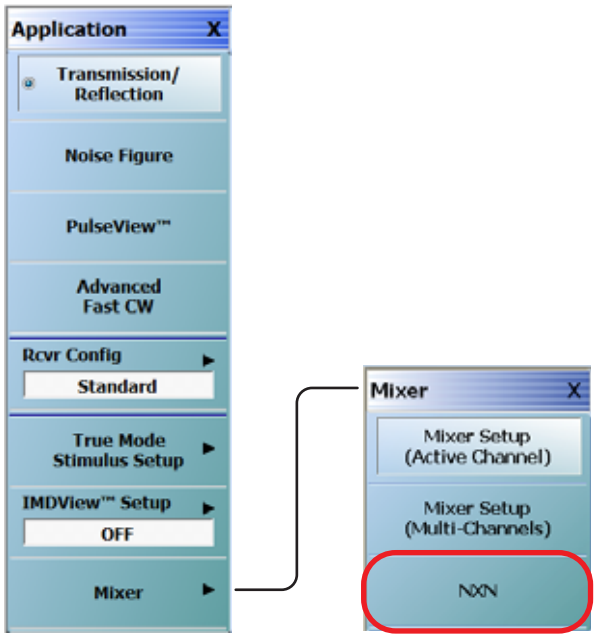
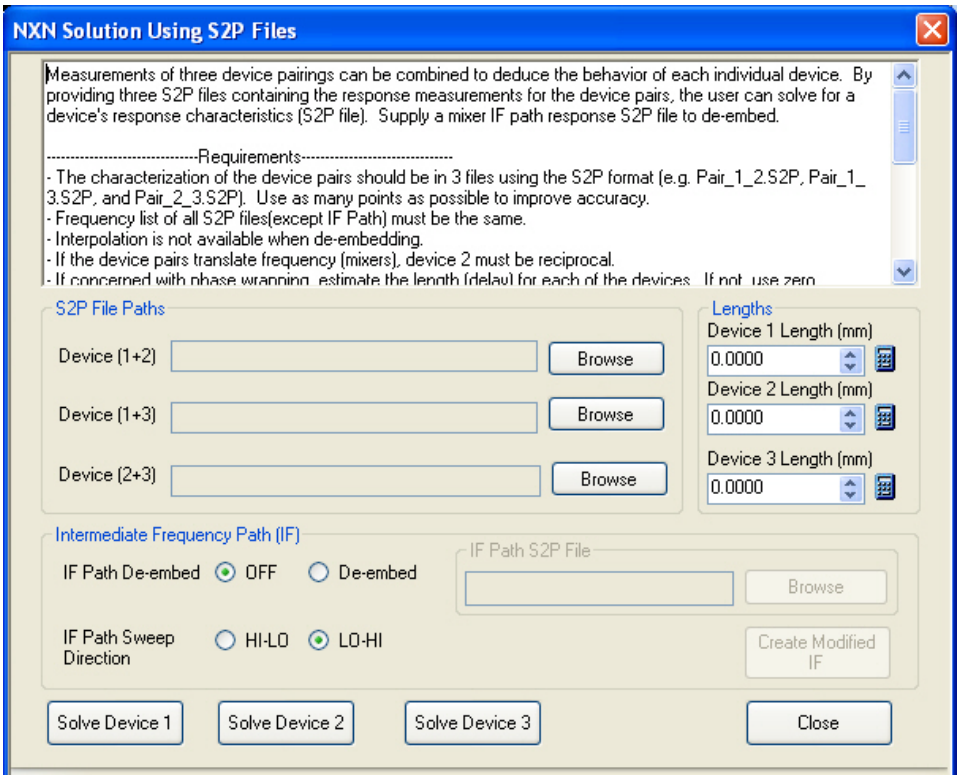


Figure 22-3. APPLICATION Menu—NXN Control Button

Selecting the NxN button will open the main NxN dialog as shown in Figure 22-4. The measurements shown in Figure 22-1 or Figure 22-2 must first be performed and those measurement results saved as S2P files. The match information in these files will be ignored, but the file format for a full 2-port measurement is used for convenience. These S2P files are entered in this dialog using the Browse buttons shown.



Main NXN Solution Using S2P Files Dialog

Figure 22-4. NXN SOLUTION USING S2P FILES Dialog Box

If the scheme of Figure 22-2 is used, then the inner or IF network must be specified. Select IF Path De-embed in this case and use the browse function to locate the S2P file for the IF network. Note that all of these S2P files must have the same number of points and the frequency range of the IF file must be appropriate for the IF sweep range between the devices. If the devices are non-converting, the frequency range of the IF file must match that of the other files.

If the devices are frequency converting, the sweep direction in the inner or IF zone must be specified. If the inner frequency sweeps up for a sweep up in the input frequency, then select LO-HI (non-inverting). Otherwise, select HI-LO.

The computation produces a phase ambiguity when it solves for the devices so some information about the electrical length of the devices should be entered. Boxes are provided for entering the approximate air-equivalent electrical length of each device. The calculator icon can be used if time delay estimates are known instead and a conversion calculator will appear. If the phase shift for the devices is not of interest, these lengths may be left at zero. If they are of interest, the estimate should be accurate to within a half wavelength at the lowest input frequency being used.

Once all of the information is entered, the Solve Device x buttons at the bottom of the dialog (x = 1, 2, or 3) will trigger the computation. Each invocation will bring up a file dialog to name the S2P file describing the device in question. If only the information for one device is needed, only that button need be selected.

If more devices are to be tested in the same configuration (e.g., one has a collection of mixers to measure of the same model), then these devices can be combined, one at a time, with a now known device (1, 2, or 3). Since the S2P file for the now known device has been generated, it can be de-embedded using the standard embedding/de-embedding utility described elsewhere in this measurement guide.

Chapter 23 — Miscellaneous Features

23-1 Chapter Overview

This chapter provides information on other features available on the VectorStar MS464xB Series VNAs including frequency blanking.

23-2 Frequency Blanking for Security

Frequency Blanking is typically used in secure environments to prevent frequency information from being displayed on the analyzer. When frequency blanking is activated, the display of all frequency information is changed from real value to "XXXX."

Procedure

1. In normal operation, the frequency information is displayed at the bottom of the display area, and frequency values are displayed for each active marker.
   - Figure 23-1 shows a typical display with frequency information displayed for all markers.
   - At the bottom of the display area, not shown here, additional start and stop frequency information is displayed.

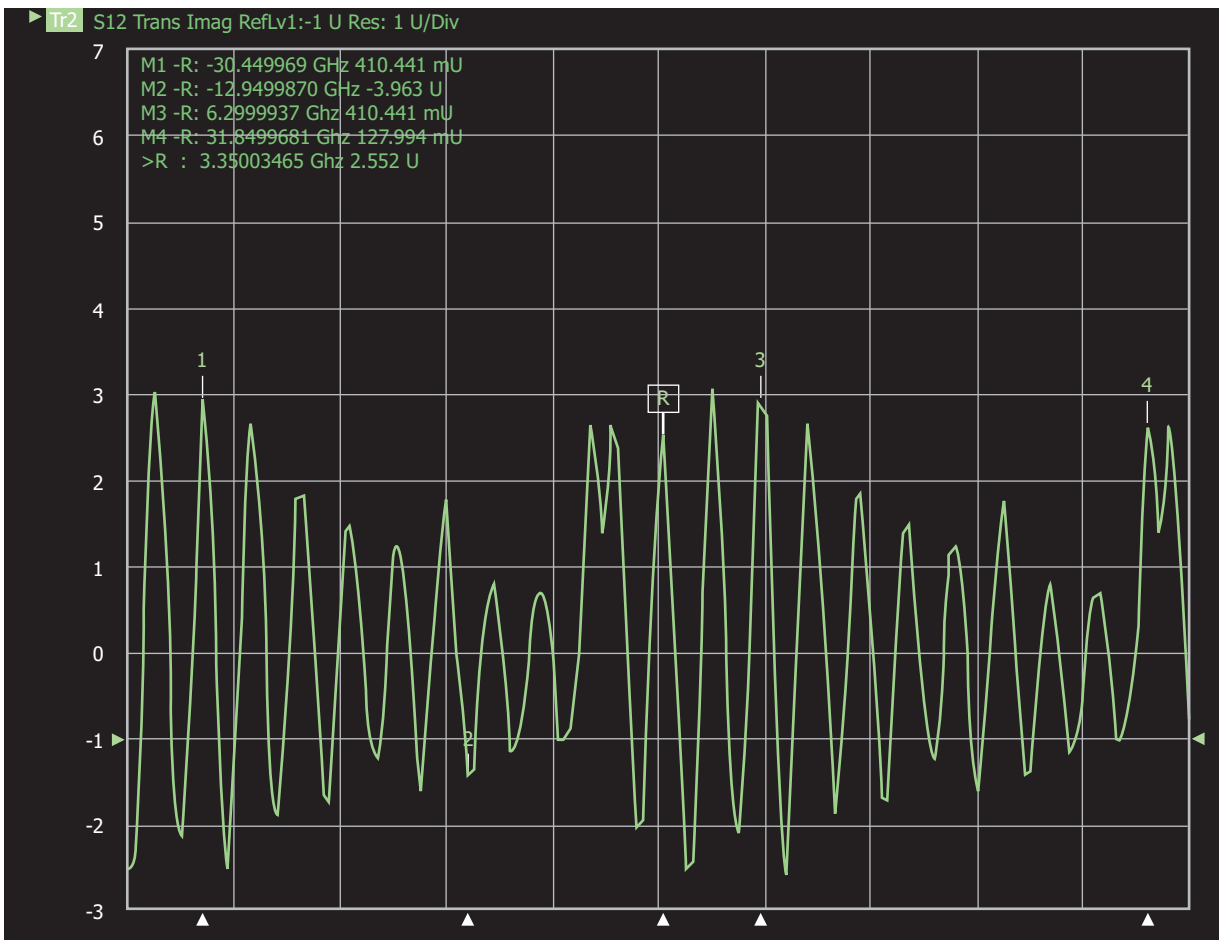


Figure 23-1. Typical Display Graph with Markers and Frequency Blanking Off)

2. A best practices recommendation is to save the current instrument setup configuration to a preset file.
   - On the top menu bar, select File | Save and assign a unique name for the CAL file.
   - Click Save.
3. To activate frequency blanking, navigate through this menu path:
   System | Utility | Frequency Blanking
4. Click the Frequency Blanking button to toggle the frequency display to Off.
   - All frequency information is replaced with "XXXX".
   - The Frequency Blanking button is now unavailable.

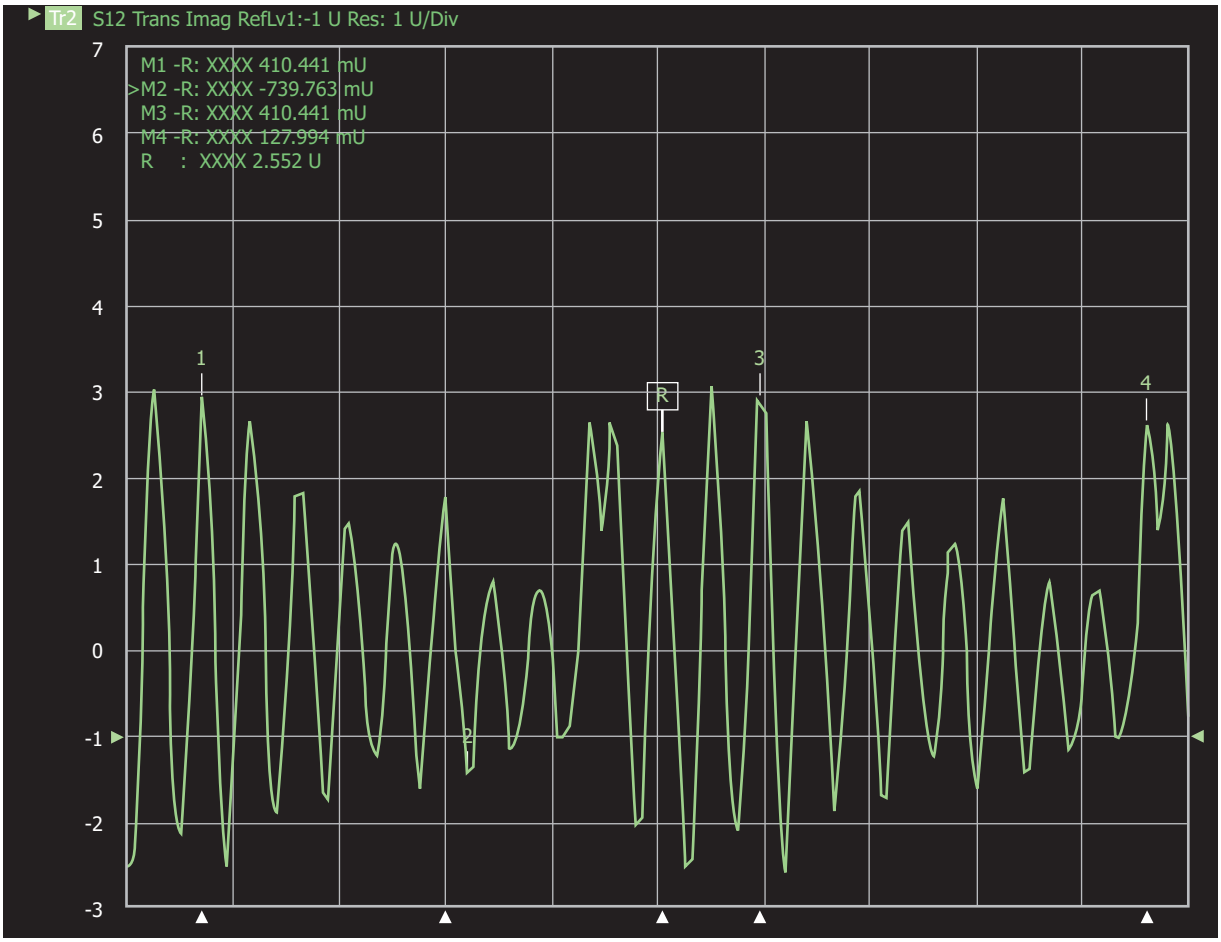


Figure 23-2. Typical Display Graph with Markers and Frequency Blanking On

5. To restore frequency information and set Frequency Blanking to Off, click the Preset icon on the icon toolbar.
   - The instrument returns to its last saved preset configuration.

Chapter 24 — Multiport Measurements

24-1 Introduction to Multiport

The MN4690B/C Series Test Set, in conjunction with a MS464xB Series VNA, allows multiport measurements in an integrated context. The test set consists of a set of test couplers along with appropriate switching to use the source and receivers of the base VNA. The 1- to 4-port measurements with this arrangement can satisfy many applications:

- Balanced/differential device measurements
- Balanced to unbalanced device measurements (including baluns and other hybrids)
- General 3- and 4-port microwave devices (such as splitters, hybrids, circulators, couplers, or distribution networks)
- Multiple measurements of one and 2-port devices for a fast, manufacturing test setup

The purpose of this chapter is to introduce the measurement setup (VNA + test set), some basic multiport S-parameter definitions, and how many of the measurement steps discussed earlier in this guide are generalized for the multiport scenario.

24-2 VectorStar VNA and Multiport Test Set

The VectorStar MN4694B/C Multiport Test Set provides coverage from 10 MHz (70 kHz if supported by the VNA) to 40 GHz and can be paired with the following VNAs equipped with K Connector Test Ports:

- MS4642B or MS4644B VNAs

The VectorStar MN4697B/C Multiport Test Set provides coverage from 10 MHz to 70 GHz and can be paired with the following VNAs equipped with V Connector Test Ports:

- MS4645B or MS4647B VNAs

Below, Figure 24-1 shows a VectorStar MN4697C 10 MHz (70 kHz if supported by the VNA) to 70 GHz Test Set paired with a MS4647B VNA.



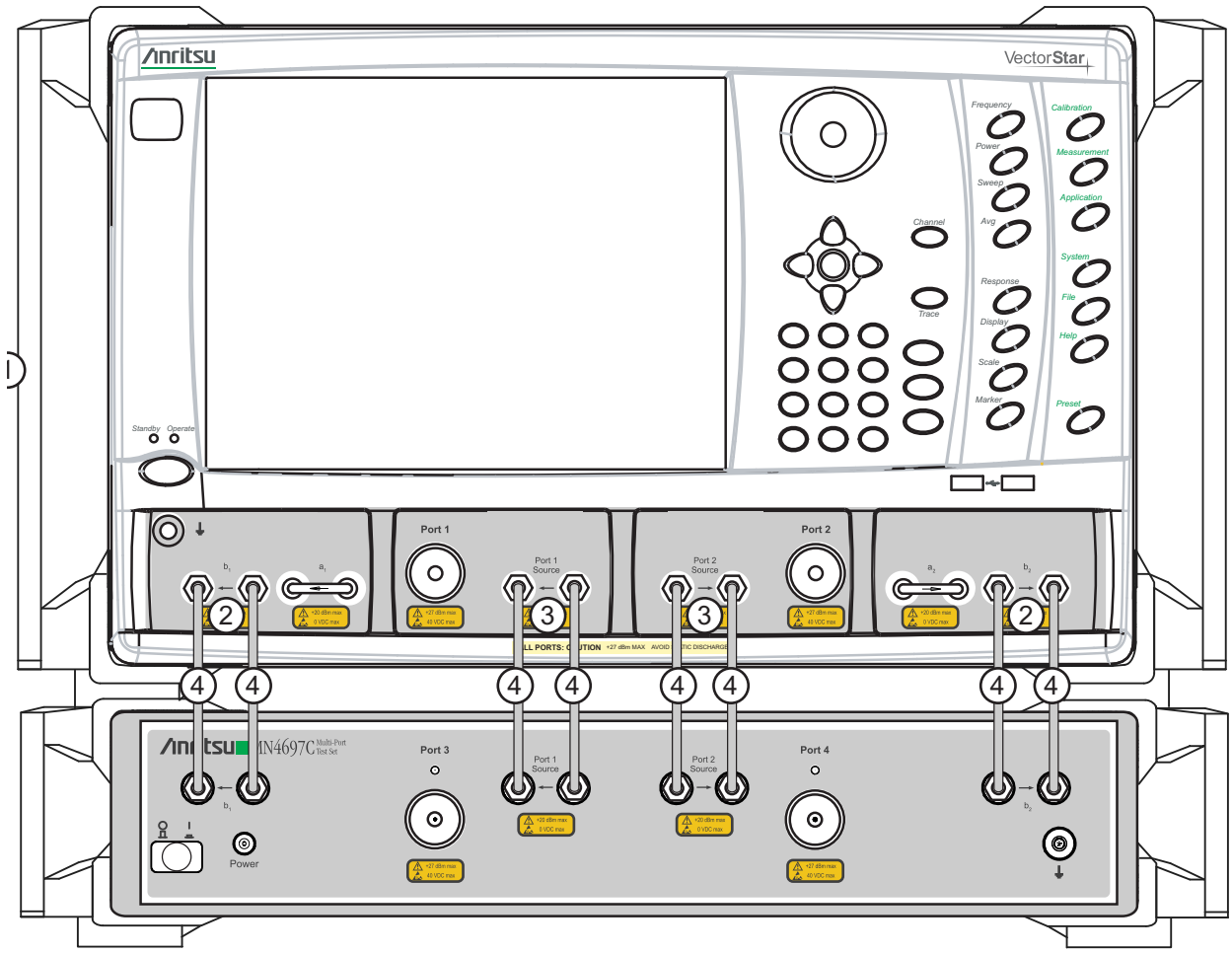
VectorStar MS4647B VNA, 10 MHz to 70 GHz, V Test Port Connectors (on top).	VectorStar MN4697C Multiport Test Set, 10 MHz to 70 GHz, V Test Port Connectors (on bottom).

Figure 24-1. VectorStar MN4697C Multiport Test Set and MS4647B Vector Network Analyzer

Cabling Connections

The front and rear panel setup configurations and system block diagrams are shown in the three figures below (Figure 24-2 and Figure 24-3). RF cables are required to link the source drive as well as the receivers. The receive path is split into two parts with some cabling on the rear panel and some on the front panel. Consult the following documents for detailed installation and cabling instructions:

- *VectorStar MN469xC Series Multiport VNA System Technical Data Sheet*—11410-00771
- *VectorStar MN469xC Series Multiport Test Set Installation Guide*—10410-00728
- *VectorStar MN469xC Series Multiport Test Set Quick Start Guide*—10410-00729.



1.Install the VectorStar VNA on top of MN4690C Series Multiport Test Set. 2. Remove VNA **b1** and **b2** loops. 3. Remove VNA **Port1 Source** and **Port 2 Source** loops	4. Install Eight (8) provided semirigid cables between the following ports: • VNA b1 input to Multiport b1 input • VNA b1 output to Multiport b1 output • VNA Port 1 Source input to Multiport Port 1 Source input • VNA Port 1 Source output to Multiport Port 1 Source output • VNA Port 2 Source output to Multiport Port 2 Source output • VNA Port 2 Source input to Multiport Port 2 Source input • VNA b2 output to Multiport b2 output • VNA b2 input to Multiport b2 input

Figure 24-2. MS464xB Series VNA and MN469xC Multiport Test Set Front Panel Interconnections

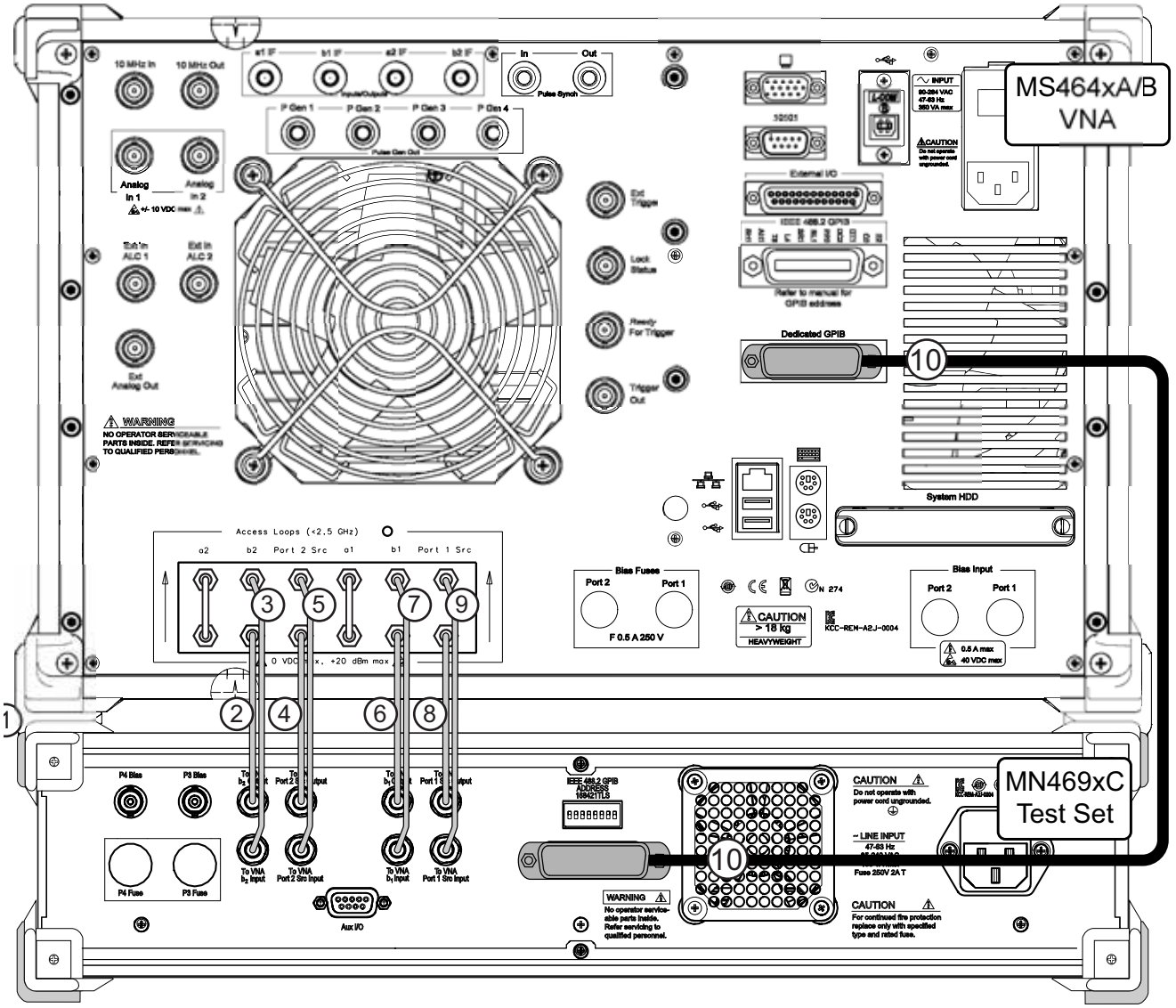


Figure 24-3. MS464xB Series VNA to MN469xC Multiport Test Set Rear Panel Connections

Table 24-1. MN469xC Multiport Semi-rigid Cable Interconnect Part Numbers and Locations (1 of 2)

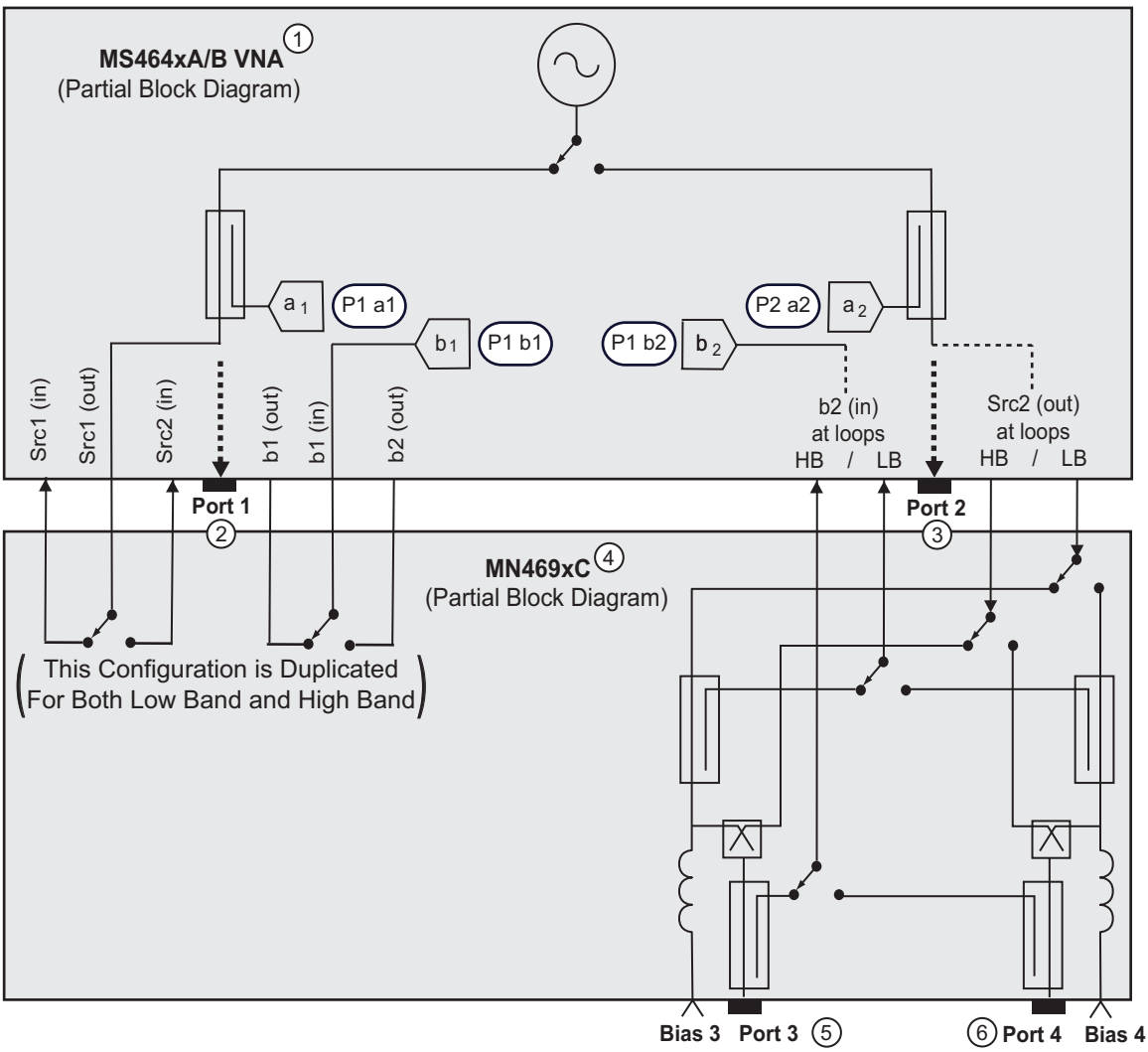
Index	Part Numbers	Description/Torque	Connection From	Connection To
1	MS469xB VNA Placed on top of MN469xC Multiport Test Set			
2	62112-141	SMA male-male semi-rigid Tighten all connections using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **b2 loop out**	MN469xC port labeled: **TO VNA b2 OUTPUT**
3	62112-140	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **b2 loop in**	MN469xC port labeled: **TO VNA b2 INPUT**

Table 24-1. MN469xC Multiport Semi-rigid Cable Interconnect Part Numbers and Locations (2 of 2)

Index	Part Numbers	Description/Torque	Connection From	Connection To
4	62112-141	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **P2 source loop out**	MN469xC port labeled: **TO VNA Port 2 Src OUTPUT**
5	62112-140	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **P2 source loop in**	MN469xC port labeled: **TO VNA Port 2 Src INPUT**
6	62112-141	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **b1 loop out**	MN469xC port labeled: **TO VNA b1 OUTPUT**
7	62112-140	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **b1 loop in**	MN469xC port labeled: **TO VNA b1 INPUT**
8	62112-141	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **P1 source loop out**	MN469xC port labeled: **TO VNA Port 1 Src OUTPUT**
9	62112-140	SMA male-male semi-rigid Tighten using an 8 mm (5/16 in) torque end wrench set to 0.9 N·m (8 lbf·in).	MS464xB port labeled: **P1 source loop in**	MN469xC port labeled: **TO VNA Port 1 Src INPUT**
10	2100-1	Rear Panel GPIB Cable 1 meter (39.3”) long	**IEEE 488.2 GPIB**	**Dedicated GPIB**

Architecture Block Diagram

Many architectures are possible (e.g., [1]) but this one was chosen for optimal high frequency performance.



The two VNA ports and the two test receivers are multiplexed to the four user ports for these measurements. The switches are of high isolation, low insertion loss, and high speed to allow for good performance even at 70 GHz.

1. VectorStar Series VNA
2. VNA Test Port 1, Port 1 Source, Port 1 a1, Port 1 b1
3. VNA Test Port 2, Port 1 a2, Port 2 b2
4. MN469xC Test Set
5. Test Set Port 3 and Bias Port 3
6. Test Set Port 4 and Bias Port 4

Figure 24-4. MN469xC Series Test Set Internal Block Diagram

Note

The VNA application must be started ***AFTER*** the Test Set is connected and powered up. If the VNA application is started before the Test Set, the VNA will remain in 2-port mode and the 4-port functions will not be available.

If this happens, exit the VNA application by selecting **MENU BAR | File | Exit** and then in the dialog box, confirm the exit. The VNA application closes to the Windows desktop.

For MS464xB Series VNAs only, if returning to 100,000 point mode, restart the VNA application by selecting: **Start | All Programs | VectorStar_100K | VectorStar**.

For MS464xB Series VNAs, if returning to 25,000 point mode, restart the VNA application by selecting: **Start | All Programs | VectorStar | VectorStar**.

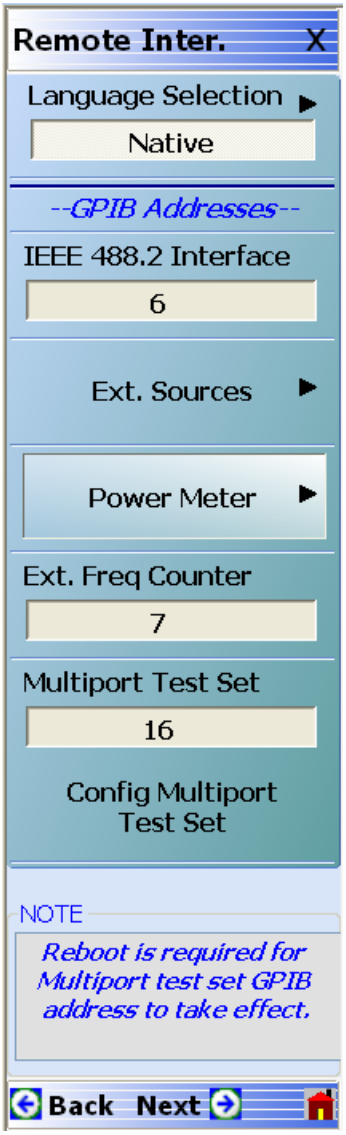
Changing Test Set GPIB Addresses

The MN469xC Series Test Set GPIB address must match the GPIB address set on the VNA (described immediately below) and is set by rear panel Dual-In-Line package switches (DIP-switches). The factory-default Test Set GPIB address is 16 (**Switch 1** ON and all other **switches** OFF) which matches the default VNA address setting. Consult the *VectorStar MN469xC Series Test Set Installation Guide*—10410-00737 for additional information on DIP Switch Settings.

Changing VNA Addresses for the Test Set

The default VNA GPIB address for the MN469xC Series Test Set is GPIB 16 and must match the address set on the Test Set rear panel DIP switches. To change the GPIB address on the VNA, navigate to the REMOTE INTER. (REMOTE INTERFACE) menu shown below (Figure 24-5) and select the Multiport Test Set button. Set the same GPIB address as set at the Test Set rear panel.

- MAIN | System | SYSTEM | Remote Interface | REMOTE INTER. | Multiport Test Set



The **REMOTE INTERFACE** menu contains the **External Test Set/Multiport Test Set** button which accesses the addressing and configuration dialog.

Figure 24-5. REMOTE INTER. (REMOTE INTERFACE) Menu

Power Up Sequence

When the VNA application starts, the GPIB bus is polled for the presence of the Test Set. If the test set is detected, the VNA enters multiport mode and the response, calibration, and certain other menus switch to those shown in this section. If the test set is not detected, the VNA application remains in the 2-port mode discussed in earlier chapters.

Blocking Test Set Combinations

One may note that not every VNA port can be connected to every test set port (termed a blocking test set). All 16 4-port S-parameters can be easily measured but not with an arbitrary setup. The VNA application handles all of these tasks automatically so this detail is normally invisible to the user. It does, however, have some impact on measurement time since there is a limitation on how many parameters can be measured during a sweep with a single switch configuration (for example, S_{11} and S_{21} at the 4-port plane cannot be measured simultaneously). In those situations when multiple switch configurations are required, the VNA application will automatically handle those tasks.

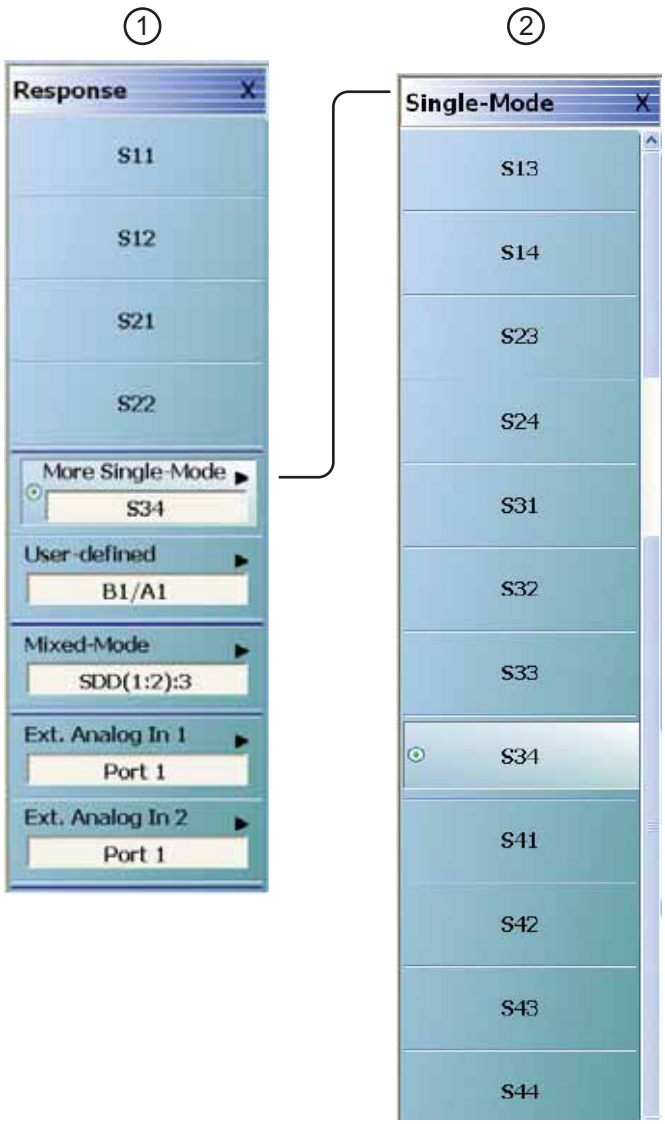
24-3 Conventional (Single-Ended) S- and User-Defined Parameters

As might be expected, there are 16 possible S-parameters when measuring a 4-port device as suggested by the defining equation (Eq. 24-1).

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{bmatrix}$$

Equation 24-1.

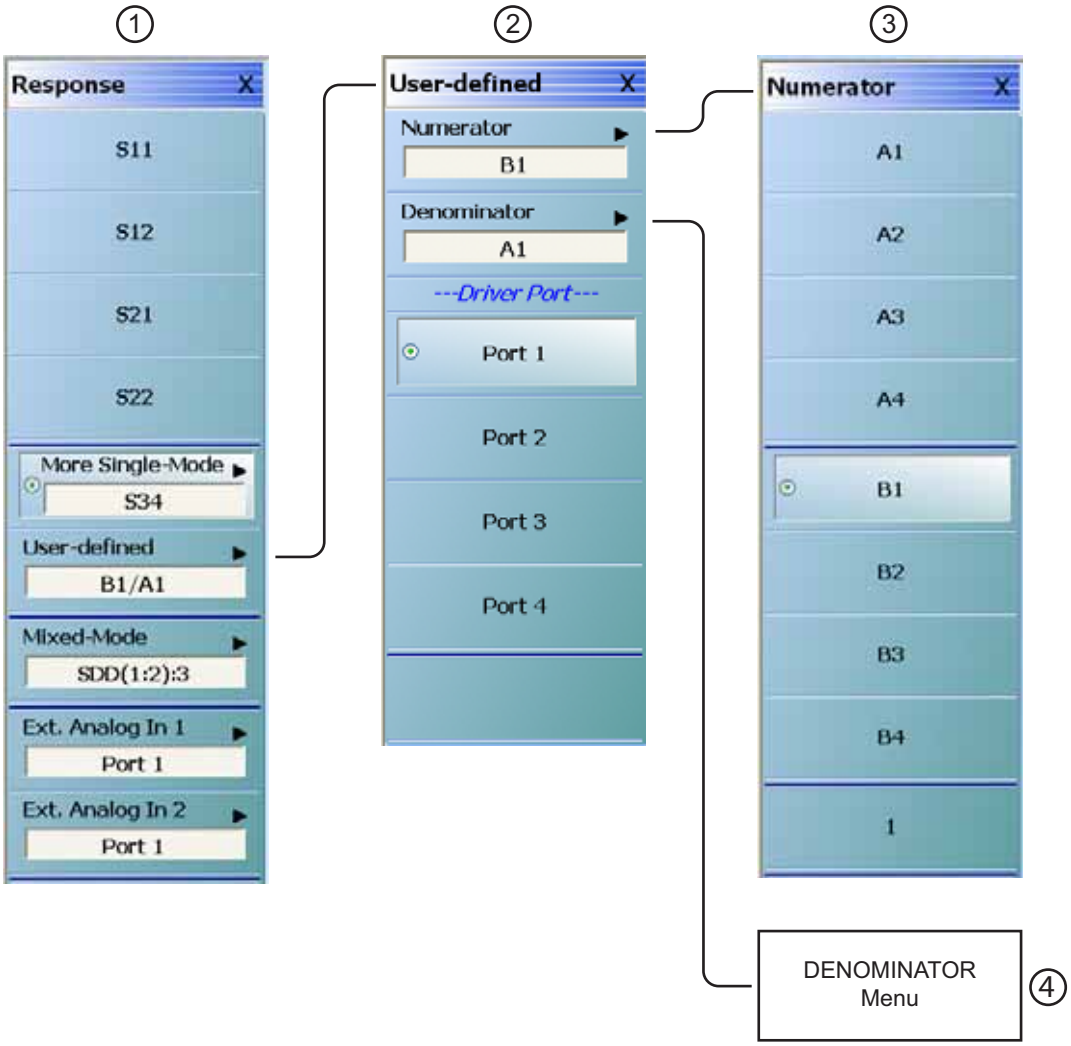
The primary RESPONSE menu and the SINGLE-MODE selection menu for the other standard S-parameters are shown below (Figure 24-6 on page 24-11). Note that the parameters after the first four 2-port S-parameters are contained on a submenu. All of the parameters are available regardless the calibration in place so obviously some parameters could be uncorrected depending on the calibration. The term “single-ended” is used here to denote conventional S-parameters. This is in contrast to “mixed mode” parameters which are explained in the next section. As in the 2-port system, the External Analog In selections are available and function the same as before.



1. The top-level 4-port version of the **RESPONSE** menu is shown here. Buttons provide access to S-Parameters for S_{11}, S_{12}, S_{21}, and S_{33}. The **More Single-Mode** button displays the **SINGLE-MODE** sublevel menu.
2. The **SINGLE-MODE** menu provides S-Parameter selection from S_{13} to S_{44}. The menu representation here is elongated, on an actual system there will be a scroll bar and not all parameters will be visible at once.

Figure 24-6. RESPONSE and SINGLE-MODE Menus

The expected user-defined parameters can be selected as suggested by the menus below (Figure 24-7). Note that any of the four ports may be defined as the driver (or sourcing) port. The allowed numerator and denominator values for user-defined are a_1, a_2, a_3, a_4, b_1, b_2, b_3, b_4, and *1* as suggested by the figure. Note that only the NUMERATOR menu is shown here for simplicity, the DENOMINATOR menu has the same selections.



1. The 4-port version of the **RESPONSE** menu and the submenus linking from the **User-Defined** button.

2. On the lower level of the **USER-DEFINED** response submenu (at center), any one of the four ports may be sourcing RF power as the “driver” port. On the upper level of the **USER-DEFINED** menu, selections are made for the numerator and denominator values of user-defined parameter.

3. The **NUMERATOR** menu (at upper right) menu for user-defined parameters is a submenu to the **USER-DEFINED** menu.

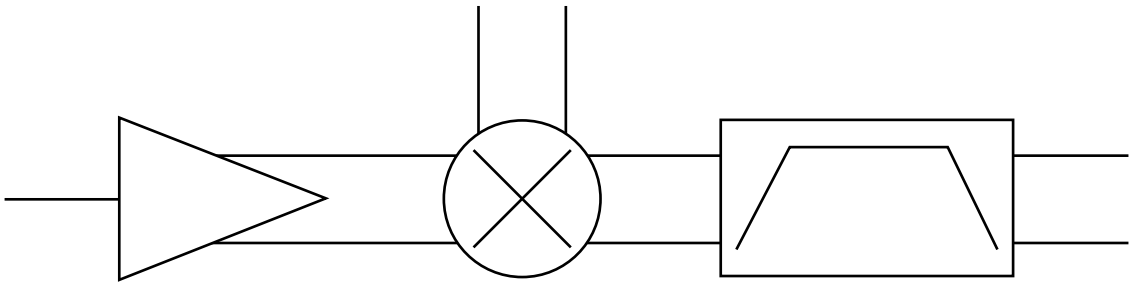
4. The equivalent **DENOMINATOR** menu (at lower right) is not shown.

Figure 24-7. RESPONSE, USER-DEFINED, and NUMERATOR Menus

When a channel is constructed of many traces measuring these parameters, the system will determine the optimal sweep combination to get at all of the required measurements as discussed earlier. The measurement time can be difficult to predict since there are a large number of possible combinations and the test set is partially blocking. Considering again the block diagram of Figure 24-4, *S1x* and *S2x* cannot be measured simultaneously nor can *S3x* and *S4x*. When those pairs (or equivalent b_i measurements) are present in a given channel, extra switching and time will be required. The switching was kept simplified to maintain low insertion loss in the test set at the higher microwave frequencies.

24-4 Mixed-Mode Parameters

The reader will have noted another submenu in the response category that describes mixed-mode parameters. One is likely familiar with the concept of a differential amplifier and the general concepts of differential and common-mode signals [2] as suggested by Figure 24-8. It is clear that it may be useful to represent S-parameter data in a form corresponding to differential drive/reception (and similarly for common-mode). Such a formulation has been developed in the past (e.g., [1]-[4]). The purpose of this section is to describe and define this representation, provide some hints on data interpretation along with explaining how to setup the desired responses.

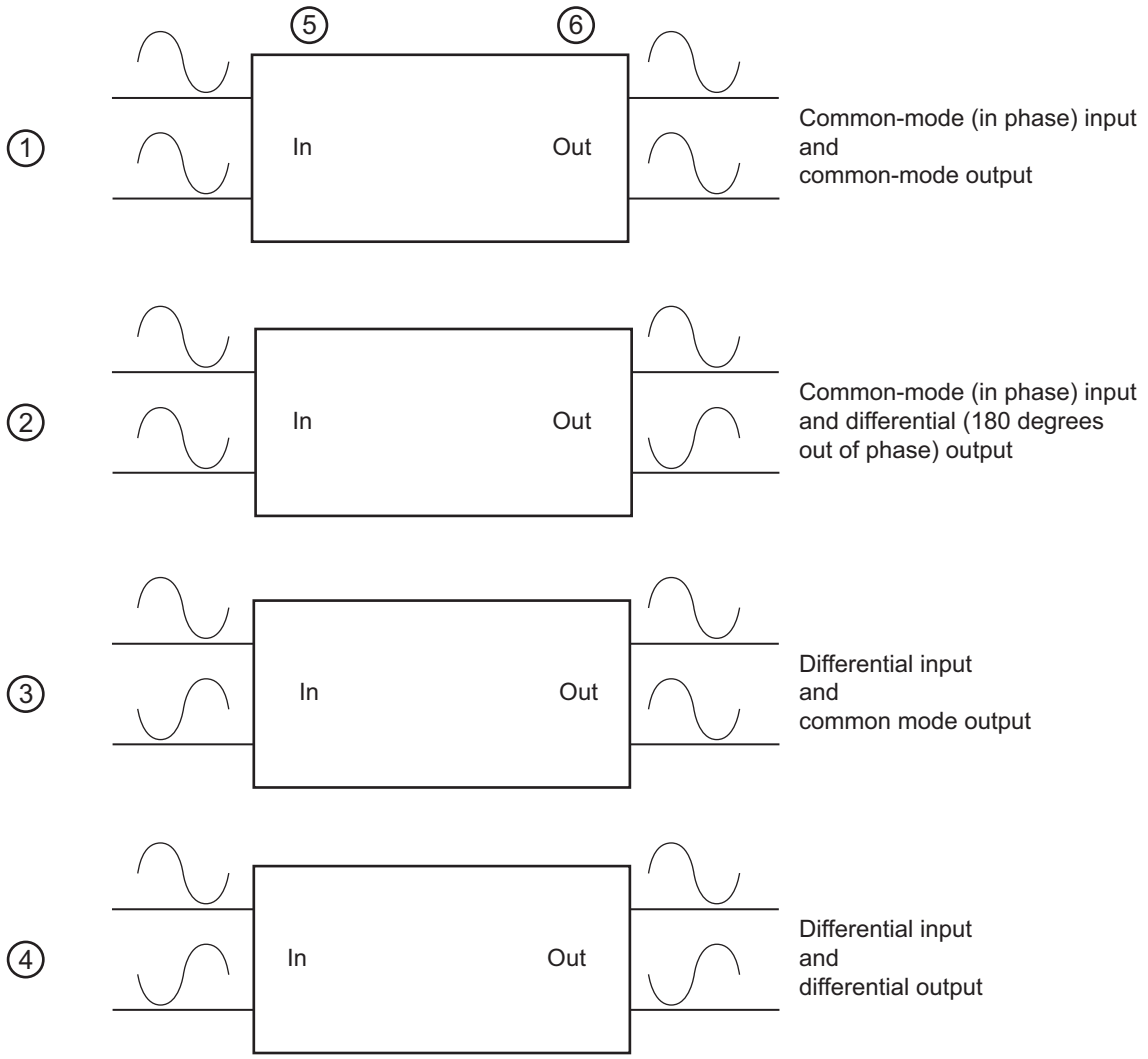


A portion of a common communications circuit is shown here where a number of components are driven differentially. It is of obvious interest to characterize these parts in terms of differential and common mode signals.

Figure 24-8. Characterizing Differentially Driven Components

The concept here is one of a pair of ports being driven at a time, either in-phase (common mode drive) or 180 degrees out of phase (differential drive). Similarly, we are interested in the received signal in a port pair context: what is the common-mode (in-phase) signal received and what is the differential signal received. In contrast to the single-ended drive and single-ended receive types of parameters are conventionally used, these parameters represent a transformation of the data.

This combined signal drive concept can also be explained pictorially. Each pair (either 1-2 or 3-4) can be driven either in phase (common mode drive) or 180 degrees out of phase (differential drive). For the transformation, it is convenient to group together single ended ports 1 and 2 together as the new Port 1 (which can be driven differentially, common-mode, or some combination). Similarly, single ended Ports 3 and 4 will be grouped as the new Port 2 (same idea: differential, common-mode, or combination). Thus the new basis is to think of a port pair as being driven in phase or 180 degrees out of phase (instead of thinking of each port of the pair being driven individually). The new input and output bases are illustrated below (Figure 24-9). The reader may recognize this as a simple shift in basis, which can be thought of as a 45 degree rotation.



The new basis for analyzing mixed-mode S-parameters are shown here. With the physical ports considered as pairs, one can analyze in terms of common-mode and differential drive and common-mode and differential output.

1. Common-mode (in phase) input and common-mode output.
2. Common-mode (in phase) input and differential (180 degrees out-of-phase) output.
3. Differential input and common-mode output.
4. Differential input and differential output.
5. Input Side (at left).
6. Output Side (at right)

Figure 24-9. Analyzing Mixed-Mode S-Parameters

A common question is whether these ports need to be driven with pure differential and common-mode signals or whether the same result can be achieved by mathematical superposition of single-ended signals. With certain assumptions on linearity, the answer is yes and the derivation will be presented next. The only practical requirement is that the bulk of any strong non-linearity be after the first differential stage or that there is reasonable common-mode to differential isolation (e.g., [5]-[6]) which is true for many devices even when gain compression is being measured.

The next step is to construct S-parameters for this type of input/output. The possible incident waveforms can be deduced from Figure 24-9 above:

- differential on the new Port 1
- common-mode on Port 1
- differential on Port 2
- common-mode on Port 2

The output waves will have the same structure, which leads to the S-parameter matrix structure depicted in Eq. 24-2.

$$\begin{bmatrix} b_{d1} \\ b_{d2} \\ b_{c1} \\ b_{c2} \end{bmatrix} = \begin{bmatrix} S_{d1d1} & S_{d1d2} & S_{d1c1} & S_{d1c2} \\ S_{d2d1} & S_{d2d2} & S_{d2c1} & S_{d2c2} \\ S_{c1d1} & S_{c1d2} & S_{c1c1} & S_{c1c2} \\ S_{c2d1} & S_{c2d2} & S_{c2c1} & S_{c2c2} \end{bmatrix} \begin{bmatrix} a_{d1} \\ a_{d2} \\ a_{c1} \\ a_{c2} \end{bmatrix} = \begin{bmatrix} S_{dd} & S_{dc} \\ S_{cd} & S_{cc} \end{bmatrix} \begin{bmatrix} a_{d1} \\ a_{d2} \\ a_{c1} \\ a_{c2} \end{bmatrix}$$

Equation 24-2.

Here the four S blocks in the square matrix to the right are actually submatrices [2]. S_{dd} corresponds to the four purely differential S-parameters while S_{cc} corresponds to the four purely common-mode parameters. The other two quadrants cover the mode-conversion terms. These are expanded in the center part of Eq. 24-2. The matrix equation can be interpreted as expressing an output wave b_i in terms of the four possible input waves a_{d1}, a_{d2}, a_{c1}, and a_{c2}. An example of one of these sub-equations is shown below.

$$b_{d1} = S_{d1d1}a_{d1} + S_{d1d2}a_{d2} + S_{d1c1}a_{c1} + S_{d1c2}a_{c2}$$

Equation 24-3.

It is straightforward to write the relationships between the single-ended incident and reflected waves and the new balanced versions. To simplify the presentation, the next set of equations will assume port pair 1 is composed of physical ports 1 and 2 while port pair 2 is composed of physical ports 3 and 4. We will show a means of generalizing this construction later. The difference and sum choices are obvious for differential and common-mode waves, respectively, and all else that is needed is a normalization factor to keep power levels equivalent.

$$a_{d1} = \frac{1}{\sqrt{2}}(a_1 - a_2)$$

$$a_{c1} = \frac{1}{\sqrt{2}}(a_1 + a_2)$$

$$a_{d2} = \frac{1}{\sqrt{2}}(a_3 - a_4)$$

$$a_{c2} = \frac{1}{\sqrt{2}}(a_3 + a_4)$$

$$b_{d1} = \frac{1}{\sqrt{2}}(b_1 - b_2)$$

$$b_{c1} = \frac{1}{\sqrt{2}}(b_1 + b_2)$$

$$b_{d2} = \frac{1}{\sqrt{2}}(b_3 - b_4)$$

$$b_{c2} = \frac{1}{\sqrt{2}}(b_3 + b_4)$$

Equation 24-4.

This linear combination of single-ended wave functions makes the transformation particularly transparent. One can define a simple transformation matrix to operate on single ended S-parameters to produce the mixed-mode S-parameters that are shown below. The parameters are grouped by the quadrants labeled in Eq. 24-2.

The differential-to-differential terms:

$$S_{d1d1} = \frac{1}{2}(S_{11} - S_{21} - S_{12} + S_{22})$$

$$S_{d1d2} = \frac{1}{2}(S_{13} - S_{23} - S_{14} + S_{24})$$

$$S_{d2d1} = \frac{1}{2}(S_{31} - S_{41} - S_{32} + S_{42})$$

$$S_{d2d2} = \frac{1}{2}(S_{33} - S_{43} - S_{34} + S_{44})$$

Equation 24-5.

The common mode-to-common mode terms:

$$S_{c1c1} = \frac{1}{2}\left(S_{11} + S_{21} + S_{12} + S_{22}\right)$$
$$S_{c1c2} = \frac{1}{2}\left(S_{13} + S_{23} + S_{14} + S_{24}\right)$$
$$S_{c2c1} = \frac{1}{2}\left(S_{31} + S_{41} + S_{32} + S_{42}\right)$$
$$S_{c2c2} = \frac{1}{2}\left(S_{33} + S_{43} + S_{34} + S_{44}\right)$$

Equation 24-6.

The common mode-to-differential terms:

$$S_{d1c1} = \frac{1}{2}\left(S_{11} - S_{21} + S_{12} - S_{22}\right)$$
$$S_{d1c2} = \frac{1}{2}\left(S_{13} - S_{23} + S_{14} - S_{24}\right)$$
$$S_{d2c1} = \frac{1}{2}\left(S_{31} - S_{41} + S_{32} - S_{42}\right)$$
$$S_{d2c2} = \frac{1}{2}\left(S_{33} - S_{43} + S_{34} - S_{44}\right)$$

Equation 24-7.

and the differential-to-common mode terms:

$$S_{c1d1} = \frac{1}{2}\left(S_{11} + S_{21} - S_{12} - S_{22}\right)$$
$$S_{c1d2} = \frac{1}{2}\left(S_{13} + S_{23} - S_{14} - S_{24}\right)$$
$$S_{c2d1} = \frac{1}{2}\left(S_{31} + S_{41} - S_{32} - S_{42}\right)$$
$$S_{c2d2} = \frac{1}{2}\left(S_{33} + S_{43} - S_{34} - S_{44}\right)$$

Equation 24-8.

The simple linear relationship between the parameters should be evident. Just to reinforce the interpretation of these parameters: S_{d2d1} is the differential output from composite Port 2 (the old Ports 3 and 4) ratioed against a differential drive into composite port 1 (the old Ports 1 and 2). Similarly S_{c2c2} would be the common-mode reflection off of composite Port 2 ratioed against the common-mode signal applied to composite Port 2.

MIXED MODE Dialog Box

The dialog for selecting a mixed mode parameter in this sense is shown in Figure 24-10.

- MAIN | Response | RESPONSE | Mixed-Mode | MIXED MODE Dialog Box

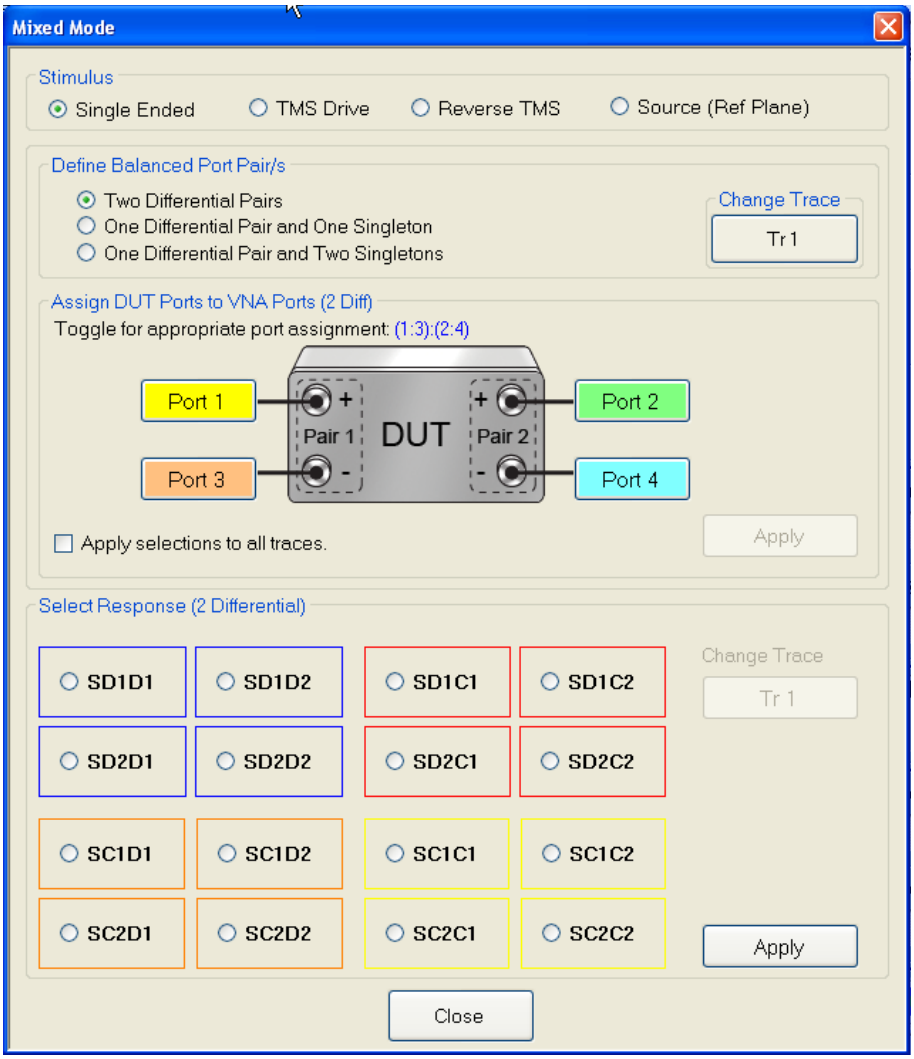
Aside from selecting among the 16 parameters described above, there is also the matter of defining the port pairs. This is accomplished in the middle part of the dialog and note that there is a polarity sense assigned to the pairing (determines the sign of the differential signals). When displaying the parameters, this port assignment must also be conveyed and is done with the following symbolism:

- (A:B):(C:D) (where A, B, C, and D are unique port numbers from {1,2,3,4})

Here, the first port pair is A-B (with A assigned the positive polarity) and the second port pair is C-D (with C assigned the positive polarity).

For example:

- (1:2):(3:4) The first port pair is measured from 1 to 2 and the second port pair is measured from 3 to 4.
- (1:2):(4:3) The first port pair is measured from 1 to 2 and the second port pair is measured from 4 to 3.



The dialog for selecting mixed-mode parameters, configured for two differential pairs, is shown here. The instrument defaults to the SD1D1 response.

Figure 24-10. MIXED MODE Two Differential Pairs Dialog Box

The 3-port version of these mixed-mode parameters is a straightforward simplification. Port X will remain single ended while ports Y and Z will form a balanced port pair. Following the analysis path from before:

$$\begin{bmatrix} b_x \\ b_d \\ b_c \end{bmatrix} = \begin{bmatrix} S_{xx} & S_{xd} & S_{xc} \\ S_{dx} & S_{dd} & S_{dc} \\ S_{cx} & S_{cd} & S_{cc} \end{bmatrix} \begin{bmatrix} a_x \\ a_d \\ a_c \end{bmatrix}$$

Equation 24-9.

$$a_d = \frac{1}{\sqrt{2}}(a_y - a_z)$$
$$a_c = \frac{1}{\sqrt{2}}(a_y + a_z)$$
$$b_d = \frac{1}{\sqrt{2}}(b_y - b_z)$$
$$b_c = \frac{1}{\sqrt{2}}(b_y + b_z)$$

Equation 24-10.

$$S_{xd} = \frac{1}{\sqrt{2}}(S_{xy} - S_{xz})$$
$$S_{xc} = \frac{1}{\sqrt{2}}(S_{xy} + S_{xz})$$
$$S_{dx} = \frac{1}{\sqrt{2}}(S_{yx} - S_{zx})$$
$$S_{cx} = \frac{1}{\sqrt{2}}(S_{yx} + S_{zx})$$
$$S_{dd} = \frac{1}{2}(S_{yy} - S_{yz} - S_{zy} - S_{zz})$$
$$S_{cc} = \frac{1}{2}(S_{yy} + S_{yz} + S_{zy} + S_{zz})$$
$$S_{dc} = \frac{1}{2}(S_{yy} + S_{yz} - S_{zy} - S_{zz})$$
$$S_{cd} = \frac{1}{2}(S_{yy} - S_{yz} + S_{zy} - S_{zz})$$

Equation 24-11.

In Equation Group 11 (Eq. 24-11), the subscript *x* remains since that port is defined solely as a single-ended port.

The three port configuration is termed a differential pair and a singleton and the dialog for selecting this type of parameter is shown in Figure 24-11. Note that one can define this parameter class even if a 4-port calibration is applied; it is a matter of port assignment. As with the dual differential case, the ports must be assigned with polarity in mind.

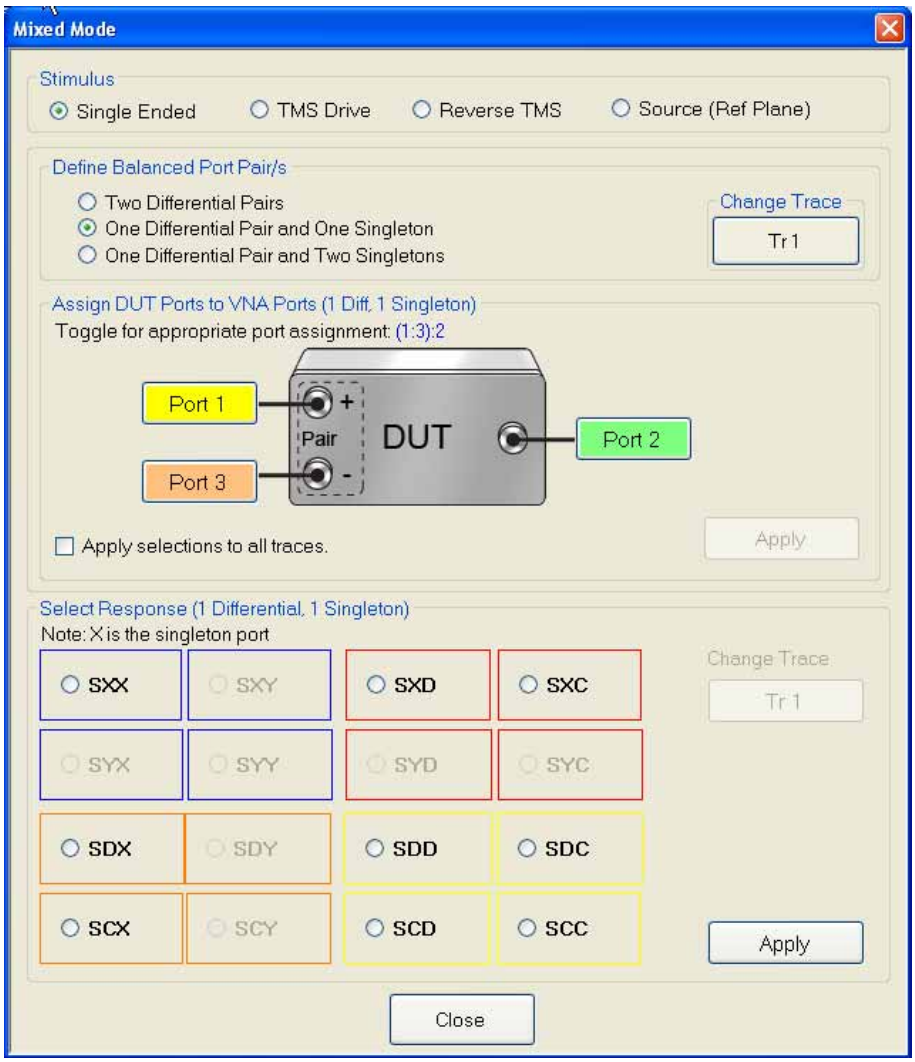


Figure 24-11. MIXED MODE One Differential Pair and One Singleton Dialog Box

For the situation in the figure above (Figure 24-11), there will be a different annotation on the graph legends to denote how the ports are configured but it is similar to the two-differential-pair format discussed earlier. (1:2):3 denotes that the differential pair is measured from Port 1 to Port 3 and Port 2 is the singleton.

Certain 4-port devices are configured as one differential pair plus two single-ended ports. The parameters can then be configured in a one differential pair and two singleton format. The mixed mode definitions are identical to those presented above for the 1 singleton case. Now there is just a different possible free port number running around in the definition and 16 parameters are again possible. This selection structure is shown below in Figure 24-12.

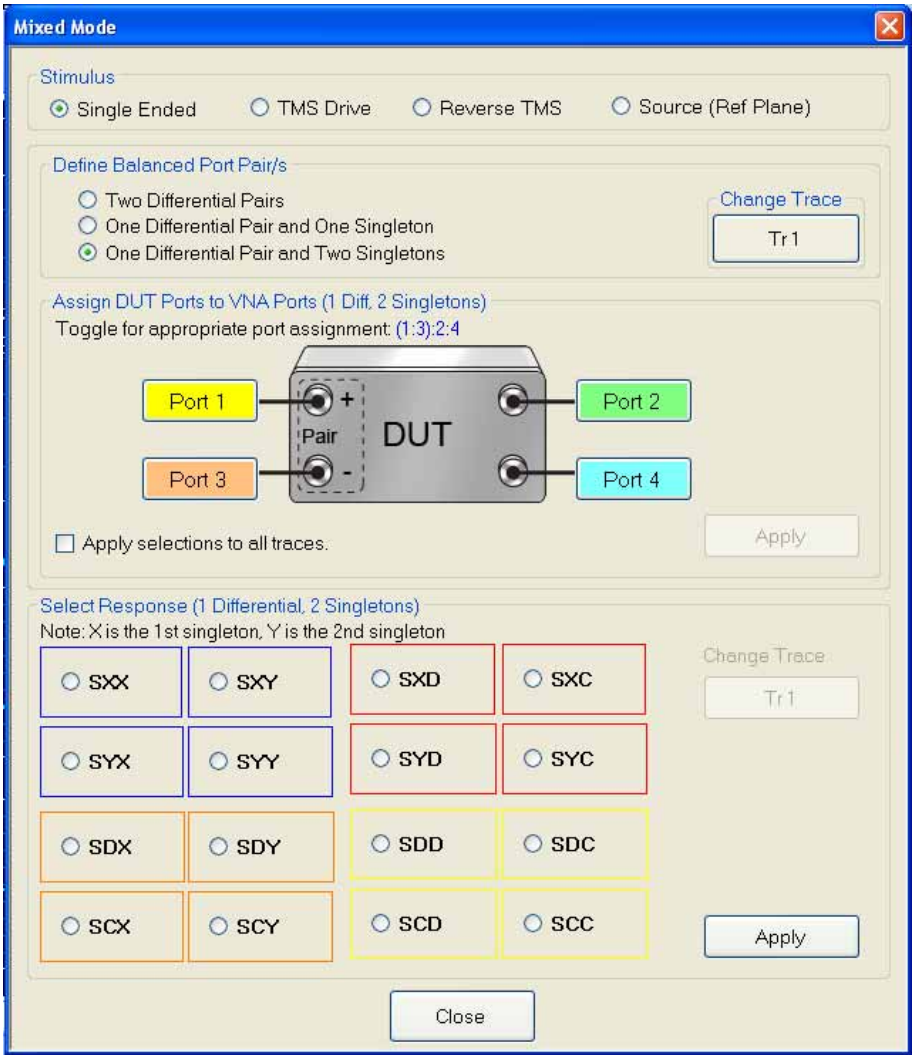


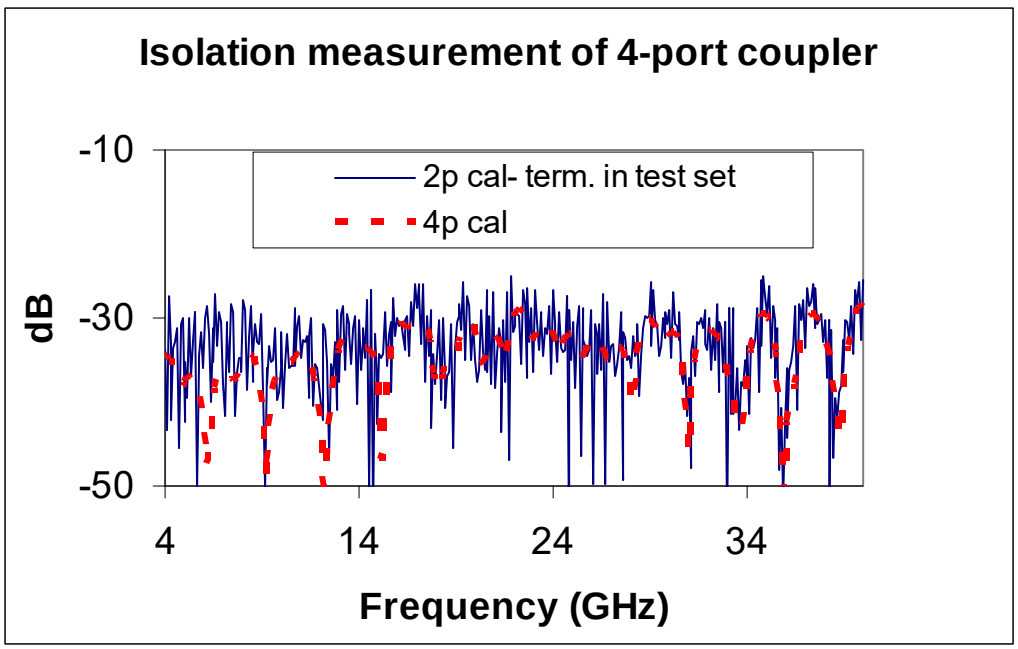
Figure 24-12. MIXED MODE One Differential Pair and Two Single-Ended Ports Dialog Box

On the graph legends, a format like (1:3):2:4 will be used to denote the port configuration. In this example, the differential pair is measured from Port 1 to Port 3, Port 2 is the first singleton and Port 4 is the second singleton.

24-5 Calibrations

In earlier chapters, a considerable amount of detail on various calibration algorithms and calibration types has been covered. The 3- and 4-port variants of these calibrations are different from their 2-port relations in fairly straightforward ways. The purpose of this subsection is to show where the differences are and how these calibrations can be properly configured.

A logical starting point is to ask what is different between measuring, say S_{41}, using a 4-port calibration versus measuring it with a 2-port calibration while properly terminating Ports 2 and 3. The quick answer is nothing, assuming the terminations are of good quality and one does not care about measurement time when measuring many parameters (while one swaps terminations around to get all of the S-parameters). If one just connected Ports 2 and 3 of the DUT to VNA ports, the results would probably not be as good since the raw broadband return loss of a VNA port is usually not better than 15 dB. To illustrate this concept, consider the measurement of isolation of a 4-port coupler (S_{41} in the previous discussion). If a full 4-port calibration is performed, the dotted line in Figure 24-13 is obtained. If instead only a 2-port cal is performed and the other coupler ports are left terminated in the test set (seeing the raw match of those ports), then the solid line of Figure 24-13 is obtained. The error here could be on the order of 10 dB at the –30 dB level.



An example showing the measurement improvement of a full 4-port calibration over a partially terminated 2-port calibration is shown here.

Figure 24-13. Measurement Improvement for Full 4-Port Calibration

By performing a full 3- or 4-port calibration, the effective return loss of those other ports is improved to that of the residual load match (typically 30–50 dB depending on the calibration algorithm and components used) thus improving the measurement result. Thus overall measurement time and accuracy are the usual drivers for considering 3- and 4-port calibrations.

The first step in understanding these calibrations is to look at the complete error model for N ports. As is apparent from the figure below (Figure 24-14), it is a straightforward extension of the 2-port error model covered in Chapter 1, “Calibration and Measurement Overview”. The directivity and match terms are basically functions of the port itself so one must just add these terms for the additional ports.

Tracking

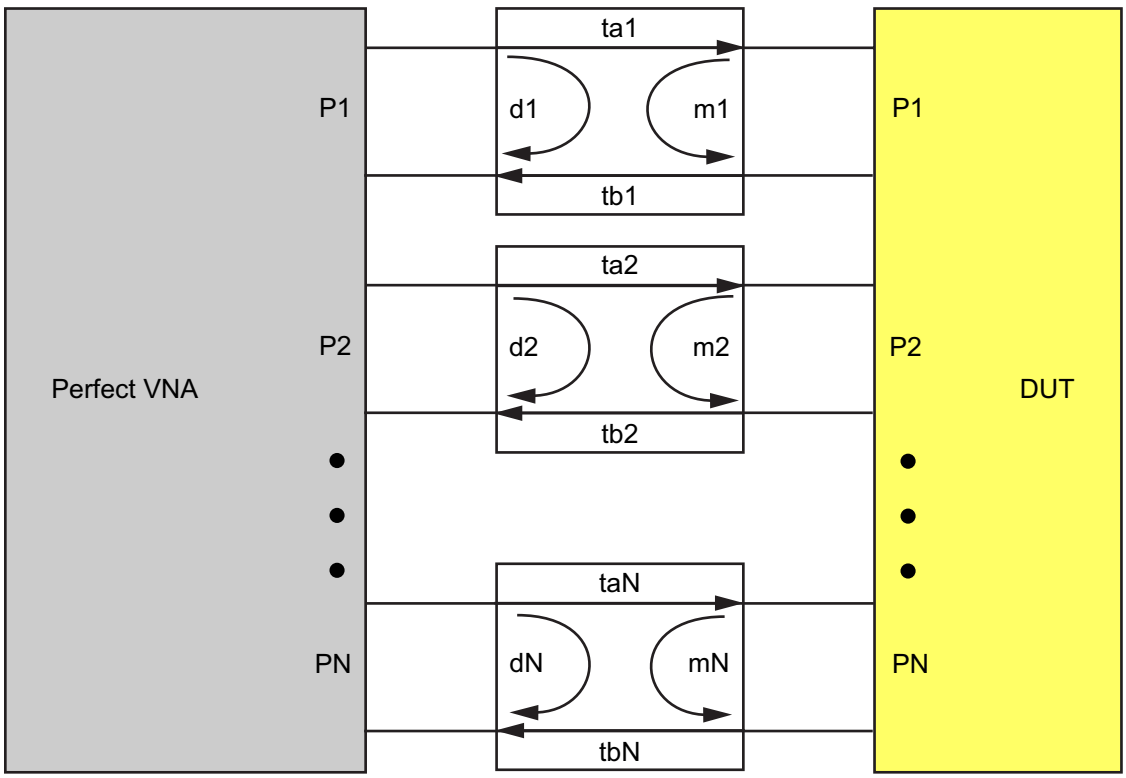
Tracking is again broken down into reflection tracking (the frequency response on a reflection measurement) and transmission tracking (the frequency response on a transmission measurement). Reflection tracking then also is a port property and can be assigned for each additional port. Transmission tracking is a property of a pair of ports (the correction to get a thru line to measure S_{ij}=1 after other port errors have been handled) hence this term must be assigned for every permutation of two ports.

Isolation

Finally isolation is usually also considered the property of a pair of ports so the same permutation will be required. The last statement is not entirely valid if isolation is dependent on DUT impedances and is normally reserved to correct for isolation deep in the instrument. This distinction will be ignored for now since the isolation step is optional.

New Error Model

Thus we have a fairly straightforward new error model as shown below (Figure 24-14). As before, the pair of lines (in the diagram) per port is to help delineate incident and reflected waves. The reflection tracking for port *q* can be thought of as the product of *taq* and *tbq*; the transmission tracking from port *k* to port *l* can be thought of as the product of *tak* and *tbl* (no new information is being added, it is just being redistributed).

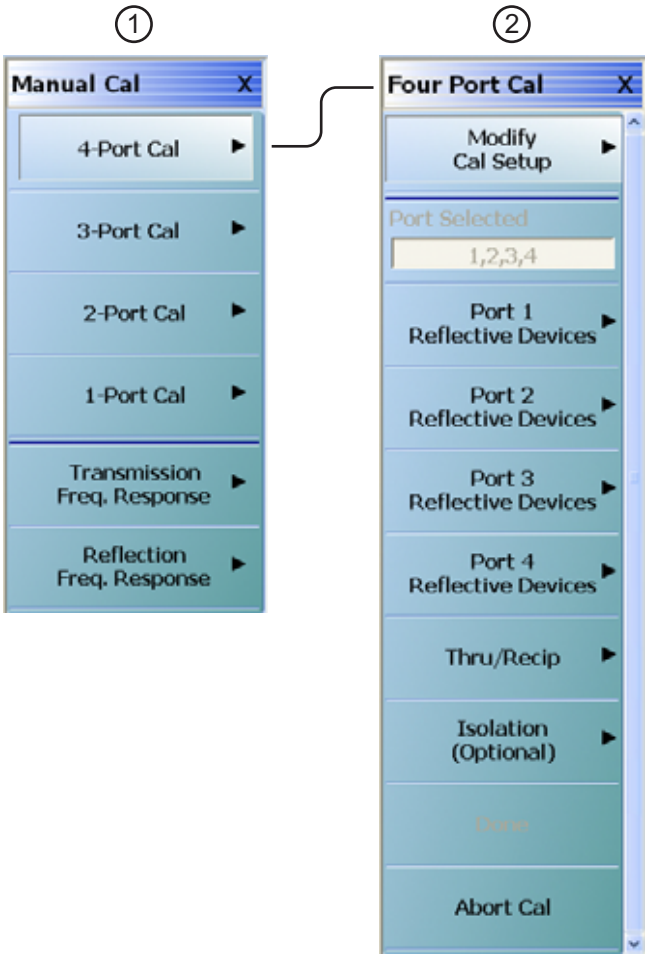


An N-port error model is shown here. Directivity, match and reflection tracking can be assigned on a per-port basis while transmission tracking can be assigned on a pair-of-ports basis. Isolation is also nominally assigned on a pair-of-ports basis but that does have problems (isolation is often ignored practically).

Figure 24-14. N-Port Error Model

Menus

As might be expected, the main calibration menu will have a selection based on the type of calibration (full or nearly so versus frequency-response) and the number of ports. The MANUAL CAL menu below (at #1 left in Figure 24-15) provides the selection path to the FOUR-PORT CAL menu (at #2 below). This menu is the calibration execution menu for a 4-Port calibration (defined standards call.



1. The **MANUAL CAL** menu (at left) provides access to all multiport calibration modes. The **4-Port Cal** button provides access to the **FOUR PORT CAL** procedure menu for a 4-Port VNA.	2. The **FOUR PORT CAL** calibration execution menu (at right) for a 4-Port calibration (defined-standards calibration).

Figure 24-15. MANUAL CAL (MANUAL CALIBRATION) and FOUR PORT CAL Menus

The menus for the lower port count defined-standards calibrations follow in a predictable fashion. On the FOUR PORT CAL menu (at 2 right in Figure 24-15 above), the second field from the top, a read-only box, will indicate which ports are in play for the current calibration. The details on selecting these ports will be presented later.

Transmission Tracking

The key complication of multiport calibrations lies in the handling of transmission tracking. Since the reflection and transmission tracking terms overlap, they are not entirely independent. The following is the relationship that one can pull directly off of Figure 24-14, “N-Port Error Model” on page 24-23:

$$etkn \cdot etnl = etnn \cdot etkl$$

Equation 24-12.

In reality, this must be modified slightly for load match/source match changes, but the redundancy concept still applies. To see how this is used, consider the redundant thru line problem (as in previous chapters, we will use the term “through (thru)” to denote any fully defined transmission line whether it be of zero length or not). Suppose we have connected thrus between Ports 1-2, 1-3, and 1-4 sequentially. Presumably, the first three term types (directivity, source match, and reflection tracking) have been computed for all four ports. The thru connects allow us to compute all of the load match terms and all transmission tracking terms except for et32, et23, et24, et42, et34 and et43. These last tracking terms can be found with the following forms of Eq. 24-12 (above) shown below in Eq. 24-13:

$$et43 = \frac{et41 \cdot et13}{et11} \qquad (k = 4, l = 3, n = 1)$$

$$et32 = \frac{et31 \cdot et12}{et11} \qquad (k = 3, l = 2, n = 1)$$

$$etc.$$

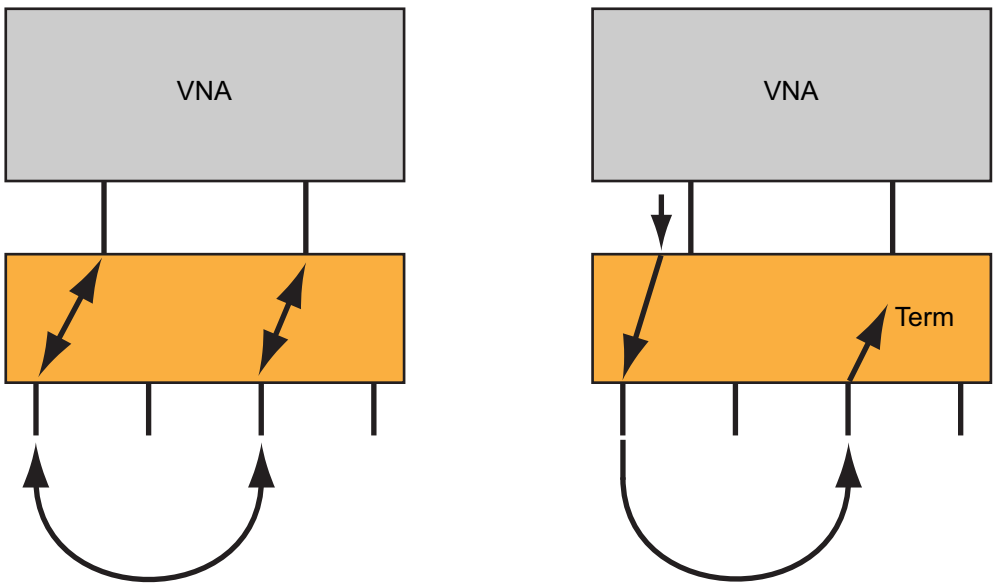
Equation 24-13.

Again, the above is modified slightly because of match differences. This process is termed the use of redundancy since the reflect measurements and the thru measurements do overlap to a certain degree. Thus, one does not need to connect all six (6) possible thru lines in a 4-Port calibration or all three (3) possible thru lines in a 3-Port calibration. The basic requirement is that all ports be “connected” in some way with a line. Now, additional thrus can always be used and can improve accuracy of the tracking terms by a certain amount. There are many cases, however, where it is difficult to construct a high quality line between ports (orthogonally positioned wafer process for example) and the use of redundancy will produce a better result than trying to use a poor thru during the calibration. Similar analyses are used for all calibration types so that the user will see considerable flexibility in how the thrus/lines are defined.

Load Match

A related topic is load match (primarily for defined-standards calibrations) since it can generally be computed in multiple different ways since multiple thrus are involved. The guiding principle is the first-in rule: the first thru measured will be used to compute the load match for the ports involved, the second thru will be used for its ports but it will not overwrite earlier computed load matches, and so on. This allows the user to optimize load match performance by measuring the highest quality thrus first.

A consistent load match is guaranteed by generally using the test set terminations to form the reference load for all transmission measurements. When the linkage to the VNA termination is required (for reciprocal calibrations and for the LRL/LRM families), the final load matches are transformed to the test set termination in order to guarantee consistency. This concept is shown below (Figure 24-16). Thus whenever a transmission parameter is measured, a unique load match is always defined and the corrections are self-consistent.

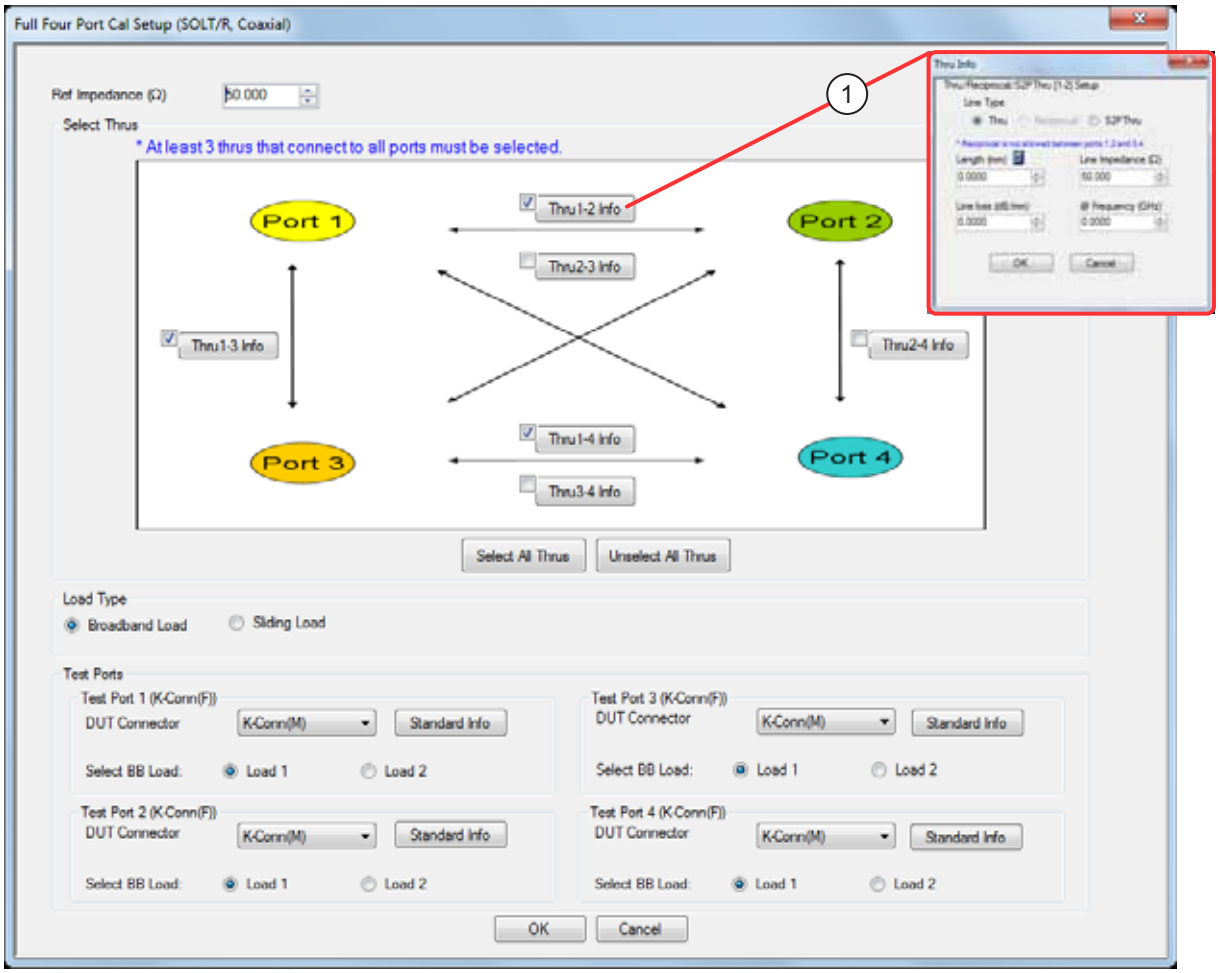


The concept of load match plane translation is shown here. Although the connection (shown on the left) may be needed for LRL/LRM and SOLR-like calibrations, the reference plane for load match will always be translated to the test set termination (as shown on the right).

Figure 24-16. Load Match Plane Translation

Setting Up the Defined Standards Calibration

The next topic is that of setting up the defined standards calibration. Consider the dialog shown below (Figure 24-17) that is used for setting up a 4-Port calibration using the SOLT/R algorithm.



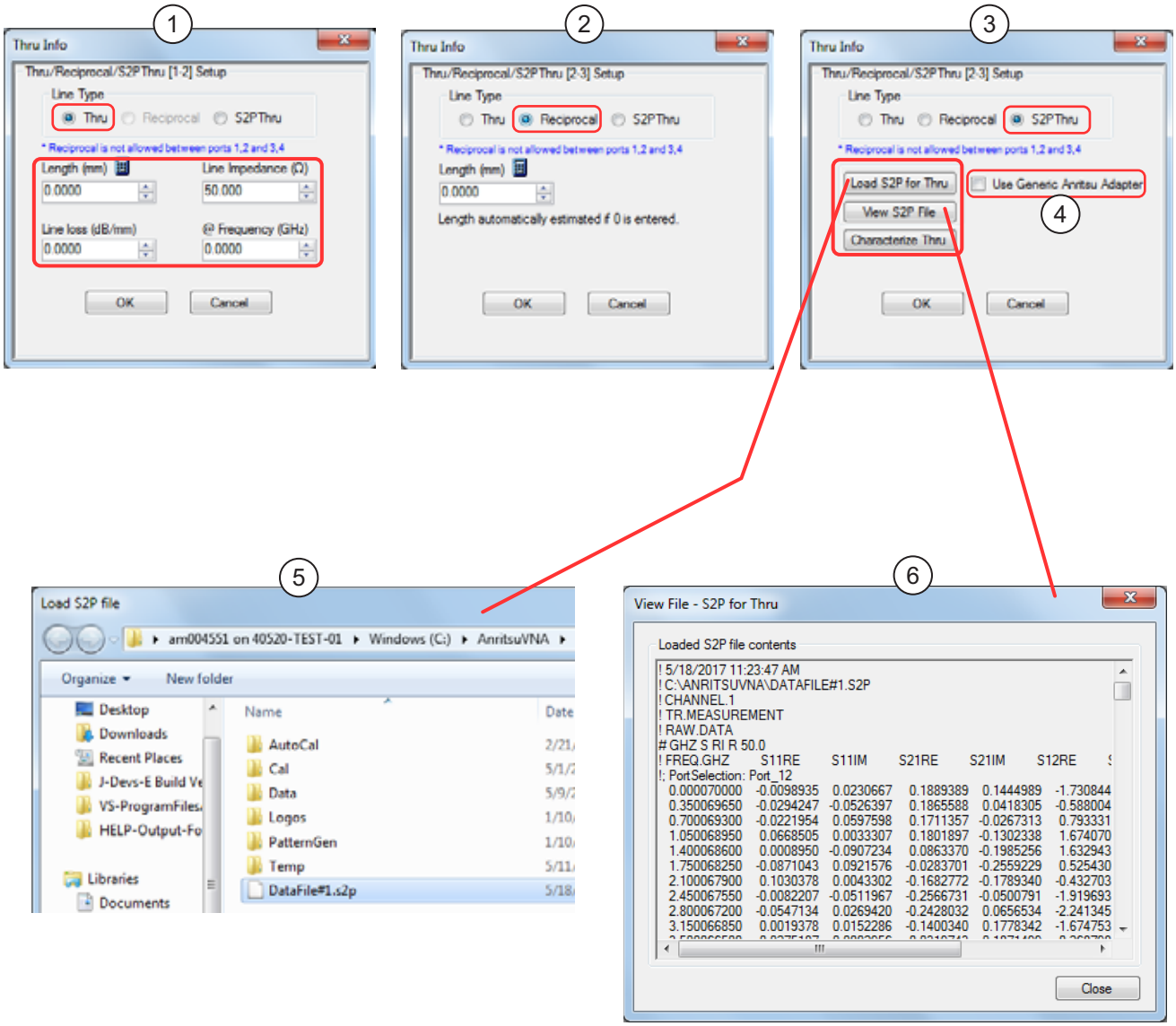
1. Selecting button opens Thru Info Dialog where the user can alter Thru, Reciprocal or S2P Thru parameters. See Figure 24-18 on page 24-28 for details.

Figure 24-17. FULL FOUR PORT CAL SETUP Dialog Box—SOLT/R—Coaxial

Selecting Possible Thrus

As might be expected, there are the usual items for selecting the port connector types, the type of load (broadband or sliding) and the reference impedance. What is somewhat different is the thru selection tableau. The six possible thrus are shown, each having a button allowing the definition of that line as shown below (Figure 24-18). As discussed earlier, at least three thrus must be selected such that all ports are "connected."

- Allowed:
  - 1-2, 1-3, and 1-4
  - 1-3, 2-4, and 1-4
  - etc.
- Not allowed (without adding more lines):
  - 1-2, 1-3, and 2-3 (port 4 is not "connected")



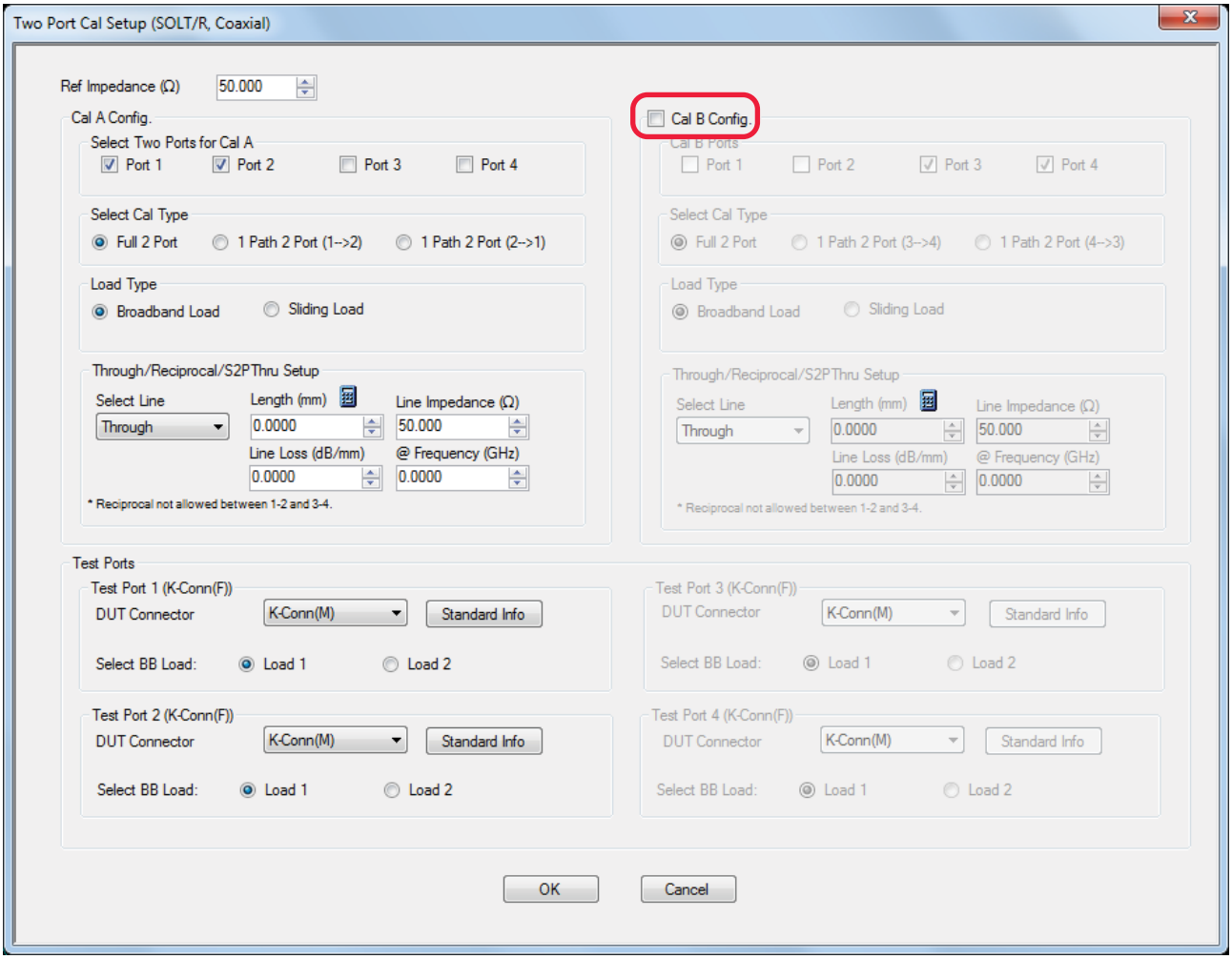
Typical thru line configuration in the THRU INFO dialog box for the defined-standards calibrations.

1. Thru Info Dialog—Line Type Thru selected, allowing configuration of Thru
2. Thru Info Dialog—Line Type Reciprocal selected, allowing configuration of Thru
3. Thru Info Dialog—Line Type S2P Thru selected, allowing loading, viewing and generation of S2P files
4. Use Generic Anritsu Adapter (check box)—Checking this box will automatically load a default .s2p file for a generic Anritsu adapter. Note that selecting the checkbox disables the Load S2P for Thru and Characterize Thru buttons.
5. File Load Dialog
6. View S2P File contents window

Figure 24-18. THRU INFO Dialog Box

As suggested by the figure above (Figure 24-18), each of the selections may be defined as a “thru” or as a “reciprocal” as discussed in earlier chapters. From the point-of-view of “connectedness”, these designations are equivalent. As discussed in an earlier chapter, however, there is some accuracy degradation (particularly with regard to load match) that can be compounded if all lines are reciprocals. This point is discussed in more detail in the uncertainty section.

The three port calibration is analogous and the menus are covered in the operation guide. One port (to include reflection response calibrations) and 2-port calibrations are a little different in that multiple calibrations are allowed. Since it is a 4-port instrument, the user is allowed to perform two 2-Port calibrations or up to four 1-Port calibrations simultaneously. The 2-port calibration setup dialog is shown below (Figure 24-19) to illustrate the option of another calibration (Cal B in this dialog).

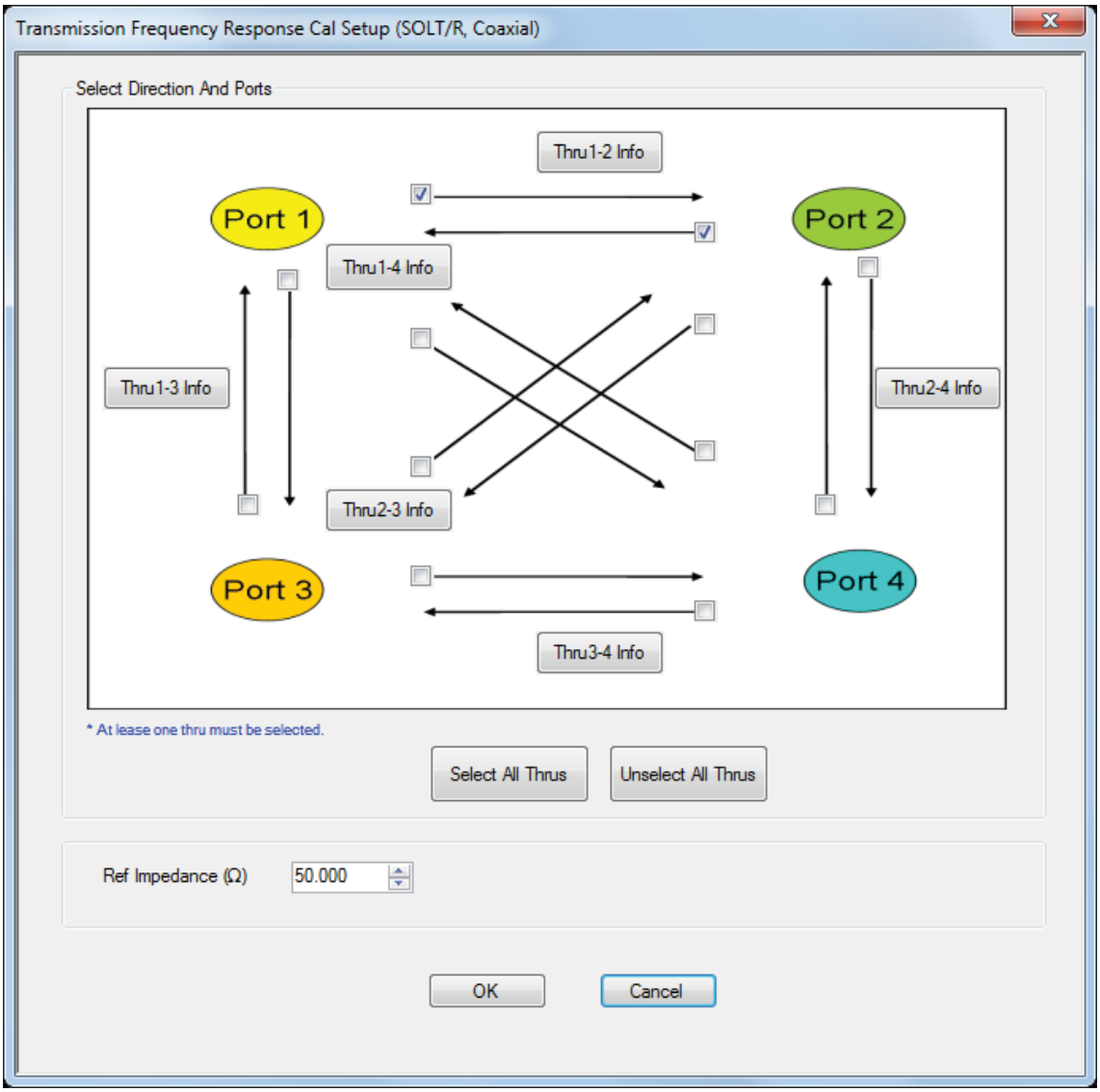


Typical calibration setup for a 2-port calibration with selections of SOLT/R, Coaxial, and Cal B Not Selected. The 2-port modify setup (SOLT/R) dialog is shown here to illustrate the multiple calibration functionality.

Figure 24-19. TWO PORT CAL SETUP (SOLT/R, COAXIAL) Dialog Box

If Cal B is activated from the check box, then the other two ports (not defined under Cal A) can be used to form an independent calibration. Port overlap is not allowed in order to avoid error-term self-consistency issues.

Transmission frequency response calibrations generalize as one might expect since there are now 12 transmission S-parameters potentially available to normalize. The dialog is shown below (Figure 24-20). As with the full calibrations, the thrus may be defined independently although the reciprocal choice is not available in this case. Note that unlike thru selections in the full calibrations, the directions can be independently selected in this case. As is the case with 2-port systems, it is allowed to normalize, say, S_{21} but not S_{12}. In the full calibration case, both directions are always required.



The transmission frequency response modify setup dialog is shown here for a 4-port system.

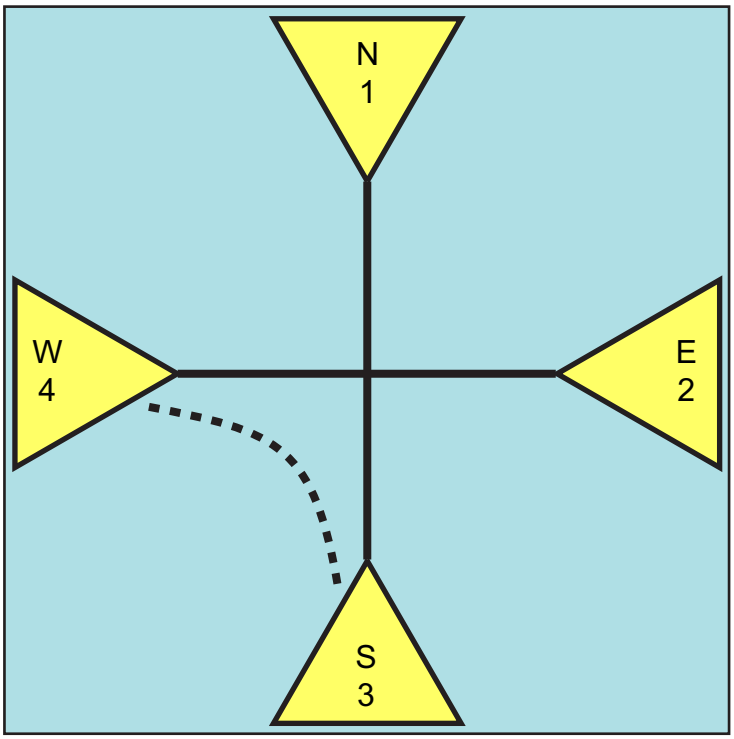
Figure 24-20. TRANSMISSION FREQUENCY RESPONSE CAL SETUP Dialog Box—SOLT/R—Coaxial

The above dialogs and setup requirements are completely analogous for SSLT/R and SSST/R algorithms. The LRL/LRM family, however, is somewhat different.

LRL/LRM Calibrations

In this implementation LRL/LRM family, there is a basic structure that is fundamentally 2-Port like in nature. As such, the 4-Port LRL/LRM calibration will be defined as a pair of basic 2-Port calibrations linked together by one or more thru measurements (to complete the path linkage as for defined-standards calibrations). This can have an additional benefit in certain cases since it allows the hybridization of reciprocal and defined-thru approaches (on the additional paths) with LRL/LRM approaches on the base paths.

A situation where this is useful is shown in the figure below (Figure 24-21) where orthogonal on-wafer probes are being used. It is relatively easier to define good transmission lines on the horizontal and vertical paths (1-3 and 2-4) but less so, at least for higher frequencies, on the other paths (bends in the required lines causing radiation or otherwise leading to reflections). Such a problem is a good candidate for hybridization.

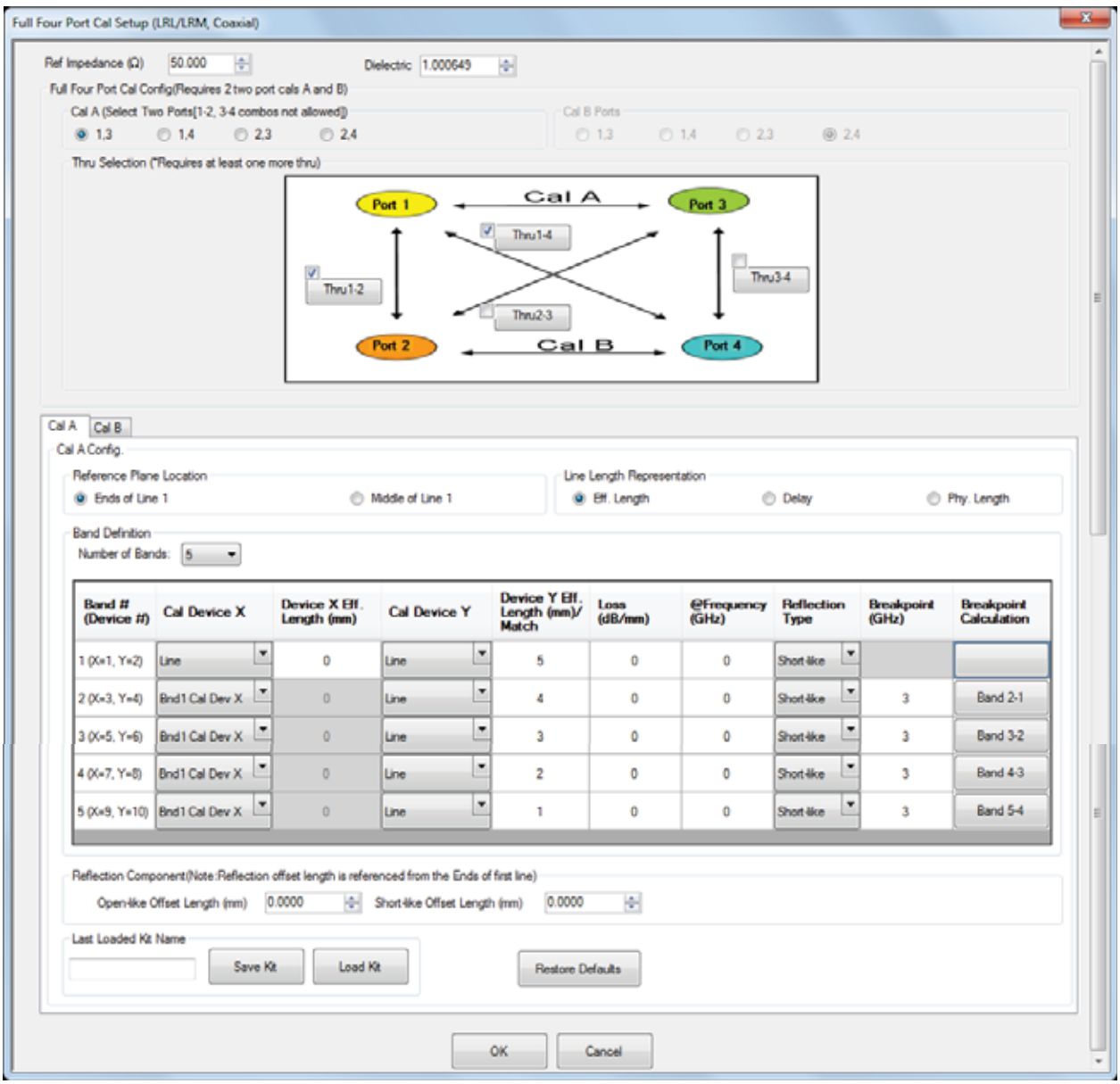


An on-wafer configuration where some line paths are more likely to be of higher quality than others. The solid lines will generally not be an issue but a transmission path illustrated by the dotted line (S-W) may be electromagnetically compromised at higher frequencies.

Figure 24-21. On-Wafer Line Path Quality

4-Port LRL/LRM Calibrations

The configuration dialog for 4-Port LRL/LRM calibrations is shown in the figure below (Figure 24-22). Note how the primary LRL/LRM ports are defined and how additional information may be required for the continuation lines. The two 2-Port calibrations are defined on the two tabs labeled Cal A and Cal B. Each tab alone looks the same as the 2-port LRL/LRM calibrations discussed in earlier chapters. Note that on the actual display all of the information may not be visible and a scroll bar may be required.



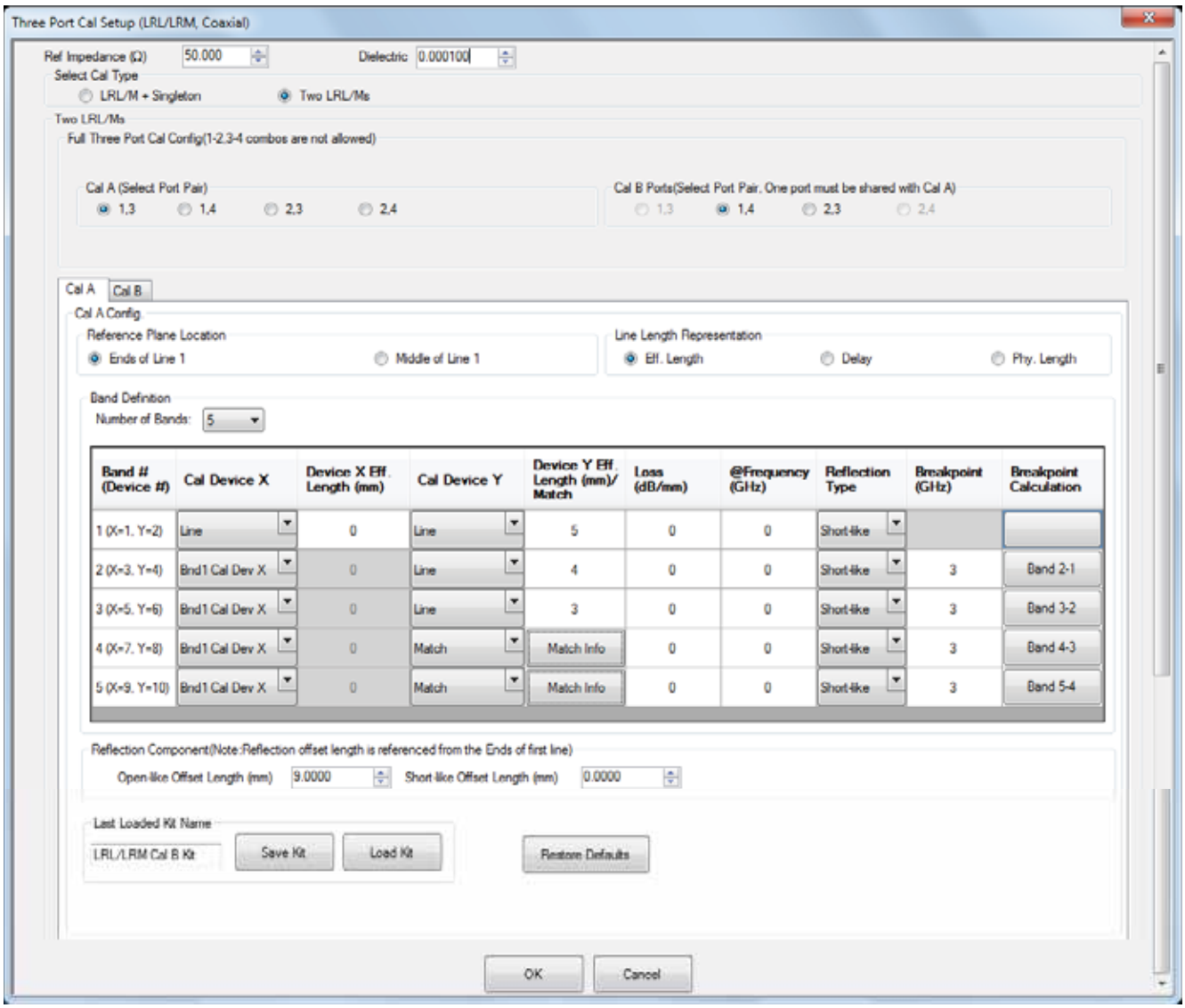
The modify calibration dialog for 4-Port LRL/LRM calibrations is shown here.

Figure 24-22. FOUR PORT CAL SETUP (LRL/LRM, COAXIAL) Dialog Box

As in the 2-port case, up to five bands may be defined with various permutations of lines and matches. Also as in the 2-port case, the transition frequencies between each of the bands are adjustable to maximize convenience and calibration accuracy. For more details on these choices, see Chapter 6 of this guide. The Cal A and Cal B band definitions do not need to be the same and, indeed, they can actually have no calibration devices in common. Such a situation may arise in a mixed media configuration but it does place some constraints on the completion thru definitions.

Three Port LRL/LRM Calibration

Much like the 4-Port LRL/LRM calibration, the 3-Port version is constructed from a pair of 2-Port LRL/LRM calibrations (unlike the defined-standards approaches). Obviously there is going to be port overlap. The first calibration will define the port reflectometer parameters on the overlap port. The modify calibration dialog for this situation is shown below (Figure 24-23).



The modify calibration dialog for the three port LRL/LRM calibration is shown here.

Figure 24-23. THREE PORT CALSETUP (LRL/LRM, COAXIAL) Dialog Box—Two LRL/Ms—Cal A Tab

The 2-Port calibration situation for LRL/LRM is similar to that shown for the defined-standards case: it is allowed to perform two of these calibrations in parallel. Aside from the additional definition required, the approach is the same as discussed in an earlier chapter.

This 3-Port calibration can also be constructed from a single 2-Port LRL/LRM cal together with a singleton definition. This version is shown below (Figure 24-24). The single LRL/LRM cal must be defined as usual and the one or two extra thrus must also be defined as before.

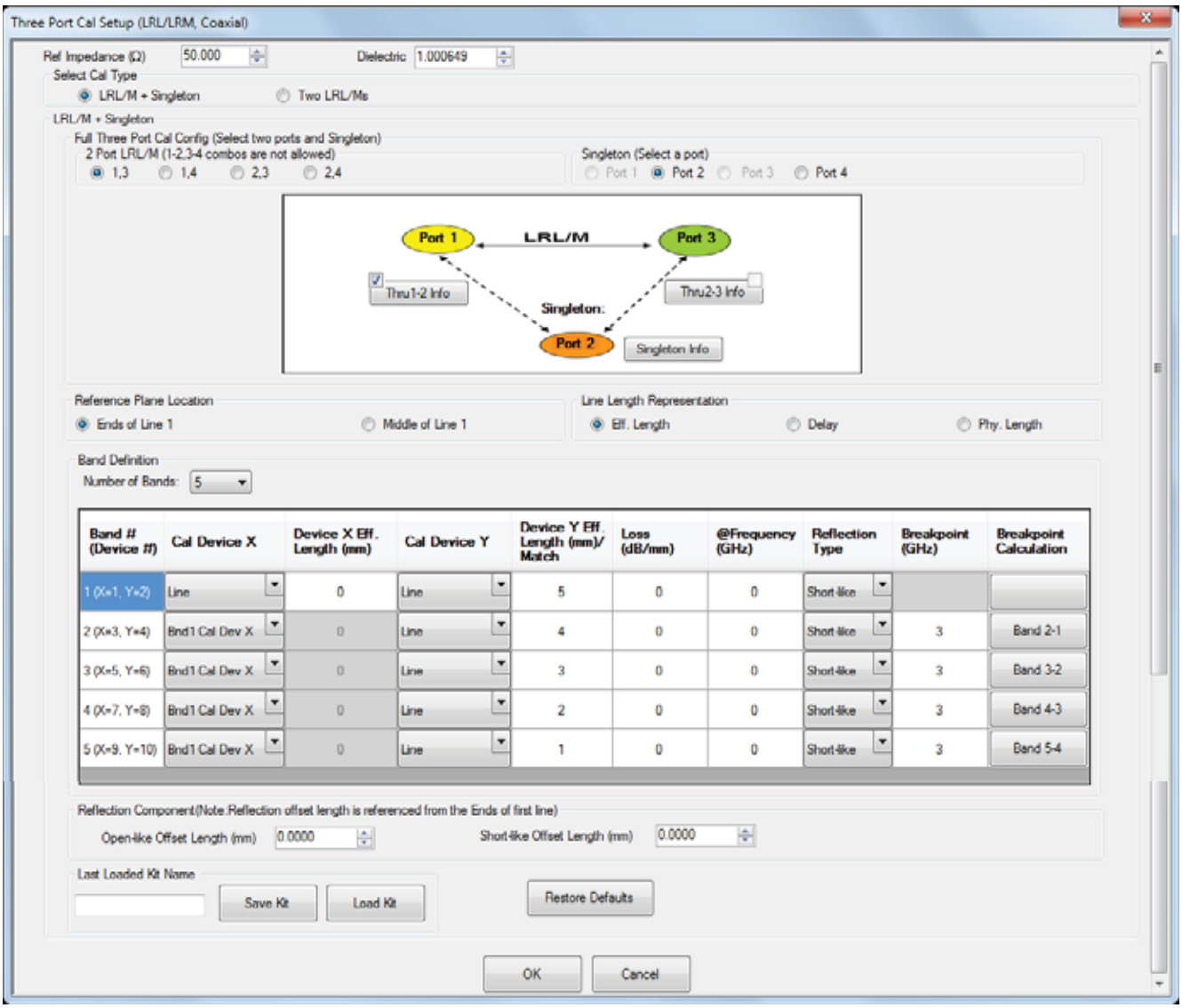


Figure 24-24. THREE PORT CALSETUP (LRL/LRM, COAXIAL) Dialog Box—LRL/M + Singleton

In addition, a single defined standard must be measured at the singleton port defined under the Singleton Info button as shown below (Figure 24-25). The standard is allowed to be an open or a short and the usual offset plus inductance or capacitance polynomials are supported. The measurement of this additional standard allows one to distinguish between load match and source match at the singleton port. Note that in this case, the connections to the singleton port must be thrus (reciprocals are not allowed). The interconnect is used to derive the singleton ports reflectometer behavior so that additional restriction is required.

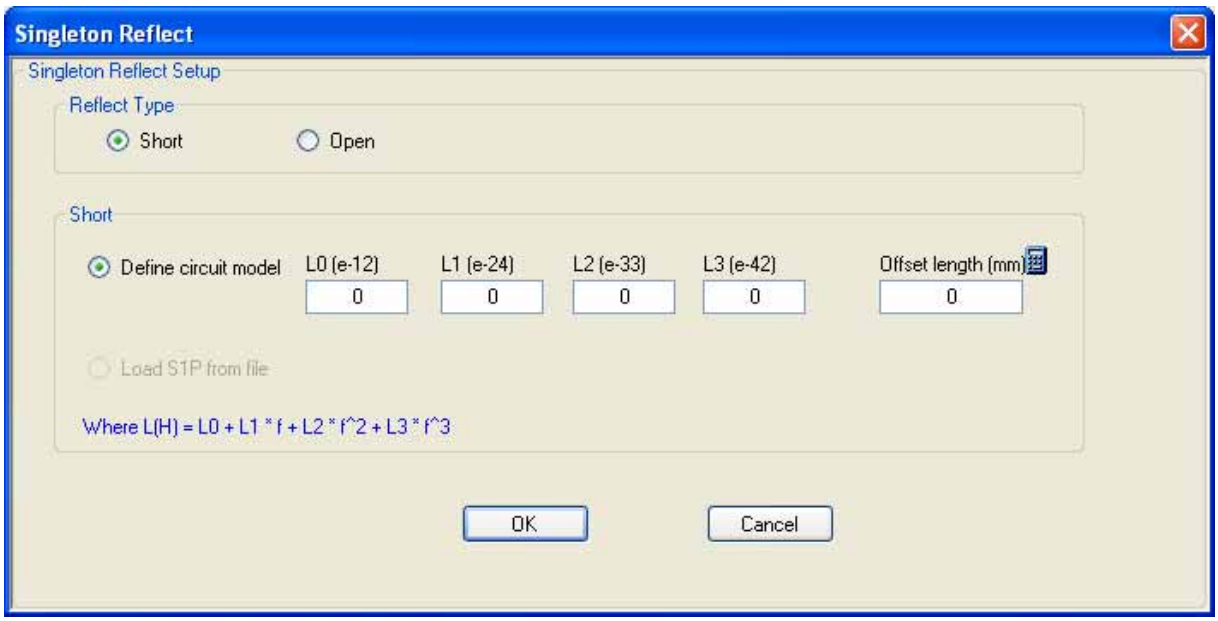


Figure 24-25. SINGLETON REFLECT Dialog Box

The calibration steps are a bit different for this type of calibration and the step menu is shown below (Figure 24-26). The extra singleton button reflects the asymmetric nature of the information required from this port.



Figure 24-26. THREE PORT CAL Menu—LRL/LRM—Coax—One LRL/M + Singleton

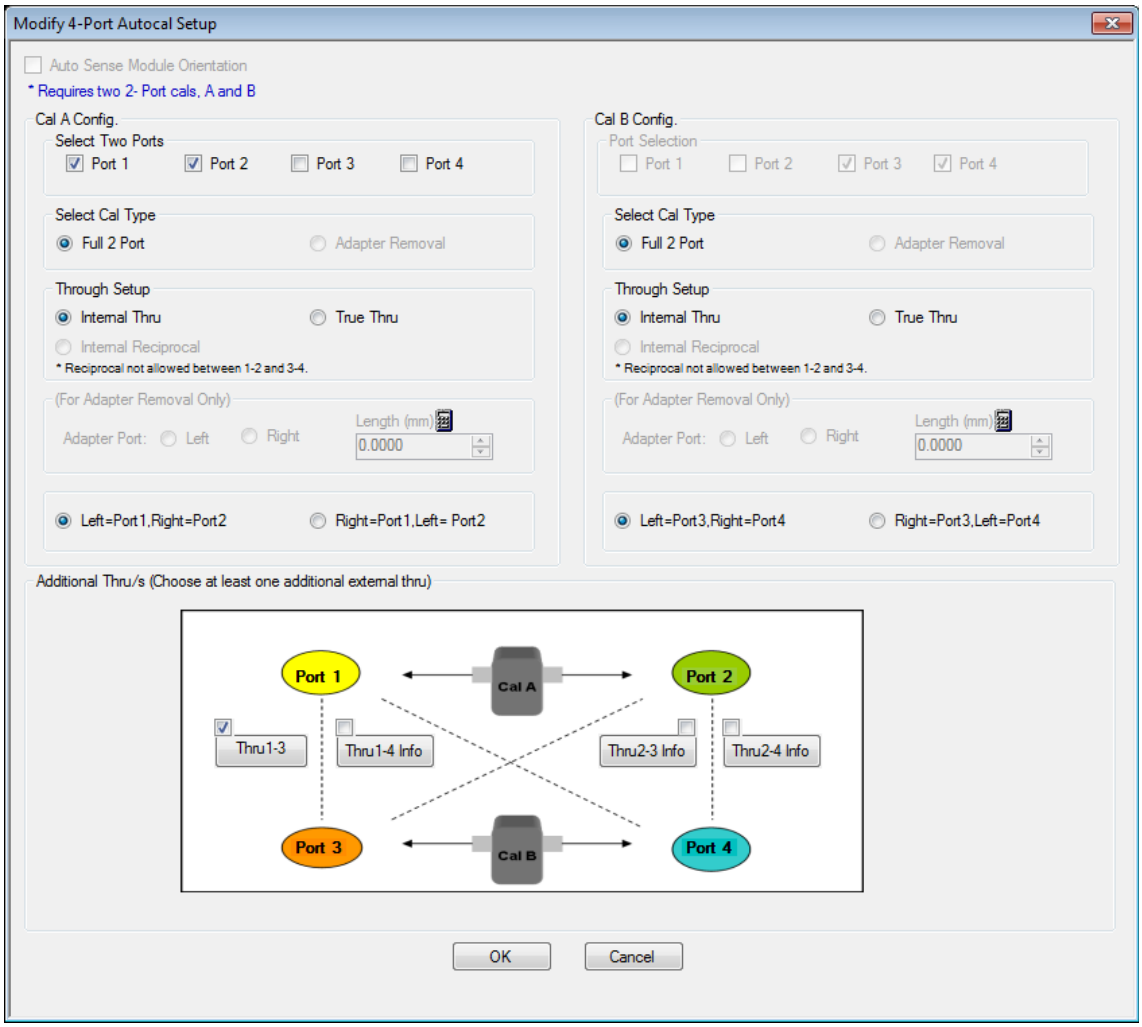
As with the 2-port LRL/LRM case, the three port calibrations can have up to five bands using various combinations of lines and matches to cover the desired frequency range. In the case of Two LRL/Ms, the two sub-calibrations do not need to use the same band definitions nor the same calibration devices. For more details on band and device selection, see Chapter 6, "LRL/LRM Calibration" of this guide.

The multiline TRL calibrations (mTRL; covered in more detail in Chapter 7, "mTRL Calibration") also have logical extensions to 4-port cases. Just as with the regular LRL/LRM calibrations, mTRL calibrations are treated as two-port constructs and the mTRL parameters can be selected independently for the two port pairs. The least-squares analysis (and the modifications for the match standard and super weighting) will occur independently for the two 2-port calibrations. Completion lines are selected in the same way as for LRL/LRM.

24-6 AutoCal Calibrations

The last main calibration category is that of AutoCal. As only 2-port AutoCal exists with this release, the AutoCal may be used to perform a pair of 2-port calibrations followed by a minimum of one additional line connection in order to complete the error model. As with the manual calibrations, additional thrus may be used to improve on the redundant transmission tracking calculations. The usual AutoCal options discussed in above in Chapter 2, "AutoCal Calibration" are still available. The modify setup dialog is shown below (Figure 24-27).

Note	A three-port automatic calibration is not available at this time.

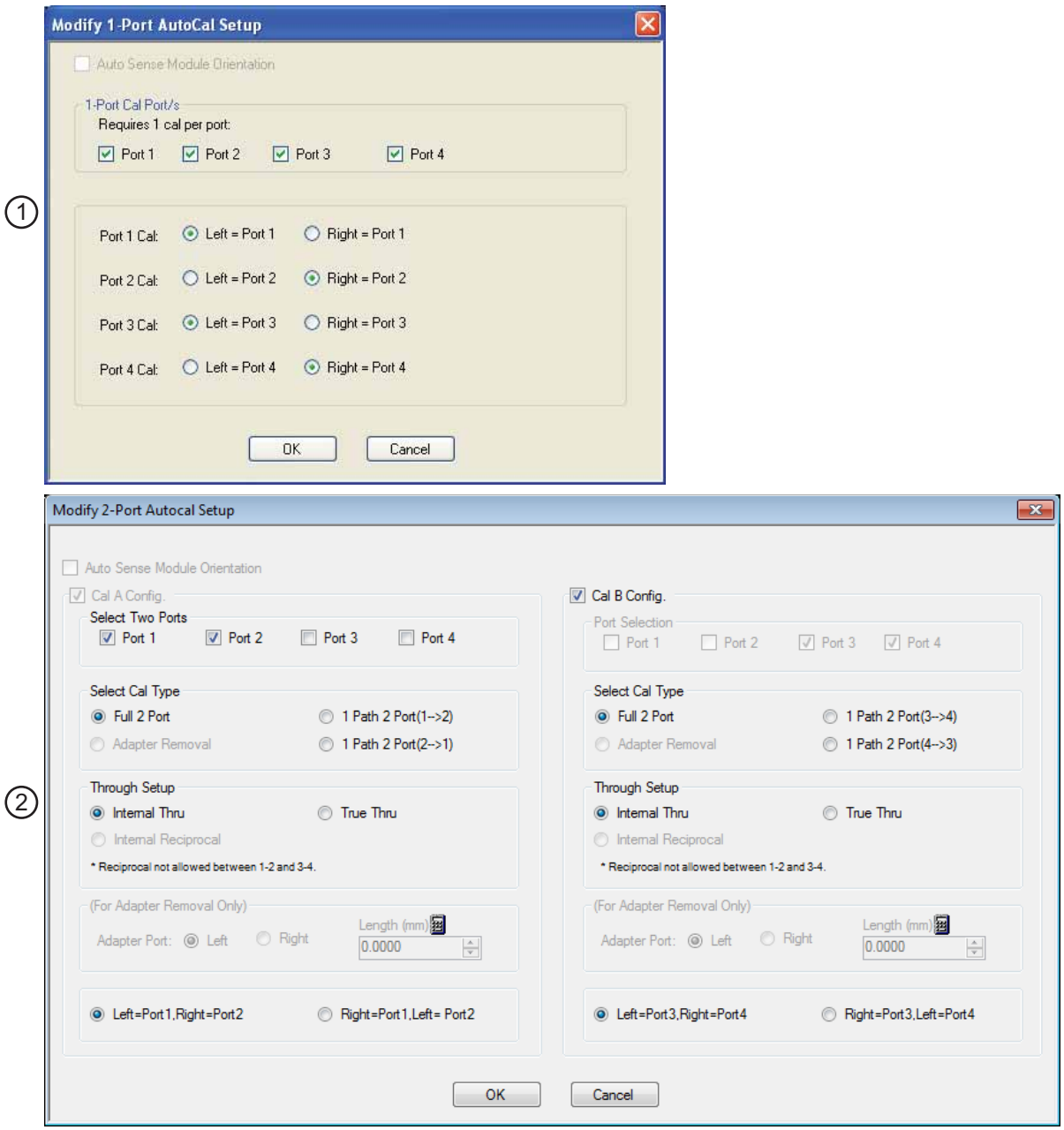


The setup dialog and menu for 4-Port AutoCal are shown here with Thru 1-3 selected.

Figure 24-27. MODIFY 4-PORT AUTOCAL SETUP Dialog Box

Note	The port assignment for Cal B (the second of the AutoCal connections as shown in the figure at upper right) is uniquely defined once the first cal (Cal A) is defined. There is flexibility on how internal/true thrus are allocated and on the use of adapter removal between these calibration steps. Only true-thrus can be used at this time to complete the calibration (those connections not involving the AutoCal itself) and not reciprocals.

At least one additional, physical thru line is required to complete this 4-Port calibration and these selections are shown as a diagram at the bottom of the dialog box (for example, the Thru1-2 or Thru2-4 buttons). It is also allowed to use true thrus during the AutoCal step so it is conceivable there could be anywhere from one to six physical thru connections involved depending on how the calibration is configured. A pair of unlinked 2-Port calibrations based on AutoCal (like that for the other algorithmic approaches) is also possible as shown below (Figure 24-28). The obvious extension continues to include up to four unlinked 1-Port calibrations.



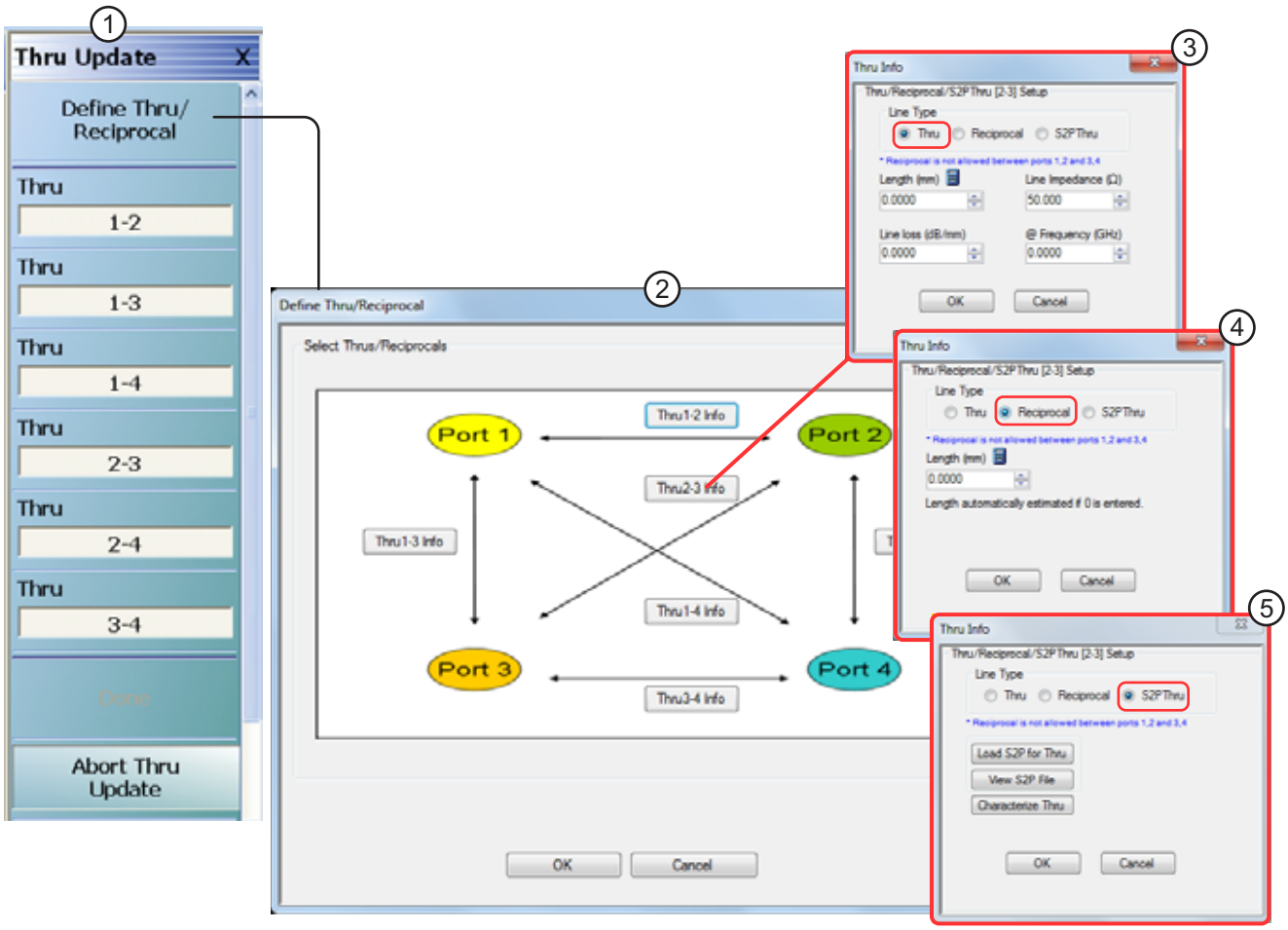
1. The **MODIFY 1-PORT AUTOCAL SETUP** dialog box (at top) for automatic calibrations using the 4-Port system.	2. The similar **MODIFY 2-PORT AUTOCAL SETUP** dialog box (at bottom).

Figure 24-28. MODIFY AUTOCAL SETUP Dialog Boxes

Port assignments and, where appropriate, the internal/true thru selections remain flexible. In the 2-port case, AutoCal adapter removal is still allowed on either calibration if two are selected.

Thru Update

Thru update would, as might be expected, have acquired additional flexibility in the 4-Port system. The menu and related configuration dialogs is shown below (Figure 24-29), with the six possible thrus which, again, may be independently defined. Any number of thrus may be selected in the update but note that only the transmission tracking and load match terms connected to the thru selected will be updated. That is, any redundantly-related terms that may have been derived during the original calibration will not be recomputed. As with the base calibrations, the first-in rule applies. If the first thru selected is between Ports A and B, the load matches for Ports A and B will be computed there. If the second thru selected is between Ports A and C, only the load match for Port C will be computed with that line (since that for Port A was already found). Thus the higher quality (in terms of match and knowledge about the loss) thrus should be measured first.



1. The **THRU UPDATE** menu (at left) is shown here along with the related configuration dialog boxes.
2. The **DEFINE THRU/RECIPROCAL** dialog (at center) allows each thru to be defined in turn.
3. **THRU INFO** dialog box (Thru/Reciprocal/S2P Thru [2-3] Setup) configures each required thru. Many of the thru definition concepts here have been discussed previously.
4. **Reciprocal** selected configures line length.
5. **S2P Thru** selected provides buttons for loading, viewing, and characterization (to generate S2P files).

Figure 24-29. THRU UPDATE Menu and Related Configuration Dialog Boxes

24-7 Uncertainty

The uncertainties in the 4-Port measurement itself are straightforward extensions of those derived from the 2-port model. Associated with each port are residuals due to source and load match, directivity, and reflection tracking. Between every pair of ports is a residual due to transmission tracking. Other terms due to isolation and repeatability are often ignored in the analysis. The only real difference from a 2-port model is that there are multiple load match terms and each is dependent on path isolation (e.g., [7], section III. A). From the point of view of classical error models, one can simulate these effects with an elevated effective load match in many cases.

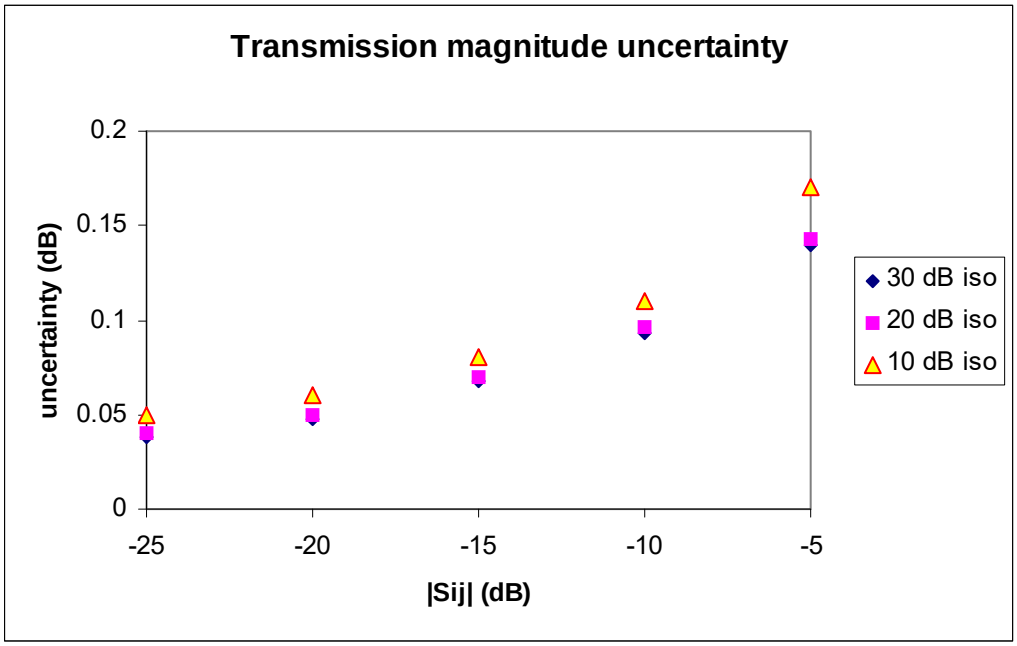
As an example, consider the measurement of S_{11}. The primary factors affecting this measurement (ignoring noise floor and compression effects) will be residual directivity errors (*Δed1*), source match errors (*Δep1S*), reflection tracking (*Δet11*) and load match (*ΔepnL*). This simplified analysis also ignores multiple reflections between a series of ports since they will be at most 2nd order effects.

$$err \approx \Delta ed1 + S_{11} \cdot \Delta et11 + S_{11}^{\,2} \cdot \Delta ep1S + S_{21}S_{12} \cdot \Delta ep2L + S_{31}S_{13} \cdot \Delta ep3L + S_{41}S_{14} \cdot \Delta ep4L$$

Equation 24-14.

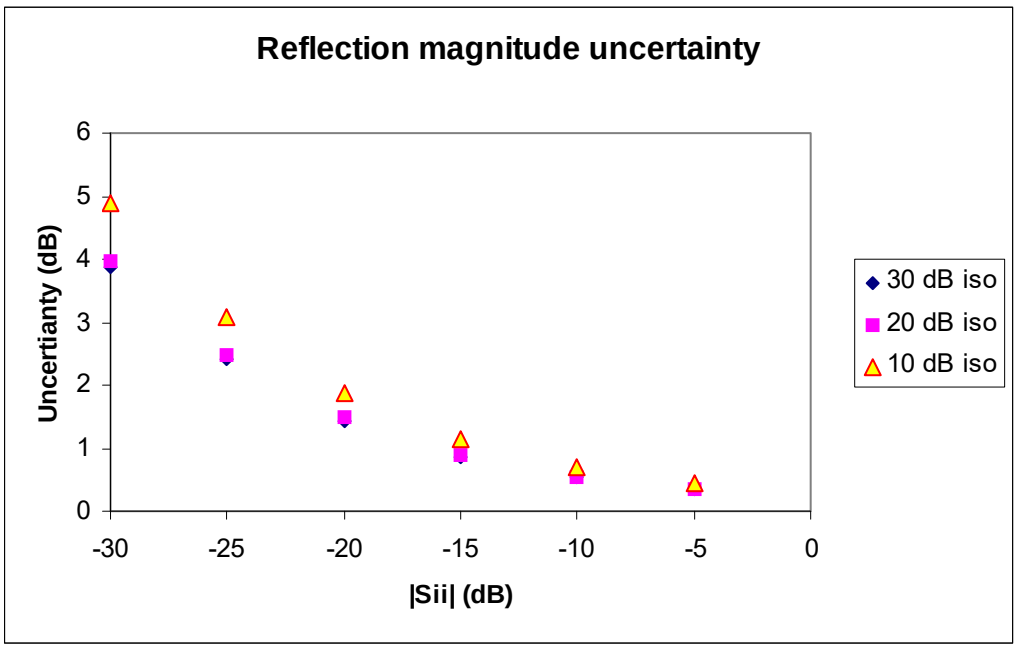
Clearly if some of the transmission terms are small, the load match quality becomes less important as observed before. Some curves are shown below (Figure 24-30 and Figure 24-31) to illustrate this dependence on isolation. These are based on calibration residuals characteristic of a reasonably high quality calibration but the main intent of the curves is to display tendencies. The transmission curve of Figure 24-30 shows little effect until the isolation gets to between 10 and 20 dB. The greatest impact will be for devices of low insertion loss (right side of the curve) as might be expected.

The reflection curves of Figure 24-31 show a similar qualitative dependency on isolation and a worsening of the relative effect as the desired return loss gets larger. Combining the two figures, it is clear that the worst case scenario is a low insertion loss, high return loss path with poor isolation to the remaining ports. This makes intuitive sense since this situation provides a maximum number of interfering standing waves within the device.



The dependence of transmission magnitude uncertainty on the loss in the desired path and the isolation to the other ports is shown here.

Figure 24-30. Transmission Magnitude Uncertainty



The dependence of reflection magnitude uncertainty on the reflection at the desired port and the isolation to the other ports is shown here.

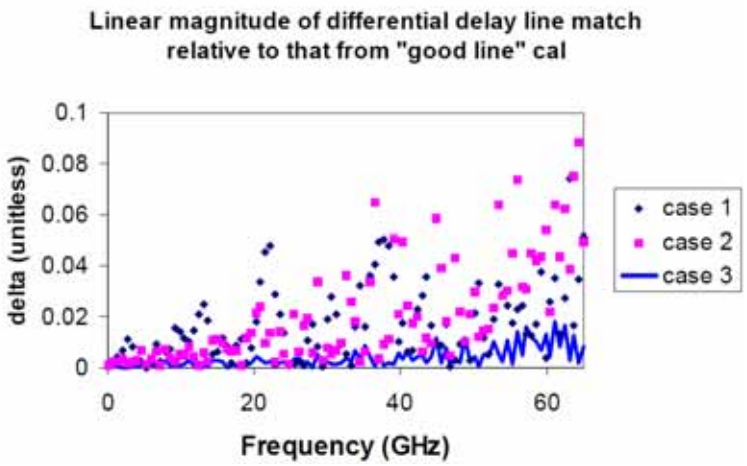
Figure 24-31. Reflection Magnitude Uncertainty

The exact configuration of the thru selections will have some impact on uncertainty as well. If one uses redundancy heavily, some of the previous calibration results (which have there own uncertainty) are used to compute other transmission tracking terms (thus elevating their uncertainties). Thus there is something of an uncertainty penalty in taking advantage of redundancy to reduce calibration time. Depending on the nature of the available thrus between some ports (electromagnetically questionable), however, there may be even more of a penalty in trying to use all of the thrus if they are of low quality. If the calibration is of reasonably high quality, the transmission tracking penalty in using redundancy will normally not exceed a few hundredths of a dB but each calibration situation can be different.

The heavy use of reciprocals in lieu of thrus can also affect uncertainties since the accuracy on load match acquisition is lower using reciprocals. Again, trying to treat the networks as thrus may make it even worse so there is an additional trade-off in play. If the reciprocal loss is under 5–10 dB and its return loss is better than about 10 dB, the uncertainty penalty will be small except for measurements of very low loss, high return loss devices (which if they existed, why were they not used as the reciprocals?).

The plot below (Figure 24-32) illustrates some of these complex trade-offs. The DUT in all cases is a differential delay line and we are interested in measuring its return loss. There are two high quality thrus available during the cal (return loss >30 dB) but the other available lines have a 15 dB return loss. The error relative to the correct return loss (in linear units) is being plotted.

- **Case 1 (not done)**—Use the poor line to establish load match and treat it as a thru.
- **Case 2 (can be done)**—Use the good lines to get load match but treat the poor line as a thru to get the remaining transmission tracking terms
- **Case 3 (preferred)**: Treat the poor line as a reciprocal.



An example of the effects of thru and reciprocal selection on measurement uncertainties is shown here.

Figure 24-32. Measurement Uncertainties for Thru and Reciprocal Selection

In this case, using a reciprocal gives a far better result than trying to force a “thru” utilization.

Uncertainty of Derived Mixed-Mode S-Parameters

Uncertainty of derived mixed-mode S-parameters is another issue of interest. Since the mixed mode terms are simple linear combinations of the normal S-parameters, the uncertainty in a mixed-mode parameter is easy to compute:

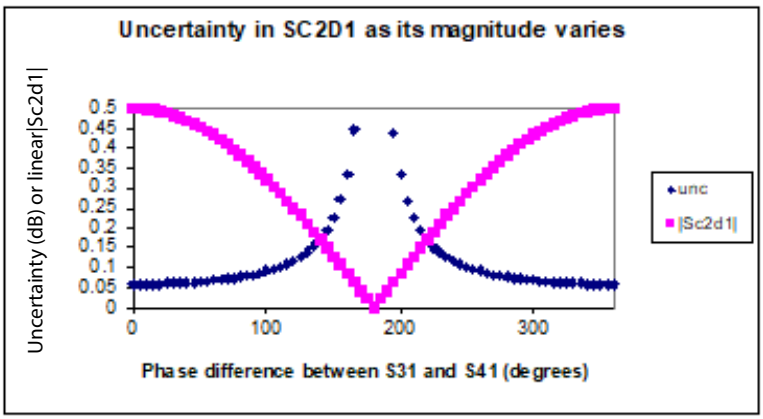
$$S_{c2d1} = \frac{1}{2}(S_{31} + S_{41} - S_{32} - S_{42})$$

Equation 24-15.

$$\frac{\Delta S_{c2d1}}{S_{c2d1}} \leq \frac{|\Delta S_{31}| + |\Delta S_{41}| + |\Delta S_{32}| + |\Delta S_{42})|}{S_{31} + S_{41} - S_{32} - S_{42}}$$

Equation 24-16.

Where linear absolute magnitudes can be used for a worst case computation. The caveat here is that this is in linear terms while uncertainties are often more useful in a log form. This has a surprising effect for some parameters. As an example, consider the case where the single-ended uncertainties are fixed at 0.03 dB, S_{32} and S_{42} are zero and S_{31} and S_{41} are allowed to vary. For the purposes of an example, the magnitude of S_{31} and S_{41} will be fixed at 0.5 and the phase will be allowed to vary. As one can see below (Figure 24-33), when the phase of the single ended parameters is such that $|S_{C2D1}|$ is small, the uncertainty (in dB terms) explodes. This follows from Eq. 24-10 since the numerator is largely fixed but the denominator can get small.



The uncertainty (in dB) in a mixed-mode parameter is a strong function of the magnitude of that parameter because of how they are computed.

Figure 24-33. SC2D1 Uncertainty in SC2D1

Since the desired quantity (S_{c2d1} in this case) is very small, the relative uncertainty can be quite high. This is somewhat intuitive since the computation is subtracting nearly equal numbers. The user is therefore cautioned that cal stability will be extremely critical on the smaller mixed mode parameters and the uncertainties will be relatively high. The biggest source of problems will often be cable changes with flexure or connector repeatability.

24-8 Hybrid Calibrations

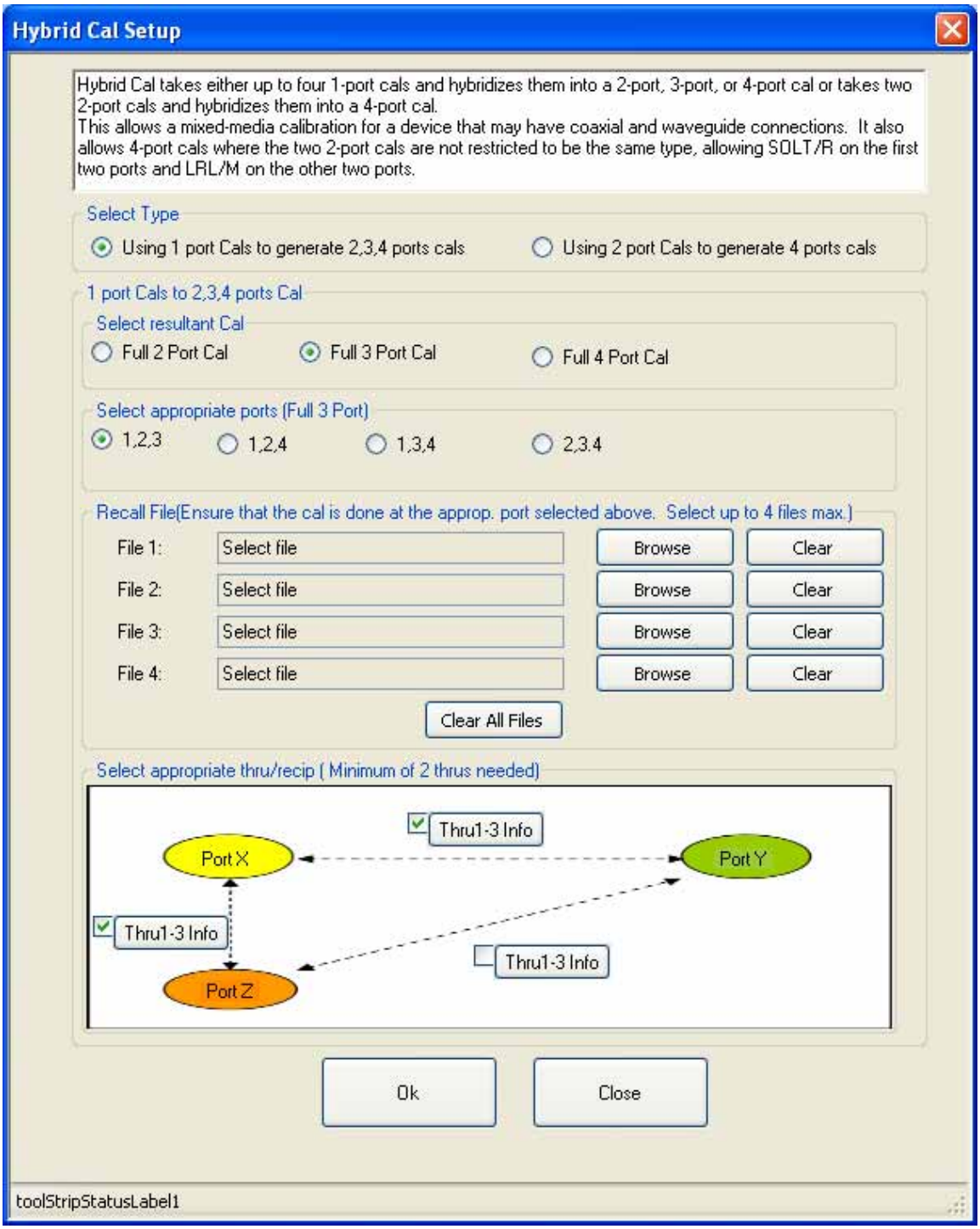
The concept of hybrid calibrations was introduced in Chapter 10, "Other Calibration Procedures" in the "Hybrid Calibrations" section for the 2-Port VNA but it takes on more importance in the multiport case. This is largely because of the geometrical complications in a multiport setup (particularly on-wafer or fixtured) that demand more flexibility in cal algorithm selection.

There are two main subcategories in the multiport case: the combination of 1-Port calibrations and the combination of 2-Port calibrations. In both subcategories, the frequency lists used in the constituent calibrations must be the same.

Combination of 1-Port Calibrations

The first subcategory is illustrated in the dialog below (Figure 24-34). The idea is a natural extension of what was discussed in Chapter 10, "Other Calibration Procedures" for the 2-Port case. A series of 1-Port calibrations are combined with a requisite number of thrus or reciprocals to generate a complete N port cal. The choices in the dialog are fairly obvious. Like the multiport calibrations discussed earlier, there is a minimum coverage requirement for the thrus/reciprocals. At least two thrus are needed for a 3-Port cal and at least three thrus are needed for 4-Port cal (and they must touch all ports). Any or all of these "thrus" may actually be reciprocals and this is commonly of interest. Consider the case of wishing to create a 3-Port cal from a series of one port calibrations on-wafer where Ports 1 and 2 are coplanar waveguide (with corresponding cal standards) with a high quality interconnecting thru and Port 3 is microstrip (with a more poorly defined "thru" linking the other ports). The 1-Port calibrations for Ports 1 and 2 would be performed using a non-dispersive (or coaxial) media type while the cal for Port 3 would likely be performed using the microstrip media type. In completing the hybrid cal, one would likely select a thru to connect Ports 1 and 2 and a reciprocal to connect Ports 1 and 3 (or 2 and 3).

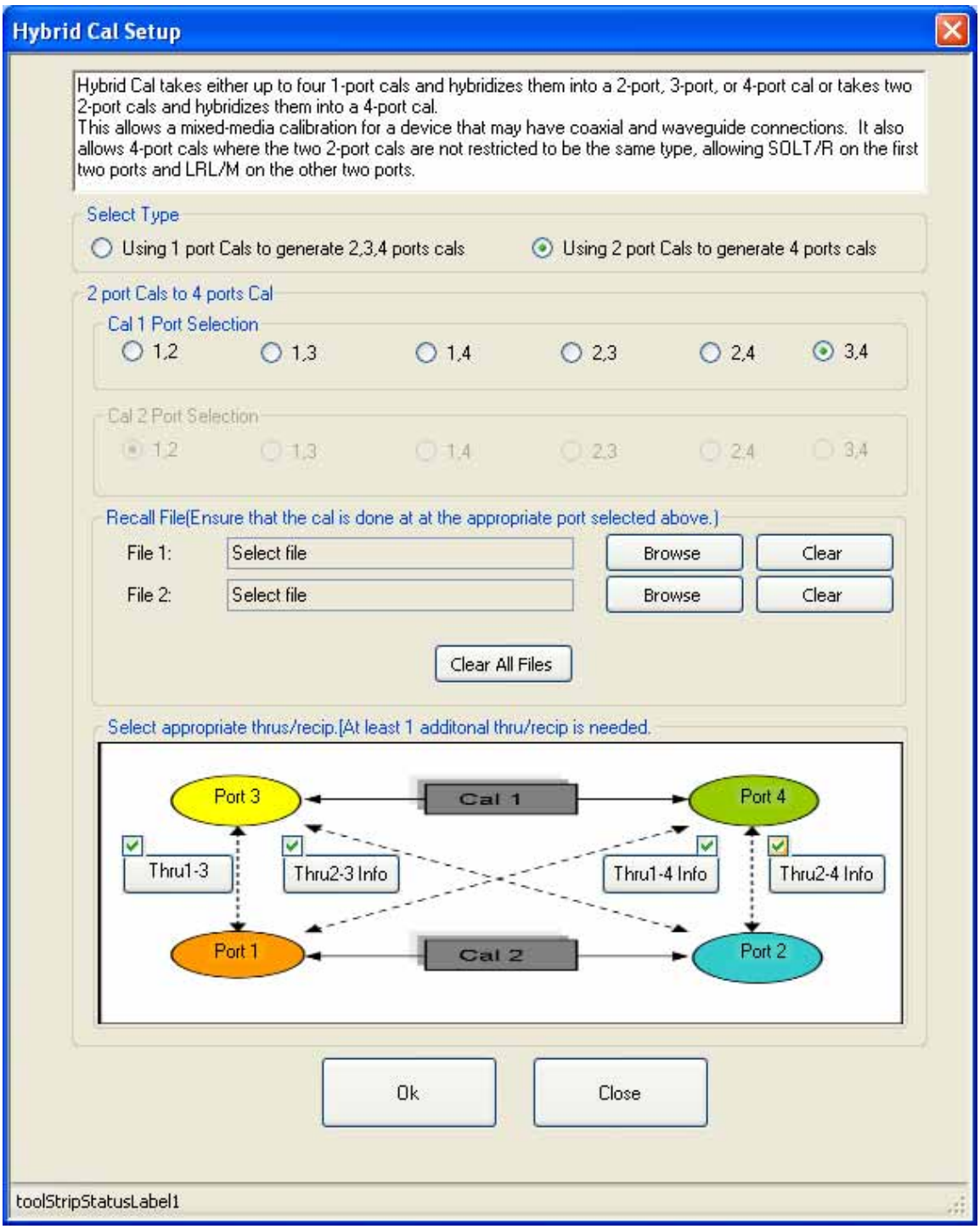
As always, there is a slight uncertainty penalty on transmission tracking for those paths not measured with thrus or reciprocals. For high quality calibrations, this is typically on the order of a few hundredths of a dB.



The Hybrid CAL SETUP dialog box using 1-port Cals to generate 2-, 3-, and 4-Ports cals—Full 3-port Cal—Thru for Ports 1-3.

Figure 24-34. HYBRID CAL SETUP Dialog Box—Using 1-port Cals to Generate 2-, 3-, or 4-port Cals

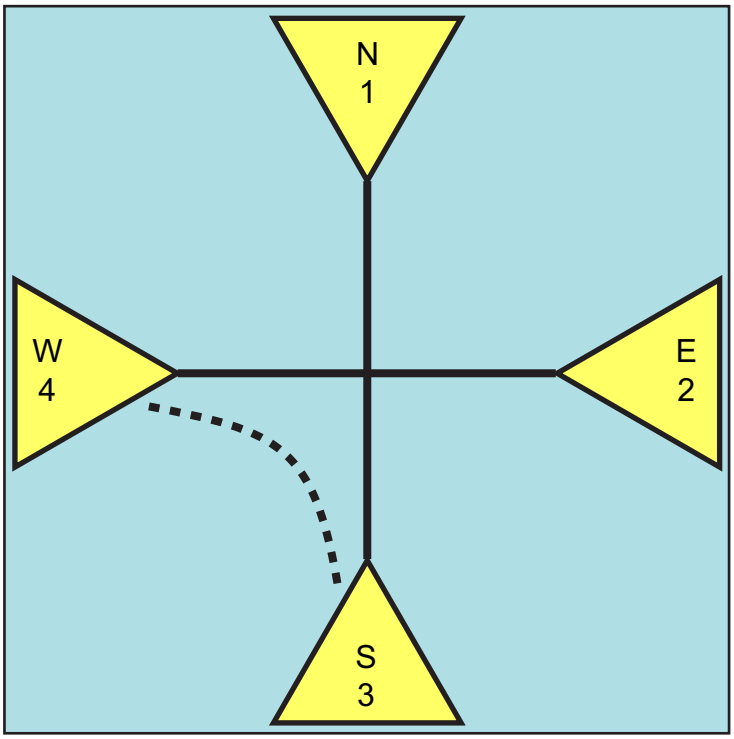
Two 2-Port calibrations can also be combined to create a full 4-Port calibration and this process is illustrated in the dialog shown below. The two 2-Port calibrations used must be disjoint (i.e., they involve different port pairs). The dialog forces this since one only actually selects the first pair. Once the constituent calibrations are selected, at least one additional thru or reciprocal must also be added. As usual, adding additional thrus (if practical) can aid certain transmission measurements.



The Hybrid CAL SETUP dialog box using 2-port Cals to generate 4-Port cals—Thrus for ports 1-3, 2-3, 1-4, and 2-4 selected

Figure 24-35. HYBRID CAL SETUP Dialog Box—Using 2-port Cals to Generate a 4-port Cals

An example where this might play a role would be the orthogonal probe setup discussed earlier and shown again below (Figure 24-36). Suppose N-S represents a clean path where LRL standards are available while E-W represents a clean path where only SOLT standards are available. Further, the corner paths (e.g., NW) have transmission lines available for calibration but they are of fairly poor quality. The two separate 2-Port calibrations would be performed with the algorithms identified. They would then be combined using the hybrid cal feature with at least one of the corner paths used as a reciprocal.

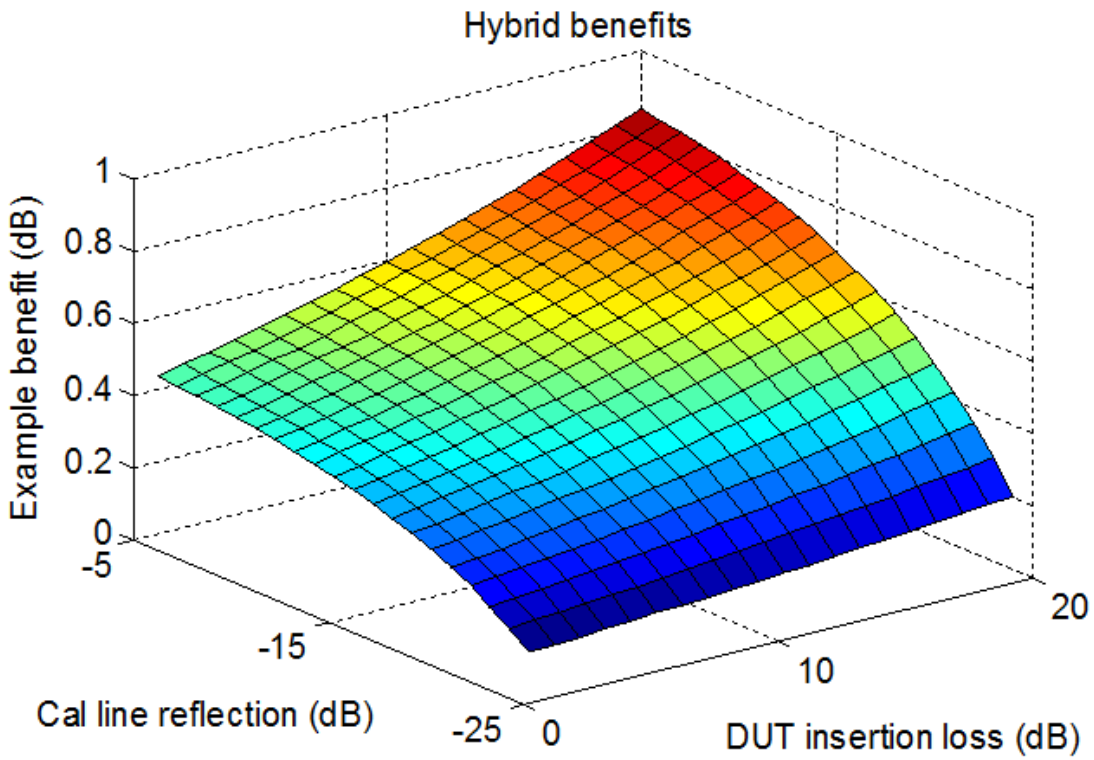


An on-wafer configuration where some line paths are more likely to be of higher quality than others. The solid lines will generally not be an issue but a transmission path illustrated by the dotted line (S-W) may be electromagnetically compromised at higher frequencies.

Figure 24-36. On-Wafer Line Path Quality

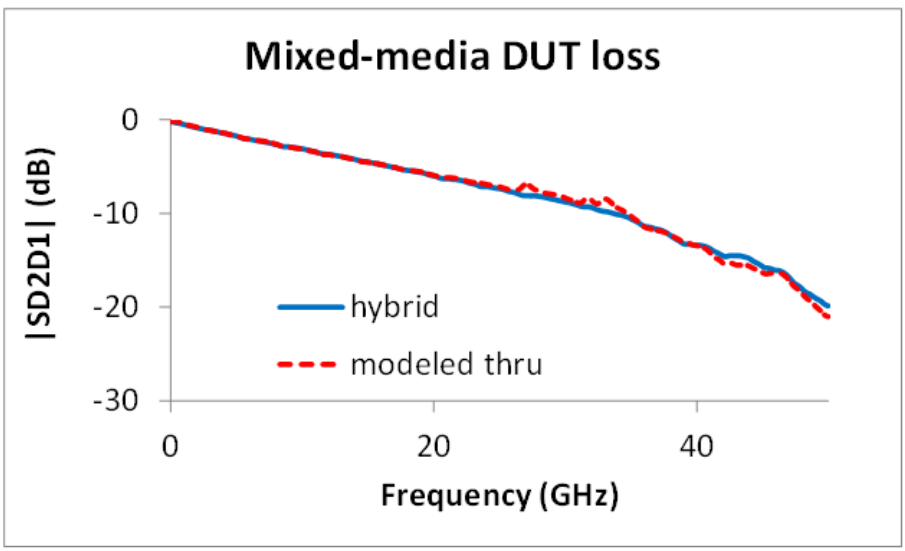
As with the 2-port hybrid cases discussed in an earlier chapter, the benefits in the multiport case depend on the media of the underlying calibrations and what interconnects are possible. A prototype case is the TRL/LRL family where some lines are good and some are not. The benefit in treating the not-as-good lines as reciprocals using the hybrid approach is calculated for an example system in Figure 24-37. The important metric here is the 'cal line reflection' which represents how well-matched the not-as-good lines are. The DUT insertion loss scale axis is perhaps misleading as a 0.5 dB benefit has considerably more value for a 2 dB insertion loss measurement than it does for a 20 dB insertion loss measurement (usually). The central point is that this hybrid has the most value for cases when some of the lines are poorly matched and when the DUT has relatively low insertion loss.

A practical example measurement is shown in Figure 24-38 where the third calibration line (TRL calibrations) had about 10–15 dB return loss at the higher frequencies. As the line reflections became worse at higher frequencies, the hybrid calibration benefits increased as one might expect.



The transmission magnitude uncertainty benefit from the hybrid approach for various reflection coefficients of the calibration lines and various DUT insertion losses is shown here.

Figure 24-37. Transmission Magnitude Uncertainty Benefit from the Hybrid Approach



A more practical measurement showing the potential benefits of a hybrid calibration is shown here. One of the calibration lines, because of its required geometry, had degrading match at higher frequencies.

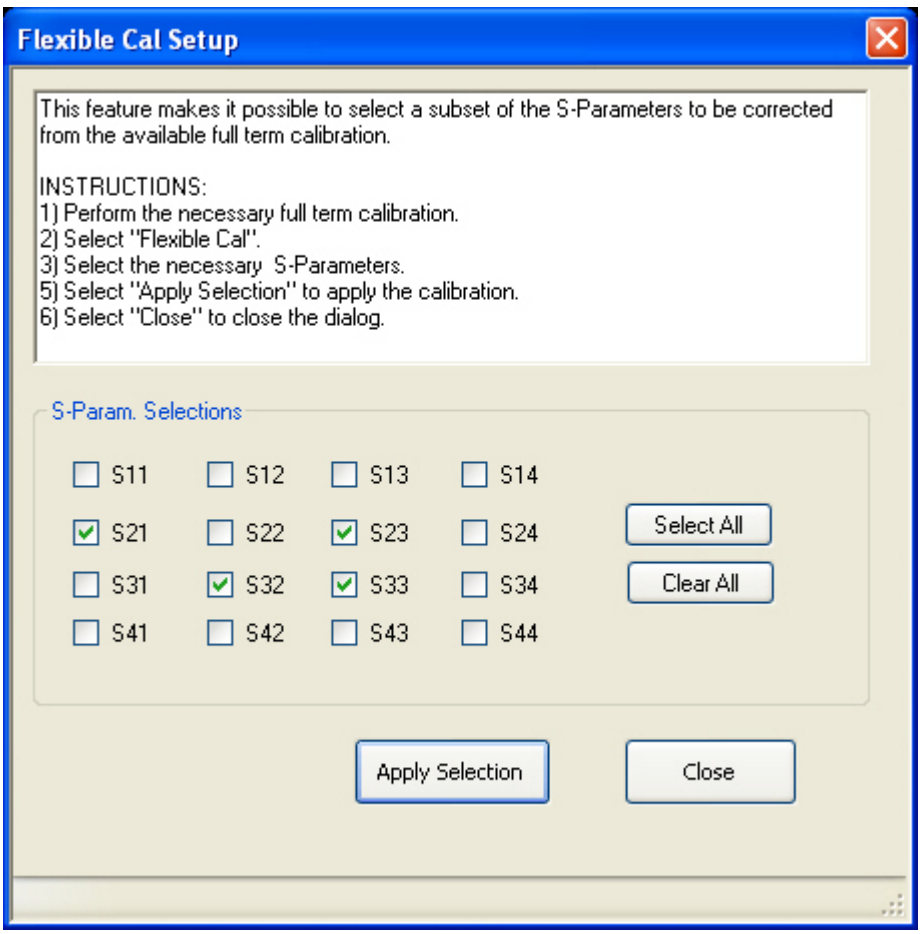
Figure 24-38. Practical Measurement Showing the Potential Benefits of a Hybrid Calibration

24-9 Flexible Calibrations

The flexible calibration implementation for multiport is a straightforward extension of that discussed in Chapter 10, "Other Calibration Procedures" for 2-port systems. The dialog shown below illustrates the 16 possible S-parameters when the parent cal is a full 4-port cal. As discussed before, the objective is to select only those S-parameters one needs for a given measurement and those that may significantly help the uncertainty of the needed measurement. As before, the system will only do the sweeps required to get the requested S-parameters and will do as complete a cal as possible based on those parameters.

Examples

- **Select S_{33} and S_{43}**
- Sweep only with Port 3 driving and use a 1-Path 2-Port cal for the two parameters
- **Select S_{11}, S_{13}, S_{31}, S_{33}, S_{42}, and S_{24}**
- All four ports will have to drive. Use a full 2-port cal between Ports 1 and 3 and a transmission frequency response cal between Ports 2 and 4
- **Select S_{11}, S_{22}, S_{14} and S_{41}**
- Ports 1, 2, and 4 will have to drive. If 1-4 is a low loss path, one may want to also add S_{44} as this will enable load match correction and could help uncertainties of S_{14} and S_{41} (it would be of less use if 1-4 had large insertion loss). If S_{44} was added, a full 2-port cal between 1 and 4 would be used along with a 1-port reflection-only cal at Port 2.



Note that the items enabled for selection are based on the last calibration performed. Also note that there is only the customize selections since the UI can get complicated with Cal A and Cal B.

Figure 24-39. FLEXIBLE CAL SETUP Dialog Box for Multiport VNAs

24-10 Network Extraction

Following onto the discussion of Chapter 9, "Adapter Removal Calibrations and Network Extraction", there are additional network extraction techniques available to help find the networks for multiport de-embedding problems. The new techniques are labeled Types E, F, and G which apply for 4-Port calibrations. The original types (Types A to D) apply for 1- and 2-port configurations as before.

- **Type E Network Extraction**

  Uses a pair of full 4-port calibrations to fully extract four S2P files describing the arms of the adapter/fixture assembly. This is a complete solution but assumes the arms of the assembly are not coupled together. This is a 4-port extension of Type C.

- **Type F Network Extraction**

  This is a 4-port back-to-back method where four S2P files are extracted and the four arms of the adapter/fixture assembly are assumed uncoupled. As with Type D (the 2-port equivalent), match is assigned to the outer planes. Port 2 of the .s2p files is assigned to the port nearer the DUT as is consistent with the de-embedding system operation.

- **Type G Network Extraction**

  This is a 4-Port back-to-back method where two S4P files are generated and the sides are assumed coupled (in a half-leaky sense). Measured cross-coupling is assigned to the outer planes. Port assignments on type G can be complex and the examples in the text should be noted.

- **Universal Fixture Extraction**

  With Universal Fixture Extraction (available with Option 21 installed, and discussed in Chapter 9 of this guide), additional choices are available for Types F and G:

  - **Multiple Standards** can be used (two lines, one line plus reflect standards, two lines plus reflect standards) instead of using just the single thru line in traditional F and G.
  - The use of **Phase Localization** (with either a line standard or a reflection standard set) to get better estimates of network parameters based on the phase slope of the measured data.
  - **Phase-localized High Crosstalk** is a special case of type G processing using the reflect standards, where the effects of a neighboring pair (with a high level of crosstalk between the primary pair and the neighbor) are handled more completely.

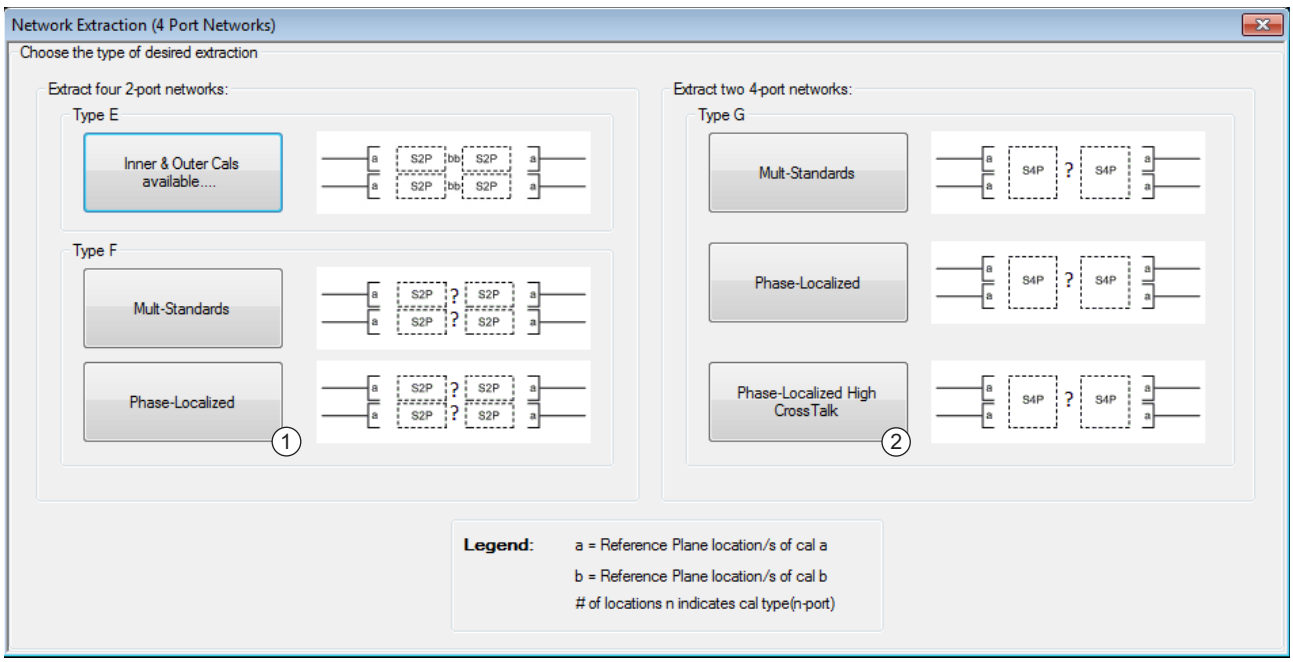


Figure 24-40. NETWORK EXTRACTION Dialog—Type Selection—4-Port VNAs (With Option 21 Installed) (1 of 2)

1. Type F Phase Localized (only available with Option 21 Universal Fixture Extraction)
2. Type G Phase Localized (only available with Option 21 Universal Fixture Extraction)

Figure 24-40. NETWORK EXTRACTION Dialog—Type Selection—4-Port VNAs (With Option 21 Installed) (2 of 2)

- **Sequential Extraction—Peeling (with Option 21)**

  This method of network extraction involves modeling the network as a collection of lumped elements. For details on this method, refer to "Sequential Extraction—Peeling (with Option 21)" in Chapter 9.

Type A Network Extraction

Before continuing with the new network extraction types, note that in the case of Type A, there is the additional requirement to specify where the adapter is (in the case of a multiport calibration). The added radio button selections are shown in the dialog below.

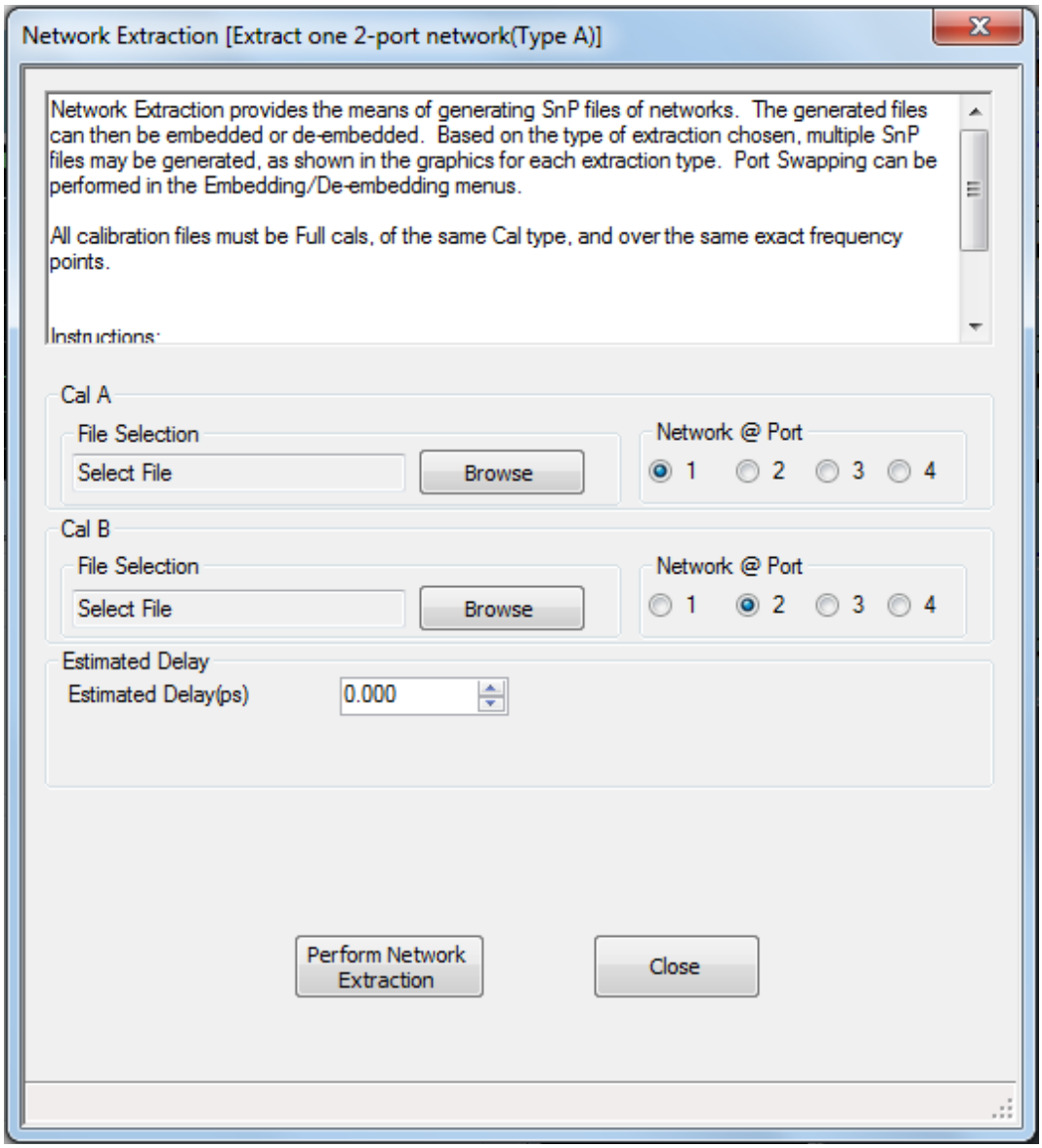


Figure 24-41. NETWORK EXTRACTION TYPE A Dialog Box

Type E Network Extraction

In terms of execution, Type E is very much like Type C. The two full calibration files must be specified (in this case full 4-Port calibrations). Upon execution, a dialog will appear allowing one to name the four S2P destination files. The files will be listed in order of absolute port number.

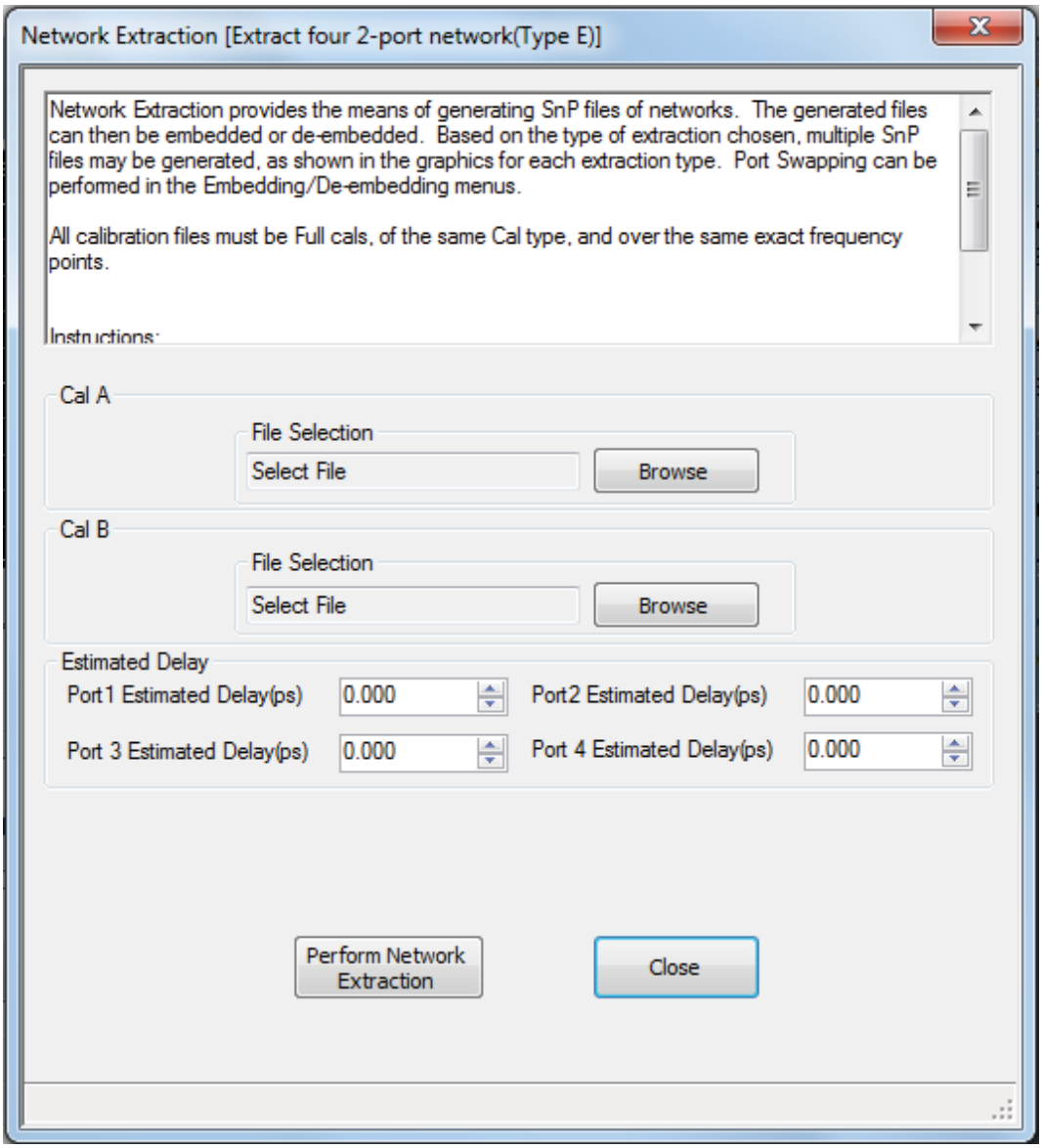


Figure 24-42. NETWORK EXTRACTION TYPE E Dialog Box

As suggested by the overall network extraction dialog, the Type E method treats the fixture as having four, uncoupled, 2-Port arms as shown below (Figure 24-43). These networks are then extracted as S2P files obviously. The required level of "uncoupled-ness" depends on expected uncertainties and other losses in the networks. If the main paths were of very low loss and there was about 40 dB of coupling between arms of the fixture, there could be an added uncertainty of about 0.1 dB from ignoring that coupling with this method. If the coupling was actually 20 dB, there could be ≅ 1 dB of added uncertainty.

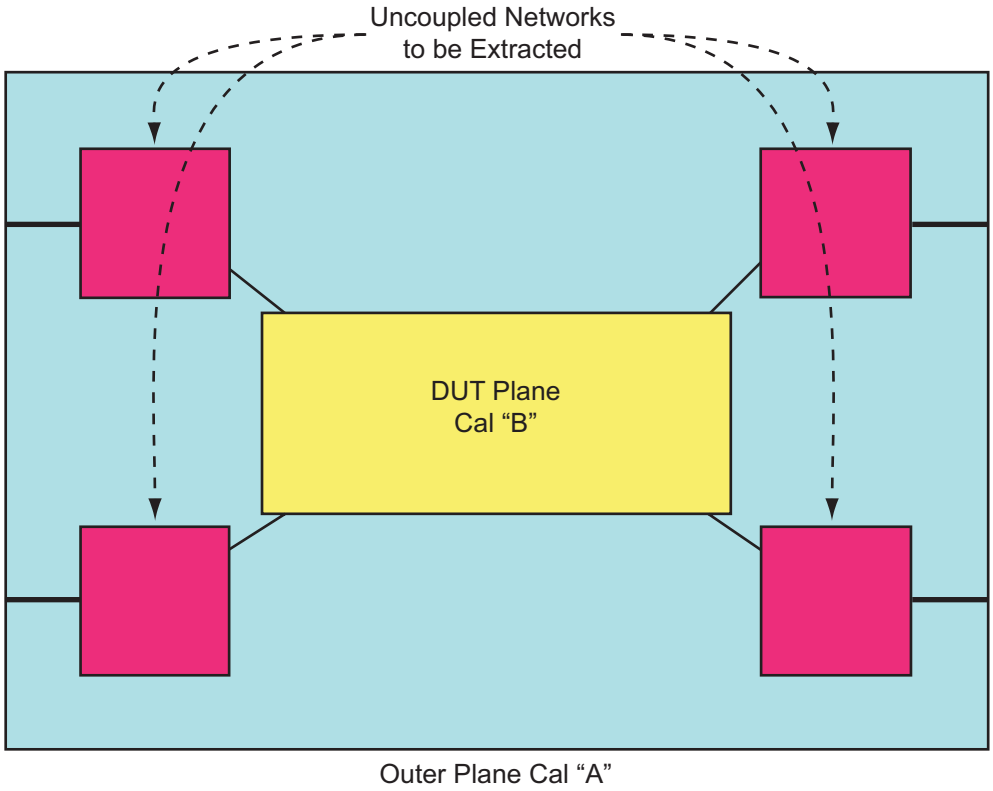


Figure 24-43. Type E Network Extraction—Uncoupled Networks To Be Extracted

Type F Network Extraction—without Option 21

Types F and G operate much like Type D except a full 4-Port cal should be applied upon entering the dialog instead of a full 2-Port cal. The usual file definition dialog will follow. As with Type D, the check box option allows one to neglect fixture match terms instead of assigning mismatch to the outer plane.

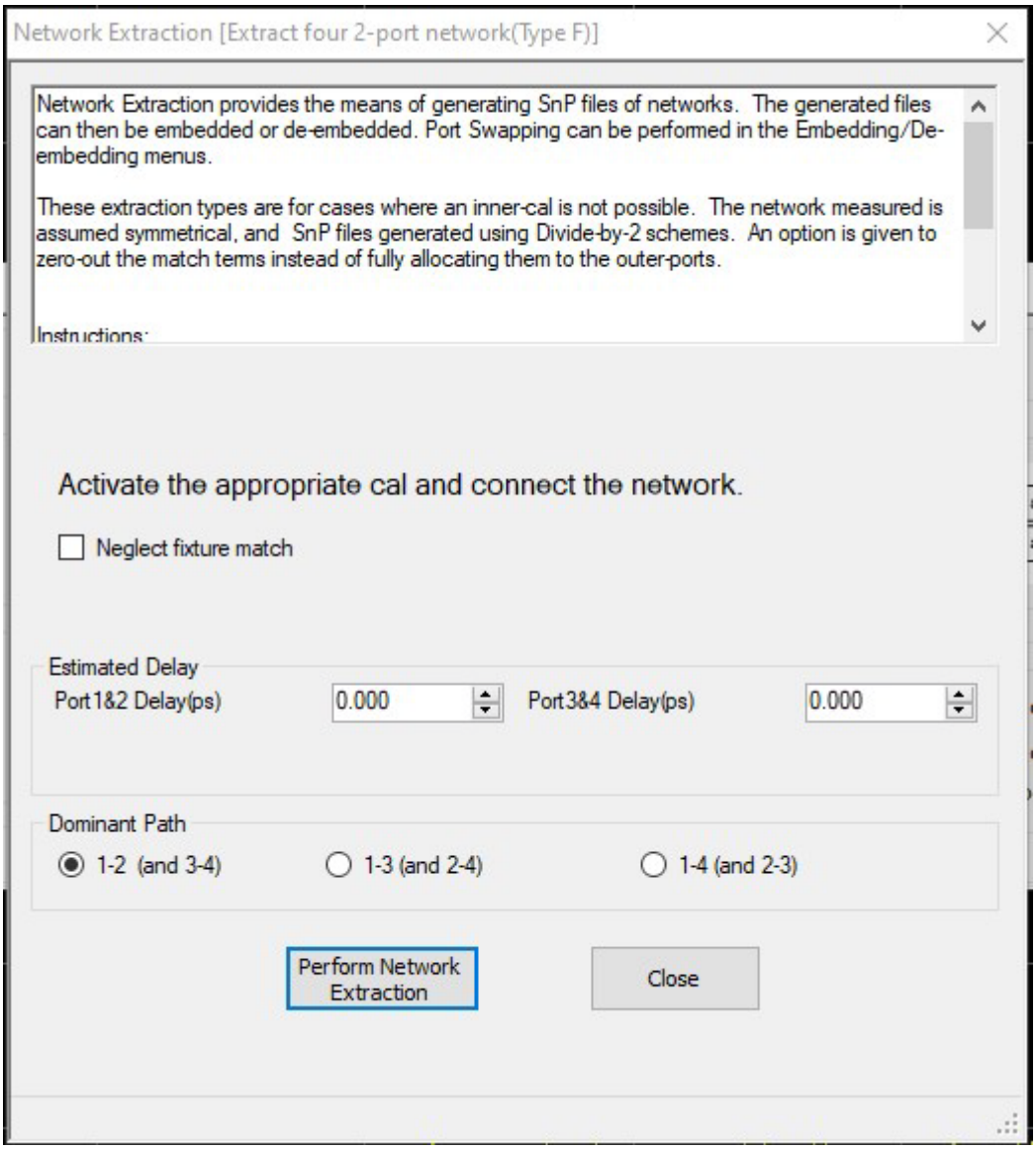


Figure 24-44. NETWORK EXTRACTION TYPE F Dialog Box—Without Option 21

In the case of Type F, the networks are again assumed uncoupled. It is a method appropriate for the same situations as Type E except when inner calibration standards are not feasible or of reliable uncertainty. The dominant path of the fixture connection should be selected as shown in the dialog. Type F assumes the fixture looks like either independent single-ended lines or a weakly coupled differential pair. The 'dominant' path describes what ports are connected during the 'thru' step or, equivalently, what the low insertion loss paths through the fixture are.

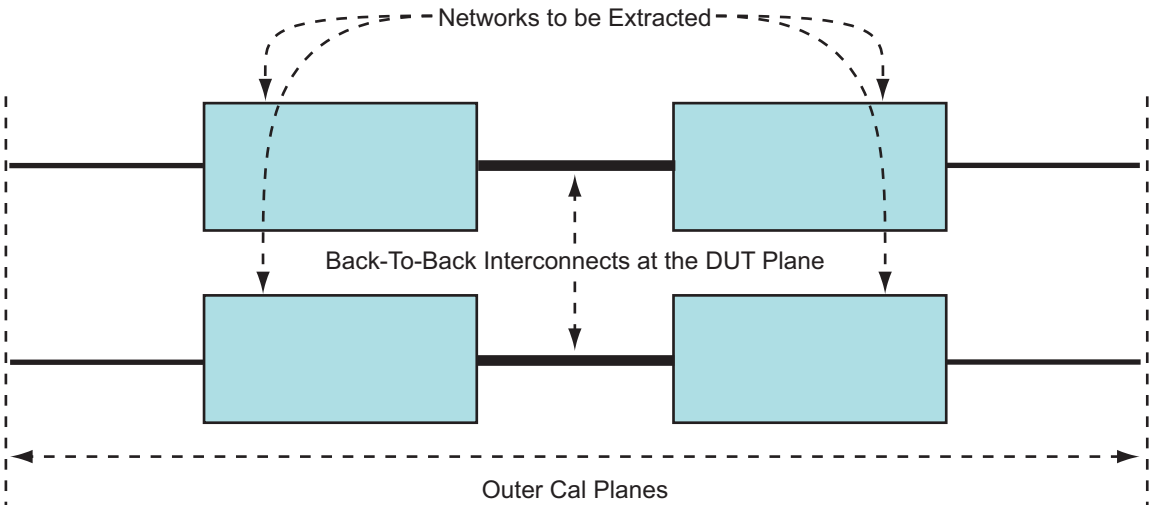


Figure 24-45. Type F Network Extraction—Back-to-Back Interconnects at the DUT Plane—Type F

Type F Network Extraction (with Option 21—Universal Fixture Extraction)

With the *universal fixture extraction* option (Option 21), additional Type F flexibility appears, including the use of multiple standards or phase localization as discussed for Type D previously in Chapter 9 of this guide. Of course, four files will be generated in this case (all .s2p) and, as stated earlier, it will be assumed any coupling between arms of the fixture is very weak. The non-Option -21 version of Type F differs in that only a single line standard can be used and its length must be 0.

The multi-standard concepts discussed in Chapter 9 for Type D do apply here, but to summarize (see also Figure 24-46 on page 24-57):

- One line standard is always required. Using only one line will set inner plane match terms always to 0 (for all four files in this case) and side-to-side symmetry (in terms of insertion loss not outer plane match) will be assumed. The line standard must consist of paths connecting both port pairs (if 1-2(3-4) is selected as the dominant path, the line standard must connect port 1 to port 2 and must connect port 3 to port 4; see Figure 24-47 on page 24-57).
- With two lines as standards, symmetry is still assumed but inner plane match will be solved for. Note that the two line lengths cannot be the same and when the phase difference between the two line lengths is 180 degrees, a singularity will occur. As an example, if line 1 is 0 and line 2 is 33 ps long (10 mm air-equivalent), a singularity will occur at 15 GHz. It is recommended to keep the excursion under 160 degrees so, for this example, 13.3 GHz would be the maximum recommended frequency. Note that at DC, there is another singularity.
- With a line and a reflect standard, symmetry is no longer forced and a coarse estimate of inner plane match is possible. The same reflect standard must be applied to all four ports during this measurement. The reflection coefficient of the standard (in magnitude and offset length) must be specified and the magnitude is allowed to be negative to allow for easy entry of a short circuit standard. Loss for line and reflect standards can be entered in dB/mm terms (that scales as sqrt(f/fref), where fref is the entered reference frequency; if fref=0, the loss is assumed to be constant with frequency).
- Using all three standard types maximizes the information possible and symmetry is not forced and a better inner plane match estimate is possible.
- The path fixture delays need not be the same in the ‘top-bottom’ sense (in the configuration of Figure 24-47, the 1-2 total fixture delay and the 3-4 total fixture delay do not need to be the same) although they commonly are equal.
- As with Type D, the Quick Extract check box will disable the file entry fields and instead the files will be save to a preset hard disk location, the de-embedding engine activated, and those files loaded into the de-embedding engine. This process can be used to save time if it one desires to immediately de-embed the fixture after extracting its parameters.

- The Neglect fixture match selection overrides contributions from multiple standards and S_{11} and S_{22} will both be set to 0 in all files.

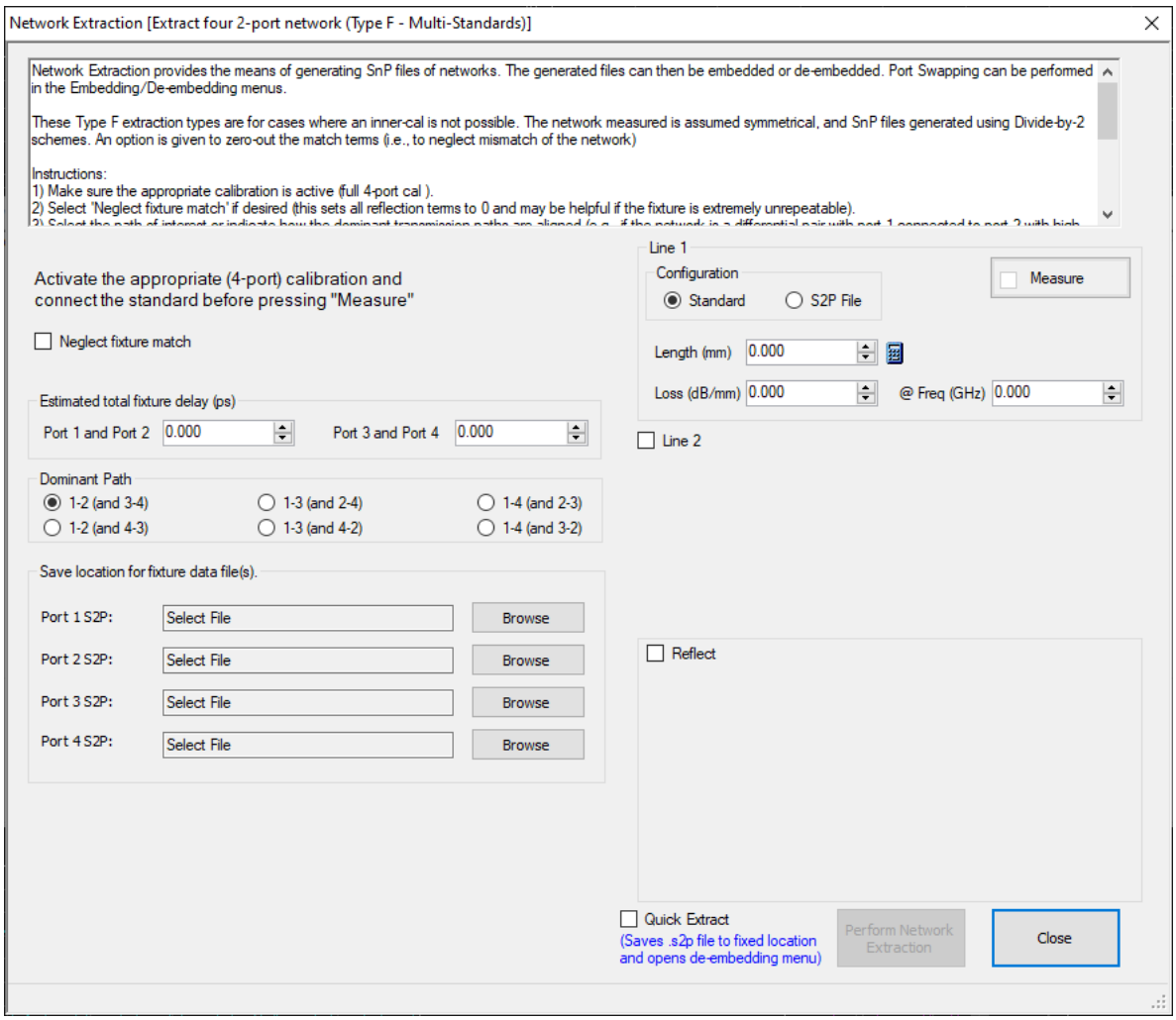


Figure 24-46. Extraction Dialog—Type F with Multiple Standards (Option 21 Enabled)

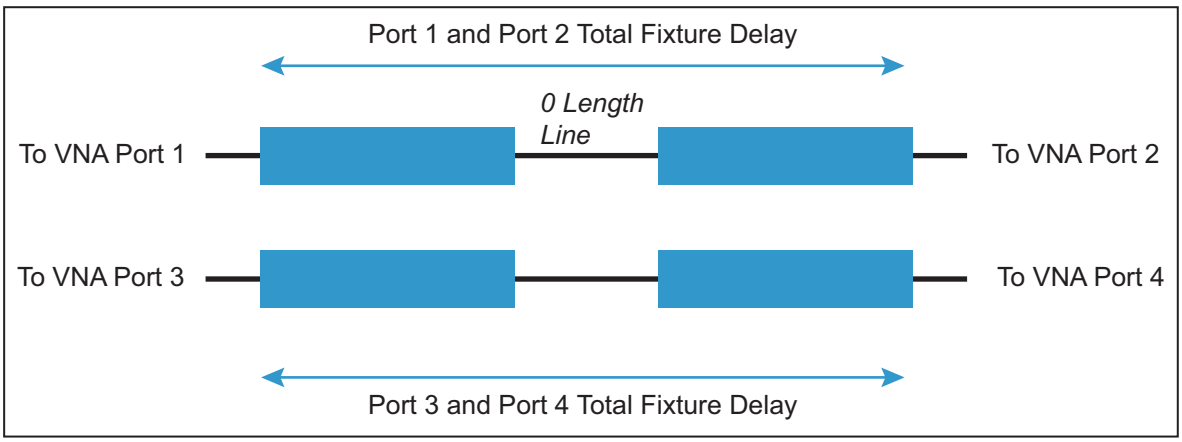


Figure 24-47. Type F Port 1/Port 2 Total Fixture Delay for Dominant Path 1-2 (and 3-4)

The configuration for the case of dominant path 1-2(3-4) with a zero length thru connected is shown here schematically.

Figure 24-47. Type F Port 1/Port 2 Total Fixture Delay for Dominant Path 1-2 (and 3-4)

Phase-localized F also works very much like phase-localized D except with additional files (dialog shown in Figure 24-48). Some of the attributes of this type:

- This method can use a thru (consisting of two interconnects as shown in Figure 24-48 for one example configuration), a single reflect (same reflect on all four ports), two reflects (different from each other but the same two reflects used on all four ports) or a thru plus a single reflect. See the phase localized D section for more details on the standards requirements.
- The fixture delay plays a more central role as that phase length is used to localize transmission and reflection behaviors for the fixture. If 0 is entered, an automatic estimation routine will be used. The 'top and bottom' delays need not be equal (1 and 2 total delay vs. 3 and 4 total delay for the example port configuration of Figure 24-48) although they commonly are. If a reflect standard is used, individual arm fixture delays can be entered (using the 'manual' mode available when the reflect standard is selected.
- There are frequency list requirements to use phase-localized techniques:
  - Log sweep and CW sweep are not allowed.
  - If segmented sweep is used, any individual frequency step cannot deviate more than 5% from the average frequency step.
  - The total fixture delay (in ns) should be at least 5/(Span of Frequency Sweep in GHz).
  - The total fixture delay (in ns) should be less than 0.3/(average frequency step in GHz).

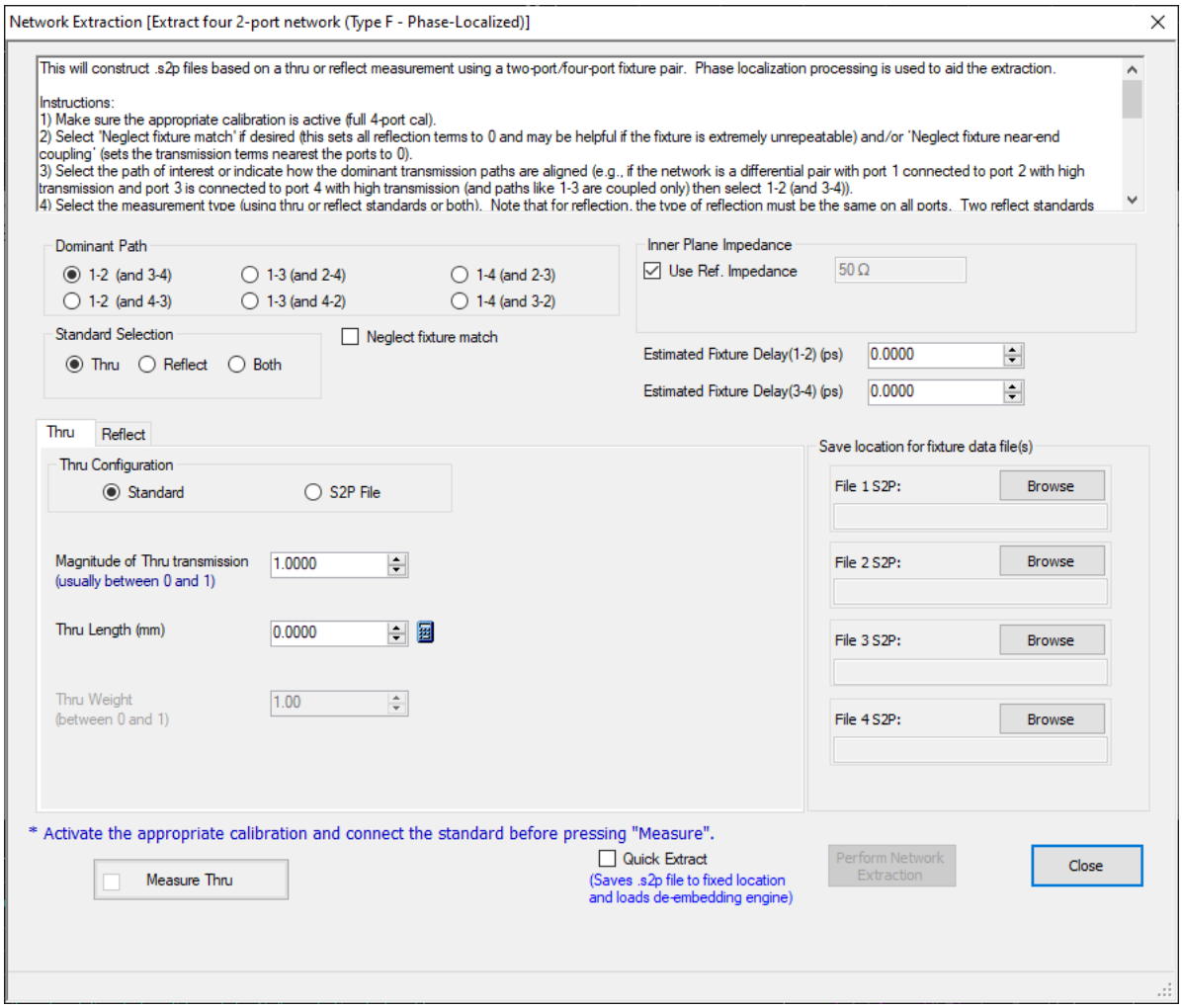


Figure 24-48. Phase-localized Type F Dialog

Type G Network Extraction

In the case of Type G, the S-parameter matrix will take on a very particular form. Ports 1 and 2 of the left network are assumed to be the outer ports and the inner ports for that network will be assigned Ports 3 and 4 for the S4P definition. Because of the match and cross-coupling assumptions, the matrix saved will take on the form below (Eq. 24-17).

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{13} & S_{23} & 0 & 0 \\ S_{14} & S_{24} & 0 & 0 \end{bmatrix}$$

Equation 24-17.

An eigenstructure-based solution is used to split the composite results in the upper right and lower left quadrants into the two “halves”. In most cases, there should not be convergence issues with the process unless the return loss of the structure gets very close to 0 dB. As shown, reciprocity is enforced. $S_{11} = S_{22} = 0$ if the check box is selected. The entire upper left quadrant is a direct map from the measurement as match and cross-coupling are assigned to the outer ports.

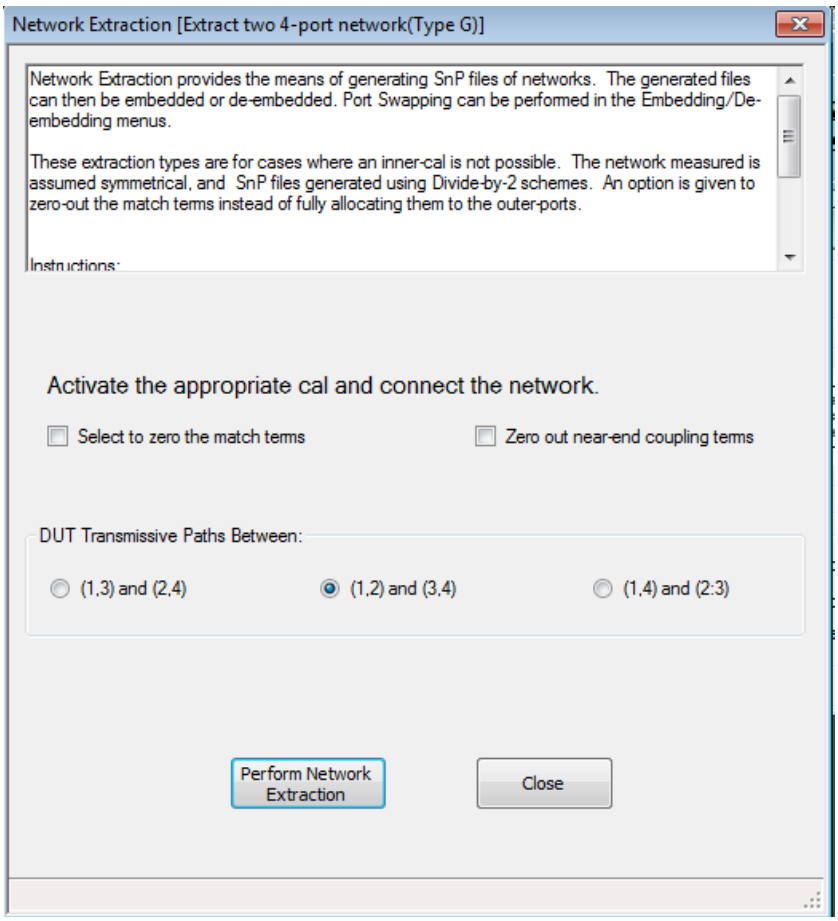


Figure 24-49. NETWORK EXTRACTION TYPE G Dialog—Without Option 21

Structurally, the measurement is the same as in Type F. The only difference is in the matrix structure of the extracted parameters.

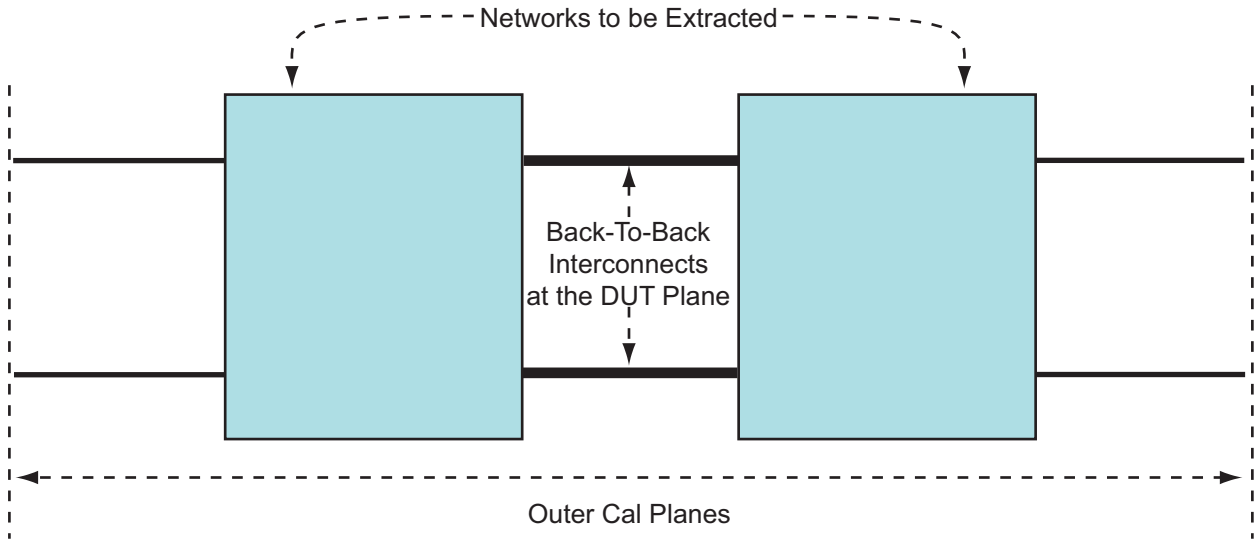


Figure 24-50. Back-to-Back Interconnects at the DUT Plane—Type G

The port assignment on Type G is more complicated than that for the other types since there are not just 2 file assignment permutations but 24. Type G will always put the port pair that has Port 1 in it on the outside (1-2, 1-3 or 1-4, depending on how the transmissive paths are excited) and it will do this for both files. So if the transmissive paths are 1-3 and 2-4; both files will have $S_{33}=S_{44}=0$.

The embedding/de-embedding engine assigns based on port pairs and assumes the files have the same port pair alignment. Since the port assignments between Type G and embedding/de-embedding could be completely disjointed (since the .s4p file may have been generated with a completely different setup), some care is required.

The VNA assumes a port concatenation type of numbering. Thus if the port pairs of the DUT are defined as 1-3 and 2-4, the system will expect ports 2 and 4 of the left EDE file to be connected to ports 1 and 3 of the DUT. Similarly, it will expect ports 1 and 3 of the right EDE file to be connected to ports 2 and 4 of the DUT. This is a little backward from 2-port embedding/de-embedding, but the requirement is to somewhat generally handle the ambiguity if the pairs are not aligned at all. Particularly for the second file, there can be some confusion about whether the ports should be swapped, but Type G offers some flexibility on port assignment to help. The *Use First File Port Assignments* will maintain this concatenation style and the files will then generally not need port swapping when used with the VectorStar embedding/de-embedding system. Other applications may expect a different type of alignment and, for these, the *Use current file port assignments* selection will assign the inner plane ports to the same location for the second file as for the first (i.e., both inner

planes would be ports 2 and 4 in the earlier example). The relationships are shown in Figure 24-51 for our example where the dominant paths are 1-2 (and 3-4) leading to outer port pairs of 1-3.

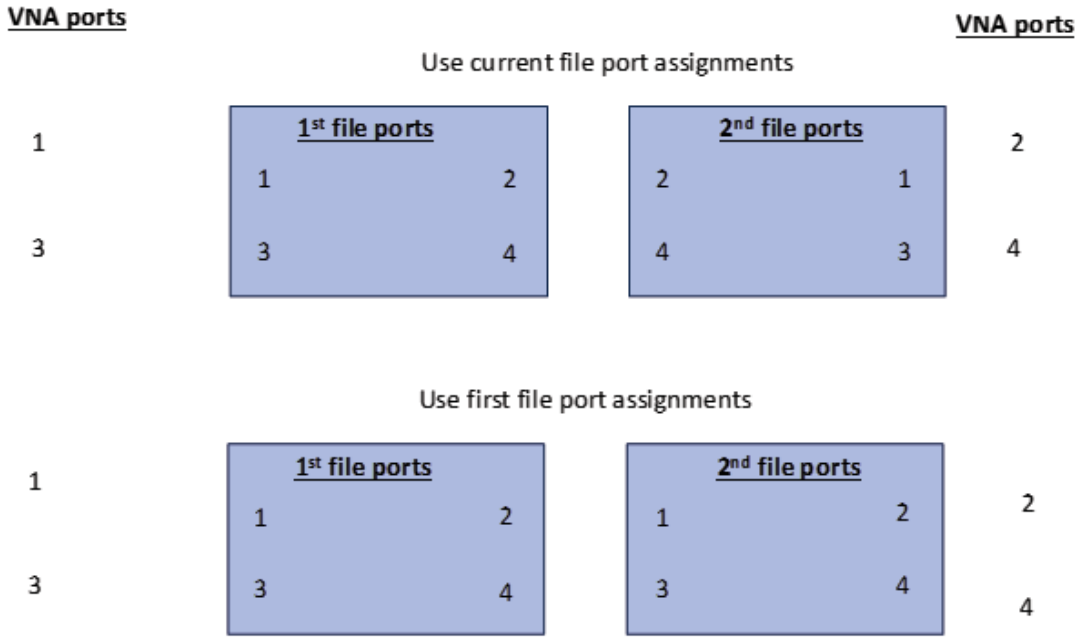
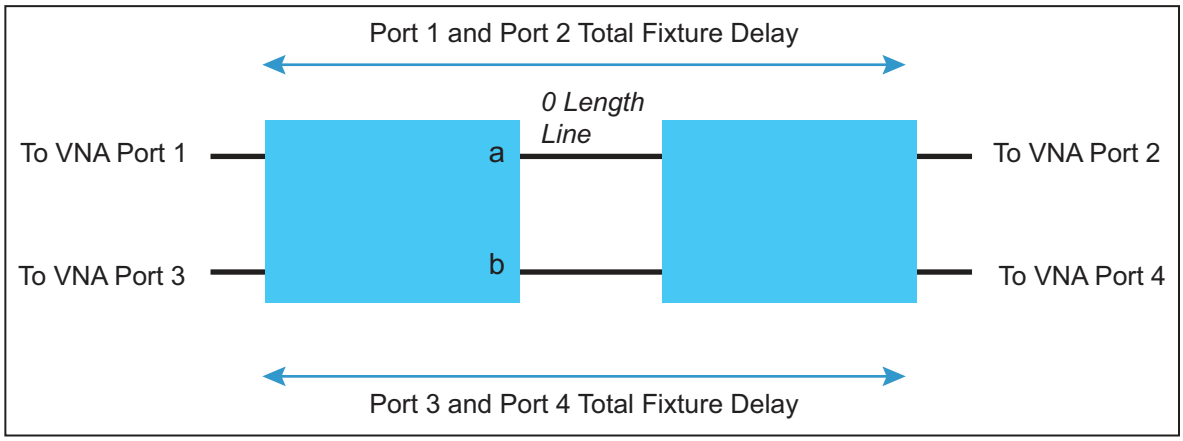


Figure 24-51. File Port Assignments

Type G Network Extraction (with Option 21—Universal Fixture Extraction)

Multi-standard Type G (again, this requires Option 21) functions much like Type F except now coupling is included in the networks and the networks are treated as two 4-port networks as opposed to four 2-port networks. The dialog is shown in Figure 24-52 and the standards choices are the same as with Type F. An example connection configuration (for one of the dominant path selections) is shown in Figure 24-53 with a zero length thru connected between the fixture halves.

Figure 24-52. Multi-Standard Type G Network Extraction Dialog (Option 21 Enabled)



For the example port configuration of 1-2 (and 3-4) dominant paths, the connection plan for a Type G extraction with a zero length thru is sketched above.

Figure 24-53. Type G Port 1/Port 2 Total Fixture Delay for Dominant Paths 1-2 (and 3-4)

As with non-Option-21 Type G, the basic one-line-standard solution involves the concept of a matrix square root to compute the insertion loss and FEXT of each half based on the composite measurement available. The interchange of energy between insertion loss and coupled paths can be large enough that this treatment is required. The inner plane match is assumed to be zero and inner plane crosstalk (S_{ab} in Figure 24-53) is assumed to be zero. Symmetry (left-to-right in the figure) in the sense of insertion loss and FEXT (but not outer plane match) is also a required assumption with the single line standard.

The addition of another line length or a reflect standard set (or both) helps add some information. As with Type F, the addition of a second line standard allows information on inner plane match and crosstalk (assumed symmetric) and some accuracy improvements on insertion loss and outer plane match. The addition of the reflect allows symmetry to be broken and allows for a coarse estimate of inner plane match. With one line plus a reflect, inner plane crosstalk is still neglected. With all three standards, symmetry is broken, the inner plane is solved for and general accuracy improves further.

As with Types F and D, the second line must have a different length than the first line and there are the same singularities at 0 and 180 degree phase differences.

The phase-localized variant is also available for Type G and its configuration dialog is shown in Figure 24-54. The construct is the same as for Type F with the difference that the localization process is also used to solve for crosstalk terms. The same restrictions on frequency list apply.

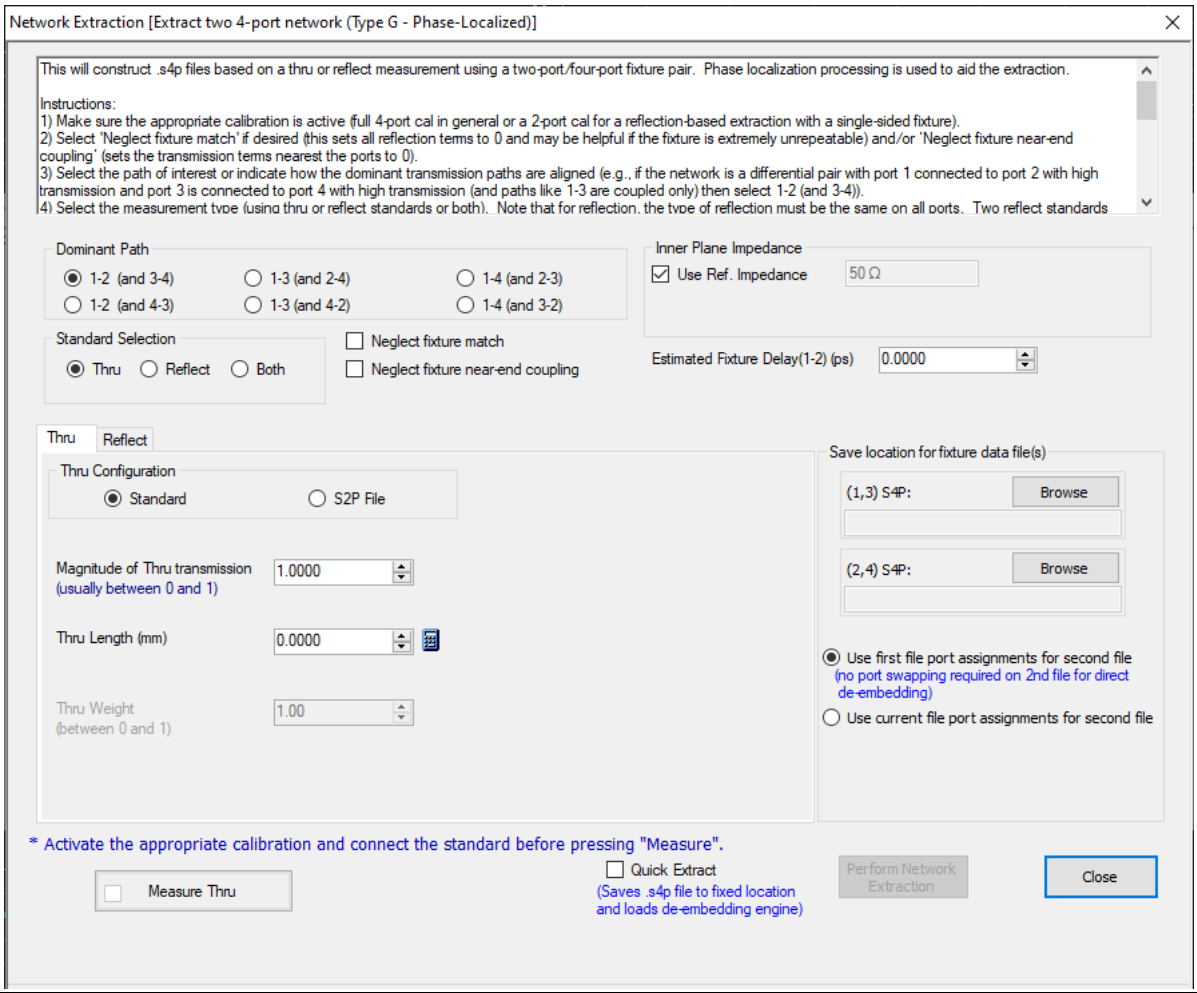
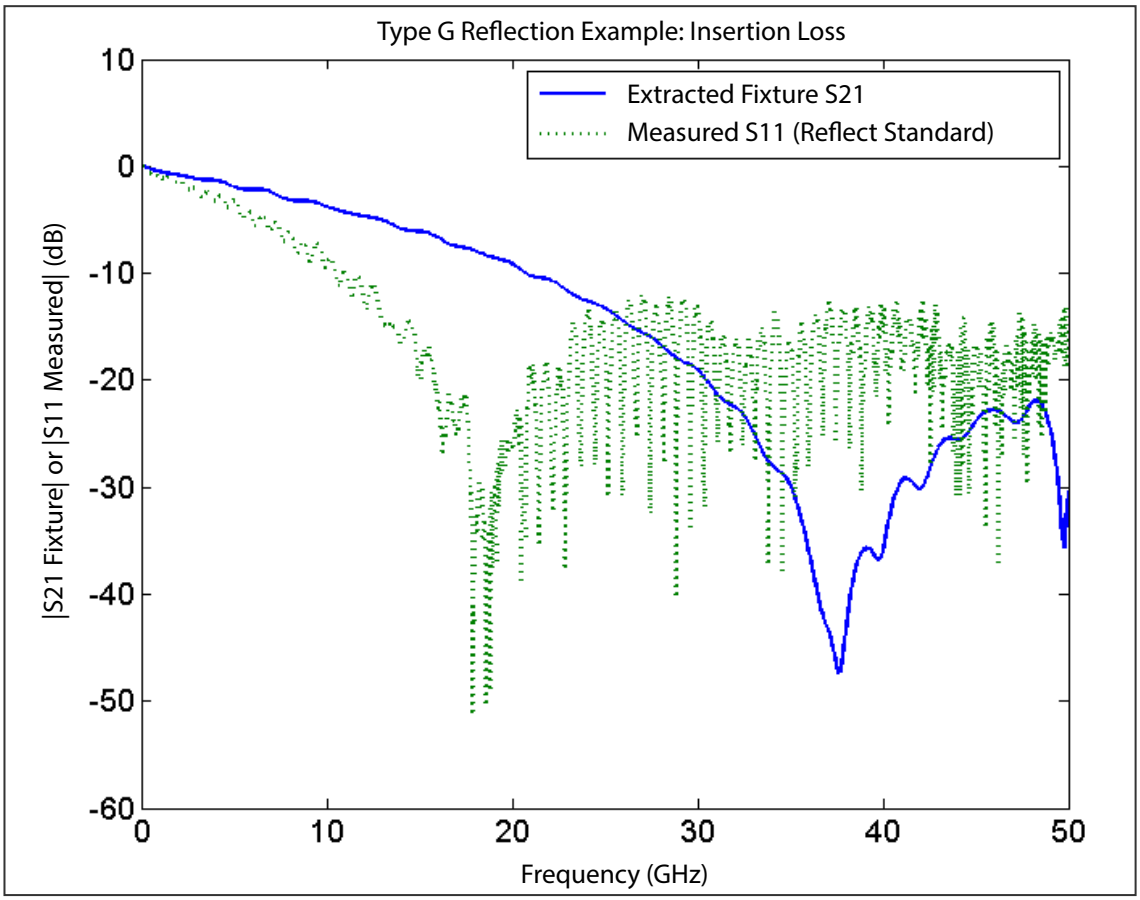


Figure 24-54. Phase Localized Type G Network Extraction Dialog (Option 21 Enabled)

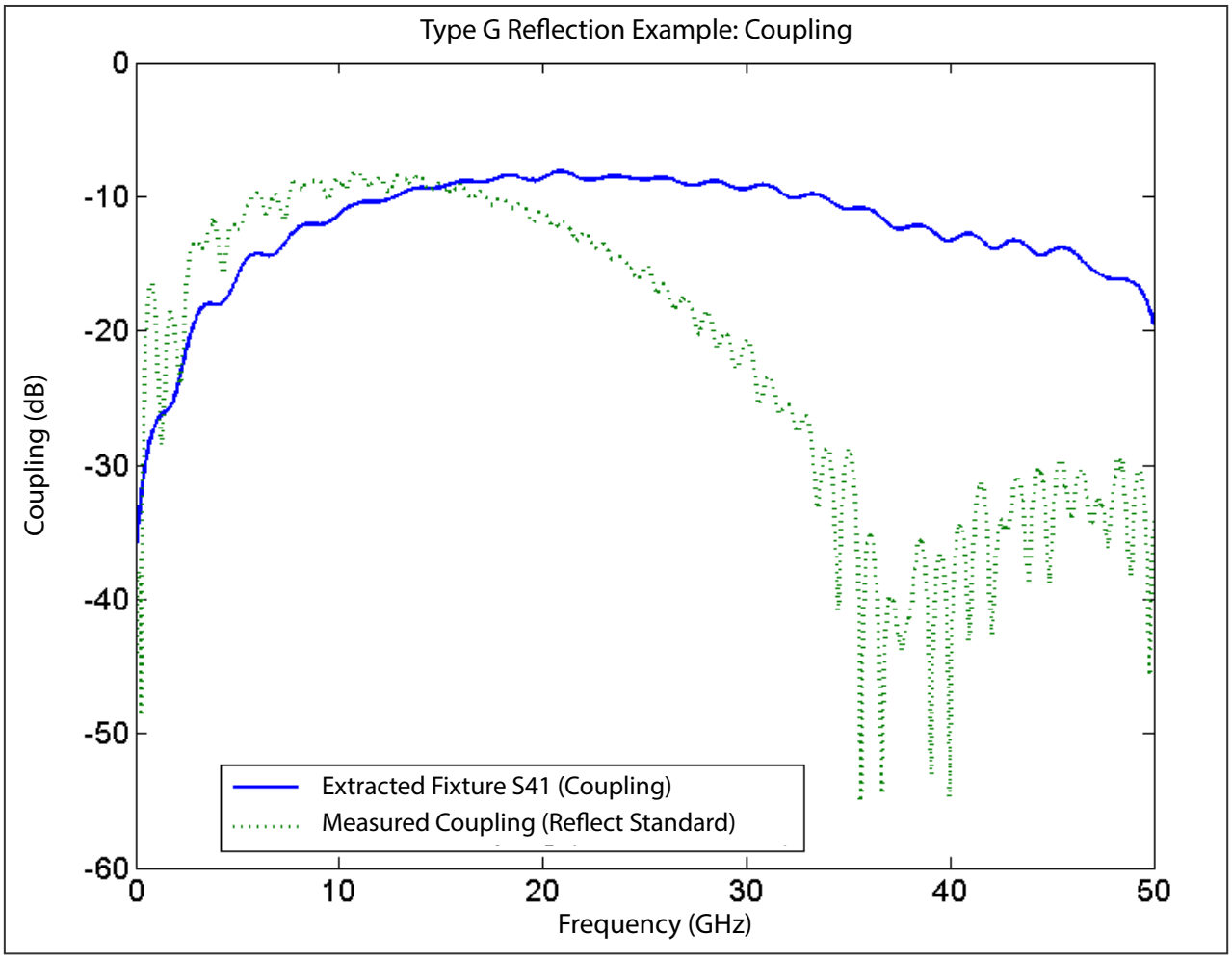
Consider an example reflection-based phase-localized Type G measurement where flush open standards are available at the inner interface. The fixture is a fairly tightly coupled differential pair with a number of internal reflection centers. The fixture length was unknown so a zero was entered for that and 1-2(3-4) was the dominant path selection. Thus the measurements needed for the extraction are based on ports 1-3 (for one half of the fixture) and ports 2-4 (for the other half (had a thru standard been selected, all 16 S-parameters would have been used—only 8 are needed when a reflect standard set is used). With the reflection standards connected, the dashed curve in Figure 24-55 was obtained which can be challenging to interpret. There is a null around 18 GHz which corresponds to a coupling maximum and there are many ripples due to interactions of the reflect standard with mismatch in the fixture itself. The extracted insertion loss of the fixture half (S21) is plotted as the solid curve and one can see the null has roughly doubled in frequency since the electrical length is about half that of what was seen in the reflection measurement with open present. Note that differential insertion loss will not experience this null. The multiple reflection centers in the fixture (which are real) contribute to the residual ripple in the result. The phase localization process allowed ripple due to interaction with the reflect standard to be removed.

The coupling (S41) of the same fixture is plotted as the solid line in Figure 24-56 on page 24-66 along with the input S31 measurement (with opens at both inner ports connected to outer ports 1 and 3). As expected for a coupling mechanism for a coupled line, there is a null at low frequency. There is another null in the input data (reflection standards connected) near 37 GHz. On the extracted data for S41, the null has moved out to twice that frequency due to the same change in electrical length. This illustrates the importance of processing coupling and insertion loss together on fixtures such as this. Treating them independently would result in the null remaining at the now non-physical frequency of ~37 GHz. As with the previous figure, the residual ripple is due to physical reflection centers within the fixture interacting.



The solid curve denotes the extracted insertion loss profile (single-ended) for the example fixture and the dashed curve is some of the measured data (with opens at the inner plane) used to compute the solid curve.

Figure 24-55. Phase-localized Type G Reflection Example—Insertion Loss



The coupling for this fixture (extracted S41 and measured (with opens) S31) are plotted in this figure. Note that the null has moved out as expected.

Figure 24-56. Phase-localized Type G Reflection Example—Coupling

As with phase localized D and F, there are some standards permutations allowed. Two reflect standards (different from each other but the same reflects are used on all ports) can be used where the added information is used to refine transmission and crosstalk results (and inner plane match to some degree). Also, a thru (in the sense of two port interconnects) can be used in conjunction with a reflect standard (same on all ports) where the additional information is used to refine knowledge of the match terms in particular. A weighting term is provided (range 0 to 1 where 1 indicates the thru measurement is used more heavily and a 0 means the reflect measurement is used more heavily). It is not allowed to use a thru with two reflects as the amount of information increase was found to be diminishing.

Type G Network Extraction Phase-localized High Crosstalk (with Option 21—Universal Fixture Extraction)

A special case of type G processing using the reflect standards is also provided where the effects of a neighboring pair (with a high level of crosstalk between the primary pair and the neighbor) are handled more completely. This can be an important problem since, when those crosstalk levels are high, how the neighboring pair is terminated will change the loading on the primary pair and, if that loading is not the same at characterization time and when the path is being used, additional uncertainty can be incurred. The approach is to characterize both the primary pair and the neighbor at the same time (using reflect standard-based phase localization) and create a complete model of the interactions. Then, different loading levels (expressed via reflection coefficients or .s1p files) can be applied to the model and the .s4p of the primary pair generated for that level of neighbor loading. As might be expected, if the crosstalk levels between the primary and neighbor are low, this method provides no real benefit and the regular methods can be used. The level of crosstalk of relevance varies with the situation

- Insertion loss effects are normally visible only on lower loss lines and are small unless the crosstalk is very high (> –10 dB) but can be on the scale of tenths of a dB for slightly lower levels.
- If intra-pair coupling levels are on the same order as the crosstalk, considerable dependence may be observed on those primary pair coupling parameters.
- If the line losses are low and the lines well-matched (relative to the crosstalk levels), variations in match can be observed.

The configuration dialog for this special method is shown in Figure 24-57 and has many of the same attributes as the usual phase localized G (dominant path selection, reflection definition for the inner plane, match term and file name selections, fixture delay estimates and inner plane impedance). The connection of the VNA to the fixture assembly during characterization is different in that one pair of ports is used for the primary pair and the other pair of ports is used for the neighbor (instead of being used for the other half of the fixture; thus this method only characterizes on 'half' of a fixture per run). The connection configuration is illustrated in Figure 24-58 for two different port configurations and the connections here must be selected under Measurement Port Assignment.

The italicized port numbers in Figure 24-58 denote how the port numbers will be assigned in the output .s4p file. As with other network extraction methods, this comes from the dominant path selection, so one should check that selection carefully. Since the embedding/de-embedding system does allow port swapping, the arrangement can be changed later. The red arrows in the figure represent the cables from the VNA with the port numbers indicated. As with other type G methods, a coaxial calibration is assumed to exist (at reference planes to be connected to the outer ports of the fixture) and the reference plane should effectively be at the tips of those arrows.

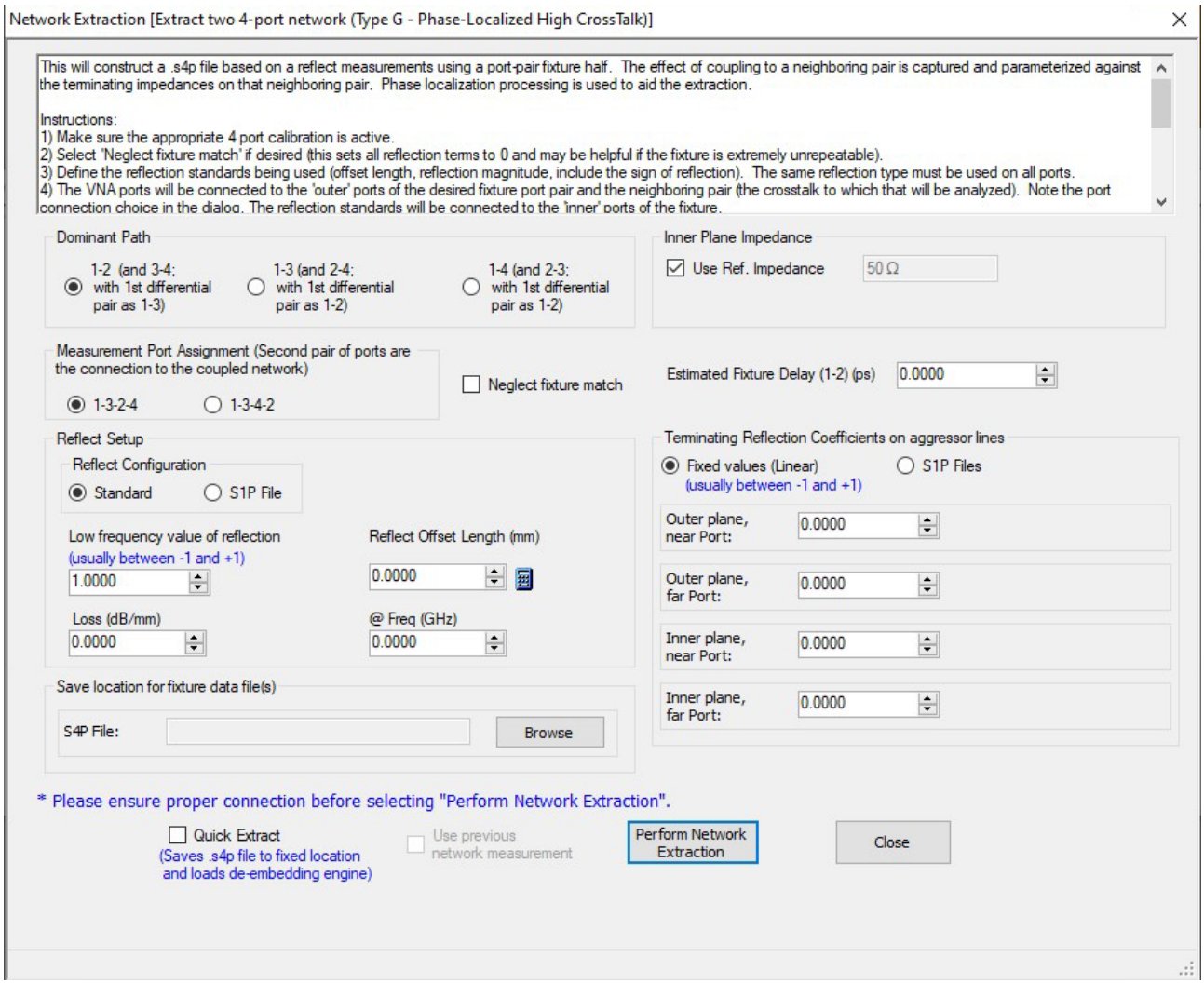
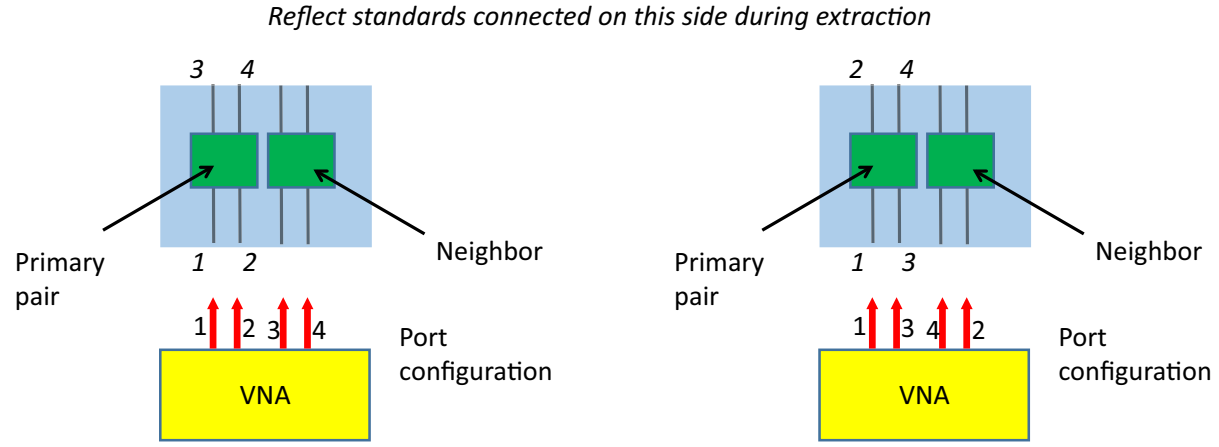


Figure 24-57. Phase-localized High Crosstalk Type G Network Extraction Dialog (Option 21 Enabled)



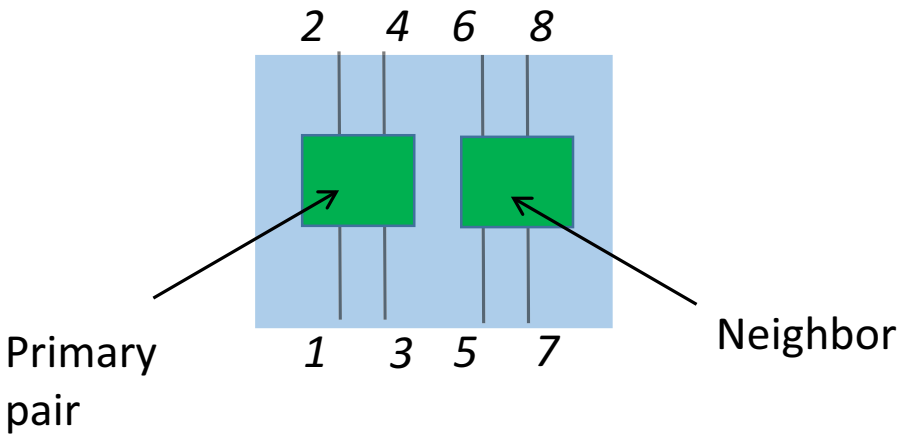
Dominant path: 1-3; port configuration 1-2-3-4

Note that one pair of ports is used to characterize the 'neighbor' pair. The port assignments are linked to the dominant path selection and the radio buttons will change in the dialog of Figure 24-57 accordingly.

Figure 24-58. Two different connection configurations for the characterization measurement.

As Figure 24-58 pictorially suggests, the characterization will result in a .s8p file which incorporates all of the crosstalk behavior (under the phase localized assumptions so the same length requirements apply as for regular phase-localized G). When Perform Network Extraction is selected, this .s8p file is generated and, if desired, can be used elsewhere. It is saved as a fixed file name at `c:\ProgramData\Anritsu Company\AC-VNA-VectorStar\Dependencies\NetworkExtraction` with file names of the form `Ch1_phaseLocalizedTypeGCrossTalk.s8p` (and the '1' would be replaced by the channel number for files generated from other channels). Every time this method is used, that file is overwritten for the currently active channel.

If one does want to use the .s8p file externally, note that the port assignments for this file are fixed (see Figure 24-59) and will not change based on settings in the dialog box. Internally, the measurements are all mapped to a fixed port assignment for processing. When the .s8p data is used to create the final .s4p file for the primary pair, the mapping is undone. This mapping process is invisible to the user unless the .s8p file is used externally.



This information is only relevant if the .s8p file is used externally to the VectorStar application.

Figure 24-59. Port Assignments for .s8p File

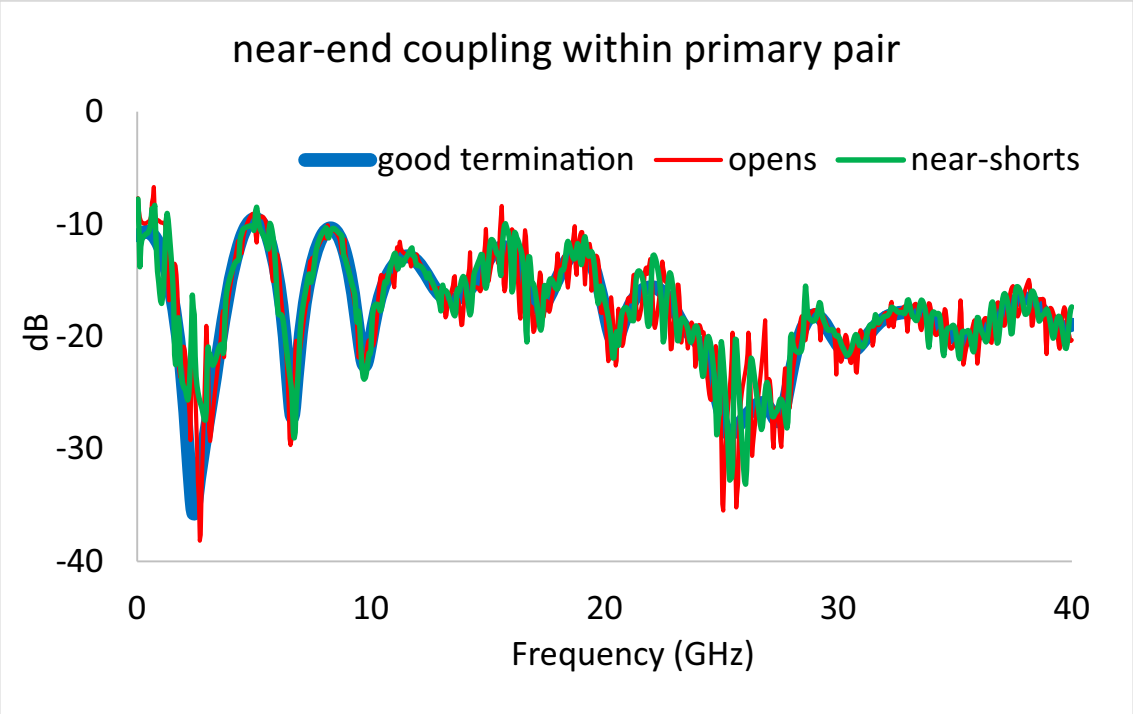
The .s8p file itself is difficult to use in practice and the intent of the feature is to assess the loading effects of the neighbor and how they can affect the primary pair. This method's primary objective is to create a .s4p file for the primary (and that is the file name entered in the dialog) for a given loading of the neighbor. This is where the reflection coefficient entry in the lower right of Figure 24-57 come into play. The reflection coefficients of the loads attached to the neighbor pair (constant with frequency and real only) can be entered here or .s1p files can be loaded for those loading elements in order to include frequency dependence and complex values. Recall that, by definition, the .s8p file will describe the primary pair characteristics when the neighbor is terminated perfectly (reflection coefficient equals zero). The first reflection coefficient entry is assigned to the outer plane neighbor port closest to the primary pair and the second entry is assigned to the other neighbor outer plane port. Similarly, the third entry corresponds to the inner plane neighbor port closest to the primary pair and the fourth entry belongs to the final (further away) inner plane neighbor port.

Now, it may be desirable to generate primary-pair .s4p files for several different neighbor loading states and it could be useful to not have to repeat the characterization every time. This is where the saved .s8p file can be used along with the Use Previous Network Measurement checkbox. If that box is checked, another measurement will not be taken but the existing .s8p file will be used with the current reflection coefficient/.s1p definitions for the neighbor pair terminations. This checkbox is only available if a .s8p file has been generated and is present in the directory described earlier (e.g., after preset, there will not be such a file).

As an example of the performance of this method, a tightly coupled set of lines (10–20 dB coupling between pairs over most of the frequency range) was analyzed. It was desired to know the primary pair S-parameters for fixture de-embedding under three different loading conditions corresponding to different ways that the fixture would be used with DUTs.

1. Good terminations (near 50 ohms single-ended). A fixed reflection coefficient of 0 was entered.
2. Opens. A fixed coefficient of 1 was entered.
3. Near shorts. A .s1p file was used with an average reflection magnitude of 0.99 and a high frequency phase of 150 degrees (179 degrees at low frequency).

Since the crosstalk was –10 to –20 dB and intra-pair coupling was only a few dB tighter on average, noticeable impact on near-end intra-pair coupling (see Figure 24-60) was expected. Indeed, the variation exceeded several dB in places. It is interesting to note that the added ripple present for the more reflective terminations is nearly 180 degrees out of phase between the open case and the near-short case and this follows intuitively. The input match (Figure 24-61) also shows some variation that is dependent on the crosstalk level at a given frequency, the mismatch level and the location along the line that is the dominant contributor to mismatch. Insertion loss for this particular example showed 0.3 dB peak variation with the neighbor terminations tried.



The solid curve denotes the extracted insertion loss profile (single-ended) for the example fixture and the dashed curve is some of the measured data (with opens at the inner plane) used to compute the solid curve.

Figure 24-60. Fixture Near-end Coupling (Primary Pair) for Different Neighbor Loading Levels

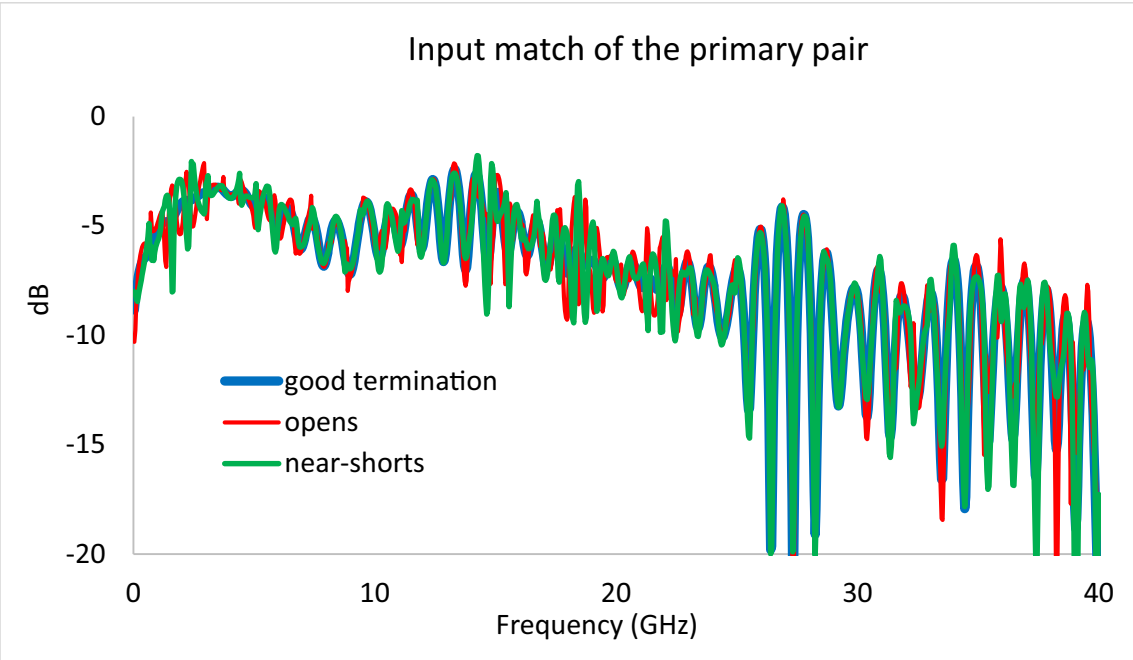


Figure 24-61. Fixture Mismatch (Primary Pair) for Different Neighbor Loading Levels

To summarize, the high crosstalk version of phase localized type G is most valuable when there is a differential pair (the neighbor) that is tightly coupled to the differential pair to be extracted and the terminations of the neighbor may vary between different DUT measurements and/or between characterization time and DUT measurement time. The loading effects of the neighbor are parameterized against neighbor terminations by this method to make it easier to generate more accurate de-embedding files for the primary pair.

Sequential Extraction (Peeling)

Another method of network extraction involves modeling the network as a collection of lumped elements. This is particularly popular for electrically small structures (e.g., on-wafer) of those with runs of transmission line punctuated by electrically small structures (e.g., PC boards with isolated vias in transmission lines).

Procedurally, this method works on one lumped element at a time. This method is covered in more detail in "Sequential Extraction—Peeling (with Option 21)" in Chapter 9, but the concept is similar for 2-port and 4-port constructions. When the lumped defect/element is treated as a simple series impedance or shunt admittance, a .s2p file is generated that can be de-embedded to allow one to get at the next element. For the 4-port case, an impedance coupling two lines can also be treated in which case a .s4p file is generated. Also transmission line segments can be separately de-embedded to get between lumped defect areas. The process is based on reflection measurements only in the case of 2-port analysis and mixed-mode reflection measurements in the case of 4-port analysis (which one can think of as a 2-port measurement within a 4-port context). A full calibration incorporating the relevant port(s) must be in force.

Uncertainties and Sensitivities

Much was covered in Chapter 9 on the sensitivity and uncertainty issues with a number of network extraction methods and little changes in the 4-port cases. Type E is just an extension of Type C in requiring two full-port calibrations and hence being sensitive to all calibration standards and measurements. Types F and G are both partial information techniques like Type D and attempt to get solid estimates of insertion loss (and crosstalk in this case) in the presence of possibly poor repeatability and will ignore inner plane match issues. To emphasize the repeatability importance to this selection process, a comparison is shown in Figure 24-62. Inner plane match is ignored also in the 4-port variants and can have an impact if sufficiently poor as suggested in Figure 24-63. For this example fixture, a Type E method would have experienced errors exceeding 3 dB due to the difficulty of making any reasonable reflection standards at a geometrically complicated interface.

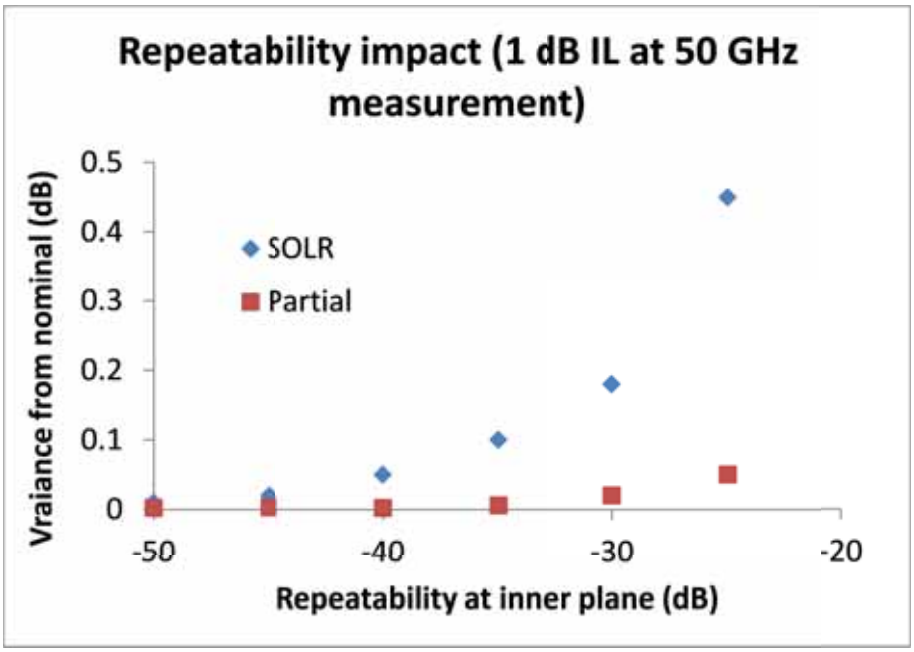
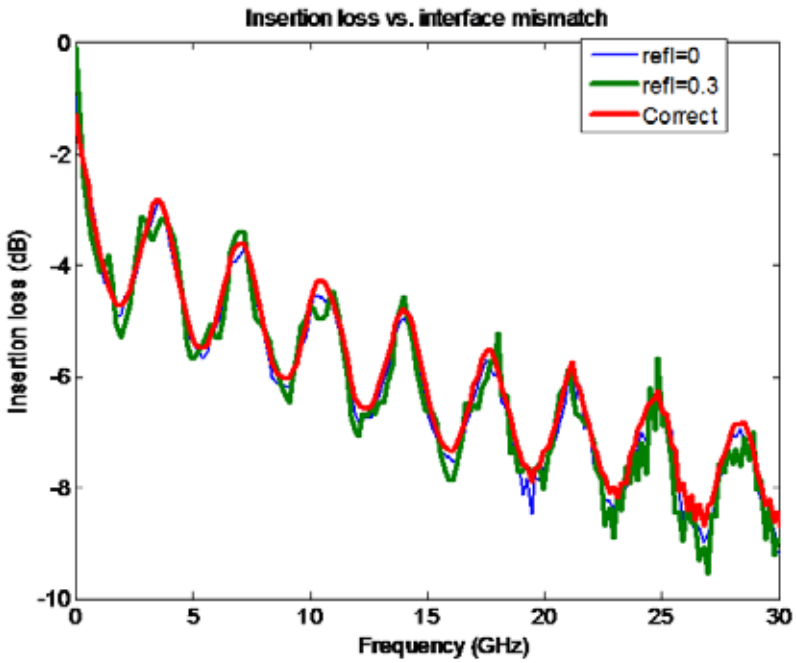


Figure 24-62. Plot of Repeatability Impact for a Partial Information Technique (1 of 2)

A plot of repeatability impact for a partial information technique (like methods D, F, and G) is shown here compared to a conventional calibration.

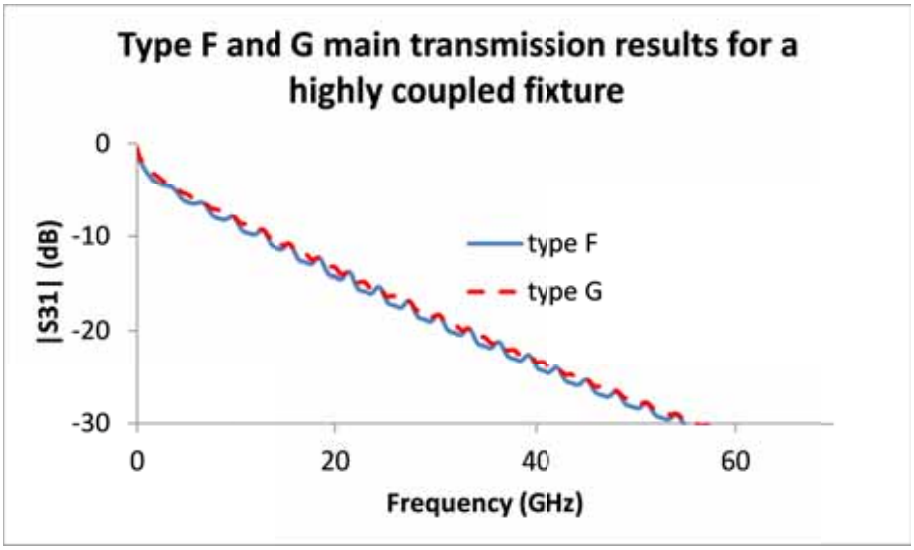
Figure 24-62. Plot of Repeatability Impact for a Partial Information Technique (2 of 2)



The effects may be noticeable but the alternative approaches may be much worse depending on the geometry and media.

Figure 24-63. Effect of Inner Match on a Type F or G Extraction

As stated earlier, the main difference between Types F and G is in handling coupling between fixture arms. Type F treats each of the 4 arms as an independent 2-port network while Type G treats each of the two macro 'sides' of the fixture as a 4-port network and hence can include coupling terms. As might be expected, the difference between the two methods becomes apparent with a coupled fixture and a measurement example is shown in Figure 24-64 for a structure with ~10–20 dB coupling levels.



The difference in measurement performance for Type F and G are shown here when the fixture arms are tightly coupled. The results generally overlay when the coupling is loose.

Figure 24-64. Difference in Measurement Performance for Type F and G when the Fixture Arms are Tightly Coupled

24-11 Embedding/De-embedding

The main addition relative to the embedding/de-embedding discussion in Figure 11 is the ability to embed or de-embed (E/DE) 4-Port networks thus allowing the use of differential and other coupled circuits. The usual single-port E/DE with 2-Port networks is still supported with a multiport cal (for uncoupled networks) and the interface works exactly the same as in Chapter 11, "Calibration and Measurement Enhancements" (with the obvious change that the 2-Port networks can now be connected to any of the four ports). As before, a check will be completed between the embedding/de-embedding configuration requested and the current calibration in place. If the current cal does not support the E/DE request, an error will be generated (for example, trying to embed a network on Port 4 when only a 1-2 two-port cal exists).

The new addition, shown in the below dialog, is that now one may specify port pairs to which a network is connected.

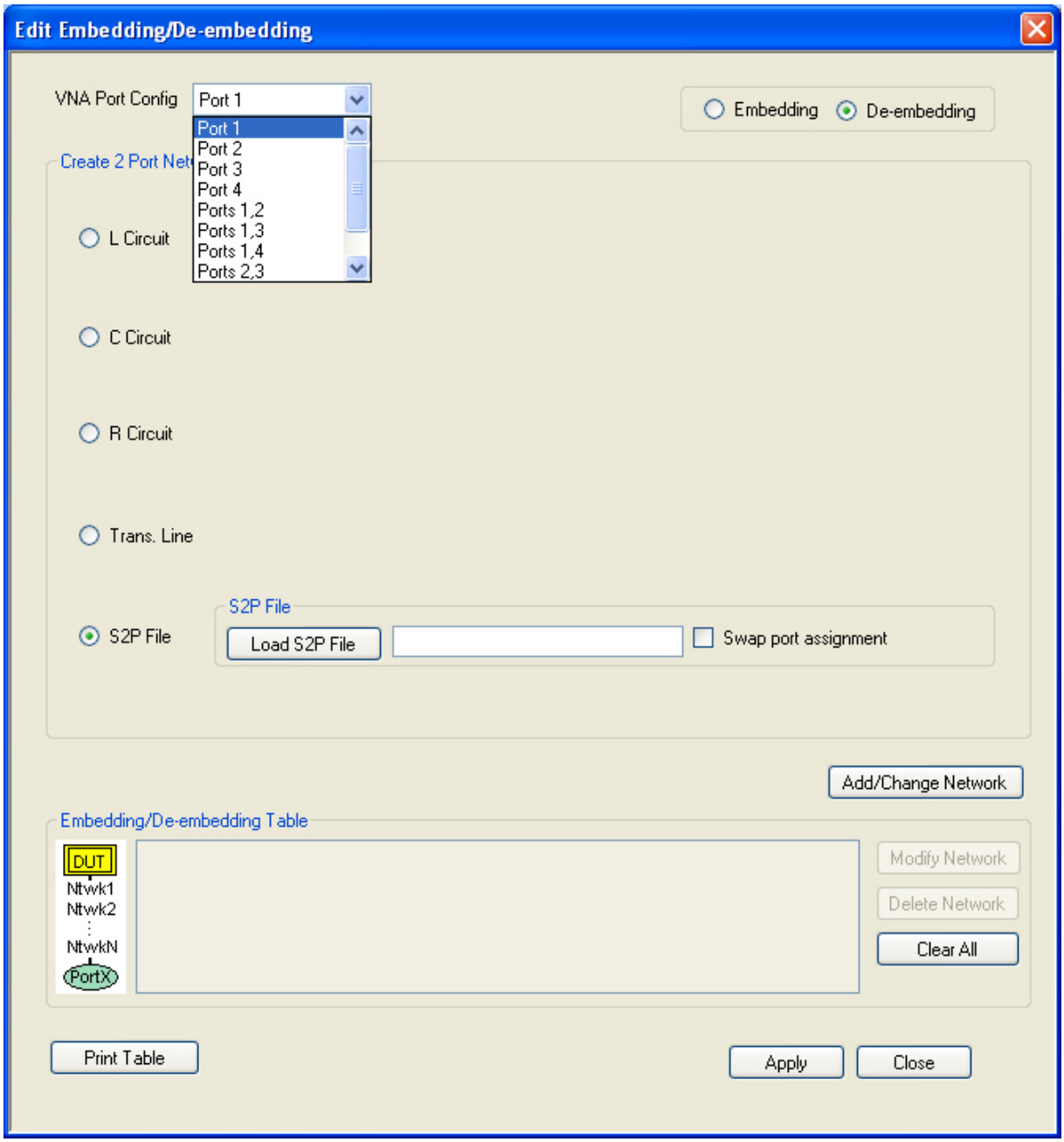


Figure 24-65. EDIT EMBEDDING/DE-EMBEDDING (E/DE) Dialog Box—S2P File—4-Port VNAs

The inductive, capacitive and resistive elements are all quite simple and the inductive entry field is shown below as an example. As before, the respective default units are nH, pF and Ohms. Note that these elements are all cross-elements between ports. For series elements or elements to ground, one would use the original single port e/de discussed earlier.

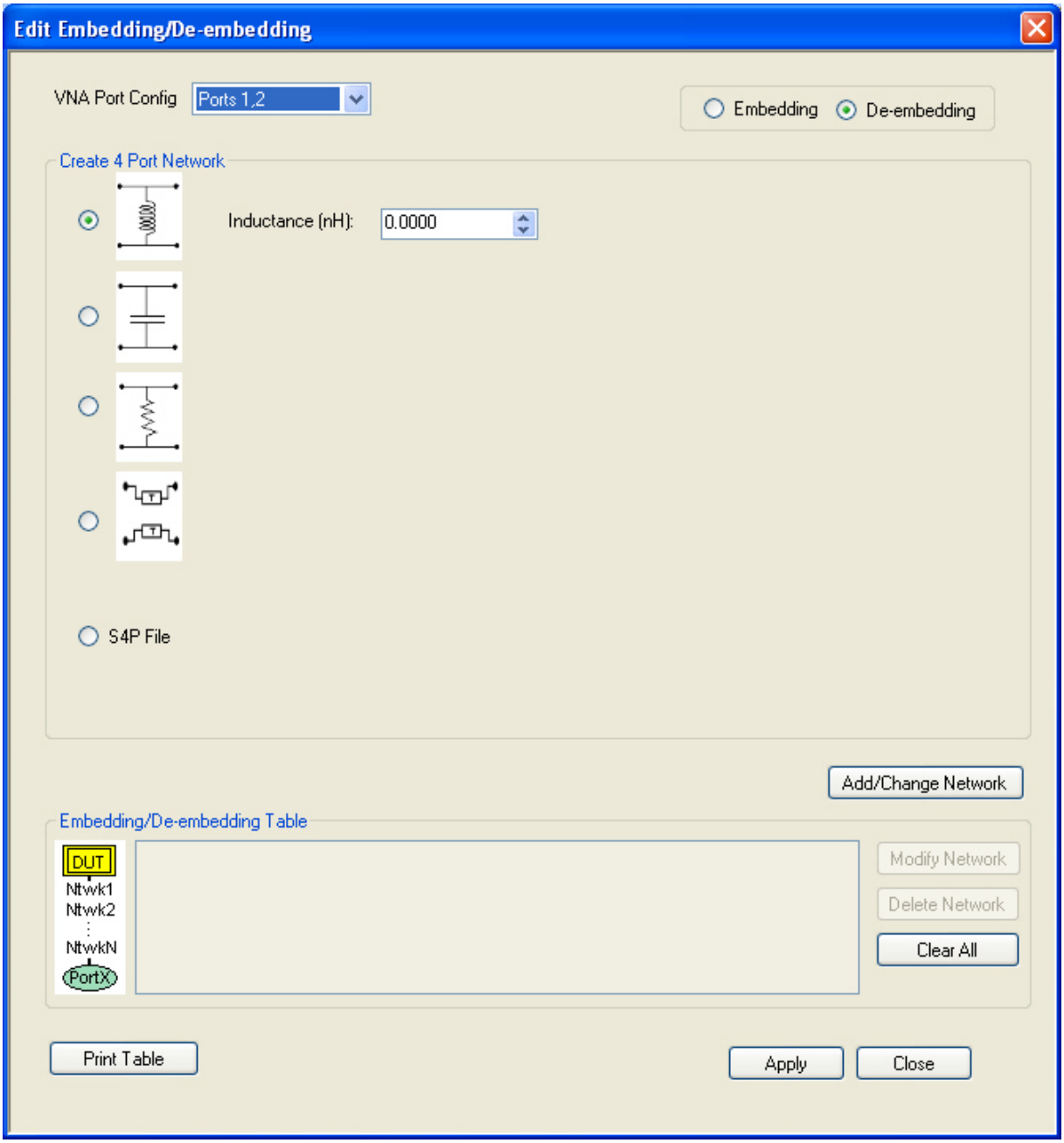


Figure 24-66. E/DE Dialog Box—S2P File—Inductance—4-Port VNAs

The transmission line element is a bit more complicated in that a general coupled transmission line was desired. As shown in the below dialog, the impedance, effective dielectric constant (and hence electrical length) and loss can be specified separately for the even and odd modes on the transmission line. The same loss scaling with frequency rules apply as with single port (2-port network) E/DE. If 0 is entered as the reference frequency (f_0, the "*@frequency*" box), the loss value will be applied uniformly at all frequencies.

$$Loss\ (f) = Loss\ (f_0) \times \sqrt{\frac{f}{f_0}}$$

Equation 24-18.

While coupled line theory is too large a topic to cover here in detail (for a brief treatment, see, e.g., D. Pozar, *Microwave Engineering*, Chapter 8, Addison-Wesley, 1990), there is a relationship between the even and odd impedances and the level of coupling that holds when the even and odd effective dielectric constants are close.

$$Z_{0e} \approx Z_0 \sqrt{\frac{1+C}{1-C}}$$

$$Z_{0o} \approx Z_0 \sqrt{\frac{1-C}{1+C}}$$

Equation 24-19.

Where C is the voltage coupling coefficient. For highly dispersive structures, the relationship is more complicated and one may want to resort to S4P file definition.

The S4P file loading works very similarly to the S2P loading discussed earlier. As in that case, the system will attempt to interpolate and extrapolate the best it can when the file frequency range and the sweep frequency range do not match exactly. If there is no overlap between frequency ranges, an error will be generated.

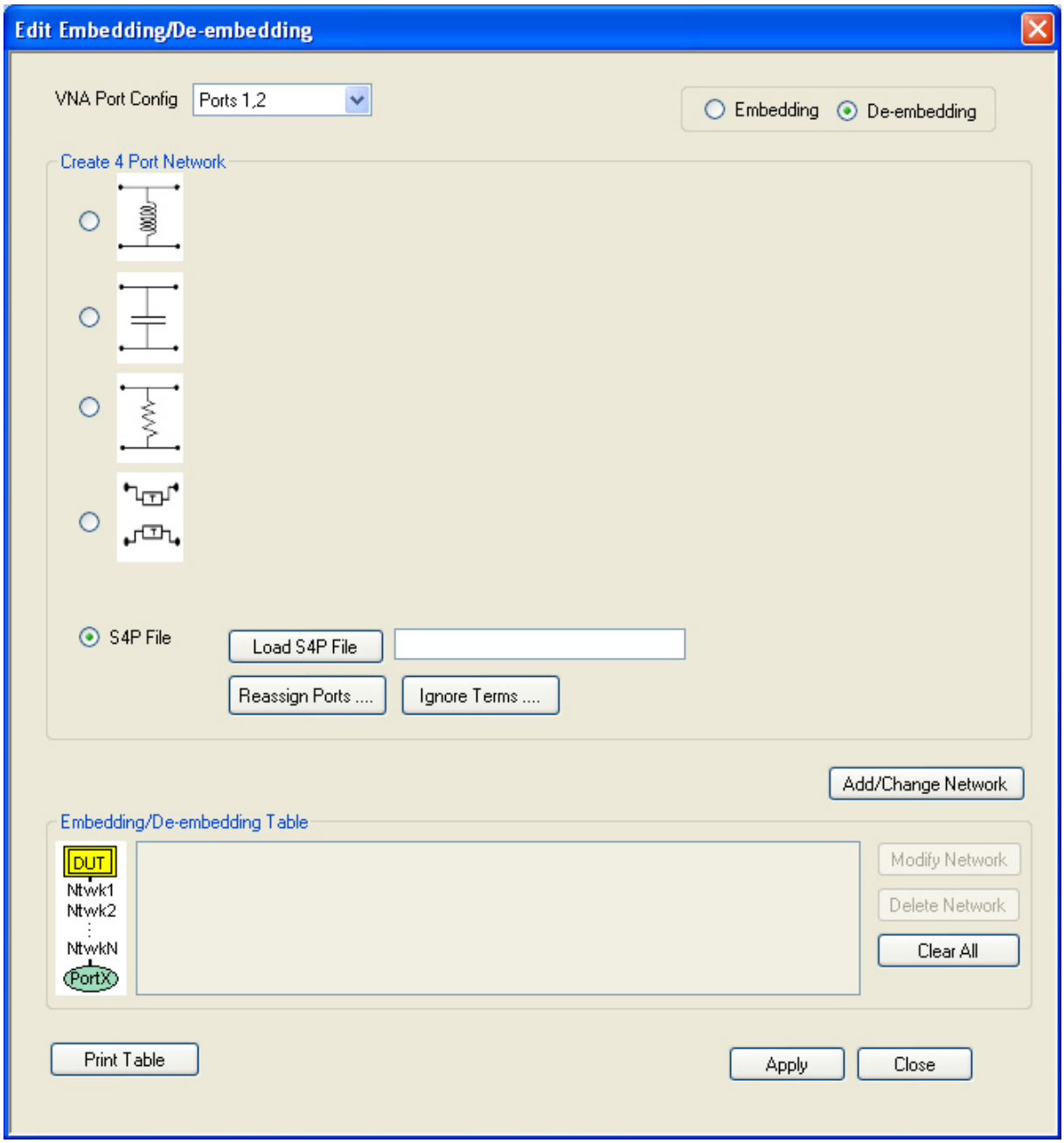


Figure 24-67. E/DE Dialog Box with Load S4P Selected—4-Port Networks

Reassign Ports Options

Instead of a *swap ports* option, there is a more general "reassign ports selection" whose dialog is shown next. This is useful in the case that the port assignment used during file generation does not match how one wants to embed or de-embed that data. The dialog allows an arbitrary reassignment but, of course, the port mappings must all be distinct (one cannot map both Ports 1 and 2 onto Port 4). The example below (Figure 24-68) shows Ports 1, 2, 3 and 4 reassigned to Ports 4, 3, 2, and 1.

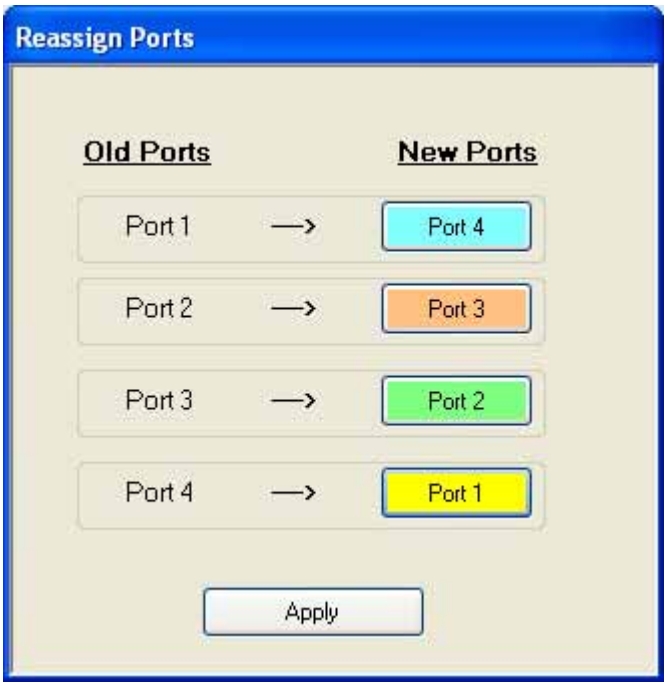


Figure 24-68. REASSIGN PORTS Dialog Box—Linked from E/DE dialog box and Reassign Ports button

A given port number in the file will be linked to the same port number of the VNA. The port assignment protocol can sometimes be confusing since the system does not know, a priori, how the .s4p was generated. If the port assignment in the file does not match how the network is connected to the VNA, surprising results can occur. An example is shown in Figure 24-69 where a network is attached to ports 1 and 3 of the VNA. The assumption is that the file will have been generated with the same connection pattern. If the file was generated with some other pattern, then the port swapping needs to be used.

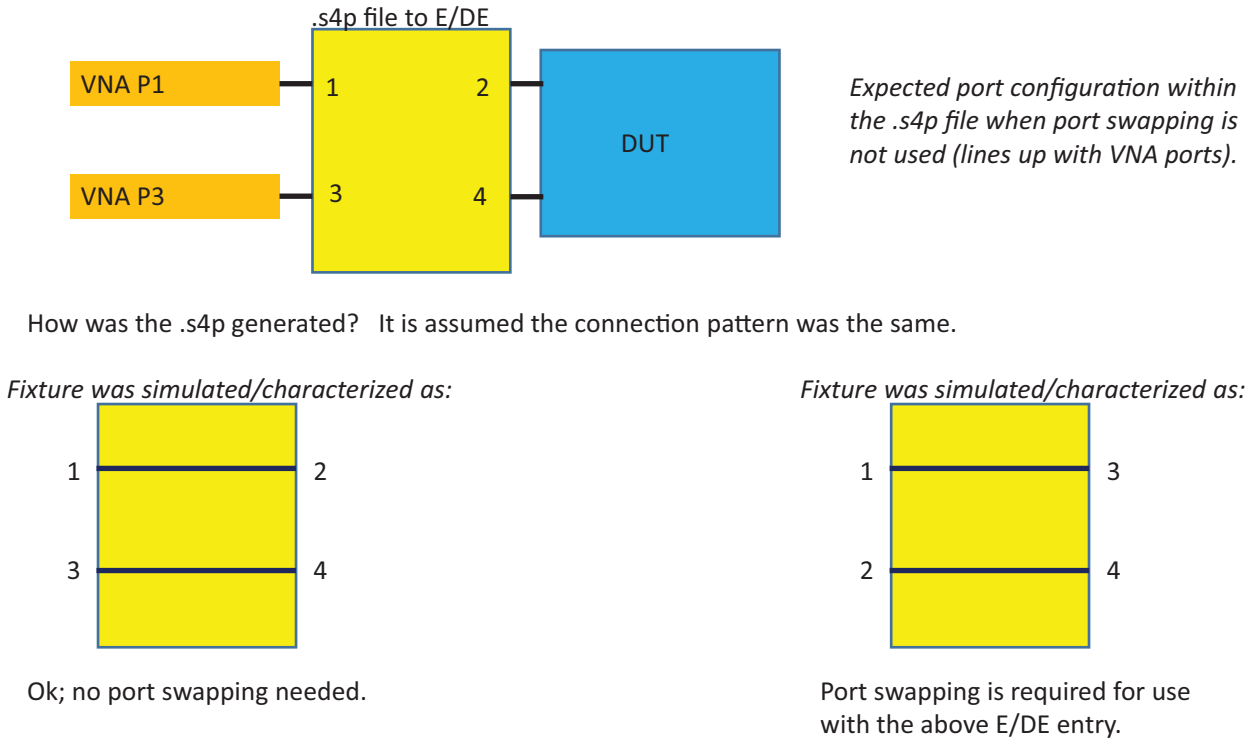
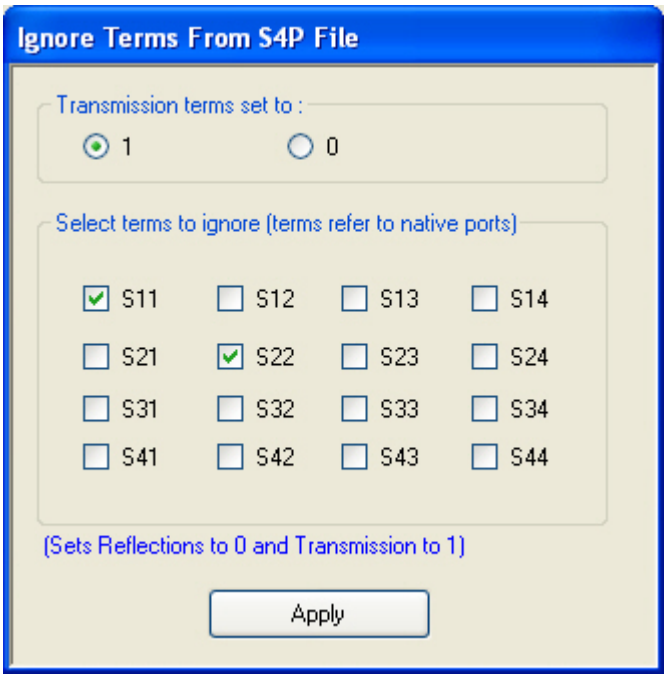


Figure 24-69. Example Showing When Port Swapping May Be Needed

Ignore Terms Options

The “ignore terms” is another dialog selection which allows one to omit match terms or leakage terms in the S4P file that may not be meaningful. A check box in this matrix will suspend loading of that parameter for that file. An example of this use might be the files generated from Type G network extraction where one now decides that the match terms (S_{11} and S_{22}) really are not that meaningful. In this case, the check boxes for those two parameters would be selected.



Transmission Terms set to 1, with Terms S_{11} and S_{22} to be ignored.

Figure 24-70. IGNORE TERMS From S4P File Dialog Box

24-12 Multiport Broadband/mmWave Measurements

Almost all of the concepts discussed elsewhere in this chapter also apply to broadband and mmWave measurements performed using variants of the ME7838x system based on the 3739 family of test sets. The software implications are minimal: all of the items discussed elsewhere in this chapter still apply. The hardware is, however, obviously configured somewhat differently. The purpose of this section is to cover the hardware configurations and the measurement implications.

Broadband/mmWave Setup Based on the 3743x/3744x/MA25XXXA Modules

Using the modules listed above, the 3739x test set, the 3736B controlling test set, and a standard MN4697B/C four port test set with the VNA, a full 4-port system can be created.

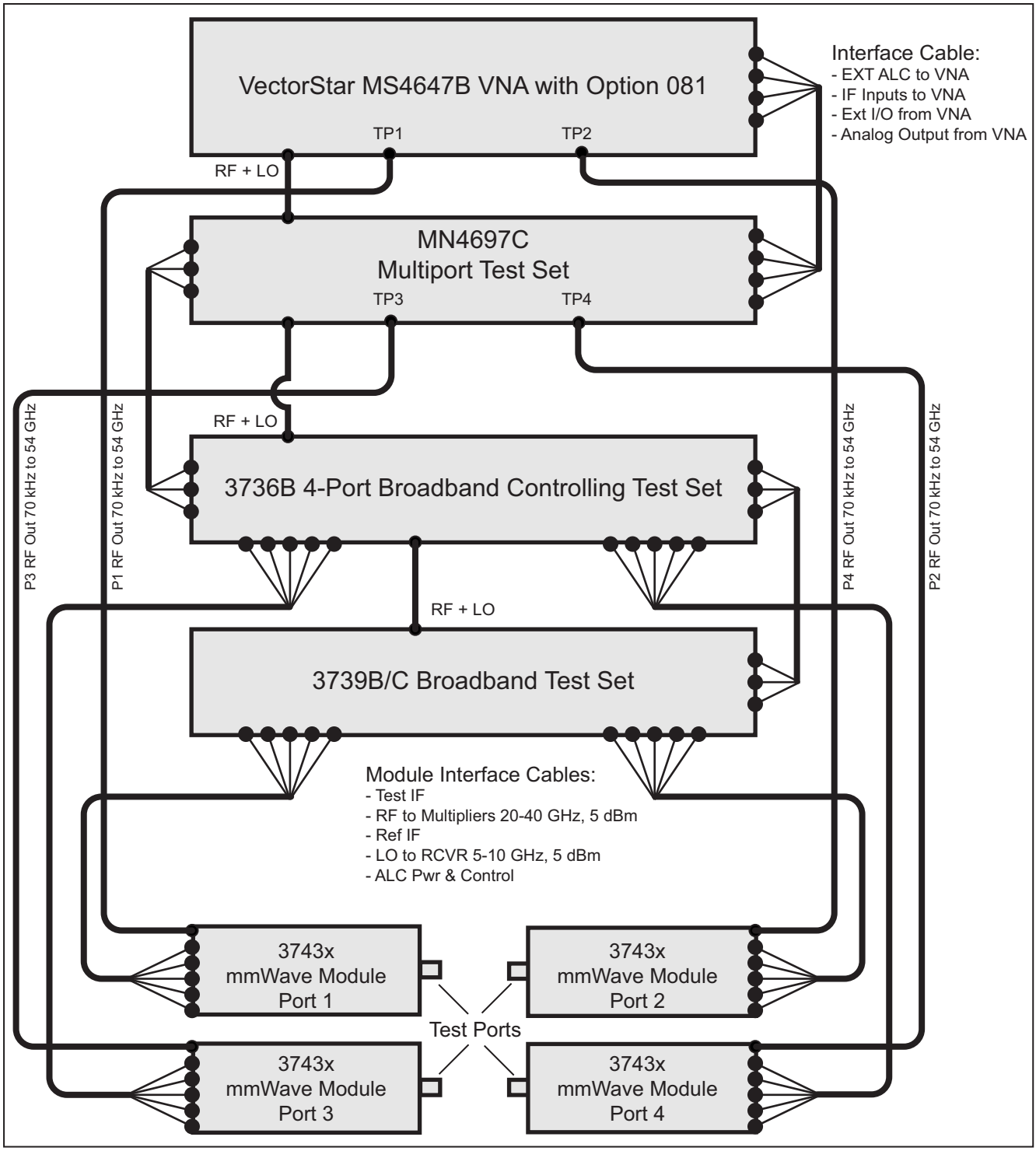
With regard to the 373x mmWave tests used in 4-port systems, we must introduce the concept of the 'controlling' test set (3736B). The other 3739x test set (for the other two modules) acts as a controlled test set and can be a regular, unmodified test set as used in a 2-port system. The 3736B controlling test set mediates control signals between the VNA and the modules, thus making the upgrade process from a 2-port system simpler. In all cases, the 3736B controlling test set feeds modules for ports 3 and 4 while the 3739x test set feeds modules for ports 1 and 2.

Two block diagrams of the system (broadband case) are shown below. Figure 24-71 is shown using an MN4697C and Figure 24-72 is shown using a MN4697B. The modules behave the same as discussed in Chapter 17 of this guide, they:

- use a direct drive path (from the 4-port test set in this case) to provide low frequency stimulus
- receive RF from a test set for high frequency stimulus
- receive an LO for all conversions within the module (>30 GHz)
- return IFs to the VNA for those local conversions
- have a control cable for power, control, and leveling signals

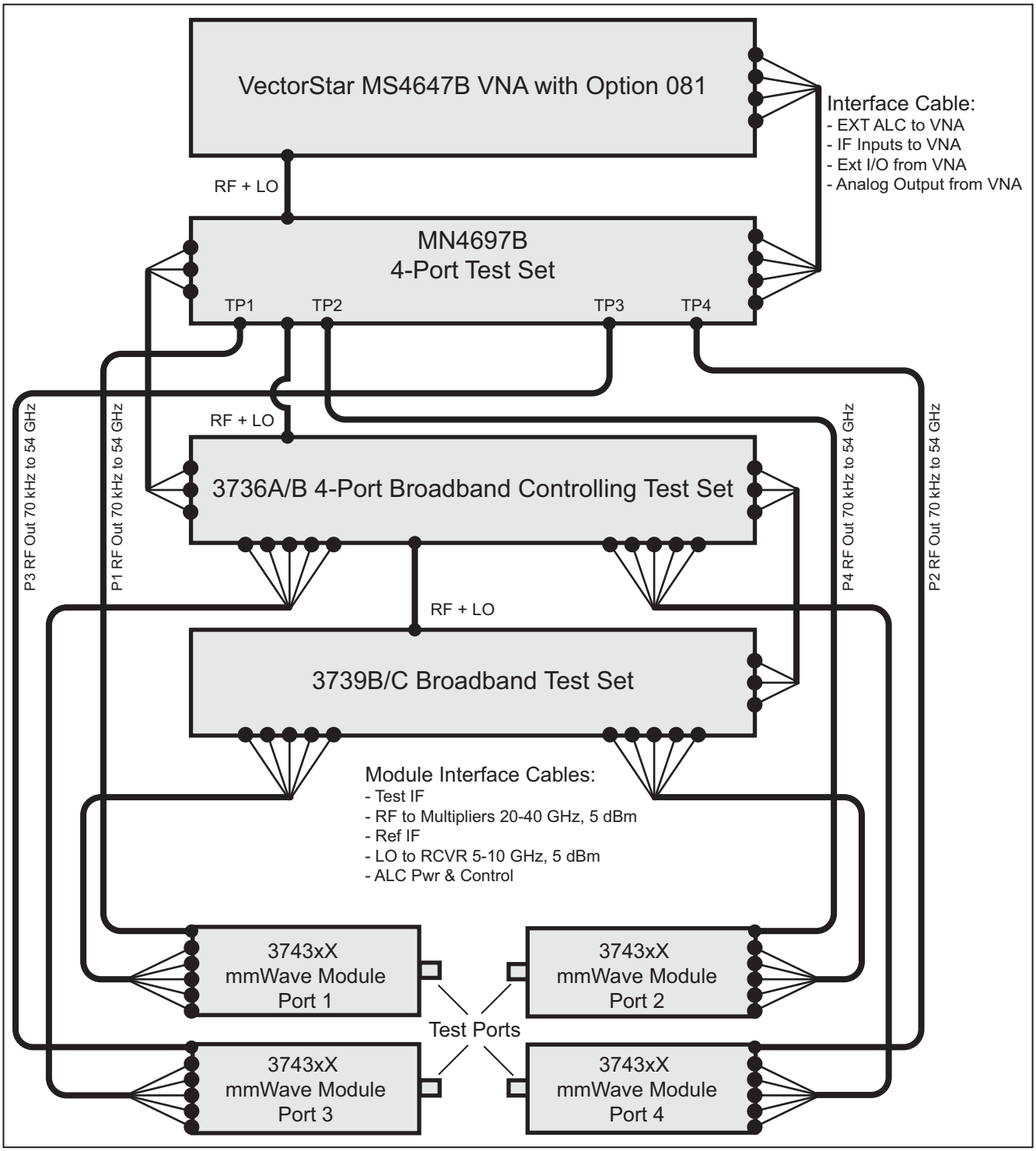
The only real difference is some multiplexing and control manipulation that now occurs in the 3736B controlling test set. From the point-of-view of the VNA and the user, this multiplexing is essentially invisible.

On the RF and LO side, there are splitters used to distribute the signals and there is enough power margin that available port power and conversion stability are not affected. The transfer switches within the test sets are configured to maintain isolation for all S-parameters. To help understand this distribution, the internal block diagram of the 3736B test set is shown in Figure 24-73.



Other ME7838x4 systems are the same except for the use of different modules.

Figure 24-71. ME7838A4/A4X—Four-port Broadband System with **MN4697C** Multiport Test Set



Other ME7838x4 systems are the same except for the use of different modules.

Figure 24-72. ME7838A4/A4X—Four-port Broadband System with MN4697B Multiport Test Set

RF and LOs come from the Option 8x connectors on the VNA front panel. A split portion of these signals is used in the controlling test set and a split portion is sent to the standard 3739x test set. IFs and ALC detection signals are returned from the controlled test set, multiplexed with those from the controlling test set, and then returned to the VNA. Additional multiplexing and LO multiplier sections are shown in this test set that are used for higher mmWave bands, as will be discussed shortly. The MN4697B/C used to provide additional control signals to the controlling test set test set in most situations via a separate DB9 connector (now these system are controlled with additional signals in the main rear panel broadband/mmWave harness; if the system does not come with a DB9 cable, the revised harness is used). These signals are the port configuration logic signals normally associated with any 4-port measurement, so they are needed to control signal separation in the mmWave modules as well.

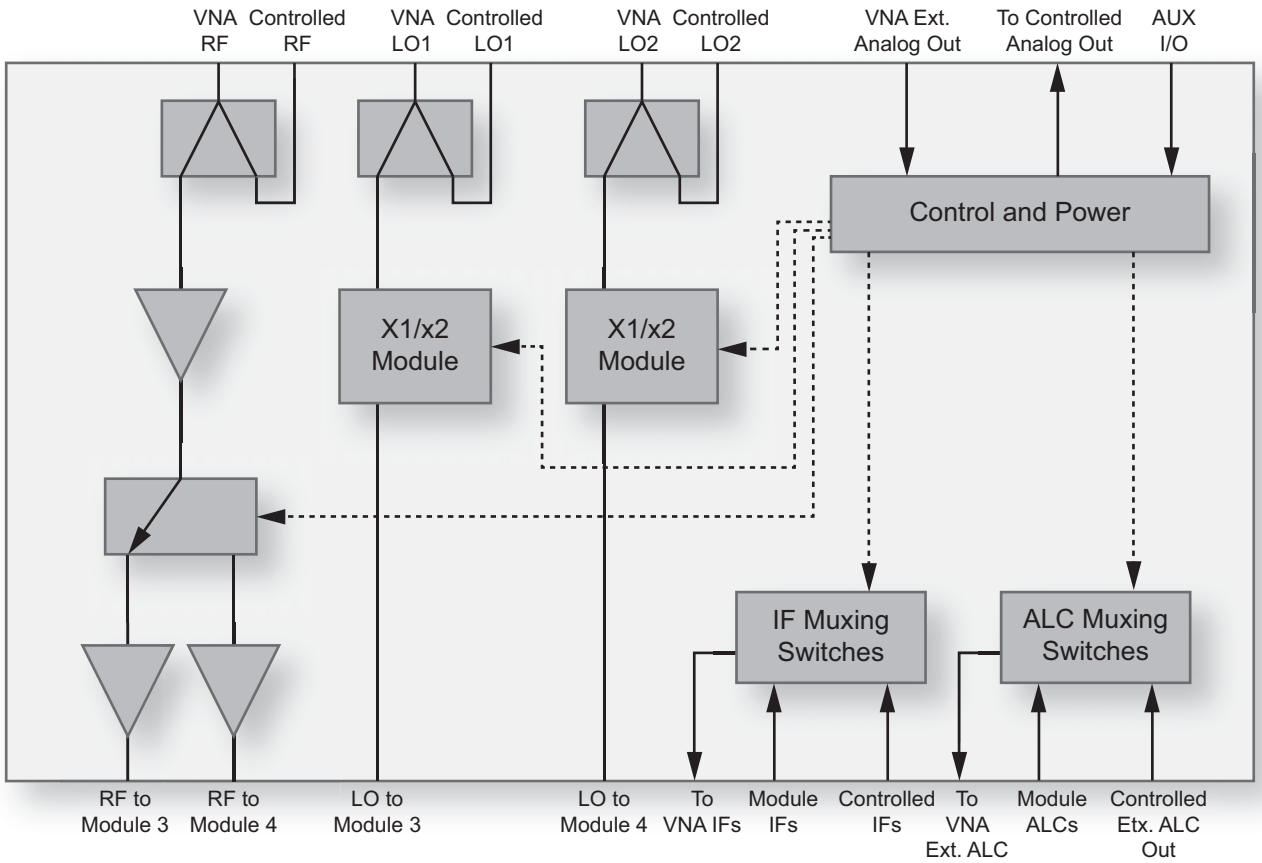


Figure 24-73. 3736A/B Controlling Test Set Block Diagram

The performance of the system is very similar to that seen in the 2-port broadband system discussed in Chapter 17. Because of the path loss through the 4-port test set, there will be less power available towards the top part of the direct drive range (54 GHz), but the higher frequency maximum power levels will be the same. A typical plot of maximum port power is shown in Figure 24-74. There is dynamic range reduction by about 6 dB at 54 GHz because of this power reduction. In addition, there is a few dB dynamic range reduction just below 30 GHz (and proportionately less at lower frequencies) due to some added loss in the baseband receiver path.

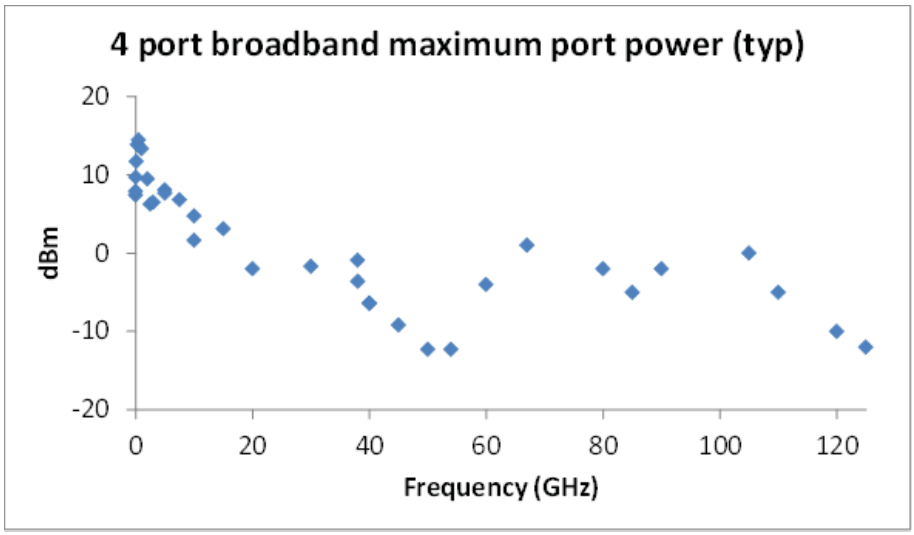


Figure 24-74. Broadband 4-Port System Maximum Port Power (typical)

Measurement stability remains excellent with the 4-port broadband system and example plots are shown in Figure 24-75 over a 24 hour period. In this case, the data was normalized at the start of the experiment and the deviations from those values over the 24 hours are plotted. As might be expected, the stability in a 4-port measurement is even more cable-sensitive than is that in a 2-port measurement. Minimizing cable runs (but not at the expense of too small a bending radius) and external temperature influences can help.

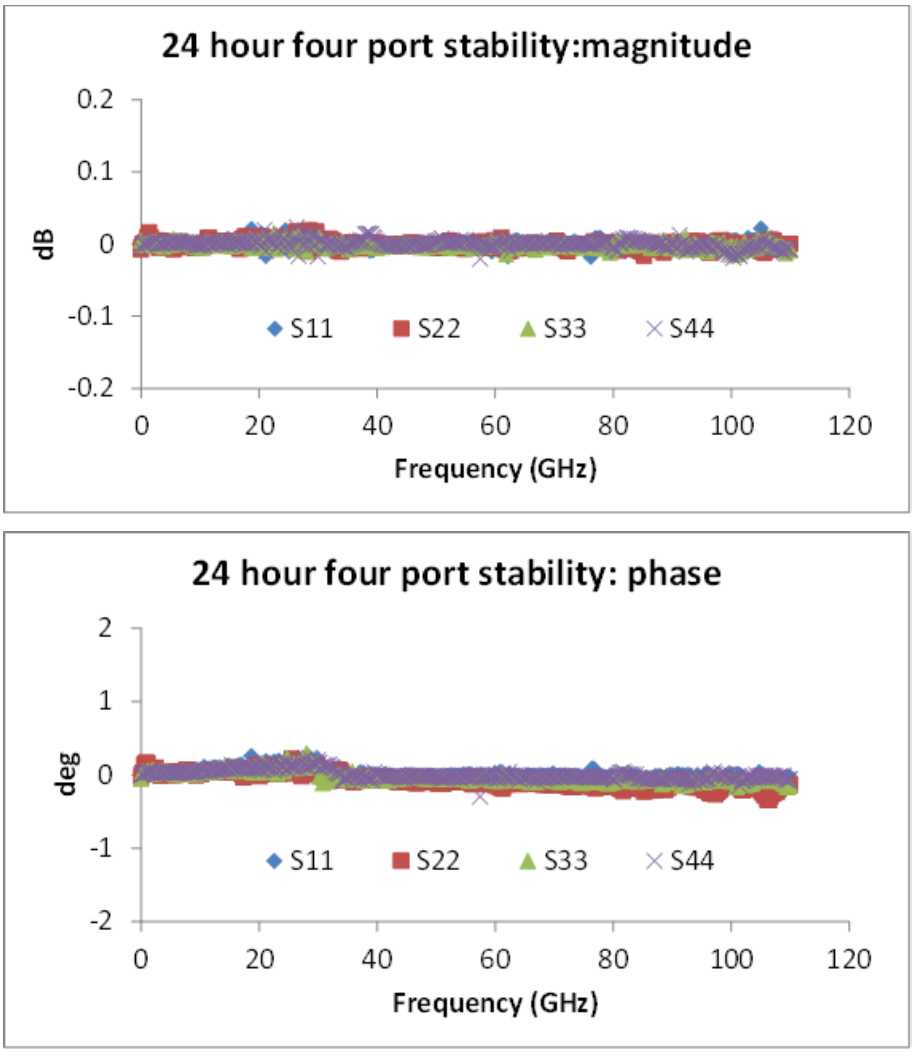


Figure 24-75. Broadband 4-Port System Stability (typical)

The calibration choices remain the same as for traditional 4-port system although in practice, there will be more SSSR related calibrations performed with the mmWave modules. The algorithmic choices were described in earlier chapters. Hybrid and merged calibrations will also be used more often in a broadband environment. The main reminder is that for both hybridization and merging operations, the underlying calibration types must be the same (i.e., full 2-port calibrations combined into a hybrid 4-port calibration or a merge of two full-4-port calibrations).

Measurement uncertainties follow the discussions presented earlier in this chapter. In general, there are no changes from the broadband 2-port case except for certain DUTs with many connected ports and certain high loss situations at lower frequencies (in which case the load match contributions of the many non-direct ports can start to become significant).

The mmWave banded systems based on the 3744A modules are constructed essentially the same way as described here except the connections from the 4-port test set to the modules do not exist (that provide baseband stimulus and response paths). The MN4697C test set may not be required for these systems. Contact Anritsu for more information.

Although the examples above did not cover it, the other 4-port ME7838x4/x4X broadband systems operate similarly to the ME7838A4.

mmWave Systems Based on OEM Modules

Much as high frequency 2-port systems are possible based on OEM mmWave modules operating to 1 THz and beyond, 4-port systems can be constructed in exactly the same way. The control paths are identical to those discussed previously, except there are no connections from the 4-port test set directly to the modules and there are no control cables directly to the modules.

An example configuration is pictured in Figure 24-76. The cables shown in the diagram are the physical implementation of many of those seen in Figure 24-71. The control signals and IF cables are routed on the rear panel (not shown).

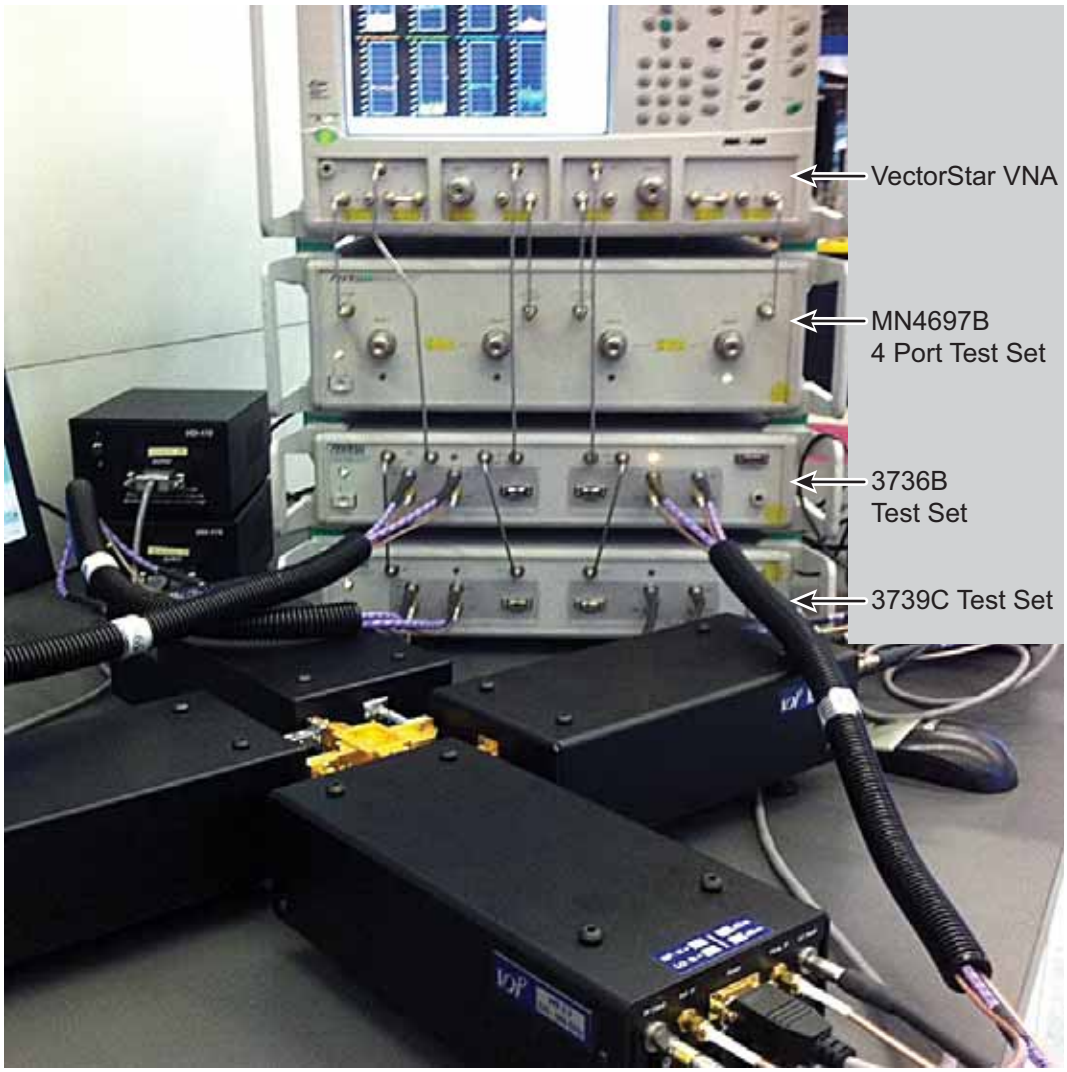


Figure 24-76. Example Four-port mmWave Configuration with 330 to 500 GHz WR-2.2 Modules

The performance of these systems is essentially the same as for the corresponding 2-port cases since there are no baseband paths involved. The RF and LO drive levels are such that saturation is maintained so port power and measurement stability will be maintained as well. Calibration choices are unchanged from the other 4-port situations, although waveguide calibrations will be predominantly used and merges are less likely. The waveguide-based reciprocal choices (often SSLR) will be helpful if the DUT configuration requires waveguide bends for connections. Uncertainties follow the discussion earlier in this chapter where the only analysis modifications occur on multiply-connected DUTs.

24-13 Saving SNP Data Files

While saving a .s4p file seems very natural with a 4-port structure, the saving of .s2p and .s3p files raises some questions (such as, which ports?) as does the desire to save possible mixed-mode parameters in the same file format.

SNP Setup Dialog

The .sNp setup dialog (see Figure 24-77) helps in the matter of which ports are used for the lower order files. Since the system cannot know which ports are necessarily of the most interest to the user (particularly when a higher order calibration is applied), the pair or triplet to be saved must be specified on this dialog. Of course, the port used for .s1p file save must be similarly specified. If a port combination is selected that is not currently calibrated, available data for the relevant parameters will be saved but may not be physically meaningful. Parameters corresponding to a trace that is currently on-screen will have current data sent to the file but it will be uncorrected. If the parameter is not currently displayed, data in an existing buffer will be saved but that measurement may have occurred previously. As with .s2p files discussed earlier, the data format may also be specified on this dialog. Navigation to the dialog: Main | System | Setup | Misc Setup | SnP Files Setup

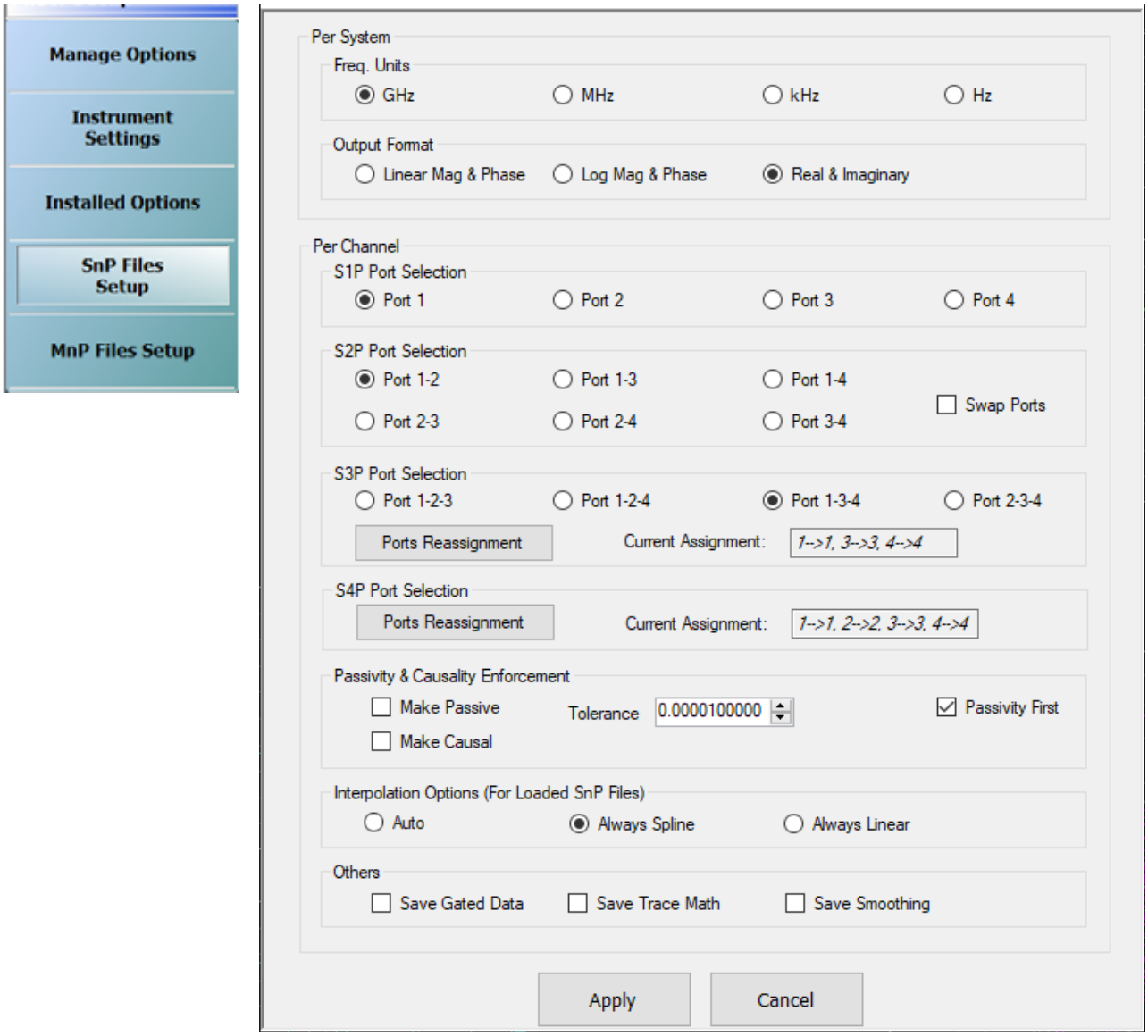
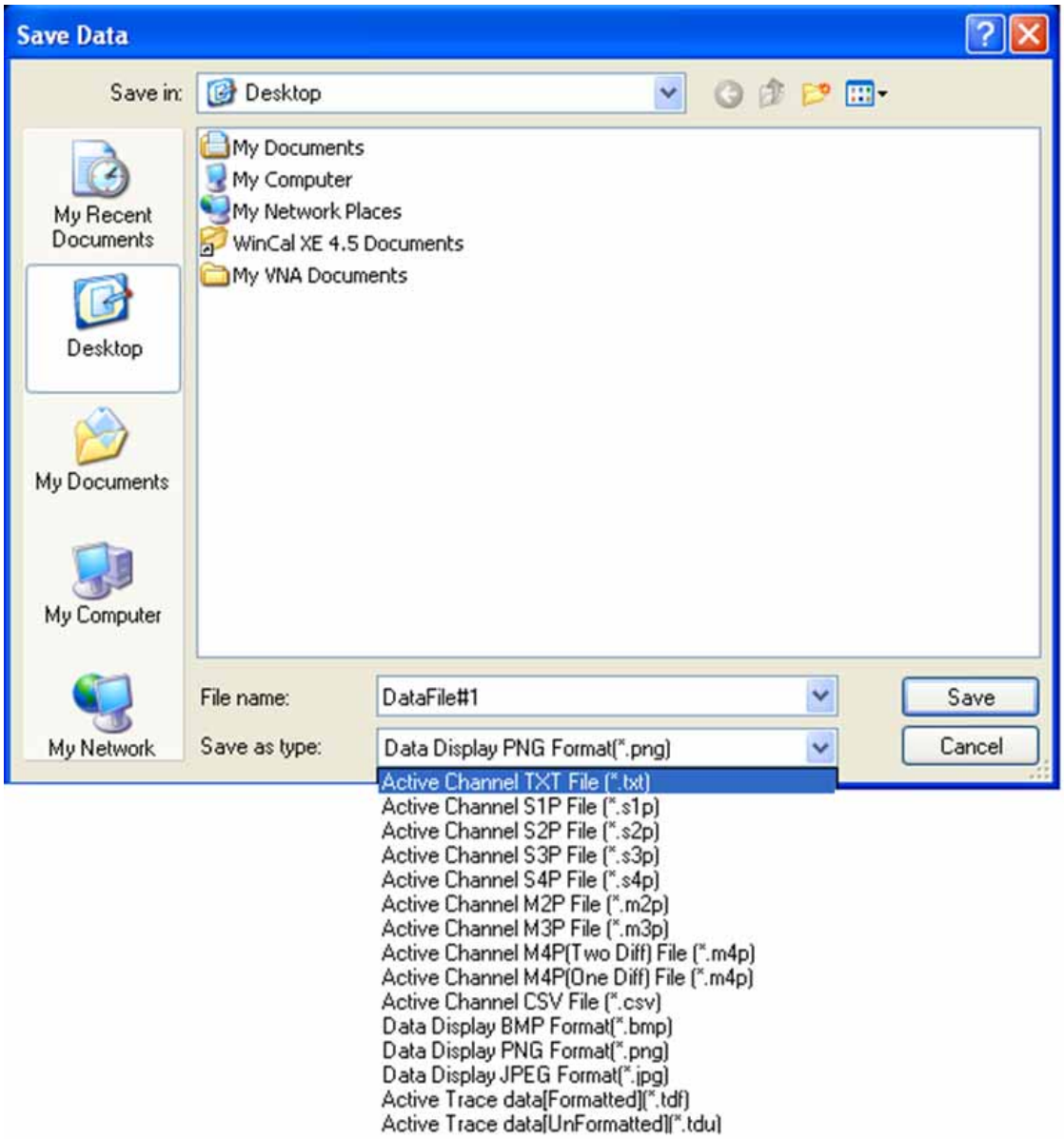


Figure 24-77. SnP Setup Dialog

MNP File Format

For mixed-mode parameters, another *file save* option is available: .mNp. The text file format is exactly the same as for the .sNp format but mixed mode parameters are used instead. The row-column orientation is what one would expect from the response menu dialogs but are summarized below.



The file save options are shown here to include the .mNp format. Note that there are two choices for .m4p depending on if the setup is for two differential pairs or one differential pair (and two singletons).

Figure 24-78. File Save Options

.m2p:	freq	SDD	SDC	SCD	SCC
.m3p:	freq	SXX	SXD	SXC	
		SDX	SDD	SDC	
		SCX	SCD	SCC	

(where the designator X refers to the singleton port)

.m4p (two diff):	freq	Sd1d1	Sd1d2	Sd1c1	Sd1c2
		Sd2d1	Sd2d2	Sd2c1	Sd2c2
		Sc1d1	Sc1d2	Sc1c1	Sc1c2
		Sc2d1	Sc2d2	Sc2c1	Sc2c2
.m4p (one diff):	freq	SXX	SXY	SXD	SXC
		SYX	SYY	SYD	SYC
		SDX	SDY	SDD	SDC
		SCX	SCY	SCD	SCC

(where the designators X and Y refer to the singletons)

Figure 24-79. The .mNp Parameter Ordering Structure

Enforcing Passivity and Causality

As discussed in Section 9-7 of this guide, enforced passivity and causality in saved .sNp files is possible. If either box is checked (Figure 24-77 on page 24-90), the data is coerced before saving so caution should be exercised. Enforced passivity uses a re-scaling of the S-parameter matrix eigenvalues, in a self-consistent way, to ensure that the 2-norm of the S-parameter matrix is less than unit (and the tolerance value correlates to how much less than unity).

Enforced causality invokes the Kramers-Kronig relations to minimize energy present before time t=0. With good calibrations and measurement practice (and no absurd de-embedding choices), these enforcement practices have minimal data impact but may aid convergence when using the data in large scale time domain simulations. When measurement issues are present, the changes to the data can be significant and the cause should be investigated.

Trace Mem Math in sNp

Another .sNp saving choice is to include Trace Mem Math operations in the saved data. When this is selected and one or more traces have Trace Mem Math turned on, the modified data for the parameters with match applied will be saved to the .sNp file. This can be useful for saving normalized results. If trace math is not applied for every parameter (which would only happen for a .s4p file if 16 traces were active and all had trace math applied and each trace had a different S-parameter), the unmodified data for the uncovered parameters will be saved instead.

24-14 Summary

This chapter has focused on the differences in 4-port versions of the MS464xB VNA system relative to the standard 2-port versions. The first part of the chapter concentrated on differences in response selection/definition and calibrations. Many of the post-processing functions from the 2-port VNA carry over unchanged into the multiport measurement context. Graph types, marker readouts, and trace data readout are all trace local and will feed off the defined parameter. The same is true for time domain since this operates on any trace-level parameter (although it will not be useful generally for an unratioed parameter such as $b_{4/1}$ since phase is not well-defined, but this is no different from the 2-port case). In particular, time domain applies to all of the mixed mode parameters. Smoothing and parameter conversion fall in the same category. Finally, the 4-port variants for the broadband and mmWave systems were discussed with a focus on hardware/configuration changes and how those changes affected system performance.

24-15 References

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6. J. Dunsmore, "New methods and non-linear measurements for active differential devices," *2003 IEEE Int. Micr. Symp. Dig.*, June 2003, pp. 1655–1658.

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Chapter 25 — 2PulseView™ (Option 35 and Option 42)

25-1 Chapter Overview

This chapter explains the background, types, and methods of pulse measurement enabled by the MS464xB Series VNA.

- Option 35—IF Digitizer includes a fast IF digitizer to accomplish a variety of measurements with very high resolution.
- Option 42—PulseView™ includes software to control the four pulse generators as well as the user interface to perform pulse measurements.
- Combined, Options 35 and 42 provide the ability to generate singlet, doublet, triplet, quadruplet, and burst pulses and to perform pulse profile, point-in-pulse, and pulse-to-pulse measurements.

Optional Pulse Modulator Test Sets are also available to pulse the RF stimulus or to provide receive-side modulation; the latter being very helpful in certain antenna and related applications.

Option 36 is a memory upgrade for Option 35 that has 8 GB for data instead of the usual 2 GB, which allows for greater record lengths and longer acquisition times at the same sampling resolution. The impact of this addition will be discussed at relevant points in this chapter.

25-2 Introduction to Pulse Measurements

Pulsed S-parameter and related measurements are critical for many common applications:

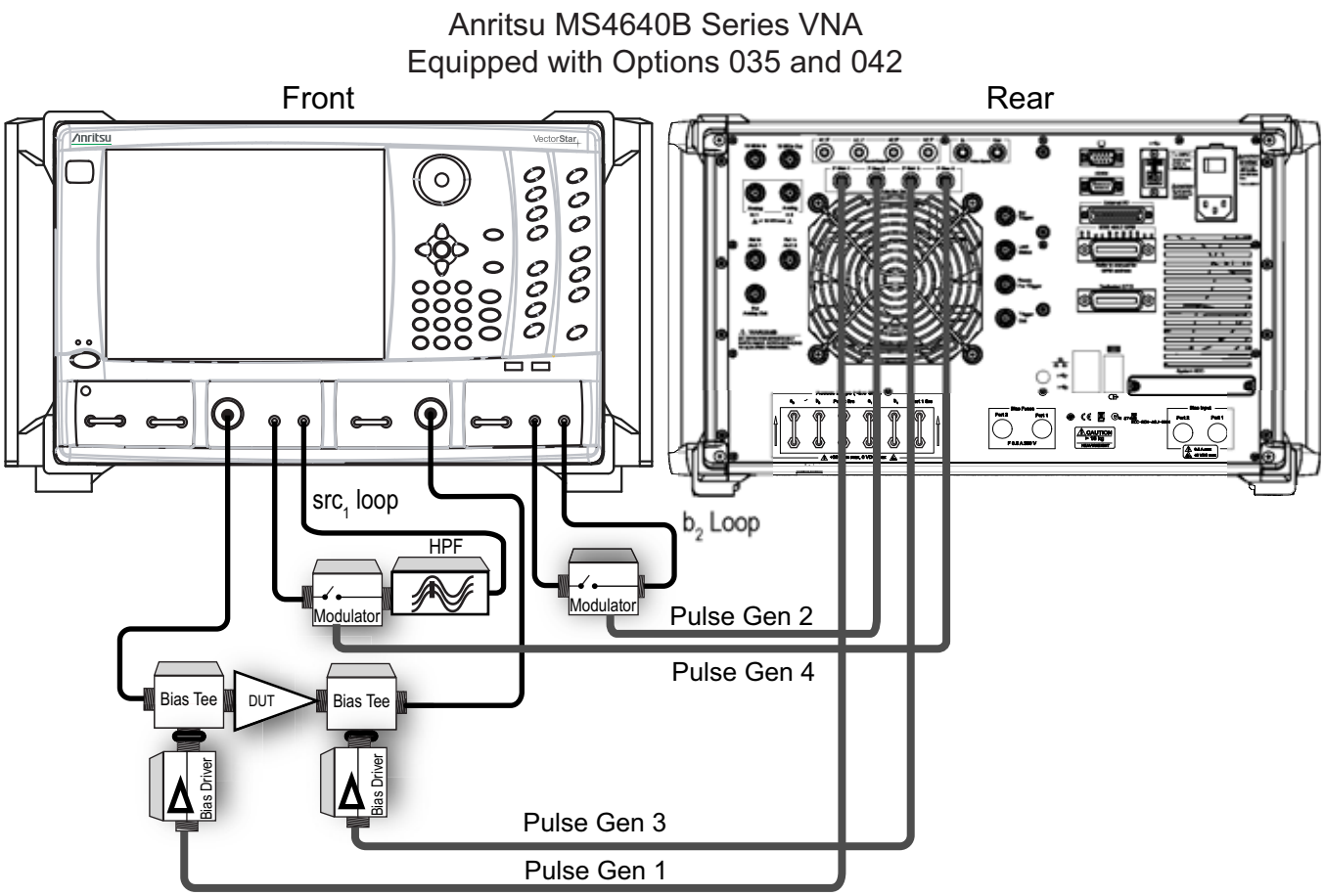
- Tests of system and subsystems designed to operate in pulsed conditions.
- Pulsed system and subsystem measurements that may be needed to examine phase and amplitude distortions at pulse onset or droop later in a pulse, as well as in cases where the subsystem was designed to operate pulsed.
- Measurement of devices and components that are designed to operate in pulsed conditions.
- Device characterization, where pulsed operation is required for thermal reasons (for example, un-mounted power devices).
- Device characterization, where pulsed operation is required to explore certain device characteristics. Trapping and thermal parameter characterization are two common tasks that benefit from pulsed measurement.

When making pulsed measurements for the applications listed above, many aspects of the test system may be of interest. A VNA is typically used when there is a need for measurement correction, phase data, or a more integrated test suite.

All of the standard VNA measurement parameters, calibration types and post-processing methods still apply to pulsed measurements, but there are many subtle factors affecting measurements that will be discussed and explained in this chapter.

Because of the wide array of applications and Devices Under Test (DUTs), many test configuration varieties involving pulsed stimulus, receiver gating, and pulsed DUT bias and control may be of interest. An example configuration is illustrated in Figure 25-1, and presumes S_{21} as the measurement of primary interest. Full correction is possible in this setup. Previous generations of Vector Network Analyzers have relied on receiver-side pulsing to control the measurement.

Note	Intra-pulse behaviors on the reflection parameters may often be ignored, depending on the measurement methodology. Figure 25-1 illustrates a setup without b_1 modulation.



Source RF pulses, multiple bias pulses, and DUT control pulses (not shown) may be needed.

Figure 25-1. A Comprehensive Pulsed Measurement Setup

25-3 Pulse Measurement Methods

Narrowband Measurements

There are many methods of performing pulsed S-parameter measurements and each has different attributes. One method is based on the spectrum of a pulsed RF signal (shown in Figure 25-2) and capitalizes on the fact that the central spectral line carries the magnitude and phase information of the underlying RF signal.

After filtering everything but this line, standard VNA processing can be applied. The time separation for the measurement (receive-side) must be done in hardware (for example, a modulator would be used for profiling on the receive-side). This approach is known by several names, including band-limited, narrowband, and others. The central spectral line is filtered off and used in the band-limited or narrowband method.

This line spacing is set by the Pulse Repetition Frequency (PRF), while the *sin(x)/x* null spacing is set by the pulse width. For low duty cycles, progressively less energy is contained in the central spectral line, which results in dynamic range reduction.

From a processing point-of-view, this method is attractive because it only involves one frequency. However, it does suffer from some significant limitations:

- Since only the central spectral line is kept, energy is being discarded, particularly as the duty cycle of the receive pulse shrinks. The result is a dynamic range penalty of 20 * $\log_{10}$ (duty cycle). Using this method, when the duty cycle reaches 0.1 %, 60 dB of dynamic range has been lost.

- Calibrations must be repeated for every different pulse configuration. (The same is true to a certain degree for all methods. However, because of the above dependence of signal level on duty cycle, the effect is extremely strong when applying this method.)
- The on/off ratios of the receive-side modulators can limit uncertainty. To illustrate the concept, consider the effect on the spectrum of imperfect isolation in a modulator between on and off states. When the pulse is supposedly off, there is information being retrieved that can be thought of as another pulse with a duty cycle equal to one minus the duty cycle of the main pulse.

  The spectra will be centered at the same frequency (the IF) and the comb spacing will be the same, but the lobe structure will be different because of the different duty cycles. The result will be an interfering (different phase) central spectral line. The interference causes an increase in measurement uncertainty. As might be expected, the uncertainty impact gets worse for low duty cycles and poor on/off ratios.
- Receive-side modulation is needed on all parameters of interest. If full term correction is needed and the DUT exhibits notable intra-pulse behavior on the reflection parameters, then multiple test-side modulators are needed, even for an S_{21} measurement. Two source-side modulators would normally be required for full term corrections, unless pulse behavior in the reverse direction was not of interest or concern. The rise-time of those modulators and their associated bandwidths can limit resolution.

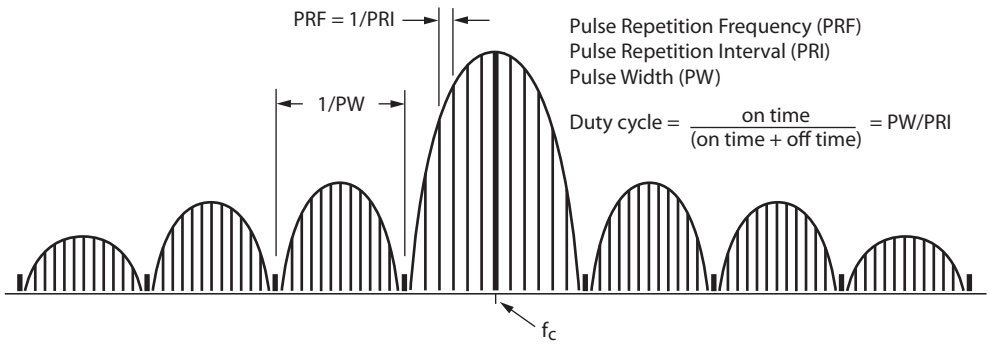


Figure 25-2. Narrowband Spectrum in Frequency Domain.

A setup for use in profiling applications is shown in Figure 25-3. While an RF modulator is shown here on the receive side (receiver gating), an IF modulator could also be used. In this case, a minimum resolution limitation is incurred if the energy distribution is limited by the IF bandwidth.

In this setup, the receive-side modulator could be omitted if the average performance over the entire pulse is of interest (for example, in some cases where pulsing is done only to prevent the device from overheating and a conservative duty cycle is employed). The source-side modulator could be omitted if bias pulsing only is being used.

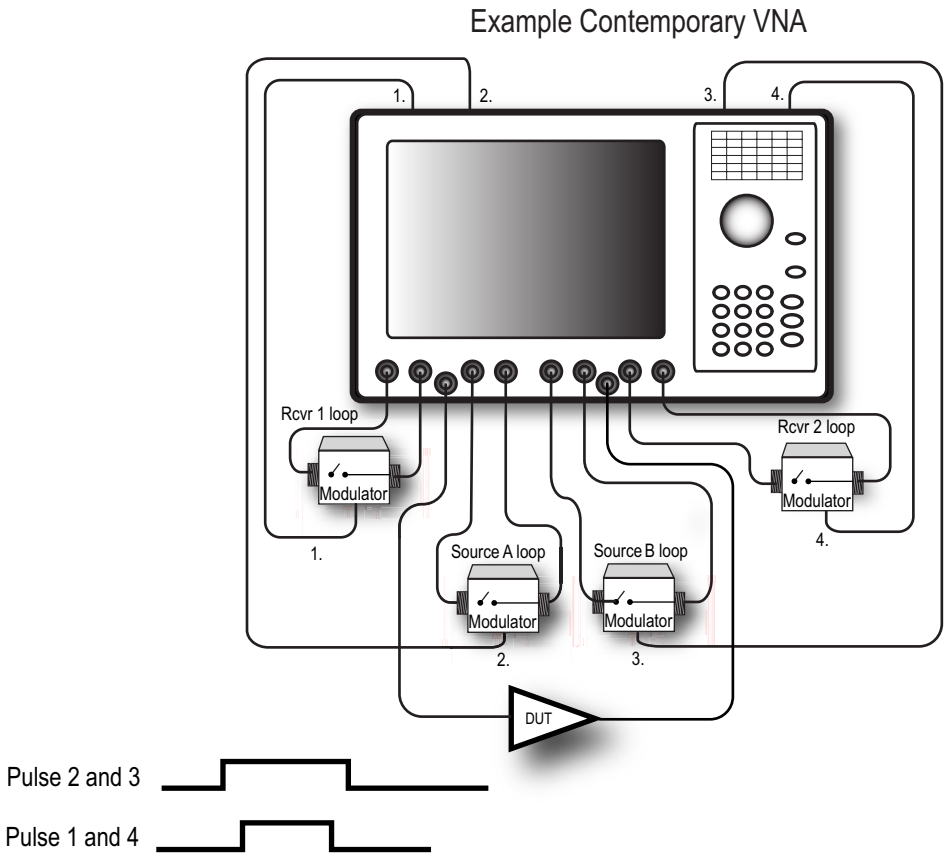


Figure 25-3. Example Setup for Band-limited or Narrowband Profiling Measurements.

Two example spectra are shown in Figure 25-4 for 80 dB and 40 dB on-off ratios (isolation) for a pulse with a duty cycle of 10 %. In this case, the error on the central spectral line amplitude is nearly 1 dB with the lower on/off ratio.

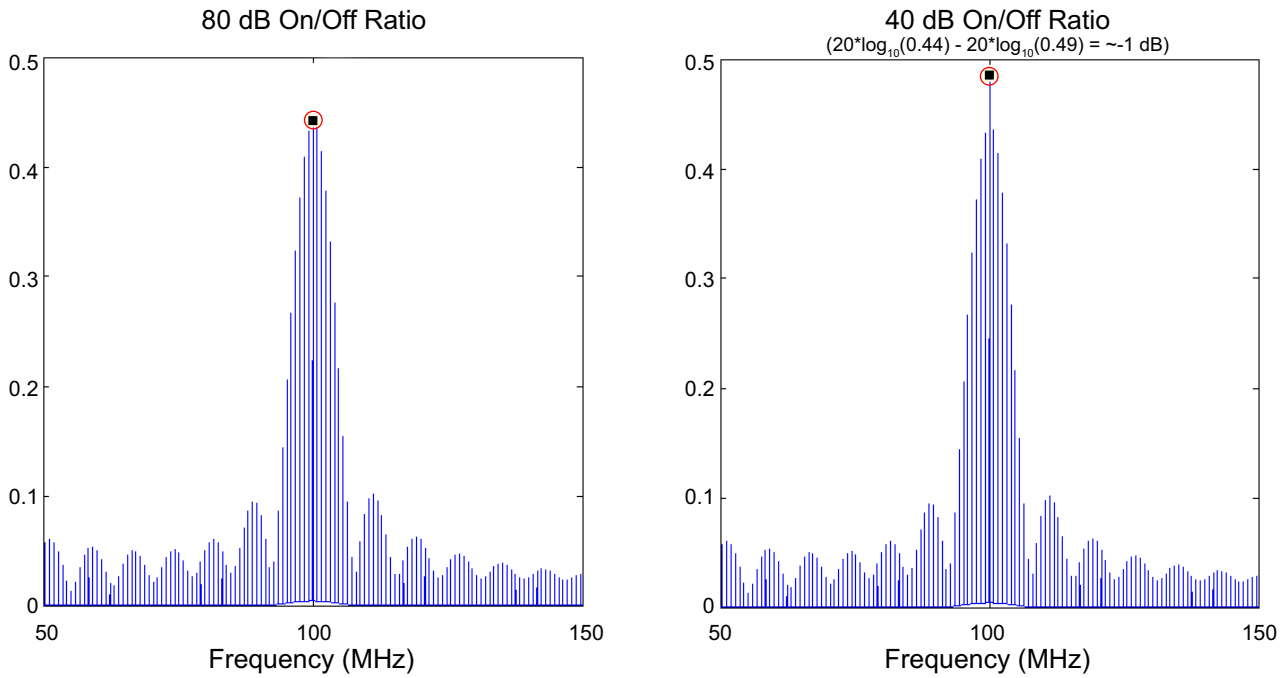


Figure 25-4. Effect of Different On/Off Ratios for Receive-side Modulation in the Band-limited Method.

Note	In Figure 25-4, the central lobe (where the information is extracted in this method) is noticeably affected and in error by nearly 1 dB.

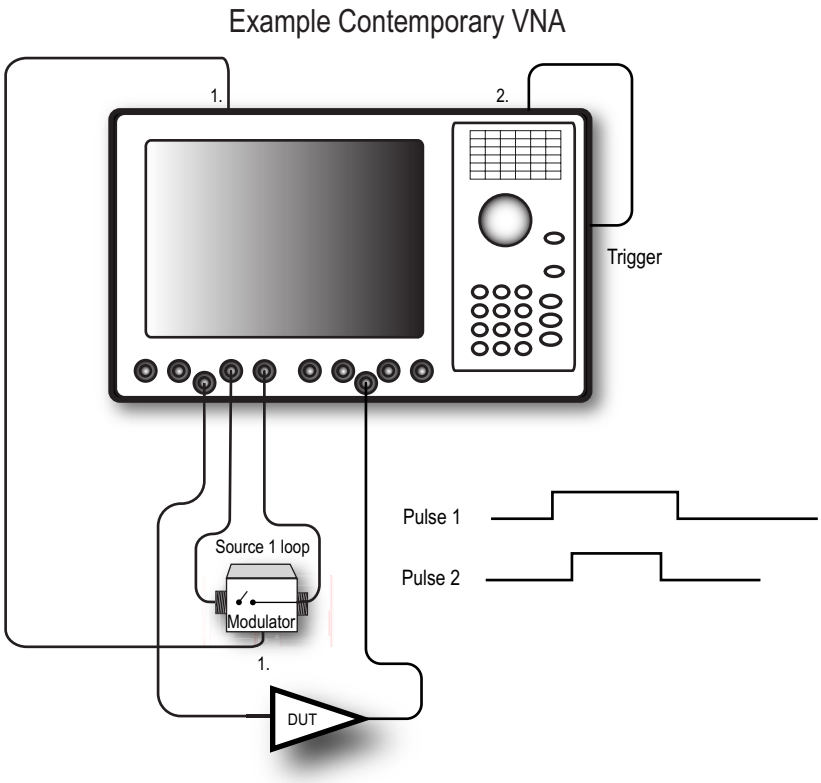
Triggered Measurements

Another method, termed a 'triggered' measurement, has been used mainly for slower repetition rate situations, where the VNA measurement is triggered to make classical measurements on sequential pulse (or synch pulse) rising edges. This method avoids the duty cycle dependence of the band-limited method and can allow for simpler profiling, if one has sufficiently precise control of the delay between trigger event and the actual measurement.

Trigger indeterminacy in contemporary VNAs is often large, resulting in a minimum pulse width that can be analyzed (often on a scale of tens or hundreds of microseconds). A second challenge lies in the data acquisition rate of many VNAs. VNAs have not traditionally incorporated precision high-speed digitizers, leading to poor timing resolution and inadequate profiling accuracy.

Calibrations often must be repeated for different parameter setups since the acquisition is being reconfigured. An example triggered measurement setup is shown in Figure 25-5.

Receive-side modulators are normally not used in this scenario, except for certain (usually antenna-based) measurements, where leakage pulses can overwhelm the VNA receivers. This class of measurement will be discussed later.



Historically, this setup has limited measurements to relatively low PRFs and wide pulses.

Figure 25-5. Typical Triggered Measurement Setup

High-Speed Digitizer Measurements

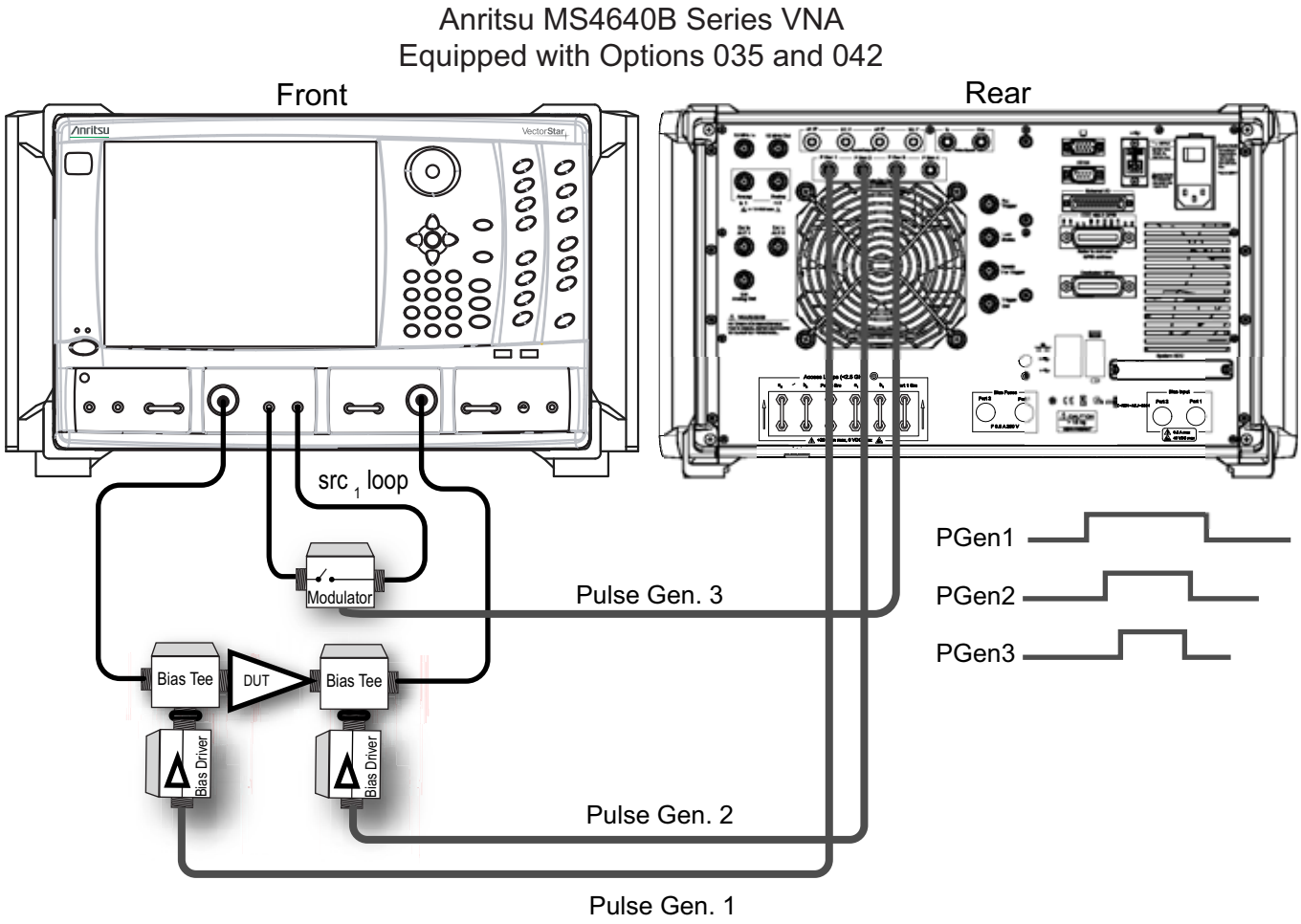
A variation on the triggered measurement technique is the method implemented in the MS464xB Series VNAs. It is also based on direct acquisition, but at a much higher data rate. As a result, resolution is enhanced and time referencing is highly accurate. The exact acquisition method differs in that the acquisition is not controlled by the pulse system, but by the portions of the data record to be analyzed. The acquired data is then correlated with the pulse pattern being used.

Note	Pulse-to-pulse measurements are an exception to this method. This type of measurement is covered in “Pulse-to-Pulse” on page 25-11.

An example test setup is illustrated in Figure 25-6. Two bias supplies were used in this example along with stimulus pulsing. However, a different number of pulsed bias supplies could be used or stimulus pulsing could be removed. Receiver pulsing may be used as well.

The MS464xB Series VNA with Options 35 and 42 provides a four channel internal pulse generator and configuration software for driving bias pulsers, DUT control, and RF stimulus modulators (available in the pulse modulator test set) or RF receive-side modulators for special applications (in the four modulator versions of the test set; for more information on the pulse modulator test sets, refer to “Setup Examples for Pulse Modulator Test Sets” on page 25-15). An input/output of the T_0 synch pulse is also provided so external pulse generators can be synchronized with the internal system (as might be needed in more complex bias situations).

No pulse generator channels are used to control the pulse measurement or profiling directly; this task is handled internally. The MS464xB Series VNA may be connected to an external trigger for pulse-to-pulse measurements and related measurements when the behavior relative to an absolute start time is important.



An example setup of the fast acquisition method of the MS464xB Series with Options 35 and 042. (Not all of the pulse connections shown must be used, and additional connections may be employed with this setup.)

Figure 25-6. Example Setup of the Fast Acquisition Method

Time records of the IF data are acquired at a high sampling rate, enabling the pulsed signals to be dissected into any of the desired pulse measurement modes. The resolution is on a nanosecond scale and the timing accuracy is very high without any duty cycle/dynamic range limitations on performance.

By using a fast digitizer and performing the alignment with pulse data in a post-processing sense, trigger latency issues associated with triggered measurements and potential jitter/inconsistency problems with that triggering are avoided. The resolution is set mainly by the acquisition rate instead, which is near 400 MHz. The jitter level is on the scale of picoseconds. When an external T_0 synch is used, there may be an inconsistency on the scale of nanoseconds as the internal system must lock onto the external pulse.

Since no energy is being discarded in this setup, there is no duty-cycle dependence and the full dynamic range (>100 dB normally depending on IFBW, power levels and averaging) is available. Even for full correction, no receive-side modulators are required and the on/off ratio, bandwidth, rise-time and video limitations of those structures are no longer an issue. A deep memory (4 GB) is used to acquire long time records of up to one-half second. For sub-Hz repeat rates, the triggered method is an adequate choice as the jitter on that time scale will not normally be significant.

Pulse Measurement Terms and Synonyms

A number of other terms for pulse measurements are commonly used. This list will help to explain terms used elsewhere in this chapter.

Duty cycle or duty factor

Pulse width relative to the pulse period, often expressed as a percentage. Common values range from 0.01 % to 50 %. This may pertain to a stimulus RF pulse, a bias pulse, a receiver gating function, or a profiling pulse, so context matters.

PRI and PRF

Pulse repetition interval (= pulse period) and pulse repetition frequency (= pulse repetition rate). **PRI=1/PRF**.

Pulse width and delay

These two numbers together with the PRF uniquely define a pulse train; however, there are cases when there are multiple pulses per period (see below).

Note	The MS464xB Series always references delay to the rising edge of the T_0 synch pulse, so that global timing can be easily defined.

Pulse patterns (singlet, doublet, quadruplet, burst)

Defines what occurs once during each pulse period of the global clock. The default is a single pulse, hence the term 'singlet'. Sometimes more than one pulse is required per global clock period, hence other settings are available. “Burst” is a user-defined number of pulses within a period. All sub-pulse widths are the same and all intra-sub-pulse gaps are the same; see Figure 25-7. (Multiple pulses per period are particularly common in radar applications.)

Record length

The amount of time to be captured within a measurement. This is dependent on the PRI and duty cycle, the amount of averaging requested, and the measurement mode.

Synch pulse (T_0)

This pulse establishes global (the pulse denotes the beginning of the global clock period) timing and can either be generated by the MS464xB Series instrument or by an external pulse generator (in which case, the MS464xB Series can accept that signal as an input). For example, when coherence is needed among pulse generators, a common synch pulse is often shared between them. Synchronization of 10 MHz timebases may be needed as well.

T_{start} and T_{stop}

These terms apply to pulse profiling setups and define the time range of interest relative to the synch pulse for profiling. Together with the number of time points, the measurement window width, and the requested averaging, these define the acquisition and processing behavior for pulse profiling. T_{start} and T_{stop} can also be used as parameters for plotting data in pulse-to-pulse measurements.

Measurement width or Aperture

This measure of time describes what portion of the pulse should be analyzed in a given measurement. For point-in-pulse and pulse-to-pulse measurements, it is the section of the pulse to be processed (with a delay parameter specifying where, relative to T_0, this window should be). For pulse profiling, this width is 'swept' across the pulse with processing across the measurement width being performed at each time step requested between T_{start} and T_{stop}.

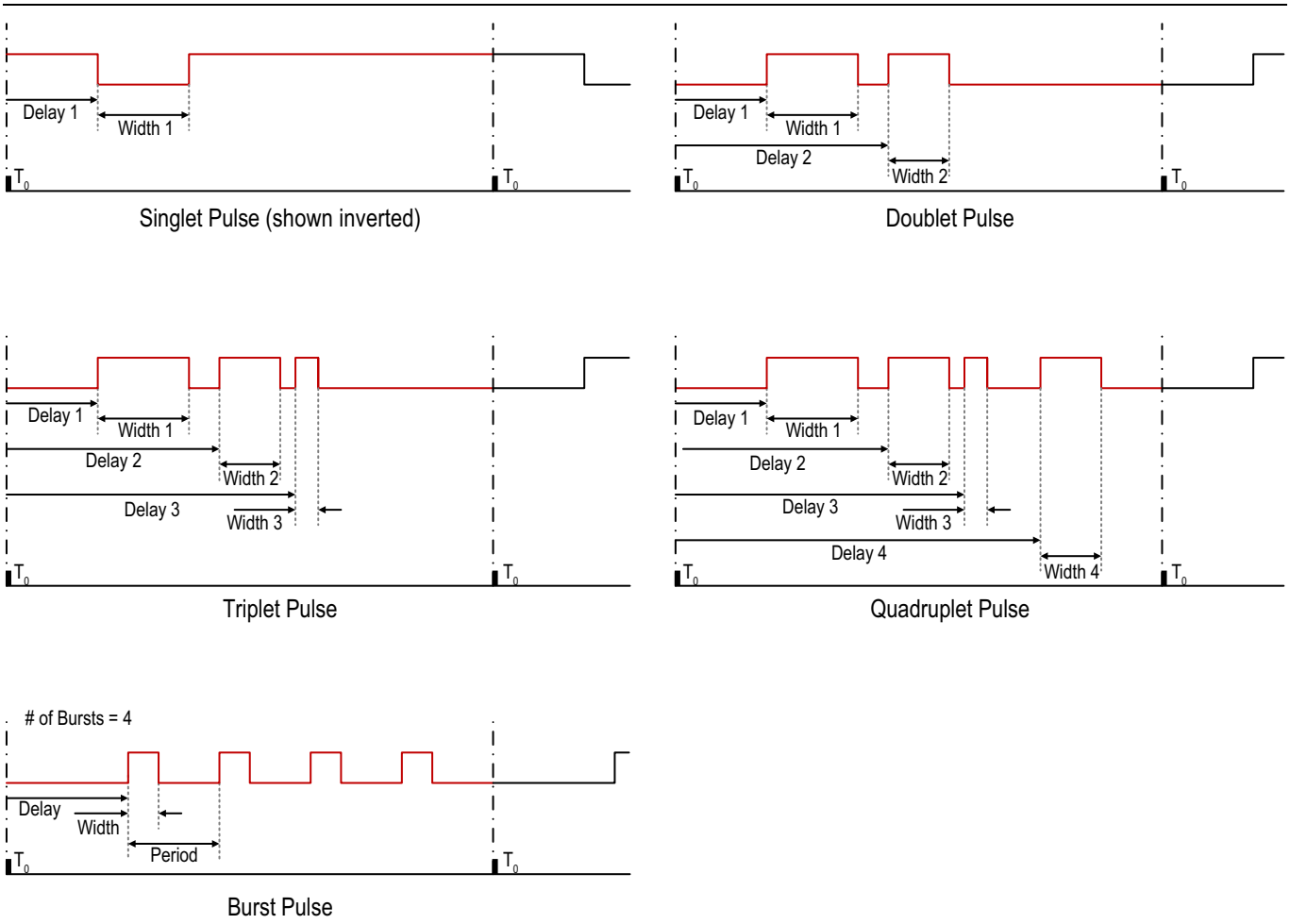


Figure 25-7. Timing Relationships for Multiple-pulses-per-PRI Signals

Measurement Modes

Pulse parameters and measurement definitions create an entire new parameter space to investigate. Many different terms have been used to describe these parameters in the literature; this document will try to conform to the most common usage. The terms will be defined in this section as used in the MS464xB Series.

Point-in-Pulse

This measurement quantifies S-parameter data somewhere within a pulse. One may want to avoid edge effects or just measure the pulse as a whole, but the structure within the pulse is not of great interest, nor is the variation from pulse to pulse.

A point-in-pulse measurement is illustrated in Figure 25-8. Data is acquired over a requisite length of time, and the user specifies the interval of interest relative to a synch pulse T_0. This interval is usually quantified in terms of a delay and a desired measurement width or aperture. In addition, a level of averaging can be specified so this same interval is analyzed on multiple pulses and the results combined. Pulses are sampled with the same coherent clock so that phase information is maintained.

The diagram in Figure 25-8 illustrates an example where the width of interest is eight samples. While only a single sample is required for measurement, about four hundred million samples are possible (limited only by the record length of the instrument). Point-in-pulse measurements are often made with swept frequency or power, and are plotted as such. In these cases, the above process is simply repeated for multiple frequencies and power levels.

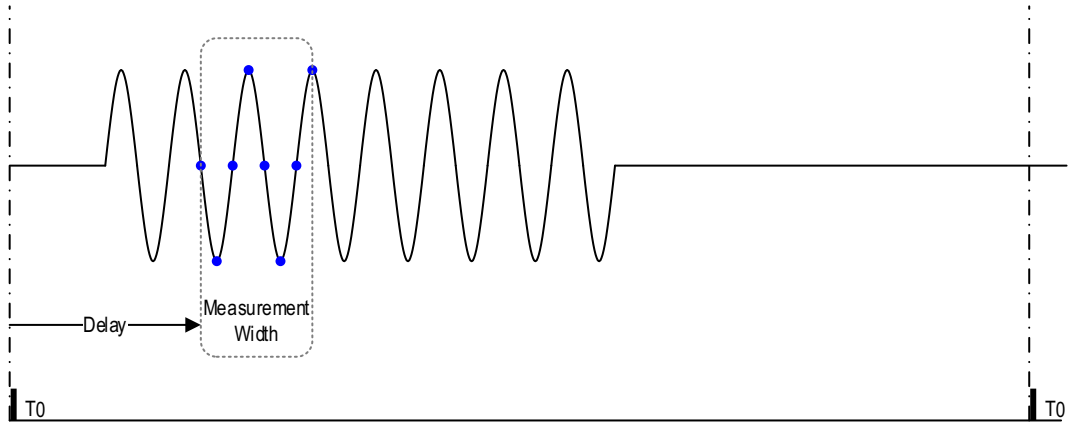


Figure 25-8. Simplified Point-in-pulse Measurement Example

Pulse Profiling

The pulse profiling measurement focuses on the structure of data within the pulse. Characteristics like overshoot or undershoot, droop, and edge response are measured, while the frequency and power are kept constant (although they can be varied between acquisitions).

This data is normally plotted vs. time to indicate position within the pulse (measurement normally is setup as constant frequency (CW) and constant power, but more complex measurements can be orchestrated using multiple channels or setups).

Variation between pulses is often not observable in this measurement, which may represent an average over a number of pulses. One can structure the measurement, however, to look at behavior from an absolute start time with no averaging in order to look at the complete evolution over multiple pulses.

A measurement example is diagrammed in Figure 25-9.

A start time and a stop time (T_{start} and T_{stop}) are specified relative to the synch pulse (T_0) along with a number of time points to describe the portion of the pulse to be profiled. The measurement may begin before the physical pulse is at the DUT and end after the pulse is no longer being applied to the DUT, but the data must be interpreted appropriately in all cases. A measurement width is also specified (illustrated as eight samples in the diagram). Additional averaging can be imposed across multiple pulses, as discussed previously. As with point-in-pulse, the range of allowed widths is large.

The measurement windows are allowed to overlap (and will if $(T_{stop} - T_{start}) / (NumTimePoints - 1) < Width$), which performs a variety of time-based smoothing operations on the data. A common setup is 10 to 20 non-overlapping (or slightly overlapping) profiling time points within a pulse, but application requirements may dictate very different parameter choices.

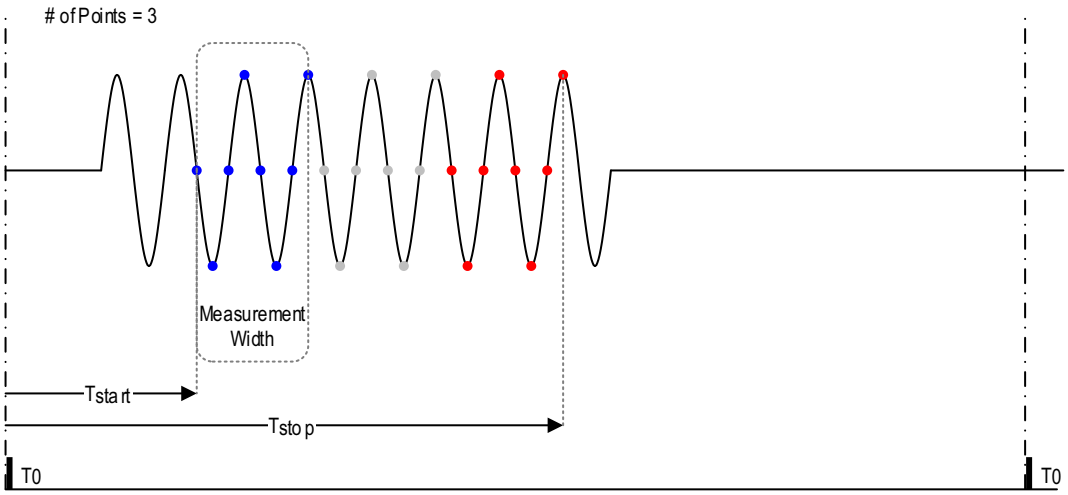


Figure 25-9. Example of Pulse Profile Measurement

Pulse-to-Pulse

Pulse-to-pulse measurements concentrate on variations between pulses that begin with an absolute start time. This type of measurement can be described as a hybrid of the first two methods. Normally one will pick a section of the pulse to measure, and then the progression over a number of pulses is plotted (versus time or index).

This process is illustrated in Figure 25-10, where the first three pulses are shown. The delay from T_0 is the same for each measurement window of interest and the window in each pulse is processed separately.

Pulses are treated in an absolute fashion, so averaging of multiple pulses is not useful and is not available. As with pulse profiling, this measurement is performed at constant frequency (CW) and constant power. One could orchestrate measurements that cycled through a variety of frequencies and power levels using multiple channels or setups.

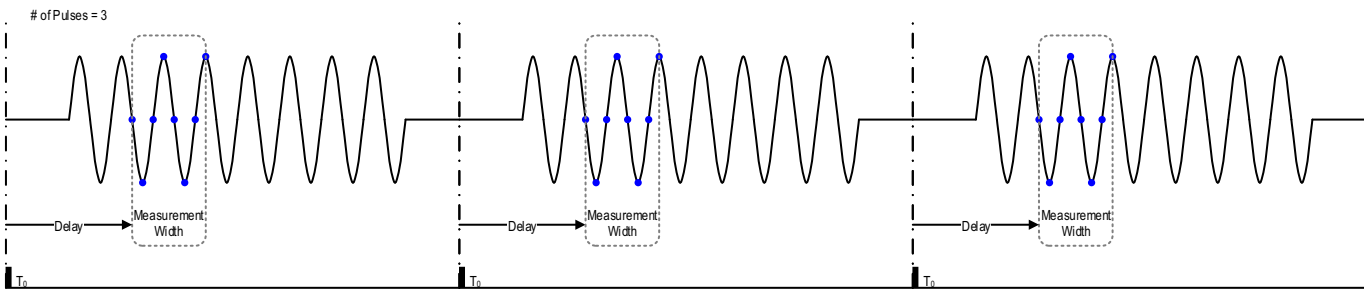


Figure 25-10. Simple Pulse-to-Pulse Measurement.

A wide range of window widths (and pulse widths) is available as in the previous modes. The upper limit is set by the record length with options for very wide pulse or low rep repeat rates.

Note	The minimum measurement for the pulse-to-pulse measurement mode is two samples.

Important Measurement Considerations

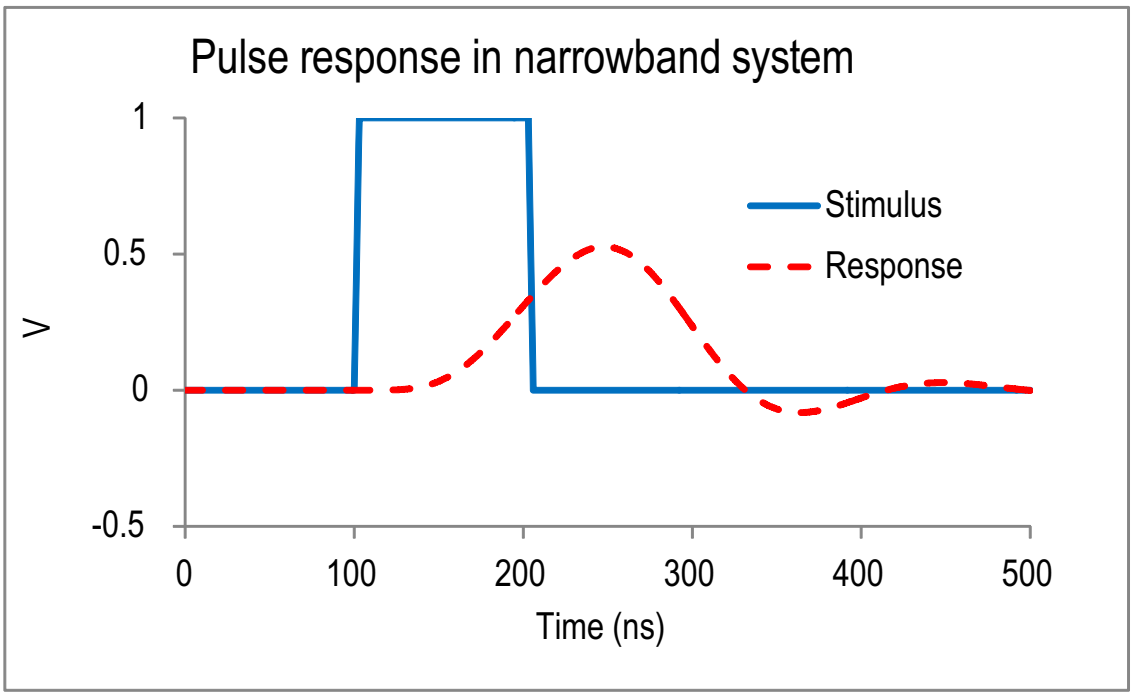
Consider the bandwidth implications of all measurement methods in the sense of the smallest measurement window or profiling width that can retain device information without being overwhelmed by system effects before moving on with measurements.

Band-limited Measurement Modes

- In the case of the band-limited method, the system bandwidth limitations are derived from the receive-side modulator response shape, which forms a minimum resolvable pulse.
- If this modulation is done at RF (as has been suggested so far), then the rise/fall time of that modulator defines this limit.
- If the modulation is done at the IF, the same concept applies but only when the modulation is considered in the context of the IF system.
- If the spectrum of the measurement setup in Figure 25-4 is distorted by band-limiting characteristics of the IF, then the system is distorting the pulse response.

An example of these effects can be demonstrated when a DUT output pulse with 100ns width (3ns rise and fall time) is sent through an IF system with a 5 MHz bandwidth.

The pulse response is illustrated in Figure 25-11. It is heavily distorted in this case, and a substantial ring-down is induced by this narrow IF system. The MS464xB equipped with Options 35 and 42 has a much wider equivalent IF and subsequent higher performance.



Example stimulus pulse and resultant response pulse distorted by a finite IF bandwidth in a narrow bandwidth system.

Figure 25-11. Example Stimulus Pulse and Resultant Distorted Response Pulse

Triggered Method

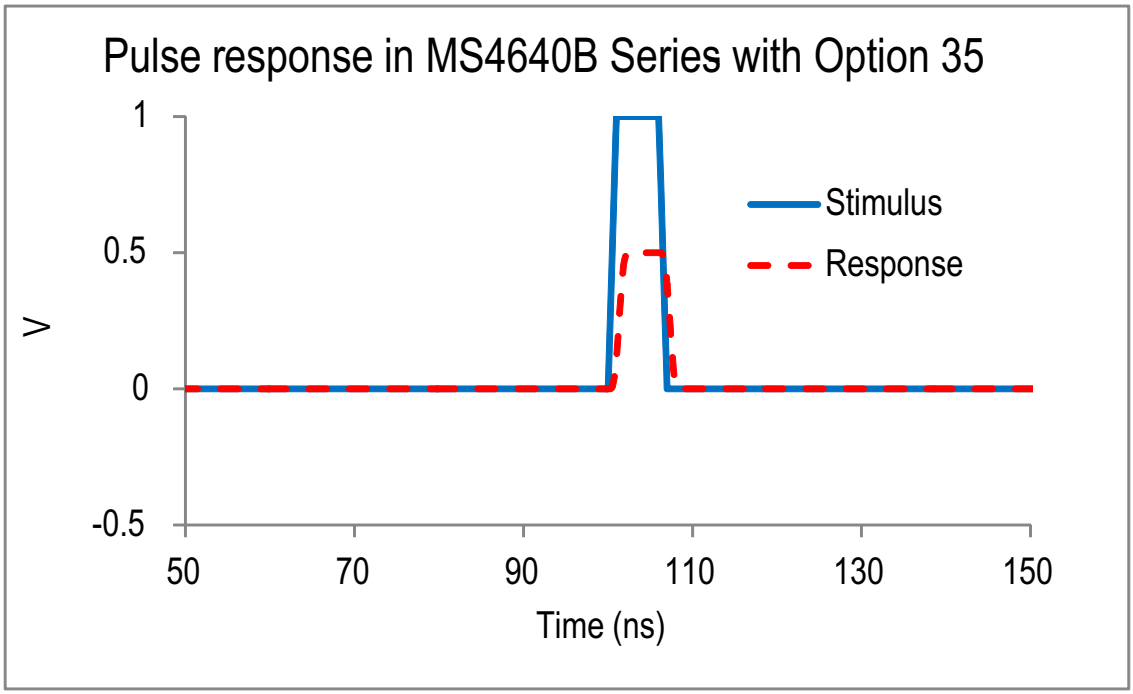
In the triggered method, resolution is generally limited by timing consistency of triggering and will be on a much longer time scale than either the band-limited techniques or the high-speed digitizing techniques of the MS464xB Series equipped with Options 35 and 42.

High Speed Digitizer Method

In the high-speed digitizer pulse measurement method, there is generally no receive side modulator (except in the special cases). In its nominal operating state, this measurement method uses an IF of 100 MHz with appropriate filtering to avoid image responses.

Figure 25-12 is a graph of the output of that system for an input pulse 5 ns wide with 1 ns rise and fall times.

Artifacts include some edge softening, but the distortion is relatively small (especially when compared to narrowband systems; see Figure 25-11). The amplitude was not scaled for this plot and the time shift is due to group delay in the system that must be accounted for in the measurement process.



The effect of the IF bandwidth on the pulse shape using the MS464xB Series -035/042 measurement method and a 5 ns wide DUT pulse with 1 ns rise and fall times.

Figure 25-12. Effect of the IF Bandwidth on the Pulse Shape

In the plots of Figure 25-11 and Figure 25-12, the output amplitude in both cases represents some loss through the test network and does not represent an effect of the processing method.

From these figures, we can conclude that a sufficiently wide IF, relative to the pulse shapes of interest, is important to ensure measurement of the DUT characteristics rather than those of the instrumentation.

25-4 Physical Measurement Configurations and Setups

The flexibility of the MS464xB equipped with Option 35 and Option 42 enables a very wide array of pulse generation and measurement configurations.

In the area of pulse generation, four internal pulse generators (with 3.3 V logic levels) are available for use anywhere in the test system. In addition, external pulse generators can be included in the test system and synchronized to the internal system to provide even more options. Pulses may be used to control bias, RF modulators, gain control systems, state control systems and a variety of other subsystems in the DUT or the measurement environment.

Pulse Generator Example

As an example, consider a power FET measurement where it is desired to pulse both gate and drain biases as well as the RF. Since this example is of a high power device, the sequencing of these events is important. Both for thermal reasons and for understanding bias transients, all of the biases will be pulsed. A drawing of the pulse relationships is diagrammed in Figure 25-13.

Three of the internal pulse generator channels are being used with the first two feeding bias switching circuitry (external to the VNA) and the third feeding an RF modulator (available in the pulse modulator test sets). All pulse generators share the same PRF but have independent widths and delays—all set with a resolution of 2.5 ns.

For this example, one may want to perform a pulse profiling measurement and T_{start} might be aligned with delay 3, or just prior (so profiling begins when RF is first applied to the DUT), and T_{stop} may be a slightly larger value than delay 3 added to the pulse generator's pulse width (profiling ending shortly after the RF is shut off).

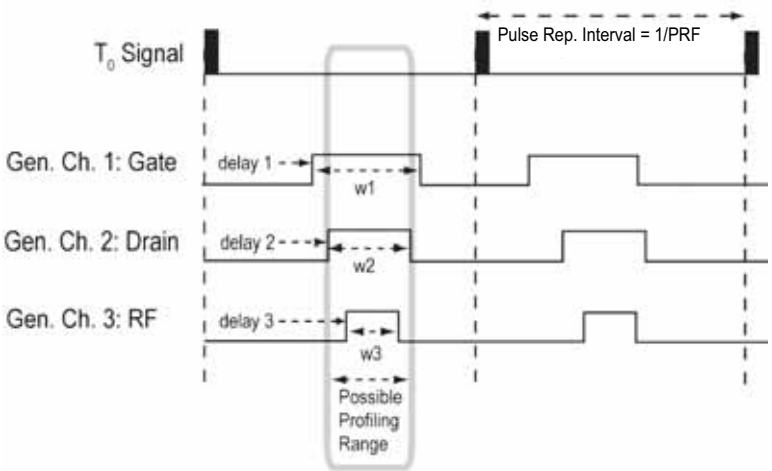


Figure 25-13. Example Pulse Generator Configuration

The pulse outputs are relatively high speed and use high current drivers, so impedances down to 50 ohms are supported. It is recommended to keep series resistance under ~10 ohms for common logic families when driving a 50 ohm load.

In the example above, each pulse was discussed as a single pulse per pulse repetition interval. Doublet, triplet, quadruplet and busted pulses are all permitted independently per generator and per channel. The timing relationships are shown in Figure 25-7.

From the point-of-view of the measurements, timing is still relative to the T_0 synch rising edge. For the doublet, triplet, and quadruplet situations, all of the widths and delays are independent (although not all are labeled in the figure for clarity). For the burst situation, all sub-pulse widths (w1) are the same and all intra-sub-pulse gaps are the same (defined by the period p and the width w1).

Setup Examples for Pulse Modulator Test Sets

For cases when RF modulation is needed, test set configurations are available. Pulse modulator test sets also include reference coupling, which can be useful in many applications. The block diagram of a fully configured test set (four modulators) is shown in Figure 25-14. The base model has only the source path modulators and is shown in Figure 25-15.

To cover a full 70 kHz to 70 GHz range with pulse modulation and to support the block diagram of the MS464xB Series, the modulation paths are split into above and below 2.5 GHz (high frequency modulator [HFM] and low frequency modulator, [LFM] respectively) with access ports on the front and rear panel for easy connection to the VNA.

As suggested by the block diagram in Figure 25-14, the source side modulators include additional reference couplers that can be fed into the reference loops of the VNA to provide ratioing against a pulsed signal. This can be useful to ratio out potential ringing effects of the modulator. In many cases, these coupled paths do not have to be used and all ratioing can be performed against the usual non-pulsed stimulus signals. Access loops are also provided to allow for amplification on the drive paths, which may be needed in higher power applications (for example several watts of DUT input power).

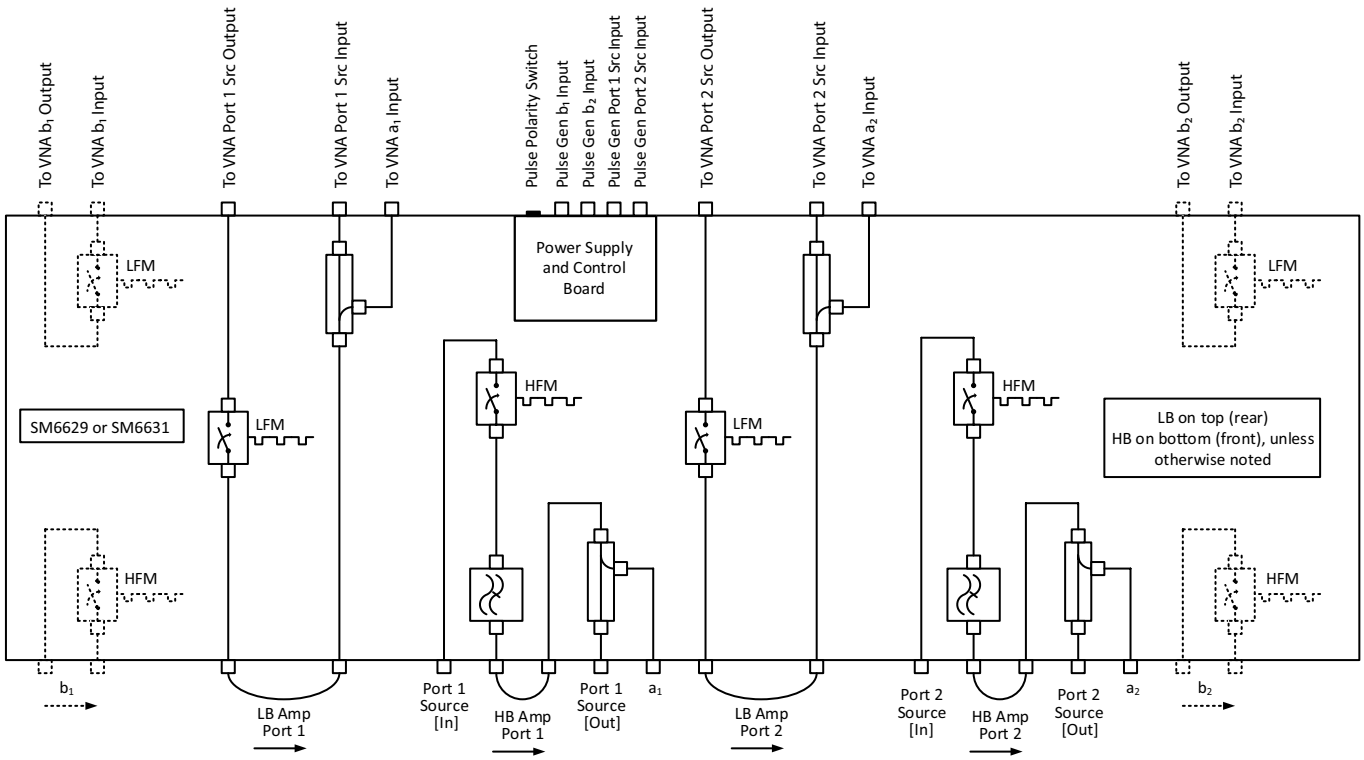


Figure 25-14. Block Diagram: Four Modulator Version of the Pulse Modulator Test Set

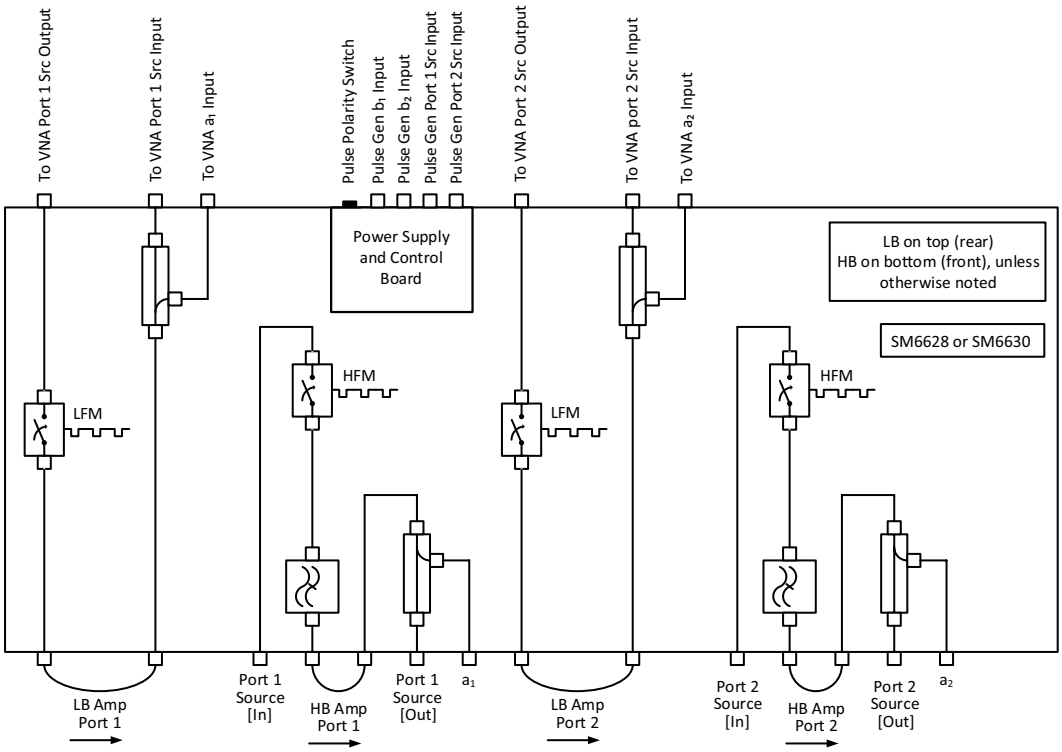


Figure 25-15. Block Diagram: Base model of the Pulse Modulator Test Set, Without Receiver-side Modulators

The most common applications will use only the source loops in order to provide stimulus modulation. In this case, only two cables per port are actually needed (two on the front for > 2.5 GHz and two more on the back for < 2.5 GHz). The source signal (post-ALC detection) is routed down to the test set and then sent back up to the VNA to pass through the test coupler. The extra reference coupler in the test set can be used to route into the reference receivers in the VNA for stimulus ratioing as discussed above.

The receiver-side paths of the pulse modulator test set are particularly useful in certain antenna applications where the transmit leakage pulse may be large enough to compress the receivers.

Note

Pulse modulator test set versions with maximum frequencies of 40 GHz and 70 GHz are available.

The video connections between the control board and the modulators in Figure 25-14 and Figure 25-15 are not shown for clarity.

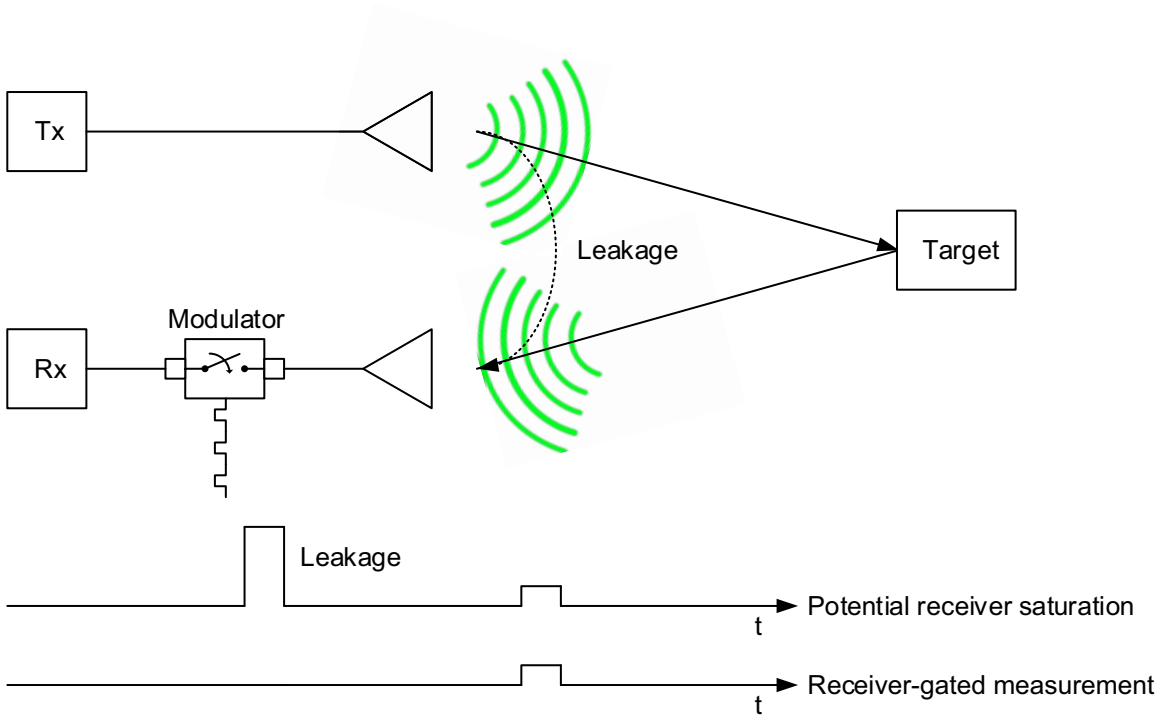


Figure 25-16. Receiver Gating Using the Receiver-side Modulators

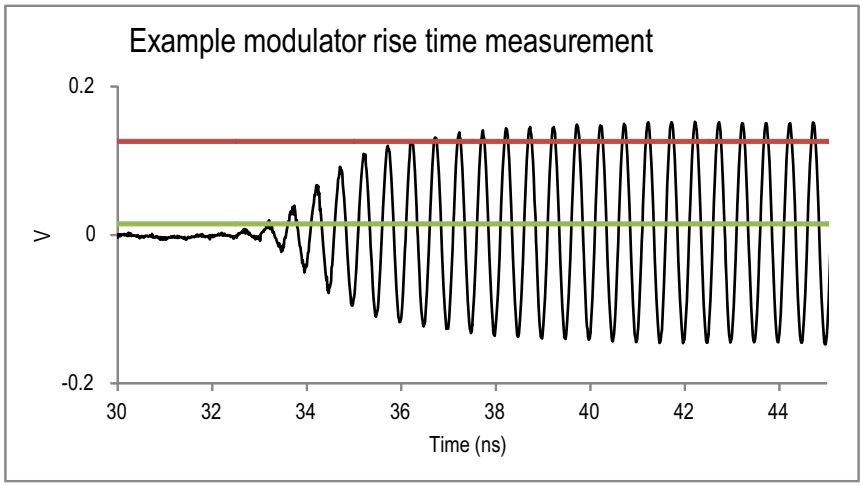
Receiver Gating

With co-located transmit and receive antennas (Tx and Rx), sometimes there is leakage of the transmitted signal into the receiver. If the power levels are high enough, they will compress/saturate the receiver. If the leakage from the transmit pulse is earlier (in time), than the return pulse that is of analysis interest, then receiver gating can be beneficial. In some sense, this is a receiver-gating process and is illustrated in Figure 25-16. An RF modulator can be used to prevent this leakage pulse from being processed by the receiver so one can concentrate on the pulse from the target or some other object.

This type of special receiver-gating is applicable to all measurement methods and is more of a tool to prevent measurement distortion rather than a tool for acquisition in itself. There may be connectorized as well as antenna-based situations where multiple pulse arrivals of different amplitudes are possible.

For any physical pulse modulator, the most interesting timing parameters are rise and fall time and latency. The former is normally set by the modulator drive circuitry and the physics of the modulator itself. For the PIN and FET modulators in the pulse modulator test set, this time is typically < 4 ns (based on 10 % to 90 % transition).

The rise/fall time limits the minimum stimulus pulse width (or receiver-gate width in the case of Figure 25-16). Latency is the delay between the incoming video pulse edge and the RF pulse edge and is largely set by cable and logic delays. For a typical setup with the pulse modulator test set, this latency is on the order of 35 ns. Since the latency is constant, it can be accounted for in the pulse setup by adjusting pulse position timing. An example detected response using the stimulus modulator in the pulse modulator test set is shown in Figure 25-17.



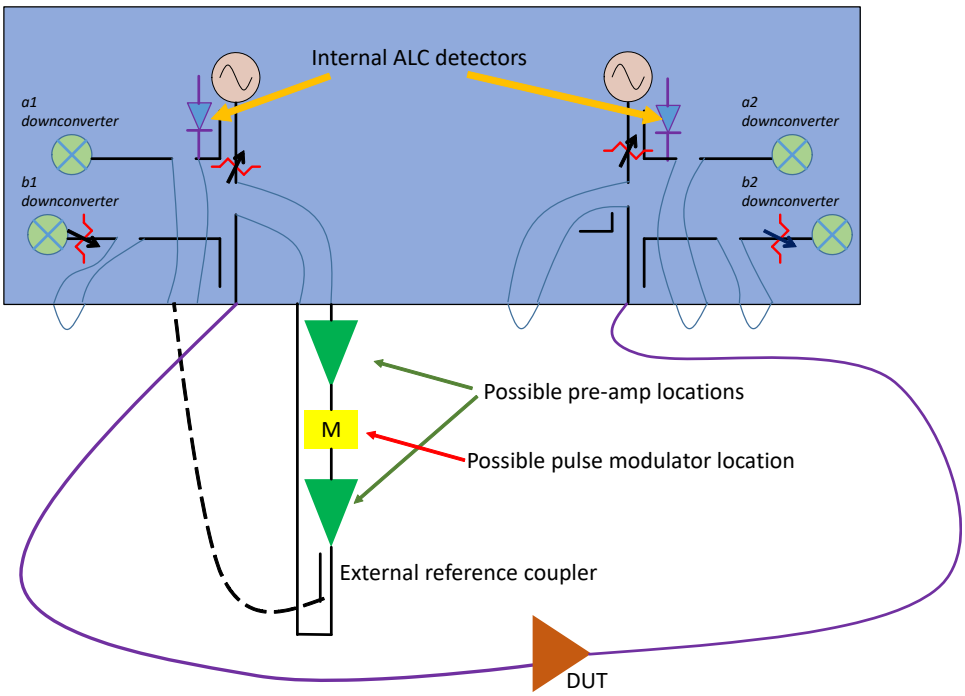
An example pulse response plot of an Anritsu pulse modulator test set (stimulus path). The latency of about 35 ns and rise time of about 3 to 4 ns are evident. The waveform was measured with a 50 GHz bandwidth sampling scope to get adequate time resolution.

Figure 25-17. Example Pulse Response Plot of an Anritsu Pulse Modulator Test Set

Note	When using the test set modulators, the polarities of the pulses applied are important. To allow the test set to remain connected when not being used, the modulators were configured to be in a pass-through state with logic low. Because of this, the pulse input must be inverted so that RF power is applied when the logic is low.

While pulsed power calibrations are discussed in "Calibrations" on page 25-24, a hardware setup consideration involves how the actual power leveling process occurs, particularly when pre-amplifiers are used. By default, the leveling detection node is within the VNA as part of the reference coupling system so it is always before the source loop and before any pulse modulation (and is not included with injection into the reference loop). This is adequate for many situations but if a preamplifier is being used, it may be desirable for the leveling loop to remove any drift of the pre-amplifier. Option 53 (external ALC) helps with this by allowing the use of an external detector. If the pre-amplifier is after pulse modulation, this can create a problem since the detected power will be based on average power which (a) may be a dynamic range problem for the detector if the duty cycle is low or (b) may not be optimal if one wants to level, for example, peak power, to minimize the chance of device damage. For these reasons, two additional software-based ALC detection schemes are available (per-point and per-sweep).

The central concept is that, if a reference coupler is placed correctly (see Figure 25-18), we have information about the signal level at every point of the pulse and can use this as the basis of leveling in a software-loop sense. Note that the pre-amplification can be before or after pulse modulation (and may change depending on power levels, pre-amplifier characteristics and modulator behavior).



A pre-amplified, pulsed measurement setup is shown where an external reference coupler is used to enable an enhanced leveling mode.

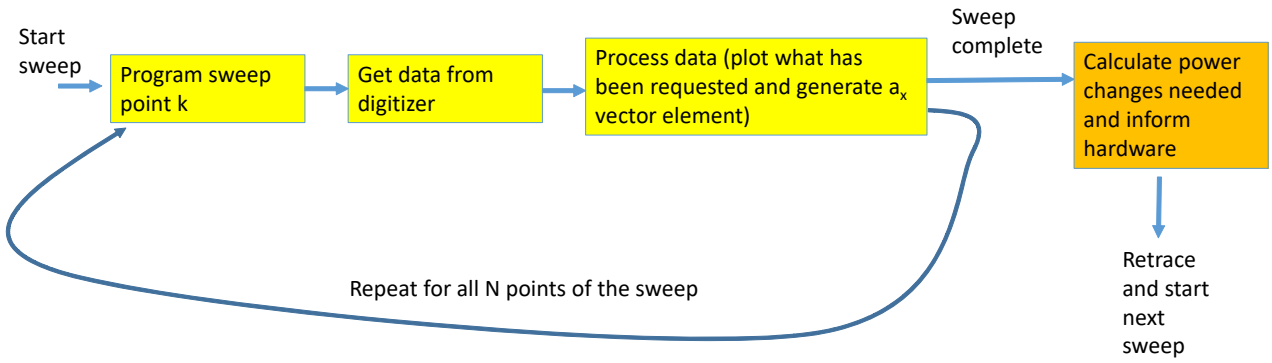
Figure 25-18. Pre-amplified, Pulsed Measurement Setup with External Reference Coupler

Other configurations are possible:

- If no reflection measurements (or full calibrations) are needed, the source loop does not need to be used and the pre-amplifier/modulator/coupler assembly could be connected to the VNA test port.
- The pre-amplifier assembly could be attached to the test port and a dual directional coupler (or two couplers) could be used with the reflected signal going to the b1 input port.
- A reverse configuration is, of course, possible.

As alluded to earlier, it would be possible to place an external ALC detector (Option 53) on this reference coupler and that would at least allow the leveling system to capture drift of the pre-amplifier(s). A potentially better option uses the internal receivers to instead help with the leveling job since the data can be analyzed at the desired time window within the pulse. This becomes a software-leveling loop of sorts but it is done in a 2-tier style: an actual diode detector (whether internal or external) is still used to control the analog loop, but now the data (from either the a1 or a2 receiver path) is used to adjust the loop settings to help maintain a desired power level at a particular point or window in the pulse.

The simplest version of the leveling is termed per-sweep and is available in all pulse modes. At the end of each sweep, the a1 or a2 values (which depends on the driving port) are then compared against the target value and the ALC hardware entries changed to being the value closer to target on the next sweep. The sequence of events is shown in Figure 25-19. It is called per-sweep since, at least in the case of the PIP mode, the power levels are adjusted only at the end of each sweep (in profiling and P2P, each measurement cycle is a sweep). There is no appreciable overhead to measurement time with this per-sweep leveling but it will only correct for power outputs drifting on the scale of sweep time or longer.



The power levels are adjusted at the end of every sweep.

Figure 25-19. Sequence of Events for the Per-sweep Method of Software Pulsed Leveling

The target value mentioned earlier requires some explanation. If no power calibration is applied, the target will be the ALC set value (so a –15 dBm entry will try to get the a1 or a2 values to –15 dB so obviously a receiver calibration helps make these values meaningful). If a power calibration is applied, the system will record the a1 and a2 values right after calibration or leveling activation and use those as targets for all subsequent sweeps. The premise is that the power calibration will be quite accurate just after calibration and we wish to maintain those values. Note that the pulsed power calibrations discussed in Section 25-6 "Calibrations" make the most sense for this setup.

If the drift rate is much faster (or a very long sweep is needed), per-point leveling may be useful and is available for the PIP mode. In this case, a pre-measurement is done at each point in the sweep for a1 or a2 and the power is immediately adjusted in an iterative fashion to get the values closer to target (iteration limit is 20 and the tolerance (leveling convergence accuracy) can be changed by the user (default is 1 dB, tighter than a few tenths may require a lower IF bandwidth)). After convergence (or iteration limit exceeded), the full set of data is used for the pulse measurement and we move on to the next sweep point. The sequence of events is shown in Figure 25-20 and the leveling selection/tolerance entry field are shown in Figure 25-21.

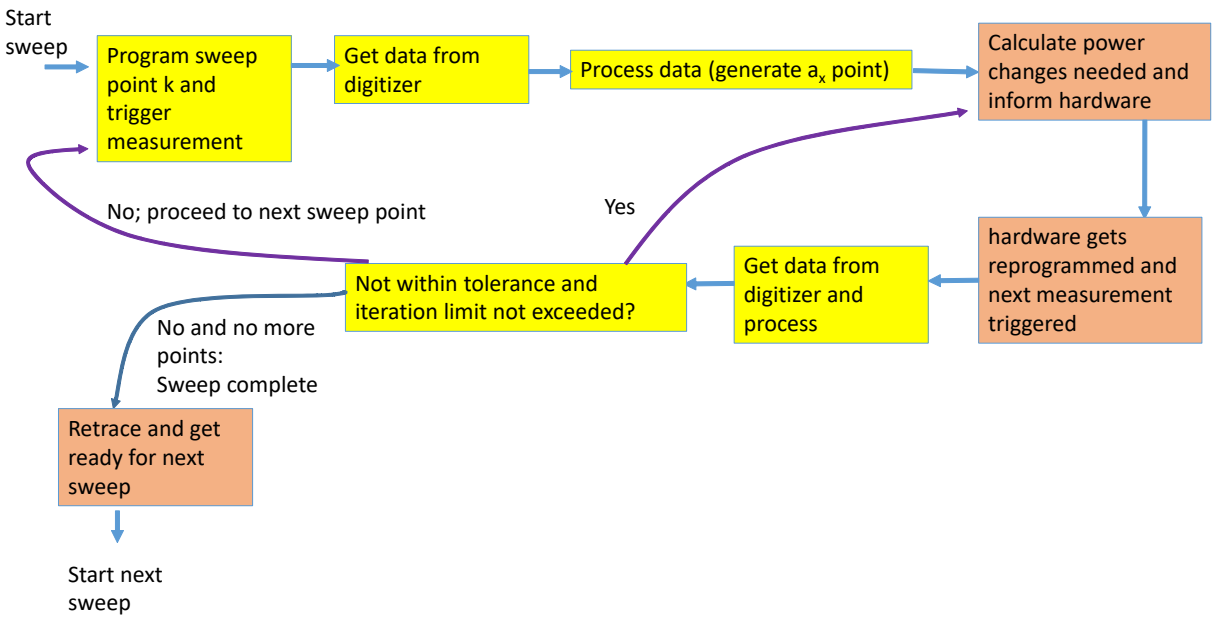
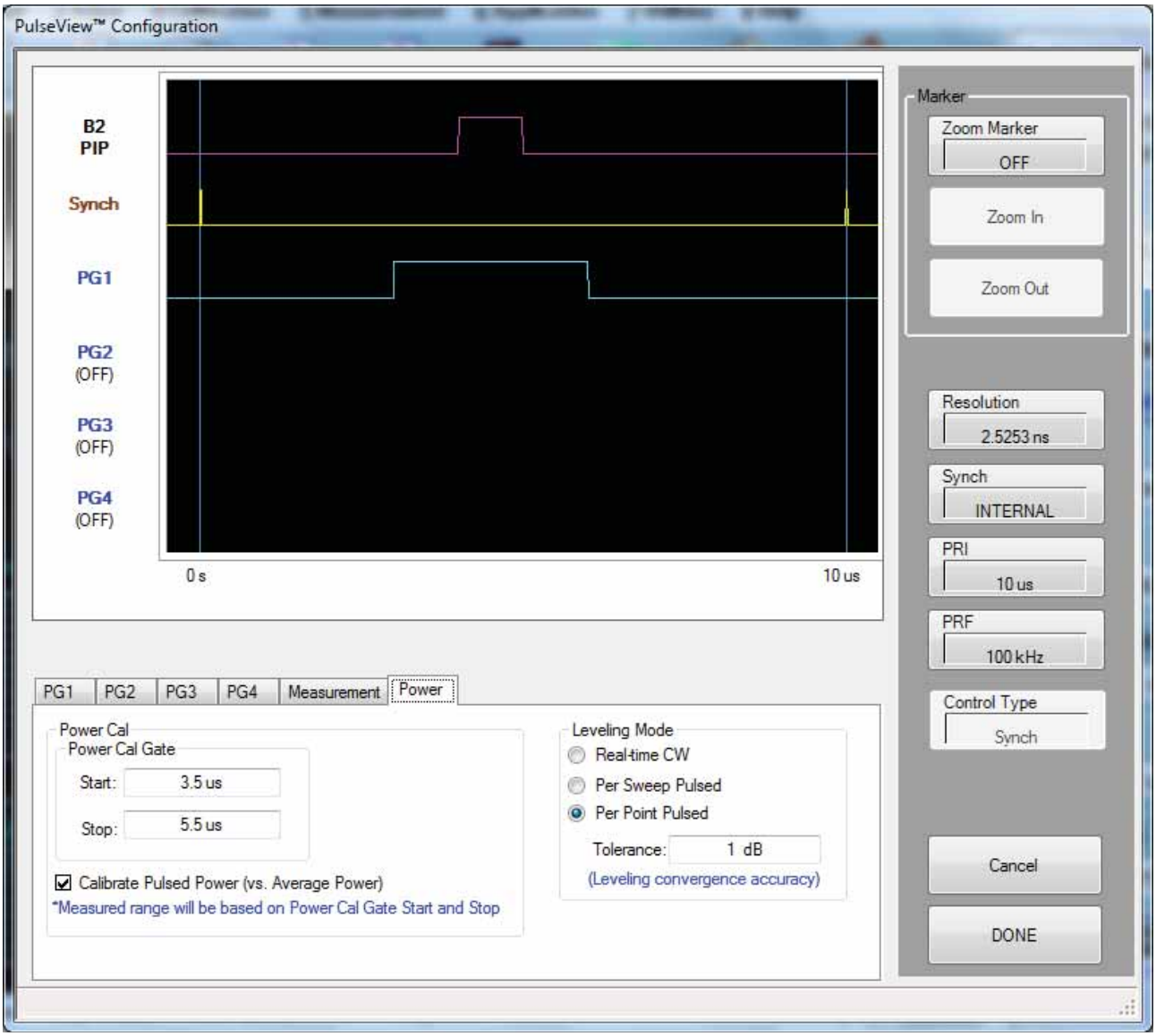


Figure 25-20. Sequence of Events for Per-point Software Leveling

The system tries to force the power level to converge to the target at each sweep point before keeping the data and moving on to the next point.

Figure 25-20. Sequence of Events for Per-point Software Leveling



Set up for per-point leveling and the power calibration set to analyze the middle of the pulse (which runs from 3–6 microseconds from synch in this example). Note that the convergence tolerance (1 dB from target here) is only shown when per-point leveling has been selected.

Figure 25-21. PULSEVIEW CONFIGURATION Dialog Box

The target values are the same as discussed for per-sweep leveling. Because multiple measurements are done per point, the overall measurement time will increase but this may be the preferred mode if the power accuracy during the measurement is critical and the pre-amplifiers are prone to significant and/or rapid drift.

Note	When recalling a setup in per-point or per-sweep leveling with a user power calibration, this system needs to re-acquire the target values and will attempt to do so on the first sweep after recall. If an appropriate setup was not in place at that time (e.g., the signal to couple off to the reference receiver on the VNA was not connected), the user can refresh the target (force a new target acquisition) using the 'Refresh Leveling Ref.' button on the POWER >> OTHER SETUP menu after the setup is correct. One may want to do a new power calibration if the setup has changed, and that will automatically store new target values. If a power calibration is recalled on its own, a new target value acquisition is also needed.

25-5 Fast Digitizing Fundamentals

To explore the high-speed digitizer method in more depth, consider the acquisition system of the Anritsu MS4640B with Options 35 and 42 shown in Figure 25-22.

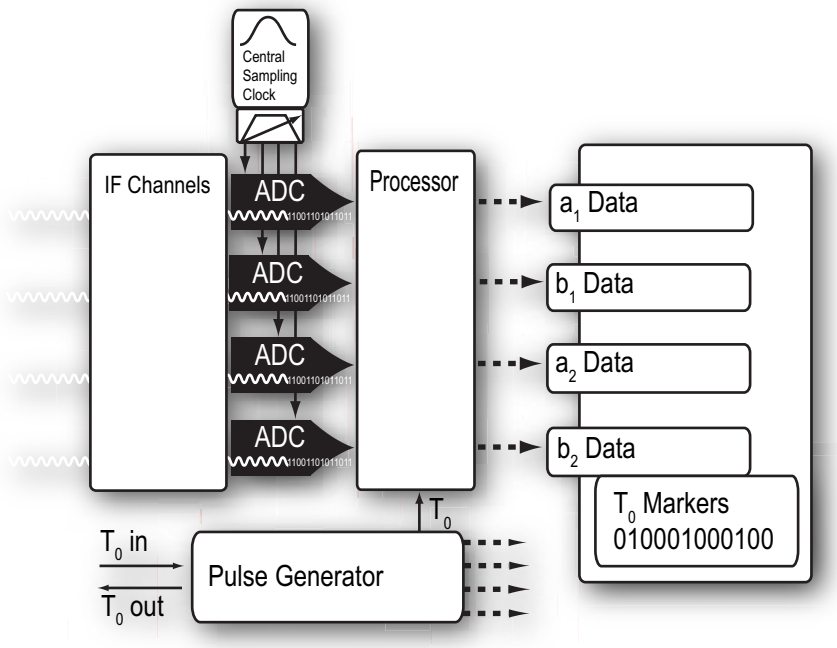


Figure 25-22. Block Diagram: Acquisition System in the MS464xB Series with Option 35 and Option 42

Details on the functionality and performance of the IF digitizer are covered in this section.

Option 35—IF Digitizer

IF signals are generated by the down converters in the MS464xB Series. When equipped with Options 35 and 042, the standard IF system is bypassed, and signals are routed to a special high-speed digitizing IF board. This board consists of analog processing (filtering, gain, calibration and other factors) in a much wider bandwidth than the standard IF system to enable the measurement of much narrower pulses. This board also houses the fast analog-to-digital converters, the pulse generators, and digital processing components.

A large block of configured memory is used to store the data from the converters. This data is keyed with the T_0 synch information. For the pulse measurement processing, the data of interest is selected (relative to T_0) and run through the appropriate conversions to the frequency domain. S-parameters are then created and the appropriate calibrations are applied.

Since the processing is relative to these time markers and not any of the pulses per se, there is considerable flexibility on where in time one can look at the results. This flexibility becomes particularly valuable in complex pulse situations where many sub-pulses are present or when multiple external generators are being used.

Narrow pulses and fast edges can be captured with a minimum resolution of about 2.5 ns. Since the memory is very deep, very low duty cycles and low repetition rates are supported without sacrificing resolution. For pulse repetition rates in the Hz range and slower repetition rates, normal triggered methods are sufficient.

Using Acquired Data

From the point-of-view of the measurement, normally only the delay (or delay and the number of measurement points for pulse profiling) and measurement acquisition width must be specified. To further understand the acquisition process, consider the point-in-pulse measurement shown in Figure 25-23. In these cases, the data of interest (denoted with x's) begins at some number of samples (delay) after the T_0 edge.

The time domain data is converted into a single frequency domain point. If a smaller width had been entered, fewer samples would have been taken. In this way, the measurement width acts as an effective processing bandwidth (for example, an IFBW).

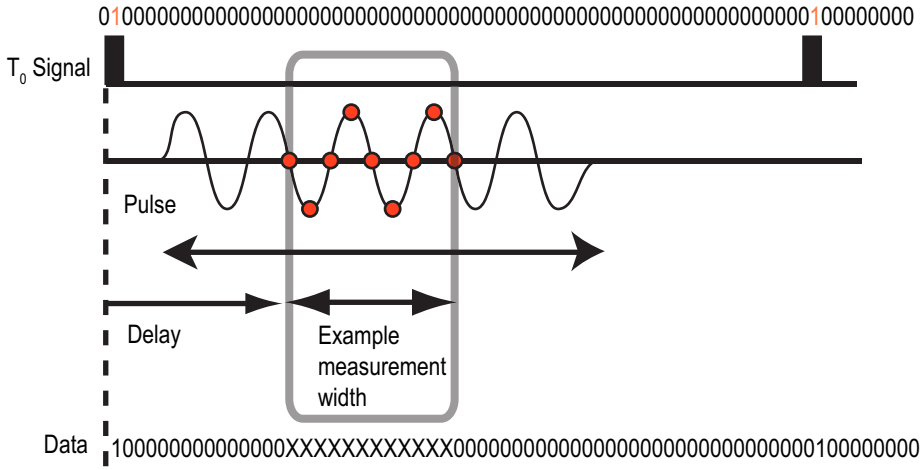


Figure 25-23. Using Acquired Data from a Pulse-to-pulse Measurement

This result can be averaged over multiple pulses to reduce trace noise or increase effective dynamic range. In this case, the x's will be collected after multiple T_0 high markers until the desired level of averaging is complete.

During the acquisition, pulse timing is coherent with the ADC clock to maintain phase. It is presumed, in the election to perform a point-in-pulse measurement, that the DUT response is consistent over multiple pulses (or that the goal is to average that behavior). The total acquisition time is computed based on the requested averaging and the PRI. If one wants to look at more individual pulse behavior, then the pulse-to-pulse measurement mode is recommended.

Pulse-to-pulse and pulse profiling measurements are processed in similar ways. For pulse profiling, a single acquisition is completed (based on time span of interest, PRI, and requested averaging). That data is sequentially processed using the selected measurement width at each of the requested time points. If averaging is enabled, the process is repeated over multiple pulses. In the pulse-to-pulse measurement mode, the data is acquired over the measurement width requested, and then the same measurement window is processed separately per period.

To understand the number of pulses that must be acquired, consider an effective IFBW (=IFBW/(pt-pt averages)….sweep-by-sweep averages are handled as a post-processing task). If this value is larger than 1/(measurement window) then only one pulse is needed although two may be acquired to guarantee synchronization. As the request IFBW shrinks, addition pulses will be captured for point-in-pulse and pulse-profiling measurements.

For example, given the following parameters:

- Measurement window = 1 μs
- IFBW = 100 kHz
- Point-to-point averages = 5

Effective IFBW is 20 kHz, so acquisition over 50 pulses will be required (with potentially an additional pulse to guarantee synchronization).

The measurement resolution in pulse mode is set by the ADC clock rate and generally defaults to ~2.5 ns (except for receiver frequencies < 110 MHz). This in turn puts a lower limit on measurement width. This also sets the maximum acquisition length which in turn limits the maximum PRI. Without Option 36, the maximum acquisition length at the default resolution is 0.5 seconds and the maximum PRI is 0.25 seconds (these values scale proportionately for wider resolutions, that is, slower ADC sampling rates). With Option 36 installed, the maximum acquisition length at default resolution is 2.5 seconds and the maximum PRI is 1 second (these also scale with resolution).

The resolution time can be increased on the PulseView menu up to 90 ns when fine time resolution is not needed but longer acquisition length is desired. The available acquisition length scales proportionately with the resolution entry as do the other pulse parameters.

25-6 Calibrations

At their root, the measurements discussed in this chapter are still S-parameter (or perhaps unratioed parameter) measurements. Even though they are conducted in a pulsed environment, for the most part, nothing has changed in terms of availability or execution of calibrations. There are, however, some subtleties to keep in mind.

Power Calibrations

There are a number of different ways that power can be calibrated:

- **Unpulsed using a normal CW power sensor.** This is the simplest method, and ensures accuracy of the average or midpoint on-state power. This is preferably done with the pulse modulator in place, but biased on so that the insertion loss of the modulator is included. The Anritsu pulsed test sets allow this to be done without external pulsing by using the invert switch on the rear panel.
- **Pulsed using a normal CW power sensor.** This will calibrate the power to the average pulsed power. For most power meters, a duty cycle can be entered that will correct this value to an average on-state power (and the VNA will not preset the power meter, so this duty cycle can be entered manually on the power meter; the VNA will not do this)
- **Pulsed using a pulse power sensor** (the Anritsu MA2411B sensor (used with the ML249X meter) and the Rhode & Schwarz NRP-Z85 sensor (used with the NRP2 meter) are supported at this time). This is the most flexible approach, since power can be calibrated to a particular portion of the pulse, so it allows, for example, calibrating to peak power with significant resolution control.
    - In the PulseView configuration dialog, a tab is available where the gate start and stop are defined. It is only the power in this time range (relative to the PRI start) that will be used for calibration. The dialog to support pulse power sensor-based user power calibrations is shown in Figure 25-24.

The time interval over which power data should be used (relative to the period start) is entered and a check box (replicated on the power calibration menu) enables its use.

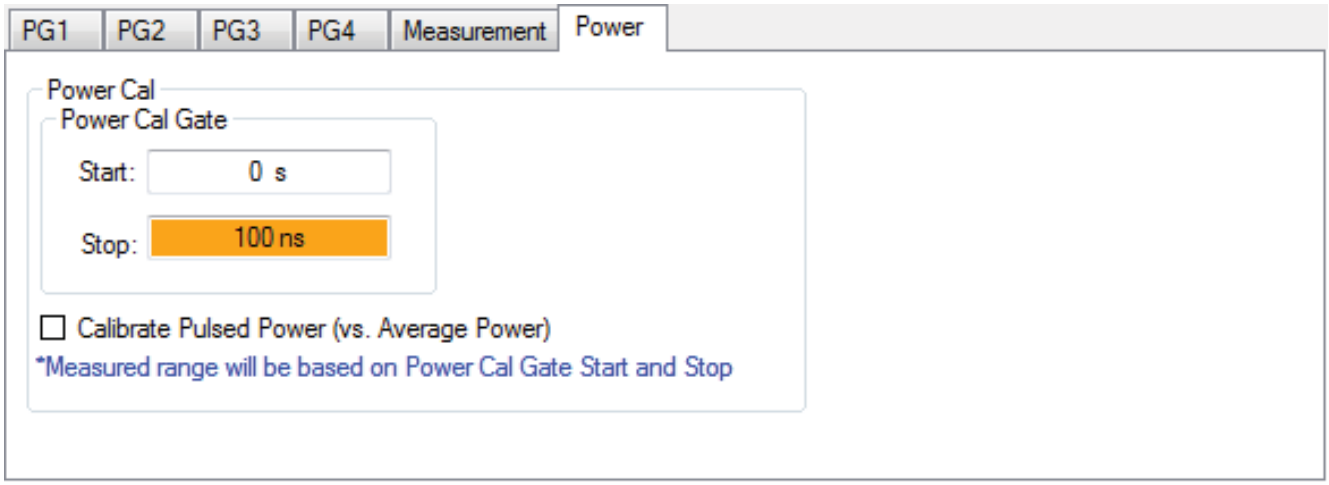
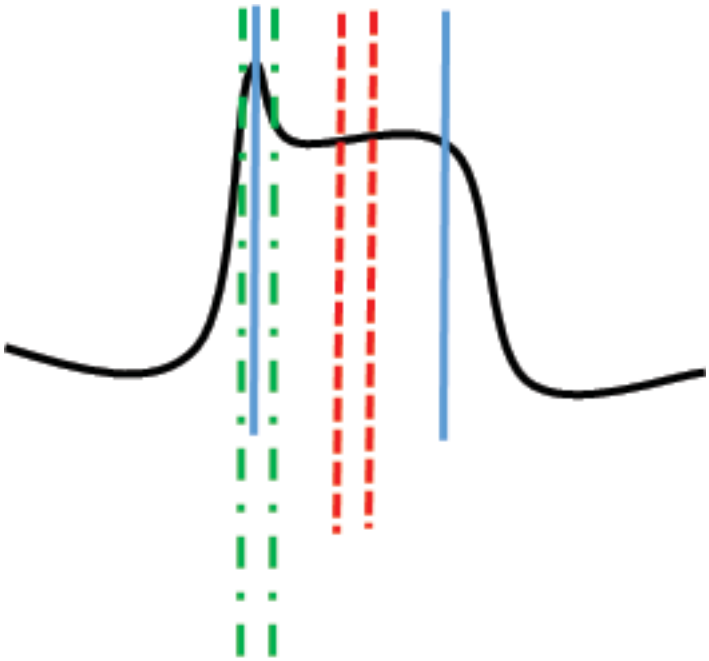


Figure 25-24. Pulse Power Sensor-based User Power Calibrations Dialog Box

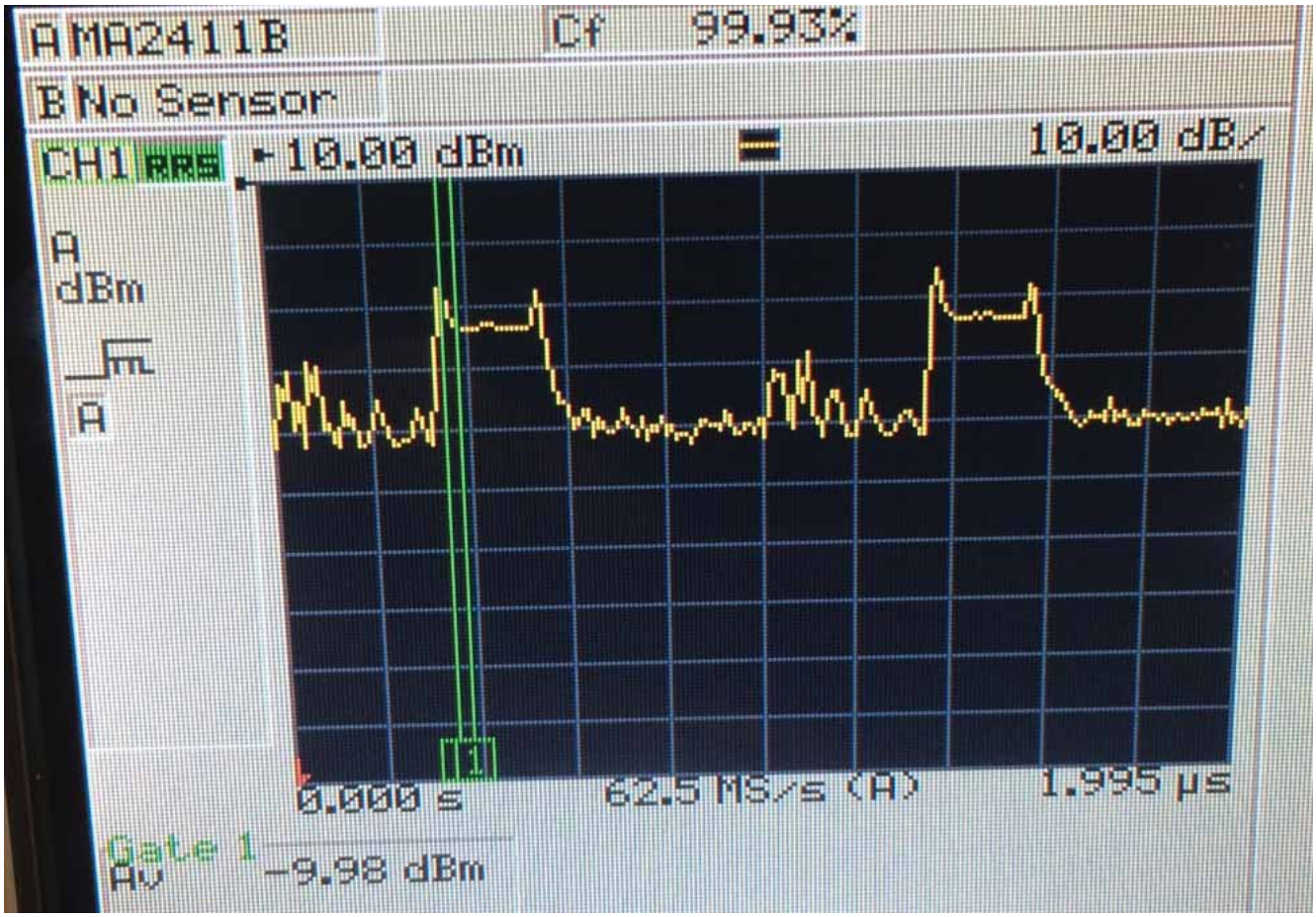
- If the Calibrate Pulsed Power (vs. Average Power) box in this dialog (also available from the Power Calibration dialog) is checked, the pulsed power meter will be directed to enter profiling mode and to measure just in the gated region. An error will occur at the power calibration stage if a pulsed power meter is not available. If unchecked, average power (over the whole cycle) will be calibrated.
- Note that the pulsed power meter must be synchronized to the beginning of the PRI and this is usually done by connecting Synch Out from the VNA to the external trigger input of the power meter. If an external pulse generator is part of the setup, it can also sometimes be used as the synch source for the power meter as long as the time alignment makes sense.
- The gate start and stop range cannot exceed `2*PRI`.

- The gate placement can be adjusted depending on the desired leveling target; some examples are shown in Figure 25-25. The entire on-state of the pulse could be averaged to create the power measurement (solid gate markers below), the midpoint of the pulse could be used (dashed gate markers below), the initial overshoot could be used (dot-dash gate markers below), etc. The gate selection relative to the pulse shape will be visible on the pulsed power meter screen to help with confirmation of the choice. An example is shown in Figure 25-26.



The solid gates would ensure that the average on-state pulsed power would be close to the target value. The dashed gate would force the mid-pulse-power to be close to target. The dot-dash gate would instead make the peak intra-pulse power be close to the target value.

Figure 25-25. Some Possible Gate Orientations That Might be Used for Pulsed Power Calibrations



The gate (marked by green lines) was selected to be near the rising edge and the power calibration was targeted at –10 dBm in that zone.

Figure 25-26. Screen Shot of ML249X Power Meter After a Pulsed Power Calibration on the VNA

- Resolution limits must be observed (2.5 ns increment) even if the power meter may use many cycles to improve resolution. Note that available gating resolution will be reduced at lower carrier frequencies just as happens with measurement resolution.

Receiver Calibrations

These are done in pulse mode and can be done with stimulus pulsing disabled (typical) or enabled. If stimulus pulsing is used, it is important to ensure the measurement window placement does not occur during an off-period in that given path.

User RF Calibrations

(For example, S-parameter calibrations such as SOLT, LRL, etc.)

User RF calibrations are done while in pulsed mode. If stimulus pulsing will be used in the measurement, it should be used in calibration as well. This will allow for correction of any minor pulse distortions that may occur in the stimulus system. If using an Anritsu pulse modulator test set, extra reference couplers are provided so that a pulsed reference can be provided to the system. Applying this reference can help reduce trace noise and minimize any video contamination that may occur in a measurement. If the stimulus is not being pulsed (for example, only DUT or bias pulsing is being used), utilization of these additional reference signals is not needed.

Note	If the reference couplers in an Anritsu pulsed test set are used, the ALC detection system will still detect the un-pulsed signal, but the extra loop port should be terminated to avoid minor power inaccuracies. The ALC detection circuitry is in the reference path of the VNA, but prior to the reference access loop where the pulsed version is injected. Reference couplers are discussed in "Minimizing Uncertainties".

All standard S-parameter calibration methods apply, the selection of which continues to be governed by the media and standards available—as discussed in other chapters of this guide.

Although not nearly as influenced by changes in pulse parameters as the narrowband or triggered measurement methods, the high-speed digitizer pulse measurement method still has some calibration dependence on the pulse parameters. If those parameters change, a new calibration may be needed, depending on the uncertainty tolerance.

To understand the effect of changing pulse measurement parameters, consider the IF waveforms shown in Figure 25-27. They illustrate a simple 2-port transmission calibration step, where a thru line is connected and stimulus pulsing is being used.

For example, take overshoot on the stimulus pulse. If the measurement window is toward the leading edge of the pulse (yellow samples), some of the overshoot may be included in the measurement. In some cases (a long thru line, for example), the overshoot may not be entirely symmetric (occurring at the same time) between measurement channels (a_1 and b_2, for example).

If the measurement window is moved later (green samples), then there is no overshoot present and the b_2/a_1 ratio has changed. Depending on the pulse and window positions, this could cause a fractional dB error in measurement unless recalibrated. Since there is partial ratioing, the maximum error is always less than the maximum overshoot, but could be in the 10 % to 30 % of peak overshoot in a worst-case scenario.

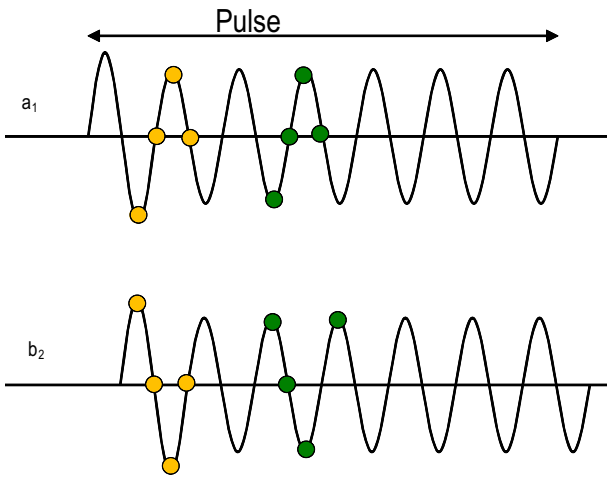


Figure 25-27. Example of Pulse Parameter Change Affecting Calibration

The decision whether to recalibrate can be affected by a number of parameters. These guidelines may help to determine whether to recalibrate the instrument:

- No stimulus pulsing is used (for example, using only bias or DUT control pulsing): With this configuration, calibrations are largely invariant to pulse parameters, and recalibration is normally not required. An exception may occur when performing an externally-triggered pulse measurement and the measurement window is close to the trigger event. In that case, experimentation with recalibration may be required to determine if there are any material effects from the measurement setup.
- Stimulus pulsing is used: As in the example of Figure 25-27, it may be desirable to recalibrate if it is believed that the measurements occur at a point for which there is some overshoot from the stimulus. If the test is made using an Anritsu pulse modulator test set, this overshoot is less than ~0.5 dB and only covers about the first 5 ns. If not using that time zone, or if that zone is used but ~0.1 dB impact is not significant, then one need not recalibrate. If using another stimulus modulator, one should consider its response pattern.
- Receive-side modulation is used (receiver-gating as shown in Figure 25-16): If one needs to operate near the edge of the modulator pulse, and if one is sensitive to small errors, then recalibration is recommended.

If pulse parameters are changed after calibration, the CORR annunciator on screen (which indicates that a calibration is applied) may change to CORR(?) to indicate the possibility that the calibration has been disturbed. See Figure 25-28. The measurements may still be entirely valid based on the exact pulse measurement configuration (see the recalibration guidelines above).

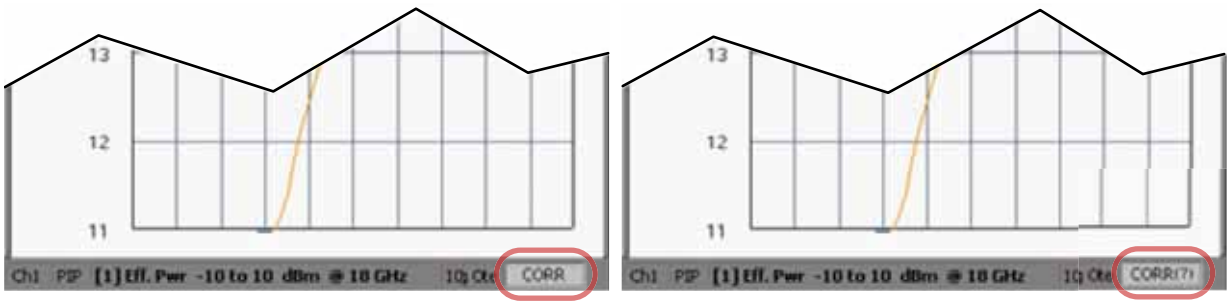


Figure 25-28. On Screen Calibration Disturbance Indication—CORR versus CORR(?)

25-7 Point-in-Pulse Measurements

This section describes point-in-pulse measurement examples in more detail and provides guidance on how to select parameters of importance. Point-in-pulse measurements are often used when some portion of the pulse is not of interest and only the averaged result over a specified measurement window is of concern.

Pulsing is often done for thermal reasons or when some measure of basic RF performance in some non-dynamic portion of the pulse is the main measurement requirement, or for both of these reasons. Typically, this measurement is done over swept power or frequency.

Multiple point-in-pulse measurements may be run on different VNA channels where the frequency (or power) sweep parameters and pulse parameters are the same, but the point in the pulse where data is taken is different. In this way, information regarding the sweep parameter and the intra-pulse behavior can be acquired nearly simultaneously.

A pair of power sweeps (perhaps set up in two VNA channels) where the effect of pulse width will be examined provides our example. This example neglects transient effects near the beginning and end of the pulse, but will reveal whether there are any effects related to an increase in pulse width (and, in this case, duty cycle).

In this example, only RF is being pulsed, and the PRI is 0.1 ms. Pulse widths of 20 µs and 80 µs are used, and the point-in-pulse measurement is set to roughly the middle of each pulse with a width of 10 µs. The amount of time between initiation of the stimulus pulse and the measurement is different in the two cases, so the effect may be a convolution of actual measurement time and the duty cycle presented to the DUT. In this simple example, only one pulse generator is used. A delay relative to T_0 is used for clarity.

Sweep 1:

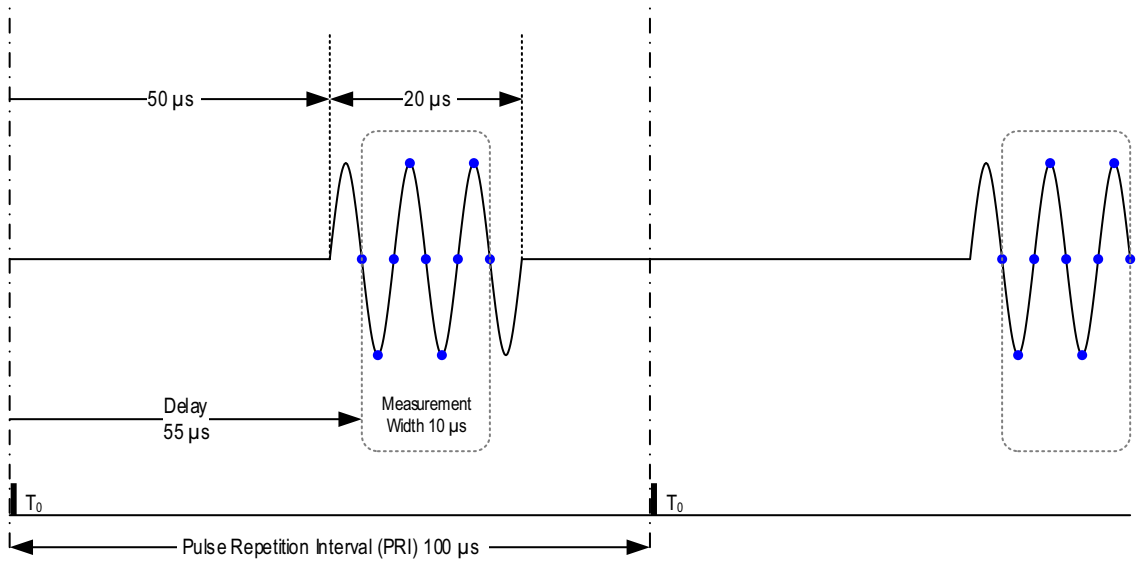


Figure 25-29. Details of Sweep 1 Timing Relationships in the Point-in-Pulse Setup Example

- PRI: 100 µs
- Pulse generator 4: delay 50 µs, width 20 µs
- Point-in-pulse position: 55 µs
- Point-in-pulse measurement width: 10 µs
- Power sweep: –10 to 10 dBm, 41 points, frequency CW 18 GHz

Sweep 2:

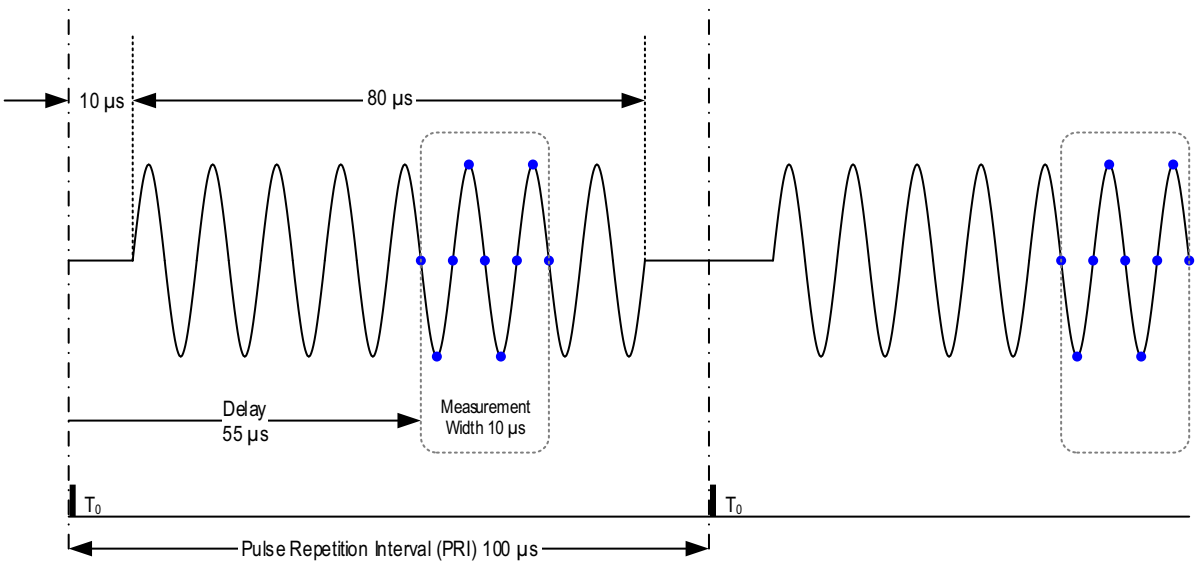


Figure 25-30. Details of Sweep 2 Timing Relationships in the Point-in-Pulse Setup Example

- PRI: 100 µs
- Pulse generator 4: delay 10 µs, width 80 µs
- Point-in-pulse position: 55 µs

- Point-in-pulse measurement width: 10 µs
- Power sweep: –10 to 10 dBm, 41 points, frequency CW 18 GHz

Approximate timing relationships for this measurement are shown in Figure 25-29 and Figure 25-30. DUT (amplifier) output power is the measurement of interest, so the pulse measurement will be processed only on b_2.

The dialog boxes and settings for these examples are shown in Figure 25-31, Figure 25-32, and Figure 25-33. Only PG4 is enabled, and it is set for inverted polarity since a pulse modulator test set is being used.

Note	If a noise floor signature is observed in the measurement, it is a possible indication that the pulse generator feeding the pulse modulator test set was not inverted.

The receiver channels remained coupled in this example, although only b_2 is being used for these measurements.

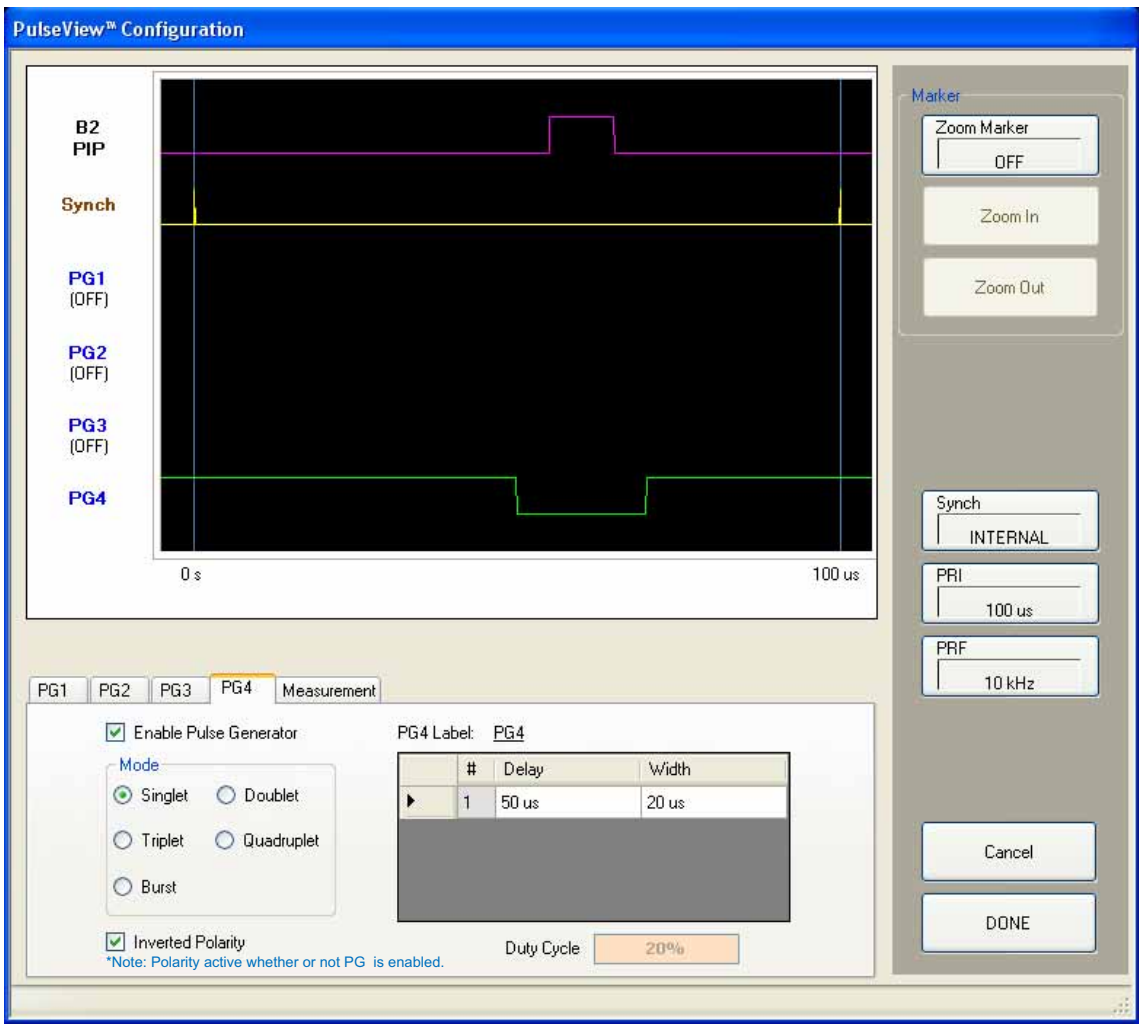


Figure 25-31. Sweep 1 Settings in Pulse Configuration Dialog Box

Note	The following receiver notations used in text and in the instrument user interface have identical meanings: A1 = a_1, B1 = b_1, A2 = a_2, B2 = b_2.

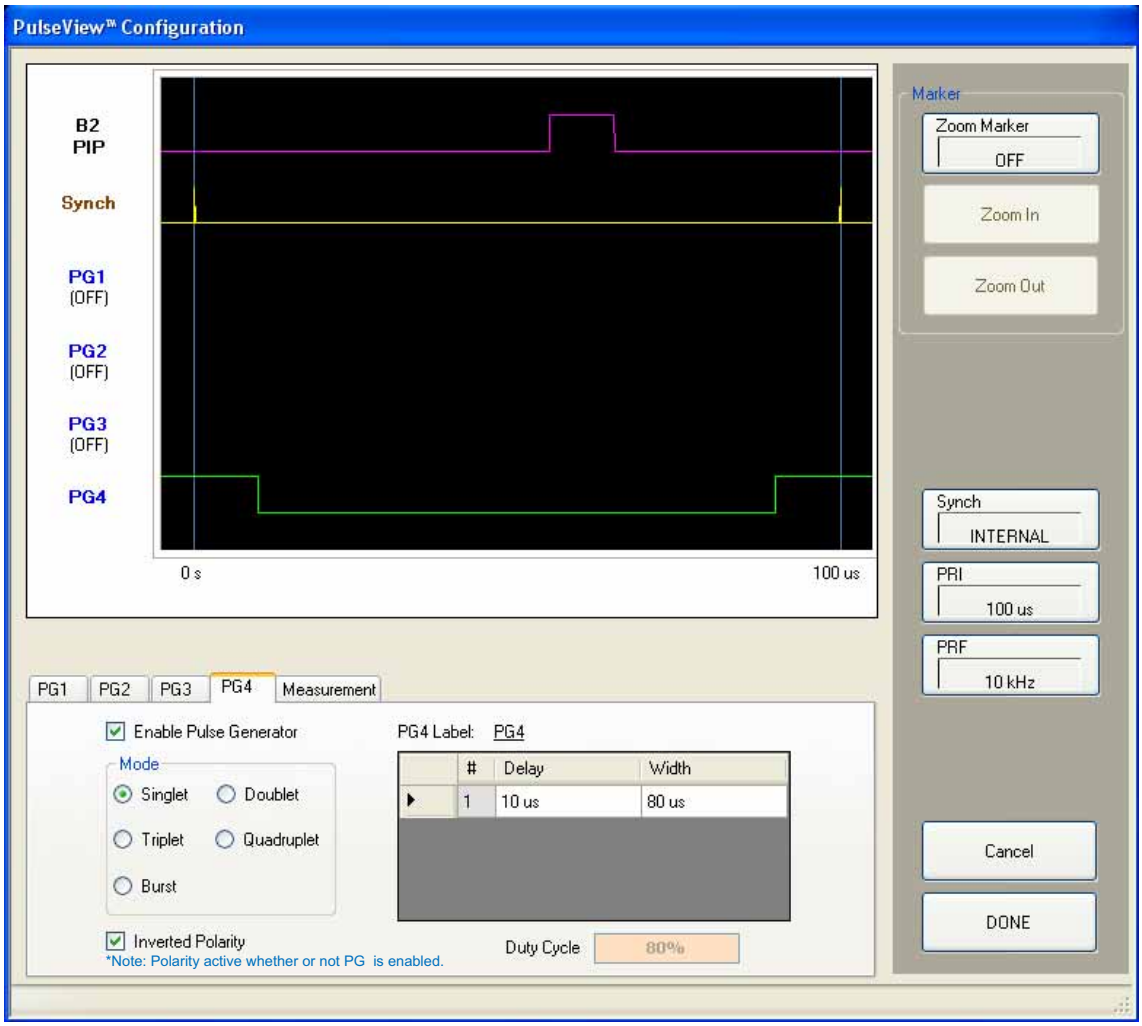


Figure 25-32. Sweep 2 Settings in the Pulse Configuration Dialog Box

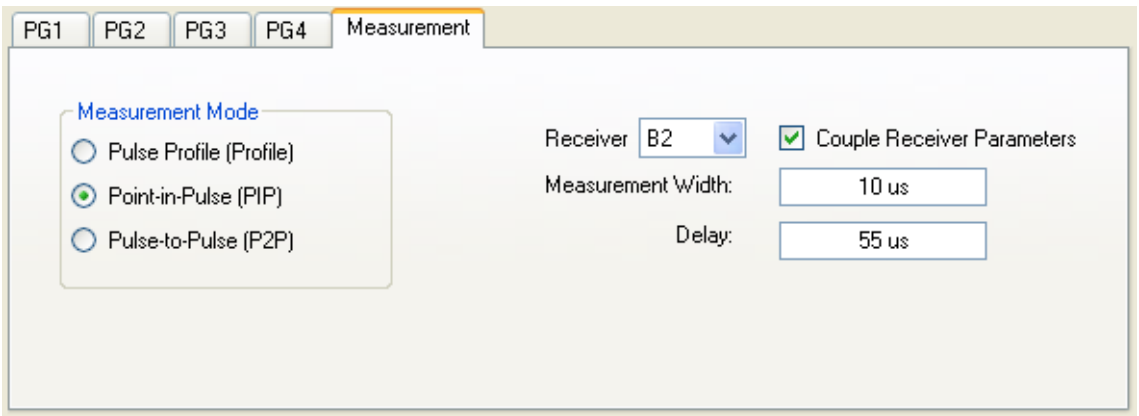
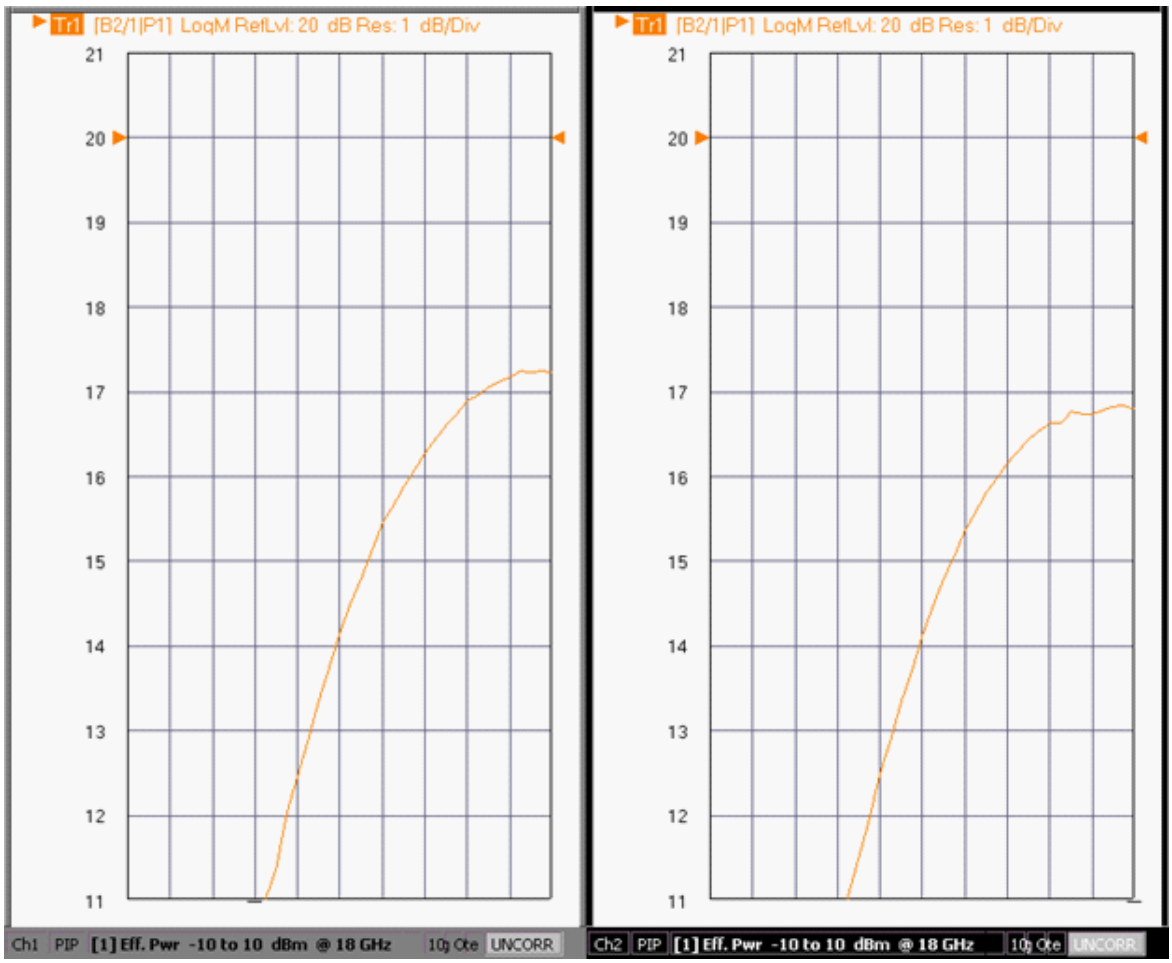


Figure 25-33. Pulse Measurement Tab Details

Point-in-Pulse Example Measurement Results

The measurement results for the example described above are illustrated in Figure 25-34.

There is a noticeable decrease in saturated power from this device as the duty cycle increases and the measurement pulse is repositioned (relative to the leading edge of the stimulus), as might be expected if there were significant thermal effects present. A receiver calibration was used for this measurement (performed with a simple thru line connection as discussed in other chapters in this guide).



Only power and receiver calibrations are applied.

Figure 25-34. Example Results of Two Point-in-Pulse Measurements

25-8 Pulse Profiling Measurements

Pulse profiling measurements are often of interest when changes in performance within the pulse duration may occur. Examples include thermal and trapping time constants or leading edge phase overshoot/undershoot and trailing droop behavior of a pulse phased array module.

For this example measurement, the PRI is 2 µs and the combined bias/RF pulse width is 500 ns. The goal is to examine the pulsed S_{21} response using a sliding 25 ns profiling window beginning just before the main pulse and ending just after it.

We are interested in S_{21}, but wish to look at the off ratio as well, so only b_2 will be profiled. The reference a_1 data will be acquired from a 25 ns window in the middle of the pulse. In order to properly calculate the ratio b_2/a_1, the profiling point counts have to match although the a_1 'profiler' is not moving (this requires that the measurement channels be uncoupled).

A point count of 91 is chosen for this example, making each of the profiling windows non-overlapping (this is not required). Sweep-by-sweep averaging over 10 pulses is used to improve trace noise. With an IFBW of 1 MHz and a profiling width of 10 ns, about 100 pulses-per-sweep are used to compute the result.

Setup

- Pulse Repeat Interval (PRI): 2 µs
- Pulse generator 1 (bias): delay 100 ns, width 500 ns
- Pulse generator 4 (RF): delay 100 ns, width 500 ns
- b_2 profiling range: 0 to 900 ns, 91 points
- b_2 profiling width: 10 ns
- a_1 profiling range: 250 ns to 250 ns, 91 points (no motion, but point count required)
- a_1 profiling width: 25 ns
- Power and Frequency are constant at 0 dBm and 20 GHz
- Sweep-by-sweep averaging: 10
- IFBW: 1 MHz

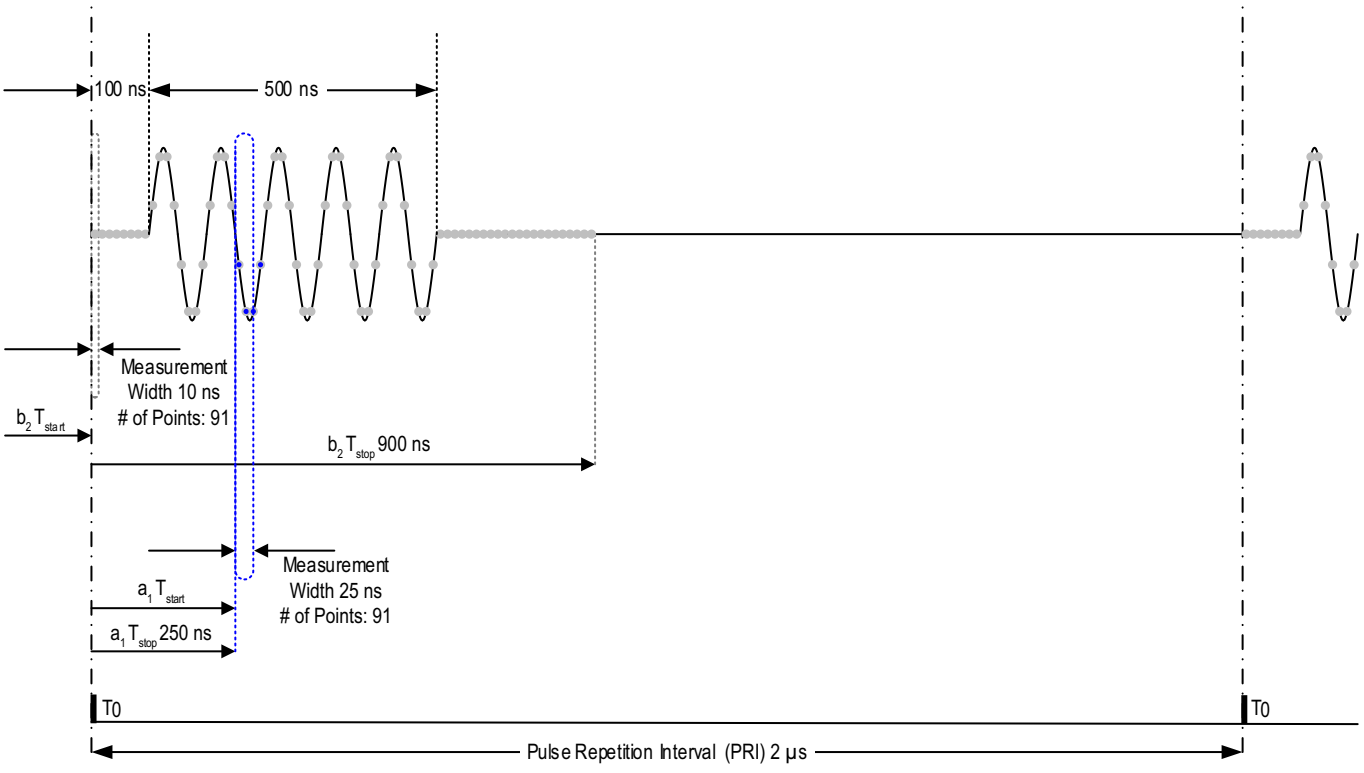


Figure 25-35. Timing Relationships for the Pulse Profiling Example

The Pulse Configuration setup dialog in Figure 25-36 shows the parameters used for the measurement. Since PG4 is driving a pulse modulator test set, note that it is selected to be inverted as discussed before. In profiling, it is usually easy to identify an error in this setting.

Internal synchronization is being used so that the VNA's internal pulse generators are providing the synchronization pulse for any external generators that might be needed in a larger setup. The PulseView™ Configuration dialog box (Figure 25-36) shows two different receiver configurations.

Determining the profiling time range for this measurement is an important consideration. A profiling time in the range of 50 ns to 650 ns might seem adequate, but in setups for this range, important information about profiling could be missed at the end of the measurement.

Note	Consider internal latency of the stimulus modulation (about 35 ns) as well as the delay through the DUT and any cables (about 50 ns in this case).

The action in this example measurement does not start until about 175 ns and does not finish until about 675 ns. Using a larger time span can ensure all of the information is captured and only adds a very small amount to the measurement time.

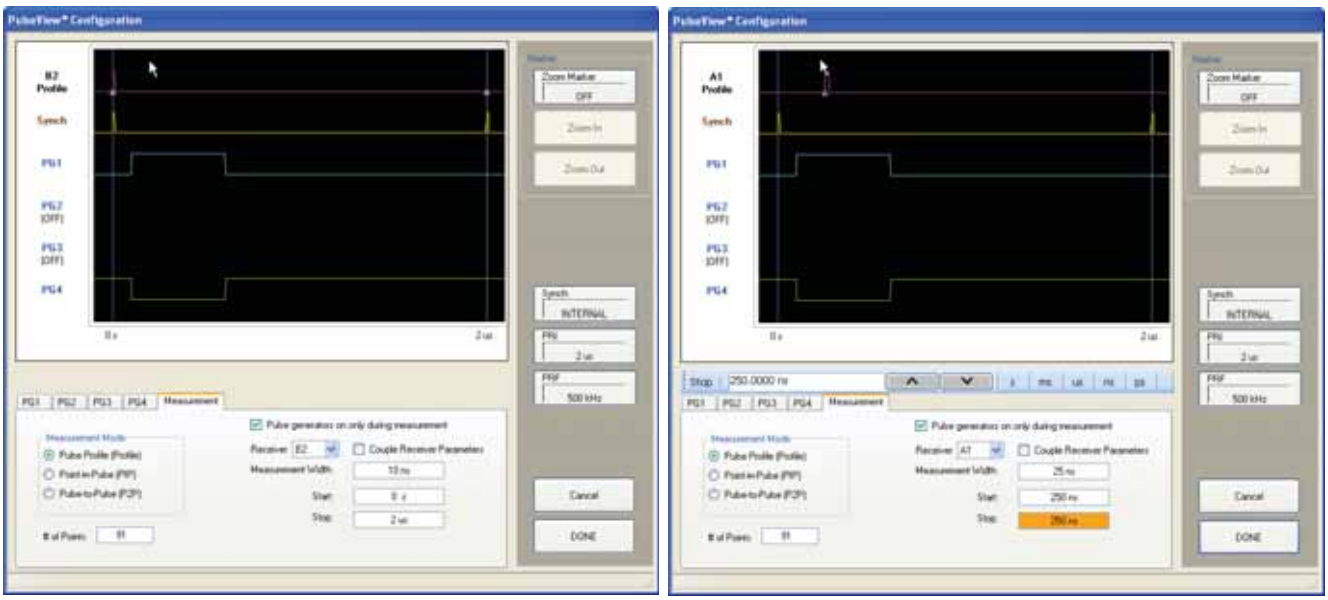


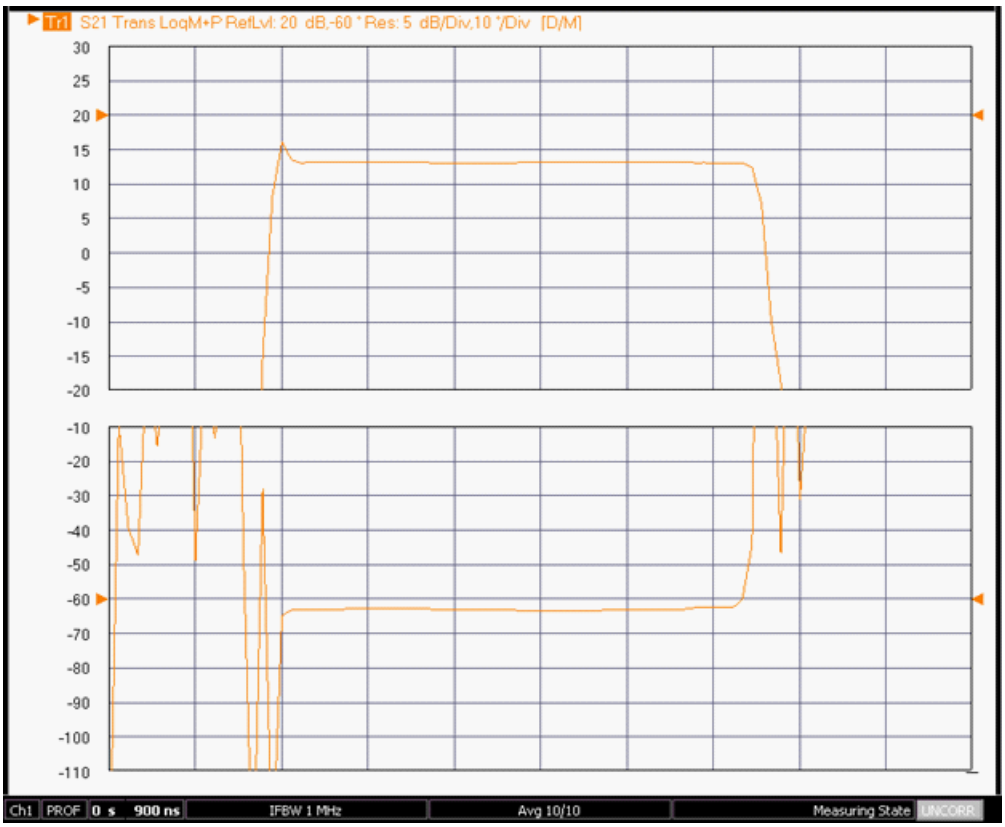
Figure 25-36. Pulse Profiling Measurement Configuration Showing Different Receiver Configurations

The measurement trace (top line in the dialog box of Figure 25-36) is labeled (on the left side) with the active path to which it corresponds (b_2 in this example) and the type of measurement (Profile in this example). It also indicates the span of the profile that has been requested (designated with triangle delimiters) and the profiling measurement width, and it indicates the measurement timing for the active path (b_2 in this case).

By selecting the Pulse generators on only during measurement checkbox, the pulse generators will only start up when making a measurement, thus avoiding early device stimulation (the user may be looking at long thermal time constant effects) and some potential overshoot conditions (from level dip release). This feature is available only in pulse profile measurement mode.

Pulse Profiling Example Measurement Results

Measurement results for the example described in Figure 25-36 are shown in Figure 25-37. A small overshoot on magnitude of ~ 2 dB and ~1 degree of phase distortion on the rising edge can be observed. The phase distortion begins very early on the falling edge (relative to the magnitude). A simple data-to-memory normalization was all that was used for this measurement example.



A fast phase overshoot is noted along with a slow gain decay. Only trace normalization (data-divided-by-memory) was used for this measurement.

Figure 25-37. Pulse Profiling Example Results

When examining the pulse edge timing, keep these points in mind:

- All of the delays are referenced to the leading edge of the synch pulse T_0,and are defined to the leading edge of the pulse being measured.
- Internally, the mapping of the T_0 edge onto the data stream will be correct to within one sample (2.5 ns). When an external T_0 is being used, the mapping will normally be correct to within two samples.
- While the pulse edges will be very accurate with regards to the pulse generator output, latency from what these edges drive (the Anritsu pulse modulator test set, other RF modulators, bias systems) can radically change the physical pulses, hence the data interpretation. With a typical cable setup, the latency with the Anritsu pulse modulator test set is around 35 ns. These latencies are normally consistent from test to test. Once latencies are known for a given setup, the results should be interpretable fairly easily. If the setup contains a re-latching or resynchronizing process, there may be a random component to the latency period.

25-9 Pulse-to-Pulse Measurements

The third type of pulse measurement is pulse-to-pulse, where transient effects from the beginning of DUT excitation may be of interest. Thermal phenomena are often studied with this method.

From a measurement point-of-view, pulse-to-pulse is something of a hybrid of point-in-pulse and pulse profiling techniques. In point-in-pulse, a single window per period (relative to T_0) is being measured. In the pulse profiling, data is being collected and plotted versus time with frequency and power held constant. By definition, during pulse-to-pulse measurements, averaging between pulses is not permitted.

An absolute time reference is often needed in pulse-to-pulse measurement applications. Consequentially, some external or manual triggering is used to begin the sequence of events. In this example, the manual trigger on the VNA will be used. When the trigger is executed, the pulse generator outputs will activate and, in this case, begin applying bias and RF to the device. The purpose of the measurement is to capture the evolution of S_{21} of the DUT over the first 10 pulses. The PRI is 1 ms, and quadruplet pulsing on both bias (3 supplies for this DUT) and RF is desired.

The relative delays are 100 µs between each leading edge of the quadruplet, and the width of each sub-pulse is 10 µs. The desired measurement window is 1 µs centered within the last pulse of the quadruplet.

Setup

- PRI: 1 ms
- Pulse generators 1 to 4: quadruplet with delays of 0, 100 µs, 200 µs, and 300 µs, and all widths at 10 µs
- Measurement delay: 305 µs
- Measurement width: 1 µs
- IFBW: 1 MHz
- Frequency: 25 GHz

The processing window will apply to a_1, b_1, and b_2 in this example (off-state behavior is not of interest). The timing relationship for this measurement is shown in Figure 25-37.

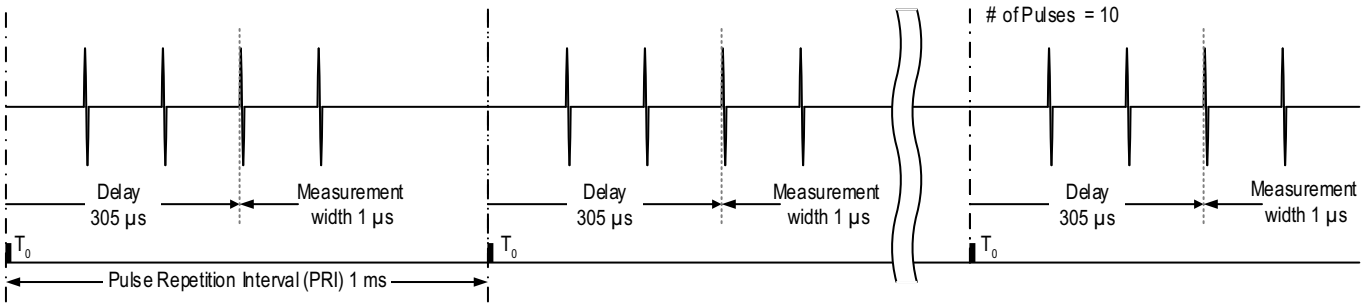


Figure 25-38. Timing Relationship for the Pulse-to-pulse Measurement Example

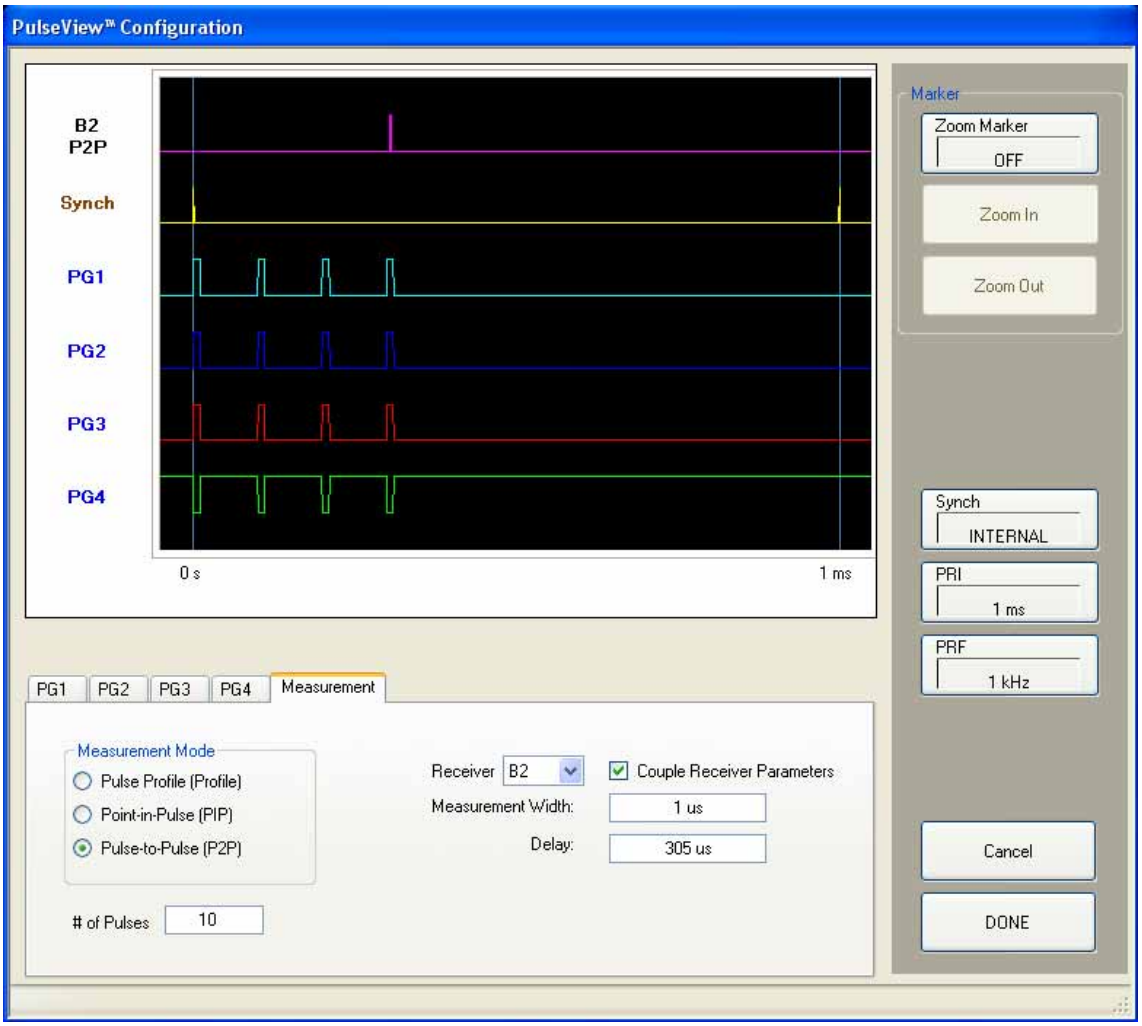
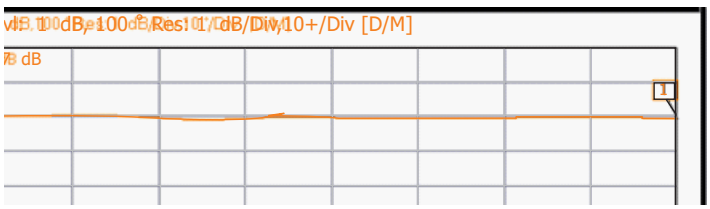


Figure 25-39. Pulse-to-pulse Example Setup Dialog

The Pulse Configuration dialog box is shown in Figure 25-39. Here, the receivers are selected to have coupled parameters, and the measurement (for all channels) is roughly centered within the final pulse of the quadruplet. The three external bias systems required positive polarity, while the RF stimulus requires inverted polarity.

The data for this measurement is shown in Figure 25-40. The magnitude shows little variation over the 10 pulses (< 0.1 dB), while the phase shows up to 5 degrees of variation towards the end of the measurement cycle.



Trace normalization was used for this measurement.

Figure 25-40. Measurement Results for the Pulse-to-pulse Measurement Example

Pulse-to-Pulse Measurement Results

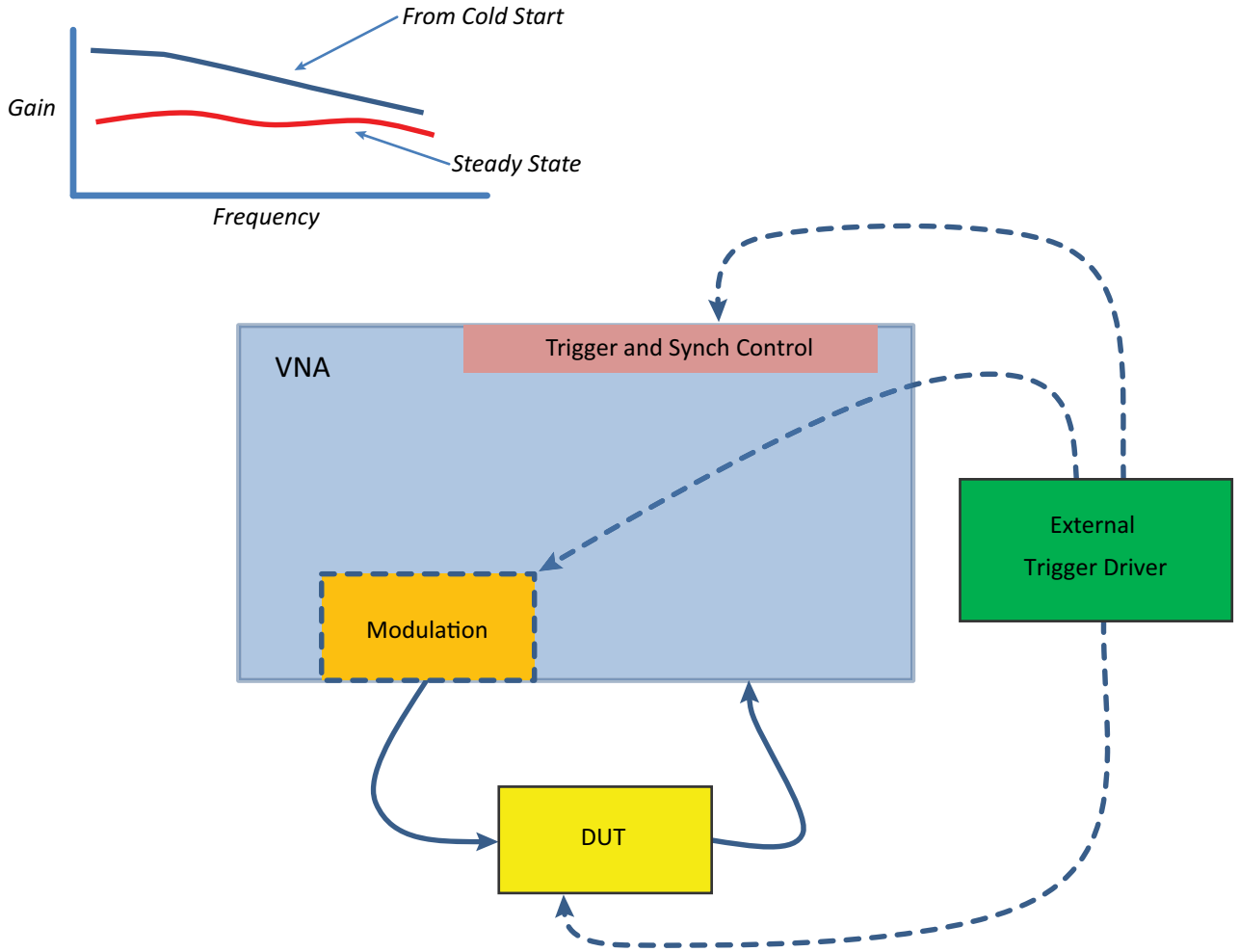
There are many ways to trigger this kind of measurement, and there are many potential variations depending on the hardware and measurement behaviors of interest. To help determine the measurement configuration, answering the following questions may be of help:

- What needs to be activated when the sequence begins? Some or all of the pulse drives? Some or all static drives (RF, bias, or control)?
- If the trigger is being generated externally, are there latencies involved and are they consistent? If being routed from the internal VNA trigger (via the pulse generators), do any external paths involved have latencies?
- Is it known how long the DUT must be ‘cold’ (time prior to the trigger event) to be in a steady state for whatever mechanism is of interest? If multiple mechanisms are of interest, is there a way of organizing the timing to isolate the mechanism of interest (sequencing biasing, for example)?

25-10 Continuous Point-In-Pulse (CPIP) Measurements

While the PIP measurements discussed earlier in this chapter may be adequate for many swept pulse measurements, there are cases, particularly related to DUT transient response, where more detailed control of the VNA sweep operation relative to the pulses and the measurements is required. This is the purpose of the Continuous Point-in-Pulse mode (CPIP) where the entire sweep of frequency or power is done during one acquisition. A use-case is shown in Figure 25-41.

The DUT may operate in a low duty cycle sense (either operationally or for thermal reasons) and then is required to operate over some frequency or power range as in a wide chirp sweep or a quick dynamic analysis sweep. How that DUT behaves during that sweep process may be very different from that when it had been operating steady state (if it could run steady state at all). The inset plot in Figure 25-41 shows an example of how such a DUT might behave under these two kinds of conditions. CPIP is designed to let one analyze the sweep response from almost any initial condition.



The DUT is not operating for a long period of time and then is required to operate over a frequency or power range immediately. One purpose of this CPIP mode is to analyze the response over that sweep while retaining transient information.

Figure 25-41. Example Use Case of CPIP

The concept is shown in more detail in Figure 25-42. Since the sweep is occurring during the acquisition, the digitizer memory will be loaded with blocks of data (denoted by the *d*s in the figure) each corresponding to a different sweep point. There will be (usually small) areas of disrupted data between those blocks where the system is changing frequency or power but the data can be "marked" (as in the external synch marking discussion earlier in this chapter) when the system is stable and the data can be analyzed from that point in the data record (or some controlled amount of delay after that).

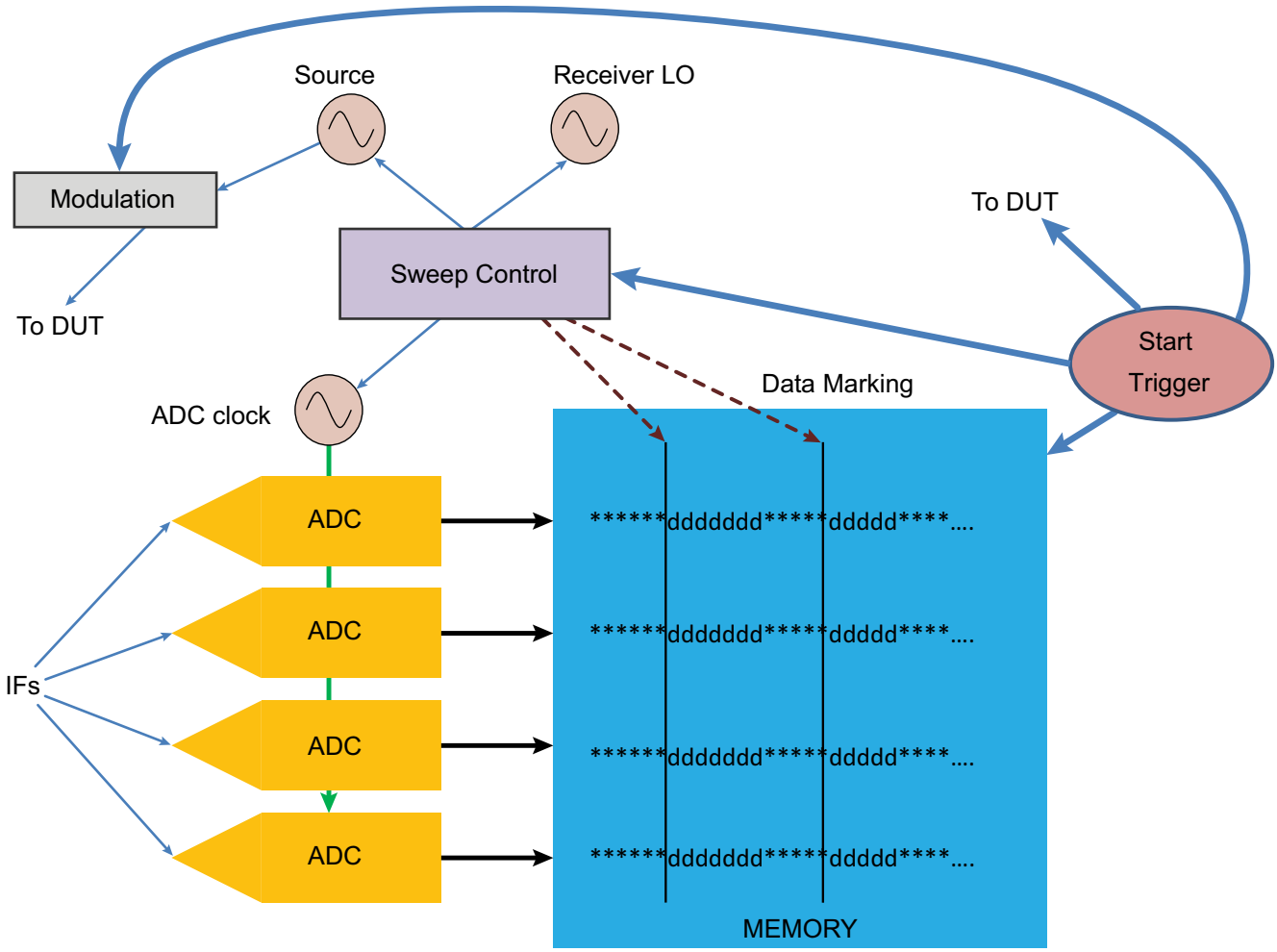


Figure 25-42. High-level Block Diagram Indicating the Marking Process for CPIP

Another key point is the triggering of the sweep since the DUT start-up transients may be great importance. Only per-sweep, per-channel and all-channel external triggering variations are allowed in this mode and the external trigger signal can be used to start the sweep, start the pulse generators and potentially bias/activate the DUT at the same time acquisition begins. Thus a controlled static starting state of the DUT and of the system is a possibility.

An important question is how does the acquisition get synchronized with the sweep? There are two main user-selectable sub-modes:

- Synch:

  An explicit synchronization signal is used in this case and a cable must be connected from the Trigger Out port on the VNA's rear panel to the Ext Synch In port (also on the rear panel). This provides the most reliable synchronization since the Trigger Out port knows about vagaries in the sweep timing (from bandswitch delays that the instrument requires, slow locking periods, etc.) and adjusts the marks in the data stream accordingly. The Trigger Out and Ext Synch In ports are required in this mode and cannot be used for other purposes. In earlier versions of software, the trigger out pulse is aligned with when the next frequency point is programmed. Thus some period of time (in the us range) is required after the synch pulse before the source is on frequency. For this reason, the delay in the UI is defaulted to 10us. Starting in 2025 versions, the trigger out is aligned to when the sources are reporting they are locked at the new frequency so the synch timing for settled measurements is more deterministic.

- Time:

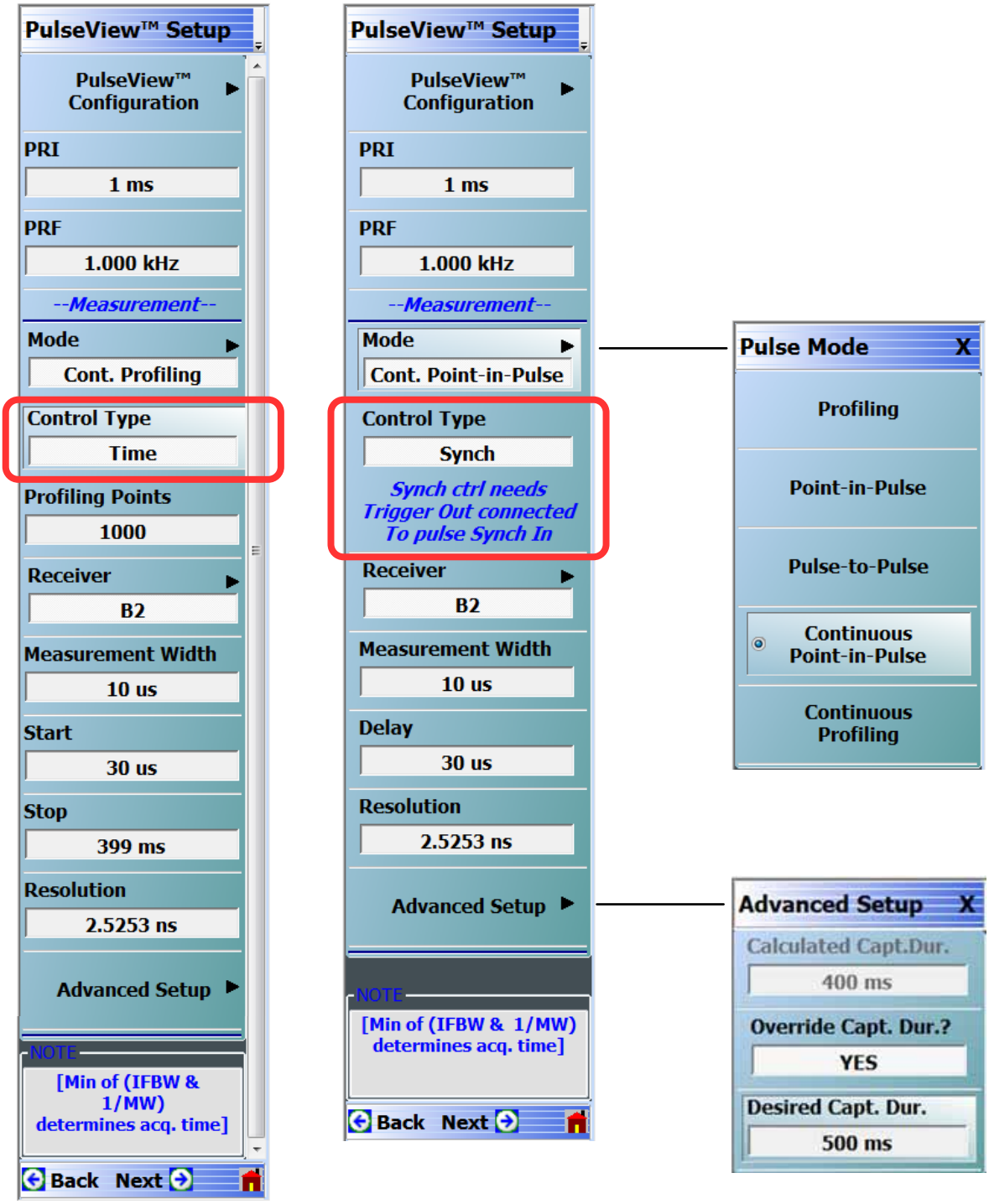
  This is a less explicit synchronization mode and does not require external connections. In this case, the acquisition system has less precise information about the sweep location so it is best suited to narrowband sweeps (not crossing bandswitch points for example) and for slower sweeps where the delay is much larger than the settling time.

In both cases, the sweep rate is set by the PRI (one pulse per sweep point unless burst is used) and the sweep timing is harmonized internally with that PRI. In Synch mode, PRIs as low as 100 µs to 200 µs are possible whereas in Time mode, PRIs of at least 5 ms are recommended (and at least 20 ms if bandswitch frequencies are crossed). As a reference, the common bandswitch frequencies are 2.5 GHz, 5 GHz, 10 GHz, 20 GHz, and 40 GHz (add 38 GHz for 50 GHz and 70 GHz systems; add 54 GHz, 80 GHz, 110 GHz, and 120 GHz for ME7838A/AX systems; add 54 GHz, 80 GHz, 110 GHz and 130 GHz for ME7838D systems). The maximum pulse widths that are practical scale with PRI. The minimum pulse width is not technically restrained beyond the system 5 ns limit. Narrow pulse measurements very close to the synch mark (i.e., low delay) may be difficult due to post-synchronization settling but may be fine for larger delays. The recommended delay (measurement and pulse) varies with frequency and power settings but at least 10 µs for power levels above ~ –12 dBm and at least 50 µs for power levels of –20 dBm and below are common (ALC final settling times are faster at higher power levels).

While there are no strict point count limits, the total acquisition time is still limited by the memory on the digitizer board (0.5 s for the standard 2 GB of memory at the default resolution near 2.5 ns, 2.5 s with Option 36 installed). Thus if 1 ms is the PRI, ~ 499 sweep points are possible for 2 GB of memory at a ~2.5 ns resolution.

The acquisition length defaults to **Number of Sweep Points * PRI**, but this can be adjusted (usually longer) if the triggering mode being used is adding inconsistent delays, or if there are related problems where the last few sweep points show drop-outs. If the synchronization marks are not received within the acquisition length, the output wave variables will be assigned the value –400 dB.

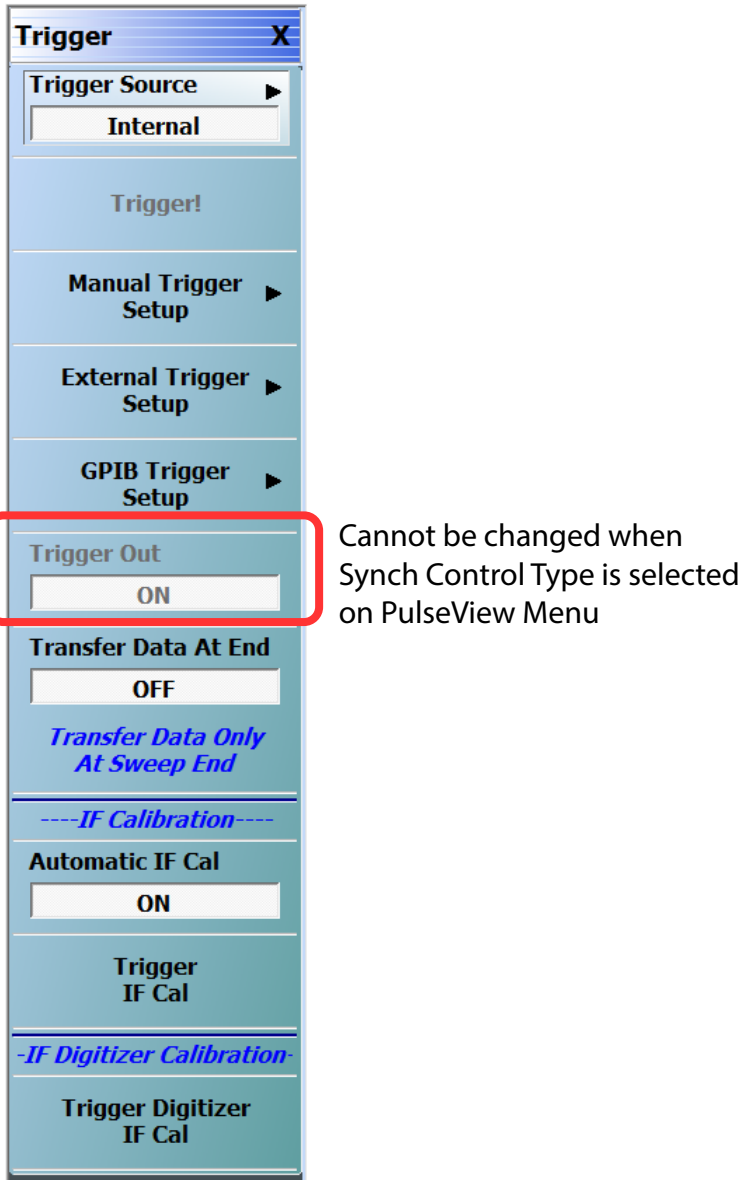
A few highlights of the menu choices when in this mode are pointed out in Figure 25-43 on page 25-43. The control type (Time or Synch) becomes available in CPIP (and in continuous profiling mode, or CProf, as will be discussed). Also the Advanced Setup button which leads to a submenu for entering acquisition length override and values becomes available.



Control type (Time or Synch) becomes available in Continuous Point in Pulse (CPIP) and in Continuous Profiling (CProf) modes. The Advanced Setup submenu also becomes available to provide additional PulseView setup items while in these modes.

Figure 25-43. Control Type Becomes Available in CPIP and in CProf Modes

When the Synch control mode is chosen, one would note on the Trigger menu that Trigger Out has been put into the ON state and cannot be changed (Figure 25-44). This would not be the case for the Time control mode.

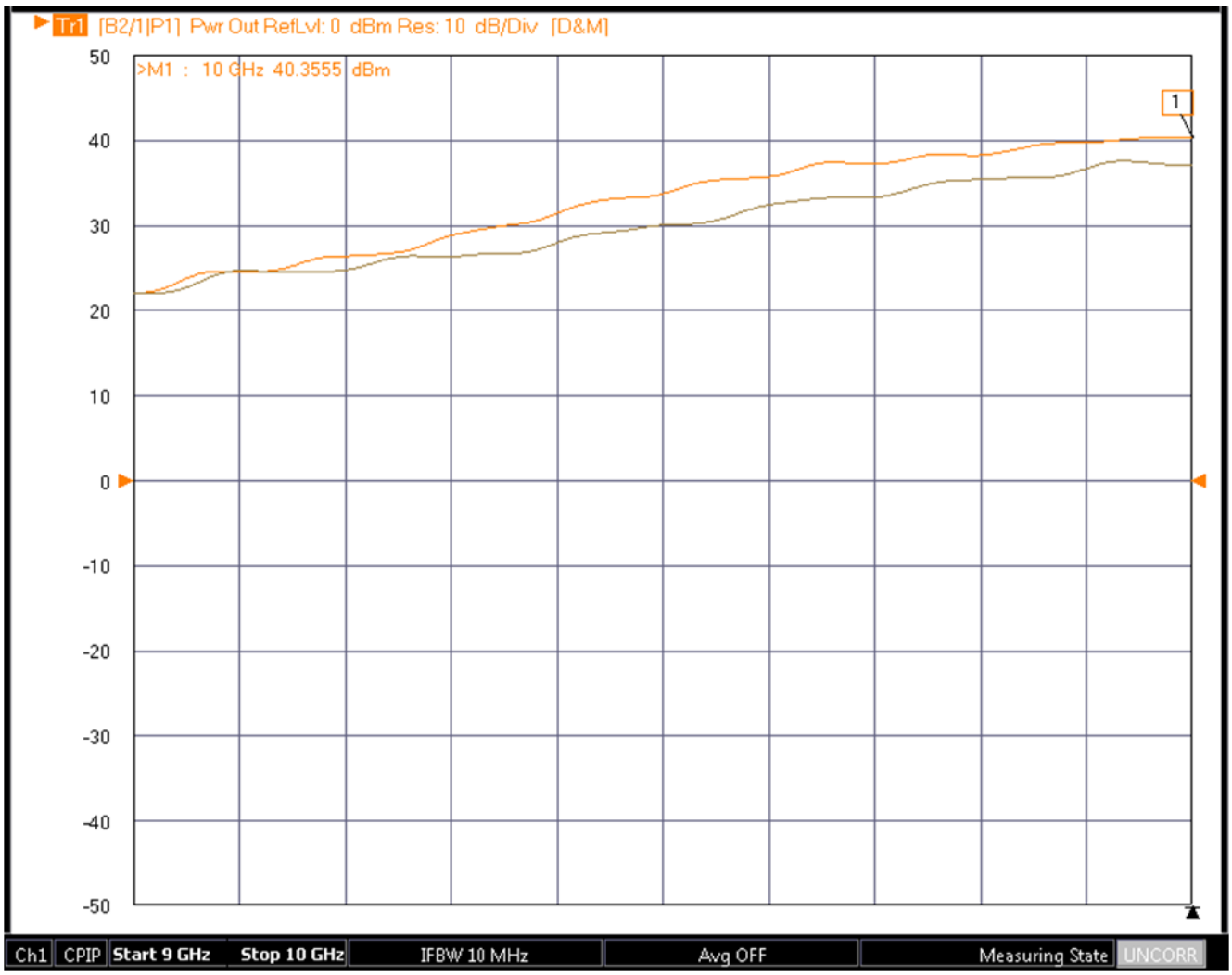


The Trigger Out function is required to be on for this sub-mode.

Figure 25-44. Trigger Menu when in CPIP Mode with Synch Selected

As with the other pulse modes, user calibrations should be performed in this mode for measurements in this mode. All of the calibrations available for PIP are available for CPIP (frequency response, 1-port, 1 path-2-port, full 2-port calibrations, all 3-port and 4-port calibration variations, receiver and power calibrations, etc.) Since the sweep dynamics and measurement profiles are quite different, calibrations cannot be shared between PIP and CPIP. Note that measurements and calibrations are still indexed with frequency or power, respectively, and not time since each measurement is still associated with a sweep point.

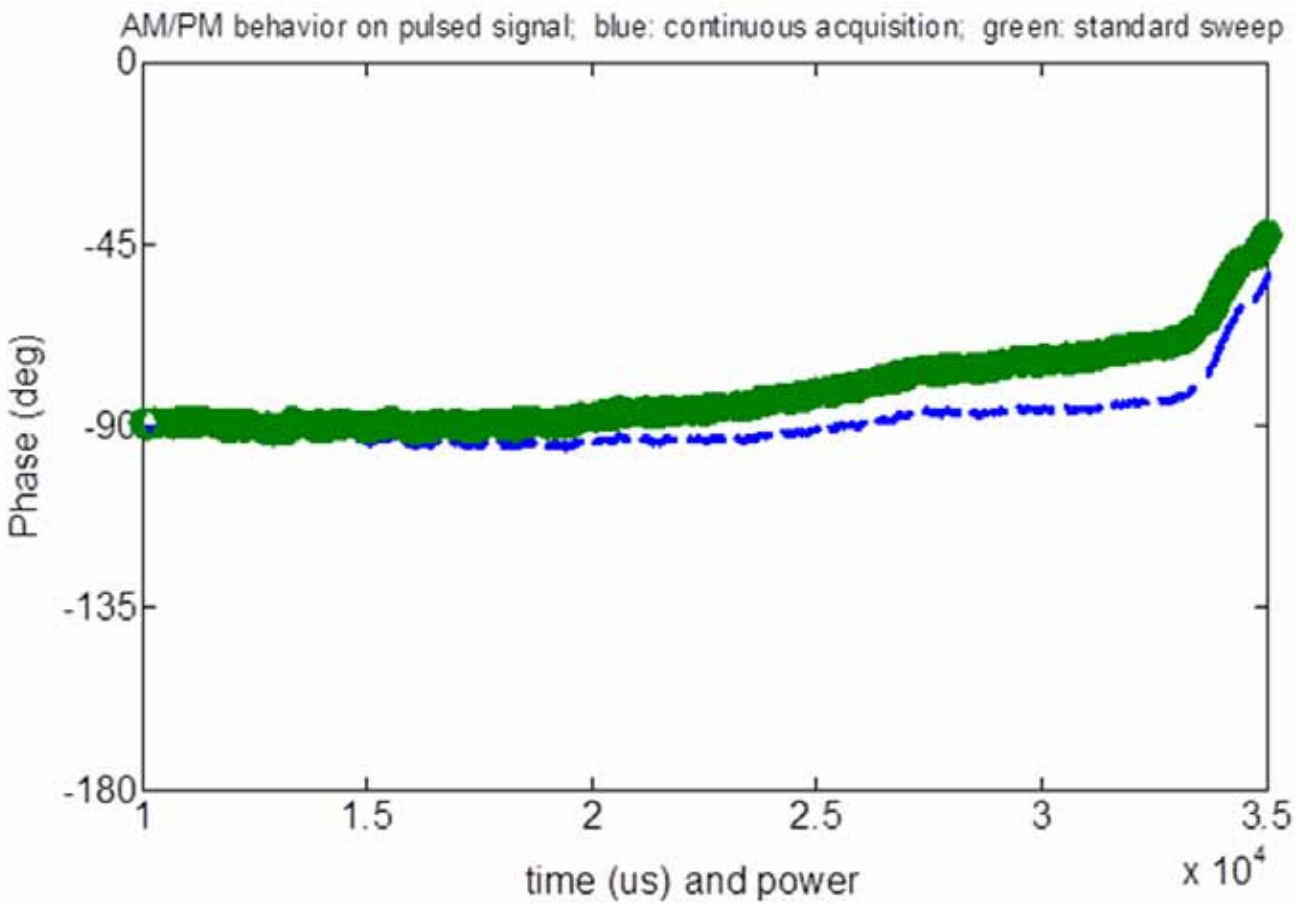
One example measurement is shown in Figure 25-45. We are interested in the device output power vs. frequency and the plot shows both the steady state PIP response (lower trace) and CPIP response (upper trace). In the steady-state response, there is some power reduction, presumably from heating, at the higher power levels available at higher frequency. This is one example where the DUT dynamics can convolve with the frequency or power response to produce an entirely different measurement result. By performing the measurement under the actual time-sequence-of-events that is of interest, a better understanding can sometimes be achieved.



An example measurement of DUT output power vs. frequency is shown here from a cold start CPIP measurement (upper trace) and from a steady-state measurement (lower trace).

Figure 25-45. Example Measurement of DUT Output Power vs. Frequency

A second example is shown in Figure 25-46 and involves an AM/PM measurement of a device. Here the power is sweeping from –20 dBm to 0 dBm over 35 ms and is plotted externally in time in order to give another vantage point. When measured with a standard PIP sweep (thick green trace), there is a general phase growth and a slight uptick at higher power levels. With the CPIP sweep (dashed blue trace), the lower phase growth is much reduced, from a different thermal and bias timing, and the high power growth is much more distinct.



An example CPIP AM/PM measurement is shown here relative to a standard PIP measurement of the same device. The different evolution again shows the potential for dependence on the exact sequence of events. The x-axis is labeled in time but corresponds to an incident power range of –20 dBm to 0 dBm.

Figure 25-46. Example CPIP AM/PM Measurement

Additional Comments Regarding CPIP Mode:

- Because the integration with sweep timing is central to this mode, measurements across multiple periods are not allowed. The IF bandwidth setting will have no effect in CPIP mode.
- For the same reason, re-measurements within a sweep are not allowed so gain ranging is automatically disabled in CPIP.
- For similar reasons, there are some limitations on triggering. Per-point manual and GPIB triggering is not permitted with the Synch sub-mode since the delays there may exceed total acquisition time.
- As mentioned in the text, sweep length is limited by the PRI (basically the per point time) and the available memory. For ~2.5 ns resolution and 2 GB of memory (standard), 0.5 seconds is available.
- Receiver frequency operation below 110 MHz is not allowed in CPIP as this would necessitate ADC clock frequency changes in the middle of the sweep.
- As mentioned in the text, sweep length is limited by the PRI (basically the per-point-time) and the available memory. For ~2.5 ns resolution and 2 GB of memory (standard), 0.5 seconds is available. With Option 36, that value increases to 2.5 seconds.

25-11 Continuous Profiling (CProf) Measurements

The Continuous Profiling (CProf) measurement also uses a single acquisition for an entire frequency or power sweep but is a generalization in data analysis relative to CPIP. In CProf, one is not restricted to a measurement window associated with each sweep point but one can have up to 25000 measurement windows dispersed over the entire sweep time or located in some subset of that range. The number of displayed points is equal to the number of measurement windows employed. The acquisition length is limited by memory again so the sweep point count * PRI is capped just as in CPIP but here the number of analysis points is an independent concept.

The central idea is sketched in Figure 25-47 where the sweep starts as in CPIP with a data mark. In CProf, the measurement windows are specified with the familiar start delay, stop delay, measurement width and number of analysis points. In the figure, only the first two measurement windows are shown but they are positioned relative to just the first synch mark when the sweep began. This gives the user considerable flexibility on where to look at data (in time) within each sweep point.

The higher level description of the application for the CProf measurement is similar to that for CPIP but where the requirement is for greater flexibility in looking at the data in time. As such, this measurement approach has applications for many time-running setups including antenna rotation/antenna pattern investigation, evaluation of DUTs with sequential operating states, sequential bias analysis problems, and many others.

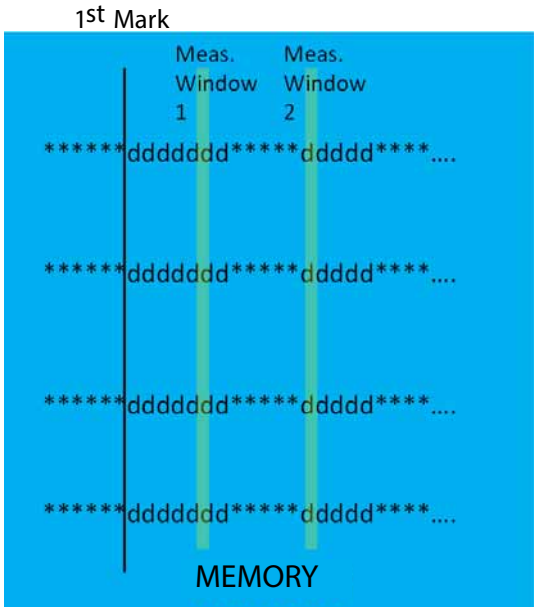
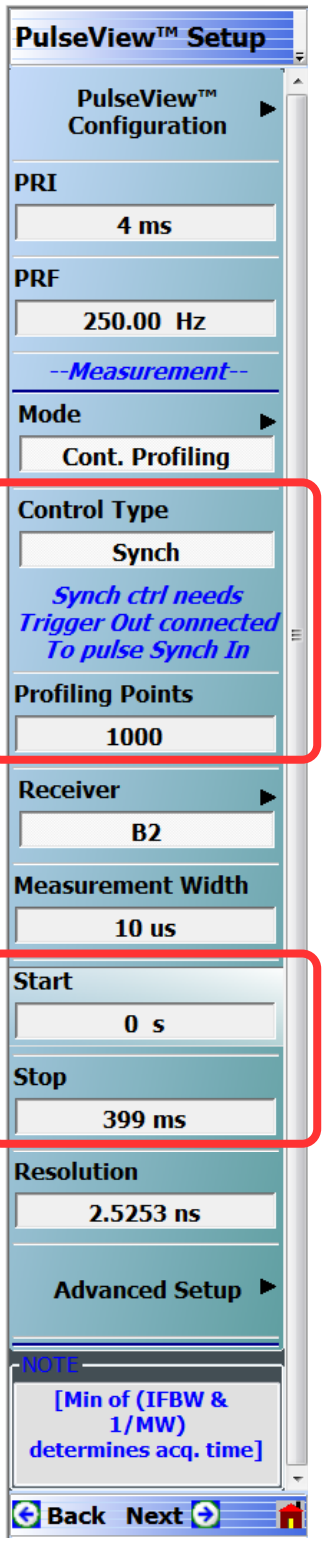


Figure 25-47. Measurement Window Positioning in CProf

From the above discussion, one can observe that synchronization control is much less critical for this measurement, at least in terms of collecting the data but the synchronization state will still determine when that first mark is recorded so the Synch mode still has value in determining where one would like to analyze data. The same triggering options and sweep selection options apply as for CPIP. To reiterate, this is in contradistinction to regular pulse profiling which is inherently a fixed-frequency, fixed-power measurement. That difference also leads to the aforementioned difference between sweep points and analysis points. The former denotes how many frequencies or powers will be swept (and is set on the frequency or power menu, respectively). The latter denotes how many measurement windows within the data record should be analyzed. There is no correlation between these two numbers. Either one may be larger than the other.

The PulseView menu for CProf only differs slightly from that for CPIP (see Figure 25-48). There is the addition of start and stop delays to indicate the profiling time interval (just as in regular pulse profiling) and there is the addition of the field for Profiling Points. As discussed previously, this is the number of measurement windows within the time record that will be analyzed and is not correlated with the number of sweep points.



Start/Stop Delays

The advanced setup addition and control type are the same as for CPIP. The number of profiling points is a new addition for the CProf mode.

Figure 25-48. Menu for CProf

Traditional S-parameter calibrations are used less often with CProf since there are points in the time record that may not correspond to a particular sweep point and the exact locations may vary with electrically long devices (working with window positioning can help as with Profiling but it becomes more complicated here due to the sweep interaction). Normalizations and related processes are frequently employed.

An example CProf measurement is shown in the top part of Figure 25-49 with a relatively small number of profiling points (100 with 50 sweep points). The points generally pick out each sweep point with some duplication. The same sweep but with 1000 profiling points is shown in the bottom part of Figure 25-49. Here the display is much more step-like since multiple measurement windows are processed while the hardware is at the same sweep point (the number of plotted display points is equal to the number of profiling points). There are also some drop-out points where the hardware is transitioning between sweep points. While more challenging to interpret, this allows one to study the interaction of DUT dynamics with the changing frequency or power of the sweep in detail.

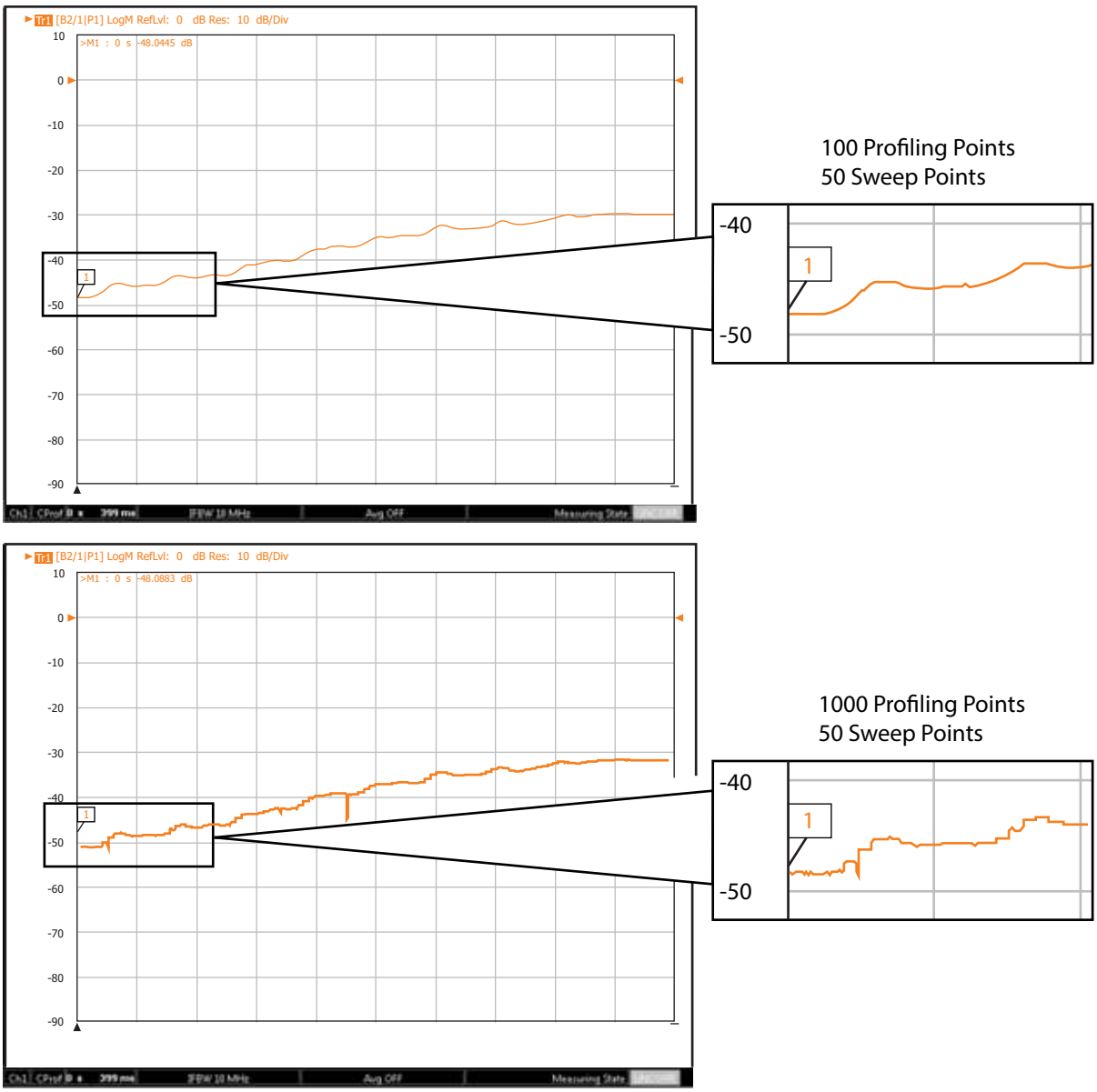


Figure 25-49. Example CProf Measurement with Same Number of Sweep Points, but Different Numbers of Profiling Points

A slightly different kind of example is illustrated in Figure 25-50. Here a surrogate antenna pattern is being evaluated while frequency is sweeping. In the first case (an omni-directional emulator) has no frequency response over the interval. In the second case, the emulator has a high pass response so one sees a spiral formation in the 'pattern'. This example helps show the flexibility possible with this measurement mode: the acquisition and analysis parameters are less constrained by the frequency/power sweep.

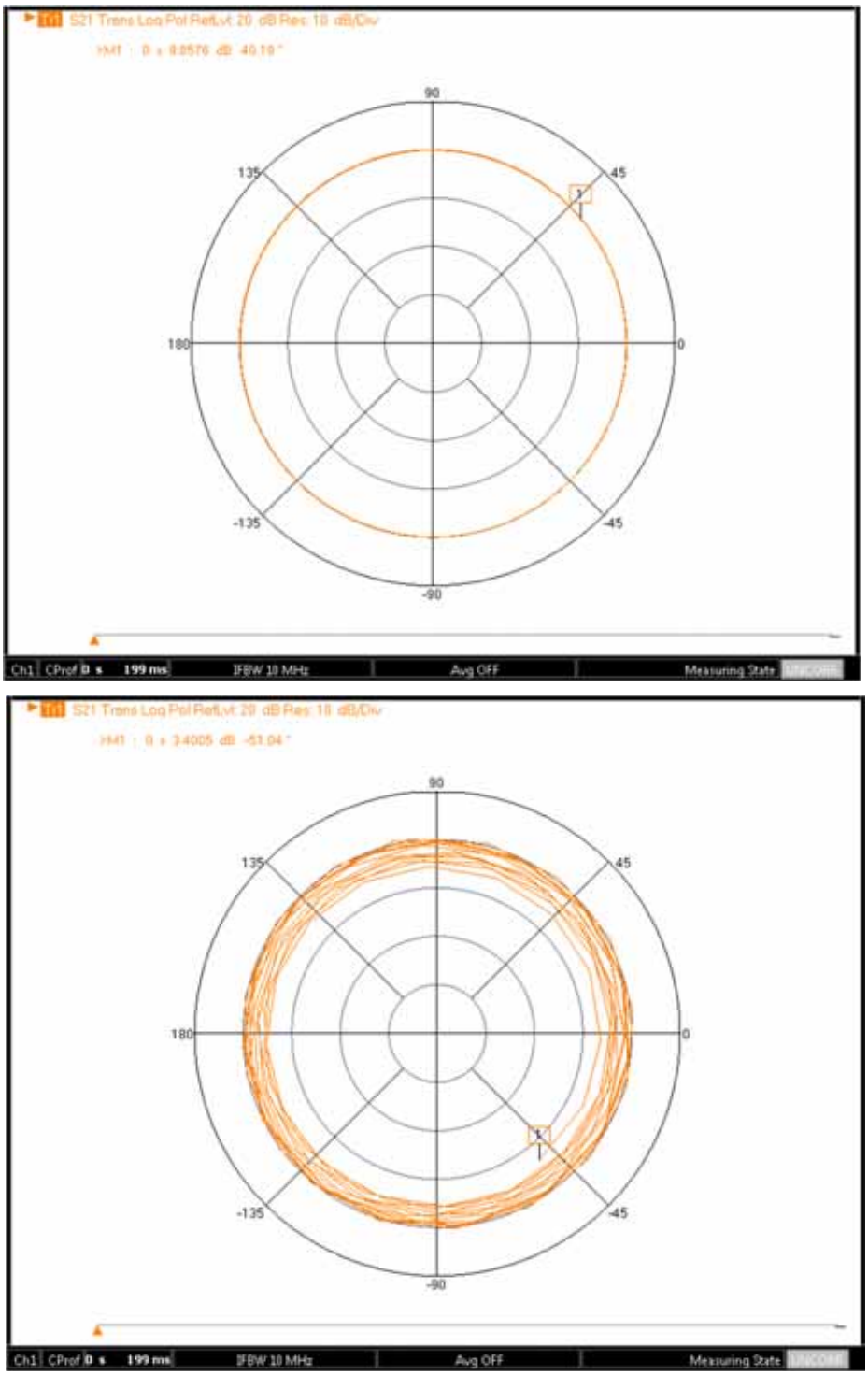


Figure 25-50. Example CProf Measurement for Two Different Antenna Emulator Structures (1 of 2)

An example CProf measurement for two different antenna emulator structures showing the flexibility of the measurement mode is presented here.

Figure 25-50. Example CProf Measurement for Two Different Antenna Emulator Structures (2 of 2)

Additional Comments Regarding CProf Mode:

- Because the integration with sweep timing is central to this mode, measurements across multiple periods are not allowed. The IF bandwidth setting will have no effect in CProf mode.
- For the same reason, re-measurements within a sweep are not allowed so gain ranging is automatically disabled in CProf.
- For similar reasons, there are some limitations on triggering. Per-point manual and GPIB triggering is not permitted with the Synch sub-mode since the delays there may exceed total acquisition time.
- Because sweep point indexing is not directly available in this mode, full calibrations are normally quite difficult although normalization approaches can be readily used.
- As mentioned in the text, sweep length is limited by the PRI (basically the per point time) and the available memory. For ~2.5 ns resolution and 2 GB of memory (standard), 0.5 seconds is available.
- Receiver frequency operation below 110 MHz is not allowed in CProf as this would necessitate ADC clock frequency changes in the middle of the sweep.
- As mentioned in the text, sweep length is limited by the PRI (basically the per-point-time) and the available memory. For ~2.5 ns resolution and 2 GB of memory (standard), 0.5 seconds is available. With Option 36, that value increases to 2.5 seconds.

25-12 Uncertainties

Uncertainties in pulse measurements are not often covered in great detail. In an ideal situation, possible uncertainties would be dominated by basic S-parameter uncertainties controlled by calibration kits and methods, noise floor and trace noise, linearity, repeatability, drift, and similar factors.

While it is not within the scope of this chapter to thoroughly analyze all of the pulse distortions possible, the most likely types of uncertainties are outlined below.

Time

When troubleshooting, address this variable first. Does the S-parameter reported at a particular time relative to T_0 represent what was intended? This can be an issue due to latency in the pulse modulators, external bias systems, the DUT itself, and the VNA. A broad profiling measurement is sometimes a way of assessing the net latency and ensuring that a given time point is indeed the one of interest.

Shaping

Stimulus shape can affect the DUT output. This shaping may arise from the behavior of the RF modulators used or the behavior of any bias system pulsing that is employed. The former can normally be calibrated out, while the latter is an intrinsic part of the measurement.

Calibration

The calibration process itself may introduce some oddities as discussed earlier, particularly when stimulus RF modulation is in play. If the edges of the pulse response are of interest, then the calibration may be acting on reduced dynamic range on parts of the pulse edge resulting in elevated net trace noise.

Minimizing Uncertainties

To help reduce uncertainties, increases in averaging or reduction in IFBW in these measurements are advised—along with not trying to measure deep into the noise floor unless necessary. Power level increases can sometime also help if the DUT can support it and one avoids receiver compression.

Nominally, 0.1 dB compression is above +10 dBm at the port and higher, if the step attenuators of Options 61/62 are engaged. If accessing the loop paths to the receivers, compression levels will be lower by the coupling factors of the couplers/bridges. Nominally this value is 13 dB.

Other artifacts can occur with RF stimulus modulation, particularly at low frequencies. These include video energy from the pulse drive itself that can appear on the frequencies of interest.

The Anritsu measurement test employs high pass filters for frequencies greater than 2.5 GHz. If at lower frequencies or using an external RF modulator, one may want to apply a high pass filter to the stimulus.

Other approaches to reduce uncertainties may include pre-filtering the pulse drives to soften the edges somewhat and reduce video harmonic content. As has been discussed earlier, the use of a pulsed reference signal (available in the Anritsu measurement test set for routing to the reference loops on the VNA) can help reduce these video effects somewhat.

To help illustrate some of the concepts and to provide hints as to how the uncertainties will vary in this vast parameter space, consider two examples. In the first, video effects and transients are neglected. The measurement may be a PIP measurement in which the measurement window is well away from the pulse edges, or it could be the middle of a pulse in a pulse profiling measurement. A plot is shown in Figure 25-51 under the following conditions:

- Pulsed bias only (stimulus modulator loss not included, nominally 5 to 6 dB at 40 GHz for high on/off ratios such as in an Anritsu Pulse Modulator Test Set; double that value if also using the receiver-side RF modulator)
- 1 kHz net IF bandwidth
- Sweep-by-sweep averaging of 5 sweeps

- VectorStar VNA default source power

As with all digitizing measurements, this result is independent of duty cycle, unlike with the single spectral line methods.

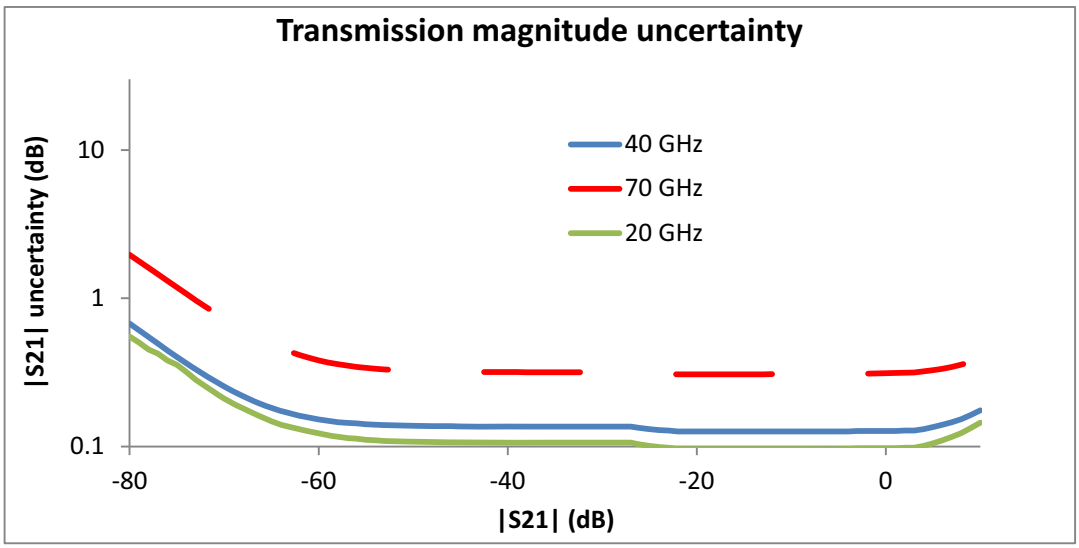
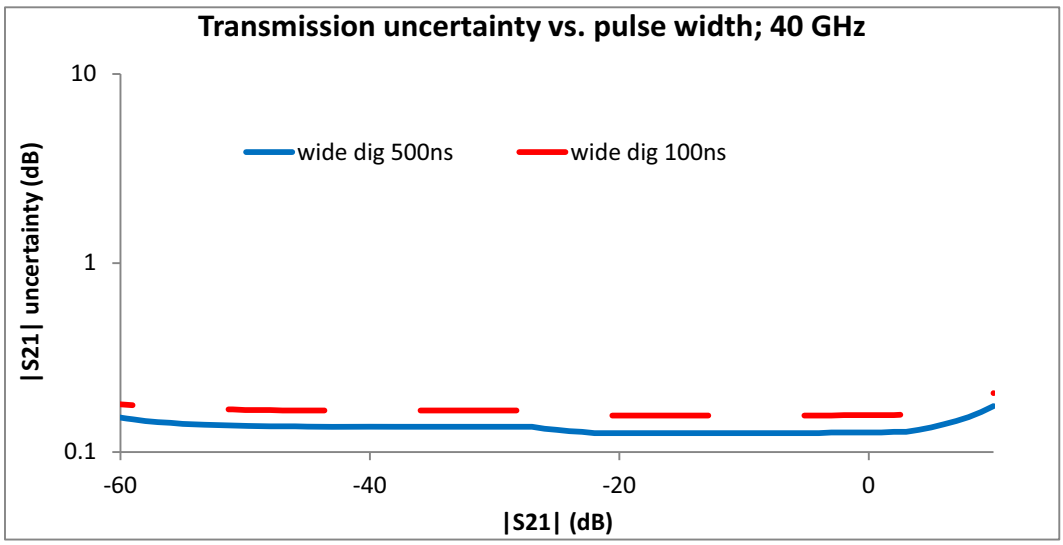


Figure 25-51. Estimate of Transmission Uncertainty for a Mid-pulse Measurement Under Certain Conditions

In the second example, consider a specific overshoot measurement, where the pulse had about a 20 % overshoot of width ~20 ns. Since the transient behavior is critical here, the pulse width relative to the IF bandwidth of the system comes into play. Correspondingly, calculations will be performed for a couple of pulse widths. The narrower pulse has a wider spread of energy, so the hardware filter edges distort the content more than for a wider pulse. The PRI was kept relatively low in this case, making the individual spectral line behavior less important. The effective IFBW was again low and the power setup from the previous example applies here. One can see the pulse width effect in Figure 25-52, but it should be emphasized that the values are specific to the particular pulse waveform under study. The central point is that all of the measurement parameters can play a significant role in pulse measurement uncertainty.



An example uncertainty computation for the value of pulse overshoot as a function of DUT transmission and pulse width.

Figure 25-52. Example Uncertainty Computation

25-13 Synchronizing with External Hardware

Aside from complete measurement triggering (as discussed earlier in the P2P example), most of this chapter has assumed the use of the internal pulse generators to perform the measurements. When using external pulse generators or when timing must be controlled entirely by an external system, there are some additional issues to consider:

- Recall from the data processing discussion that a central concept is the marking of the data stream with markers corresponding to T_0 so that the correct data segments can be analyzed. Normally, that T_0 timing is generated internally and also used to drive the pulse generators. The T_0 timing signal is provided to the rear panel so it can be used on external equipment (Pulse Synch Out). The internally generated T_0 signal is frequency locked to the data acquisition system, so jitter is minimized.
- If connecting the Pulse Synch Out to an external system, that system must be set for the same PRI for any reasonable timing and, for it to be precise, the 10 MHz timebases must also be linked. If timebases other than 10 MHz are used in the external system and are accessible, contact the factory for more information on how to link them to the MS464xB Series VNA.
- A Pulse Synch In connection is also available to allow an external system to provide the T_0 timing. This can be used in two ways:
    - If the VNA and external 10 MHz timebases can be linked, then the default mode is to let both systems run (set for the same PRI of course). The Pulse Synch In signal is used to phase align the T_0 signals. This is enabled by selecting either EXT RISING EDGE or EXT FALLING EDGE on the PulseView™ Configuration dialog (under Synch). The internal pulse generators continue to be available with their timing controlled by the internal T_0 system that has been aligned to the external system.
    - If it is not practical to link timebases, then the external T_0 signal can be used to directly mark the data. This is enabled by selecting either EXT RISING EDGE or EXT FALLING EDGE and clicking the Ext. Synch Marking checkbox on the Measurement tab of the PulseView™ Configuration dialog. Again, the external PRI should match the internal PRI setting, although the accuracy of this agreement is less important (accurate settings will help to avoid missing some pulses or being overly limited in profile or delay ranges). In this mode, the internal pulse generators are disabled (since the internal T_0 system is disabled) and there may be increased jitter since the data marking is incoherent with the data acquisition system. As a result, this method is not recommended for very narrow pulses (< 200 ns).

25-14 Pulsed Millimeter-Wave Measurements

Pulsed measurements are also quite relevant at millimeter-wave frequencies, particularly when striving for isothermal operation of on-wafer devices. The purpose of this section is to explore how the setups and measurements change when going to the broadband/millimeter-wave realm with ME7838 broadband systems (with maximum frequencies currently ranging from 110 GHz to 226 GHz).

Basics

Since the pulse measurement options are based on wideband IF processing, one would expect that, on the surface, little would change when going to millimeter-wave frequencies. Indeed, this is largely the case. Pulse timing control and synchronization remain the same as for the base VNA, but one may notice added jitter for the finer timing resolutions because the available IF bandwidth in the modules for the broadband systems is slightly reduced. Still, high resolution measurements are possible; an 81 GHz example profile measurement is shown in Figure 25-53 with a time range of 250ns, and considerable detail is evident.

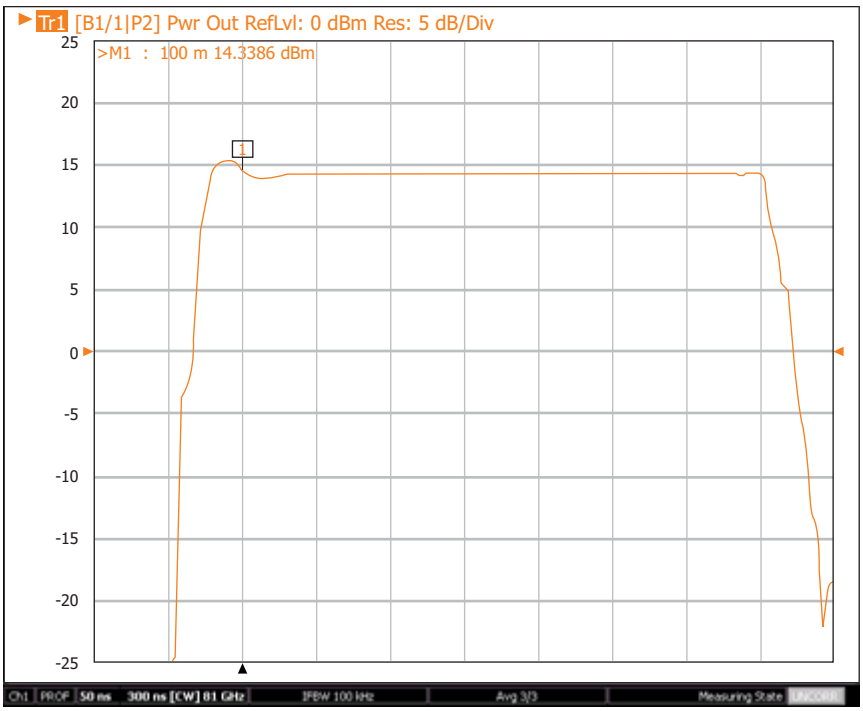


Figure 25-53. Example Millimeter-wave Profiling Measurement Illustrates Available Resolution

If the measurements do not require RF drive modulation, there are not many other differences from pulsed operation with the based VNA. Time-gated power calibrations, which might have been done at DUT output even when the drive was not pulsed, are not available. This is mainly because of the lack of wide instantaneous bandwidth pulsed power sensors at millimeter-wave frequencies (as of this writing).

Millimeter-Wave Measurements with RF Drive Modulation

As with lower frequency pulsed measurements with modulated drive, a pulsed test set (or other modulator) is required. The setups do vary slightly because of how the stimulus signal flows to the port. Recall the structure of the broadband system (see Chapter 17 of this guide for more information) as shown in Figure 25-54. For frequencies nominally below 54 GHz, the VNA synthesizer directly supplies the drive through (essentially) a pass-through transmission line in the module. Thus, pulse modulation applied in a source loop of the VNA or between port 1 of the VNA and the module will work as would be expected. Accessing through the source loops is preferable if match correction or reversing measurements under 30 GHz are desired (30 GHz is nominally when the module receivers take over; the base VNA receivers are used below that frequency).

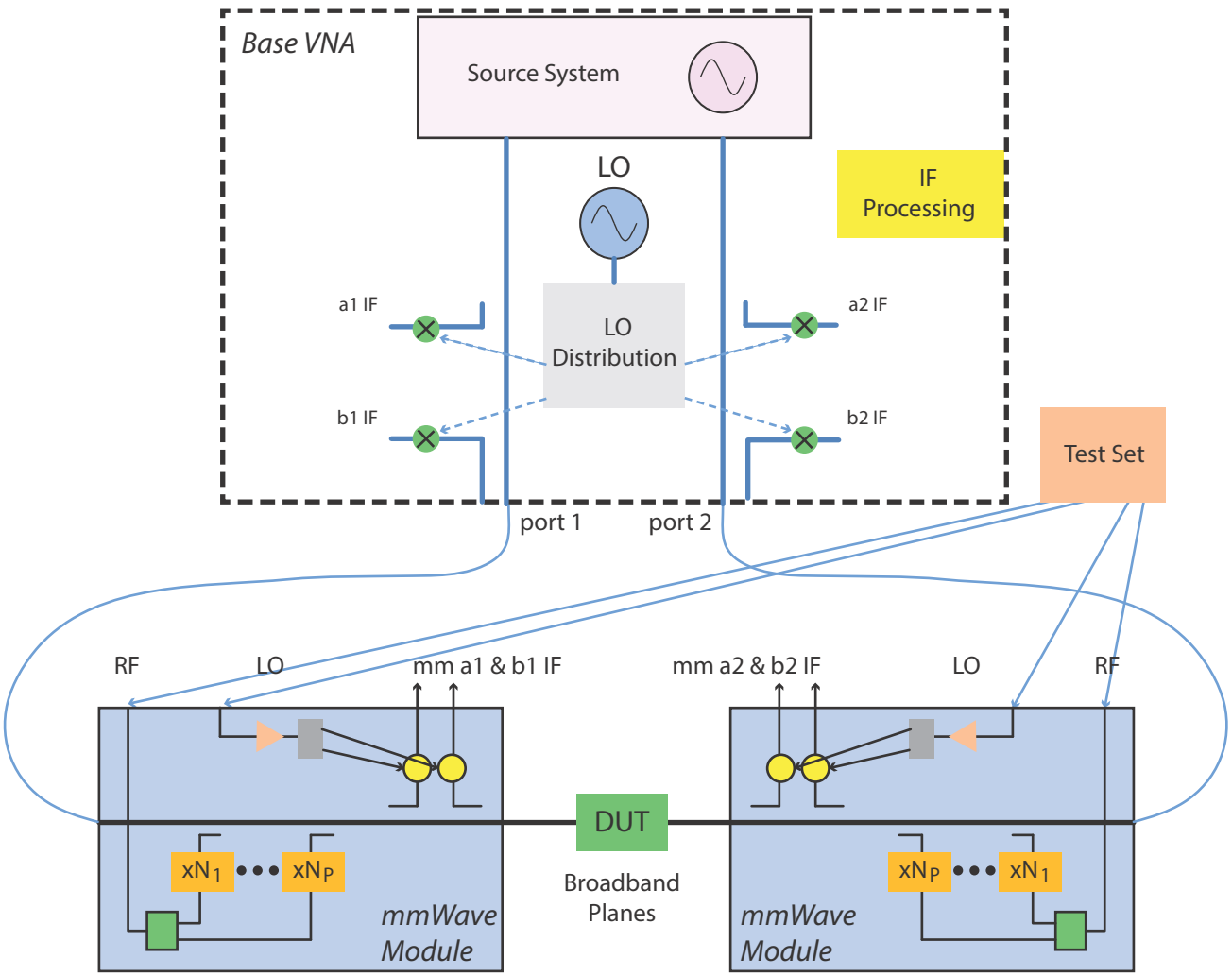


Figure 25-54. General Broadband System Architecture for Reference

Notice that the RF entry into the modules, for the multiplied bands above 54 GHz, presents a bit more of a challenge:

- The modulation has to occur before the module (no later access points are available) so pulsing will be done before the final multiplier. This has both advantages and disadvantages. The P_{out}-P_{in} slope of the multipliers is always >1 (an example is shown in Figure 25-55) so there is an improvement in on-off isolation. However, the multipliers may introduce some video distortion.

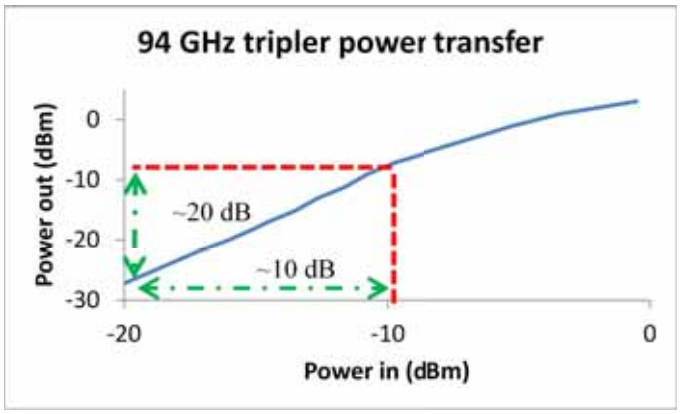


Figure 25-55. Example Multiplier Transfer Curve

- Because the 3739X test set (part of the broadband system) uses a transfer switch to route the RF drive and has some amplification, it is usually desirable to modulate before the test set when operating above 54 GHz. This is suggested in Figure 25-56.
  - A source modulator is here placed in the RF drive path and both module port 1 and module port 2 will be modulated when they are driving. The modulator could be placed between the test set and the module as well (slightly less flexible but, since later in the chain, there may be less video distortion; note that the levels in both locations may exceed 5 dBm).
  - The RF frequencies going to the modules do not exceed 40 GHz. The minimum frequency on this RF drive path for the module is 26 GHz for 110/125 GHz systems, 18 GHz for 145 GHz systems and 12.75 GHz for 220 GHz systems. The frequency ranges for banded millimeter-wave modules from 3rd-party vendors vary (consult the source multiplier values from the vendor).
  - The receive-side modulators in the test set are not particularly useful for millimeter wave as there are no RF loops in the receive path and the frequency range availability would not match anyway. The receive-side modulators could conceivably be used on the IF path (low-band modulators) but there is little de-sense that could cause problems at that level in the architecture.

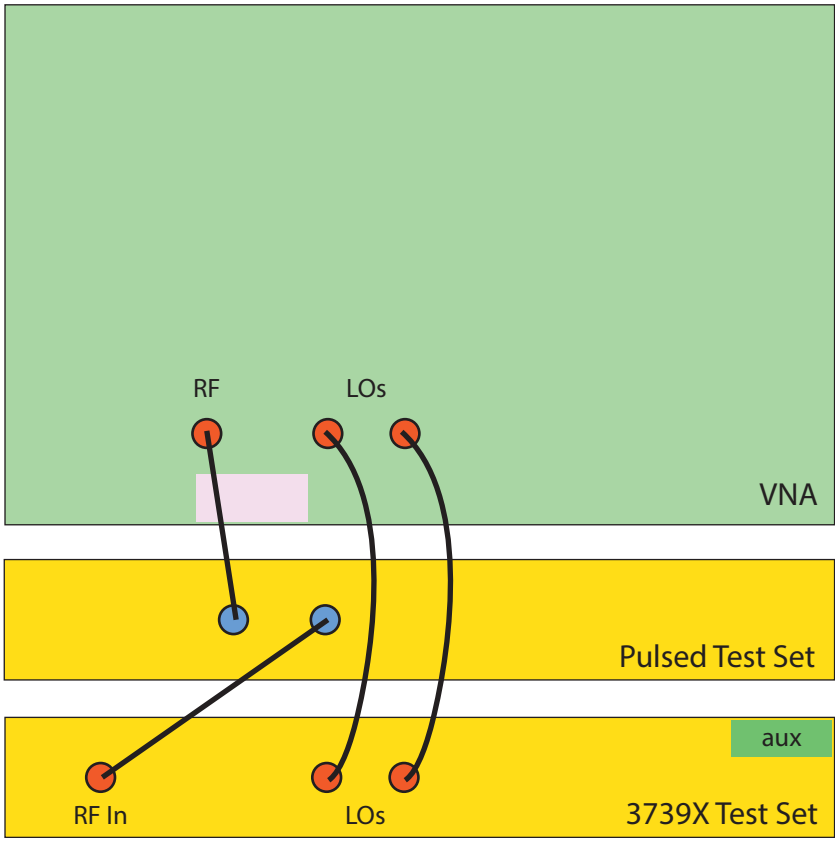
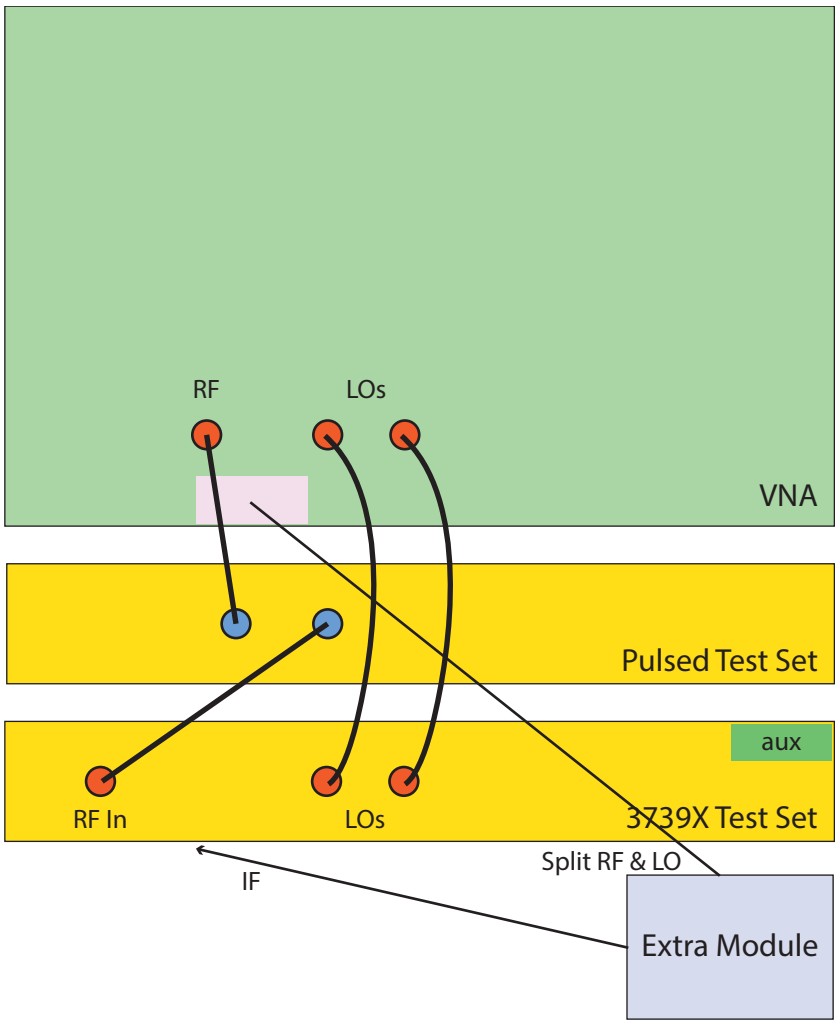


Figure 25-56. Normal Position for Drive-side Pulsed Modulation for Millimeter-wave Frequencies

- In normal 3739 modes, the system uses IF leveling so the pulsed RF presents some changes in how ALC control works.
  - The average signal level at the detection point will be down by `20*log10(duty cycle)` so the leveling loop will compensate by increasing the drive level. Thus, if one does nothing, the entry on the power menu will now represent an average power rather than an 'on state' power and the value may have to be dropped considerably to avoid unleveled messages and very high peak powers.
  - User power calibrations can be done while pulsed and it will help to manually enter the duty cycle/duty factor on the power meter. With such a calibration, the power menu will again be reading in terms of ‘on state’ power.
  - Because of the drop in detected signal level for a given on-state power, the power control range will be reduced from non-pulsed conditions (generally by about the dB form of the duty cycle).

- An auxiliary module can be used to level the system based on CW power (see Figure 25-57) which will simplify the power entry interpretations and calibrations at the expense of additional hardware. This extra module would use a split off sample of the pre-modulator RF drive and a split off sample of the module LO to generate a local IF for leveling. Its control cable would connect to the port 1 position. The main port 1 module's control cable would connect to the aux module connector shown at the right on the test set.



An extra millimeter-wave module can be used as a leveling hub to simplify power control in pulsed drive scenarios above 54 GHz (IF leveling). This module's job is to sense and level the pre-modulator drive.

Figure 25-57. Extra Millimeter-wave Module Used as a Leveling Hub

- In millimeter-wave multiple source modes, most of the same rules apply. RF leveling is often used here (for frequency converting measurements, for example) but the same issues as discussed above apply since the detection point is in the module (after modulation). A variant, termed VNA-RF leveling, uses a detection point back in the VNA and this can simplify the situation (just as the extra module did for IF leveling). In this case, the power menu and power calibrations would again represent on-state power without duty cycle considerations. Because the detection is before the modulator and the multiplier, the power control will be sensitive at low levels. As with all RF leveling modes, however, power control is limited and will be slightly more so in the pulsed case.
    - VNA-RF leveling is not available above 125 GHz so for higher frequency systems, classical remote detection is usually used although the *auxiliary module* option is available. This gets a bit more complicated in that this situation normally arises in frequency converting measurements so the

system LO is not in the right place to create a leveling signal. The DUT output can sometimes be used for leveling or an additional LO can be generated with an external synthesizer.

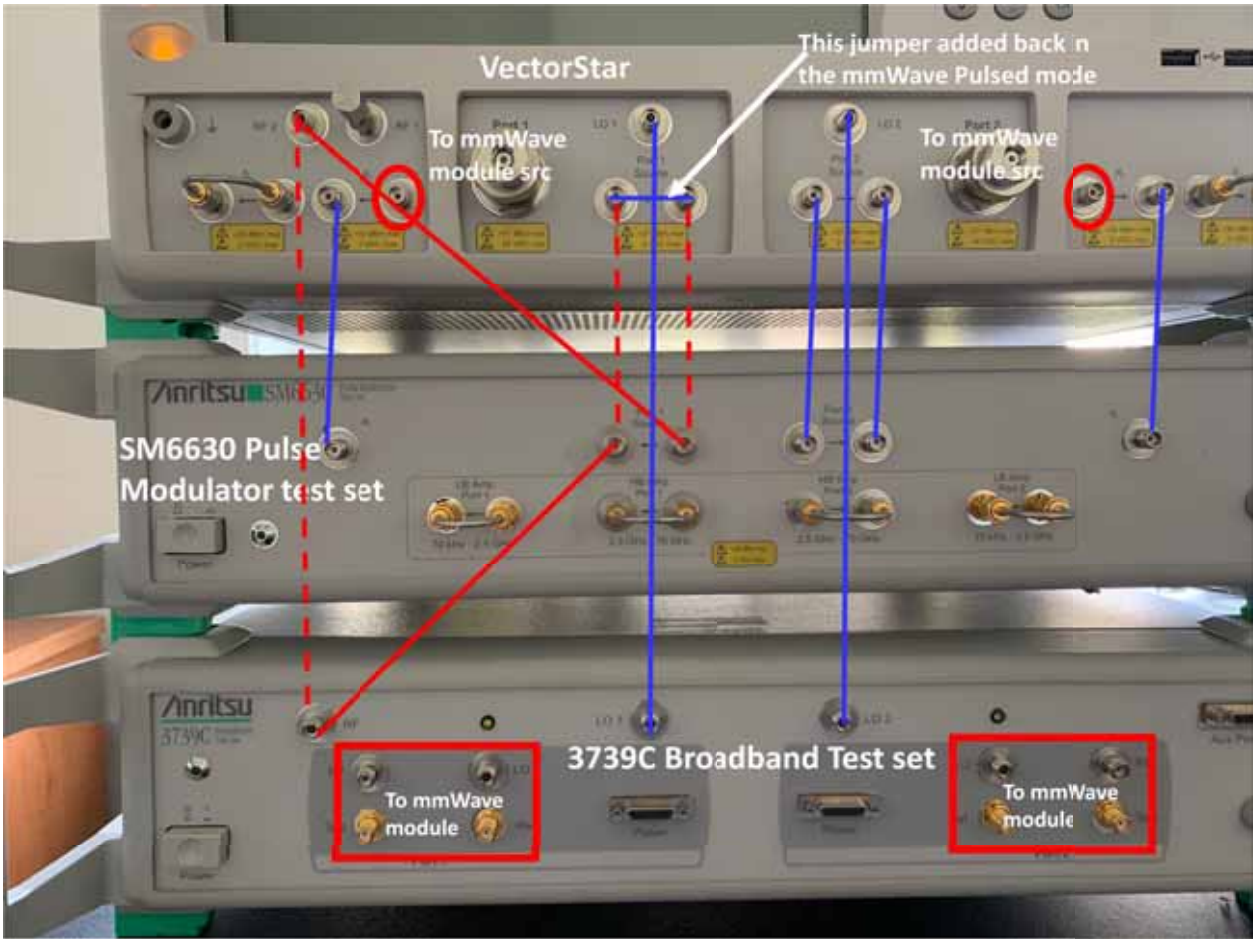


Figure 25-58. A more complete cabling diagram of the concepts in Figure 25-57

- The detection locations are summarized in Figure 25-59. The methods all have their advantages. IF leveling has the largest control range and best stability but is not available (without added modules as discussed earlier) in frequency converting measurements and there are pulsed power definition complications. RF and VNA leveling are available for frequency converting measurements but have less control range and stability (detection linearity and noise). VNA leveling is more easily defined than RF in a pulsed system and has better detection linearity than RF detection based on the final multiplier, but it is more sensitive since it is early in the chain.
- Software-assisted leveling (see the discussion earlier in this chapter) is available where post-processed reference measurements are used to servo the drive power but are less useful than at microwave frequencies since the transient time constants are different and the starting power accuracies are not as confined.

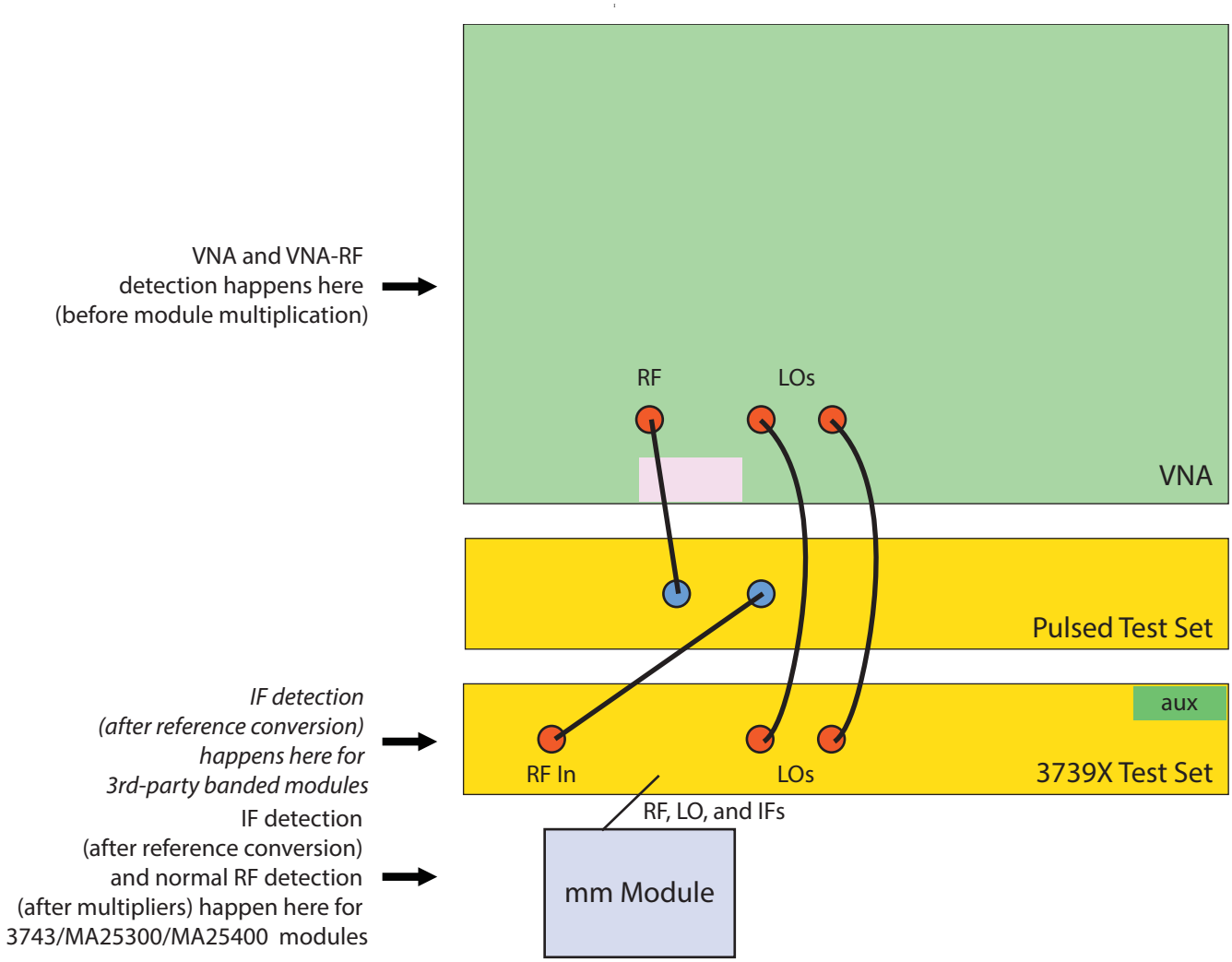
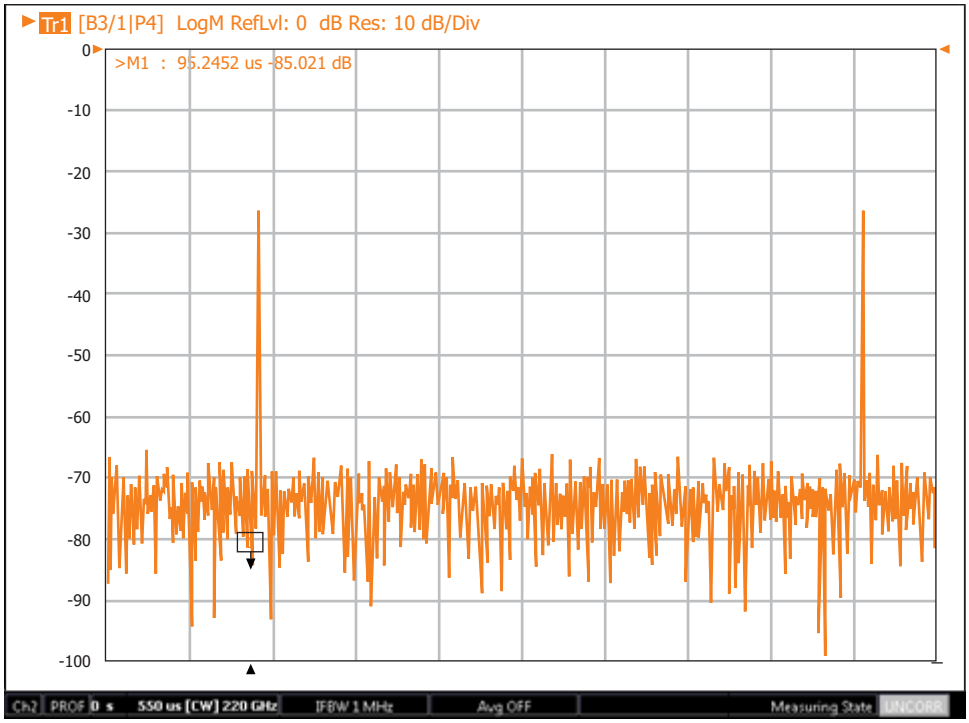


Figure 25-59. Detection Locations

- On four port systems, modulation can be added on paths to both test sets if desired (or to individual modules). The same leveling comments discussed earlier apply here as well. An example ME7838G4 measurement at 220 GHz is shown in Figure 25-60 where a profile over two periods was being evaluated For this setup, VNA leveling was used, the CW frequency was set for 220 GHz, and the period was set for 400 µs. An internal pulse generator fed the external pulse test set connected in the RF drive path as discussed earlier. The pulse width was set for 5 µs and the delay for 100 µs. The profiling range was set for 550 µs with 551 points and a measurement width of 1µs. This resulted in adjoining measurement windows but this, of course, is not required.
- A receiver calibration only was used for this measurement (on the test 3 receiver) as output power of the DUT was of primary interest. The receiver cal was performed while there was no RF pulsing. Average

power was monitored with a power meter (with RF pulsing on and before the DUT was connected) on the on-state power checked manually as the G-band power meter being used did not have a duty cycle entry (as might be seen on lower frequency meters).



The period was relatively long in this example.

Figure 25-60. Example 220 GHz Profiling Measurement with RF Drive Modulation

- For measurements on banded millimeter-wave setups using 3rd-party millimeter-wave modules, many of the same considerations discussed so far will apply. RF drive modulation would normally be done before the test set.
  - The default leveling mode is based on detection in the VNA so the pulse drive would not change things immediately. IF leveling modes are available which would have the same implications as those discussed for the broadband setup.
  - There are calibrated VNA-RF leveling modes (detection in the VNA) also as discussed for the broadband systems.
  - Anritsu has not validated the pulsed video performance (e.g., are the IFs wide enough, do the multipliers induce video distortion, etc.) on all 3rd-party modules, so we cannot predict behavior in general, but the majority of the modules tried did not show anomalies.

Chapter 26 — Dual Source and DifferentialView™ (Option 31 and Option 43)

26-1 Chapter Overview

This chapter examines the architecture changes inherent in Option 31 and Option 43, how existing measurements are affected, and what new measurements are possible.

- Option 31—Dual Source Architecture provides a second signal source.
- Option 43—DifferentialView™ provides True Mode Stimulus (TMS) analysis (requires Option 31).

Configuration choices, measurement design, and measurement behaviors are also discussed for this new measurement category.

26-2 Dual Source Architecture

Option 31 introduces a second source into the MS464xB structure which has applications for mixer measurements, true mode stimulus measurements for differential/common-mode circuits, IMD and other two-tone measurements, and faster multiport measurements, etc. This particular option only introduces the second source which drives port 2 of a 2-port instrument. This section explores the basic structural and performance changes of the instrument, how common measurements might be configured, and the implications of more complex setups.

The block diagram illustrated in Figure 26-1 shows what one would expect: parallel signal generation structures feeding two independent ports. Like the rest of the MS4640 family, there are separate paths for above and below 2.5 GHz. Relative to the non-Option 31 system, a transfer switch has been removed so that the two sources independently drive their ports. Everything in the source chain (including high band/low band components and multipliers) is replicated.

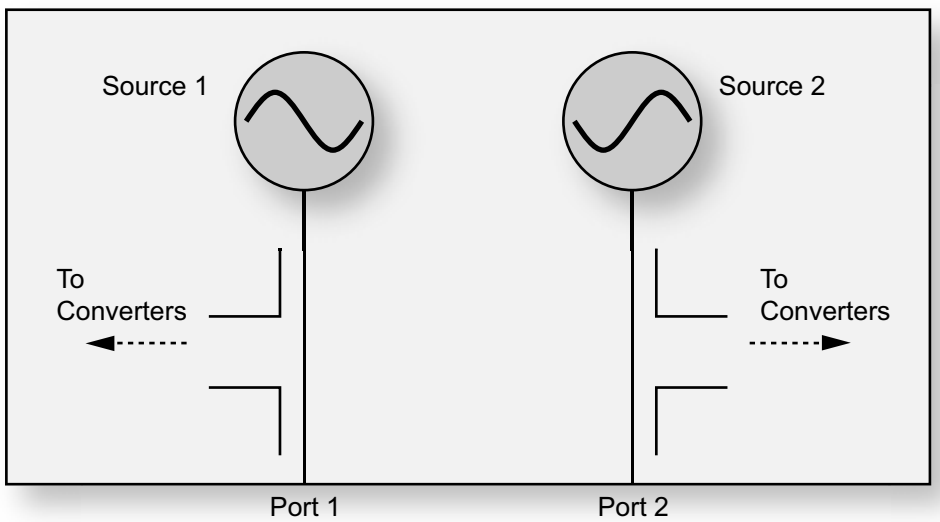


Figure 26-1. Dual Source Option 31 Block Diagram (simplified)

Option 31 removes the transfer switch, which enables higher port powers and more drive flexibility for applications such as mixer and multi-tone measurements. From a standard S-parameter measurement point-of-view, we must introduce the concept of an 'inactive' source. When, for example, Port 1 is driving in an S_{11} measurement, we do not want Port 2 driving to retain the definition of the S-parameter. As there is no longer a switch in position to block that source, it must be powered down as much as is practical and tuned to an innocuous place. The 'safe' place is possible since cross-aligned filters can be used to minimize any signal leakage. Fully powering down the source would not be advisable for reasons of measurement stability and speed. In normal S-parameter modes, this configuration is handled automatically for the user so nothing separately must be done; the S-parameter measurement acts as if the option was not present. When in multiple source mode, however, it may be desirable to take manual control of the process. The use of multiple source control (Chapter 18 of this measurement guide) also requires the concept of 'active' and 'inactive' sources.

- **Active:** the source is on frequency and is driving at the requested power
- **Inactive:** the source is parked at a safe frequency and cross-filtered and the power is modulated down to minimum levels

When in standard S-parameter mode and Port 1 is driving, then the state is {source 1 active, source 2 inactive}; when Port 2 is driving, the state is {source 1 inactive, source 2 active}.

The multiple source setup screens change somewhat since the second source is completely independent. The new setup screen is shown in Figure 26-2. The internal second source entries function exactly as those for the first internal source as discussed in the multiple source Chapter 18 of this guide. The new function is that of internal source control (see submenu in Figure 26-3) which is a parallel of the external source control submenu that exists without Option 31.

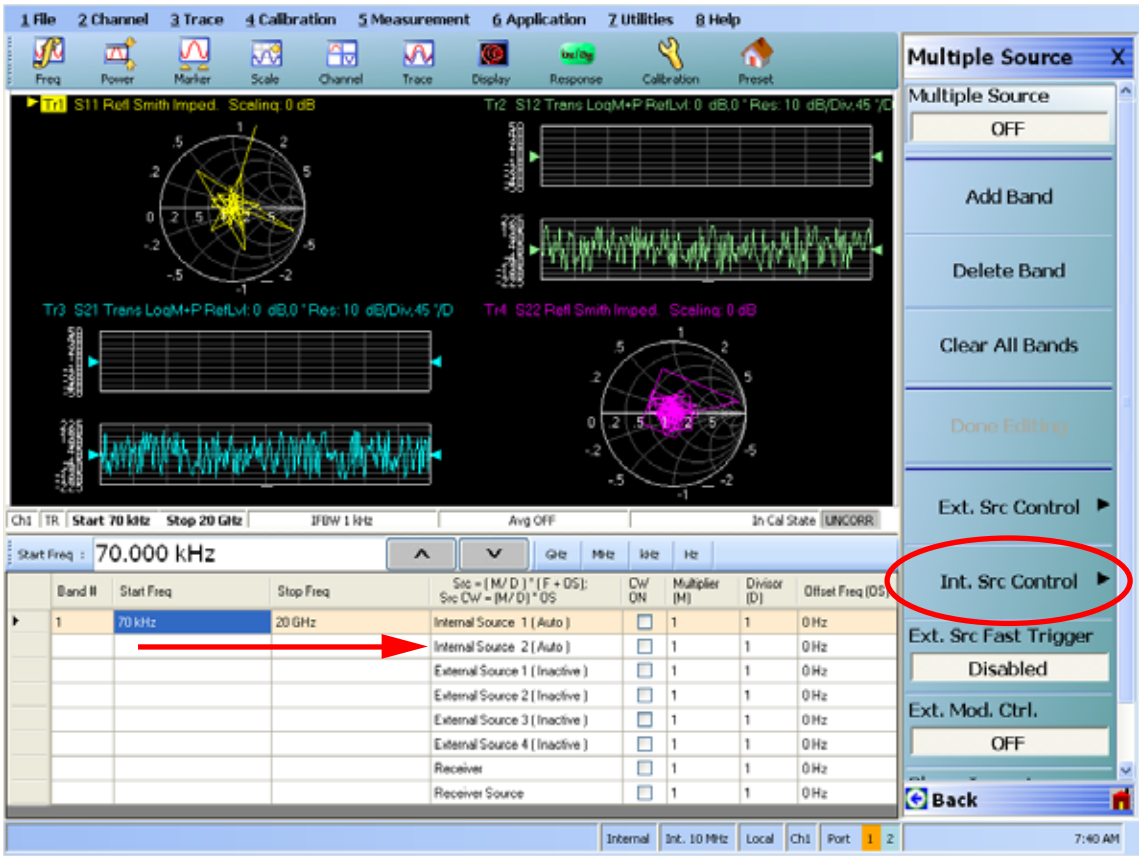


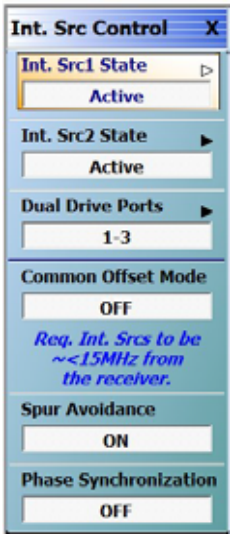
Figure 26-2. Multiple Source Setup Screen (instrument with Option 31)

The second internal source equation is an obvious and necessary addition. The Int. Src. Control button is added for this option and leads to the menu of Figure 26-3.

The internal source control submenu is shown here that allows one to independently activate or deactivate the two internal sources.



Dual 'Auto' Mode



Double 'Active" Mode

Figure 26-3. Internal Source Control Menu

The 'Auto' state is how the sources behave in Standard receiver mode: the source is on-frequency and driving at requested power if that port needs to be driving for the given measurement. If, for example, all traces and applied calibrations are based on Port 1 driving, then Source 2 will always be inactive. If the user calibrations or the trace selections cover both ports driving, then sequentially the system will be in the following two states:

- Source 1 active, Source 2 inactive
- Source 1 inactive, Source 2 active

The double 'Active' state is useful for mixer and IMD measurements that will be discussed, while the double 'Inactive' state is useful when only external synthesizers are being used for a measurement. All selections are independent for the two sources so there are nine allowed permutations.

The nine possible internal source state configurations within multiple source control (with Option 31) are shown in Table 26-1.

Table 26-1. Internal Source State Configurations

Source 1 state	Source 2 state	Example use
Auto	Auto	Standard S-parameter mode
Auto	Inactive	Generally Port 1 stimulus, but want the ability to shut it down
Auto	Active	Generally both ports driving, but want the ability to shut down Port 1
Inactive	Auto	Generally Port 2 stimulus, but want the ability to shut it down
Inactive	Inactive	External sources only used
Inactive	Active	Always using Port 2 stimulus
Active	Auto	Generally both ports driving, but want the ability to shut down Port 2
Active	Inactive	Always using Port 1 stimulus
Active	Active	Mixer (using 2nd source for LO), IMD When a 4-port test set is connected and active, setting both sources active drives ports 1 and 3.

Additionally, when both sources are available and at the same frequency, relative phase control between them makes sense and a phase entry field applies on the Frequency menu as shown in Figure 26-4. This control is only available if both sources are active and at the same frequency. A factory calibration of phase applies in this case and is referenced to the test ports. When one needs the phase control at a more remote reference plane, it is important that equal length (in the electrical sense) paths be used on the two ports.

While phase control could be offered if both sources are active but the frequencies are not the same, the interpretation of the data becomes much more difficult as does a calibration, so phase control is not offered in this case. This is a precursor to the true mode stimulus functionality of Option 43 that is discussed in a later section. The phase control allows the direct measurement of phase-sensitive DUTs such as IQ modulators, certain power amplifiers, and other balanced active structures.

In Figure 26-4, the phase control entry field is shown for use in the cases when both sources are driving. The title of the entry box will show 'Inactive' if both sources are not turned on and at the same frequency.

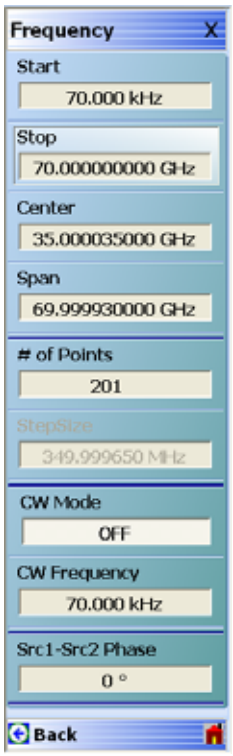


Figure 26-4. Frequency Menu

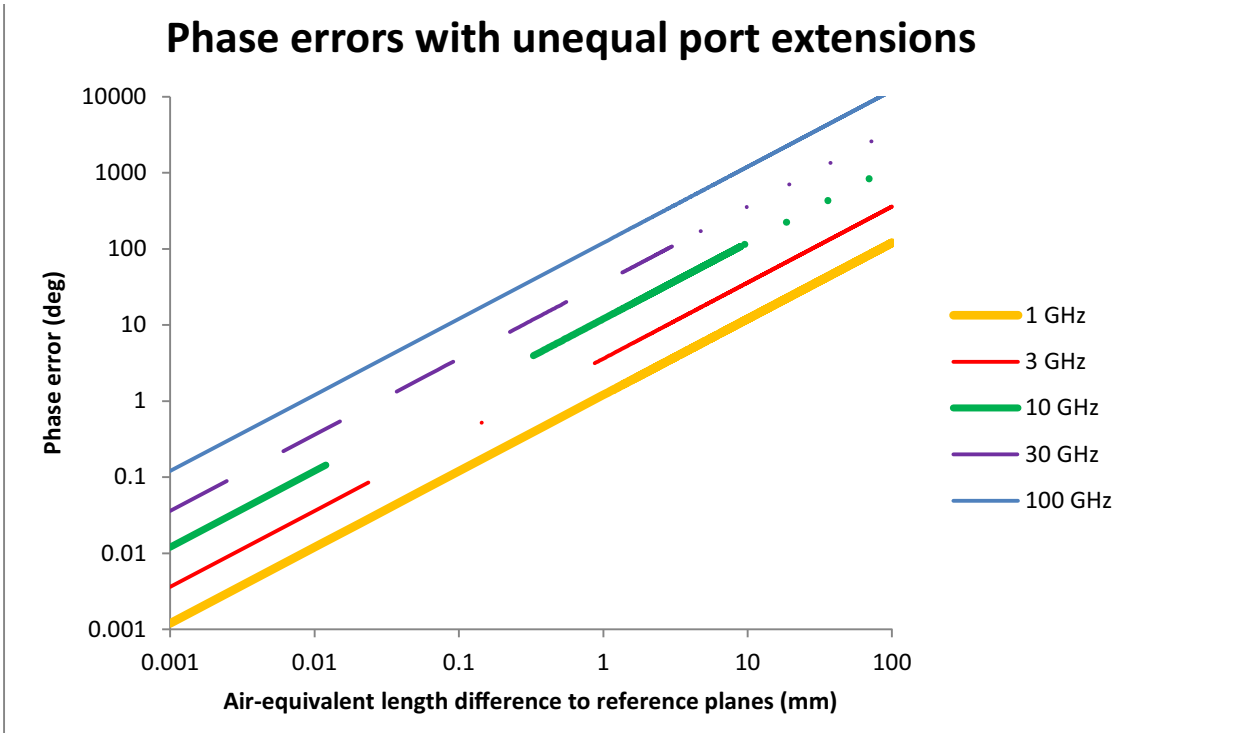
The factory phase calibration is an important concept, particularly with regard to its reference plane placement—at the test ports. If equal length cables are used (to within needed tolerance), then reasonably phase-accurate stimulus can be applied to sensitive DUTs and for cases such as IQ modulators. To measure the phase relationship, very linear and sensitive receivers are already available within the instrument, but we must remove the effects of path and match asymmetry in order for the phase calibration to be precise. Fortunately, this concept is already implicit in the user RF calibrations performed for ordinary measurements in the VNA (the directivity, tracking, and match error coefficients contain just the information needed). If both ports are terminated, then the actual delivered signal ratio (between the two ports) can be quickly calculated from the following (where the error terms were described in Chapter 1 of this guide):

$$\frac{Delivered_1}{Delivered_2} = \frac{\left[1 + \frac{ep1S \cdot (b_1 / a_1 - ed1)}{et11}\right] et21}{\frac{a_2}{a_1} \cdot (et22 - ep2S \cdot ed2) + \frac{b_2}{a_1} ep2S}$$

Equation 26-1.

This equation then forms the basis of a factory calibration algorithm to get the desired delivered phase difference between the ports. With terminations connected, the indicated a_1, a_2, b_1, and b_2 parameters are measured and Equation 26-1. is used to calculate what the delivered phase to the reference plane actually was. The phase registers can then be adjusted to provide the correct phase at each frequency. This set of phase register settings forms the calibration.

Variations of this equation are used for different port pair combinations. An obvious question is what level of length matching is required for a given phase error and a given frequency. This relationship is shown in Figure 26-5 in terms of air-equivalent line length difference. For other media, multiply the physical length difference by the square root of the effective permittivity (for media such as coax cables, microstrip or coplanar waveguide in fixtures, etc.). For rectangular waveguide, the air-equivalent difference can be computed using the propagation constant [1]. As would be expected, the required care in length matching is a direct function of the frequency ranges involved for a given tolerable phase error.



The phase error for different line length differences when extending from the test ports and using the factory phase calibration.

Figure 26-5. Phase Error for Different Line Length Differences

On the subject of the factory phase calibration and its accuracy, two additional toggle buttons from Figure 26-3 require discussion. The factory phase calibration is performed over a certain frequency list and interpolation is used to get to the frequencies in use as discussed. The operating mode of the hardware during calibration and measurement will have some bearing on the accuracy of that interpolation, since the interpolation process requires some knowledge of how the sub-synthesizers in the system are programmed regarding their phase transitions and other properties. Two aspects of the factory phase calibration are 'Common Offset Mode=On' (which describes how some of the internal synthesizers are referenced) and 'Spur Avoidance=Off' (which means little effort is put forth in realigning internal synthesizers to avoid interaction spurs with the DUT.

Spur avoidance is normally turned off in multiple source mode since many of the spurs generated by a DUT, such as a mixer or multiplier (which are the typical multiple source mode DUTs), overwhelm the internal synthesizers and cannot be corrected. By maintaining these settings (Common Offset Mode=On and Spur Avoidance=Off), the best possible interpolation success of the factory calibration is ensured. There may still be small errors in places due to synthesizer frequency hops. If the other modes are needed (such as Spur Avoidance=On for a near-noise-floor measurement with a non-spur producing DUT or Common Offset Mode=Off for a frequency converting DUT), then interpolation will still be attempted, but it may be less successful near some source transitions (such as the particularly dense < 5 GHz system frequency). A key aspect of Common offset mode is that it can only be used when the base source frequencies and LO frequency are within ~320 MHz of each other when referenced to the 10 to 20 GHz frequency range (~160 MHz in the 5 to 10 GHz range, ~80 MHz below 5 GHz, ~640 MHz in the 20 to 40 GHz range, and so on).

To illustrate basic Option 31 control via multiple source mode, consider an upconverter measurement where internal Source 1 will be used as the IF stimulus (ranging from 0.01 GHz to 1 GHz) and internal Source 2 will be used as the LO (fixed at 3 GHz). The RF output of the DUT will be fed directly to a b_2 loop input. In this case, the RF will be ranging over 3.01 to 4 GHz. The setup is shown in Figure 26-6 and the multiple source screen is shown in Figure 26-7. Both sources are put in the Active state since we want them both to always be driving (with careful selection of the response parameters, one of the sources could have been left in the Auto state). A receiver calibration (Chapter 8 of this measurement guide has more information) was used for calibration and the conversion loss of the passive upconverter is also shown in Figure 26-7. More details on mixer measurement calibration approaches are discussed in Chapter 21 of this measurement guide. For the desired drive levels, the port powers were uncoupled so that Source 2 could be set at the required +7 dBm for this measurement and Source 1 could be set to the desired 0 dBm.

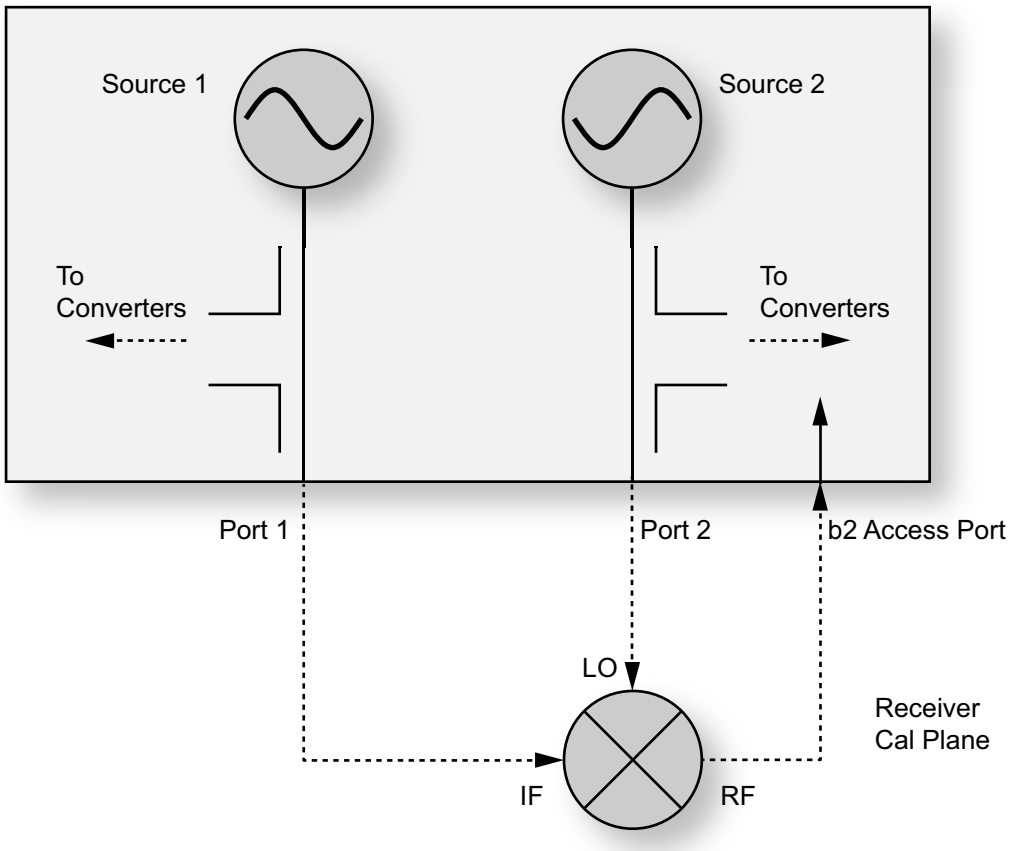


Figure 26-6. Mixer Measurement Setup Example

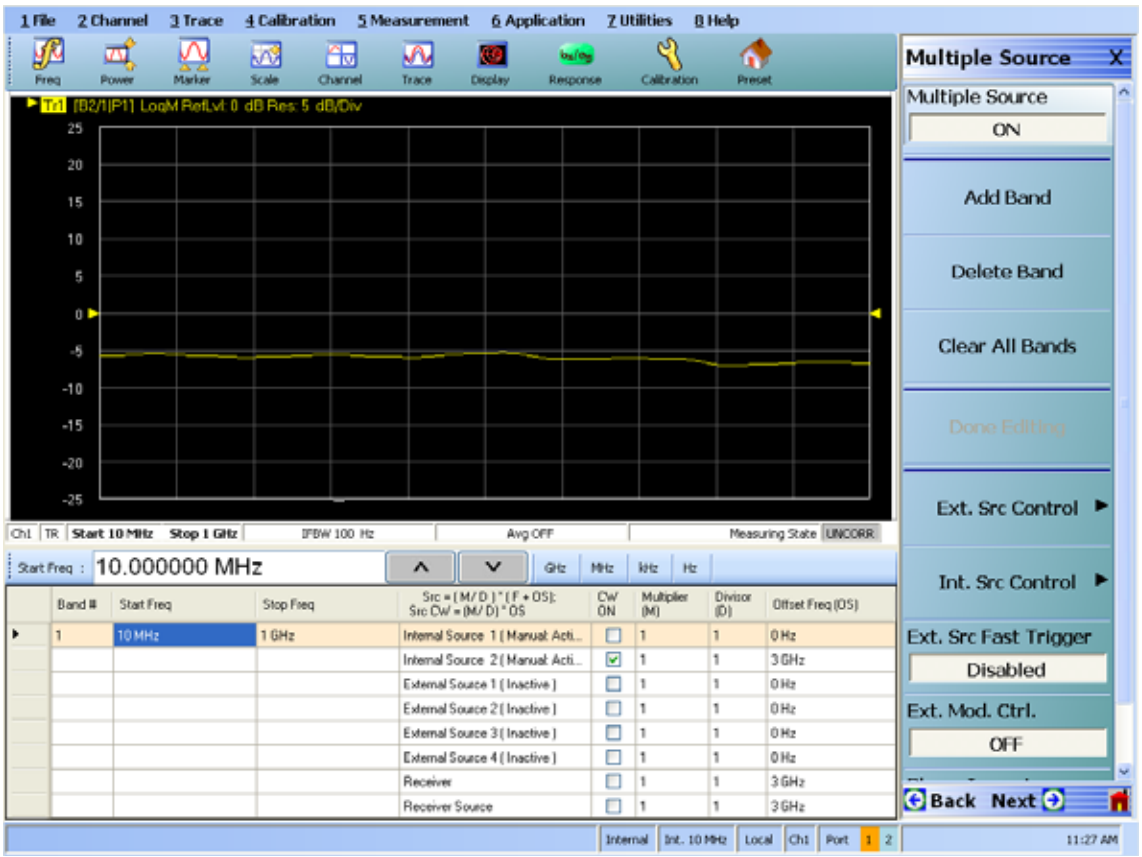
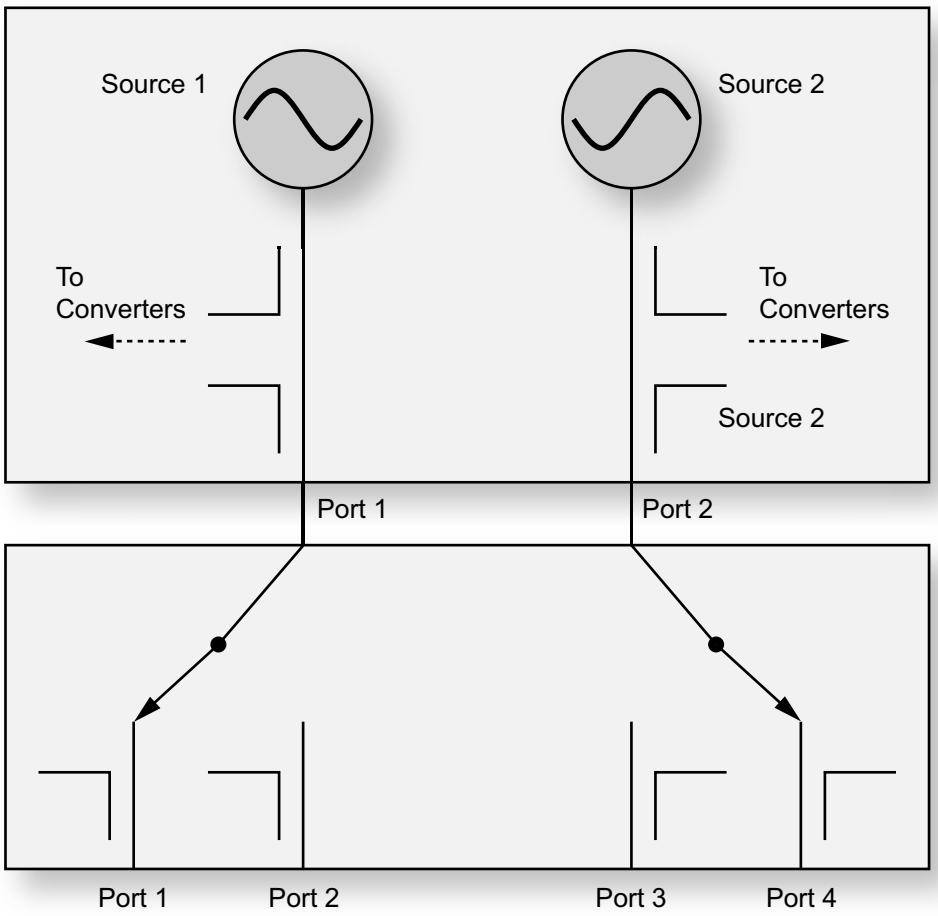


Figure 26-7. Multiple Source Screen for the Example Mixer Measurement

Additional points to consider when using Option 31:

The broadband and millimeter wave options (Chapter 17 of this measurement guide) continue to be available with the dual source modes and most of the operational aspects will be available. In 2-port systems based on the 3739x test set, however, simultaneous source drive in the millimeter wave bands is not possible because of how the RF is routed. In 4-port systems based on this test set, however, simultaneous (both sources Active) drive is allowed. The relevant options for configuring for use with the 3739x test set are Option 84 (when the base VNA does not have loops) or Option 85 (when it does have loops). A second RF tap is provided with Option 84/85 (relative to Option 80/81 for MS4647B, and relative to Option 82/83 for MS4642B, MS4644B, and MS4645B); and this is used for dual drive in the 4-port cases. When used in a 2-port configuration, assembly is done as it is for a non-Option 31 VNA with the RF1 spigot connected to the 3739x test set. When used with a 4-port system, RF1 on the VNA is connected to the controlled test set and RF2 is connected to the controlling test set. LO, IF, and Control connections are unchanged.

When connected to the 4-port test set (MN469xx), dual source drive is also possible. If sources are in the Auto state, the 4-port measurements will proceed as they do in single source instruments. If both sources are Active, then two of the four ports will be driving. Because of the partial switching fabric, not all permutations of driving ports are possible (1-3, 1-4, 2-3, and 2-4 are possible; 1-2 and 3-4 are not). This concept is illustrated in Figure 26-8.



One dual drive configuration when using the 4-port test set is shown here (with Ports 1 and 4 potentially driving at the same time).

Figure 26-8. One Dual Drive Configuration when Using the 4-Port Test Set

The port power controls are unchanged when adding Option 31. Without Option 31, two calibrations were stored, one for each port, since the RF paths were different. With Option 31, two calibrations are again used but now they pertain to entirely different source paths. The power settings are coupled by default (i.e., power levels change together), but this can be turned off. User power calibrations (discussed in Chapter 8 of this guide) also apply to the two source paths independently.

Note	When both Source 1 and Source 2 are set active in the Multiple Source Control menu (Int. Src. Control) and the 4-port test set is active, a driven port pair selection will be available (by default, ports 1 and 3 will drive).

26-3 True Mode Stimulus

DifferentialView™ is True Mode Stimulus (TMS) and requires the Option 31 dual source architecture. As has been discussed in the literature (e.g., [2]-[3]), there are certain measurements of certain differential devices where it is critical that the device be driven with true differential and/or common-mode signals. That is, the superposition method discussed in Chapter 24 is not adequate. There is a nonlinear phenomenon that can occur when the early stages of the DUT are compressing in an unbalanced fashion. If the DUT is not compressing or if the apparent front-end compression occurs in a balance fashion, then true mode stimulus and single-ended stimulus produce the same results [2]. The purpose of this option and this section is to cover the other case.

Option 43, working with the dual sources of Option 31, provides a phase-calibrated stimulus for this situation. The option integrates drive changes with the user RF calibrations to ensure an accurate phase relationship is delivered to the DUT. The main change relative to the usual configuration of Option 31 is that the dual sources are phase synchronized and the phase registers programmed according to a calibration process. This concept is illustrated in Figure 26-9. Hardware clocking and synchronization ensure repeatable phase programming and measurements on the measurement channels will enable real-time correction of the phase angles as needed.

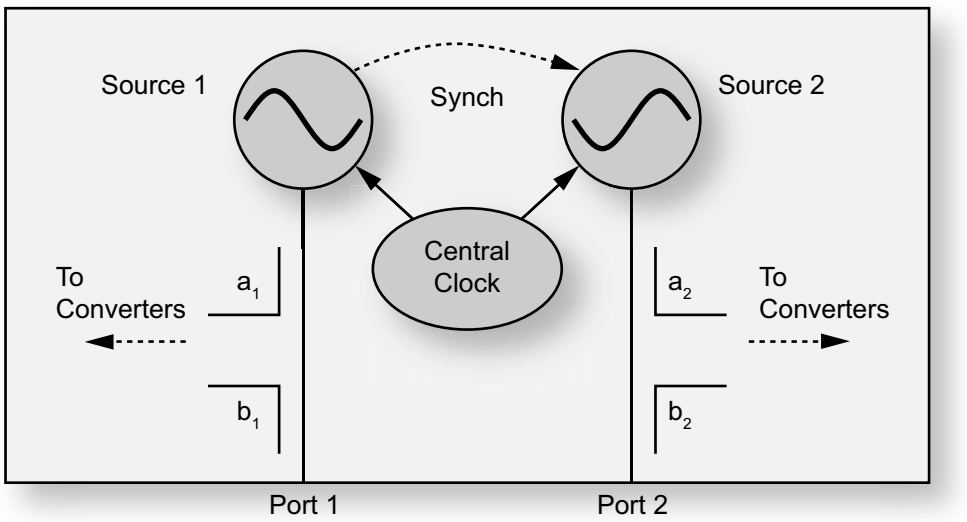


Figure 26-9. Phase Control and Synchronization Concept of Option 43

The factory phase calibration discussed in the Option 031 section may be adequate for some applications. In the class of unbalanced front-end compression relevant for this section, more is needed. Since the reflections of the DUT may be very unbalanced (particularly when it is experiencing unbalanced compression), the net phase relationship delivered to the DUT may be different from the incident phase relationship. Such a delivered phase discrepancy can significantly change the response of the device if it is sufficiently nonlinear [2]. A solution to this is to adjust the incident phase relationship so that the delivered phase relationship is at the desired value based on the S-parameters of the DUT. This is the core of Option 43 along with phase sweep features.

A natural approach is to build this into the user RF calibration routines since the S-parameters and needed error coefficients are already available. This is something of an extension of the factory calibration discussed earlier except now the reflections are arbitrary. The phase relationship would be set up based on the factory calibration to get the incident arrangement correct, and then the delivered relationship can be computed from the below (where a_1, a_2, b_1, and b_2 are measured when the DUT is driven with the desired common-mode or differential signal).

$$\frac{Delivered_1}{Delivered_2} = \frac{\left[1 + \frac{ep1S \cdot (b_1 / a_1 - ed1)}{et11}\right] et21}{\frac{a_2}{a_1} \cdot (et22 - ep2S \cdot ed2) + \frac{b_2}{a_1} ep2S}$$

Equation 26-2.

The phase and amplitude drives are then adjusted as necessary to get the correct delivered signal relationship. Obvious variations of Equation 26-2. are used for different port pairs in a 4-port system.

The application of the calibration then proceeds normally. The differential and common-mode measurements are made (subject to any user requested phase and amplitude offsets), the single-ended parameters calculated from those measurements, and the calibration error coefficients applied. All calibration algorithms are supported. It is important to note that the calibration is done in a single-ended sense (since the error terms are the same) so the same calibration is used for both single-ended and TMS measurements. If one is in TMS mode when a calibration is started, the system will automatically switch to single-ended mode for the calibration and then revert to TMS at the conclusion of the calibration. This duality of the calibration, applying to both single-ended and TMS modes, makes it very convenient to switch between them. Note that interpolation is very problematic for TMS due to the phase correction requiring very good error coefficients, so interpolation is not allowed when in TMS mode.

The main menu of the true mode stimulus application for a 2-port system is shown in Figure 26-10. The main purpose of the menu is to give, at a glance, some of the most important TMS parameters (such a port configuration) and allow offset changes to be entered. The port reference is to help establish the sign convention. The phase and power offsets are applied to both the common-mode and the differential mode states when in TMS drive. When in single-ended but double-active drive (discussed earlier), they both apply to the static state at every frequency. If the phase sweep was active (available for CW only), the start and stop limits would be enterable here.



Figure 26-10. Two Port True Mode Setup Menu

The configuration panel for True Mode in the 2-port case is shown in Figure 26-11. The Single Ended state is the default and is exactly the standard VNA mode discussed before where multiple source controls can override. The Source (Ref Plane) selection turns on both sources (same frequency) and allows one to perform measurements without the phase correcting link to the user RF calibration (to be discussed), thus saving sweep time. The existing factory phase calibration is used and the phase and power offsets can be used for adjustment or manual correction. The phase sweep is available when in CW and can be useful for identifying null points and to balance the state of a variety of devices, and can be very useful as a troubleshooting tool (hence why the reduced sweep time in Source (Ref Plane) mode is helpful).

As an example of this, consider the measurement of the differential return loss of a balun-based device input stage; it may be interesting to see how centered the return loss null is relative to 180 degrees and how abrupt the change is with phase deviations from that value. Such a measurement result is shown in Figure 26-12. A full port RF calibration is generally required.

There is not much to the port assignment in the 2-port case except for providing the sign reference. The assignment can be changed by clicking on the port buttons.

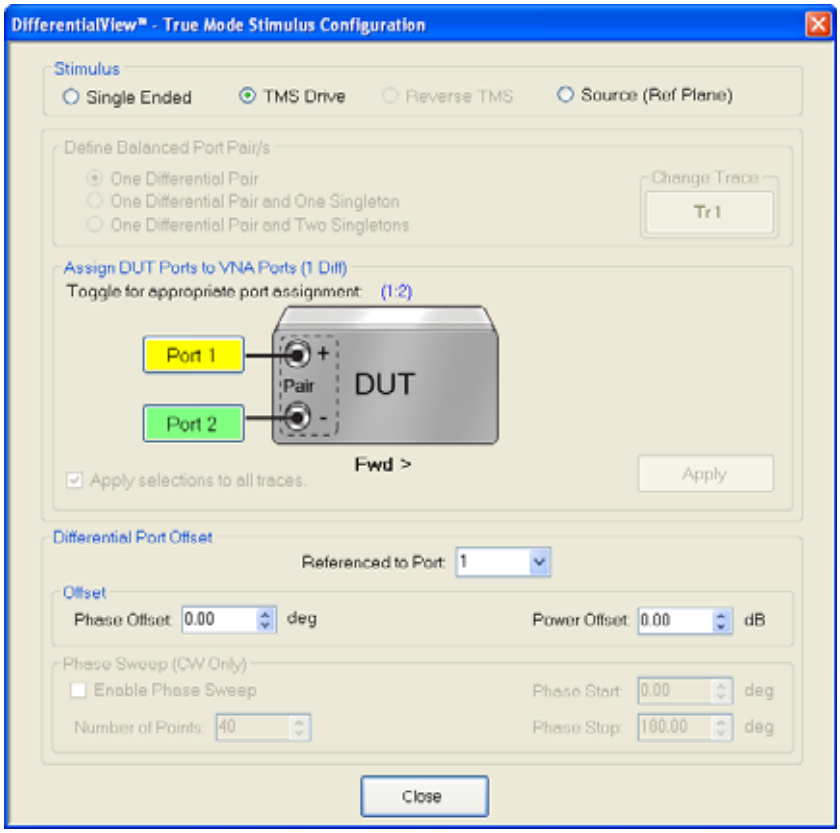
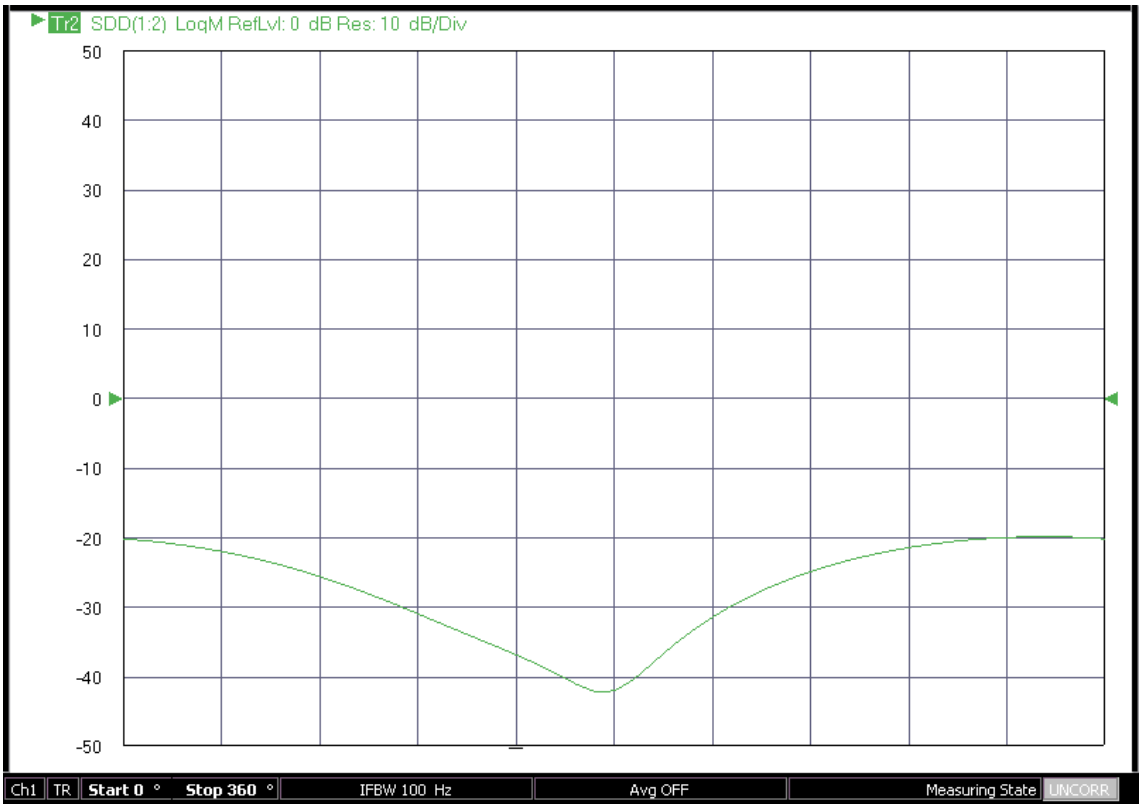


Figure 26-11. TMS Configuration Dialog (2-Port Case)

Note	A full 2-port calibration must be applied for TMS drive and Source modes.

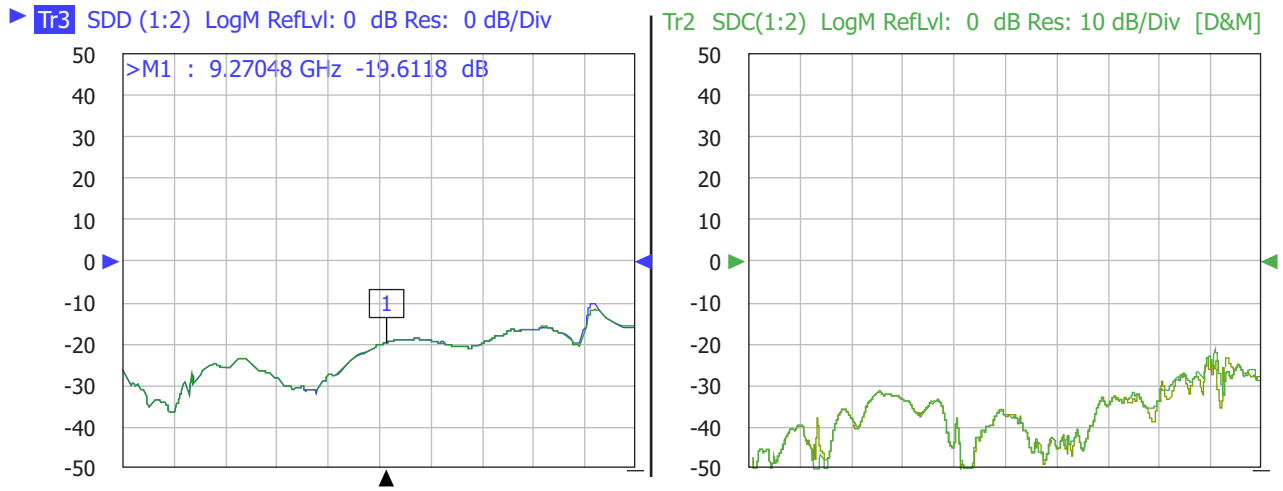


An example phase sweep measurement is shown here of the differential return loss (SDD) of a balun-based front-end. The match null in this case is roughly centered at 180 degrees as would be expected and the response is relatively broad.

Figure 26-12. Example Phase Sweep Measurement

In terms of measurements, single-ended responses can be plotted since they are the base result of the calibration process, but it is more common in TMS mode to use the mixed mode parameters SCC, SDD, SCD, and SDC for the 2-port case (see Chapter 24). As discussed in the pre-amble to this section, the single ended and TMS versions of these parameters will be the same for linear devices and some nonlinearly operating ones.

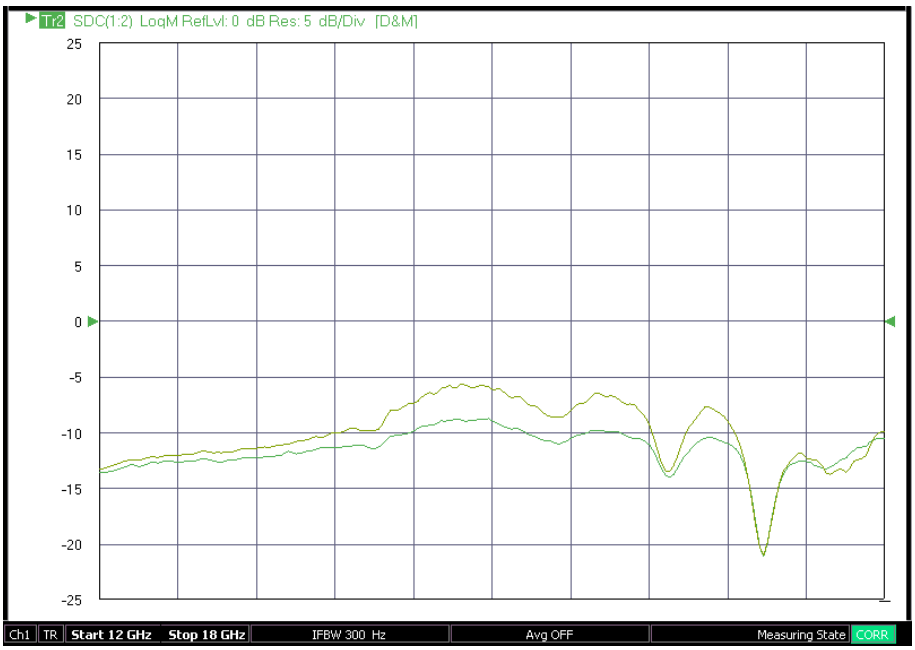
As an example, the measurement of the input differential match of a balanced transimpedance amplifier is shown in Figure 26-13 for both single ended (green trace) and TMS (blue trace) drive.



A measurement of differential input match ($|S_{DD}|$ at left) and mode-converting match ($|S_{DC}|$ at right) of a transimpedance amplifier is shown here in both single-ended and TMS modes. There is no appreciable difference in this case.

Figure 26-13. Measurement of Differential Input Match and Mode-converting Match

But another device shows a substantial difference with the two drive modes, partially because of a radically different common-mode behavior. In this case, the difference particularly shows up in reflected mode conversion and is displayed in Figure 26-14. The absolute mode conversion is much higher than when a similar device is in a linear state and single-ended drive overstates the value by several dB.



The measurement of mode converting reflection of a more unbalanced nonlinear device is shown here with single-ended data in the lighter, upper trace and TMS data in the darker, lower trace.

Figure 26-14. Measurement of Mode Converting Reflection of a more Unbalanced Nonlinear Device

The measurement steps to achieve the above result might include:

1. Select the frequency and power range of interest (12–18 GHz, 201 points, 0 dBm drive in this case).
2. Perform a full-port calibration on ports 1 and 2. SOLT using a 3652A calibration kit was the selection in this case.
3. Select the SDC response parameter for the trace and first observe the measurement result on the DUT while still in single-ended mode. Save the data to trace memory and enable Data and Memory display.
4. Switch to TMS mode and allow for the requisite number of sweeps to complete (a '+' sign on the status bar flashes to indicate each sequence is complete). Observe the differences from the single-ended results in memory.

As suggested by Figure 26-11, there is also the possibility to add amplitude and phase offsets to the stimulus. Thus in the common-mode measurement step, instead of applying tones of equal amplitude and phase, the tones will be slightly shifted. This capability can be used to account for asymmetries in the test fixture (if one had to calibrate outside the test fixture) when one wants to maintain true mode drive at a different reference plane (the die itself, for example). The correction cannot be complete, however, since match information between the calibrated and desired reference plane is not known, so some caution is advised. During the differential measurement step, the same offsets are applied.

When moving to 4-port measurements, many of the concepts are the same, but there are considerably more configuration choices. The true mode configuration dialog for this case is shown in Figure 26-15. Of particular note are the port selection section and the drive mode choice. Three port cases are covered as a subset here, where one pair can be treated as a TMS pair that can be configured on the right or left side of the diagram for convenience (using the reverse drive choice). In the 4-port case, the drive can be tested as all single-ended, two pairs, or one pair and two single-ended. As would be expected, the TMS corrections are only going to be applied on the pairs. The port configuration is somewhat arbitrary and choices can be made to match the desired sign convention. The one exception to flexibility is that 1 and 2 cannot be a pair and 3 and 4 cannot be a pair due to internal switch configuration limitations.

A full port RF calibration corresponding to the diagram in Figure 26-15 must be applied for TMS and Source (Ref Plane) modes. That is, a full 4-port calibration is required for *Two Differential Pairs* and a full 3-port calibration is required for *One Differential Pair and One Singleton* (and that 3-port cal must cover the appropriate ports). The Reverse TMS state is for the 3-port case and inverts the diagram so the differential pair is on the right. Operationally, regular TMS Drive and Reverse TMS are the same for the 3-port case.

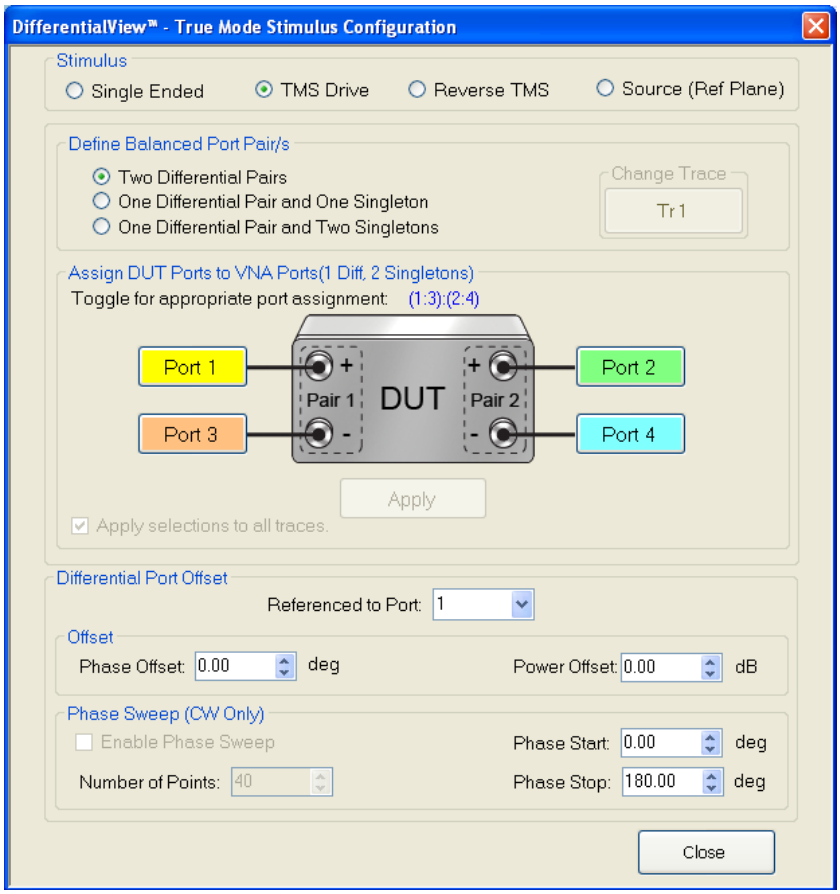


Figure 26-15. TMS Configuration Dialog (4-Port Case)

Much like in the 2-port case, Source (Ref Plane) allows one to drive dual ports with phase and amplitude offsets and with CW phase sweeps without the coupled phase correction steps. This speeds sweep time and can help with troubleshooting and analysis.

On display selection, the mixed-mode parameters will most commonly be used and the parameter selection process is the same as discussed in Chapter 24 of this guide. This dialog is repeated in Figure 26-16 here for convenience and the configuration from the TMS configuration will map by default to the response selection dialog. Although it is an unusual situation, it is allowed to use different port pairings for the mixed mode parameter definition compared to the drive configuration of TMS. The most common time this would be used would be to shift polarity within a pair to match a particular display requirement for final data. In graph legends and in the menu display, the port pair configuration is labeled as (x:y):(z:w) for pairs x-y and z-w. For the three port case, the singleton port will be outside of the parenthesis (e.g., (x:y):z). Note that the order within the pair designation denotes polarity.

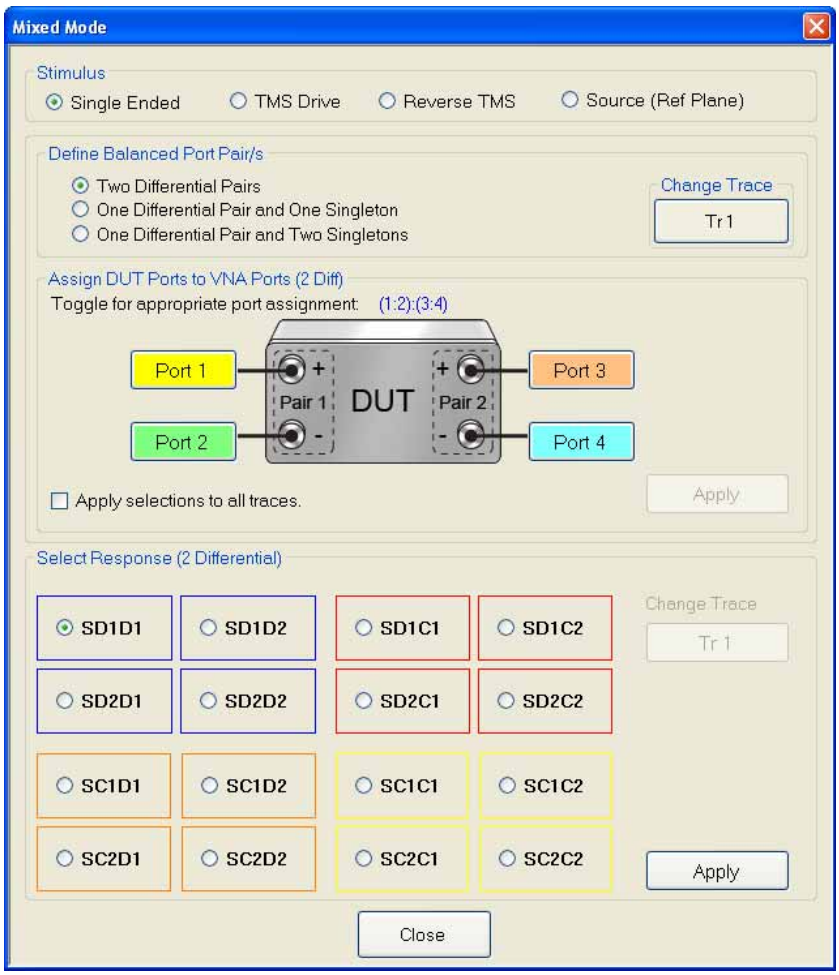
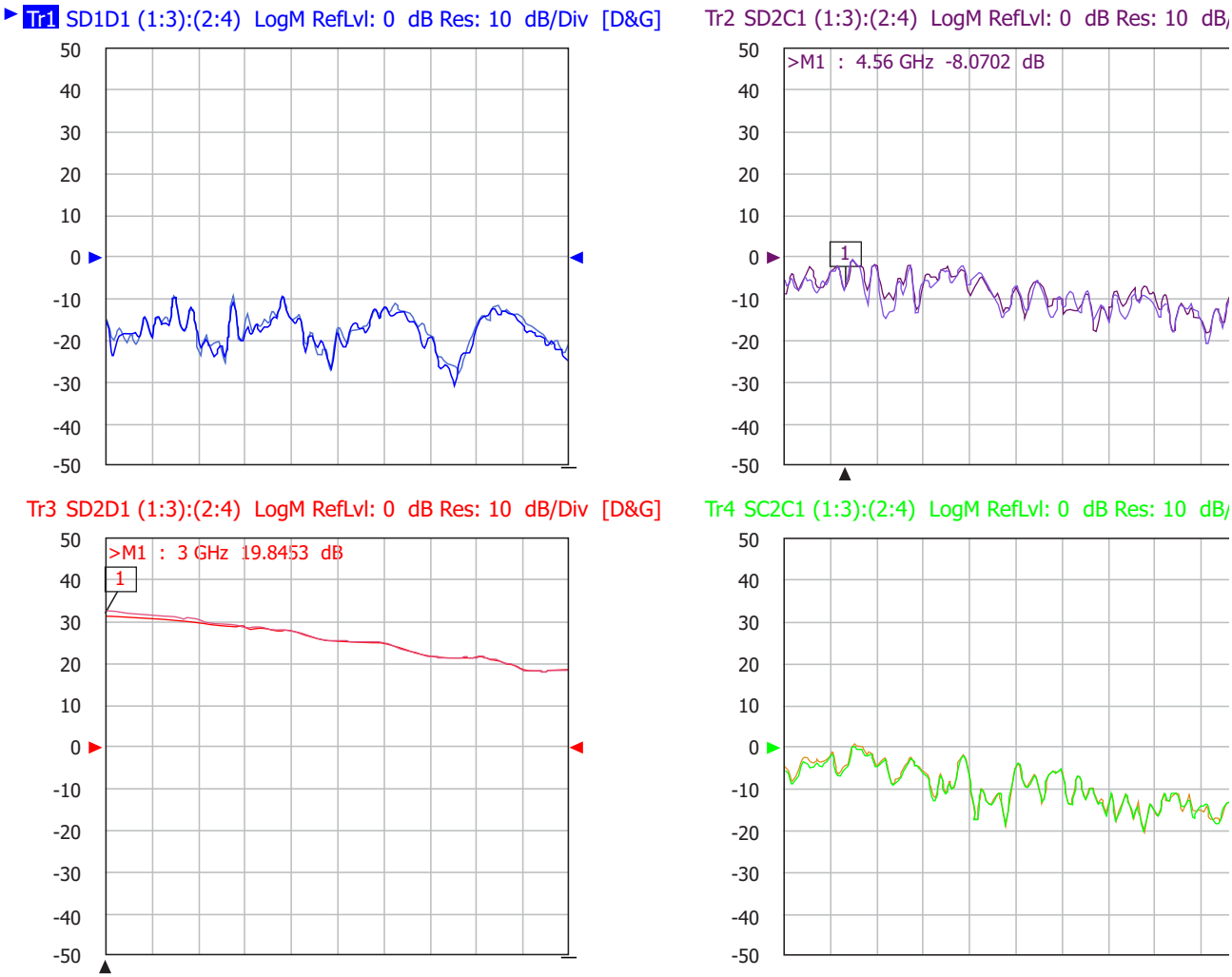


Figure 26-16. Mixed-mode Parameter Dialog

As a first example, consider a fairly linear amplifier measurement. Setups and measurements were done in both single-ended and balanced-balanced true-mode stimulus configurations. Several mixed-mode parameters of the amplifier are shown in Figure 26-17. This particular amplifier had unbalanced compression mechanisms due to its topology, but was being driven at a low enough level that drive sensitivity was weak.

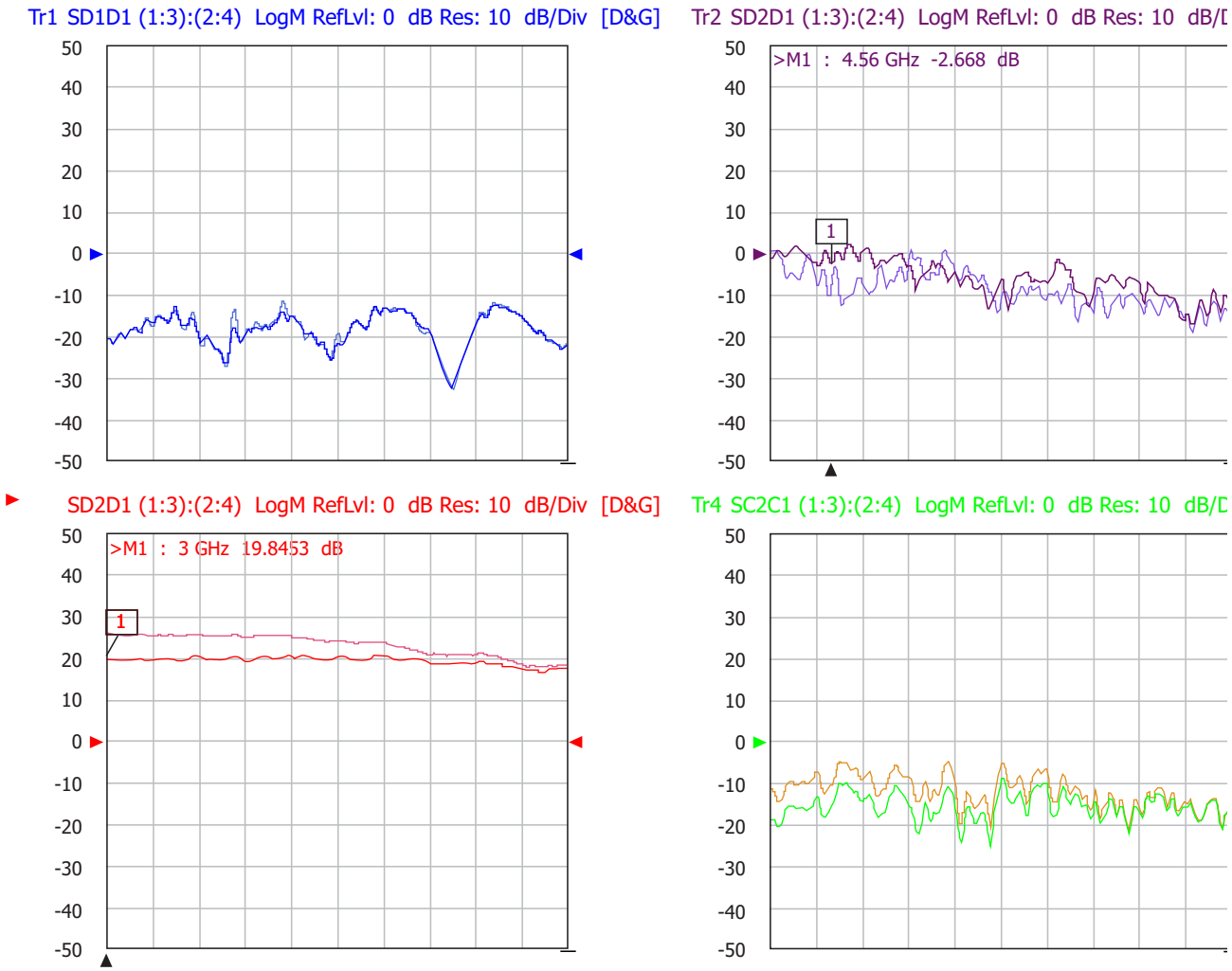


Some mixed-mode parameters of an amplifier in both single-ended and true-mode stimulus modes are shown here at a low drive level. In this case, there was no significant difference.

Figure 26-17. Mixed-mode Parameters of an Amplifier

The power level was then increased to the level of several dB of compression. Since this amplifier had a very unbalanced compression response, significant differences were observed between single-ended and TMS drive. These measurements are shown in Figure 26-18. Note that while the transmission parameters (including mode-conversion parameters) show differences between the drive types, the input differential match shows little change. They are not plotted here, but the mode-converting match and common mode match terms were changing rapidly.

The procedure used here was similar to that in the previous 2-port example, except a full 4-port calibration was used. This also happened to involve an SOLT calibration with a 3652A calibration kit. The response parameter selections are, as always, per trace, but the port configuration employed for mixed-mode was the same on all traces.



SD2D1| values of a second amplifier (for single-ended and true-mode stimulus drive) are shown here. In this case, there is a divergence at compressive drive levels.

Figure 26-18. |SD2D1| Values of a Second Amplifier

As with the 2-port version of TMS, phase and amplitude offsets are available. Some caution is required on amplitude since the default ALC calibrations are referenced to the VNA's base two ports. Additional user power calibrations may be needed at the final reference planes to have an accurate read on absolute power. The TMS process will equalize power delivered to the DUT (in absence of an entered offset) relative to the starting value for Port 1 or Port 2 (in the 4-port case).

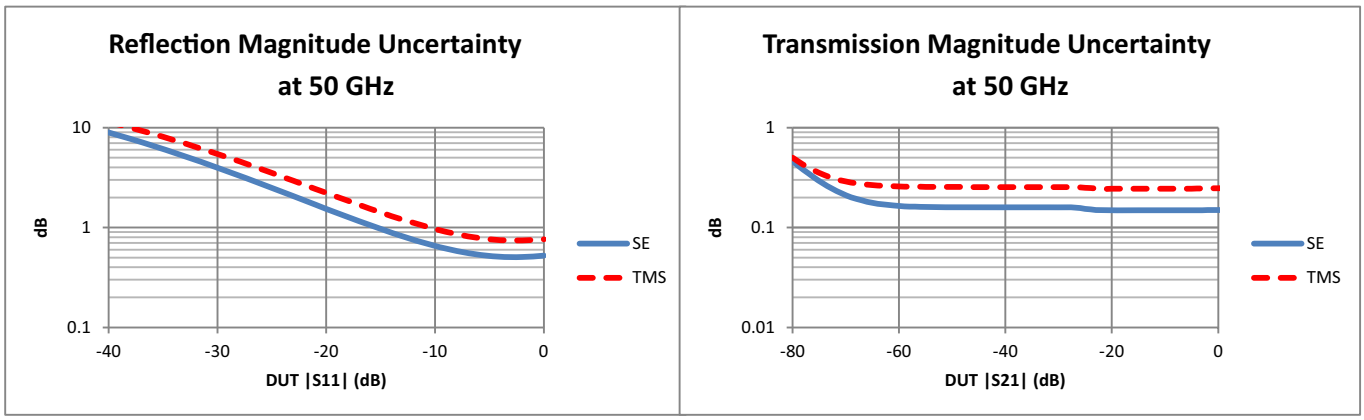
An important question is how do the uncertainties of the S-parameters derived with true-mode stimulus differ from those measured with single-ended stimulus? Since the error coefficients are derived from exactly the same measurements, the calibration kit-related uncertainty terms are the same. The difference lies in how the raw S-parameters from the DUT measurement are fed into the calibration. In the case of single-ended measurements, the input is obviously direct. In the true-mode stimulus case, however, raw single-ended S-parameters for use in the calibration are extracted from differential and common mode drive measurements.

Consider the simple 2-port case, the raw S_{11} is found from:

$$S_{11}^{raw} = \frac{1}{\left.\frac{a_2}{a_1}\right|_c - \left.\frac{a_2}{a_1}\right|_d} \left[\left.\frac{b_1}{a_1}\right|_d \cdot \left.\frac{a_2}{a_1}\right|_c - \left.\frac{b_1}{a_1}\right|_c \cdot \left.\frac{a_2}{a_1}\right|_d \right]$$

Equation 26-3.

In the limiting case of perfectly in-phase (and equal amplitude) a_2 and a_1 for common-mode drive and perfectly out-of-phase for differential drive, this reduces to the average of b_1/a_1 in the two drive states. These measures are, however, subject to their own uncertainties in the stochastic terms and in the deterministic terms (while macro issues have been largely removed by the pre-measurement adjustments, the residuals still apply). Thus there is something of a compounding of uncertainties with true-mode stimulus. For the quasi-linear situations of relevance here (unbalanced compression), this uncertainty increase is small in comparison to the functional correctness, but it does suggest that true-mode stimulus will present something of a detriment to uncertainty, however modest, when linear or balanced compression measurements are called for. An example comparative uncertainty plot is in Figure 26-19.



Example comparisons of uncertainties for single-ended vs. true-mode stimulus in a linear S-parameter measurement are shown here. The calibrations in this case were all sliding load SOLT and the uncertainty evaluation is at 50 GHz. Drift differentials were not included.

Figure 26-19. Example Comparisons of Uncertainties for Single-ended vs. True-mode Stimulus

Not included in Figure 26-19 are any differences in stability due to drift from thermal or temporal changes. Since the correction process for true mode stimulus (using Equation 26-2. for example) is dependent on the error coefficients, any changes in those error coefficients will cause a delivered phase error along with an error in extracting the raw S-parameters. The former can be quite germane for the unbalanced nonlinear cases being discussed in this section and primarily argues for high cable quality, using as controlled an environment as possible, and performing periodic recalibrations. In the sense of the latter, this is another uncertainty adder for measurements longer after calibration time.

26-4 Other Notes on Using Option 43

Simultaneous Drive Availability in 2-Port BB/mmWave System

As mentioned in the Option 31 section, most broadband and mmWave configurations are compatible with dual source systems. The exception is that simultaneous drive, which includes true-mode stimulus, is not available on 2-port 3739x-based broadband and mmWave systems since the ports are linked by a transfer switch on that model. True mode stimulus is available with 4-port 3739x-based broadband and mmWave systems.

TMS Coherence Flag and BB/mmWave Systems

When in broadband/mmWave modular 3739 modes, the effective trace noise at higher frequencies can become important. While not ideal for low-signal-level measurements (which are not common in true mode stimulus applications), there is a frequency plan modification that can improve this trace noise particularly in phase above 54 GHz. This is selected with the TMS Coherence flag (available via System | Setup | Misc. Setup) and should be turned ON for both the 4-port calibration phase and for the TMS measurement. If one later decides to turn this off, the 4-port calibration must be redone.

TMS Availability

The 4-port configuration has been the subject of most of this section. True mode stimulus is available with 2-port standard MS464xB VNAs for use in measuring balanced reflection.

True mode stimulus is available in 4-port broadband and mmWave configurations using Option 84 or Option 85. The required connections were described earlier in this chapter.

26-5 Summary

The dual source option family (Option 31 and Option 43) has been discussed in this chapter. Both are centered on the availability of two sources that enable simpler mixer and IMD measurements, true mode stimulus balanced-differential measurements, and a variety of other possibilities. The hardware parts of the options have been configured for maximum connection flexibility and the very open multiple source control system allows for maximum measurement setup flexibility. The TMS calibration approach is as transparent as possible so switching between dual-drive and single-drive measurements is easier.

26-6 References

1. C. T. A. Johnk, “Engineering electromagnetic fields and waves”, *Wiley*, 1975, chp. 8.

2. J. Dunsmore, K. Anderson, and D. Blackham, “Complete pure-mode balance measurement system,” *2007 IEEE MTT-S Int. Micr. Symp. Dig.*, June 2007, pp. 1485–1488.

3. J. Simon, “True differential stimulus gives additional insight into nonlinear device behavior”, *69th ARFTG Conf. Dig.*, June 2007, pp. 1–9.

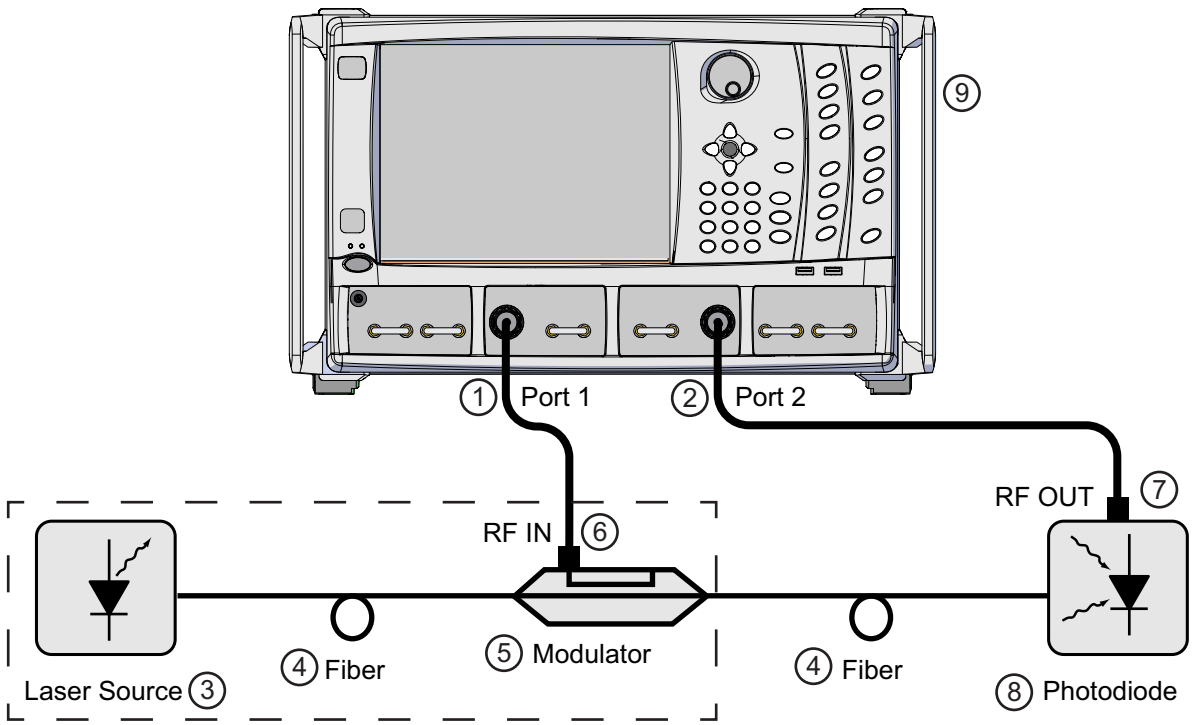
Chapter 27 — E/O and O/E Converter and Optical Measurements

27-1 Chapter Overview

As fiber and free-space optical communication bandwidths increase, the need for very high speed optical modulators and detectors has also increased. The frequency response characterization of these ***electrical-to-optical*** (E/O, modulators sometimes integrated with lasers) and ***optical-to-electrical*** (O/E, detectors and receivers) converters can be important in terms of such parameters as bandwidth, flatness and phase linearity. In addition, the microwave frequency response characteristics of certain purely optical (O/O) components may be of interest as their optical bandwidth may be narrow (couplers, amplifiers, filters, etc.) or may have unusual dispersion characteristics. The MS464XX VNA has a number of measurement utilities to facilitate this kind of analysis and, coupled with the MN4765X O/E calibration module or some other calibration device, some level of measurement traceability is possible. The ME7848A series of systems incorporate the MS464XX VNA, MN4765X O/E module and, optionally, a MN4775X laser/modulator unit. This chapter will discuss some of the measurements of interest, setup considerations, possible measurement performance, and examples of execution procedures.

27-2 Introduction and Background

Conceptually, the job of the optical modulator is to place a microwave signal as modulation onto an optical carrier. Similarly, the job of the photodetector or receiver is to recover that modulation and regenerate the microwave signal. For a VNA-based measurement, both directions of conversion are required so that the processing can occur in the microwave or modulation signal domain. The result is a setup like that shown in Figure 27-1 (a MS464XX VNA is shown in the image, but a broadband system ME7838XX can be used as well). The optical carrier is generated (usually) by a coherent laser source (which may be integrated with the modulator), modulation is applied, and then the modulation is recovered. Optical fiber is shown as the media in Figure 27-1 but it could be some other optical guiding medium or free-space in some cases. The VNA acts as a microwave stimulus (Port 1 in the figure) and receiver (Port 2 in the figure). 3- and 4-port cases are also possible, involving multiple converters or differential ones, which will be discussed later in this chapter. Although a fiber is shown as the only element between the detector and modulator in Figure 27-1, some optical DUT may be there for O/O measurements.

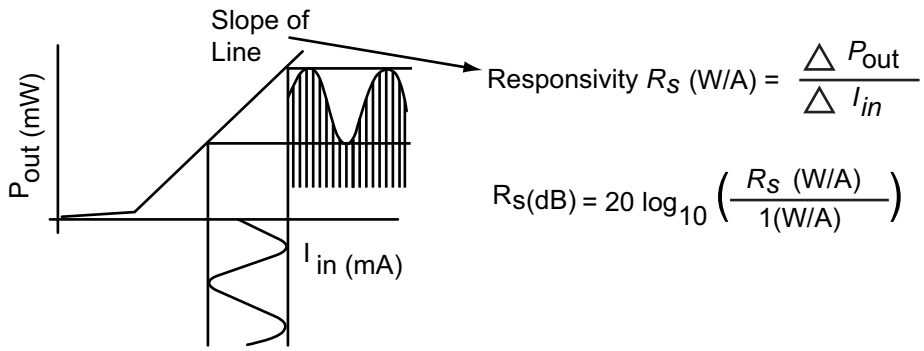


Index	Description	Index	Description
1	VNA Port 1	6	Modulator RF IN
2	VNA Port 2	7	Photodiode RF OUT
3	Laser Source	8	Photodiode
4	Fiber	9	MS464xB VNA
5	Modulator		

In some cases, the laser and modulator may be one assembly (as is the case in the MN4775X converter). The photodiode may be integrated with amplifiers or other components into a photoreceiver.

Figure 27-1. General 2-Port E/O or O/E Measurement Setup

Since the measurements results will be classed as normal S-parameters, one may wonder how these relate to the actual optical behaviors of the components. In some sense, they all become relative because the conversion between domains introduces dependencies on optical laser power, optical path losses (usually small) and other absolute shifts. Thus the real measure of conversion is essentially a responsivity slope between the optical and electrical domains as illustrated in Figure 27-2. The S-parameters that appear on the instrument display for an O/E or an E/O component then represent a relative responsivity measure (in both magnitude and phase). Often, the frequency response of this quantity is of interest as that determines bandwidth and magnitude vs. frequency plot gives this information. The phase linearity and group delay are both ways of looking at the deviation from a purely linear phase function that can be an important assessment of potential phase-related modulation distortion. Return loss of the component may also be of interest but that is a purely microwave measurement.

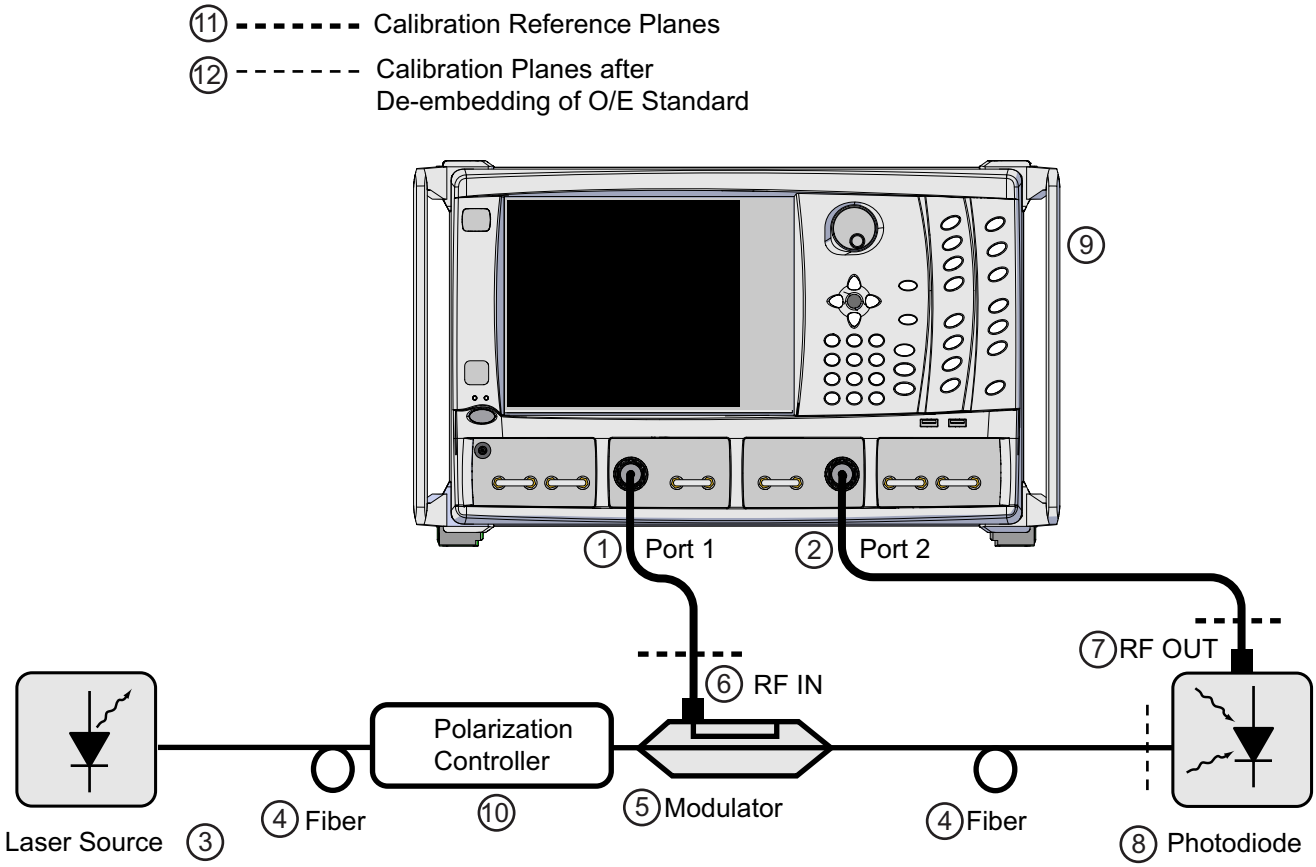


The conversion parameters of the O/E and E/O devices measured with the VNA are essentially measures of responsivity.

Figure 27-2. Responsivity Concept

A next question may be how the measurements are conducted. The starting point is a 2-port VNA calibration as has been discussed in earlier chapters of this measurement guide. This calibration establishes reference planes at the microwave ports of the optical devices as shown in Figure 27-3 on page 27-4 (often coaxial ports but could be waveguide, in a fixture, etc.). The next step is the use of an O/E calibration device such as the MN4765X. That particular model is a wide bandwidth photodetector housed in a thermally controlled module with carefully designed bias circuitry. This module is characterized at a traceable facility using electro-optic sampling techniques (or references derived from that) so its frequency response (in magnitude and phase) is well-known with established uncertainties (e.g., [1] -[2]). If such a calibration device is the detector in Figure 27-1, then its effects can be de-embedded (see the embedding/de-embedding sections of Chapter 11 in this guide for more general information) since those behaviors are known. This then moves the reference plane to the optical side of the photodiode/calibration detector as shown in Figure 27-3. Now a measurement of S21 will describe the loss and phase of the modulator alone (plus some effect of the fiber which will be discussed). This frequency response (magnitude and phase) gives the required performance information discussed earlier when combined with the microwave reflection measurement (S11 in the diagram) of the modulator RF port that comes for ‘free’ with the calibrated VNA measurement.

A polarization controller is shown in Figure 27-3 but, technically, this is optional. Many modulator structures (including Mach-Zehnder modulators such as those used in the MN4775X converter), are polarization sensitive so at least a polarization-maintaining fiber is recommended on the modulator input.

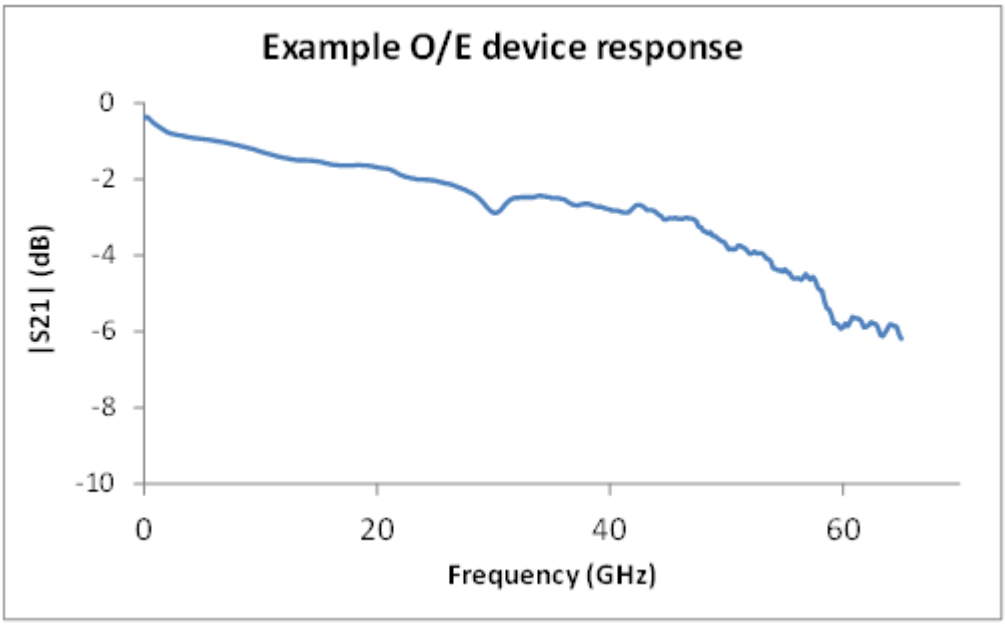


Index	Description	Index	Description
1	VNA Port 1	7	Photodiode RF OUT
2	VNA Port 2	8	Photodiode
3	Laser Source	9	MS464xB VNA
4	Fiber	10	Polarization Controller
5	Modulator	11	Calibration Reference Planes
6	Modulator RF IN	12	Calibration Planes after de-embedding of O/E Standard

A calibration is done in coax, waveguide, or some other electrical media and the de-embedding function, via the O/E-E/O measurement utility, of the VNA are used together with optical data to move a reference plane into the optical domain.

Figure 27-3. Reference Plane Placement

An example plot of the conversion response of a 50 GHz photodetector is shown in Figure 27-4. From the magnitude response, the 3-dB bandwidth is indeed around 50 GHz but the roll-off is sufficiently slow that this device is commonly used beyond 65 GHz. The phase response is also shown in the figure but the linear portion has not been removed. The group delay plot (derivative of phase with respect to frequency) is often a more convenient way of looking at the phase behavior. Towards 50 GHz, there are some deviations from flat group delay (equivalent to deviations from linear phase) that are not surprising in view of the device's bandwidth. If being used as a characterization device, these deviations are not important since they can be well-characterized (to beyond 65 GHz in this case).



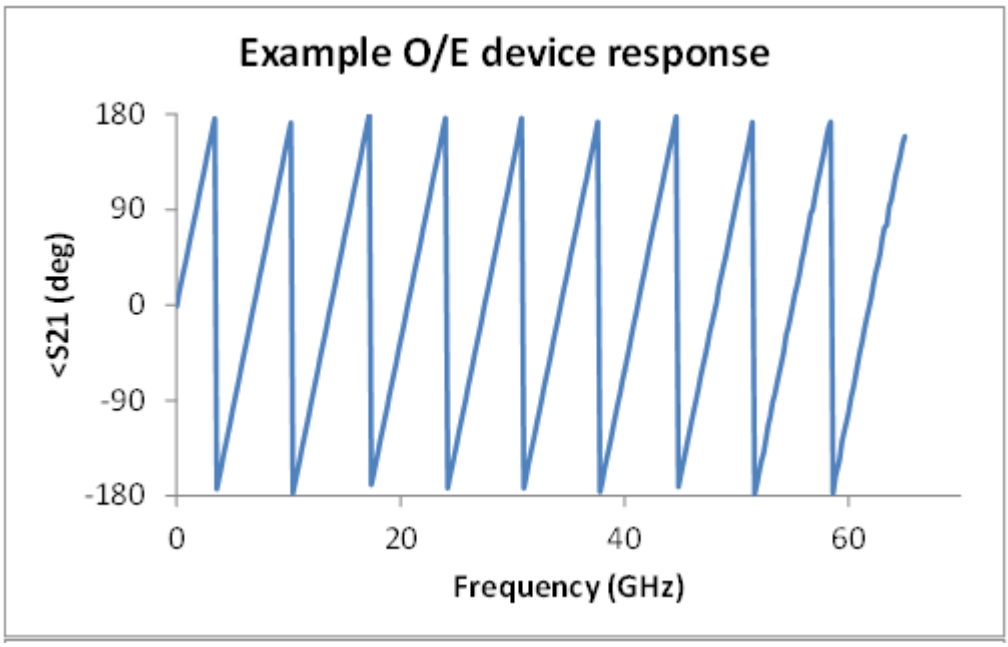


Figure 27-4. Characteristics of Example O/E Device

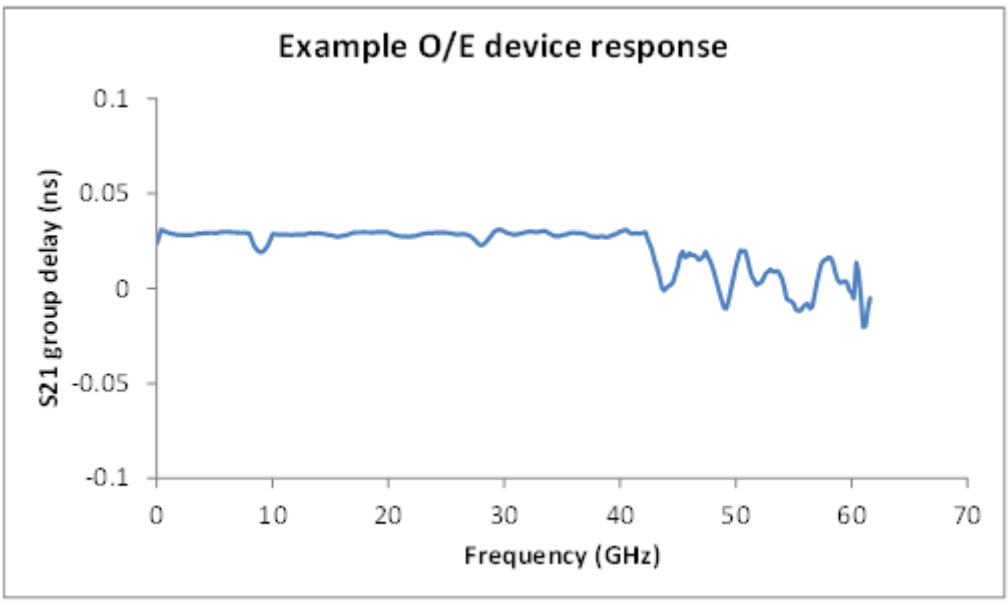


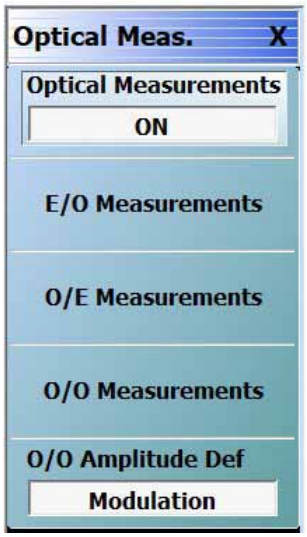
Figure 27-4. (Continued)Characteristics of Example O/E Device

If one now wanted to measure a different O/E device (not a calibration module), one could then insert that detector into the setup of Figure 27-3 on page 27-4 and instead now de-embed the modulator response that was just found. In this case, because it is a second level de-embed, there may be some elevation of uncertainties that will be discussed. One could also obtain an E/O calibration device and use that instead in a one-step process.

The de-embedding (or sequential de-embedding) steps form the basis of this O/E-E/O measurement utility. The key points are controlling traceability and uncertainties throughout the process when multiple devices are being used, to control match so minimal additional artifacts are introduced, and to not try to de-embed what cannot be de-embedded. This last point is important in that inner-plane (optical) match is not known and the transmission path is unilateral anyway, so there are no multiple reflections within the DUT assembly to remove.

Optical Measurements Menu

All of these measurement aids are available under the Perform Optical Measurements button located on the MEASUREMENTS menu (Navigation: MAIN | Measurement | MEASUREMENT | Optical Measurements | OPTICAL MEAS. menu). As might have been guessed from the previous discussion, these approaches naturally separate based on whether the target is an O/E device (e.g., detector or receiver), an E/O device (e.g, modulator), or an O/O device (e.g, coupler, amplifier, or filter, etc). The menu selections, as shown in Figure 27-5 delineate that choice.



On the Optical Measurements menu are the three selections based on which type of device is to be solved for.

Figure 27-5. OPTICAL MEASUREMENTS Menu

A certain class of purely optical (O/O) measurements can also be made where both the detector and the modulator have been de-embedded. One essentially performs the steps for both the E/O and O/E measurements in series to place the reference planes in the optical domain. This is still a microwave/millimeter wave frequency response measurement, so the O/O measurement is really a measure of the response to changing modulation bandwidths and, as such, is most suited to relatively narrowband optical devices such as amplifiers, filters, couplers, resonators, etc. For example, it would be unusual for a patch cord to show variation over a millimeter wave modulation bandwidth (unless it had some particularly unusual distortion properties). Also, mismatch in the optical domain is not correctable with this technique and, if the fiber runs are long, it is possible that ripple will occur on the scale of millimeter wave frequencies. An additional complication with the O/O measurements is the difference between modulation domain power detected by the VNA receivers and optical power. The issue comes about because the O/E detector or calibration module is essentially a square-law device taking optical power changes to root-power (or normalized voltage). The default O/O readings will be in the modulation domain, which may represent double the loss (in dB terms) from what would be seen with an optical power meter. Thus an optical power choice is available for O/O measurements that will scale the results appropriately when thinking in terms of optical carrier power. Note that this correction happens after all calibrations have been applied, but before post-processing. So if one is using trace math normalization to help evaluate, for example, optical device loss, the normalization should be done in the same state (modulation domain or optical power) as the measurement.

All of the measurements (up to this point) in this chapter work by using information about optical converters to modify an RF calibration coefficient set on the VectorStar VNA. This is exactly how conventional de-embedding works but the optical variants of this process take additional steps to handle the unilateral nature of the optical system (one cannot apply RF to the detector and usually expect RF to be produced by the modulator) and some differences in match terms. As with regular de-embedding, the optical modification to the calibration coefficients can be turned on and off. Turning the Optical Measurements to OFF will cause the original RF calibrations coefficients to be loaded. Returning Optical Measurements to ON will reload the modified coefficients. Both sets of coefficients are stored as part of the setup file.

While the bulk of the optoelectronic functionality is intended to measure the RF (modulated) response in converting devices and optical systems, it is also possible to measure effects on average optical power (the 'carrier' in this measurement class). This can be useful to get a quick estimate of laser or OE/EO behavior, optical path losses, or interconnect issues. The simplest method using the VNA is to plot average optical power directly. While akin to what an optical power meter can do, this measurement is based on OE module DC output. The accuracy will not be as high as for a dedicated optical power meter but it does have the ability to reveal effects from the RF system as well be discussed.

The measurement of optical power is orchestrated from the response menu when optoelectronic measurements are enabled as shown in Figure 27-6. The **Analog In** ports of the VNA are used to measure the low frequency response of the OE module and convert this to optical power.



Figure 27-6. The lower part of the RESPONSE menu when in an optoelectronic mode is shown here. Mean optical power can be selected as a response variable (two inputs are available). The RF driving port must be selected and the detector responsivity entered.

Note that are two Mean Optical Power selections available since there are two **Analog In** ports on the VNA. Thus, the optical power from two different points in a setup, assuming detectors are available, can be measured simultaneously along with the S-parameters. External Analog In response parameters can still be selected and these will plot raw voltages at the analog in ports.

Several points may be important:

- A DC path from the OE module to one of the Analog In ports is needed. This is often done using a bias tee (the internal bias tee in VNAs equipped with Opt. 06x or an external one) or possibly a splitter or combiner. A sketch of one such arrangement is shown in Figure 27-7. The bandwidth of this path does not have to be high (kHz to low MHz range depending on sweep speeds).
    - If the second Analog In connector is to be used to process the voltage from a second detector, the internal bias tee may be difficult to use (unless that detector is connected directly to the EO device in the Figure 27-7 example). This second detector output could be connected directly to the other analog in port if the detector is not processing an RF signal or it can be connected via an external bias tee.

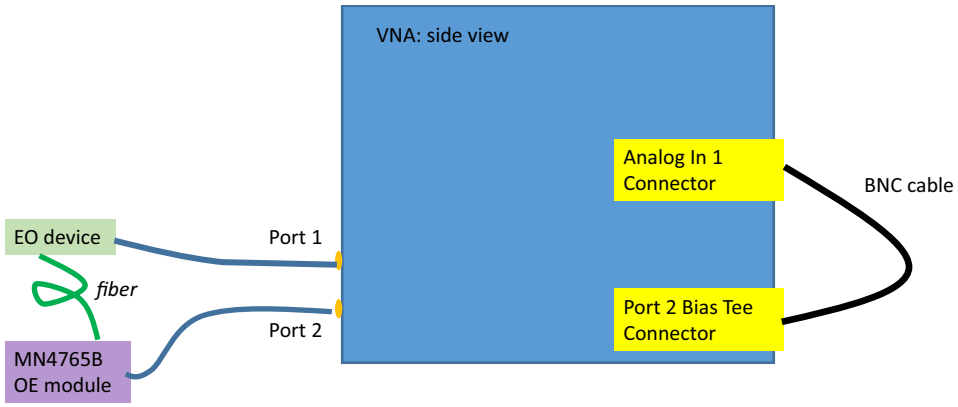


Figure 27-7. One connection arrangement for optical power measurements is shown here. The DC detector output must get to one of the Analog In rear panel connectors and using the internal bias tee and a BNC cable is one approach.

- The responsivity of the detector must be known (in A/W). Typical values for the Anritsu MN4765B are shown below but this can also be computed by using the straight analog in measurement (not the optical power response selection) and a measurement with an optical power meter if greater accuracy is desired. See Figure 27-8 for the sub-menu where responsivity is entered.

OE module	Typical responsivity (A/W)	Comments
MN4765B-004x	0.2 (at 850 nm)	0.6 @ 1060 nm, 0.7 @ 1310 nm, 0.8 @ 1550 nm
MN4765B-0070	0.7	
MN4765B-0071	0.45	
MN4765B-0072	0.65 (at 1550 nm)	0.45 @ 1310 nm
MN4765B-011x	0.55 (at 1550 nm)	0.25 @ 1310 nm

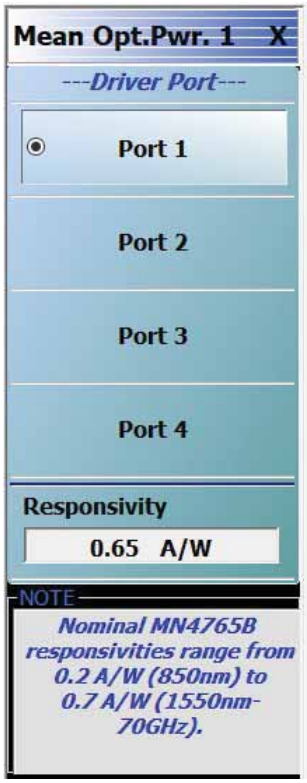


Figure 27-8. MEAN OPT. PWR. submenu where responsivity can be entered.

- Other OE devices can act as the power detection conduit, but the responsivity must also be known, the output DC levels must be in the range of the Analog In port (-5 V to +5 V), and a 50 ohm terminating resistor in the OE device is assumed (since only voltages are measured by the analog in port). Accuracy of the DC measurement is on the order of 0.5 mV (additive) so very low responsivity detectors will produce mean optical power measurements with higher uncertainties.
- As suggested by Figure 27-8, the driver port must be selected. This is the RF driving port that will be associated with the DC measurement and will almost always be the port to which the EO device is connected.
- The signal levels measured are typically in the mV range for common optical power levels so the calibration of the Analog In port can be relevant. While this calibration happens automatically at application launch, the values can drift somewhat over time if the instrument is left running for extended periods. A calibration button is available on the response menu to refresh this state. This calibration re-zeros and recomputes the gain corrections of the low frequency analog in paths.
    - In some extended-length setups, ground voltage imbalances can occur which complicate the mV scale measurements at low optical power. Additional grounding paths (e.g., ground straps or cables) may be needed from the detector plane to the VNA to get accurate low optical power readings.

The optical power measurement is, like any other response variable, plotted versus the current sweep parameter (frequency or power). In many cases, the result will be a very flat line as shown in Figure 27-9. The power out graph type is particularly useful for this measurement as the markers read-out explicitly in dBm.

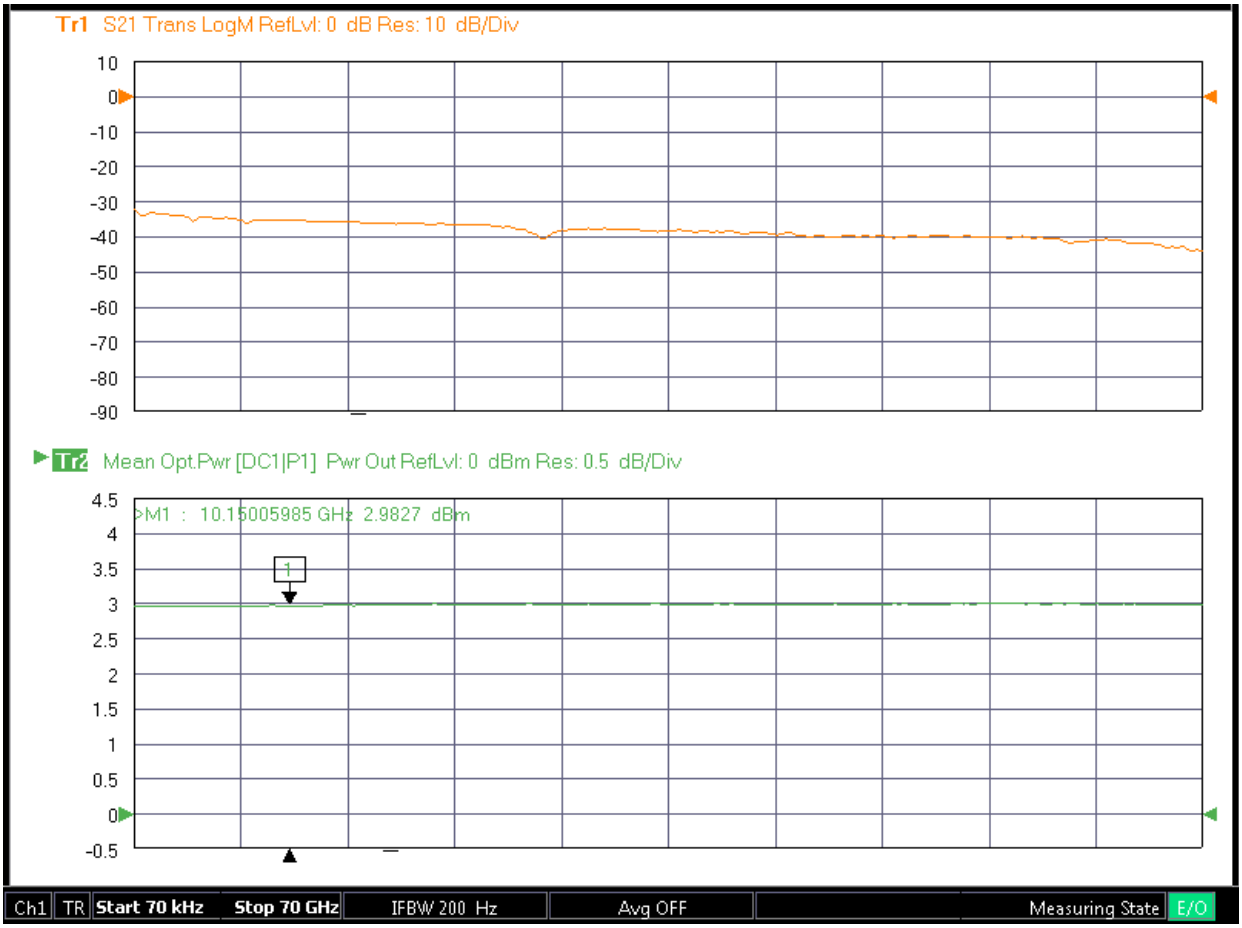


Figure 27-9. An example optical power measurement (trace 2) is shown here as part of an EO measurement setup. The Power Out graph type is particularly convenient.

If some part of the item being measured is interacting nonlinearly with the RF signal, variation over the sweep may be observed. One example is with a Mach-Zehnder EO device where the RF amplitude is on the same order as the V_{π}, or modulation swing range, of that modulator. In this case, the RF can alter the bias state of the device and the mean optical power getting to the detector can vary widely. An example of such a situation is shown in Figure 27-10 where the RF was applied through an amplifier that had more gain at lower frequencies and, for the high drive example, was able to compromise the bias state of the modulator at those frequencies. The net optical power deviation was not large in this example, but multi-dB swings are possible for some configurations.

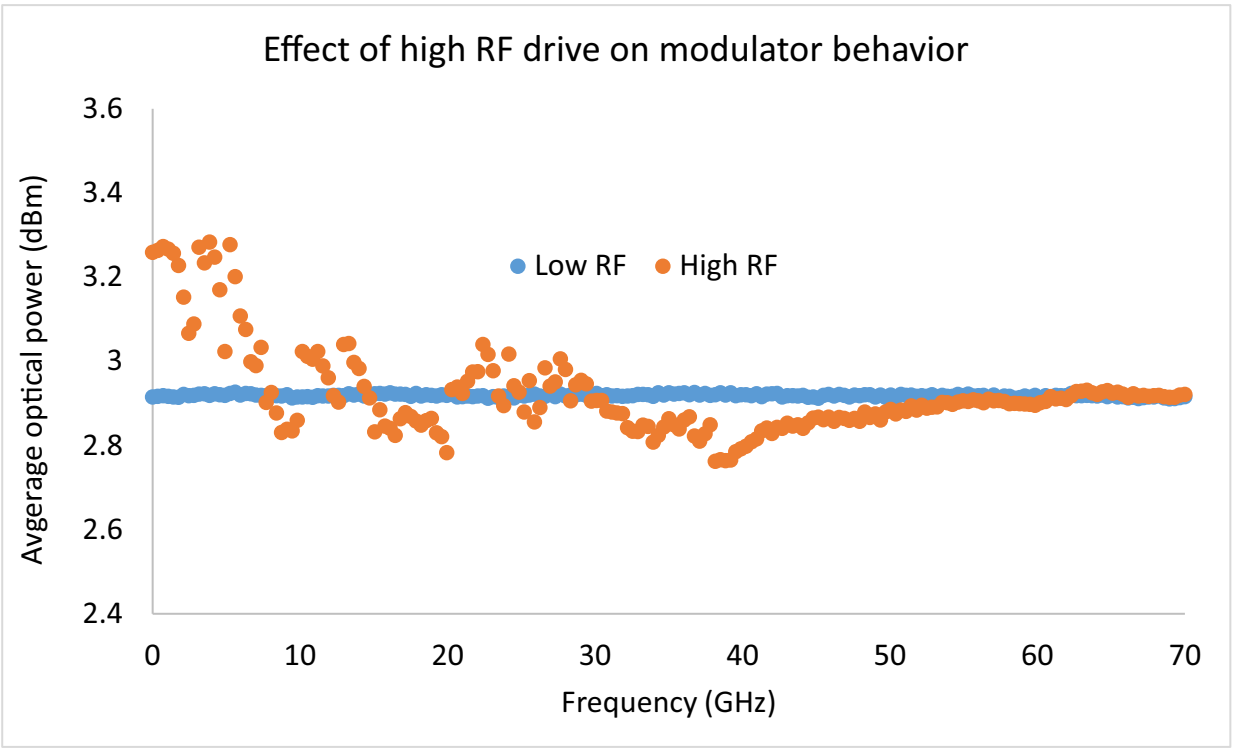


Figure 27-10. An example where the swept nature of the optical power measurement can be useful: observing RF-related nonlinear effects of the DUT structure.

The optical power uncertainty from this measurement is limited by the accuracy in the knowledge of the responsivity and the accuracy of the DC voltage measurement. For an Analog In port with a current calibration, the latter uncertainty is typically 0.5 mV assuming no ground distortions are present. The optical power in mW is computed as

$$P_{o,mW} = \frac{20V_{analog_in}}{Responsivity\ (\frac{A}{W})}$$

So, if the responsivity is very low, the Analog In voltages will be lower (for a given optical power) and the DC measurement accuracy will have more of an effect on the computed optical power.

27-3 Measurement Setups and Considerations

It is not the intent of this chapter to fully cover the optical setup details. However, some common issues and concerns will be discussed. More general information of fiber optic measurement setups can be found elsewhere (e.g., [3]-[4]).

Optical Linearity

The linearity of the characterized photodiode directly affects the accuracy of the measurement and the optical input power at which RF photocurrent will remain linear is important. One way to check that the measurement path is still linear is to normalize the S21 plot against itself and increase the optical power to see if there is a gross sensitivity (the absolute level will move, but the important question is if the shape of the response changes with optical power). For example, set the optical laser power to 4 or 5 dBm with the setup like that of Figure 27-1 on page 27-2. Store the resulting S21 to memory (under the Display/View Trace menu) and then view DataMemMath (data divided by memory). Now increase the laser power in 1 dB steps until some compression (in this case, meaning that the deviation from flatness is visible on this scale) is seen in the normalized plot on the scale of 2 dB/div. At that point, decrease the optical power level until it is out of compression. Make sure that the maximum DC current for the photodiode is not exceeded.

It is often useful to maximize optical power to improve dynamic range as long as the linearity of the DUT or the system optical components is not compromised. Every 1 dB increase in optical power will improve dynamic range by 2 dB due to the power-law detection characteristics of the photodiodes.

Laser Power and Photodetector Bias Sequencing

Always make sure the photodiode is biased properly before turning the laser on. Improper bias or no bias can degrade photodiode performance and can also result in damage. Instructions on handling and biasing of the photodiode are shipped with the characterized photodiode accessory. Always observe ESD precautions as these devices are very sensitive to static discharge.

Optical Fiber Lengths

The measurement setup will typically require optical fibers to interconnect optical components with different connectors. For example, a modulator with an FC/PC connector at the output will require an optical patch cord to adapt to the FC/APC connector on the input of the characterized photodiode accessory. Optical fibers have negligible frequency-dependent loss over microwave modulation bandwidths. Thus, adding short lengths of optical patch cords to the setup does not affect the accuracy of transfer function measurements. To avoid certain polarization-related issues, it is recommended to keep the patch cord length under 10 m. The use of patch cords appropriate for the wavelength being used is recommended, as multi-mode propagation is possible in some cases and may result in some measurement instability.

For the higher RF frequency applications (>70 GHz), polarization dispersion may start to play a role particularly if polarization-maintaining fibers are used after the modulator. This happens because the modulation sidebands on the optical carrier (whose spacing is determined by the RF frequency) can experience sufficiently different phase velocities that the incidental phase modulation can become noticeable. This can change the amplitude response from the photodetector if the distortion is large enough. Some caution is advised with longer runs of polarization maintaining fibers or long runs of fiber under mechanical stress.

Modulator Bias Control

Lithium niobate modulators are generally biased using a modulator bias controller (MBC) to control the operating point of the modulator. When biased in quadrature, the input RF signal linearly modulates the optical carrier. Note that when an MBC is applied, it must be designed for small signal operation. The default power from the Port 1 test port is –10 dBm for the MS4647B when equipped with Options 51, 61 or 62. This level results in a modulation depth of <10% for many commercially available modulators. If a different model VNA or a different option configuration is being used, it may be required to drop the VNA power level from the default setting to ensure that the modulator is operating linearly. Different technology modulators may have different linearity limits so it may be required to consult the manufacturer of that device.

A DC power supply can be used in place of an MBC. However, the stability of the measurement may be degraded due to drift in the modulator's bias point.

These concepts apply to the MN4775X converter; the unit defaults to quadrature bias maintained with a closed-loop dithering system (a small modulation is added to the bias and the detected response to that modulation used for feedback). The dithering frequency is 3 kHz by default, but can be changed (should generally be higher than the current measurement IF bandwidth). There are cases when trace noise, on a fine scale, can be improve by reducing dither amplitude or turning it off, but, again, stability will be affected. Other bias states (modulator full on, full off...) may be useful for certain measurements (particularly DC). An internal calibration of the bias system is required on the MN4775X and normally this happens automatically at first activation and at other times. The calibration may need to be triggered manually when switching wavelengths in multiple wavelength versions (e.g. Option 72).

Connector Care

It is important to establish proper cleaning procedures when connecting fiber optic devices together. Fiber optic cores are made of glass and can easily be scratched or chipped if care is not taken. The connectors found on the MN4765X are of the FC/APC type. APC (Angled Physical Contact) is chosen to help minimize back-reflection. The 8° angle at the end-face of the APC connector has an optical return loss of better than 50 dB. DFB lasers require large amounts of reflection isolation to function properly. The optical return loss from a common PC connector can 30 dB or worse, depending upon the polish and cleanliness of the connector.

The following are some tips to help ensure quality connections:

- Always clean connectors after every connection.
- Use a fiber optic scope often to ensure there are no defects on the connector end face that can cause damage to other connectors.
- Use insertable patch cords for expensive devices that require many connections.
- Always use a cloth that is free of fiberglass to clean the connectors. If necessary, use alcohol to remove stubborn dirt and oil. Thoroughly remove any alcohol residue before reconnecting.
- Avoid using any oils for connecting two cables together. Oils are messy and very difficult to clean up.
- Optical connectors do not need torque. Some connections are better when the two fibers are barely touching. Tightening the connector too much will result in higher insertion loss, more reflection, and in some cases damage to the connector.
- Always observe proper mating to APC connectors. Connecting APC to PC connectors will damage the connectors.

Specification Items

In the ME7848A system Technical Data Sheet, there are a number of parameters unique to the O/E, E/O, and O/O classes of measurements. Repeatability is defined to include trace noise and any short-term instability in optical connections. Because of the relative low received signal level to the VNA (often < –40 dBm), trace noise is generally worse than in a traditional VNA measurement with the same IF bandwidth and averaging.

Dynamic range or noise floor is defined with losses of one or both optical converters included (depending on which measurement is being referenced). As an example, if one is measuring an E/O device, the dynamic range or noise floor metric of interest includes effects of both the VNA and the O/E module. The noise floor can be expressed in dBm terms (power delivered to the instrument) or in absolute responsivity terms of the DUT (A/W or W/A).

27-4 Example 2-Port Procedures

With the physical setup described, the next task is really how to interface with the measurement utilities. Consider first the case of a 2-port E/O measurement where one has a characterized O/E device (such as the MN4765X). The dialog for this measurement is shown in Figure 27-11 (the system automatically detects if it is a 4-port configuration in which case one of the later dialogs would be displayed instead). One choice that has to be made is the port configuration. The default is to have the E/O device connected to port 1 of the VNA, but this can be changed.

Navigation to 2-Port OPTICAL MEASUREMENT Dialog Box

MAIN | Measurement | MEASUREMENT | O/E-E/O-O/O button | OPTICAL MEAS. | E/O Measurements | 2-PORT E/O MEASUREMENT dialog

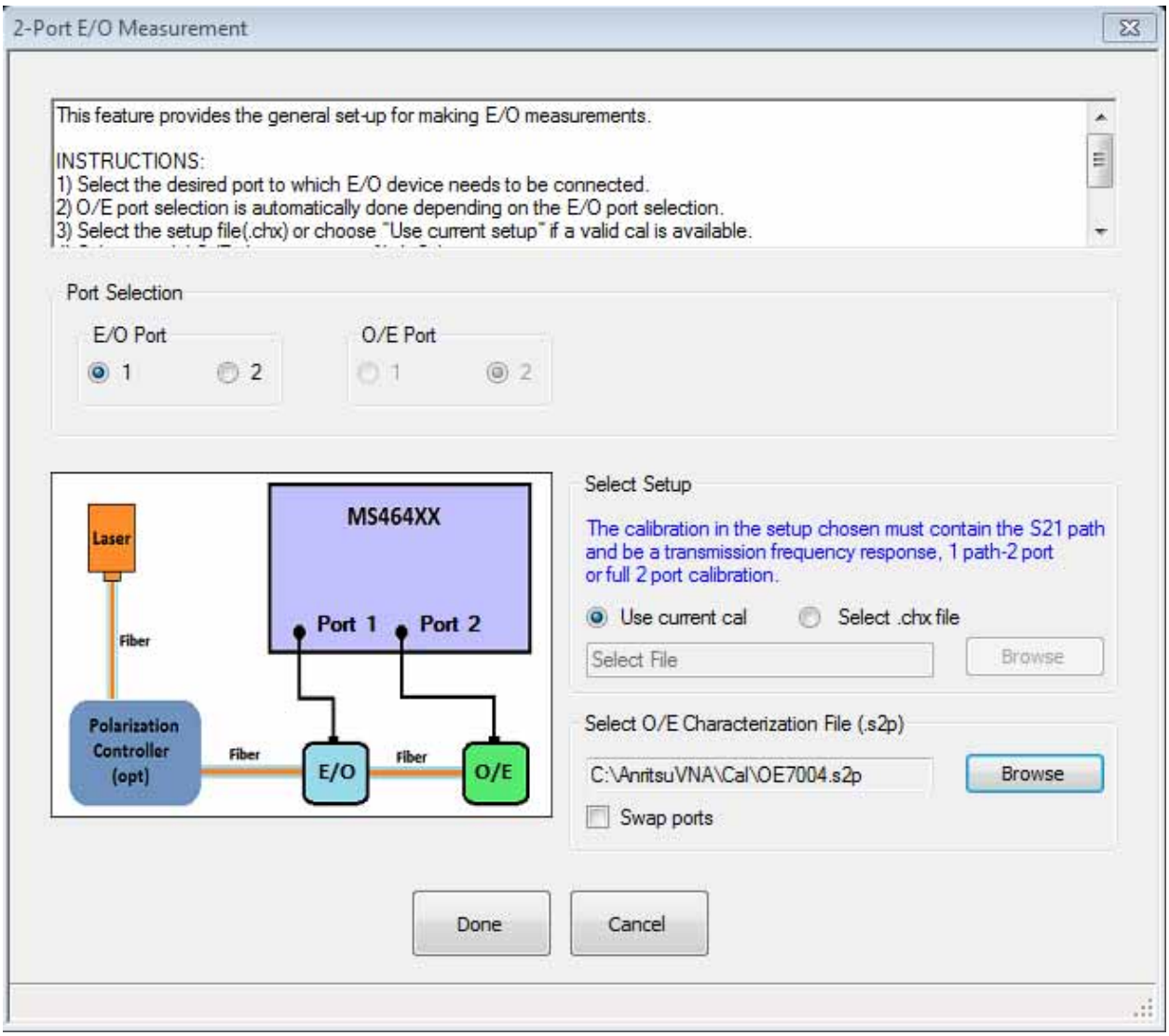


Figure 27-11. Dialog for 2-Port E/O Measurements

The other major requirement is a 2-port VNA calibration. This dialog assumes that the calibration has already been done with the desired method, frequency range, etc. and saved (using Save Setup under the File menu) as a .chx file or is available in the current setup. Most of the setup parameters are obvious:

- **Frequency Range and Number of Points** (dictated by the bandwidth of the DUT among other test requirements)

- **Power Level** (this depends on the linear drive range of the modulator as discussed above and the maximum available power of the VNA, which varies with options. Generally, the higher the better within those limits for best signal-to-noise ratio)
- **IF Bandwidth and Averaging** (trade-off between measurement time and trace noise. Generally 10–100 Hz IFBW is recommended for very wideband devices since the conversion losses tend to be high and the signal-to-noise ratio is stressed). Averaging can be added for additional trace noise reduction but there is a point of diminishing returns beyond about 10 sweep-by-sweep averages or if IFBW/(pt-by-pt averages) falls below 1 Hz.
- **Calibration Method** The only requirement is that the calibration contain a transmission path that match the E/O → O/E path (anything beyond that is acceptable). Some of the choices are:

  E/O Port = 1

  Forward transmission tracking (1 → 2)

  1 path – 2 port forward (1 → 2)

  Full 2 port calibration

  E/O port= 2

  Reverse transmission tracking (2 → 1)

  1 path – 2 port reverse (2 → 1)

  Full 2 port calibration

Any calibration algorithm can be used (SOLT, SOLR, LRL, LRM, etc.) as appropriate.

If the current setup is not to be used, the desired setup file can be loaded in the dialog of Figure 27-11, as can the O/E characterization file (in a .s2p file format). Note the Swap ports checkbox availability near the characterization portion of the dialog box. It is always assumed that the dominant path in the characterization file is the S21 parameter. If the file was constructed differently (e.g., S12 was the measurement path for the calibration device), then selecting the checkbox will force the instrument to reorder the ports in the file before processing.

When Done is selected on this dialog, the de-embedding is applied and the resultant calibration left on the system includes the shifted reference plane as suggested by Figure 27-3 on page 27-4. This state can be saved as a new setup file (again using the Save Setup command under the File menu) if desired. Note that to go back to the non-de-embedded state, one can recall the old setup file that was just loaded in the dialog box. At this point, one can make measurements of any number of E/O devices and save the results using any of the usual techniques.

The O/E measurement setup is quite similar and that dialog is shown in Figure 27-12 on page 27-18. The difference, of course, is that an E/O characterization file must now exist. One can generate it using the previous procedure and saving the measurement result as a .s2p file or one may already have a characterization file for the E/O device. If this is not the case, a shortcut is provided to help generate an E/O characterization file with less work. This path is triggered with the Go Measure E/O button which will bring up the sub-dialog of Figure 27-13 on page 27-19.

Assuming one has an O/E calibration device (such as the MN4765X), this Go Measure feature allows one to load that photodetector characterization file and use the setup file found on the main E/O dialog to do a quick measurement of the modulator (or assembly). The sub-dialog also has a field to save this new E/O characterization file.

When completed, one can go back to the previous dialog and select Done. The resultant setup will now have the Port 1 reference plane moved to after the modulator so one can now measure any number of new O/E devices. Again, the final setup (with shifted reference planes) can be saved as can any new measurement data of the O/E devices.

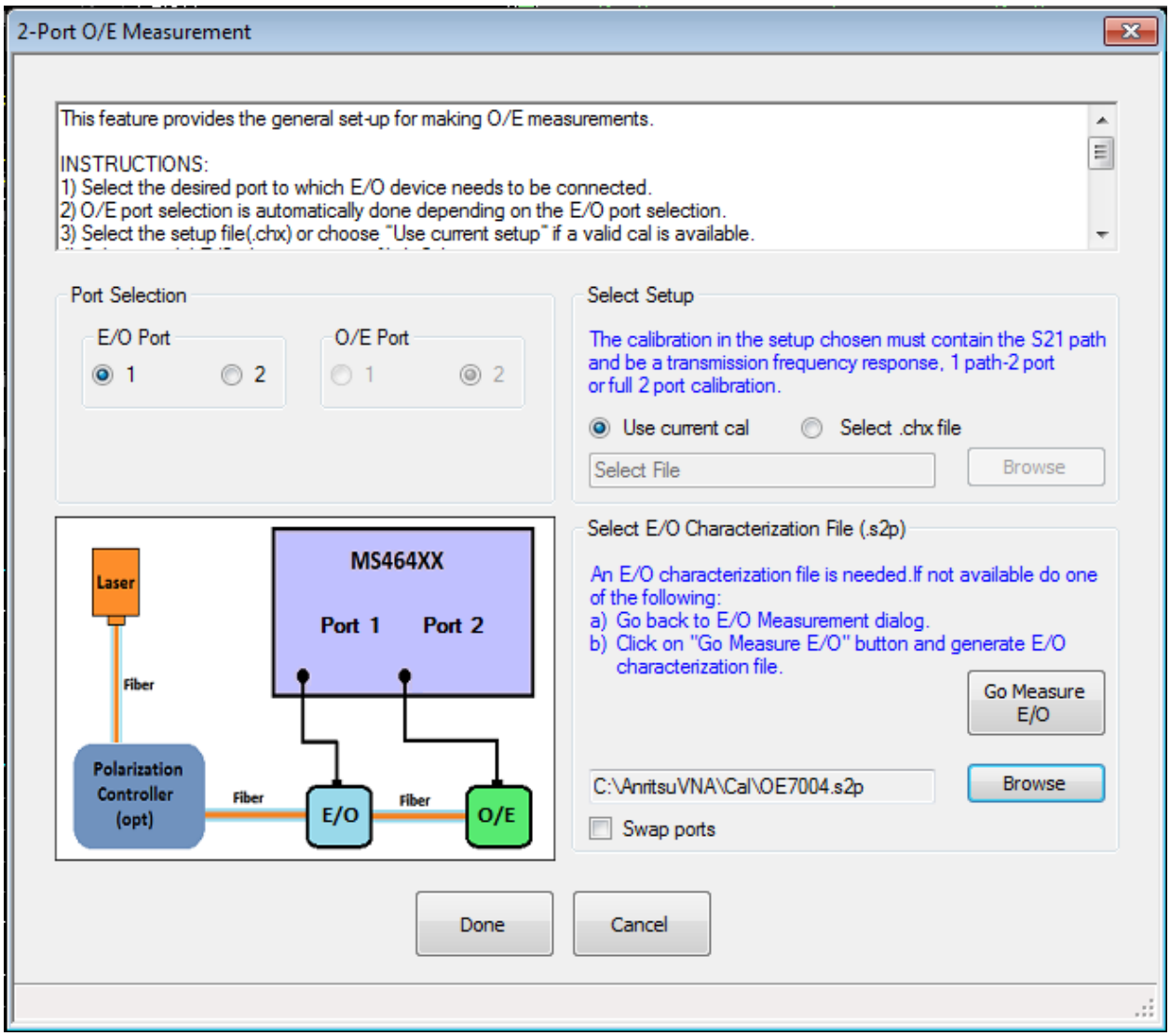
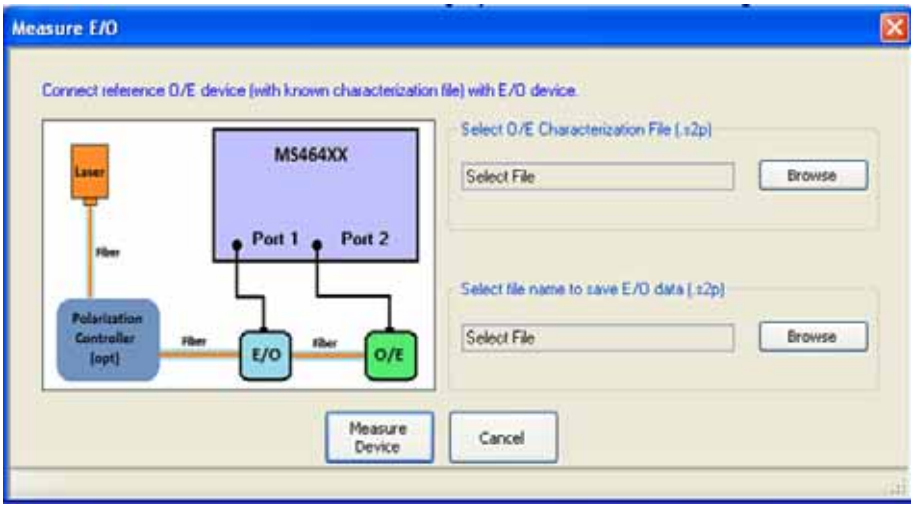


Figure 27-12. Dialog for 2-Port O/E Measurements



When measuring O/E devices, the characteristics of the E/O device must be known. If a file does not already exist, this dialog can help in doing the intermediate measurement with the help of a calibration O/E device such as the MN4765X.

Figure 27-13. Help in Doing the Intermediate Measurement with the Help of a Calibration O/E Device

The O/O measurement is somewhat like the E/O and O/E measurement setups in series: both the detector and the modulator must be de-embedded to leave the reference planes in the optical domain. As suggested by Figure 27-14, if both .s2p files exist, their file names can be entered directly and the de-embedding will proceed. If one of them does not exist (usually the modulator file but either is allowed), it can be measured on the fly much like with the Go Measure process in the O/E measurement previously. This Go Measure process (dialog shown in Figure 27-15 on page 27-21) allows one to enter the known device's file and to define the file name for the newly created file. Note that this Go Measure process assumes the modulator and detector are directly connected together in an optical sense and this defines the reference plane locations. In some sense, this step is like an optical normalization.

Note that the .s2p file name for the known device is assumed to be the same as that used on the main O/O dialog as this device normally doesn't change between Go Measure and O/O configuration steps. If a different device is to be used, the file name on the main O/O dialog can simply be changed after the Go Measure process is completed.

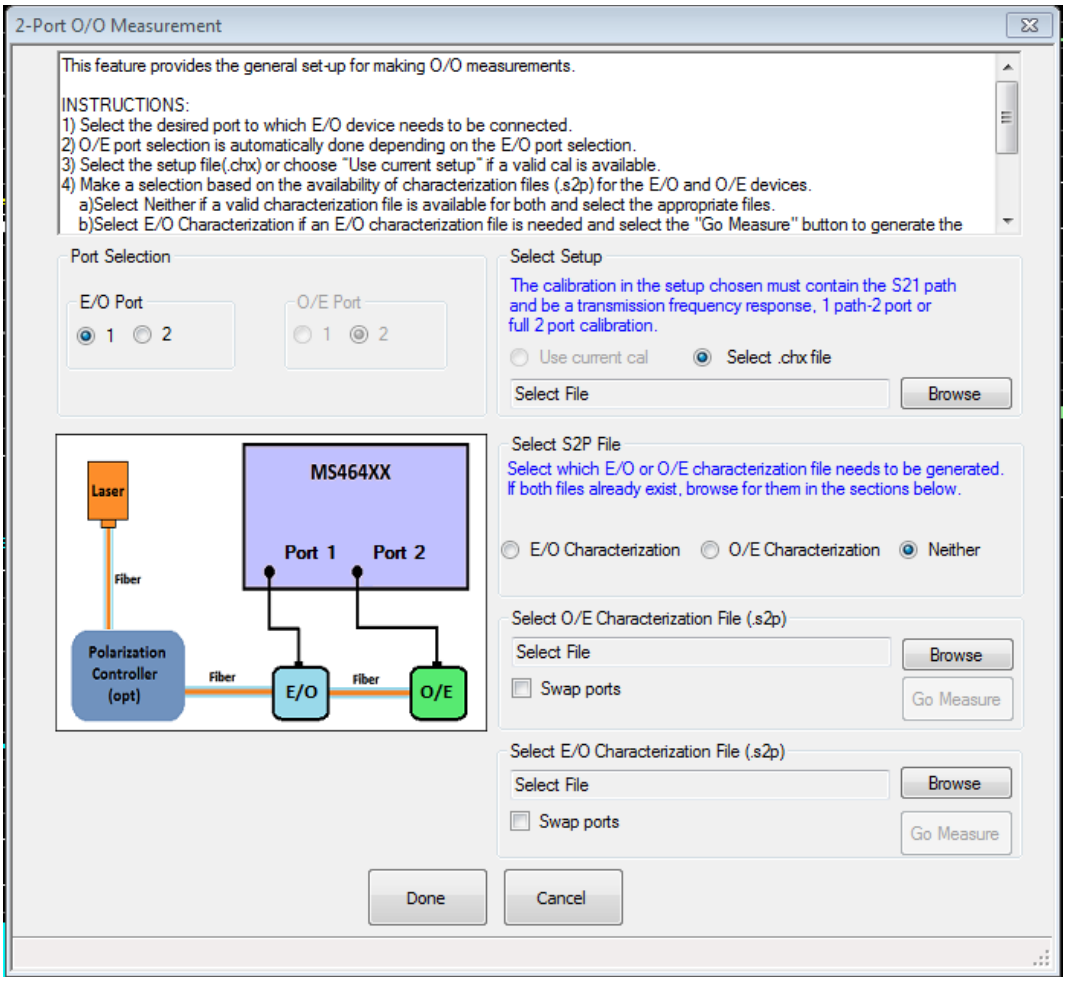
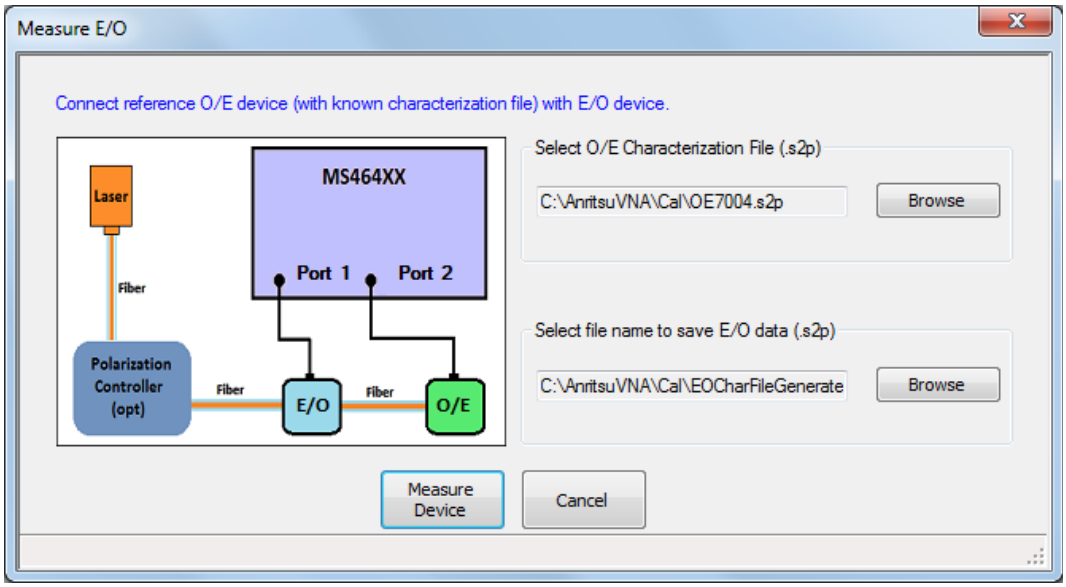


Figure 27-14. 2-Port O/O Measurements Dialog



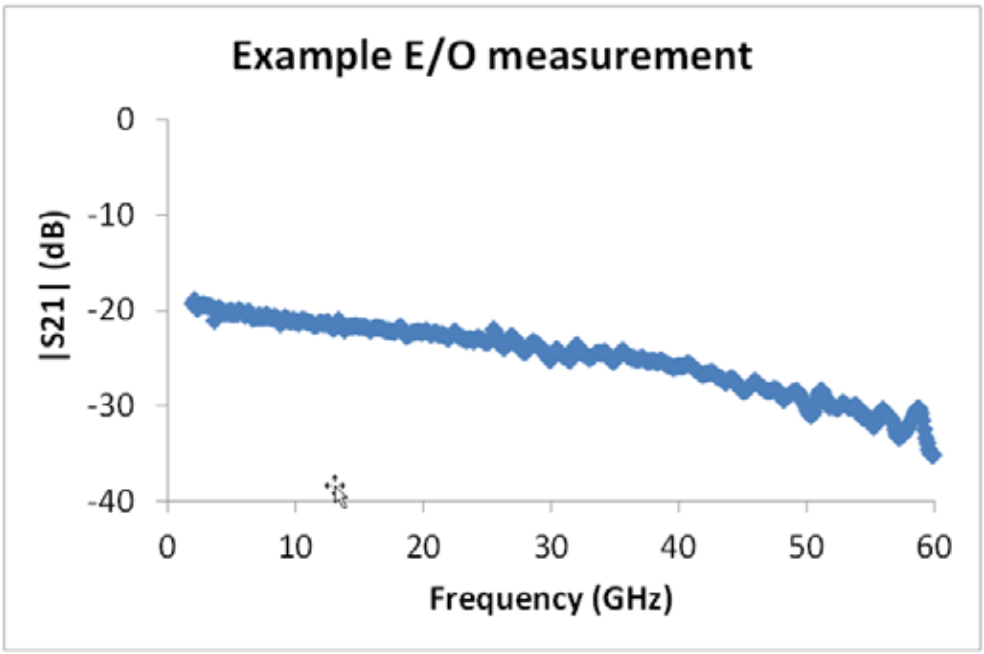
When measuring O/O devices, the characteristics of both O/E and E/O devices must be known. If a file for one does not already exist, this dialog can help in doing the intermediate measurement with the help of the file for the other device (usually a calibration O/E device such as the MN4765X). At least one converter must have a .s2p file to do the measurement.

Figure 27-15. Help in Doing the Intermediate Measurement with the Help of the File for a Second Device

Example E/O Measurement

As an example, consider the following setup:

- 2 GHz to 60 GHz, 10 Hz IFBW, power –10 dBm, no averaging.
- A full 2-port SOLT calibration is performed using a 3654D Calibration Kit.
- The resulting setup file is saved as setup.chx. This setup file will include all of the above parameters including the calibration data.
- The optical assembly is hooked up with the modulator-under-test connected to Port 1 and an MN4765X O/E calibration module connected to Port 2 of the VNA. The laser is powered up after setting up bias control on the modulator and applying bias to the MN4765X (and, of course, connecting the fiber).
- The E/O measurement utility is invoked using the default port assignment, loading the .chx setup file and loading the characterization file of the MN4765X. Upon selecting Done on the dialog, the resulting S21 measurement reflects the conversion behavior of the modulator and is shown in Figure 27-16 on page 27-22. In this case, there is about 10 dB of roll-off over the bandwidth of the measurement.



The results of an example 2-port E/O measurement are shown here using the procedure discussed in the text.

Figure 27-16. Results of an Example 2-Port E/O Measurement

27-5 Example 4-Port Procedures and Variations

When a 4-port VNA is used, there are obviously more port permutations possible for connections but also differential or multiport O/E and E/O devices can be handled. The measurement processes are essentially the same as for the 2-port cases just discussed but there are some additional logistics to handle. For the case of solving for single-ended E/O device when used with a single-ended O/E device (Case 1 Configuration), the dialog of Figure 27-17 applies and the main choice is which ports are to be used.

Navigation to 4-Port E/O Measurement Dialog Box

MAIN | Measurement | MEASUREMENT | Perform Optical Measurements| OPTICAL MEAS.

The same rules discussed in the previous section apply on the type of calibration used (in the .chx file to be loaded or the current setup), except that the path selected (using the port checkboxes) must be covered by that calibration. In addition, full 3-port and 4-port calibrations can be used if desired (where the 3-port calibration must cover the ports being used). The characterization file in this case is again a .s2p format file with S21 as the parameter of interest.

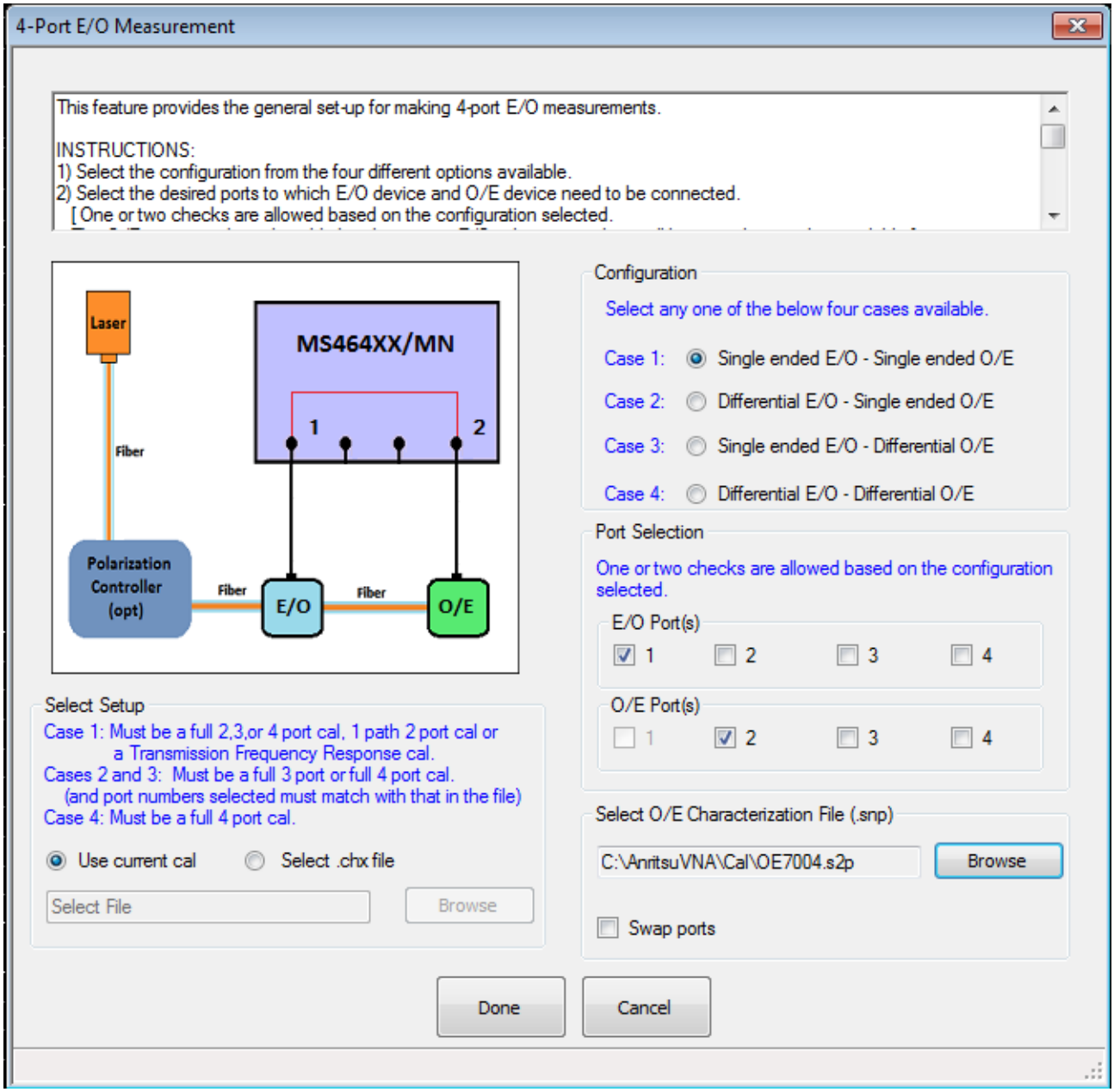


Figure 27-17. Case 1 Dialog—Single-ended 4-port E/O Measurements

For the case of one or more differential (or multiport) optical devices, the additional "cases" apply and the dialog changes slightly as shown in Figure 27-18, Figure 27-19, and Figure 27-20.

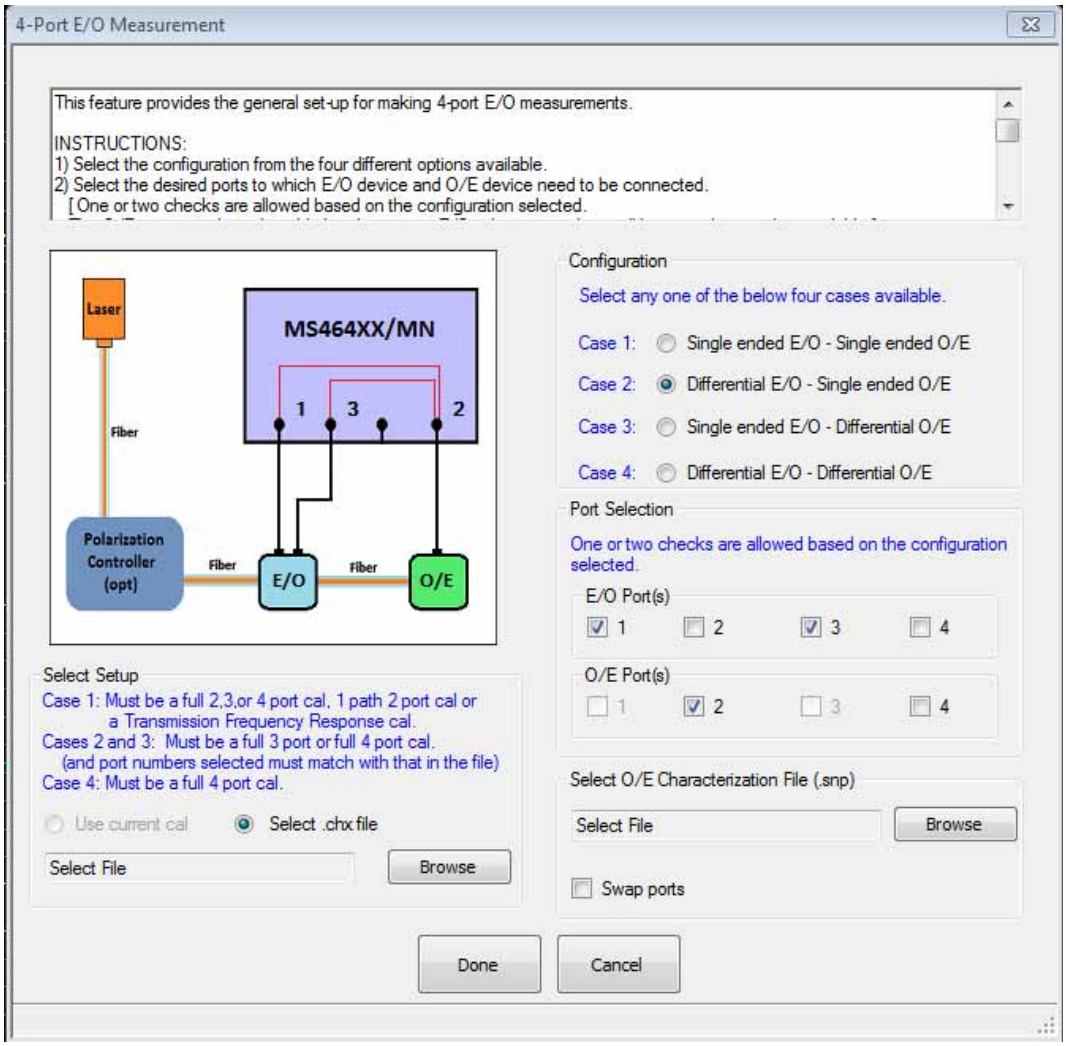


Figure 27-18. Case 2 Dialog—Variations for Different DUT Port Structures

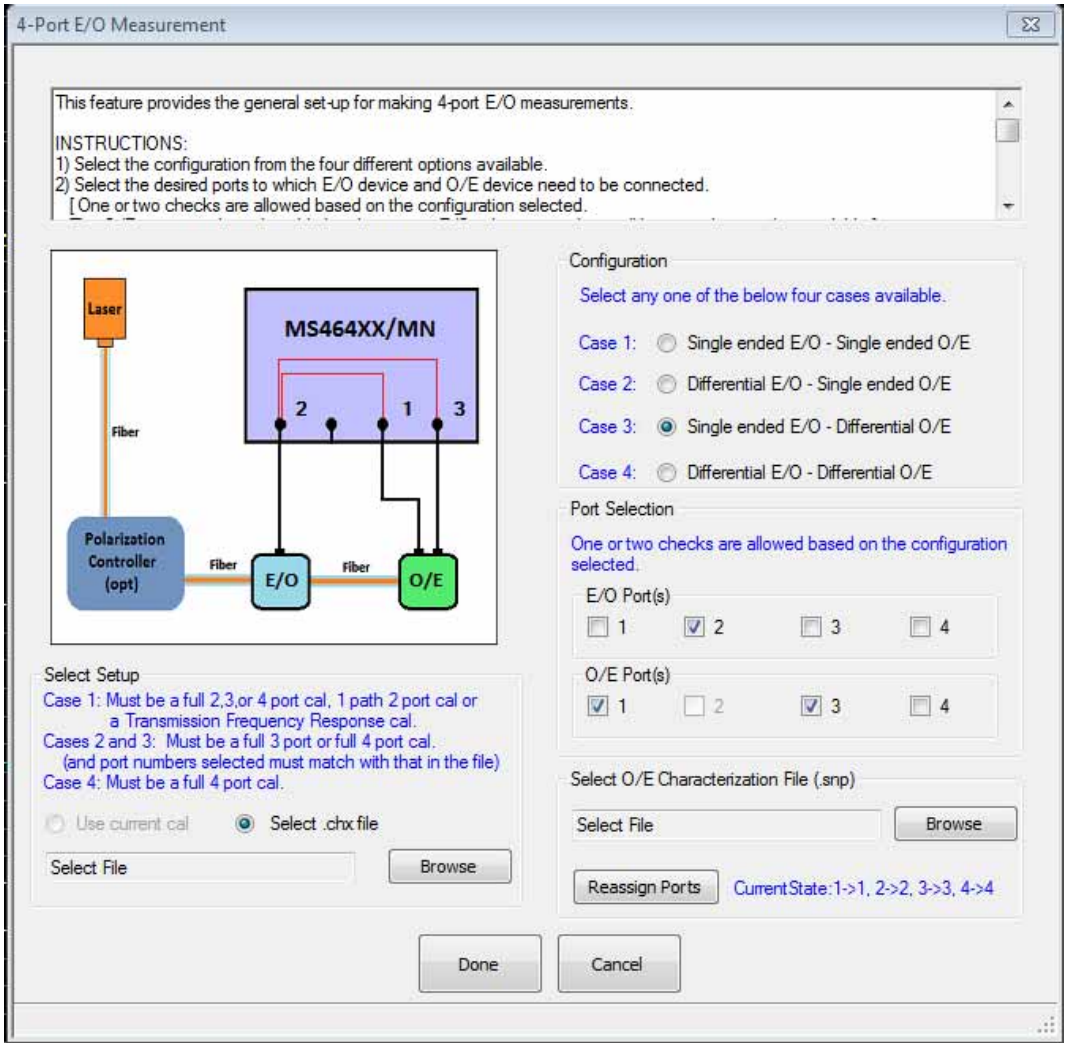


Figure 27-19. Case 3 Dialog—Variations for Different DUT Port Structures

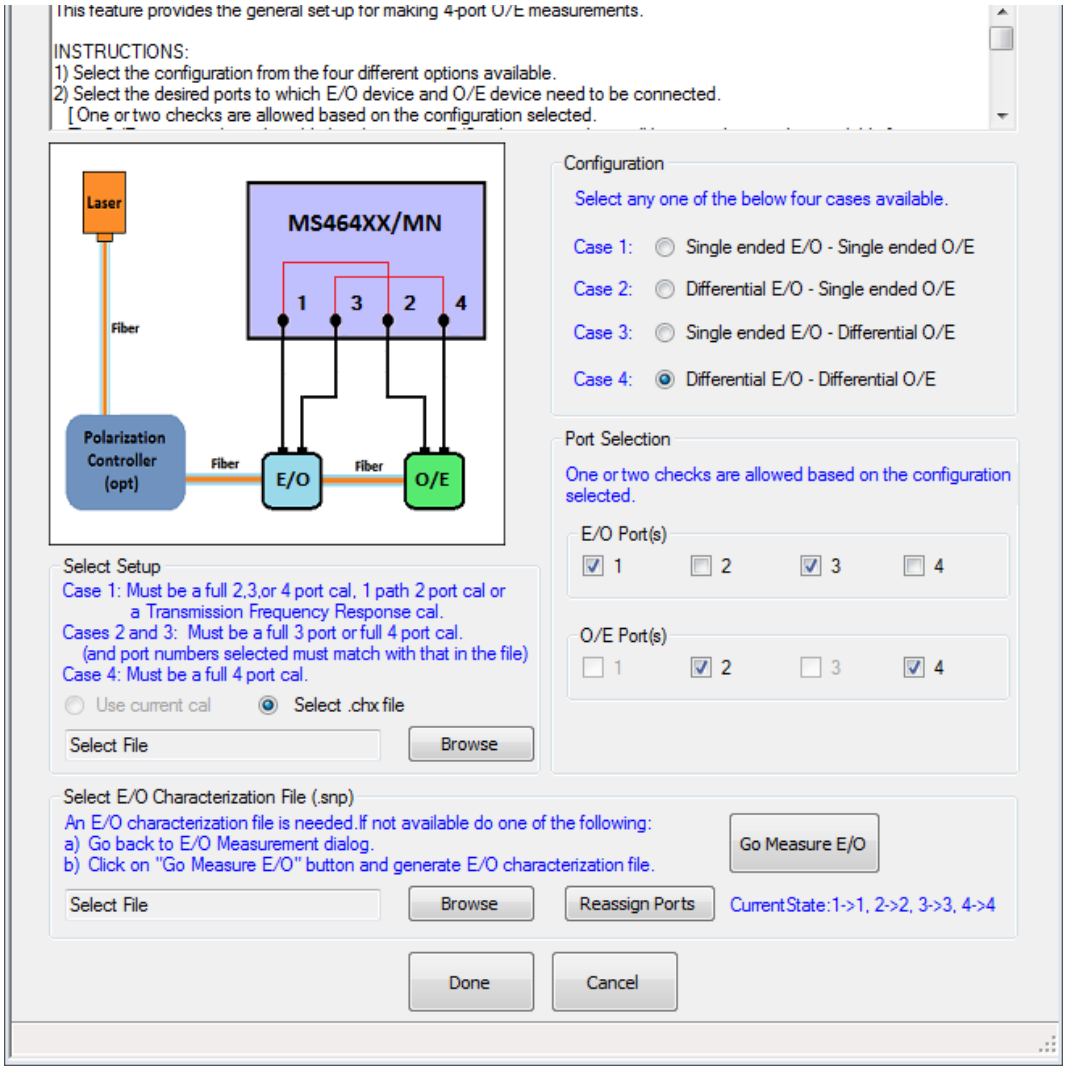


Figure 27-20. Case 4 Dialog—Variations for Different DUT Port Structures

A few things change in these other cases:

Case 2: Differential E/O—Single-ended O/E

The .chx file (or current setup, if that is used) must contain a full 3-port calibration (using the ports in question) or a full 4-port calibration of any algorithm (including hybrids).

The O/E characterization file is still of the .s2p form (S21 is the dominant parameter).

Case 3: Single-ended E/O—Differential O/E

The .chx file (or current setup, if that is used) must contain a full 3-port calibration (using the ports in question) or a full 4-port calibration of any algorithm (including hybrids).

The O/E characterization file is of a .s3p form (S21 and S23 are dominant) or a .s4p form (S21 and S43 are dominant).

The *reassign ports* option is available for assigning the ports in the characterization file to match the path expectations. The current state of port reassignment is shown as a read-only field in the dialog. The default is no reassignment ($1 \rightarrow 1$, $2 \rightarrow 2$, $3 \rightarrow 3$, $4 \rightarrow 4$).

Case 4: Differential E/O—Differential O/E

The .chx file (or current setup, if that is used) must contain a full 4-port calibration of any algorithm (including hybrids).

The O/E characterization file is of a .s3p form (S21 and S23 are dominant) or a .s4p form (S21 and S43 are dominant).

A *Reassign Ports* option is available for assigning the ports in the characterization file to match the path expectations.

In this case, there is also potential confusion as to how the DUT is exactly connected if it is "2 paths in parallel" instead is a true differential device pair. It is always assumed that the lower numbered E/O port is connected to the lower numbered O/E port. If two DUTs are being measured in parallel, make sure the port connections match this assumption.

As in the 2-port case, when one clicks Done to leave this dialog, the resident calibration will now have the O/E device de-embedded and the live measurements will reflect the behavior of the E/O device. Data and setups can be saved from this state as usual.

Caution	The .s3p and .s4p files loaded as characterization files have assumed transmission paths as detailed in the text. Use the **Reassign Ports** feature to make your file match those assumed paths.

For the O/E measurement case, the permutations are essentially the same and these are shown in Figure 27-21. The only real difference is that case 2 and case 3 swap roles in terms of file types required as is obvious in the figure.

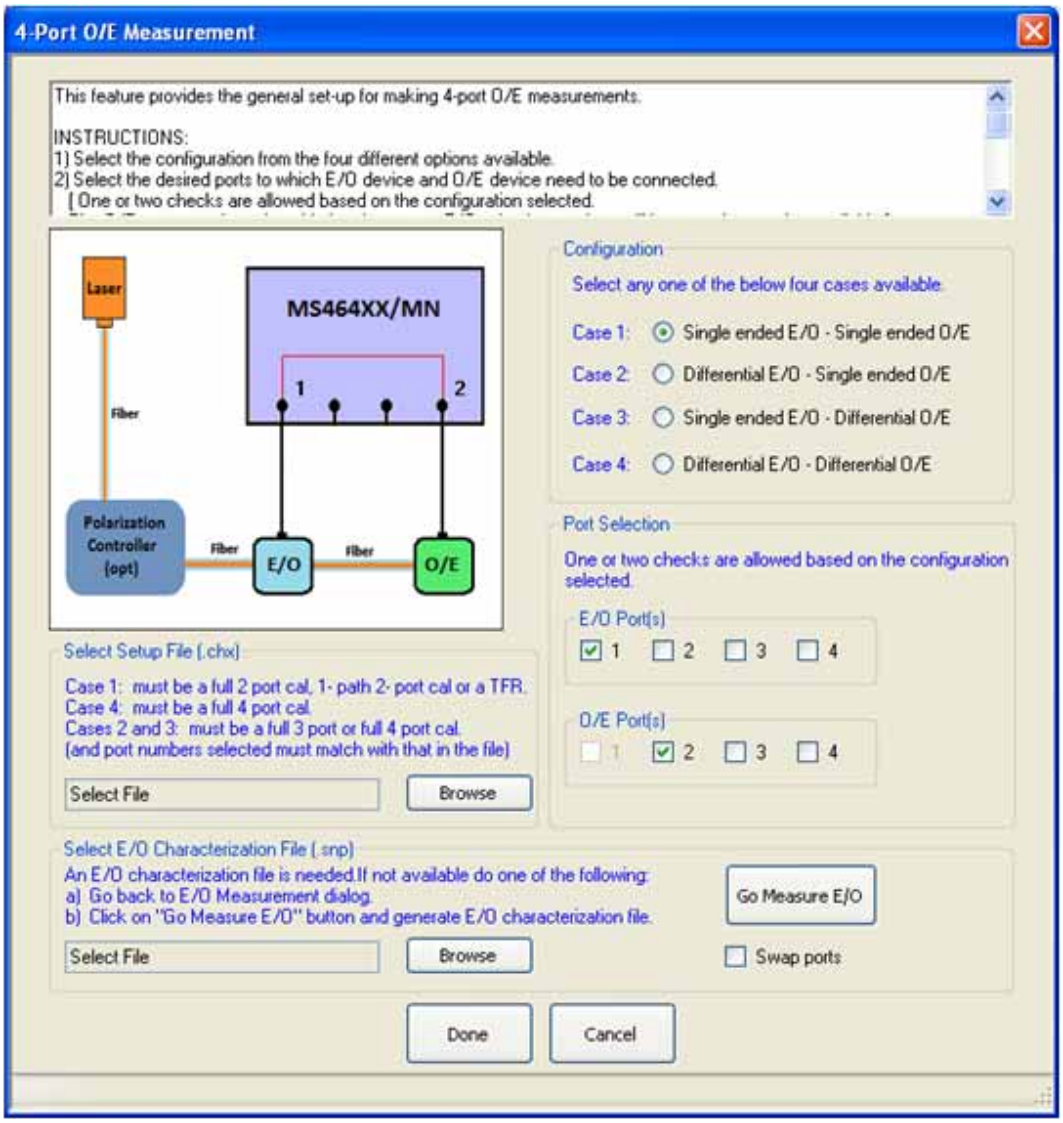


Figure 27-21. Dialog for 4-Port O/E Measurements (1 of 4)

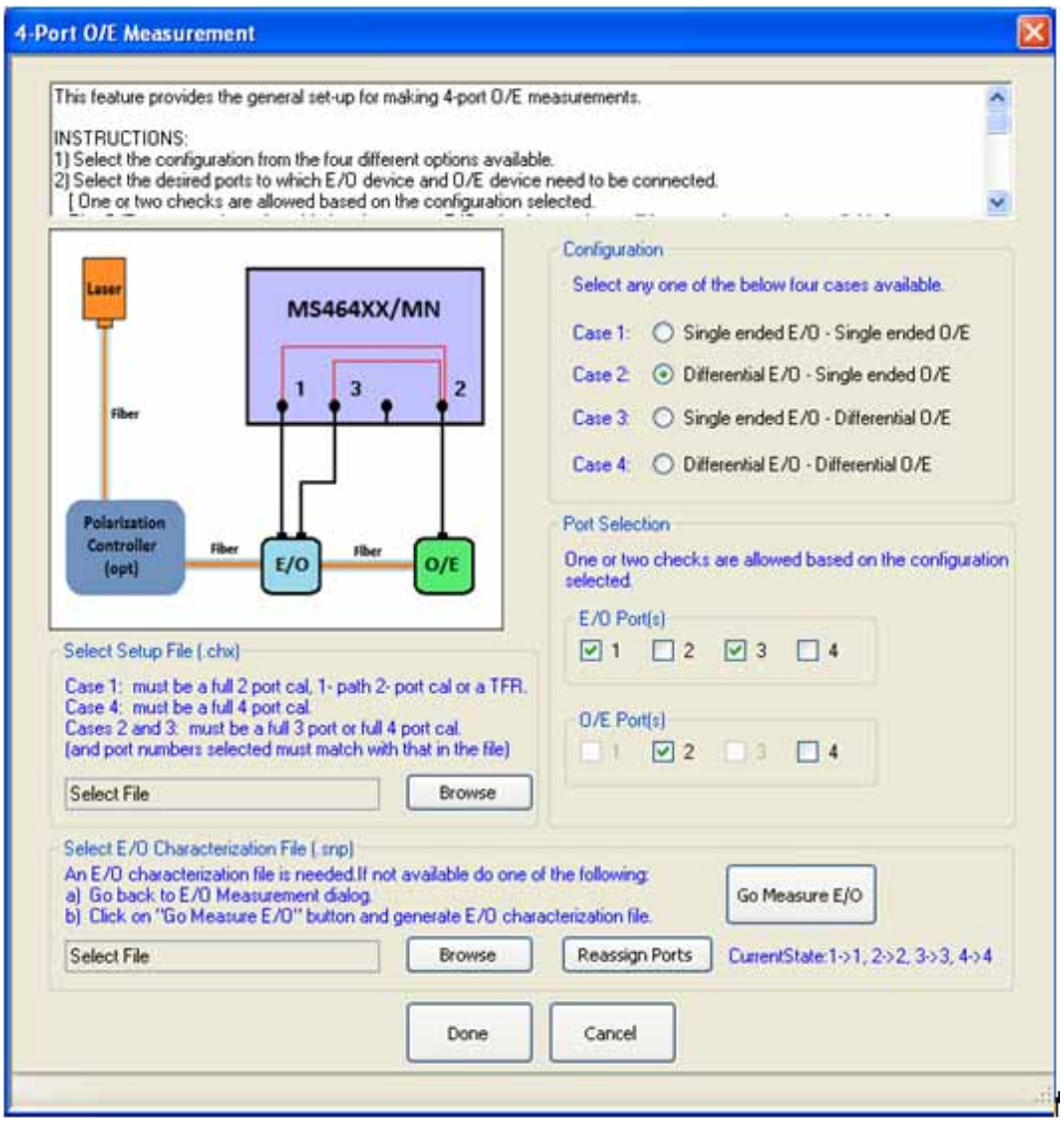


Figure 27-21. Dialog for 4-Port O/E Measurements (2 of 4)

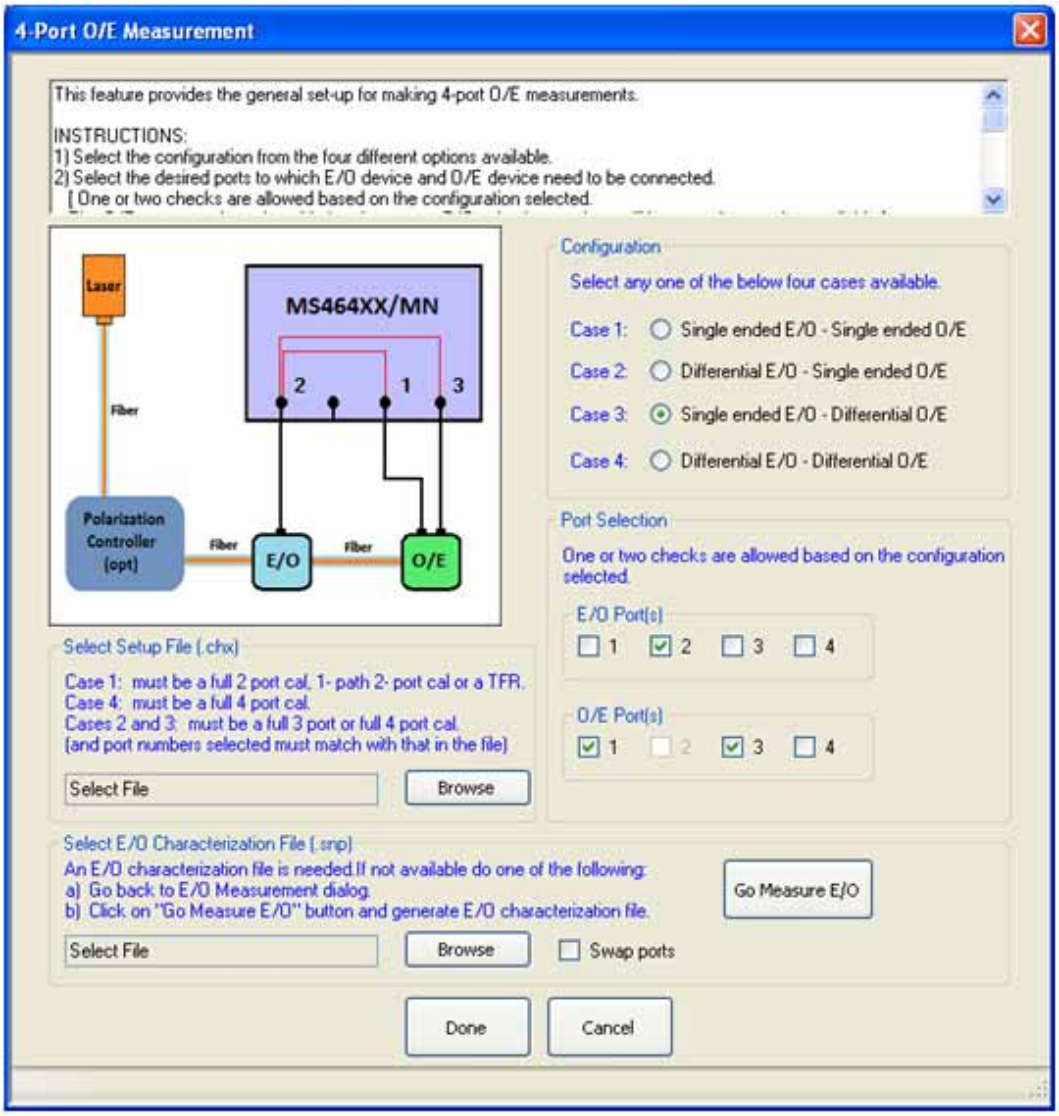


Figure 27-21. Dialog for 4-Port O/E Measurements (3 of 4)

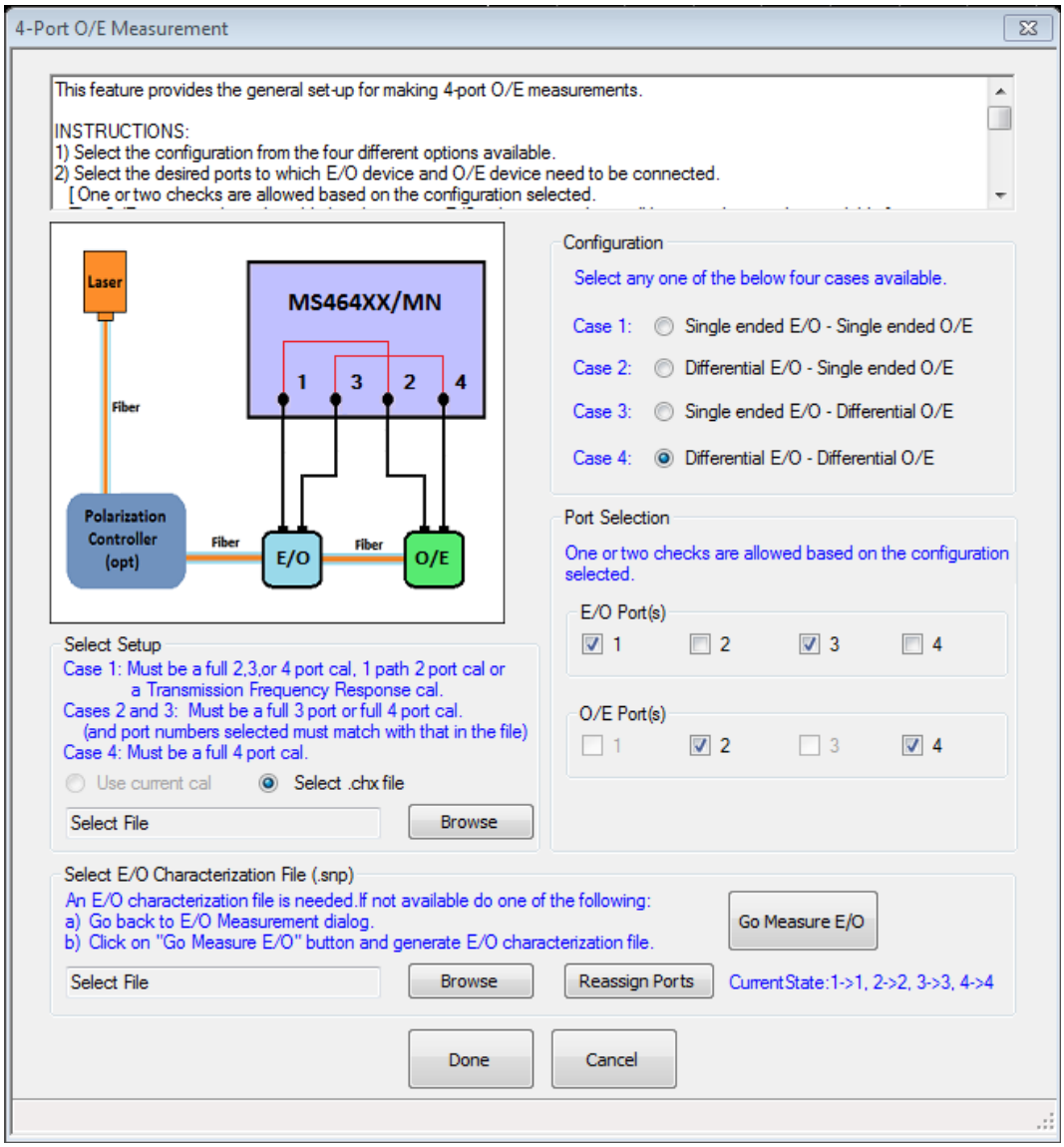
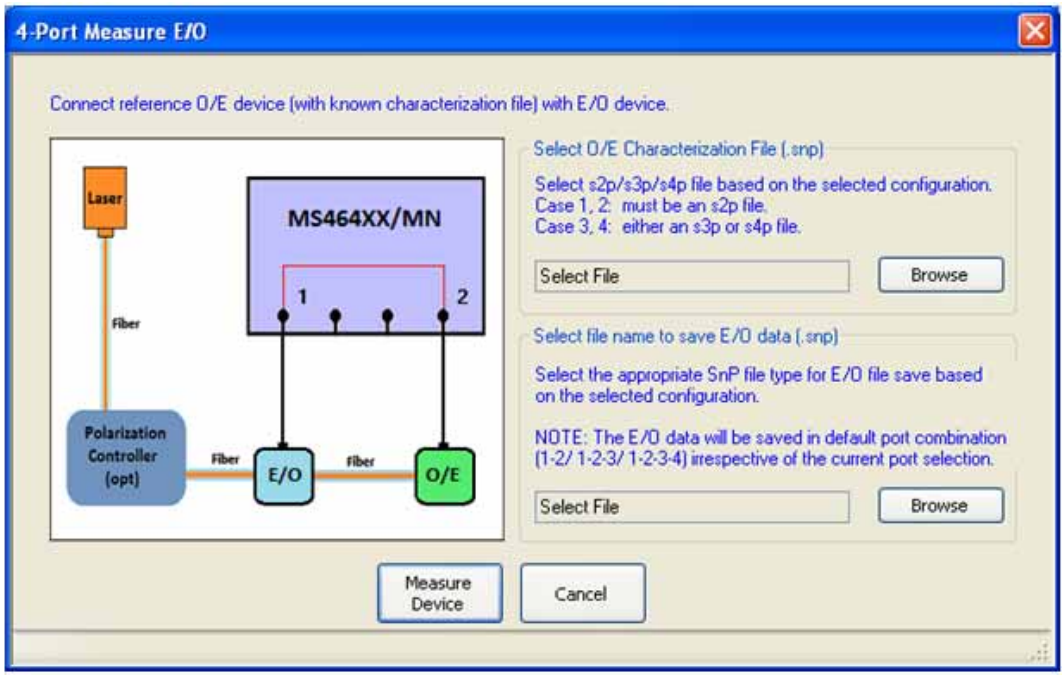


Figure 27-21. Dialog for 4-Port O/E Measurements (4 of 4)

As with the 2-port O/E measurement case, there is a shortcut to go measure the E/O device to get its characterization file (assuming one has a characterized O/E device to start with such as the MN4765X). This shortcut sub-dialog is shown in Figure 27-22 and it must use the same case and port assignment as the parent dialog. In addition, the E/O file that is generated from this sub-dialog will follow the dominant port assignment paths that have been detailed in this section.



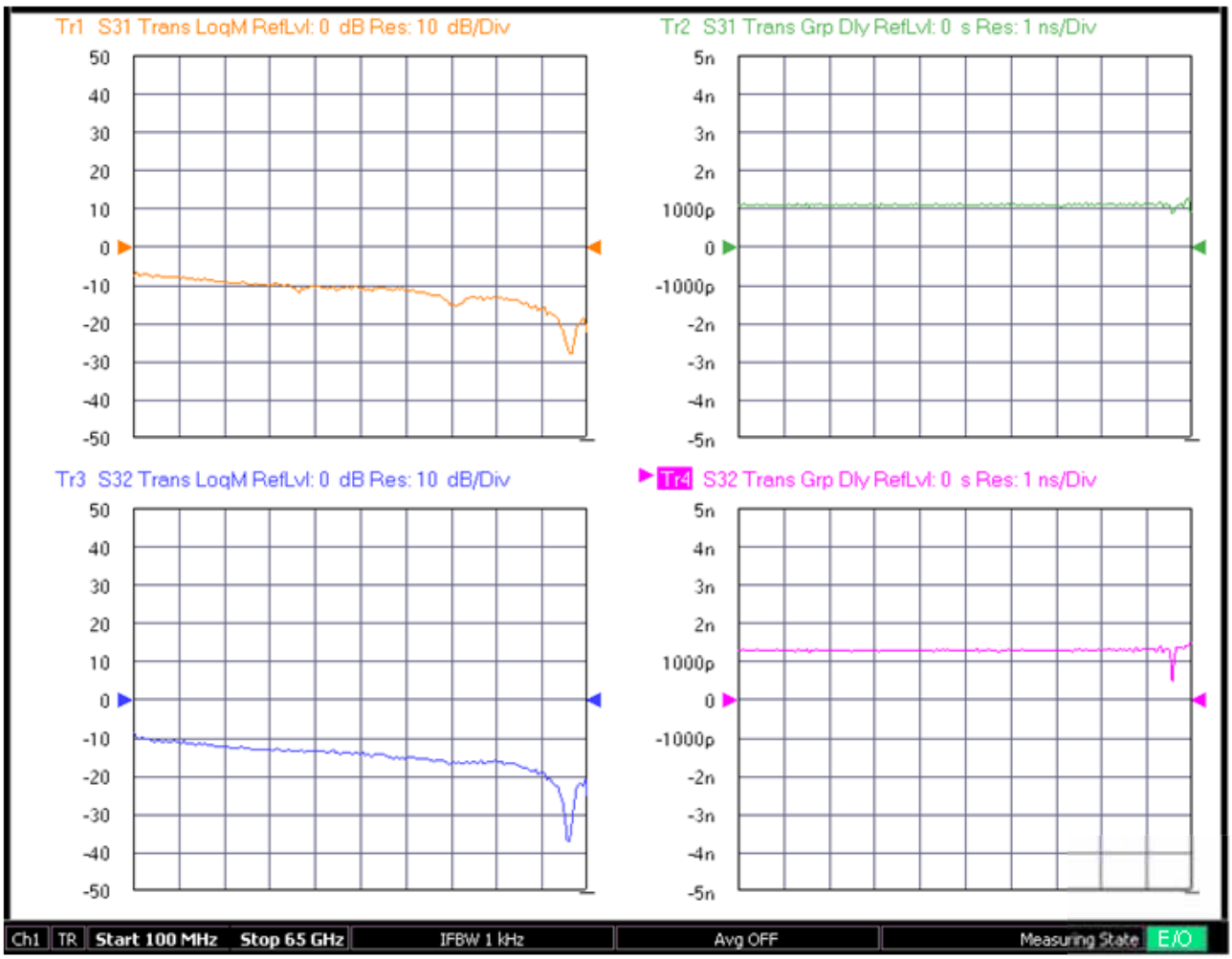
The 4-port intermediate measurement dialog is shown here for O/E measurements when the E/O characterization file does not already exist.

Figure 27-22. Dialog for 4-Port Intermediate Measurement

4-port O/O measurements proceed analogously. The dialogs are very similar to those seen previously and will not be shown here for brevity.

As an example, consider the following setup:

- 100 MHz–65 GHz, 1000 Hz IFBW, –17 dBm drive power
- A double-path E/O device feeding a single-ended MN4765X (so this becomes case 2 under E/O measurements). The E/O device will be connected to ports 1 and 2 while the O/E device will be connected to port 3. The measurements of interest are then S31 and S32. The O/E characterization file is still of the .s2p format where S21 is the controlling parameter.
- A full three port calibration is performed using the 3654D calibration kit (SOLT) and the setup file is saved (.chx).
- The optical components are connected and powered in the appropriate order as discussed previously.
- The E/O dialog is configured as described above and, after 'Done' is selected, the resulting measurements are those of the E/O device. Some example possible measurements are shown in Figure 27-23 on page 27-34. In this case, the device is not as symmetric as hoped (note the differences between S31 and S32) in terms of both conversion magnitude and group delay. A strong resonance is observed in both paths near 62 GHz and lower-Q resonances are seen in the S31 path at lower frequencies. This type of data could be useful in device optimization/redesign.



In this case, the data represents a dual-path E/O device being analyzed.

Figure 27-23. Example of what could be a Common ‘case 2’ Measurement

27-6 Uncertainties

In the measurements just described, the uncertainty can be broken into two broad categories:

1. Uncertainty associated with the characterization of the calibration device (such as the MN4765X)
2. Uncertainty in the measurement with the DUT

Typically the user will purchase a characterized photodiode and receive a data file describing that device's transfer function. There is some uncertainty associated with that data based on the characterization technique.

There are different levels of characterization possible. A direct characterization using such optical techniques as electro-optic sampling (e.g., [1]) or the heterodyne method (e.g., [2]) would be termed 1st tier. It is outside the scope of this note to go into detail on these characterization techniques but they are discussed in the references and elsewhere in the literature. These 1st tier characterizations are typically performed at National Metrology Institutes (NMIs, such as NIST or NPL) or large private laboratories.

A 2nd tier standard is characterized by a laboratory based on a 1st tier standard. It would be generated using a process similar to the techniques discussed here but under carefully controlled conditions (bias, temperature, wavelength, etc.) using a 1st tier standard as the characterization device. This 2nd tier device is much more commonly available (the MN4765X is an example) and the uncertainties for a 2nd tier standard will be used in later calculations. The uncertainty penalty in going to 2nd tier device is typically small (on the order of 0.1 dB additional).

When measuring the DUT, there is an uncertainty associated just with the VNA measurement which is discussed elsewhere (e.g, [5]). Since the characterized photodiode response is then de-embedded, the characterization uncertainty must be combined with the VNA measurement uncertainty to obtain an overall value. In the case of an O/E measurement, there are actually two user measurements involved (one with a modulator and the characterized photodiode and one with that modulator and the DUT) so an additional uncertainty must also be included (note the distinction between this second level de-embed and a 2nd-tier calibration device). Typically these uncertainties are all added on a root-sum-square basis since the measurements are assumed to be dominated by uncorrelated quantities.

Before proceeding to some uncertainty values, it may be useful to examine dependencies. On the VNA side, S21 uncertainty is typically quite low for medium power levels but will deviate at high signal levels (receiver compression, not an issue in these measurements) and at low signal levels (effects of the receiver noise floor). Thus the overall uncertainty will be a function of detector output signal level (getting worse as the signal level gets closer to the noise floor). Results can be improved by using higher RF drive levels (keeping all devices linear) and high optical power levels (same caveat). The RF match of the modulator and photodiode will also influence uncertainties to some degree (at all signal levels) but the dependence is relatively weak as long as return loss is better than a few dB. For the plots in Figure 27-24 and Figure 27-25, a modulator match of –24 dB at low frequencies to –15 dB at 65 GHz and a detector match of –20 dB at low frequencies to –9 dB at 65 GHz were assumed as is typical for some commercial devices. Making both matches worse by 5 dB causes an uncertainty degradation of about 0.15 dB.

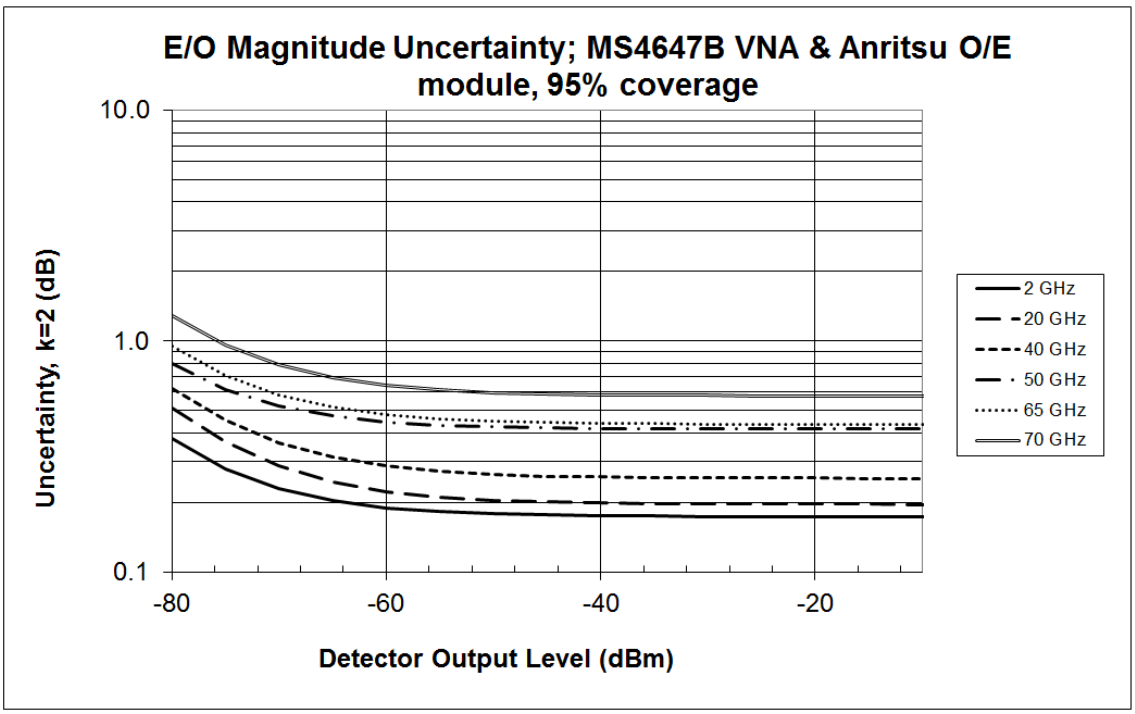
On the optical side, there is little high level signal dependence as long as the devices are operating linearly. The characterized photodiodes are usually chosen to be very linear over wide power ranges to keep this from being an issue. The characterized detectors typically also have very weak wavelength dependencies; usually less than a few hundredths of a dB over 40 nm. When using a modulator as a transfer standard (as in an O/E measurement), however, it is important that the same wavelength be used in the different measurements with that modulator since much greater wavelength sensitivity usually exists in that component (although this will vary with technology). Specifications for the Anritsu MS4647X VNA and 3654D V Calibration Kit were used to calculate VNA uncertainties at various frequencies. The use of different VNAs and/or calibration kits may result in slightly different values. Characterization uncertainties of the MN4765X O/E calibration device were used for the standard part of the model. Optical system drift is included in the error model but it is assumed that all components are mechanically and thermally stable. Connector repeatability is also included in the model but all connectors are assumed to be in very good condition. The results are shown with an independent variable of photodiode output power and plotted for both magnitude and phase for the two different types of measurements. Frequency is used as a parameter. While the results for a 2-port measurement are shown here, there are normally no significant deviations in going to 4-port structures.

Generally, the MN4765X is treated as the traceable optical converter element in the uncertainty analysis since it has the most direct path to a standards laboratory. When measuring an OE device, the characteristics of the MN4765B are first transferred to a modulator (like the MN4775X converter), and then that modulator response is de-embedded from the DUT measurement. As will be seen, this additional transfer adds some small amount of uncertainty (for repeatability and additional VNA measurement reasons). A separately characterized modulator can, of course, be used as well, but, usually that will have a slightly higher starting uncertainty since an OE transfer probably occurred earlier (and one would have to include the aging of that characterization). An OO measurement will typically involve two such transfers, so the uncertainties increase a bit more.

Example uncertainty curves have been generated here that are based on using 70 GHz/1550nm version of the MN4765X as the 2nd tier calibration device. As can be seen in Figure 27-24 and Figure 27-25, the uncertainty reaches an asymptote for detector output levels above about –60 dBm. Above this level, the dominant source of uncertainty passes from measurement signal-to-noise ratio to characterization uncertainty. The x-axes in these plots are indexed in received VNA power, so optical power, modulator optical path loss, other optical path losses, and responsivities all play a role. The ME7848A system uncertainties (see the appropriate technical data sheet) are slightly different since actual converter mismatch values and repeatabilities are used, but the trends are largely the same for all configurations within a wavelength family. There are more substantial differences between wavelengths because the first tier calibrations (by the metrology institutes) have different uncertainties per wavelength and the MN4765B behaviors vary significantly with photodiode used therein.

There is some frequency dependence for several reasons:

1. Noise floor is higher at higher frequencies.
2. The characterization uncertainty goes up with frequency.
3. The basic VNA uncertainty (due to residual mismatch, etc.) goes up with frequency.



Example E/O measurement uncertainty plots are shown here for the conditions discussed in the text.

Figure 27-24. Example E/O Measurement Uncertainty Plots

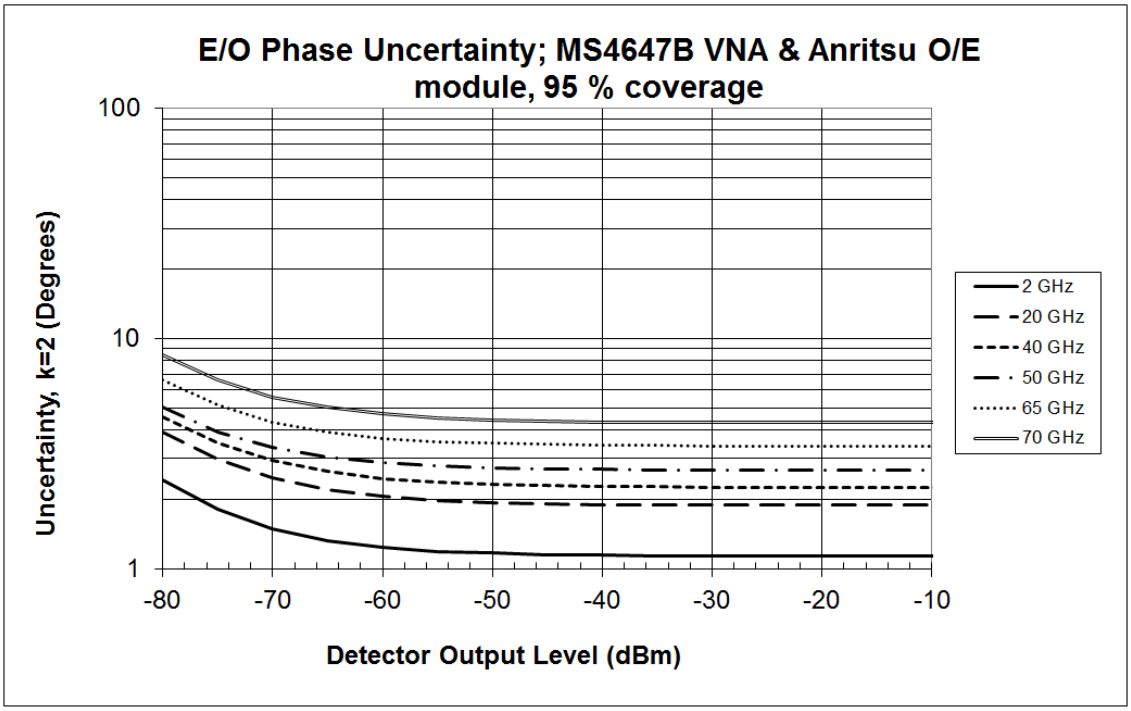


Figure 27-24. Example E/O Measurement Uncertainty Plots

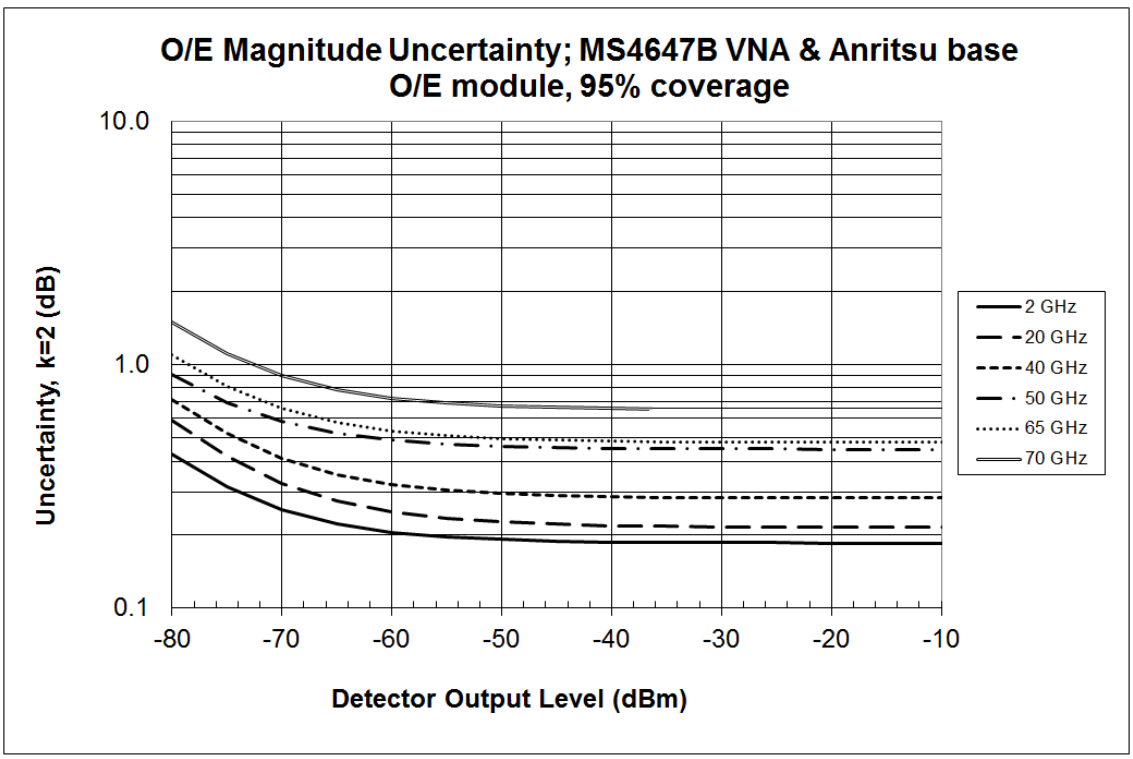
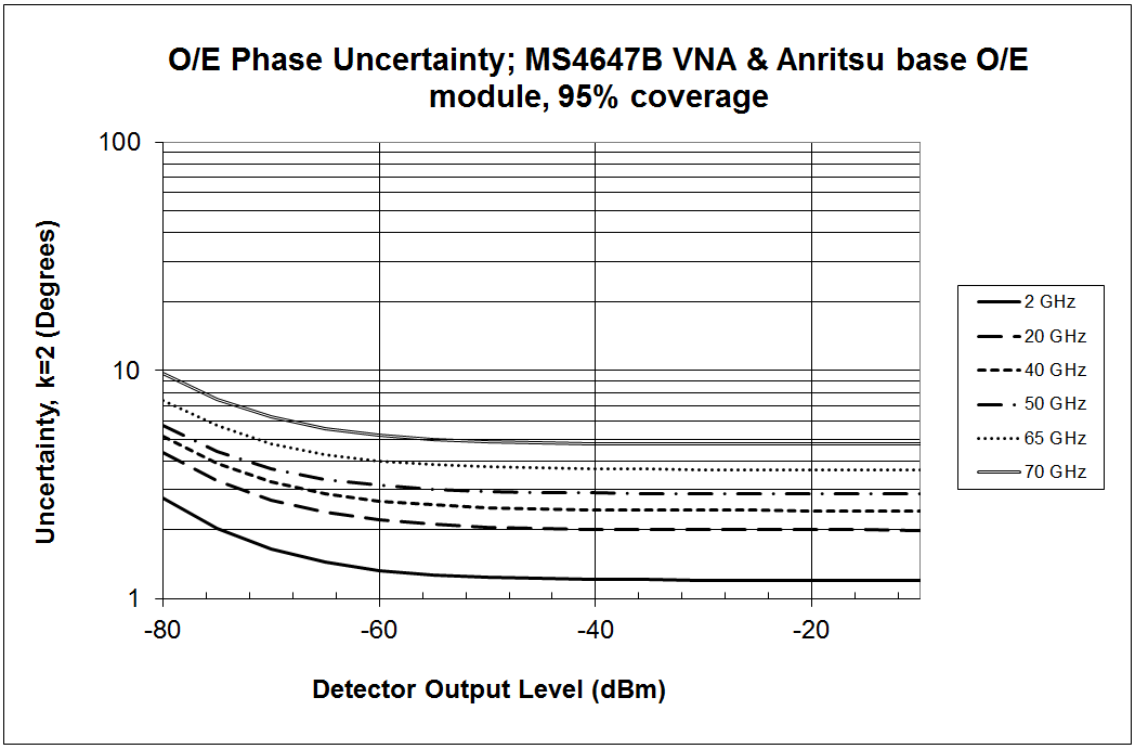
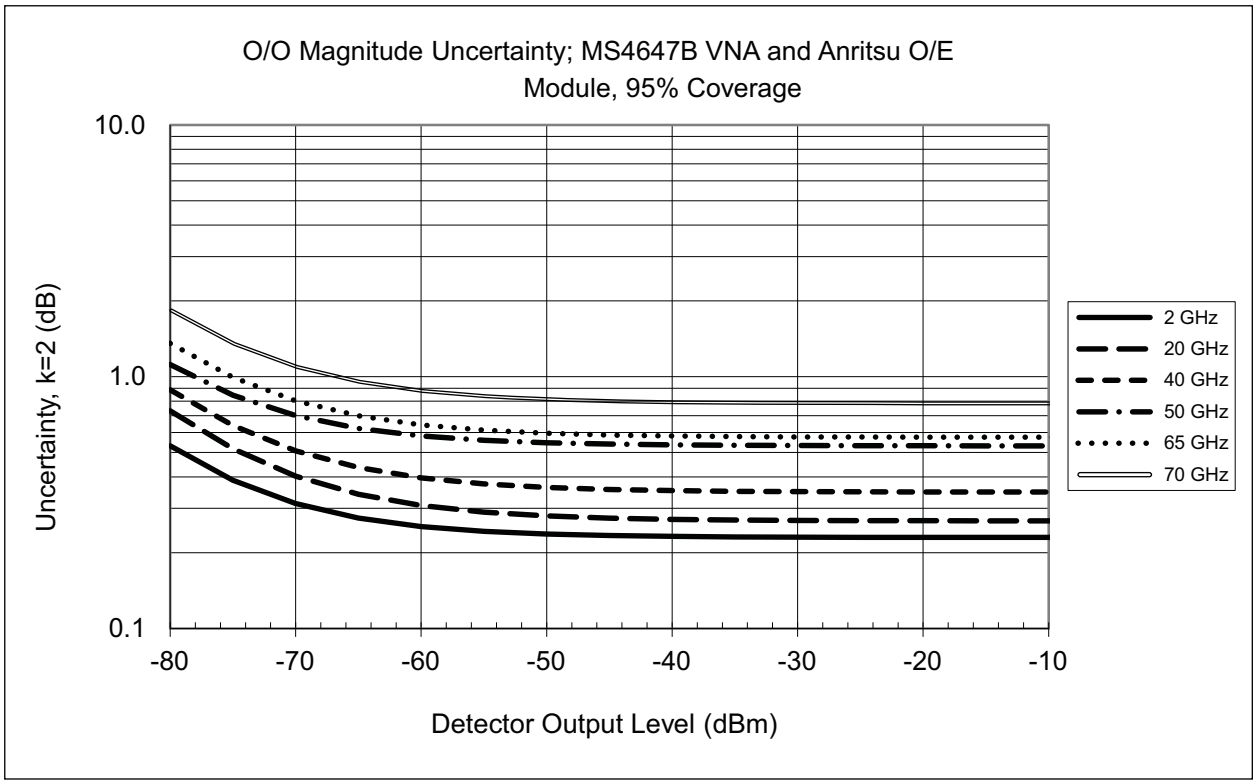


Figure 27-25. Example O/E Measurement Uncertainty Plots (1 of 2)



Example O/E measurement uncertainty plots are shown here for the conditions discussed in the text.

Figure 27-25. Example O/E Measurement Uncertainty Plots (2 of 2)



Example O/O measurement uncertainty plots are shown here for the conditions discussed in the text. Optical component mismatch effects are not included.

Figure 27-26. Example O/O Measurement Uncertainty Plots

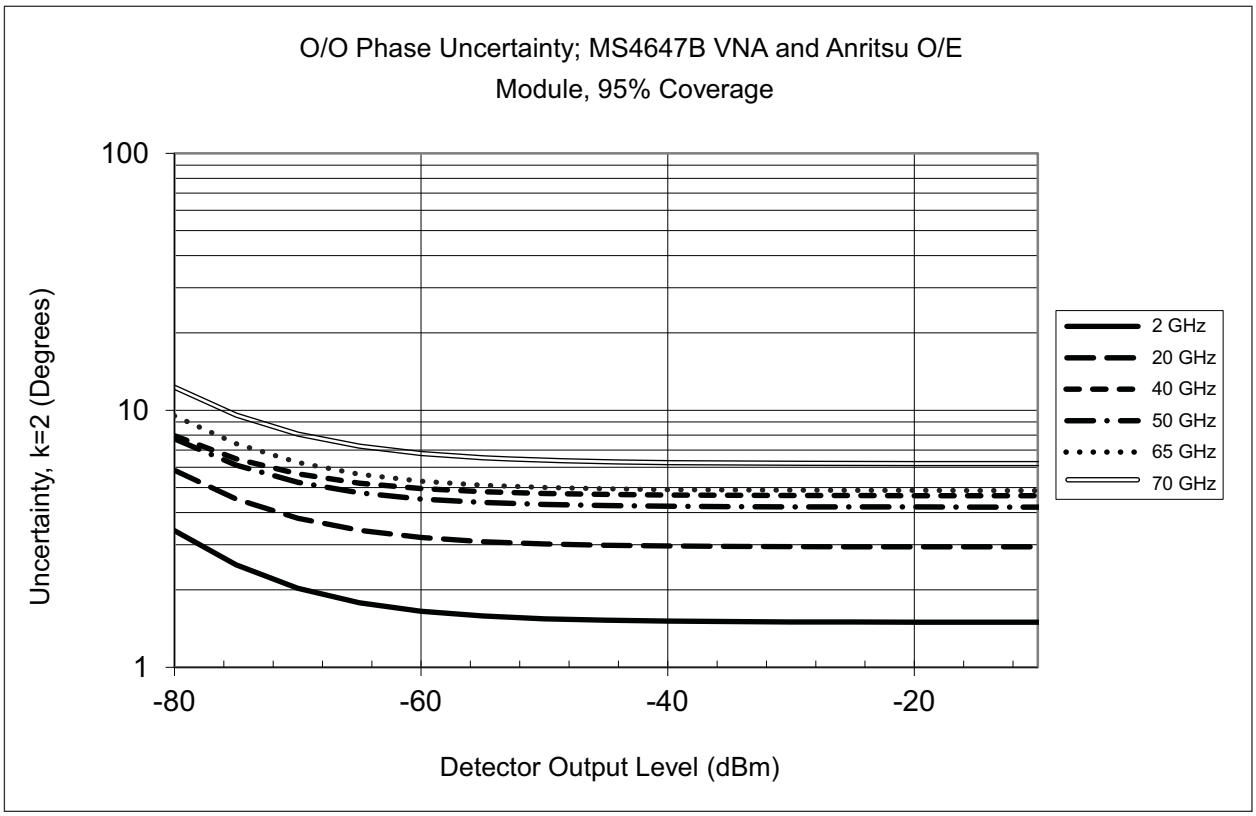


Figure 27-26. Example O/O Measurement Uncertainty Plots

Additional Comments on the Uncertainty Curves and Calculations

- Older MN4765As were characterized on a different VNA platform (37XXX) that had different base uncertainties particularly in terms of the noise floor at higher frequencies. The characterization data for those modules has that uncertainty built-in and hence, the optical converter measurements, even if performed on a MS464XX VNA will be somewhat elevated (by about 0.5 dB in E/O magnitude at –50 dBm and 65 GHz; less at lower frequencies and higher power levels). Uncertainties for 37XXX-based characterization and measurement are published in an older application note (e.g., [6]). The uncertainties for a MS464XX-measured, but 37XXX-characterized setup will be part way between the curves published in that older application note and the curves published here.
- MN4765As were characterized to 65 GHz. There are a variety of configurations for B and later revisions including those characterized to 70 or 110 GHz for 1550 nm applications and those characterized to 70 GHz or 110 GHz for both 1550 and 1310 nm wavelengths. Consult the technical data sheet for these models for more information on their uncertainty behaviors. In general, the trends will be similar to those discussed in the earlier example. The 110 GHz module usage is obviously of most benefit with the ME7838X broadband VNAs.
- While individual models were characterized with specific wavelengths in mind, typically the uncertainty degradation over neighborhoods of 50nm or more is very minor (less than a few tenths of a dB in magnitude). At lower RF frequencies at least, 1550-characterized devices can be used at 1310 nm with similar degradation although behaviors changes above ~ 30 GHz to 40 GHz. Of course, for devices characterized at multiple wavelengths (e.g., Options -72 and -43), this previous statement is not needed. Consult the factory for more information.
- Different uncertainties apply for different wavelength MN4765B models (and many of those models support multiple wavelengths and the uncertainties are different for each) in large part because of differences in how those modules are characterized and in part because of differences in RF performance.

- Errors due to polarization maintenance problems or multi-moded fiber (explicit multimode fiber or single-mode fiber being used in a wavelength range that multiple modes propagate) are not included in these uncertainties. It is assumed that appropriate optical components for the application and for the intended wavelength are being used.
- O/O uncertainties do not include optical mismatch, dispersion or other forms of optical distortion. These are only representations of uncertainties related to converter characterization, de-embedding and the RF calibration process.
- All of the uncertainties are relative in the sense they are not parameterized against optical power. The characterization optical power level for the MN4765X can be provided for one to calculate absolute uncertainties but this will require a characterization of the user's optical components. The ME7848A system does provide absolute uncertainties (see the appropriate technical data sheet), as the E/O and laser behaviors are known. An absolute responsivity uncertainty will include all of the above effects plus uncertainty in optical power and uncertainty in absolute responsivity of the paired converter (e.g., the O/E module when measuring an E/O device).

27-7 Summary

The E/O, O/E, and O/O measurement utilities have been discussed in this chapter. This utility is essentially an advanced de-embedding tool allowing one to remove the effects of one optical conversion device to learn the properties of the other. There are a large number of configuration choices, particularly in the 4-port cases, but the measurement principle is the same: use a good VNA calibration to characterize the converting pair and then de-embed the effects of a calibrated/characterized device to analyze the DUT. Signal-to-noise ratio is often a limiting uncertainty factor so it is important to carefully choose drive levels (as high as possible without being nonlinear), IF bandwidth, and averaging in order to optimize the measurements. The analysis of this chapter pertains mainly to the use of the MN4765X OE calibration module with the MS464XX VNA and, optionally, with the MN4775X E/O converter (which together all form a ME7848A optoelectronic measurement system).

27-8 References

1. D. F. Williams, P. D. Hale, T. S. Clement, and J. M. Morgan, "Calibrating electro-optic sampling systems," *2001 Int. Micr. Symp.* Dig., pp. 1527–1530, May 2001.

2. P. D. Hale and C. M. Wang, "Calibration Service For Evaluating and Expressing Optoelectronic Frequency Response at 1319 nm for combined photodiode/rf power sensor transfer standards," *NIST Special Publ. 250–251*, 1999.

3. D. Derickson, *Fiber optic test and measurement*, Prentice-Hall, pp. 252–263 and 621–638, 1998.

4. J. Hecht, *Understanding fiber optics*, Prentice-Hall, pp. 28–31, 1999.

5. "What is your measurement accuracy," *Anritsu Application Note 11310-00270*, 2001.

6. "E/O and O/E measurements with the 37300c series VNA," *Anritsu Application Note 11410-00311*, 2003.

Chapter 28 — IMDView™ (Option 44)

28-1 Chapter Overview

Intermodulation distortion (IMD) measurements are a powerful tool to assess low level (and sometimes high level) nonlinearities of a DUT. The origin of the measurement was a very practical one: when a modulated signal is passed through a nonlinearity, there is spectral broadening from some mixing products of components in the input signal and that broadening could result in the violation of the bandwidth of a communications channel (e.g., [1, 2, 3]). The use of the metric has since broadened considerably to represent a general linearity measurement but it does have its areas where caution might be required, both in terms of how the data should be interpreted and in terms of ways of setting up the measurement with reasonable uncertainties.

The purpose of this chapter is to explore the meaning of the measurement, how it can be setup using the MS464xB instrument using the IMDView™ application (Option 44), some suggestions on setup processes, and an uncertainty analysis. Most of the discussion will revolve around 2-port systems although IMD measurements can be made on 4-port systems (particularly the manual variants in the final section). Option 44 support for 4-port configurations is available starting with software version V2020.11.1.

28-2 Measurement Background

Intermodulation distortion measurements grew up from a need to understand the impacts of nonlinearities on spectral occupancy. While a simplification (with regard to communications signals), the use of two tones proved to be a reasonable proxy and, in most cases, is still actively used. While multiple tones and fully modulated signals are used in certain industries and certain cases, the two-tone IMD measurement remains very popular and powerful. Because of its history, certain conventions and nomenclature are used and it may be helpful to start by exploring some of these.

If one has two sinusoidal tones and one models the DUT linearity (or, rather, lack thereof) with a polynomial expansion, the signal would take the form ([1]-[3] and others):

$$K_0+K_1[A\cdot\cos(\omega_1 t)+B\cdot\cos(\omega_2 t)]+K_2[A\cdot\cos(\omega_1 t)+B\cdot\cos(\omega_2 t)]^2+\ldots+K_n[A\cdot\cos(\omega_1 t)+B\cdot\cos(\omega_2 t)]^n+\ldots$$

Equation 28-1.

Where ω_1 and ω_2 represent the angular frequencies of the two tones (separated by 2π*Delta). The term beginning with K_2 will generate a DC value, signals at the second harmonic of each of the two tone frequencies, and signals at the sum and difference frequencies (these are termed the 2nd order intermodulation products; since Delta is usually small, these signals will be at low frequency and near the 2nd harmonics). The term beginning with K_3 will generate signals at frequencies:

ω_1	$2\omega_1+\omega_2$
ω_2	$2\omega_2+\omega_1$
$3\omega_1$	$2\omega_1-\omega_2$
$3\omega_2$	$2\omega_2-\omega_1$

Of particular note are the last two frequencies. Since ω_1 and ω_2 are relatively close together, these new frequencies will be near the fundamental tone frequencies but offset by Delta (2π*Delta in angular frequency space). Skipping ahead, for the 7th order term, there will be frequencies at:

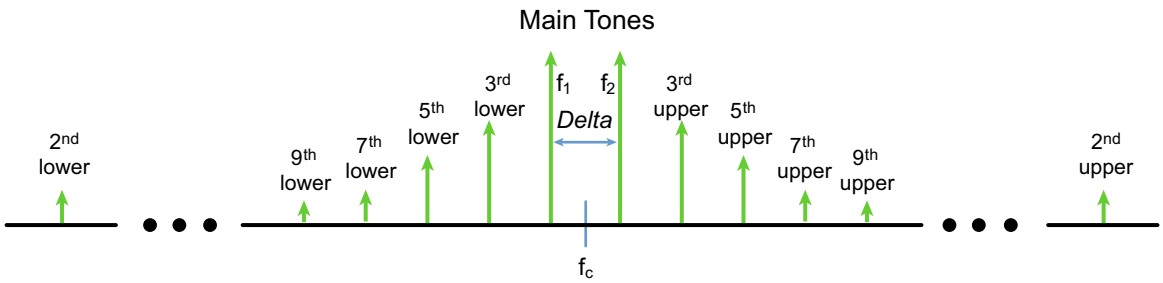
$$4\omega_1 - 3\omega_2$$

$$4\omega_2 - 3\omega_1$$

These frequencies will be near the fundamentals but offset by 3*Delta (6π*Delta in angular frequency space). Even order products above 2 are generally not analyzed since the difference products are further away from DC (and the 2nd order difference product can cause issue for homodyne systems where a zero IF is used) and the harmonics are often not of as much interest. The diagram in Figure 28-1 on page 28-2 shows some of the more interesting products. Here f_c denotes the center frequency and the main tones are at:

$f_1 = f_c - Delta/2$ ($\omega_1 = 2\pi f_1$)

$f_2 = f_c + Delta/2$ ($\omega_2 = 2\pi f_2$)

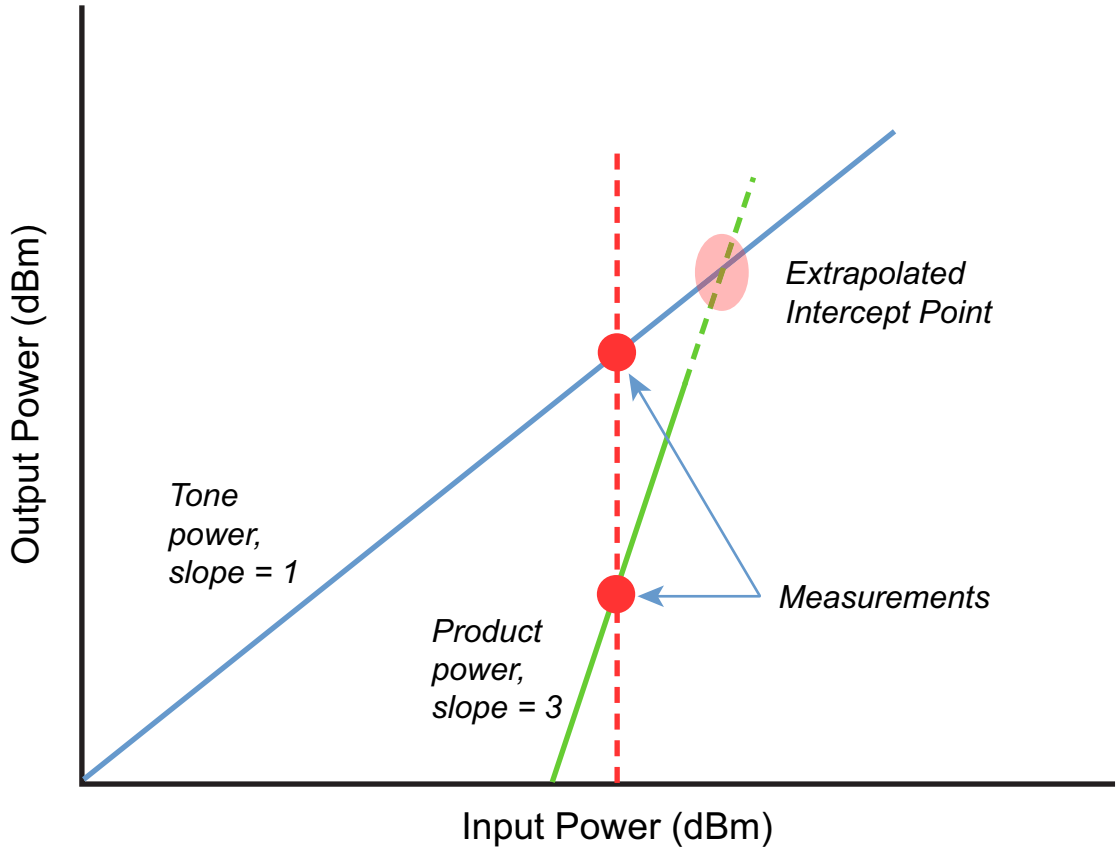


Some of the intermodulation products commonly of interest are shown here. The most critical odd-order products are clustered near the main input tone frequencies and the 2nd order products of most interest are usually at low frequency and near the 2nd harmonics.

Figure 28-1. Intermodulation Products

Sometimes the absolute tone powers are of interest and sometimes the intermodulation product power relative to the main tone power may be specified (and this may be in terms of an average of upper and lower, in terms of maximum or minimum….). The asymmetry between upper and lower intermodulation products may also be specified. One of the more subtle variations is a quantity called intercept point. This is a constructed quantity based on assumption of the power dependence of output main tone power on input power being a slope of 1 (not a bad assumption until the DUT compresses) and the power dependence of the product amplitude having a slope (relative to input power) of N for an Nth order product (often not a reasonable assumption). If these slopes are assumed and a measurement is made at one input power point, one can then theoretically compute at what power level the product and main tone amplitudes would be equal. This intersection is called the intercept point (and if it is in terms of output power, it is called output-referred and if in terms of input power, it is called input-referred). Now, this intersection is often not physically realizable since the DUT may compress or the slopes may otherwise change before it can be reached but it is a common metric. As long as the power at which the measurements were made is specified, it is a reproducible metric of some part of the DUT's nonlinearity. These concepts are illustrated in Figure 28-2 on page 28-3 for the 3rd order case.

The point about the importance of the power at which the IPn point is computed discussed in the previous paragraph arises from the often ill-behaved power profile of the intermodulation products. In Figure 28-3 is an example of a relatively well-behaved case where the main tone (right plot) shows some signs of gain expansion and the 3rd order product has many different slopes. Choosing different measurement points for computing IP3 in this case can result in answers differing by more than 10 dB. Certain higher efficiency amplifiers are even more unusual and can be non-monotonic.



The idealized relationships leading up to an intercept point computation are illustrated here for the case of IP3.

Figure 28-2. Intercept Point Computation Example for Case of IP3

Shown below in Figure 28-3, the 3rd order IMD product (left plot) and the main tone power (right plot) versus input power are plotted here for a relatively well-behaved amplifier. The 3rd order product slope is only occasionally near 3 and the even main tone power slope deviates from unity.

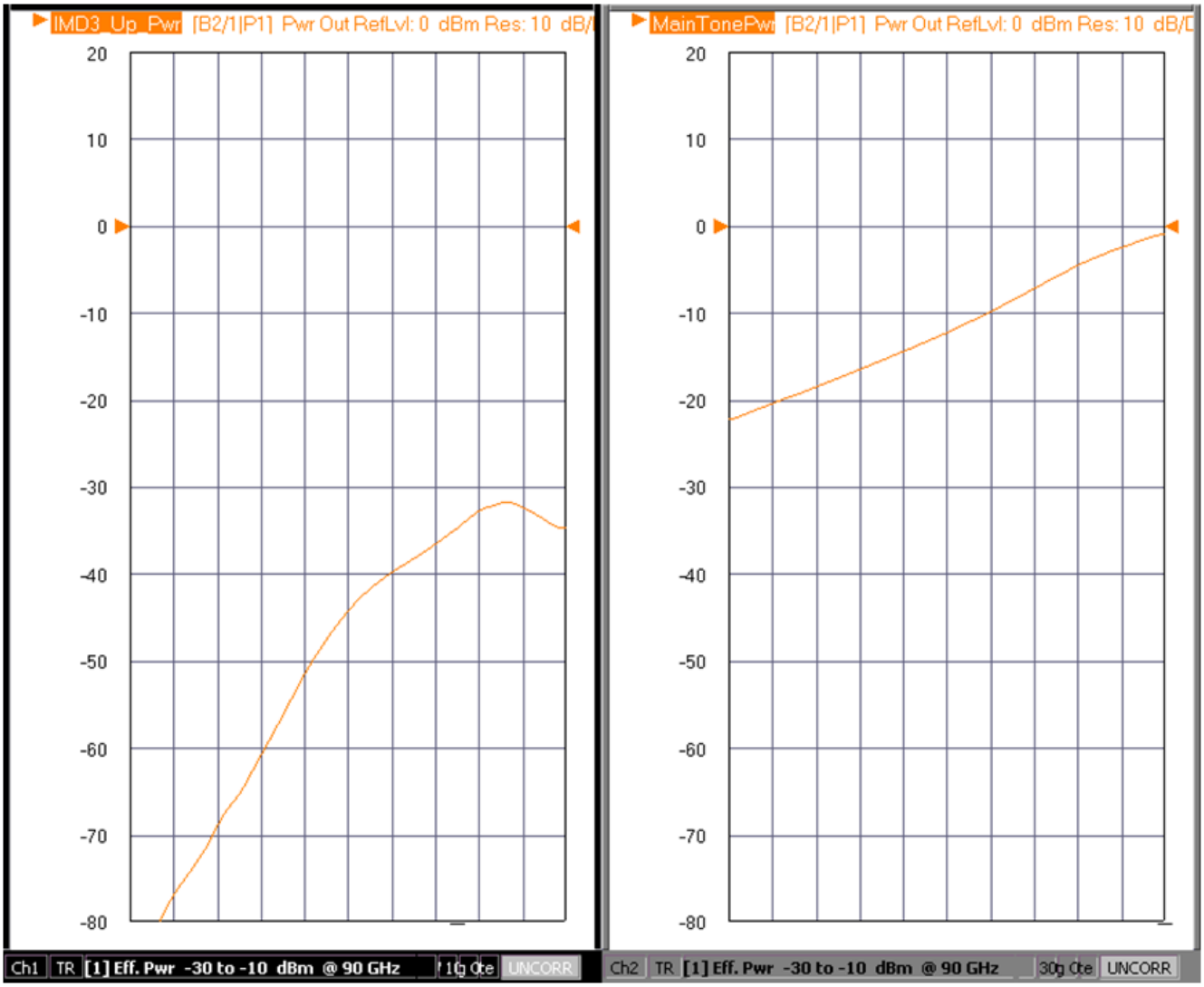


Figure 28-3. 3rd Order IMD Product (left) vs Main Tone Power (right)

Definitions of IMD Measurement Terms

Below are definitions of some terms used in discussing IMD.

f_c (Center frequency):

In the IMD mode, it refers to the frequency half-way between the two main tones.

Delta:

The spacing between the main tones.

In other literature, it is sometimes called the 'offset' or 'tone spacing'.

IPn:

Nth order intercept point.

This is usually a calculated point in power where the main tone power and Nth order IMD product are equal based on linear (of dB quantities) extrapolation. This is an absolute-power-like quantity so power calibrations take on additional significance.

OIPn/IIPn:

Output or input referred Nth order intercept point (respectively).

The calculation of intercept point can be done at the DUT output or input planes with the difference being the DUT gain. Generally the output referred value is more directly found.

Asym:

Product asymmetry or the difference in amplitude of the upper vs. the lower product of a given order.

This can be a useful measure of bias network or thermal interactions in the DUT and can also be a hallmark of some system nonlinearities (source contamination or certain IF network issues).

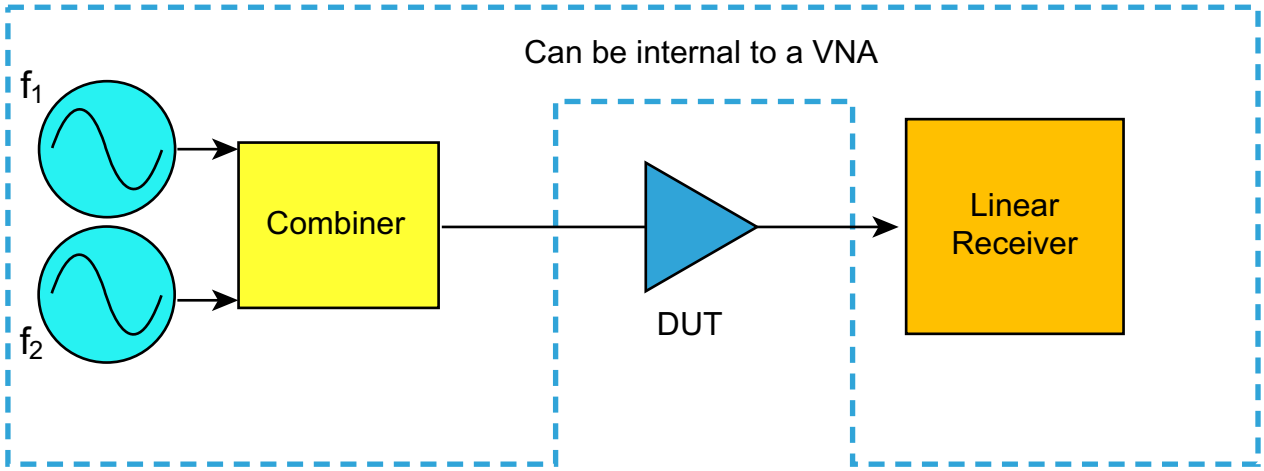
Order:

The level of nonlinearity that is being observed in a given product.

The most interesting odd order products are localized around the main tones at one Delta from a main tone (for 3^{rd} order) to 4 Deltas from a main tone (9^{th} order). Second order nonlinearities produce direct mixing products at the sum and difference of the main tones.

28-3 Typical Measurement Configurations And Considerations

As two tones, combined into one drive, and a linear receiver form the basis of the measurement, there are many possible configurations. Classically, two separate synthesizers, a coupler/combiner, and a spectrum analyzer have been used. A VNA consisting of two internal sources (or using one or more external sources), some combining structure and the VNA's own receiver can be used as well and this structure can be helpful in performing S-parameter measurements at the same time. Many different measurement structures for IMD are possible. The key elements are shown in Figure 28-4.



Many different measurement structures for IMD are possible. The key elements are shown here.

Figure 28-4. IMD Measurement Structure Key Elements

Performance Attributes

As will be seen in later sections, some of the important performance attributes include the following:

Receiver Linearity

If the receiver generates its own IMD products, these will convolve with the products generated by the DUT and make results difficult to interpret.

Source Contamination

If the sources are remodulating each other, unintended intermodulation products can actually be generated at the input plane and are injected into the DUT where they can linearly pass through or mix with the DUT's own products. In either case, interpretation of the results becomes more complex. In practice, this effect often occurs because of an output nonlinearity of a source system (the final power amplifier for example) or an interaction of the leveling system of the source. In either case, a tone from one source leaks over to the other source and there is some nonlinear combination occurring (e.g., [1]). The system engineer can help by minimizing these nonlinear effects. On the setup side, high isolation combiners are helpful. In some very high IPn scenarios, isolators or isolation amplifiers are sometimes placed in the source chains.

Another aspect of contamination is that of spurious signals (whether they are on the source signals themselves or there are spurious receiver responses that create data fluctuations that look like it). A relevant point is how well the hardware can avoid problem areas by readjusting internal synthesis stages or if parameter modification is required.

Power Accuracy

This has two different aspects:

a. Often the DUT IMD products will be extremely power sensitive so an inaccurate drive level will put the DUT into an unintended state and

b. Some expressions of intermodulation distortion have absolute power dependencies (intercept point and IMD (dBm) among them) so power inaccuracies will translate directly to the result. The latter effect happens in part since the drive power accuracy is coupled to the receiver calibration. A tertiary effect can occur with these calibrations if interpolation is invoked by changing the frequency list after calibration; these interpolation effects are most severe for highly mismatched test structures and a low density of calibration points.

Mismatch

Aside from simple issues of how much power is actually delivered to the DUT (reduced by mismatch at the input plane) and inaccuracies induced in the receiver calibration (from mismatch between source and receiver), there can be issues of different mismatch levels at a product frequency versus that at a main tone frequency. This is mainly an issue for large tone Deltas and longer setup electrical lengths (which follows from this situation allowing a major standing wave change on the scale of a tone Delta). A well-matched combiner and receiver structure can help with this effect.

There are a number of different dependencies that may be of interest in the IMD class. Obvious ones include bias and tone power. Since this is a quasi-linear measurement (usually), these dependencies will often be far more severe than for S-parameter measurements. Another obvious dependency is that with frequency (of the main tone average). Less obvious perhaps is the dependence on the separation between tones. Some interesting behaviors can be seen where the tone difference frequency can excite bias, thermal and trapping behaviors causing and intermodulation response at other orders. These effects often show up in intermodulation product asymmetry (e.g., [4]-[5]) and this measurement will be discussed further in later sections of this chapter.

28-4 IMDView Measurement—Hardware

There are a wide variety of possible configurations for the IMD measurement when using the MS464xB depending on measurement requirements, installed options on the MS464xB, and available other equipment. Some considerations:

Option 31

Installs a second source so both tones can be generated within the VNA. The signals can come out the main test ports (or source loops if Options 51/61/62 are installed) and be combined externally before driving the DUT or they can be combined internally if Option 32 is also installed.

Option 32 (requires Option 31)

Internal combiner and switching. This option installs an internal combiner (with >20 dB of isolation) over the frequency range of the instrument. Because of the switching, S-parameter measurements can be made without disconnecting the DUT.

External Sources

If Option 31 is not installed, at least one external source is required (although 1 or 2 external sources can be used with any options configuration).

The use of external sources is covered in some detail in Chapter 18 of this measurement guide but some of the information will be repeated here for convenience. Up to four external sources can be controlled and each must have a unique GPIB address. The external source number and the GPIB address are connected on the System/Remote Interface/Ext. Sources menu that is shown in Figure 28-5.

The GPIB bus is polled for these addresses upon entering the IMDView configuration dialog (as well as on entering the multiple source setup screen, on start-up and on certain other occasions associated with mm-Wave operation) and those synthesizers that are 'found' will be activated on the tone selection pull-downs. Note that the synthesizers should be set to RF ON prior to use with the VNA; otherwise the sensing and activation may not occur properly.

If a synthesizer is disconnected from the bus after it has been selected, bus communication errors and sluggish VNA response is possible. Going to all internal sources or turning off IMDView until the source can be reconnected may help in regaining communications. In extreme cases, restarting the synthesizer and/or the MS464xB may be required. Another side note: for reliable measurements, the 10 MHz references of the synthesizers and the VNA should be connected (both are auto-sensing) and any of them can be used as the parent, or they can be connected in a daisy-chain fashion (although there is some noise penalty in doing so). If many instruments are connected in a star-fashion to a single 10 MHz out, the signal level may fall too low so it is important to monitor the 'Ext Ref' annunciators on all instruments connected as 10 MHz recipients.

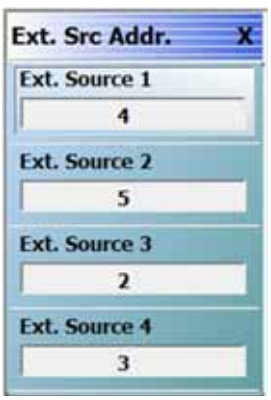
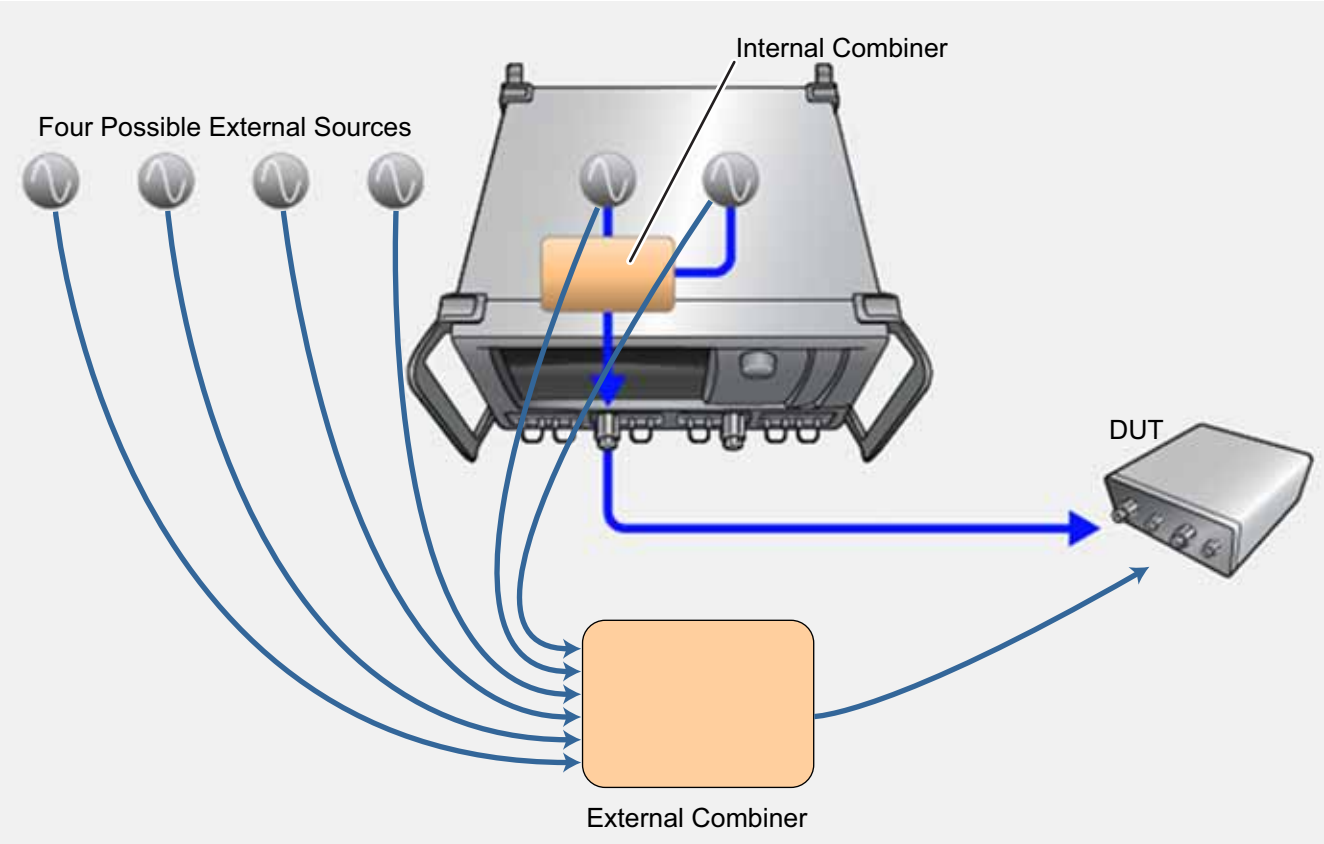


Figure 28-5. External Source Address Menu (1 of 2)

The GPIB addresses of the various numbered external synthesizers are set on the menu shown above. None, some, or all of the synthesizers listed may be connected but, if they are to be used, the address here must match that on the synthesizer and the identifier on this menu will be used elsewhere in the VectorStar application.

Figure 28-5. External Source Address Menu (2 of 2)

Most of the possible ways of applying the two tone stimulus to the DUT are illustrated in Figure 28-6. Some of the variations not shown include using the internal sources routed to an external combiner using the source loop outputs (when those loops are in the instrument, Options 51/61/62) instead of the ports. Such a choice may be useful to increase power delivery or if it is desired to save a port for access to the receivers.

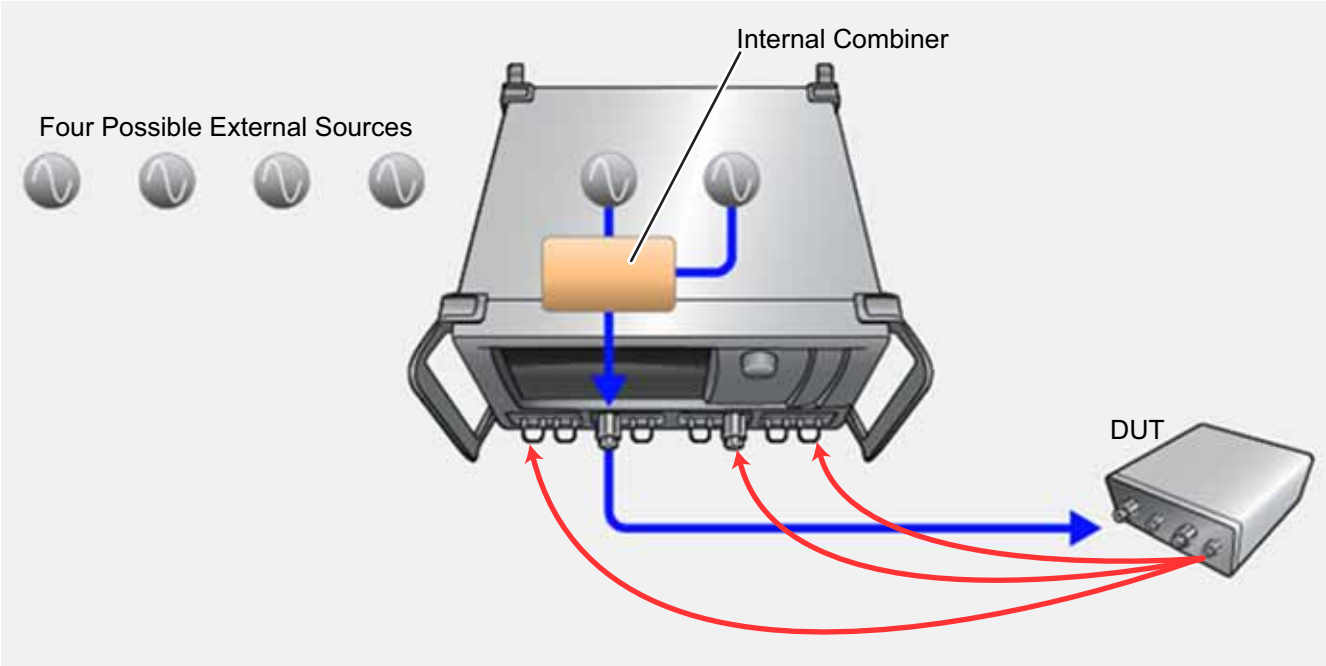


Most of the possible sourcing combinations (not all of which may be available based on the number of external synthesizers available or the options on the MS464xB) are illustrated here. Only two source paths are definable per channel although more are shown here.

Figure 28-6. Possible Ways to Apply Two Tone Stimulus

A few additional comments are warranted on the use of the internal combiner. As suggested in Figure 28-6 (darker blue line paths), the combiner routes the source 2 signal out to port 1 (in combination with the signal from source 1). Because of the loss of the combiner, the available power for the signal from source 2 will be about 5 dB to 10 dB lower than what is available from the port (depending on frequency). A factory ALC calibration will handle the corrections when the internal combiner is used so the power delivered will be reasonably accurate. Also, user power calibrations are available for any of those routings for improved power accuracy at the final DUT plane. These are discussed in more detail in a later section.

Moving to the receiver side of the measurement, there are also a number of locations that can be used for the DUT output. If Option 32 is available and the internal combiner is being used, port 2 is a logical choice as this facilitates single connection measurements. The b2 or b1 access loops can also be used as suggested in Figure 28-7 (does require one of the loop options). If two external synthesizers are used, then there are four choices: port 1, port 2, b1, and b2 (and more on a 4-port system). For other combinations of sources, there will be a different, but fairly obvious set of receiver access points. From the instruments point-of-view, it does not matter if a port or the b_i access point is used but it is up to the user to make sure the receiver calibration and normalization calibration (if needed, more on that later) are done to the appropriate reference plane. The MS464xB does need to know which 'side' is being used (port 1 side vs. port 2 side) so the data from the appropriate internal receiver path is used in the computations.



The possible receiver paths (red arrows denote possible output paths from the DUT) that could be used here are shown above for one source configuration. The choices available depend on the instrument options and the sourcing choice.

Figure 28-7. Possible Receiver Paths

When using an external combiner, there are of course many choices satisfying the desired bandwidth.

Wilkinson Combiners

These reactive combiners have ~3 to 4 dB insertion loss per arm and can have bandwidths exceeding 4 octaves. Broadband isolation may be on the order of 15 to 20 dB and narrowband isolation levels can exceed 25 to 30 dB.

Resistive Splitters

The available bandwidths can exceed 70 GHz (and many decades) but insertion losses can exceed 6 dB. While a number of terms are in use (splitter vs. divider for example), the basic structures are 2- and 3-resistor versions. 3-resistor versions are quite match-invariant, but only have ~6 dB of isolation. 2-resistor versions can be quite sensitive to match, but can have 10 to 12 dB of isolation.

Directional couplers

While asymmetric, directional couplers can cover 70 GHz of bandwidth with isolations exceeding 20 dB. The insertion loss asymmetry requires some additional care on power calibrations and ensuring that target powers can be met (particularly if the coupling is weak).

There are occasions when more isolation is desired (particularly for very low IMD levels; generally not before the –80 dBc level with typical setups) to reduce source interaction. Classically this can be done by inserting an isolation device (an isolator, a pad, an isolation amplifier, or any combination of these) in each arm of the setup as suggested in Figure 28-8. The use of isolation devices in the setup to reduce source-to-source contamination is illustrated here. As isolators only cover (usually) an octave of frequency and may not always be available for the band of interest, broadband amplifiers plus attenuators are often used in practice. The power capability of the isolation component must be taken into account. Although it is not often a problem, the harmonic generation by the isolation device can sometimes affect the measurement, particularly if the DUT's operating state is sensitive to harmonic input. Thus it is usually recommended to keep any isolation amplifiers well away from compression.

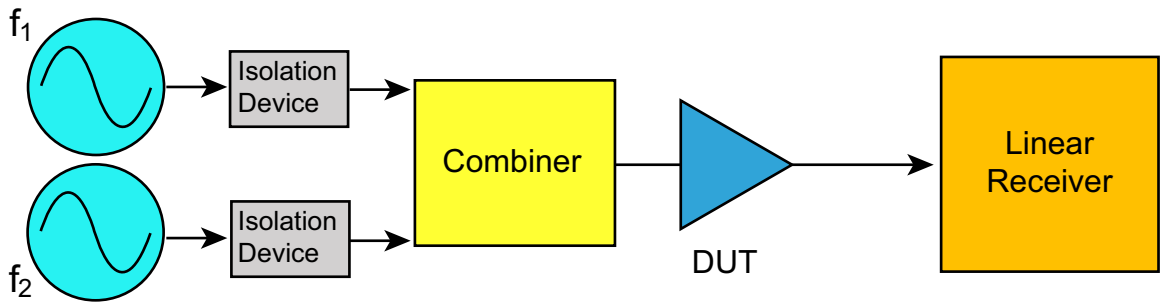


Figure 28-8. IMD Measurement Structure with Isolation to Reduce Source to Source Contamination

If using the Option 32 internal combiner, the source loops on the VNA can be used to insert additional isolation devices or to insert amplifiers just for more signal power. The source-side step attenuators (if so equipped) also precede the combiner so they can be used to increase isolation as well.

With a 4-port system, many of the configurations are similar. Some notes:

- In single-source systems, the choices for stimuli are the same as for the 2-port single-source systems.
- In dual source 4-port systems, the internal combiner is not available. Both internal sources can be used as stimuli from the ports. Because of internal switching, allowed driving port pairs are 1-3, 1-4, 2-3, and 2-4 (1-2 and 3-4 are not allowed). Of course, external synthesizers can be used.
- Any port can act as the receiver.

28-5 IMDView™ Measurement—User Interface

The IMD measurement is enabled from the application menu but is not an orthogonal application (like Transmit/Reflect (i.e., S-parameters), Pulse, and Noise figure are). As such, many of the setup parameters (like frequency, power, IFBW…) carry right into the IMD setup without having to re-enter them. Certain variables are unique to IMD (such as Delta, IFBW during product measurement, the response parameters…) and hence will have to be entered.

IMDView Menu

The IMDView menu allows one to quickly change certain parameters while still observing the data. Changing the power levels or Delta, for example, could allow one to quickly spot the response of DUT to that kind of stimulus change. The menu (when in the swept frequency mode, to be discussed) is shown in Figure 28-9. The main frequency and power entries are just replicated fields from the frequency and power menus. Values can be entered in either location and they will copy through.

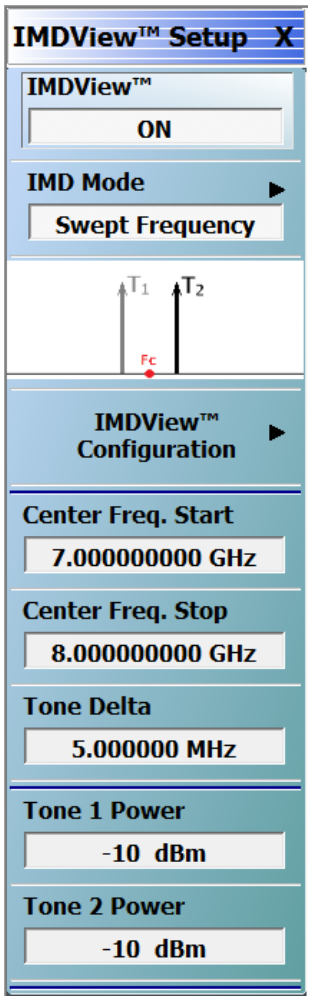


Figure 28-9. IMDView Setup Menu (When in Swept Frequency Mode)

IMDView Dialog

Most of the setup tasks are handled in the IMDView configuration dialog that is pictured in Figure 28-10. Some of the fields here are the same as in S-parameter measurements (frequency range, number of points, power level setting fields and attenuator settings) but others are new and will be discussed in some detail in this chapter.

Some of the generic items are:

Receiver Port

This is the port where the DUT output will be connected (by default, port 2). Note that the test receiver loop can be used in lieu of the actual port connector (on systems with any of Option 51, 61, or 62) for this connection and that may be desirable for certain low output power situations or for other setup reasons. In 4-port versions, four choices will be available.

Tone Selections

This is where a synthesizer is assigned for each tone. In Figure 28-10, the two internal sources and internal combiner were selected as an example (Option 32 required). Instead of 'Int Src 2->Port1', one could have selected 'Int Src 2->Port 2' (Figure 28-11 on page 28-14) in which case the source 2 signal would have routed to the port connector and an external combiner would be required (practically, note that either a source loop or a receiver loop access would normally be required to make the measurement in this case). Any of four external synthesizers can also be selected as tone sources if they are connected.

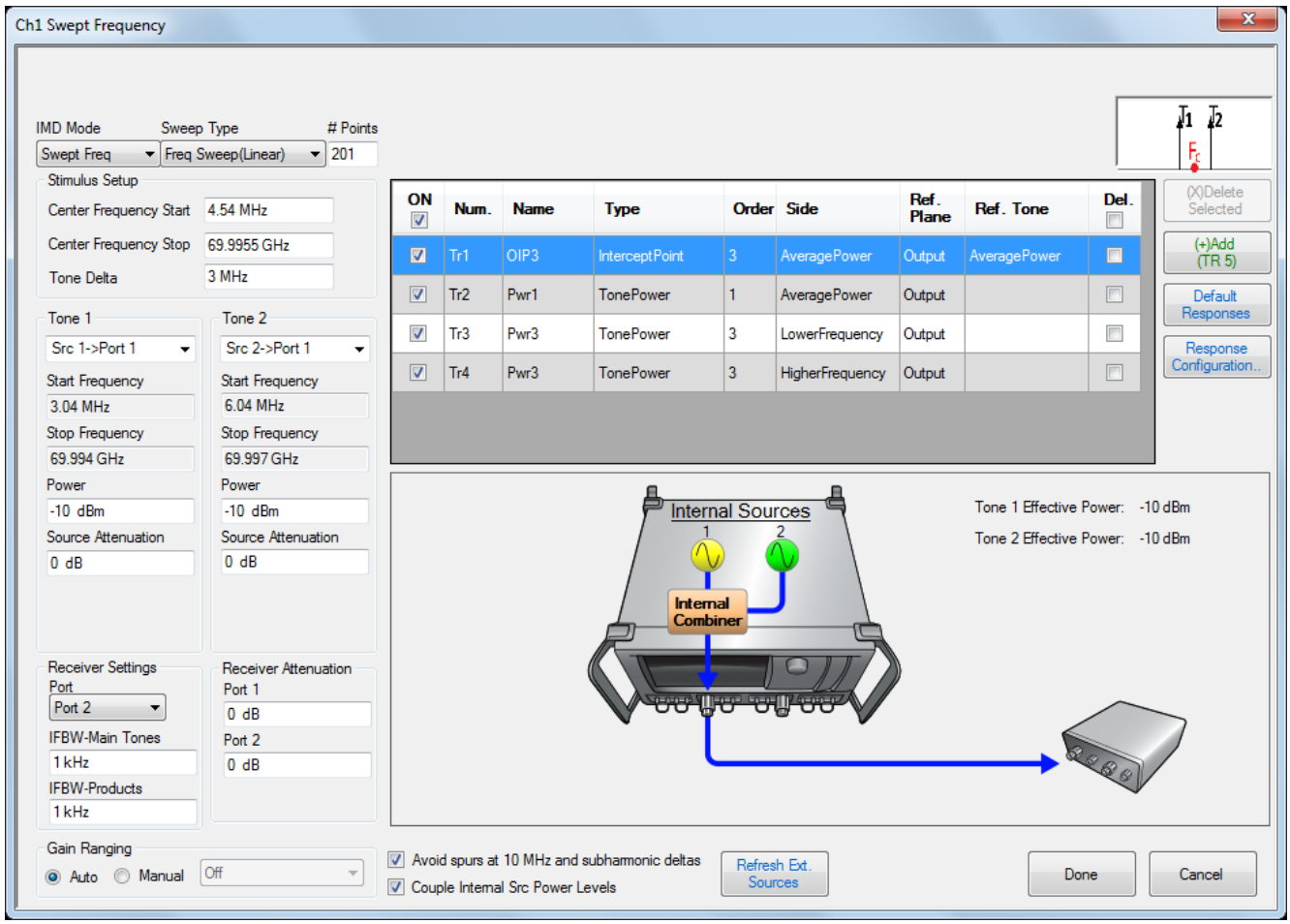


Figure 28-10. IMDView Configuration Dialog Using Option 32 (Internal Combiner)

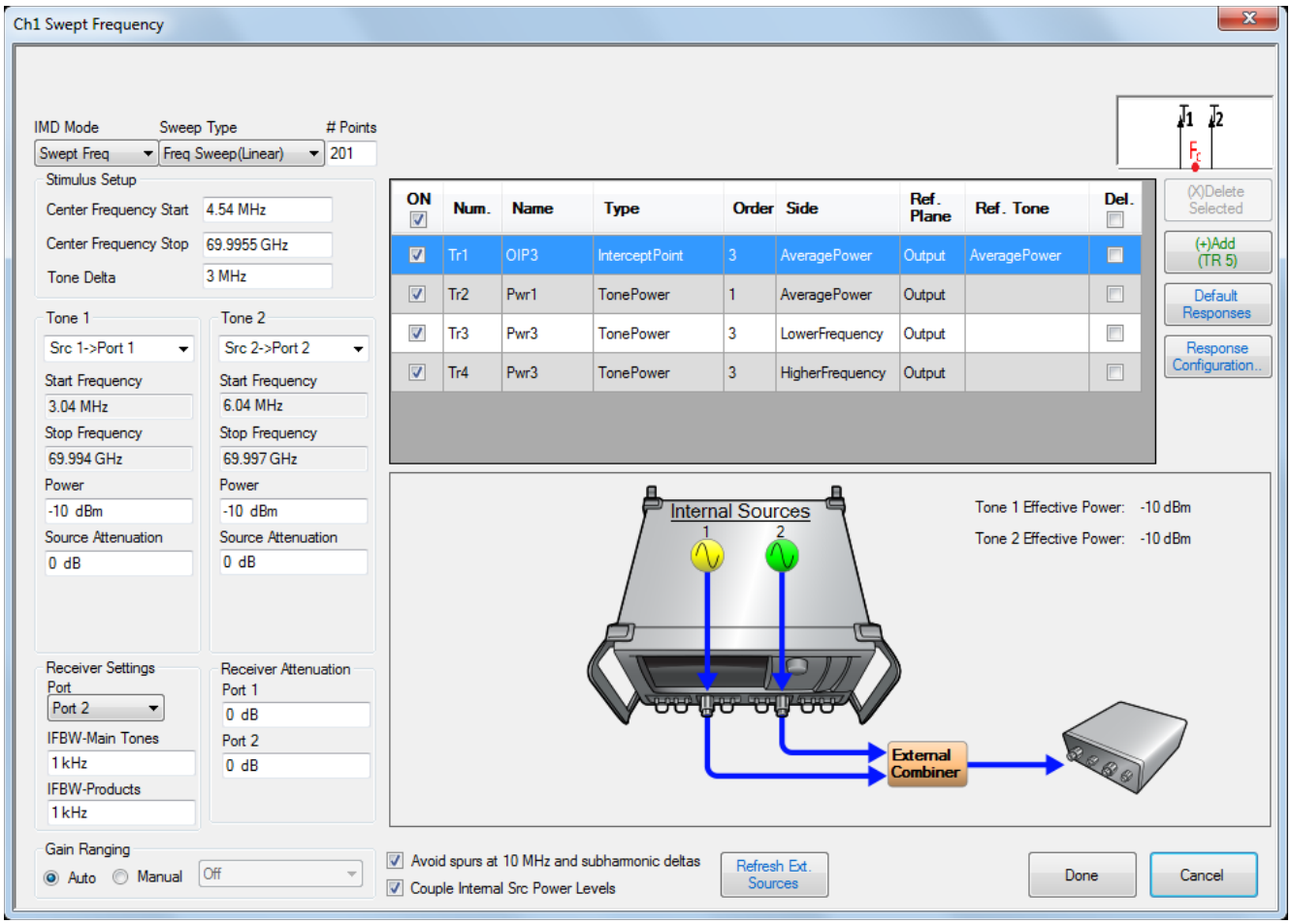


Figure 28-11. IMDView Configuration Dialog Using External Combiner

For 4-port systems, the dialog is similar and the measurement choices are the same. Figure 28-12 shows a dual source system using ports 1 and 3 to drive an external combiner.

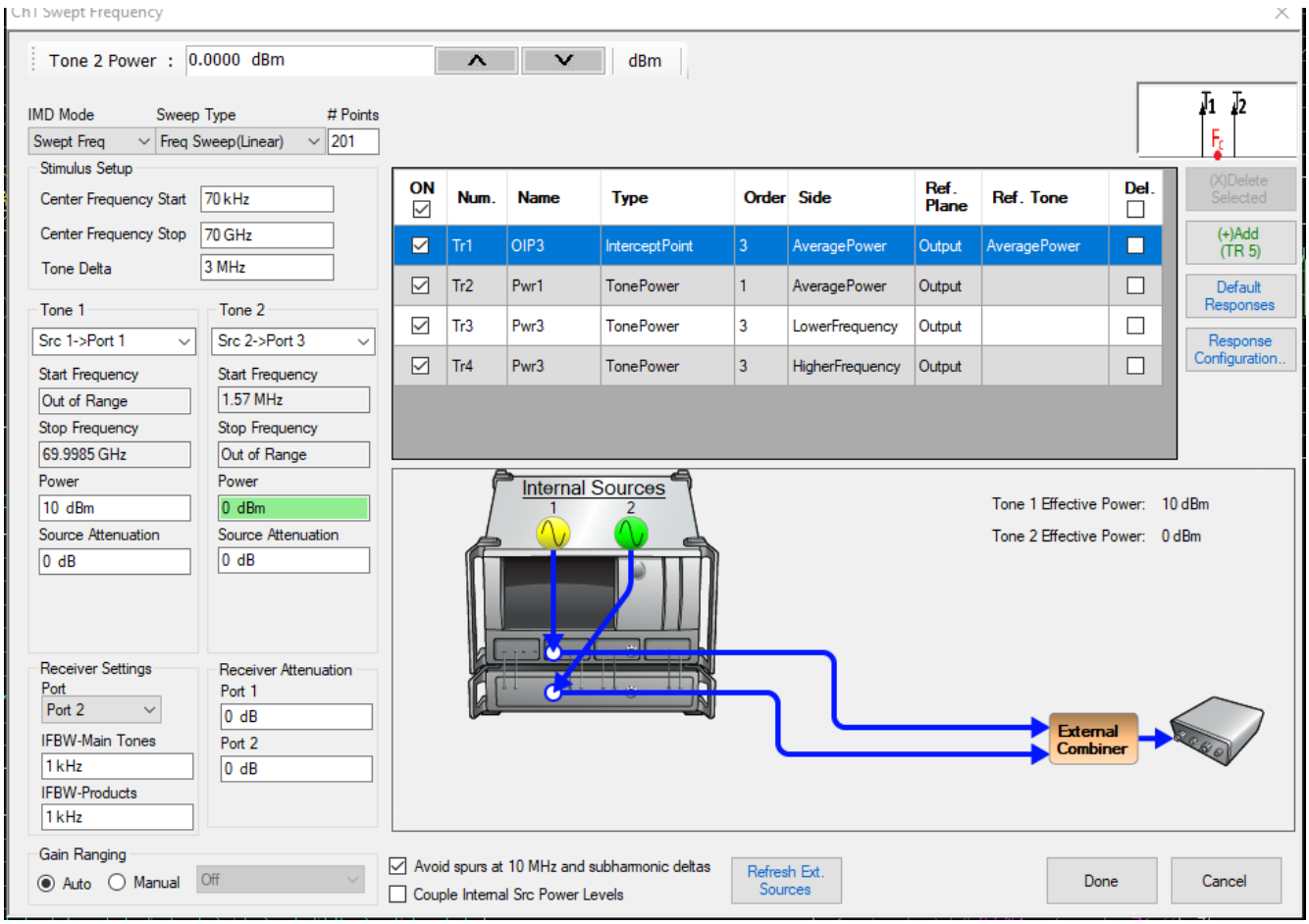


Figure 28-12. A Setup Using Two Internal Sources (From Ports 1 and 3) as the Tones (4-Port System)

A setup using two internal sources (from ports 1 and 3) as the tones (4-port system) is shown in Figure 28-13.

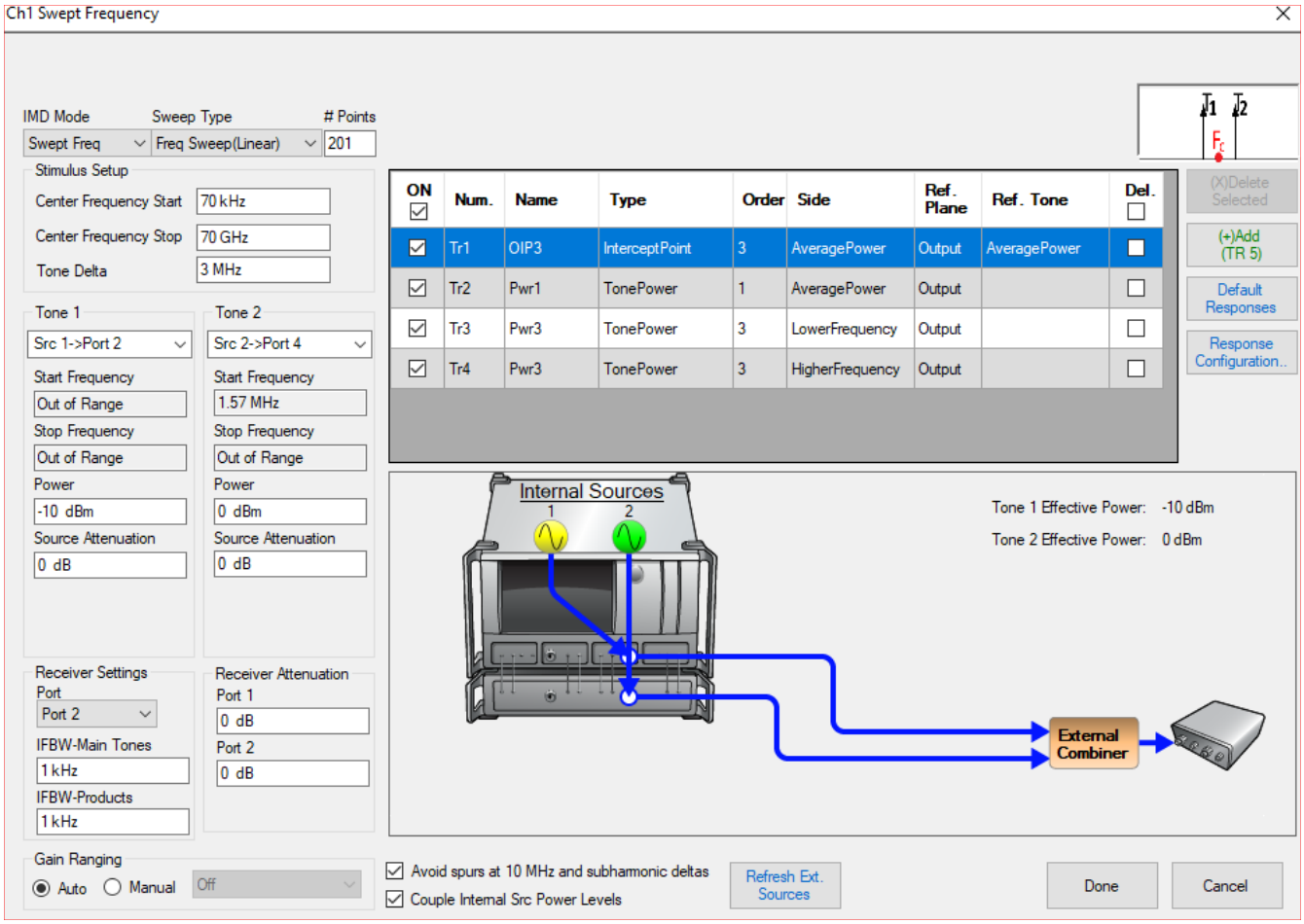
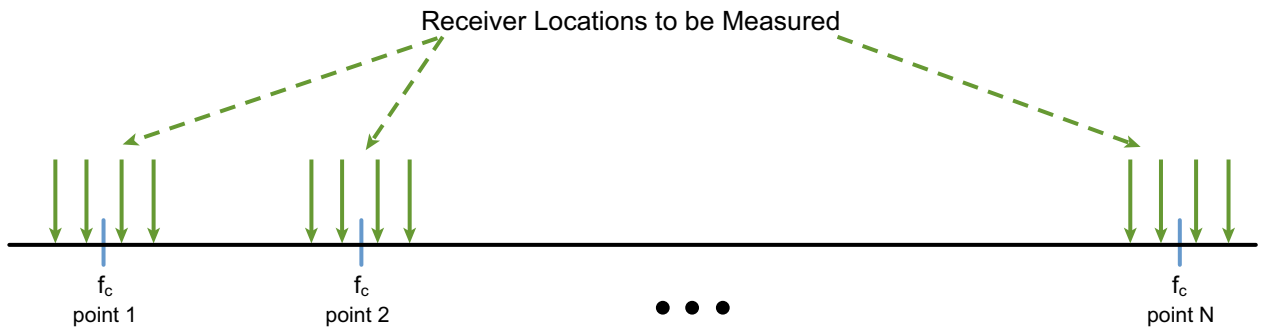


Figure 28-13. Two Internal Sources 4-Port Using Two Internal Tones

28-6 IMDView Application—IMD Modes

Swept Frequency IMD Mode

The swept frequency IMD mode causes the center frequency f_c to sweep over the indicated range meaning the two tones from the sources are also sweeping. At each frequency 'point', the sources are stationary and the receiver will scan the tones and products as desired. As an example, consider the case of N frequency points and we wish to measure both main tones and both third order products. The sweep concept is illustrated in Figure 28-14. At each of the N center frequency points, the main tones are positioned and the receiver measures one of four locations. On the next sweep, another of those receiver locations per f_c point is measured and so on. In this case, four sweeps would be required. In certain modes, the sweep combination would be different (the receiver would sweep for a fixed f_c and then the system moves to the next f_c).



The receiver measurement locations are illustrated above for an N point swept frequency measurement where both main tones and both third order products had to be measured. If additional product measurements were required, there would be additional receiver locations shown.

Figure 28-14. Receiver Measurement Locations—Swept Frequency IMD Mode

The receiver behavior requires some additional explanation as it will affect measurement time. Depending on which orders are requested by the user (through the trace configurations), different product locations will be measured. Depending on how the measurements are configured (see Section 28-7), one or both of the products of a given order will have to be measured. Also, for the referenced measurements (relative IMD (dBc) and IPn), one or both of the main tones will also have to be measured. The number of sweeps will be revisited after all of the possible configurations have been discussed.

The range of allowed f_c (and hence the start and stop frequencies on the frequency menu) will be dependent on Delta and the measurements requested. This is illustrated in Table 28-1. Since the required measurements will exceed the highest used f_c by some multiple of Delta/2 (and will be below the lowest-used f_c by some multiple of Delta/2), the receiver could conceivably be asked to measure outside the range of the instrument. Thus one will see new limits and frequency imposed on the menu entries to ensure the system can make the measurement. Table 28-1 shows the receiver frequency range that is needed depending on f_c, Delta, and the intermodulation product orders requested (via the response types). These limiting frequencies must be within the range of the instrument and parameters will be coerced to make measurement possible (generally f_c except for the lower 2nd order case where Delta will be limited).

Table 28-1. f_c Range Dependencies

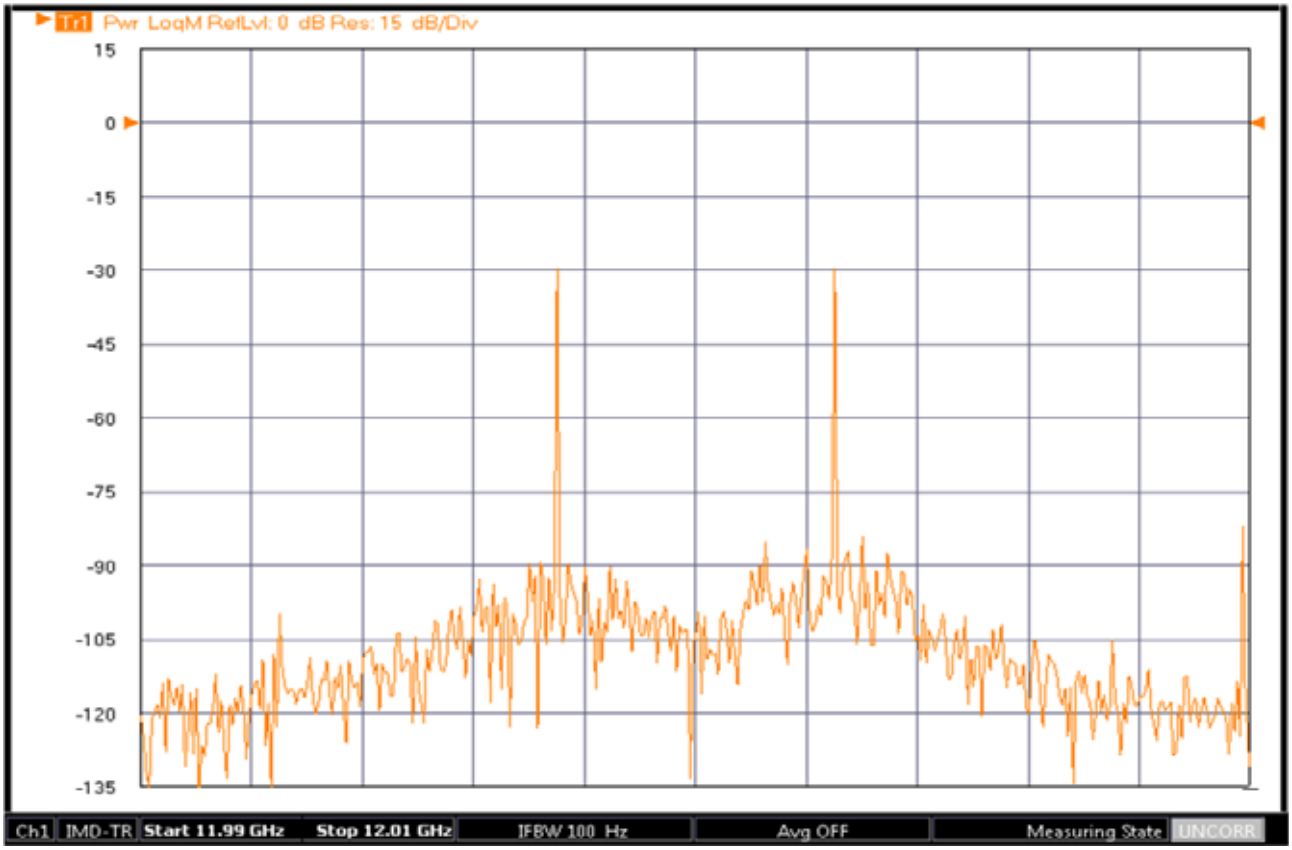
Receiver Frequency Range Element	Frequency Limit if 2nd Order Included	Frequency Limit Without 2nd Order (N=maximum odd order requested)
Min frequency needed	*Delta*	f_c – *N*Delta/2*
Max frequency needed	$2f_c$	f_c + *N*Delta/2*

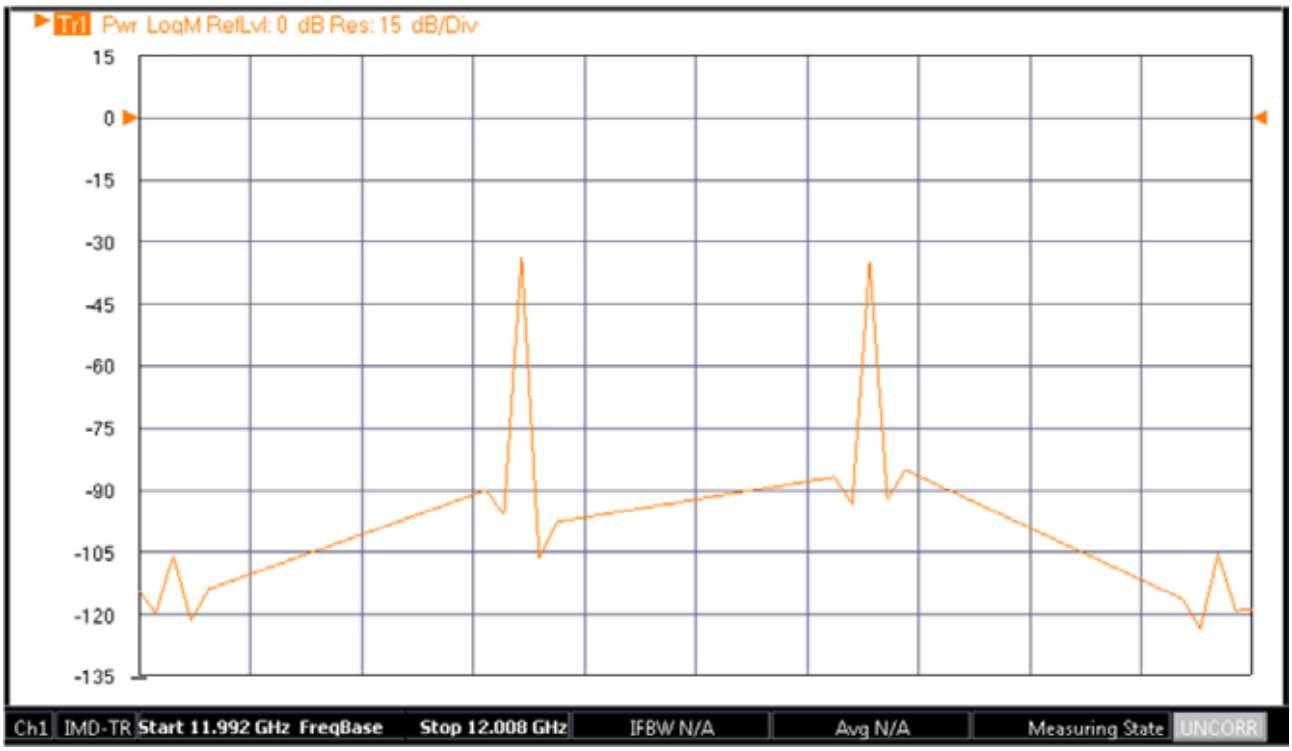
As an example, suppose only a measurement of a lower 3rd order product is requested and the Delta is 20 MHz. In this case, the one receiver location measured is 30 MHz below f_c so, for an instrument equipped with Option 70 (low frequency coverage), the minimum f_c is 30.07 MHz (40 MHz without Option 70). As another example, consider the case of the only requested measurement being the upper 5th order product and the instrument is a MS4647B (70 GHz upper frequency limit) and the Delta is 6 MHz. Since the measurement must be at 2.5 Deltas above f_c, the maximum f_c is 69.985 GHz.

The measurement of 2nd order products is something of a special case since the measurements occur at Delta (difference of the main tones) and 2*f_c (sum of the main tones). Thus for lower 2nd order coverage, the minimum Delta is 70 kHz (with Option 70). For the upper 2nd order product, the maximum f_c is 35 GHz (for a MS4647B).

The SpectrumView measurement mode is somewhat simpler in that the main tones are fixed and the receiver simply sweeps over some specified range to look at a maximum product order (with the special case of 2nd order products handled a bit differently). The measurement protocol is then something like what one would see when using a spectrum analyzer. There are two different sweep modes: linear and segmented. The linear sweep uses equally spaced receiver frequencies with the spacing chosen so that all requested odd-order products will be measured. If the IFBW is low, the sweep time can be fairly long but some additional DUT spurs can sometimes be detected. Note that the number of frequency points is set to a useful default number when leaving the IMD configuration dialog but it can be changed afterwards. Note that arbitrary changes in the frequency list can result in some products not being at measurement frequency points and errors can occur.

The segmented sweep version only measures frequencies in small neighborhoods of the tones and products. A segmented sweep table is generated with these small neighborhoods (5 points with a local spacing of Delta/20) when leaving the IMD configuration dialog. These segments can be edited manually after leaving the dialog but, again, some care is required to ensure that the desired product frequencies are in the list. Since far fewer points are measured, the sweep time can be much faster for a given IFBW. In an example measurement covering only 3rd order products with an IFBW of 100 Hz, the segmented sweep version of SpectrumView generally takes less than 1/10 the time per sweep than does the linear frequency sweep version. Examples of these classes of measurements are shown in Figure 28-15 on page 28-19.

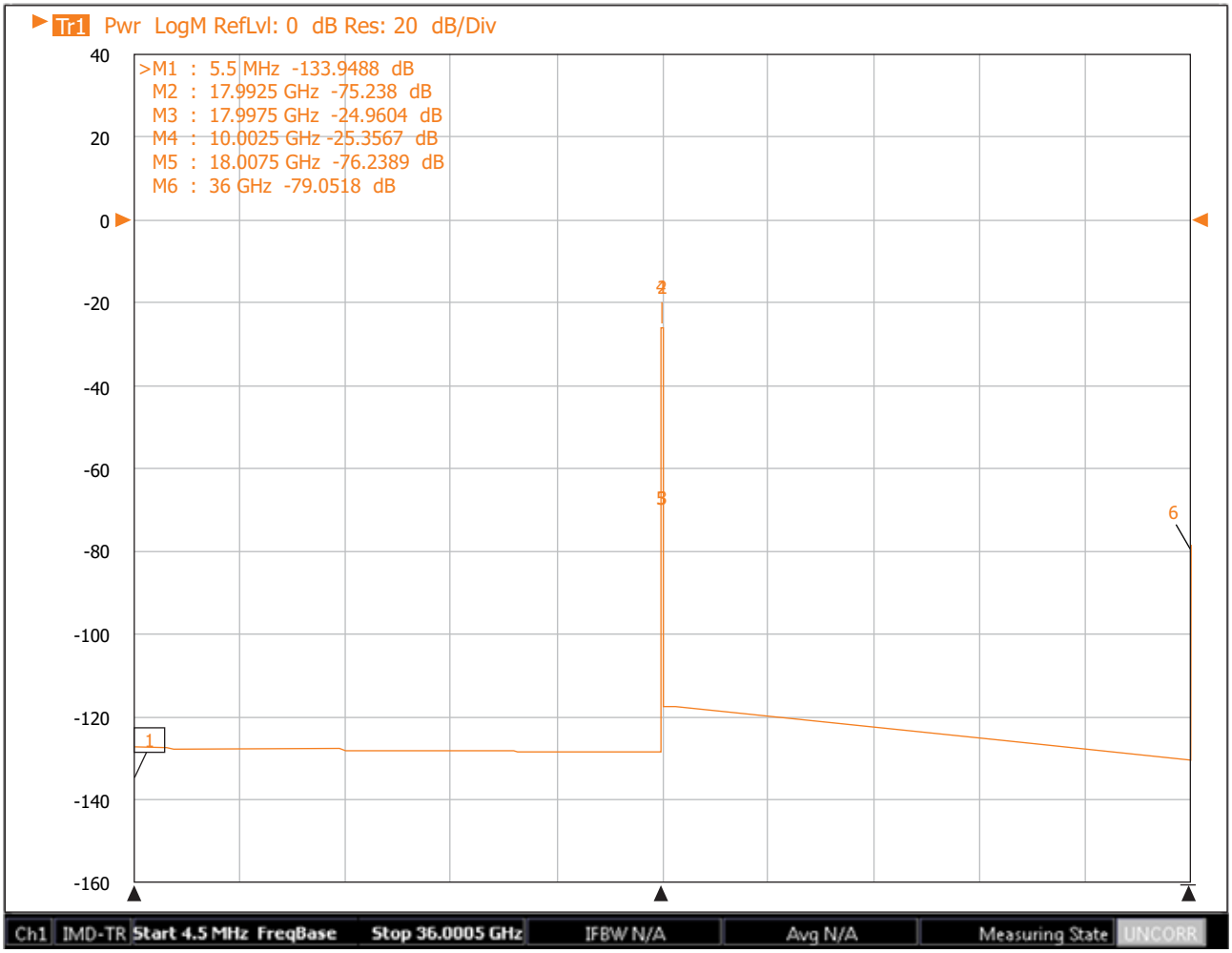




Example plots of the linear frequency sweep and segmented sweep versions, respectively, are shown here to illustrate the point density and frequency ranges. A maximum IMD order of 3 was selected for both.

Figure 28-15. Example Plots of the Linear Frequency Sweep and Segmented Sweep

When the 2nd order products are included, it can typically be difficult to visualize on screen since the frequency groupings (the odd order products and main tones vs. the second order products) are usually very disparate unless Delta is significant relative to the center frequency. An example plot with f_c = 18 GHz and Delta = 5 MHz is shown in Figure 28-16. The markers, however, will allow readout of all of the values. Second order products can only be measured in SpectrumView-segmented sweep (not linear sweep SpectrumView) and swept frequency modes. Note that in this case the lower 2nd order product is quite small but the upper 2nd order product is much higher; part of this is from the high pass nature of the DUT.



A segmented sweep SpectrumView plot where 2nd order products are included can be difficult to visualize. The marker readouts, remote data retrieval, and file data saves will capture all of the requested product and tone data.

Figure 28-16. Segmented Sweep SpectrumView Plot

The .txt file save of the above measurement is shown in Figure 28-17 on page 28-21 with all of the tone and product amplitudes highlighted.

The .txt file corresponding to the segmented sweep SpectrumView measurement with 2nd order products (and odd-order products up through 5th) is shown below.

MS4642B
!10/13/2014.8:59:38.AM
!CHANNEL.1
!TR.MEASUREMENT
!RF.CORRECTION.OFF
!AVERAGING.OFF
!IF.BANDWIDTH: 1KHZ
!NUMBER.OF.TRACES: 1
!TRACE: TRACE.1
!PARAMETER: S11
!PORT: PORT1
!GRAPH: LOGMAG
!SMOOTHING: OFF
!TIMEDOMAIN: OFF
!SWEEPTYPE: SEGMENTED SWEEP (Freq-based)

PNT	FREQ1.GHZ	LOGMAG1	
1	0.004500000	-1.192536E+002	
2	0.004750000	-1.223562E+002	
3	0.005000000	-1.249968E+002	Lower 2nd order product
4	0.005250000	-1.203293E+002	
5	0.005500000	-1.209533E+002	
6	8.987000000	-1.312447E+002	
7	8.987250000	-1.319516E+002	
8	8.987500000	-9.720628E+001	Lower 5th order product
9	8.987750000	-1.364294E+002	
10	8.988000000	-1.269367E+002	
11	8.992000000	-1.279696E+002	
12	8.992250000	-1.207548E+002	
13	8.992500000	-1.074044E+002	Lower 3rd order product
14	8.992750000	-1.287776E+002	
15	8.993000000	-1.261067E+002	
16	8.997000000	-9.622281E+001	
17	8.997250000	-1.064584E+002	
18	8.997500000	-3.406046E+001	Lower main tone
19	8.997750000	-9.706224E+001	
20	8.998000000	-9.951480E+001	
21	9.002000000	-8.510971E+001	
22	9.002250000	-9.651289E+001	
23	9.002500000	-2.913247E+001	Upper main tone
24	9.002750000	-8.721240E+001	
25	9.003000000	-8.677116E+001	
26	9.007000000	-1.219051E+002	
27	9.007250000	-1.191280E+002	
28	9.007500000	-9.859796E+001	Upper 3rd order product
29	9.007750000	-1.264688E+002	
30	9.008000000	-1.376964E+002	
31	9.012000000	-1.225438E+002	
32	9.012250000	-1.299418E+002	
33	9.012500000	-1.126324E+002	Upper 5th order product
34	9.012750000	-1.214378E+002	
35	9.013000000	-1.251717E+002	
36	17.999500000	-1.228497E+002	
37	17.999750000	-1.290535E+002	
38	18.000000000	-9.406280E+001	Upper 2nd order product
39	18.000250000	-1.169812E+002	
40	18.000500000	-1.295389E+002	

Figure 28-17. Segmented Sweep SpectrumView Text File

28-7 IMDView Application—IMD Response Types

Tone Power

The simplest measurements are just of tone power in absolute terms (dBm). The receiver simply positions at the desired tone/product location. The upper or lower one can be explicitly requested. The measurement can be set up to display the higher or lower power version (in which case both are measured so the highest/lowest can be evaluated) or the average (again, requiring that both be measured).

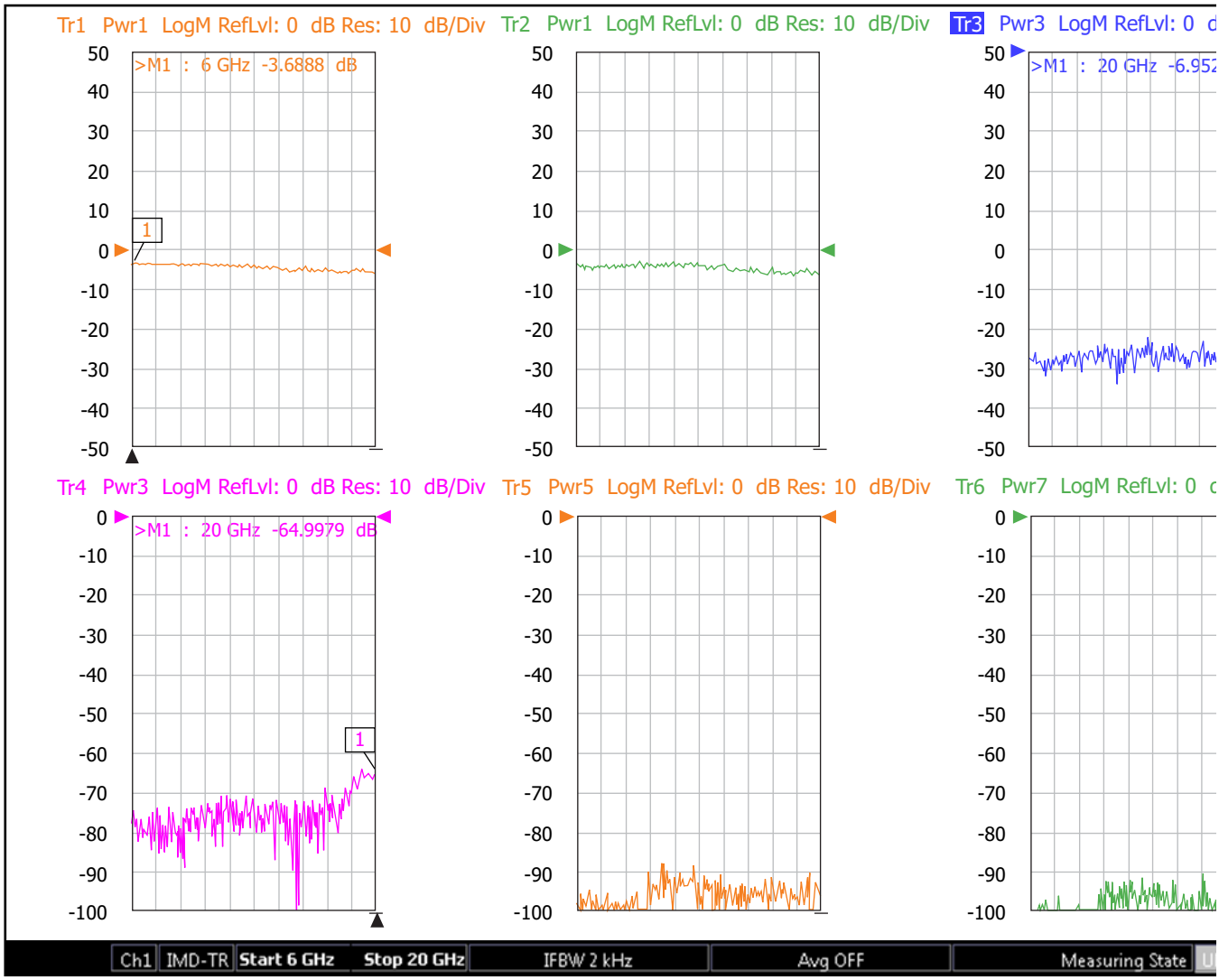
Since this is an absolute power, obviously the power and receiver calibrations will have direct impact on the result. This is a 1:1 effect in addition to any effects from changing the actual DUT operating state.

Also, since this is an absolute power quantity, the reference plane (input or output referred) is important. The native result of the measurement is output-referred. If input-referred is selected, the result is divided by (in linear terms) the DUT gain. The normalization calibration is used to get enough information to compute DUT during the measurement cycle. If the normalization calibration was not performed, the DUT gain cannot be evaluated and input-referred values will be inaccurate (will be computed using a gain of unity). Note that the normalization calibration will automatically use interpolation and extrapolation if the frequency list is changed after calibration. If the initial calibration had low point density, if the setup is electrically long and mismatched, the interpolation error can be on the dB scale.

Also, note that the tone power level cannot be changed after calibration to maintain input-referred accuracy. The gain computation is based on tone 1 so if the Delta is large, there can sometimes be frequency-dependent errors. As Deltas are normally on the scale of 10s of MHz or less, tone-to-tone frequency response is usually not an issue except in some unusually frequency-dependent structures.

Some example measurements are shown in Figure 28-18 on page 28-23: Both main tone powers (Tr1 and Tr2), both 3^{rd} order product powers (Tr3 and Tr4), the maximum 5^{th} order product power (Tr5) and the maximum 7^{th} order product power (Tr6). For this DUT, the 3^{rd} order product starts to become noticeable at the upper end of the sweep range. Log mag graph types were used here (with dB marker readouts) but Power Out graph types could also have been used (with dBm marker readouts). These measurements were all output-referred so no normalization calibration was needed. Note the different scales that were used for the various response variables.

At the bottom of the dialog is a checkbox (checked by default) to enable tone power coupling so that both tone powers will change in tandem. Note that this control is linked to the coupled source power item in the Power menu system.



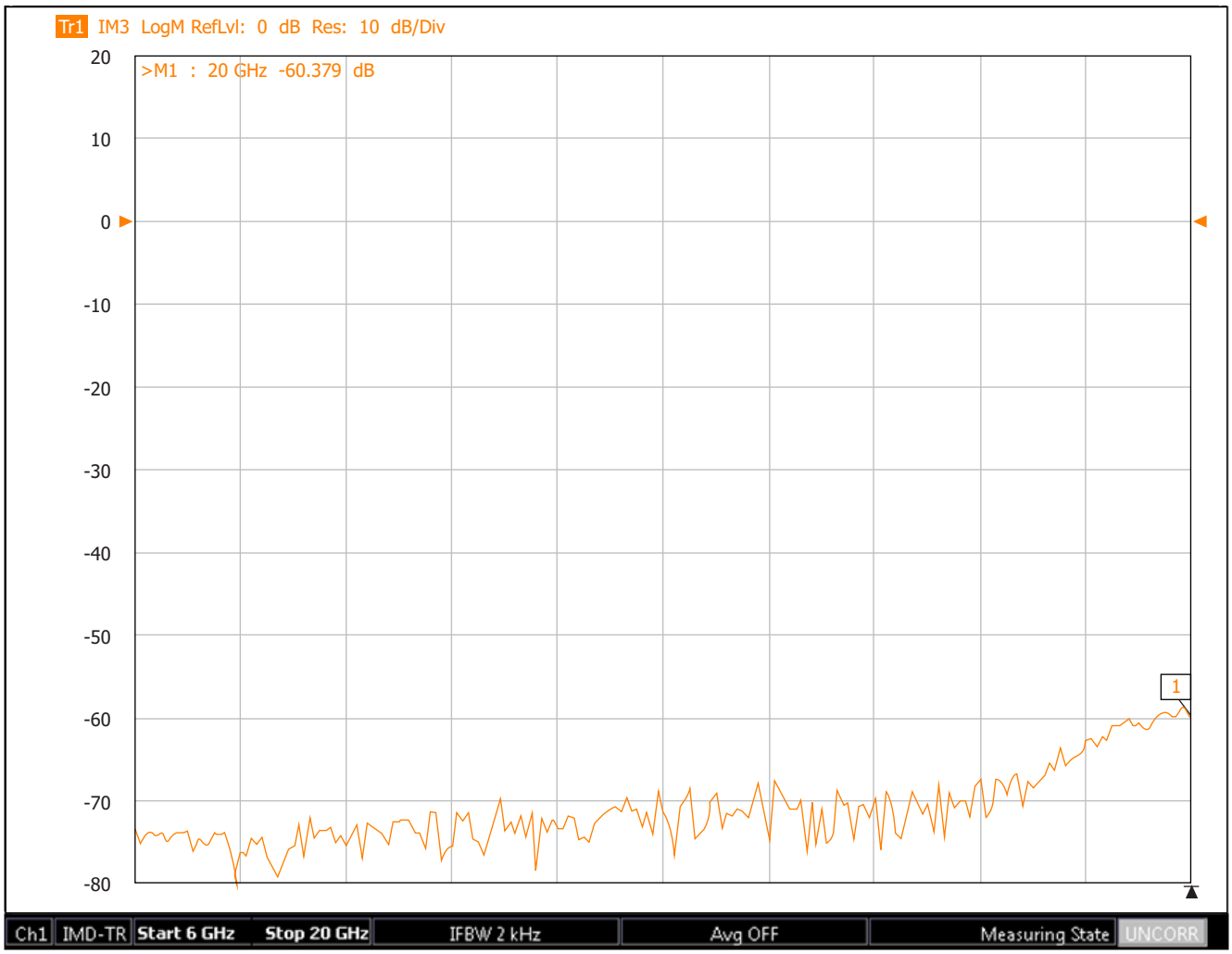
Example tone power measurements are shown here for a variety of product orders. Log mag graph types were used here (marker readouts in dB) but Power Out (or Power In) graph types can also be used (marker readouts in dBm).

Figure 28-18. Example Tone Power Measurements

IMD (Relative to a Main Tone—dBc Terms)

The IMD product in dBc terms is another commonly used format. Here the product level relative to the main tone level is the result. Not only can one select higher frequency/lower frequency/maximum power/minimum power/average tone power for the product but also for the main tone. Since this is a relative measurement, there is less direct effect of the power calibration/receiver calibration aside from DUT operating point considerations and any frequency dependence of the receiver between tone and product (that latter generally only being an issue for very large Deltas).

An example measurement of IM3 is shown in Figure 28-19 where the average of both product and main tones were selected. As might be expected from the previous section's example, the relative product also only becomes noticeable at the higher frequencies. The lower frequency results are being affected somewhat by the measurement noise floor and this effect could be reduced further by dropping the IFBW (particularly for the product measurements since it is those values that are closer to the noise floor). For this measurement type, Log Mag graph types are generally used.



An example IM3 measurement (dBc terms) is shown here. Although the log mag graph types have marker readouts in dB, the measurement response (IM3) indicates that it is a relative dBc-class result.

Figure 28-19. IM3 Measurement in dBc

Intercept Point (IPn)

The intercept point is the next level of complexity and, as defined above, represents a calculated intersection of the main tone power and product power based on the measurement at one power level. This calculation assumes the Pout/Pin slope for a main tone is always 1 and that for the N^{th} order product is N. This assumption has a large number of physical issues and the calculation is mainly still used as a matter of convention. As a result, the result is very dependent on exactly what power the measurement is performed relative to the internal nonlinear mechanisms in the DUT.

The calculation, in dBm terms, is quite simple:

$$OIPn(dBm) = P_{main_tone}(dBm) - \frac{P_{product}(dBm) - P_{main_tone}(dBm)}{n}$$

Equation 28-2.

Because of the leading term, this is an absolute power quantity causing some of the power/receiver calibration issues discussed earlier along with the DUT operating point dependency. As with the relative IMD product, the same product selection questions come into play. The input-referred version has the same normalization calibration dependency discussed earlier.

An example measurement is shown in Figure 28-20 for an output-referred IP3 result. The average of products and main tones was used for this measurement. Since it is output-referred, a normalization calibration is not required. A log mag graph type was used here (dB readouts) but a power out graph type can also be used (dBm readouts). There is some noise floor influence in this measurement at lower frequencies which could be reduced by dropping the IFBW for the product measurements.

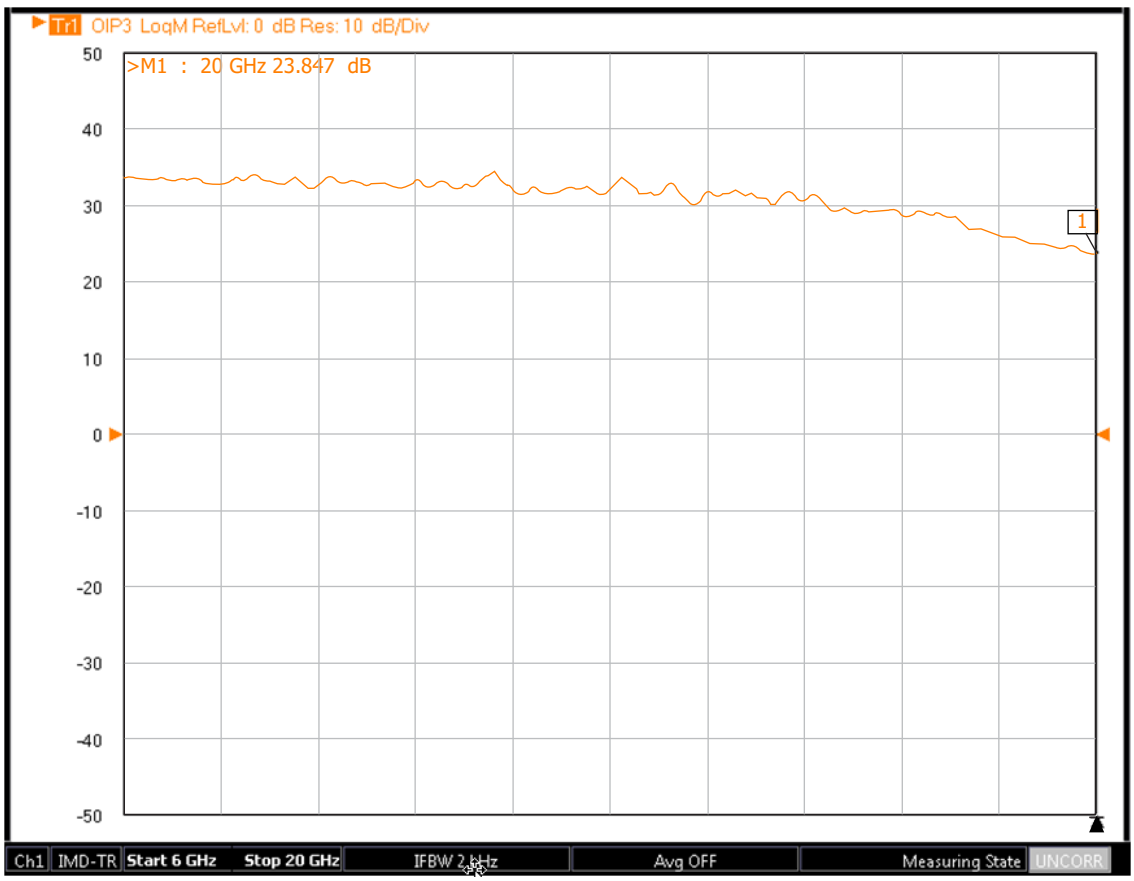


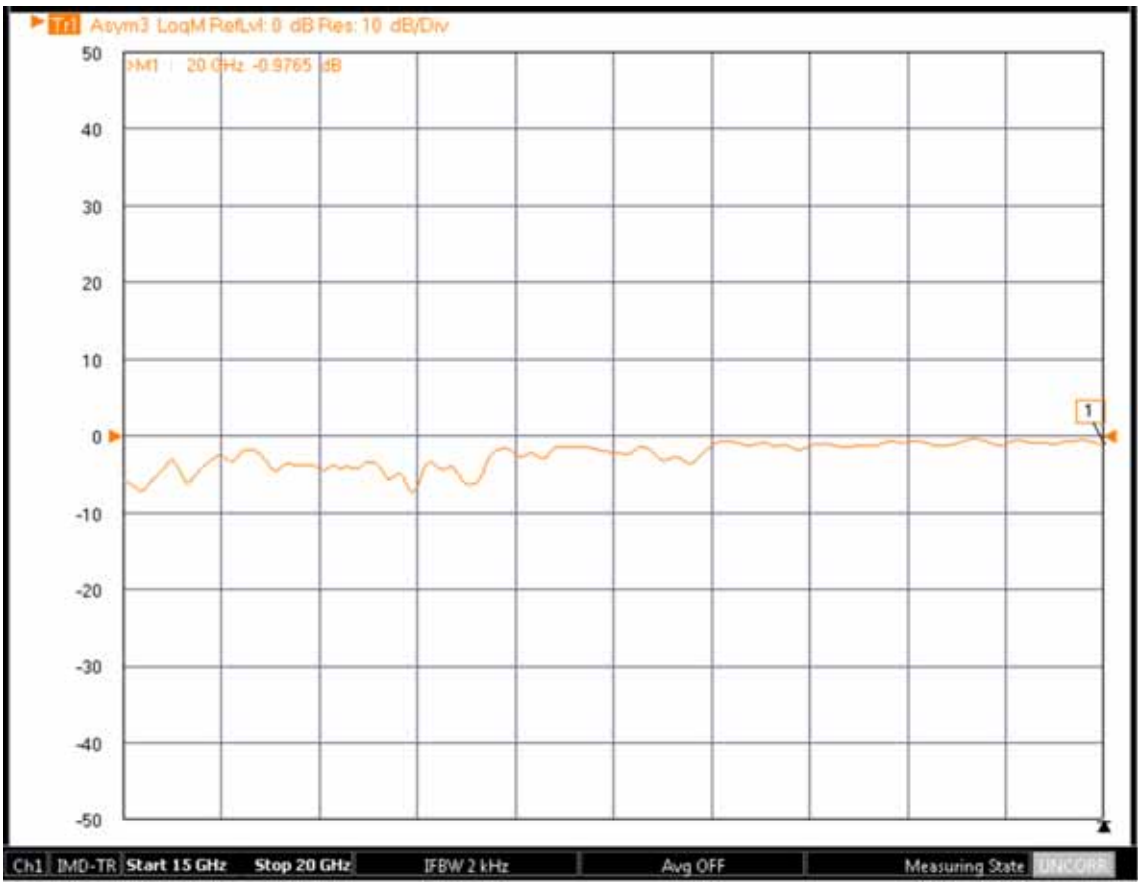
Figure 28-20. OIP3 Measurement Example

Tone Gain

The gain of the incident tone is sometimes useful to know without also activating an S-parameter channel. This can be measured using a tone gain response parameter and requires that a normalization calibration be performed (if no such calibration exists, the value of gain will default to unity). This measurement only applies to the main tone frequencies as there is, ideally, no incident power available at the product frequencies. The input signal amplitude is found during the performance of the normalization calibration and this is then simply divided into the measurement of the main tone output. This can be a useful quick check of proper device operation and behaviors such as bias and control dependencies. Log mag or Lin mag graph types are typically used for this measurement.

Asymmetry

Asymmetry is the difference between upper and lower amplitudes of a given order (including order 1 for the main tones). The reference (upper/lower/minimum/maximum) defines the sign of the asymmetry calculation. This is a relative measurement so the power/receiver calibration dependency is somewhat weaker as with the IMD (dBc) measurement. Again, if the Delta is very large there can be some frequency dependence between the two measurements which makes the receiver calibration still of value. An example measurement (over a different frequency range and with a different DUT than for some of the other examples) is shown in Figure 28-21. Log mag or lin mag graph types are typically used for this measurement.



An example product asymmetry measurement is shown above. Some of the difference here is due to a slight asymmetry in drive tones as a user power calibration had not been performed. Tone asymmetry can sometimes be an indication of 2nd order intermodulation products interacting with DUT bias or thermal time constants.

Figure 28-21. Product Asymmetry Measurement

As stated earlier in this chapter, asymmetry measurements (e.g., [4]-[5]) can be particularly useful in identifying memory effects in devices that may arise from bias network issues, thermal or trapping effects of other causes. The typical mechanism is that the lower 2nd order product (which will be at the frequency Delta) may be generated and is low enough in frequency to get into the bias networks or excite intrinsic DUT time constants. The excitation may then be able to remodulate onto the carriers at an offset of Delta or its harmonics, thus adding signal energy to the odd-IMD-product locations. Commonly, these remodulation processes are not symmetric so product asymmetry can be a hallmark of such behavior. It should be noted that instrument compression and source remodulation can also produce asymmetric responses so following the setup hints elsewhere in this chapter may be needed to avoid false indications.

Number of Sweeps Required

Now that all of the response types have been detailed, it may be useful to revisit how many measurements will be required per point in swept frequency mode. For the available orders with Option 44, there are a total of 12 possible receiver locations. One way to visualize this is with a series of check-boxes as suggested by Figure 28-22. Every trace response will have one or more measurements required so one or more boxes could be checked. If two traces, however, need the same measurement, that measurement can be shared by the two traces so excess measurements are not made. After all of the traces have their required measurements checked, one is left with the required receiver positions for that point.

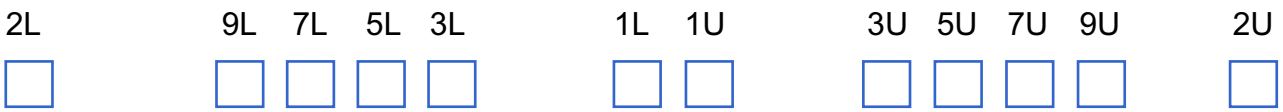


Figure 28-22. Possible Receiver Positions at a Swept Frequency IMD Point

A series of checkboxes are shown above for each of the possible receiver positions at a swept frequency IMD point (the number is the order and the letter is 'L' for lower and 'U' for upper thus 3L refers to the lower third order product and 1U refers to the upper main tone at f2). After highlighting the measurements that each trace requires (and no duplicate measurements are needed by definition), one will know how many measurements are needed to update everything.

Tone Power:

If 'higher frequency' or 'lower frequency' is selected for the Side, only one measurement is needed. If 'average power', 'max power' or 'min power' is selected for the Side, then both measurements of that order must be made.

IMD (relative) or Intercept Point:

If 'higher frequency' or 'lower frequency' is selected for the Side, only one product measurement is needed. If 'average power', 'max power' or 'min power' is selected for the Side, then both product measurements of that order must be made. If 'higher frequency' or 'lower frequency' is selected as the reference type, then only one main tone measurement is needed. If 'average power', 'max power' or 'min power' is selected for the reference type, then both main tone measurements must be made. Thus for these response types, up to 4 measurements may be needed per trace.

Tone Gain:

This only applies to the main tones. If 'higher frequency' or 'lower frequency' is selected for the Side, only one measurement is needed. If 'average power', 'max power' or 'min power' is selected for the Side, then both measurements must be made.

Asymmetry:

By definition, both measurements of that order must be made.

As an example, suppose the traces are configured as follows:

Tr 1: OIP5, side is average power, reference type is average power (required 1L, 1U, 5L and 5U)

Tr 2: Tone power 5, side is higher frequency (requires 5U)

Tr 3: Tone power 5, side is lower frequency (requires 5L)

Tr 4: Tone power 3, side is average power (requires 3L and 3U)

Tr 5: Asymmetry 3 (requires 3L and 3U)

The checkbox array for this example is shown in Figure 28-23. A total of six measurements will be required (6 sweeps normally before updating of the display occurs) as there is considerable re-use of the measurements amongst the traces.

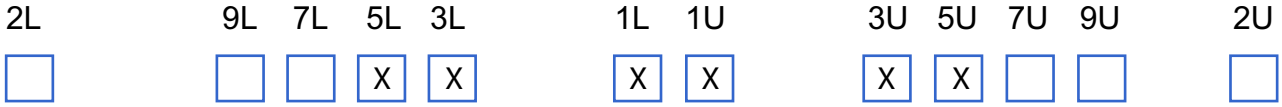


Figure 28-23. Required Receiver Measurements (for above example)

Since the displayed data is only updated when all measurements of a cycle are completed, a progress indication is given on the status bar at the bottom of the screen. A series of dots will appear as measurements are being made and a '+' will appear when a cycle is completed. These messages are shown in Figure 28-24. Messages for other channels (if being used in IMD) will appear sequentially.



Figure 28-24. Status Bar Messages—IMD Measurement Status

IMDView Measurement Variables

Table 28-2 describes IMD measurement variables.

Table 28-2. IMD Measurement Variables

Measurement	Valid Orders	Valid IM Side	Reference Plane Has Effect?	Reference Tone Has Effect?
Tone Power	All	All	Yes	No
IMD Relative to Carrier	All (Except order 1 will be trivial or an asymmetry measurement)	All	No	Yes
Intercept Point	All except 1	All	Yes	Yes
Tone Gain	Only 1	Only lower frequency	No	No
Asymmetry	All	All except average	No	No

28-8 IMDView Application—Other Parameter Selections

Delta

While often a test protocol will specify the Delta to be used, there can be some effects of Delta selection on the measurement. As discussed earlier, the DUT response may change with Delta based on internal behaviors (often involving bias systems, thermal effects, trapping, control signal interactions (as with automatic gain control)….). Also, as discussed, the Delta choice can affect the allowed f_c frequency range to ensure that the instrument can make the measurement.

Less obvious perhaps, is a dependence on measurement IF bandwidth (IFBW). If one is measuring the upper 3^{rd} order product (for example) and the IFBW is on the order of Delta, the upper main tone can easily corrupt the measurement. Depending on the product level being measured, one often desires that the IFBW be no larger than 1/10–1/100 of the Delta.

Also, as the Delta gets very small, the measurement of a product can occur on the phase noise skirts of a main tone. This becomes most obvious for a Delta below 100 kHz and going to even lower IFBWs can help (<1/1000 Delta). Source contamination also becomes harder to control at smaller Deltas as leveling loop interaction becomes more favorable. The MS464xB will work to optimize synthesizer and leveling loop configurations at these lower Deltas but these Delta dependencies can be important concepts to consider.

Among the actions taken by the MS464xB to improve low Delta measurements:

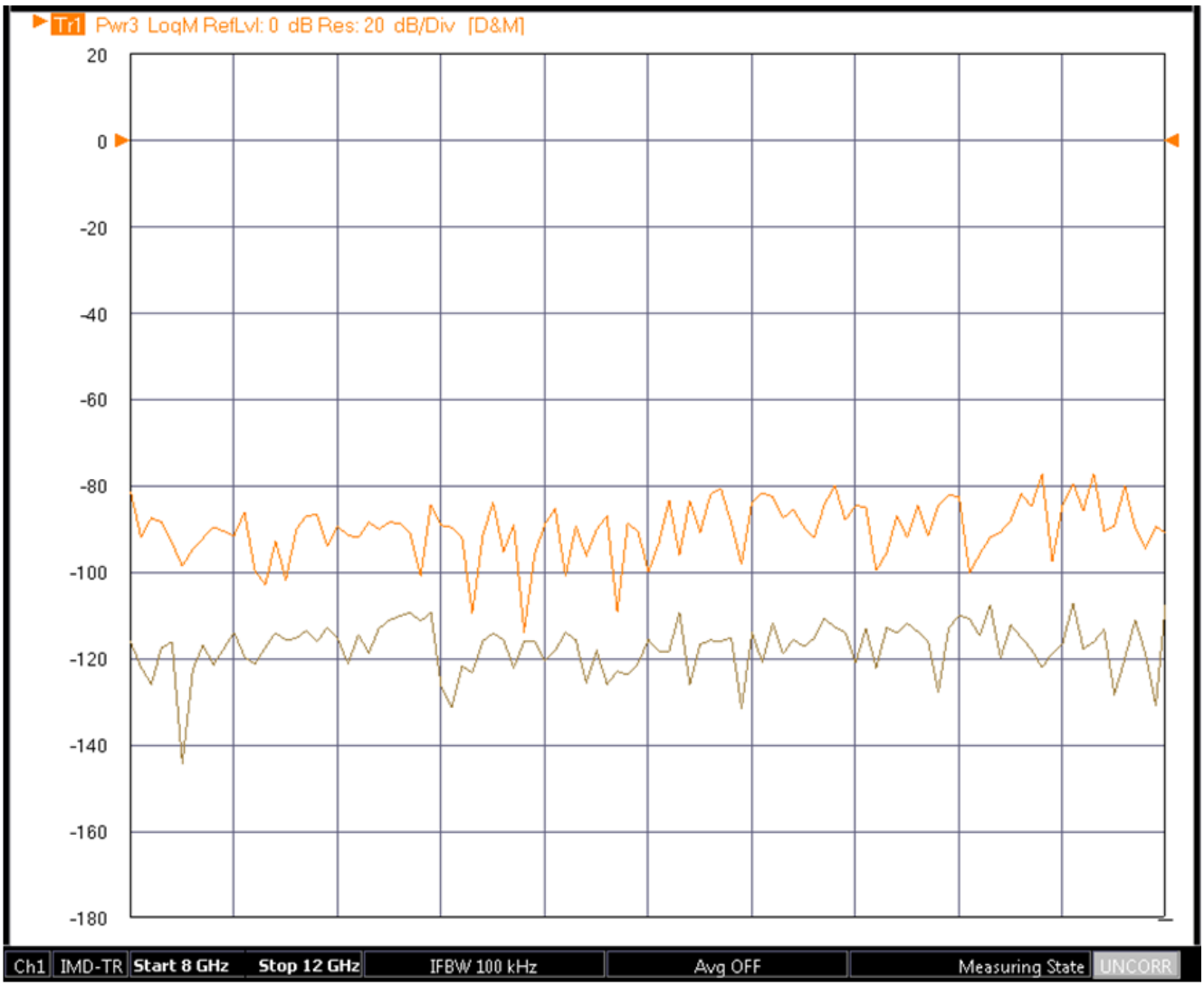
- The ALC loop bandwidth is automatically reduced as the Delta decreases. This prevents second order products from entering the leveling system and remodulating in the locations of odd order products on the sources. This does slow maximum sweep speed somewhat but typically low IFBWs are used anyway in low Delta measurements for noise and dynamic range reasons.
- The synchronization level between internal sources and the internal receiver is increased also as Delta drops. This increases phase noise correlation between source and receiver and reduces net noise levels at the system IF which can increase dynamic range. This function has the largest benefit when two internal sources are used but will help with single-internal-source measurements as well (increasing dynamic range by 10 dB or more for Deltas near 100 kHz, it offers less benefit for Deltas < 10 kHz or > 1 MHz). There is no benefit when two external sources are used. This mode is termed Common Offset and automatic control of this mode can be disabled using multiple source control (discussed later in this chapter).

In general, the measurement floor will start increasing for Delta values below 100 kHz but should still be lower than –80 dBm (for low IFBW, nominal main tone amplitudes and combiners providing at least as much isolation as the Option 32 internal combiner) for Delta values larger than a few kHz. Operation down to Hz scale Delta values are allowed but the measurement floor will continue to increase (exception: if 2^{nd} order lower products are involved, the minimum Delta is 70 kHz on systems with Option 70 since the Delta frequency must be directly measurable by the instrument in that case).

IF Bandwidth

As stated in the previous subsection, the IFBW and Delta interact to a considerable degree. Other factors, however, may also impact the choice of IFBW such as required measurement speed, desired noise floor, and location of other DUT spurious signals. In SpectrumView segmented sweep and in swept frequency IMD, it is possible to select different IFBWs for main tones vs. products. This can help in optimizing measurement speed since the main tones are generally at a higher signal level so the noise floor during those measurements may be less important to accuracy. By sweeping faster during the main tone measurements, time is saved without much of an uncertainty penalty.

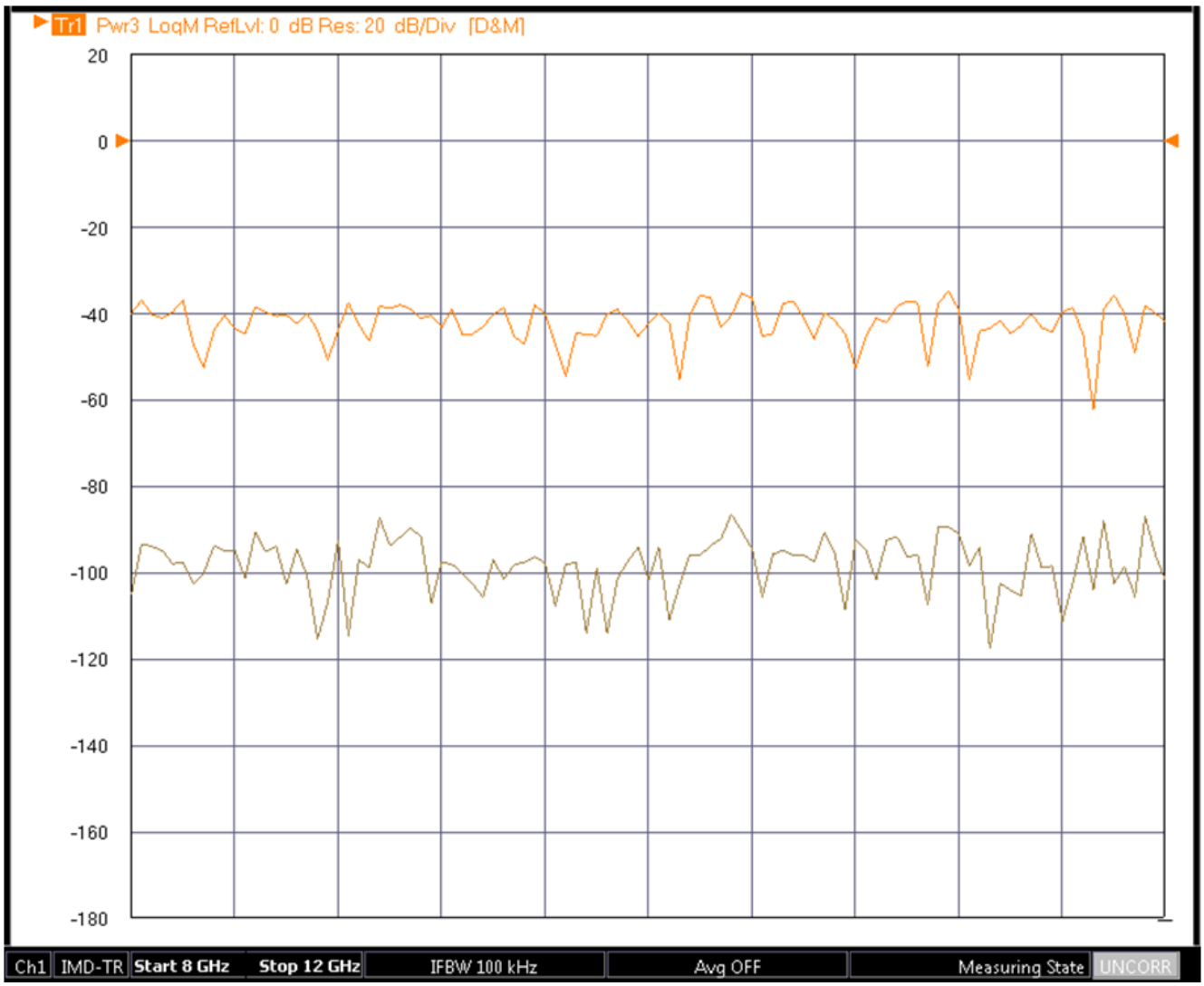
To see how some of these effects can interact, consider the measurement of a 3rd order product using 100 Hz and 100 kHz IFBWs. In one example, the Delta will be 30 MHz and the results are shown in Figure 28-25. The measurement is far away from the main tones in both cases so there is little contamination and the phase noise skirts cause little interference. As such, the IFBW change causes some change in integrated noise power but, at 100 Hz, the measurement runs into a lower system floor so there is not as much difference as one might classically expect ($10*\log_{10}$(IFBW ratio)).



Upper trace is for the 100 kHz IFBW.

Figure 28-25. Measurements of 3rd Order Product (Delta=30 MHz) for 100 kHz and 100 Hz IFBWs

When the Delta is reduced to 30 kHz (using the same two IFBWs), the differences are much more stark as shown in Figure 28-26. Now the phase noise skirts are more significant: the 100 Hz IFBW floor is raised somewhat and the difference to the 100 kHz trace is much larger due to contamination of the main tones on this measurement. Note that a manual gain ranging selection of 'Low' was used for all of these measurements (to be discussed further in Section 28-9). If Gain Ranging was left in Auto, the effects would have been even more dramatic as the instrument's IF system may have entered compression due to the presence of the contaminating tones.



The measurements of a 3rd order product (Delta=30 kHz) for 100 kHz and 100 Hz IFBWs are shown here. The upper trace is for the 100 kHz IFBW.

Figure 28-26. Measurements of 3rd Order Product (Delta=30 kHz)

Spur Avoidance

There are several levels of spur avoidance that the MS464xB performs and some are completely done behind the scenes where various sub-synthesizers are re-arranged on a per-point basis so that the resulting measurement has the lowest residual spurs. For IMD, there are two additional classes of interest since the receiver is scanning product locations while in the presence of large main tones:

IF-Related Issues

If the Delta or its multiples or sub-multiples gets near to the system IF frequency, to the image frequency, or to the clock frequency, other residual spurious receiver responses can occur. In these cases, the system will automatically change to a new IF frequency. In SpectrumView, where this is harder to do on a point-by-point basis, the frequency list is altered to minimize the effects of the primary image.

Reference Issues

The 10 MHz system reference is omnipresent (and, indeed, in many laboratories, the 10 MHz rides on mains ground) so using this as a Delta can be problematic. If the 'Avoid spurs at 10 MHz and subharmonic deltas' box is checked (and it is by default), and this issue is detected, the Delta will automatically be shifted by 1 product IFBW with a maximum shift of 10 kHz. If this shift is unacceptable, spur avoidance can be disabled with the checkbox on the IMD configuration dialog but note that the available dynamic range for product measurements will be reduced.

28-9 IMDView Application—Receiver Configuration (Broadband/mmWave)

The instrument receiver configuration can be an important practical choice: are the broadband/mmWave measurement modules (374xx, MA25x00A) being used (e.g., [6], [7]). Such a setup can allow measurements to 110, 125, 145, or 220 GHz depending on the configuration. Some setup details are different in this case as the internal combiner (if Option 32 was installed) is not available to those frequencies so an external combiner is required. Also, the cable routing is a bit different for this measurement as suggested by Figure 28-27. Note the use of the aux power connector on the 3739C test set is used to power a third module (usually termed the Aux Module) and to provide the control logic.

When using two internal sources, full broadband coverage is normally not possible, since the port 1 cable routings would have to be multiplexed in 2-port systems. External setups making use of port 1 (for <54) and the RF1 connector (for >54) are possible but these would be switched or filtered to prevent spur introduction.. The sketch in Figure 28-27 is for the >54 GHz mmWave case (for lower frequencies, the modules will be acting in pass-through mode, as discussed in Chapter 17 of this guide). Port 1 drives the tone 2 module in this case and feeds the RF IN port of the module (multiplier path). The power calibration of this tone is done using a power calibration submenu present when this configuration is activated (more details in the power calibration section). The tone 1 power calibration (coming from source 2 via the 3739C test set) is done using the normal power calibration menus. Because of these connection requirements, the application will only allow sweep ranges entirely above or entirely below 54 GHz when in this receiver configuration class.

The numbering in this diagram may be somewhat counter-intuitive but in 2-port broadband/mmWave applications using a dual source VNA (Option 31), it is source 2 that drives the test set for mmWave frequencies (and that signal is switched to the two normal modules for S-parameter measurements). Source 1 is normally unused in that configuration making it available to help generate the 2nd tone (via the third module) in an IMD measurement. In 4-port configurations, ports 1 and 2 cannot drive simultaneously and ports 3 and 4 cannot drive simultaneously (a common setup is to use port 1 and 3 driving together).

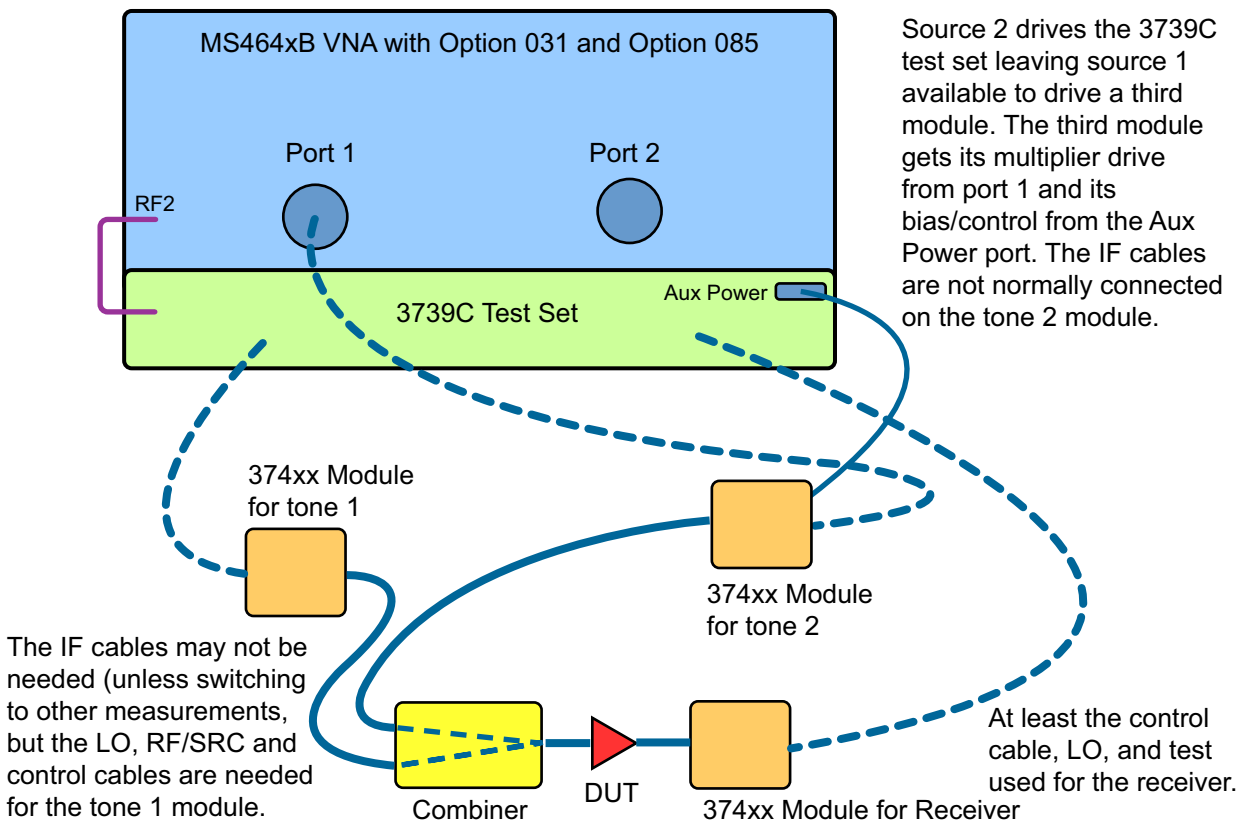


Figure 28-27. IMD Setup for mmWave (1 of 2)

The module for tone 2 in this picture is often called the 'Aux Module' since it is powered by the Aux Power connector of the test set.

Figure 28-27. IMD Setup for mmWave (2 of 2)

The mmWave IMD setup is shown above in Figure 28-27 for the case of using two internal sources. The 374xx modules are shown here for 110 or 125 GHz coverage. The MA25300A modules can be used for 145 GHz coverage. The MA25400A modules can be used for 220 GHz coverage. The modules for tone 1 and the receiver have standard connections to the test set. The module for tone 2 is connected to the Aux Power connector (for power and control) and to Port 1 (RF to drive the multiplier). Although a 3739C test set is shown, a 3739B test set can also be used. The older 3739A test set does not have the Aux Power connector required for the third module.

The situation using an external synthesizer to generate one of the tones is described in Figure 28-28. The external synthesizer would be connected to the RF In port (the multiplication path) of the mmWave module. The power calibration for the external synthesizer must be at the pre-multiplication plane using the power calibrations internal to the MG369xx or MG362xxx synthesizers (discussed elsewhere in this chapter). Generally, the 10 MHz references (VNA and synthesizer) must be connected as has been discussed and either may serve as the generator.

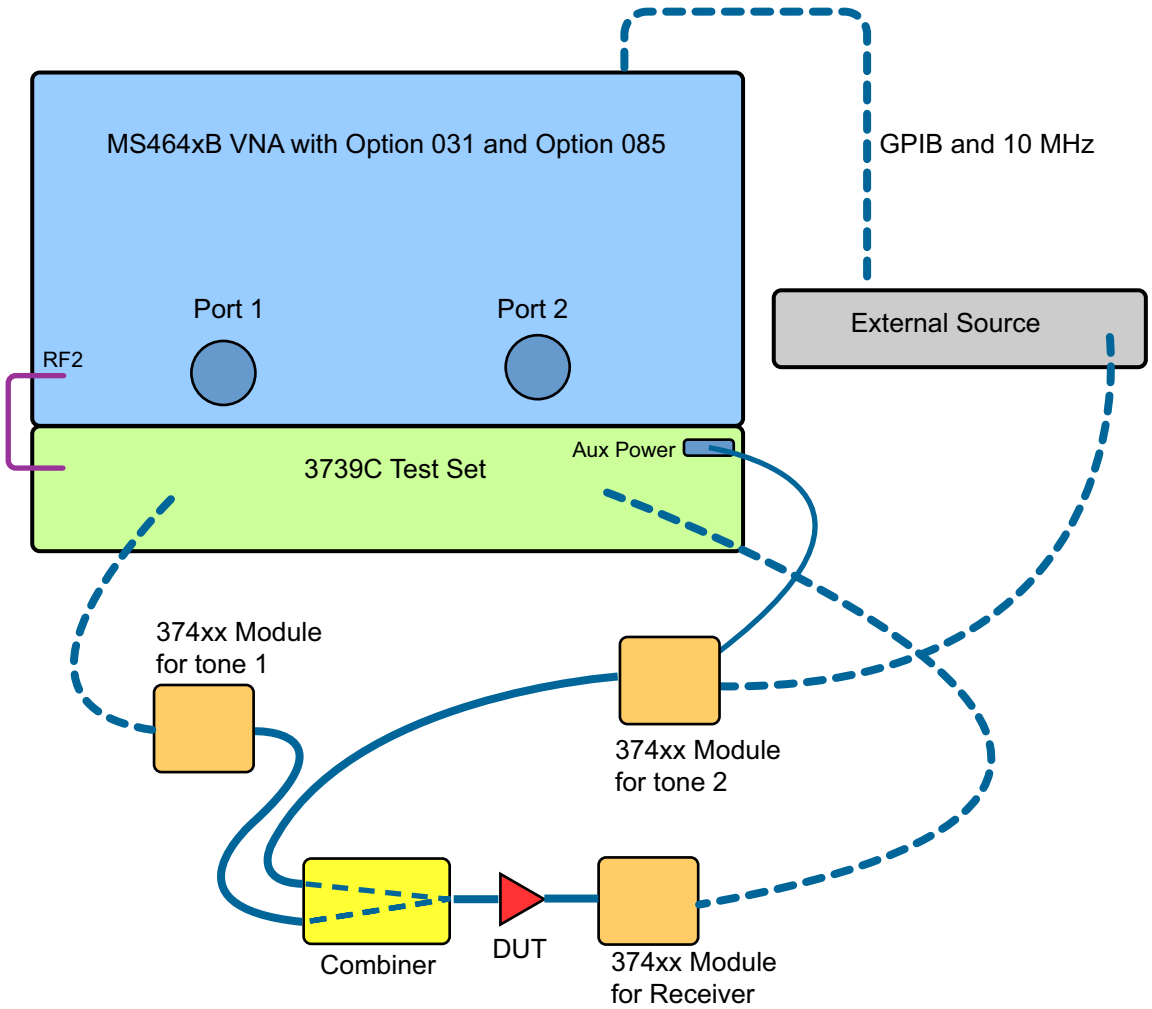
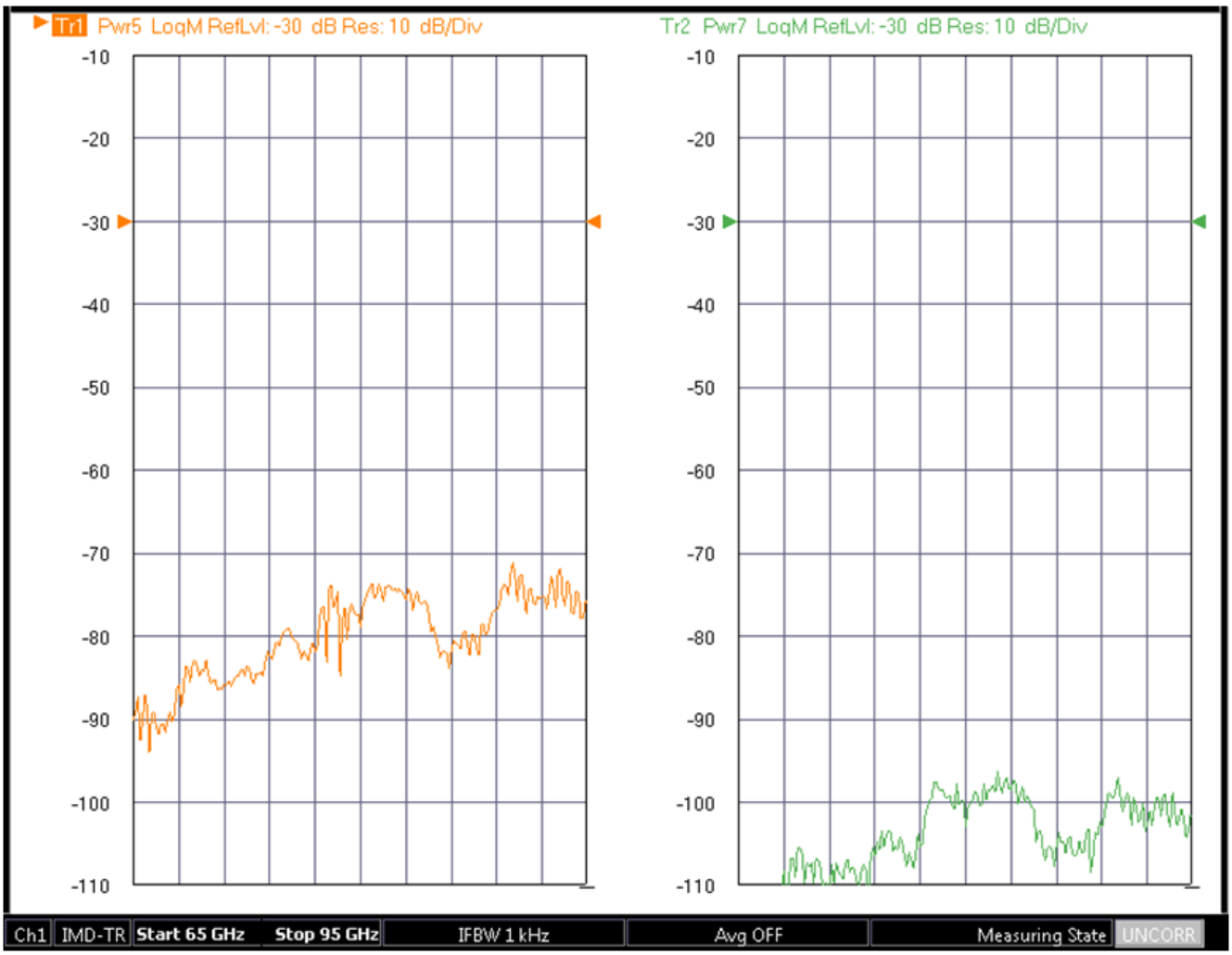


Figure 28-28. IMD Setup for mmWave—Using External Source

Up to this point, little has been said about the combiner used in the mmWave configurations except that it must be external (combining prior to multiplication would result in very high levels of source-intermodulation products making that approach impractical for most measurements). As at lower frequencies, it is mainly isolation and insertion loss that are of interest. Because of the way the leveling circuitry is constructed in the mmWave modules, the isolation requirements are a bit less severe (it is difficult for remodulation to occur). A combiner covering 30–110 GHz is available from Anritsu providing >15 dB of isolation over the band (contact Anritsu for more information).

In the above discussion, it has been assumed that a standard transmit/receive mmWave module was being used for the receiver (3743x, for example). It is also possible to use the receive-only module (3744A-Rx) for this application. This module has no coupler on the front-end so the compression levels are lower (0.1 dB at –3 dBm nominal) and the target main tone input power should be in the –23 dBm range.

As an example measurement, consider a DUT covering 65–95 GHz. The two internal sources were used with an external combiner after 3743A/AX modules as suggested in Figure 28-27. The fifth and seventh order tone powers are shown in Figure 28-29. The profiles are similar but offset in this case suggesting perhaps a rather linear final stage with some significant gain variation with frequency.



Using some higher order products as the metrics.

Figure 28-29. Example mmWave Measurement

The procedure to obtain the above measurement is relatively straightforward:

1. Configure the hardware as in the diagram of Figure 28-27.
2. Configure IMDView for the desired frequency range (65–95 GHz), tone spacing (3 MHz), tone power (–10 dBm) and products (5^{th} and 7^{th} order power levels, output referenced and on the higher frequency side).
3. Perform power and receiver cals using those respective menus. Two power sensors were used for this sweep range (A ML-series Anritsu meter with a SC7770 sensor and a W8486A sensor on a HP437 power meter). The receiver calibration was accomplished with a FF adapter and its loss was neglected (<0.5 dB at these frequencies).
4. Observe the measurement results.

With 4-port broadband/mmWave systems, the measurement concepts are the same. For dual-source systems, the setups can be a bit simpler in that the aux module does not have to be used. Instead, any allowed pairs of port modules can drive the combiner directly (ports 1 and 3, 1 and 4, 2 and 3 or 2 and 4). Of course, the aux module control is still allowed (and there are two aux connectors on 4-port systems) and external synthesizers can be used. An example setup is shown in Figure 28-30 where the port 1 and port 3 modules are used to provide the tones (using standard cable harnesses) and the port 4 module is used as the receiver. The thick blue lines in the diagram denote the ME7838 cable harnesses. The thinner blue lines to the module denote the single RF cables coming from the ports that would be used below 54 GHz. Unlike the 2-port case, both bands can be connected in the 4-port IMD setup

Manual configurations are possible where two modules are used as receivers (e.g., for differential-output DUTs) but this has to be configured directly with multiple source control. Note that the receivers are time-coherent so differential and common-mode IMD components can be resolved in that case.

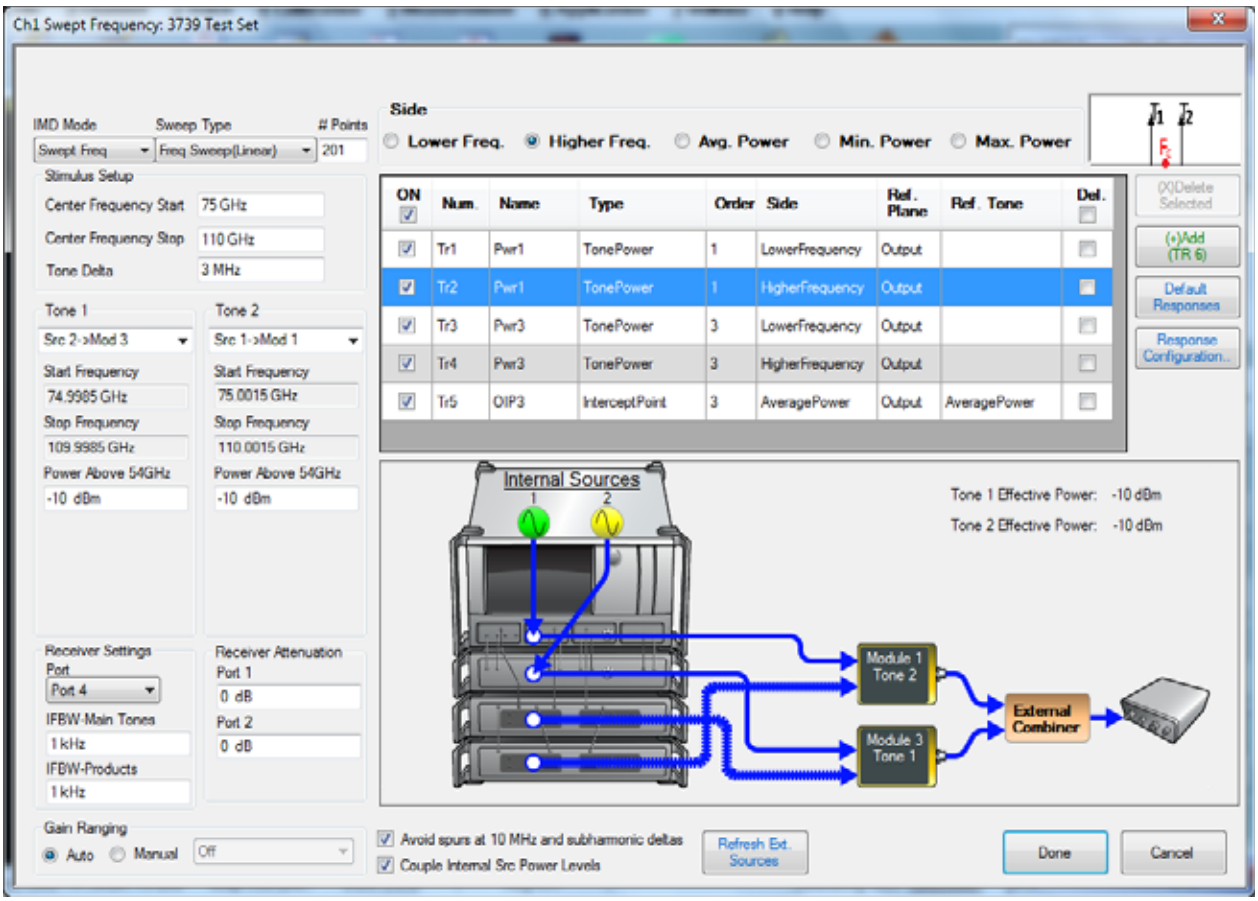


Figure 28-30. Port 1 and Port 3 Modules Provide the Tones Port 4 Module is Receiver

28-10 IMDView Application—Calibrations and Settings

Power Calibrations

As can be gathered from prior sections, power control and power accuracy can be quite important in many IMD measurement scenarios. In addition, combiner losses are often not negligible and, when using external combiners, getting some confidence in the actual power at the DUT will often require a power calibration. Some additional comments on power control and how power calibrations are accomplished might be useful.

In instruments with Option 61 or 62, step attenuators are available for positioning source power for the internal sources. Generally when using external sources, the step attenuators in those units are used for power control. One can conceivably route an external synthesizer into the source loop but, note that the step attenuators are before the loop access so they will not be of use in attenuating an external synthesizer level.

User power calibrations orchestrated from within the VNA do not apply to external synthesizers, but those synthesizers have their own power calibration routines that can be used. While the *MG369xC Operation Manual* (10370-10373) and the *Rubidium MG362x1A Series Low-Noise RF/Microwave Signal Generators Operation Manual* (10370-10383) contain more complete information, the general procedure will be summarized here. Up to five tables with 2–801 frequency points can be stored and these are constructed with a calibration procedure much like that used for the MS464xB user power calibrations. The synthesizer must not be under VNA control when the calibration is performed. A power meter is connected to the synthesizer GPIB bus and the sensor connected to the relevant reference plane (usually after the combiner). The frequency sweep should be manually set up on the synthesizer to correspond to the tone range to be used for the IMD measurement. The power level should be manually set to somewhere around the level that will be used for the IMD measurement (the penalty for missing the exact level is quite small due to the linearity of the synthesizer ALC system). The power meter GPIB address does need to be correct for what the synthesizer expects although the default values are common between the VNAs and the synthesizers from Anritsu. The setup is shown in Figure 28-31.

After performing the calibration (and assigning it to one of the user offset tables) and activating the calibration, the synthesizer can be returned to VNA-control (after reconnecting GPIB cables). Note that calibration (offset) tables can also be loaded onto the synthesizer directly from a computer using GPIB (see the MG369xx (10370-10375) or MG362xxx (10370-10385) Programming Manual for more information). When the VNA requests that the synthesizer program a specific power level, the calibration (offset) table in the synthesizer will automatically be used. Note that if the synthesizer is reset, the calibration (offset) tables may be lost and the process will need to be repeated if they have not been saved.

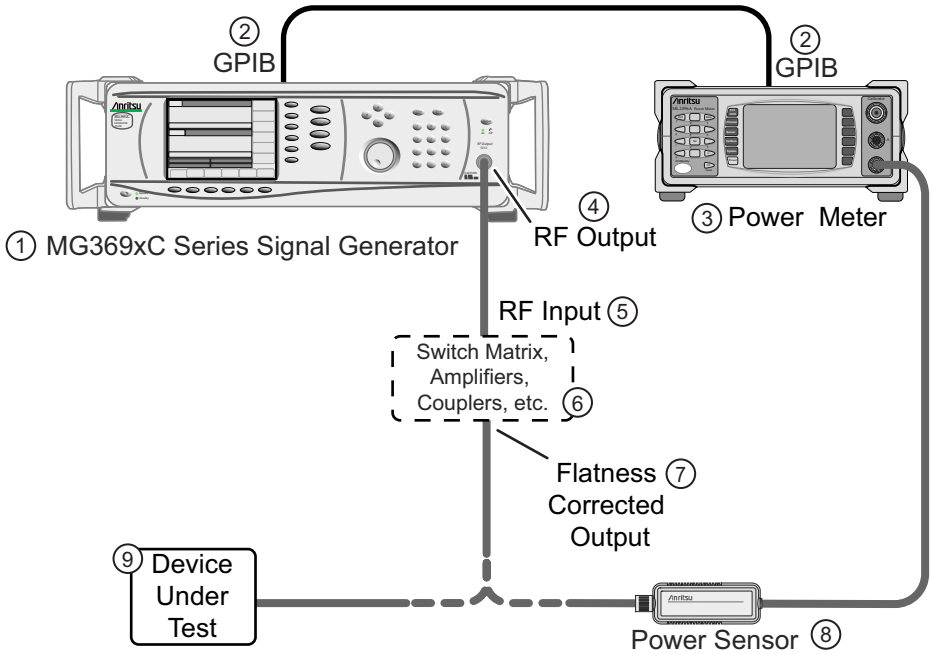
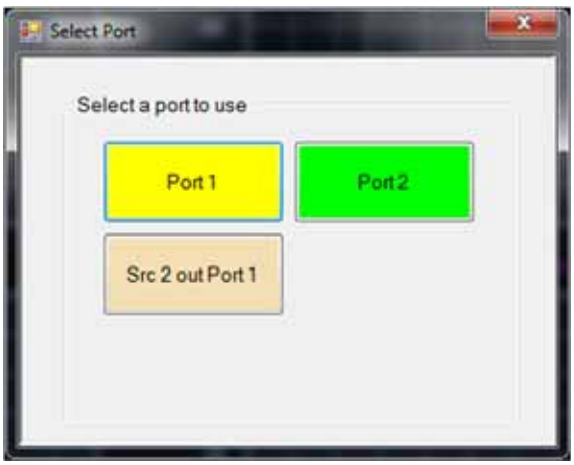


Figure 28-31. Power Calibration for External Synthesizer

When Option 32 is installed and the two internal sources are being used, some of the power issues are handled for the measurement. Note that the step attenuators of Option 62 are before the combiner so they can be used to independently adjust tone level if both internal sources are being used. There is an additional factory ALC calibration for source 2 when routed to source 1 so the power at the port for both tones will be reasonably accurate (to within nominal specifications). User power calibrations, used to get additional accuracy, also make a distinction on source 2 whether it is driving port 2 or is being routed to port 1. The dialog for this selection is shown in Figure 28-32 and the power calibration menu when this path has been selected is shown in Figure 28-33 on page 28-40.



When equipped with Option 32, the power calibration port selection menu has 3 choices (2-port system) as shown above. The ‘Src 2 out Port 1’ selection enables the redirection path for the calibration and the corresponding calibration will store values for use when that path is active.

Figure 28-32. Power Calibration Port Selection (Option 32 Installed)

The relationship between the power entry fields and the Target Power field of the power calibration is important. To review a few concepts from the Receiver calibration and Power calibration chapters (Chapter 8 and Chapter 15 of this guide, respectively), a few definitions:

ALC Entry Field (Port n Power on Power Menu):

Where the requested power is entered on the main Power menu for Port 1, Port 2, Src2 out Port 1, etc. (this will also apply to the >54 GHz power fields for the 3739-test-set-based mmWave modes and to the Aux Module power levels for mmWave IMD measurements).

Effective Power Field (on Power Menu):

The read-only fields below the ALC entry fields. These normally reflect the settings of source attenuators (if present) but play an additional role when power calibrations are active.

Target Power Field (on Power Cal Menu):

The entry area on the Power Cal submenu where the objective of the power calibration is entered.

The normal procedure is to enter the desired power at the user reference plane (where the power sensor will be connected for the power calibration). If the gain/loss between the VNA port and the user reference plane is small, the ALC entry field can be set to the same value as the Target Power. If the gain/loss is significant, the ALC field should be offset by an estimate of that gain/loss (ALC entry field should be higher if it is a loss, and lower if it is a gain). This offset need not be extremely accurate; it serves to inform the system of where roughly to set the internal sources in order to achieve the desired user reference plane power. After the power calibration is completed, the Effective power field will match the Target power. If the ALC entry field is later changed (from the value at the time of the calibration), the Effective power field will also change to reflect an estimate of the new user reference plane power. This new value is only an estimate since the calibration was not performed at that power level and the accuracy will decrease as the change from the calibrated level increases.

Example Calibration

An external combiner is being used with two internal sources so that there happens to be about 8 dB of loss between the VNA port and the DUT reference plane. The desired power for each tone is –20 dBm.

Start with Source 1:

1. Set Target Power to –20 dBm and set the ALC entry field (for that port) to –12 dBm. Perform the power calibration.
2. After the calibration is successfully completed, the Effective power will read –20 dBm and the ALC entry field will remain at –12 dBm.
3. Change the ALC entry field to –15 dBm. The Effective power field will change to –23 dBm (the estimate of the new power at the reference plane).
4. Repeat the above steps for Source 2 (using the appropriate ALC entry field and Target power for that source); the power numbers will be the same as above.
5. Turn the power calibrations off. If the ALC entry fields are still at –15 dBm, the Effective power fields will change to also be at –15 dBm. With the power calibration no longer applied, the system's best estimate of the power being applied is that based on the internal power calibrations.

Note that the source attenuator settings (if Options 61/62 is installed) will be reflected in the Effective power values even if no power calibrations are applied.



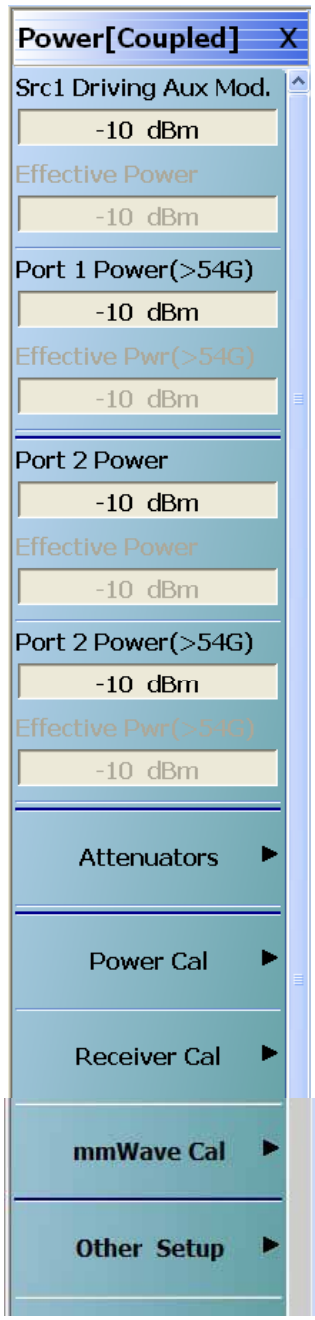
The power calibration menu is shown here when the redirection path has been selected. The power calibration will attempt to get –10 dBm at the power sensor location with src 2 directed to port 1 for all frequencies in the current frequency list. Note that because of the internal loss on the redirect (5 dB to 10 dB), –10 dBm may not be achievable at all frequencies on 50 GHz and 70 GHz systems.

Figure 28-33. Source Selection in Power Cal Menu (Option 32)

Again, a reminder is that there is additional loss from source 2 when routed to port 1 so the maximum power is lower (see the Technical Data Sheet for more information).

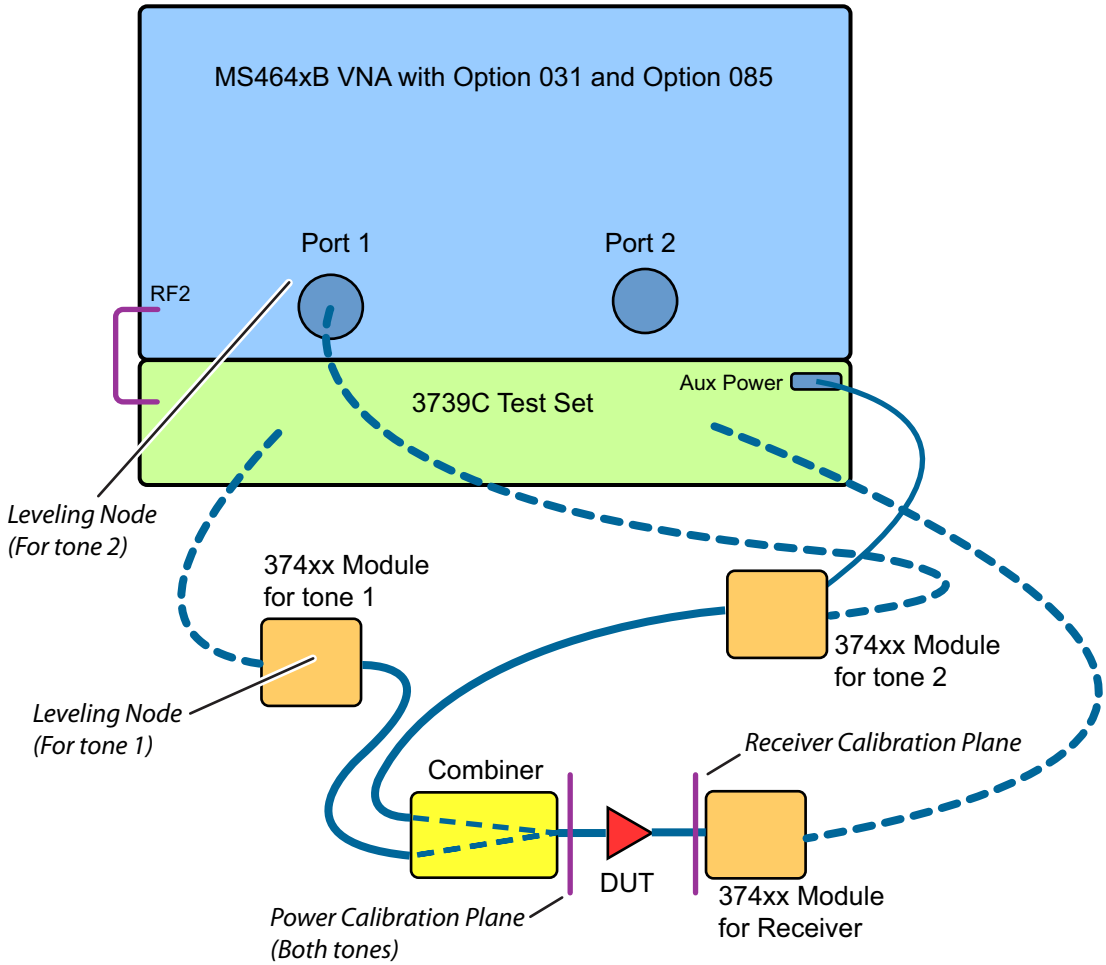
When using the broadband/mmWave modules, there are some additional changes to power calibrations but these are largely the same as for ordinary broadband/mmWave modes (see Chapter 17 of this guide for more information). As the 1.85-mm connectorized power sensors only operate to 70 GHz, a different sensor (e.g., W8486A on a 437-compatible meter) must be used to 125 GHz and a third sensor (ELVA-01) for frequencies between 125 and 145 GHz. If a sensor change is required for a user power calibration over a swept frequency range, the system will automatically prompt for a sensor connection change. Note that the three meters must all be on different GPIB addresses (by default, at 13, 15, and 17 respectively).

The tone sourcing from the traditional module labeled port 1 (connected to port 1 of the 3739x test set) will use the port 1 user power calibration field. The tone sourcing from the third module (when using a dual source VNA) will use a new user power calibration field (see Figure 28-34) and this power calibration will reference the input to the multiplier path of the module (since no leveling is available through the module). The power calibration will, however, proceed in the same way. Since the leveling is at an earlier point in the chain, the stability over time may be slightly worse than for the module being driven by the 3739x test set. These concepts are annotated in Figure 28-35 on page 28-42.



The power calibration and indication fields for the dual source/mmWave scenario are shown here.

Figure 28-34. Power Menu



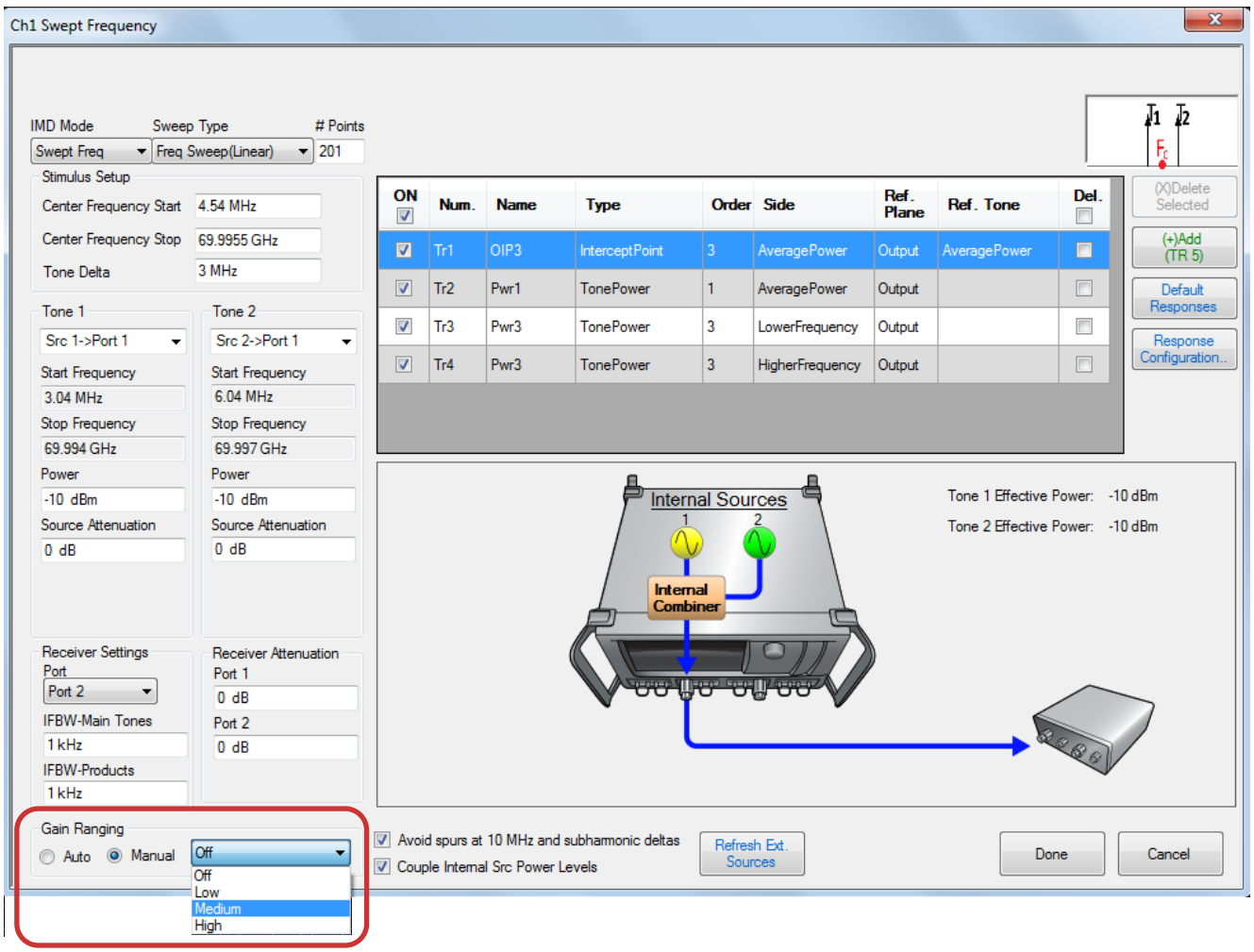
An annotated version of Figure 28-27 is shown here with the calibration planes. The power calibration planes are normally in the same place for both tones (even when using two internal sources) but, for architectural reasons, the leveling nodes may be in different locations. This does not affect instantaneous accuracy but can affect longer term power calibration accuracy.

The dual source connection plan for >54 GHz measurements is repeated here with notes regarding the power and receiver calibration planes and the leveling locations.

Figure 28-35. IMD Setup for mmWave—Using Dual Source

IF Gain

A Gain Ranging selection is available for controlling IF gain when in IMD measurements (see Figure 28-36). In most measurements (S-parameters for example), the IF gain is automatically selected based on the current amplitude in order to maximize dynamic range of the system. When another signal is nearby (not at the receiver location, as when measuring a product and a main tone is nearby), the decision process can put the IF chain into too high a gain and the system produces higher level of intermodulation products itself. At Deltas <3 MHz, the system will automatically use controlled gain but this can be overridden by a manual gain selection. At larger Deltas, one may wish to use a manual gain setting if incident tone powers are approaching instrument compression limits. The highest gain that can be used without introducing distortion will optimize dynamic range. One can check for this by doing an initial measurement with gain off and then increase gain until products increase (and then back off a step). The gain steps are nominally 10 to 14 dB.



The IMDView configuration dialog with manual Gain Ranging control is highlighted here. Particularly for small-Delta cases, going to Manual gain control can minimize main tone contamination problems (Off and Low settings are most often used).

Figure 28-36. Gain Ranging Control in IMDView Configuration Dialog

Receiver Calibrations

The receiver calibrations are covered in some detail in Chapter 8 of this guide but portions are repeated here for convenience. The concept of a receiver calibration is to transfer power calibration accuracy to the receiver so that absolute power measurements are possible. For the highest accuracy results, a user power calibration often precedes the receiver calibration so that cable and environmental losses at least can be included (the factory ALC calibrations are referenced to the test port). Then a thru connect is established between the power calibration reference plane and the receiver reference plane as suggested by Figure 28-37.



The receiver calibration menu is shown here for the default path of source 1 driving port 2. Just a thru line connection to the desired receiver plane from the power calibration plane is all that is required.

Figure 28-37. Receiver Cal Menu

The source and receiver being used are specified prior to running the calibration. Normally for IMD, source 1 will be used as the stimulus (although it does not have to be) and commonly the test receiver on port 2 is the target to calibrate.

When in IMD mode, the receiver calibration frequency list will automatically be configured to include the measurement locations (which product orders, upper vs. lower, etc.) requested. As such, the point count can increase significantly relative to what the frequency menu says. The higher order product locations will be discarded first as the maximum number of points is reached (25000 or 100000) with the 2nd order points discarded last (if they have been requested) since they are furthest away from the main tones. The reasoning on the others is that the odd orders are closer to the main tones and hence likely to suffer less interpolation errors. The highest odd orders are likely to be dominated by the noise floor in an actual measurement so any interpolation error on those will have the lowest net effect.

The receiver calibration can still be performed outside of IMD and will be applied correctly but may utilize additional interpolation depending on the IMD frequency requirements. As long as the receiver reference plane position has not changed, there would no further accuracy penalty.

Normalization Calibrations

As discussed earlier, the normalization calibration is intended to allow on-the-fly DUT gain computations to allow for input-referred tone power and intercept point measurements. This calibration is not required for other measurements and is not used in SpectrumView. The calibration is performed with a thru line connect between the power calibration and receiver reference planes just as with the receiver calibration (menu shown in Figure 28-38). The difference is that a different (simpler) frequency list is used and the normalization calibration can only be performed within IMD and uses the current power level (nominally the power level at which the test was specified; recall the importance of this particularly for intercept point responses). The receiver calibration can be performed in non-IMD modes and can be performed at other power levels (as long as there are no linearity or noise issues induced by that choice).

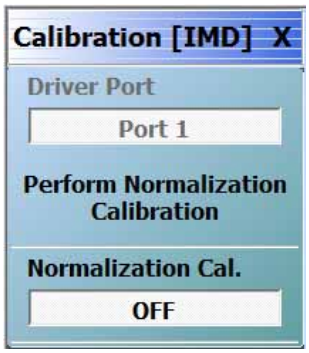
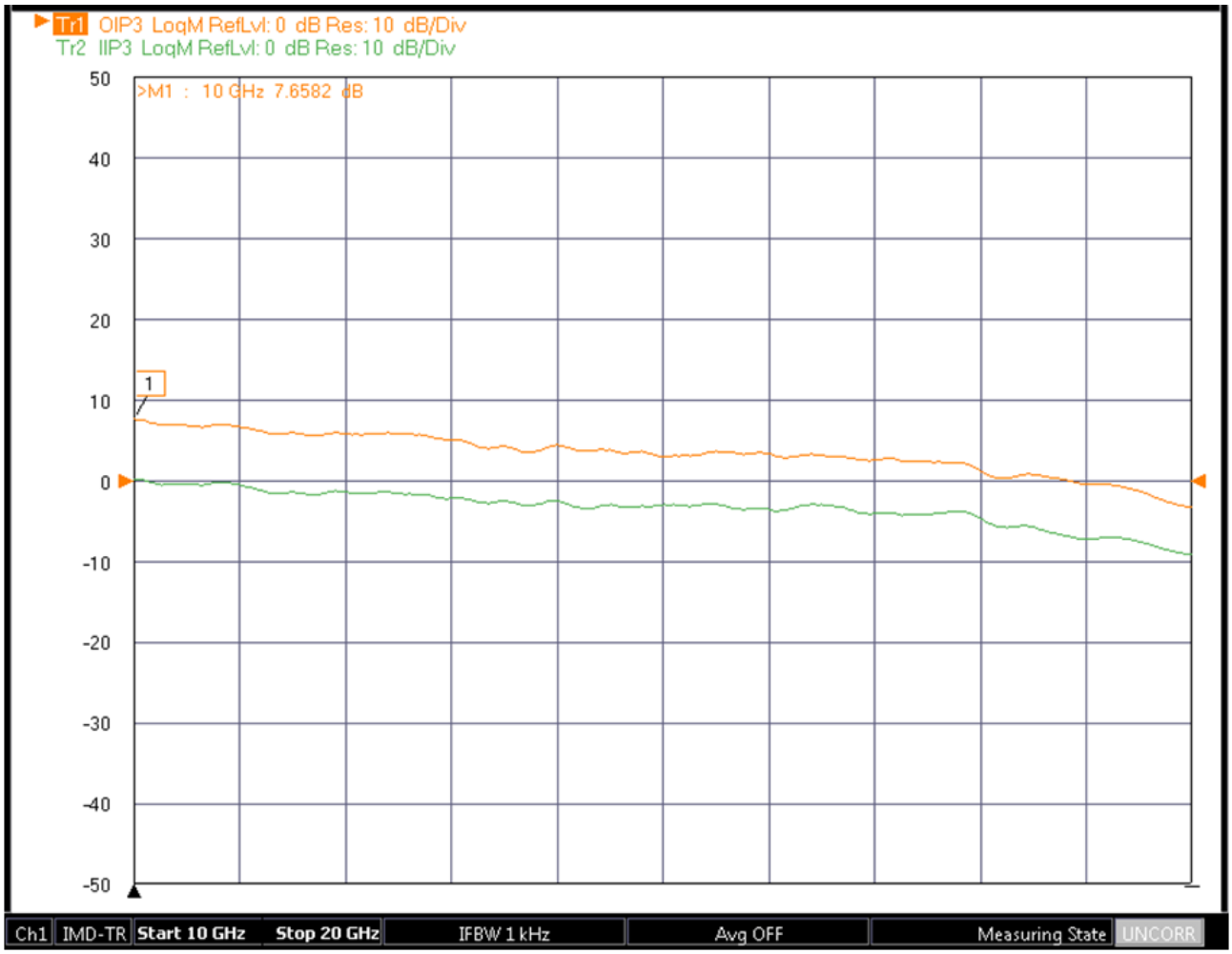


Figure 28-38. IMD Calibration Menu

The Calibration menu when in an IMD mode is shown above. The normalization calibration (requiring only a thru line connection like the receiver calibration) is performed with the first button and the application of the calibration is controlled with the second. This calibration is only used for input-referred tone power and intercept point measurements. An internal source must be defined for tone 1 for the normalization calibration to be available. Otherwise (or if no normalization calibration has been performed in general), input-referred IPn and Tone Power will return the same values as their output-referred counterparts.

To show the effect of this calibration and the behavior of the input- and output-referred parameters, consider the measurement of a reasonably broadband amplifier with about 7 dB of gain across 10 GHz to 20 GHz. The input and output-referred IP3 values for this amplifier are shown in Figure 28-39 (using averages among both the upper and lower products and the upper and lower main tones) and one can see that the difference between the values is the gain of the DUT. If the normalization calibration was not performed or was turned off, the OIPn and IIPn values would be the same (the same is true for input- and output-referred absolute tone values).



IIP3 and OIP3 are plotted here for an example amplifier to show the effect of the normalization calibration to find DUT gain real-time.

Figure 28-39. Normalization Calibration Effect—Broadband Amplifier

An important aspect of this process is that the gain of the DUT is evaluated real-time and used to appropriately generate the IIPn (or input-referred tone power). Thus if one is adjusting the DUT bias, tuning the DUT, or drift in the DUT is occurring, errors from gain misconceptions will not affect the input-referred results.

28-11 IMDView Application—Measurement Setup Hints

The intent of this section is to summarize some of the observations earlier in this chapter in the context of some practical measurement setups. Particularly for very high IPn measurements (or, equivalently, very low absolute product levels), some care on the power and setup plan can be quite important. While it varies with frequency and absolute tone levels, a threshold of difficulty is often in the –75 to –85 dBm product range for –10 dBm main tone output levels.

Targeting Receiver Power Levels

When referenced to the port, the 0.1 dB receiver compression point of the MS464xB and the broadband systems (ME7838x) is around +10 dBm (for center frequencies below a few MHz, it declines to ~ +5 dBm). If the incident tone levels to the instrument are in that range, the receiver IMD products will often be quite large. One does not want to drop the level too low or noise floor effects will dominate. A good target value for tone amplitudes at the instrument port is about –10 dBm (or –23 dBm if using the b1 or b2 access). This level can be achieved using the internal step attenuators or external attenuators. Those attenuators should of course be placed/set prior to performing the receiver and normalization calibrations. This target level is not particularly sharp in definition and there is only minor degradation going 5 dB higher or 10 dB lower. The receiver IP3 value is generally in the 35 dBm range at the port for the MS464xB and in the 30 dBm range for the ME7838x operating at mmWave frequencies.

Source Combining and Isolation

A wide variety of different source combination options have been discussed in this chapter. The choice may be dependent on available hardware or options configurations. External sources may be able to provide higher power in some cases than the base VNA (ranging from –10 dBm for a 70 GHz single source system with attenuators at some frequencies to +13 dBm for a 20 GHz system) and may have lower close-in phase noise for low offset measurements. The internal sources can allow measurements of IMD and S-parameters with a single connection and have very good ALC-immunity which helps reduce the need for additional isolation devices in the chain. In the end, a useful metric for both source isolation and receiver behavior is to look at the IMD products with just a thru line as a DUT. Such a measurement can allow one to optimize output signal levels, combiner and source selections, and other elements of the physical setup.

For mmWave measurements, internal or external sources can again be used but the combiner must be external. The same thru-line check measurement can be very useful for these frequencies as well.

Other Settings Selections

Delta selection was discussed earlier and the given test protocol may not give the user flexibility with what Delta to use. If a choice is available, larger Deltas will generally be easier to measure as source noise skirts are less of an issue. Round numbers can sometimes be an issue for spurs but the instrument's spur avoidance algorithms will help that situation.

As the product levels or Delta decrease, so should the measurement IFBW (at least of the products). Generally the IFBW should be 10x to 100x smaller than Delta as discussed previously to avoid tone contamination. As the product levels fall below –70 dBm, the IFBW should generally be at or below 100 Hz to avoid noise floor influence.

Gain Ranging (IF gain settings) can be left in Auto for offsets larger than several MHz, but for smaller offsets or when the main tone levels into the receiver are relatively high, going to manual is desirable. Low is generally a recommended setting if the main tones incident on the port are below –10 dBm but above –25 dBm.

On the receiver and normalization calibrations, interpolation and extrapolation are automatically used if the frequency list is changed after the calibration was performed. An error is only generated if the calibration and measurement frequency ranges are disjoint. The errors can usually be small (<< 1dB) if the calibration frequency list was relatively dense. If the calibration list was sparse (and particularly if there is a lot of mismatch in the setup), interpolation errors can exceed 1 dB. A recalibration is advised if there is a significant change in the frequency list (e.g., measurement points are more than 100 MHz from calibration points for test setups on the scale of a few meters in size including cables).

In terms of the calibrations, an ideal sequence of events would be something like:

1. Use the IMD configuration dialog to set up the frequencies of interest and, for swept frequency mode, the response parameters of interest. Turn on IMD.
2. Perform a user power calibration. The frequency range of the power calibration will automatically cover the product ranges. In mm-wave situations, multiple power meters/sensors may be needed and the instrument will ask that they be connected to the power reference plane at the appropriate time. All required meters must be on the bus when the calibration starts.
3. Perform a receiver calibration (may not be needed for relative measurements such as IMD in dBc terms or asymmetry, very important for absolute power and intercept point measurements). Receiver calibrations are automatically performed over a wide enough range to cover the product frequencies if IMDView in on and configured at the time.
4. Perform a normalization calibration (if needed because of the use of input-referred parameters).
5. Measure the DUT(s).

28-12 IMDView Application—Uncertainties

There are a number of uncertainty mechanisms of interest for the IMD measurements and most have been discussed elsewhere in this chapter. To summarize these issues and to put some coarse bounds on their quantitative effects:

Receiver Linearity and Residual IMD Products

Due to the linearity of the MS464xB VNA and the mmWave module (374xx and MA25x00A) receivers, this is usually not a dominant factor unless the incident levels are too high. With –10 dBm incident tones (ordinary coupled ports), the residual products are generally < –100 dBm. Thus if one is measuring –80 dBm tones from the DUT, the receiver induced uncertainty would be under 1 dB.

Source Linearity and Residual Products

If the sources communicated with each other, there could be remodulation and spurious IMD generation on the source side. Because of the system structure, this effect should be insignificant for product levels above –100 dBm with the use of even a relative low isolation (<10 dB) combiner. These levels will increase for Delta values below a few kHz.

Receiver Calibration Accuracy

If reading out the product values in dBm or converting to an intercept value, the absolute power accuracy is important and is dependent on the power calibration behind it. Using the power calibration techniques discussed, this uncertainty should normally be under 0.5 dB. The base factory ALC calibration accuracy is on the order of 1 dB to 2 dB but the more significant issue is the inclusion of the effects of cabling and combiner (if external) between the port and the DUT input reference plane.

It will not be included in the below analysis but the power accuracy can have a significant effect on the operating point of DUT in some cases and that is very DUT-dependent.

Return Loss

Match interactions between the combiner and DUT or between DUT and receiver can have impacts on the order of the product of the reflection coefficients (in linear terms). Padding on the receiver or combiner (before the reference planes) can help reduce these effects if dynamic range allows. Also, this effect is more important on the absolute parameters (product power or intercept point) than the relative parameters (IMD in dBc terms or asymmetry). For absolute parameters where the DUT has 10 dB return loss levels and the port reference planes have 15 dB return loss (combiner output or receiver input), the match-induced uncertainty can be up to ~0.5 dB. For the relative parameters, it will generally be <0.1 dB unless Delta is very large and the electrical length of the setup is large (the issue is if a standing wave forms whose period is comparable to Delta such that a product and a tone see different portions of that standing wave structure).

Noise Floor

The noise floor of the instrument can be the limiting factor in some cases but the high dynamic range of the system helps in this regard. In a 10 Hz bandwidth, the noise floor is typically in the –110 dBm to –130 dBm range depending on frequency and the use of a lower IFBW can help further. As the Delta descends below 10 kHz, there are additional noise floor factors related to the skirt noise of the sources and the internal LO. Some synthesizers have lower phase noise levels that can help with very low Delta measurements.

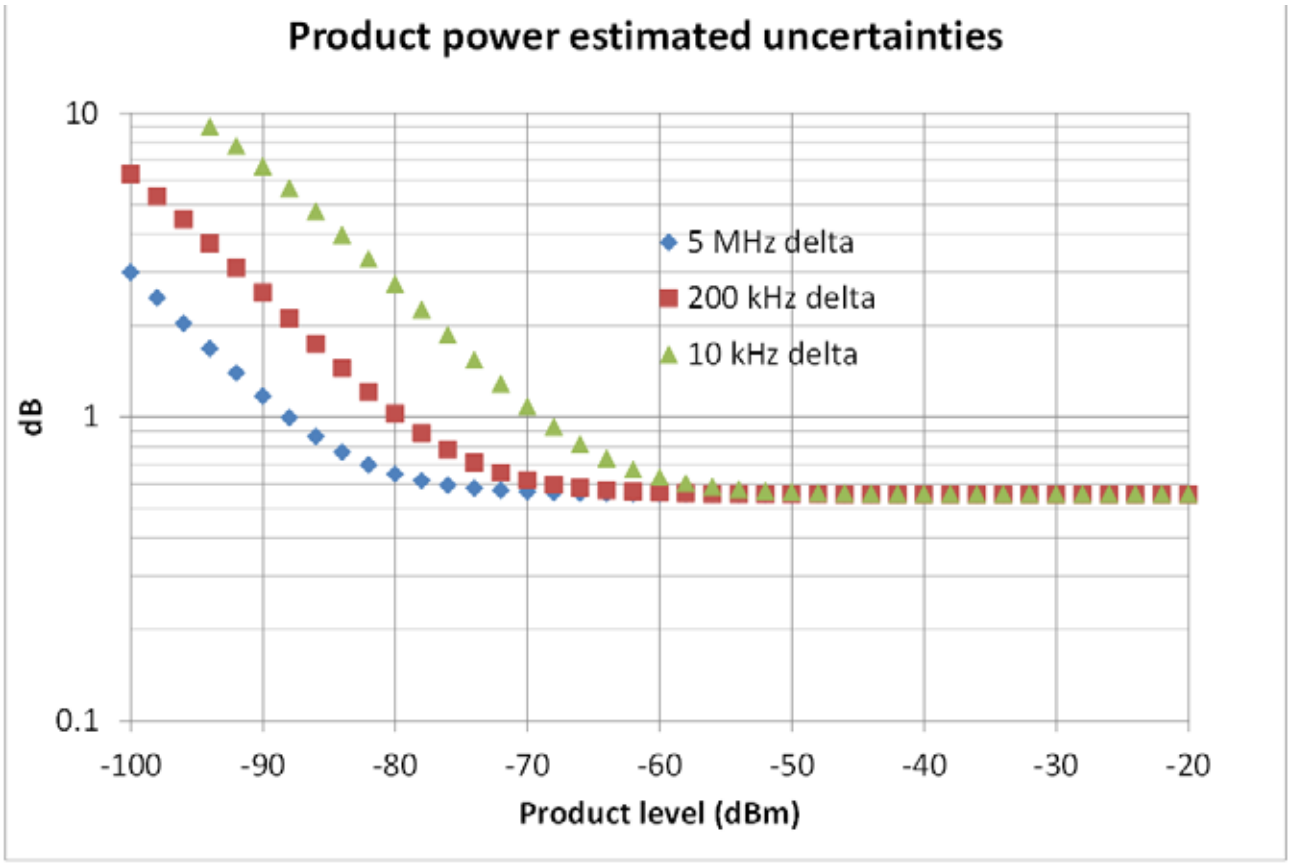
Gain Analysis and Input-referred Measurements

The gain measurement used to make input-referred measurements is not match-corrected so an additional mismatch term will add for those cases. Also, the frequency variation between main tone and product locations is not considered. If the DUT gain changes rapidly with frequency, and Delta is large, then there could be additional uncertainty terms. For most DUTs and Deltas ≤10 MHz, this normally is <0.1 dB of effect.

While it is not practical in this document to produce graphs to illustrate the uncertainty levels under all possible conditions, it may be useful to explore some subset of the parameter space. Consider measurements bounded by the following:

- Main tone levels into the VNA receiver port (non-mmWave setup) at –10 dBm
- DUT match is –10 dB. Reference plane match from the rest of the system is –15 dB.
- Two VNA internal sources and internal combiner are used (Option 31 and Option 32).
- DUT gain is flat with frequency over the scale of Delta.
- User power calibrations and receiver calibrations are performed and no interpolation is used.
- A 10 Hz IFBW is assumed.
- Gain-ranging is assumed to be in an appropriate state (manual low or off for smaller Deltas).
- Absolute product power will be used as a metric.

A plot of estimated product power uncertainty for these conditions is shown in Figure 28-40. If an intercept point is the response, the uncertainties will rise somewhat due to the uncertainty in the main tone power adding in. Since that is a high level signal, its uncertainty is dominated my mismatch and power calibration effects and is generally < 0.5 dB. If a relative measurement is being performed (IMD in dBc terms or asymmetry), uncertainties will be lower than shown in the graph since mismatch effects largely cancel.



An example plot of product power uncertainty vs. product level for several values of Delta are shown here. The calculations are based on the assumptions in the text and the uncertainty values can be higher or lower if those conditions are not met.

Figure 28-40. Product Power Estimated Uncertainties

28-13 Advanced Topics—Multiple Source Control with IMDView

The standard receiver configuration selection within the IMD configuration dialog includes the standard mode as well as various 3739-based mmWave modes. The 3738-based modes and the VDI/OML-based mmWave modes are not available with IMD. Multiple source control, however, is available as an overlay. A typical use case where this might be helpful is in a mixer IMD measurement (for which a fully manual measurement was discussed in Section 21-15 of this guide). The tone positions (and power levels, etc.) are easy enough to define for the input of the DUT but there must be a way of configuring the receiver location correctly to account for frequency translation and, perhaps, to provide for LO control. There are few things to observe about this measurement class

- The sources used for IMD tones in the IMD configuration dialog will not be available for control within multiple source control.
- The band definition must contain the IMD frequency range of interest and should represent the input frequency of the DUT.
- Any other sources available can be used as LO.
- The receiver and receiver source equations should be used to point to the output frequency of the DUT.

To explore this further, consider an example where the DUT is a downconverter with an input frequency range of 16 GHz to 18 GHz, a fixed LO of 13 GHz, and an output frequency range of 3 GHz to 5 GHz. The IMD configuration dialog would be setup for a 16 GHz to 18 GHz frequency range with the desired number of points, Delta value, response variables, etc. We will use external synthesizer 2 as the LO for the example and will configure the receiver and receiver source equations with an offset of –13 GHz to indicate the DUT conversion. The complete multiple source overlay equations along with the example data are shown in Figure 28-41.

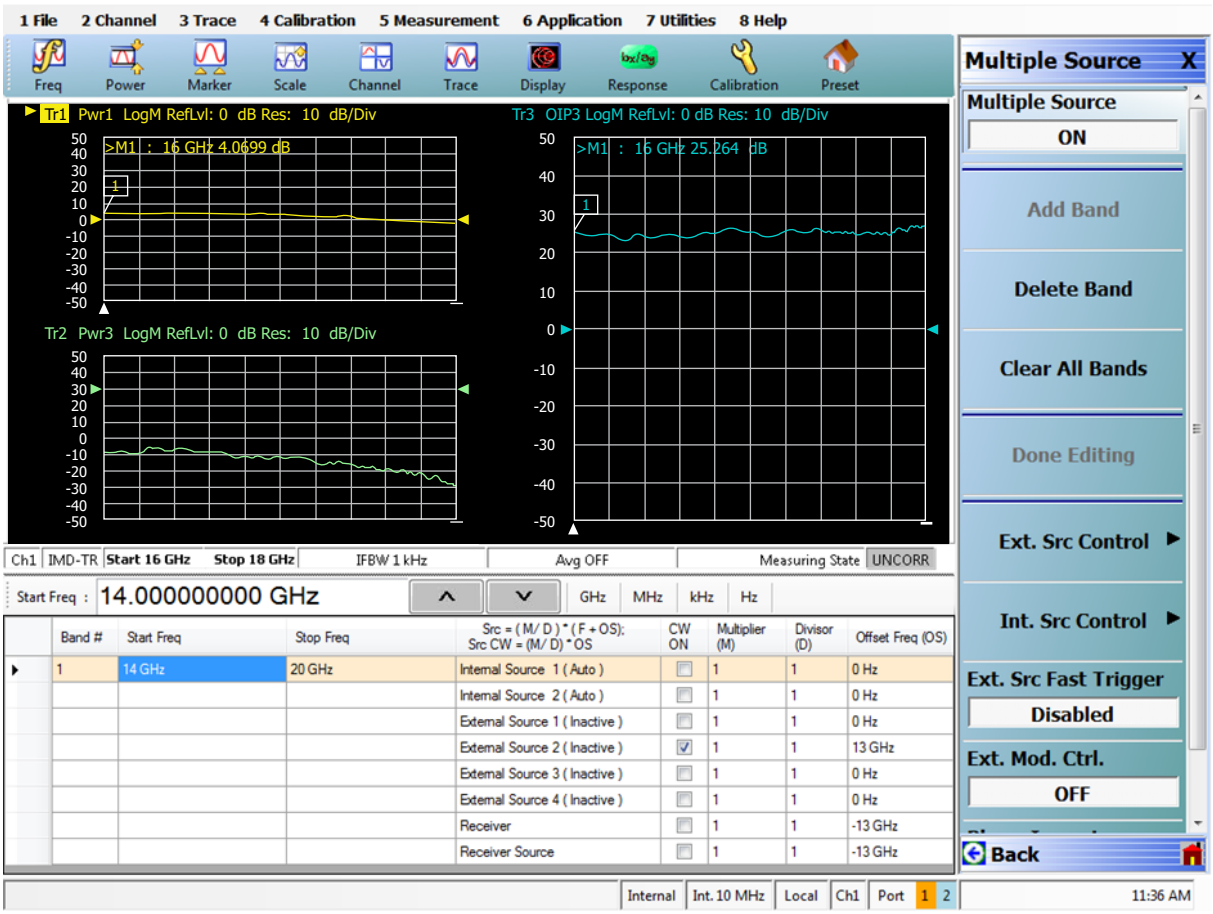
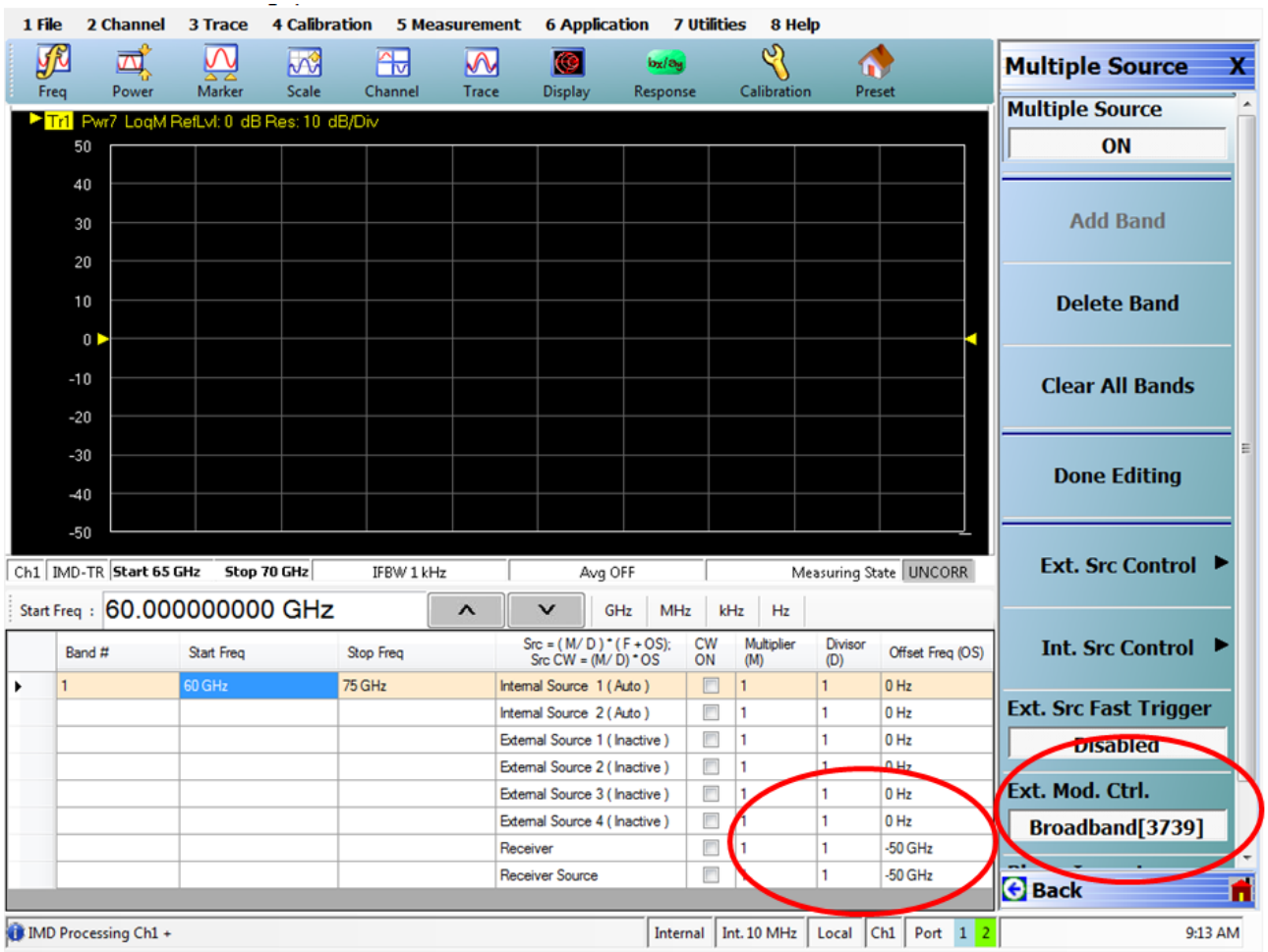


Figure 28-41. Example Mixer IMD Measurement Using Principle of Multiple Source Control Overlay

Hints and Other Comments

- For the equation setup, it may be helpful to just think about how f_c will be processed by the DUT. Does it upconvert f_c by X GHz or does it downconvert it by Y GHz, etc.? The various measurement locations for the IMD tones and products will be handled by the IMD engine if multiple source control places the center correctly.
- Receiver calibrations should be performed prior to multiple source activation and over a wide enough range to cover the output frequencies as well as (usually) the input frequencies.
- The receiver source equation just serves to index the receiver calibration so, in this measurement, it is important that it be pointing in the same place as the receiver itself.
- If there are any conflicts between the multiple source entries and the IMD configuration entries (frequency range, etc.), IMD status will take precedence and multiple source control will be turned off with a warning message.

Multiple source overlay is also possible using the broadband/mmWave modules (374xx and MA25x00A). The receiver configuration selection in the IMD configuration dialog and that in the Ext. Mod. Ctl. dialog of multiple source control must match. In the example of Figure 28-42, this happens to be with the 3743A/AX broadband module. For the example measurement, the DUT is a downconverter with an input range of interest of 65 GHz to 70 GHz and a fixed LO of 50 GHz. The measurement of interest was upper 7th order product power. The equations are shown in the figure. Note that the band definition exceeds the frequency range being measured by IMD and that is acceptable (it cannot be smaller than the measurement range).



An example mmWave IMD measurement using multiple source overlay is illustrated here. The receiver equations, as with all of these overlay setups, focus on where f_c translates to.

Figure 28-42. Example mmWave IMD Measurement using Multiple Source Overlay

Fully manual multiple source IMD measurements (like those discussed in Section 21-15) are still possible even in instruments equipped with Option 44. The calculated response variables will not be available (most tone powers with relative IMD (dBc) measurements possible using trace math normalization) and there will be no control orchestration. To facilitate these measurements in instruments equipped with Option 32, there is an additional control to allow one to switch source 2 output to route through the combiner to port 1. This switch, under the Power/Other Setup menu, is available when IMD is off. When IMD is on, the button is grayed out as shown in Figure 28-43. Note that the Apply Slope function is disabled as well when IMDView is on.

When this function is used in a manual multiple source measurement, the src2 → port1 ALC calibration will automatically be activated and user power calibrations employing that path will be available. Note that when using both internal sources, they must be manually both made 'Active' using the Int. Src. Control menu within multiple source control.

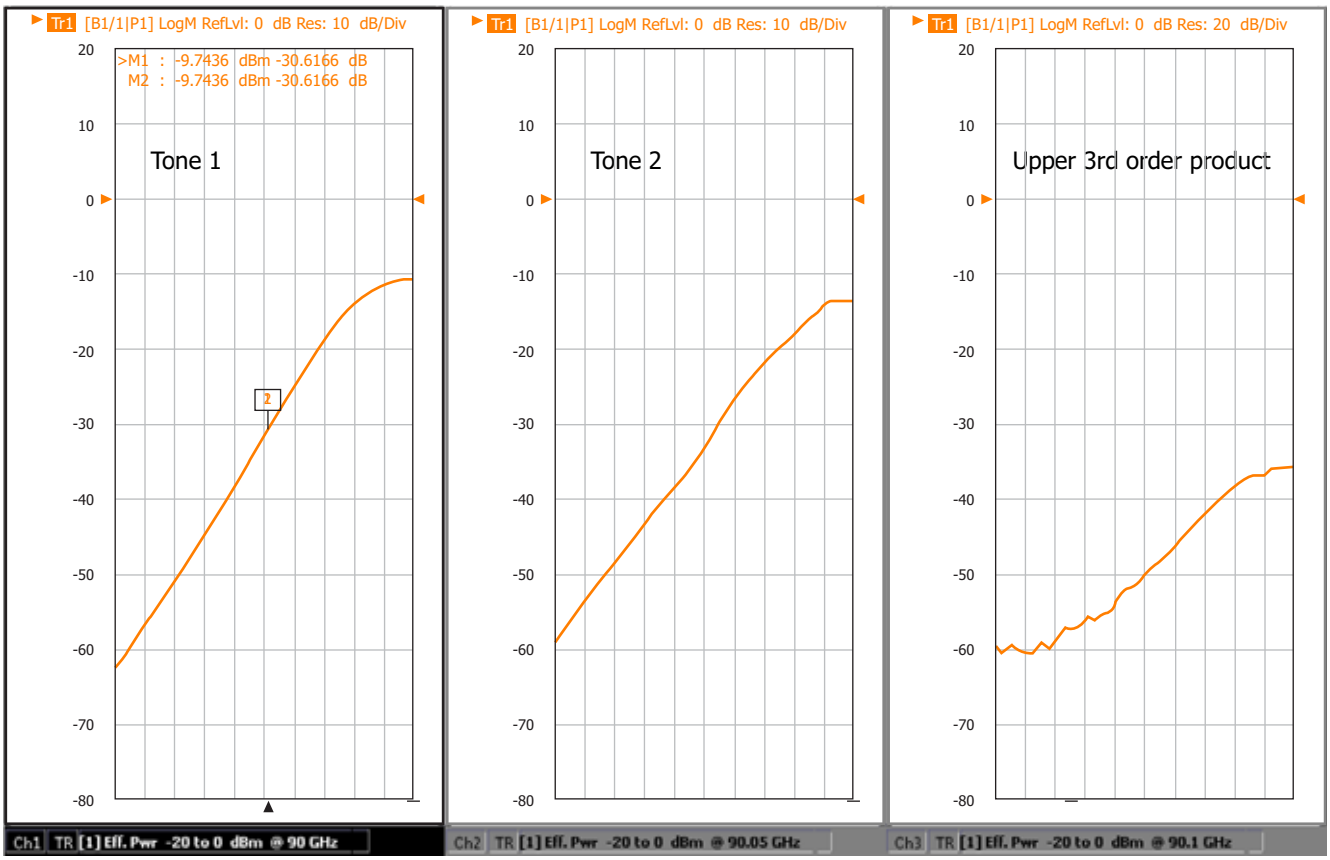


The power/other setup menu is shown here when IMDView is ON (on a system with Option 32). The SRC 2 Out Redirection button is greyed out because it can only be controlled from the configuration dialog when in IMD. For manual IMD measurements (not using IMDView but configured ONLY with multiple source), this button will be available and can be used to employ the redirection and the internal combiner.

Figure 28-43. Power Setup Menu—IMDView ON

In the fully manual mode, more elaborate measurements are possible including those with the VDI/OML mmWave modules, power sweep product measurements, Delta-sweep measurements and others. As examples, a power sweep mmWave measurements (using the 3743A/AX modules and the 3739B test set) are shown in Figure 28-44. Channel 1 was configured to look at the lower main tone, channel 2 was configured to look at the upper main tone and channel 3 was configured to measure the upper 3rd order IMD product. The measurements were around 90 GHz with a power sweep range of –20 dBm to 0 dBm in this particular case.

Another example is that of Delta sweep where f_c is constant but the tone spacing is varied. This can be a powerful tool when looking at DUT video quasi-linear responses (e.g., with bias networks, thermal or trapping effects). For an example measurement, the Delta was swept from 10 kHz to 100 MHz in a log sweep fashion with f_c fixed at 12 GHz to measure an amplifier. Two channels were used to look at the lower and upper 3rd order products. This data is shown in Figure 28-45 along with the multiple source equations for the upper product.

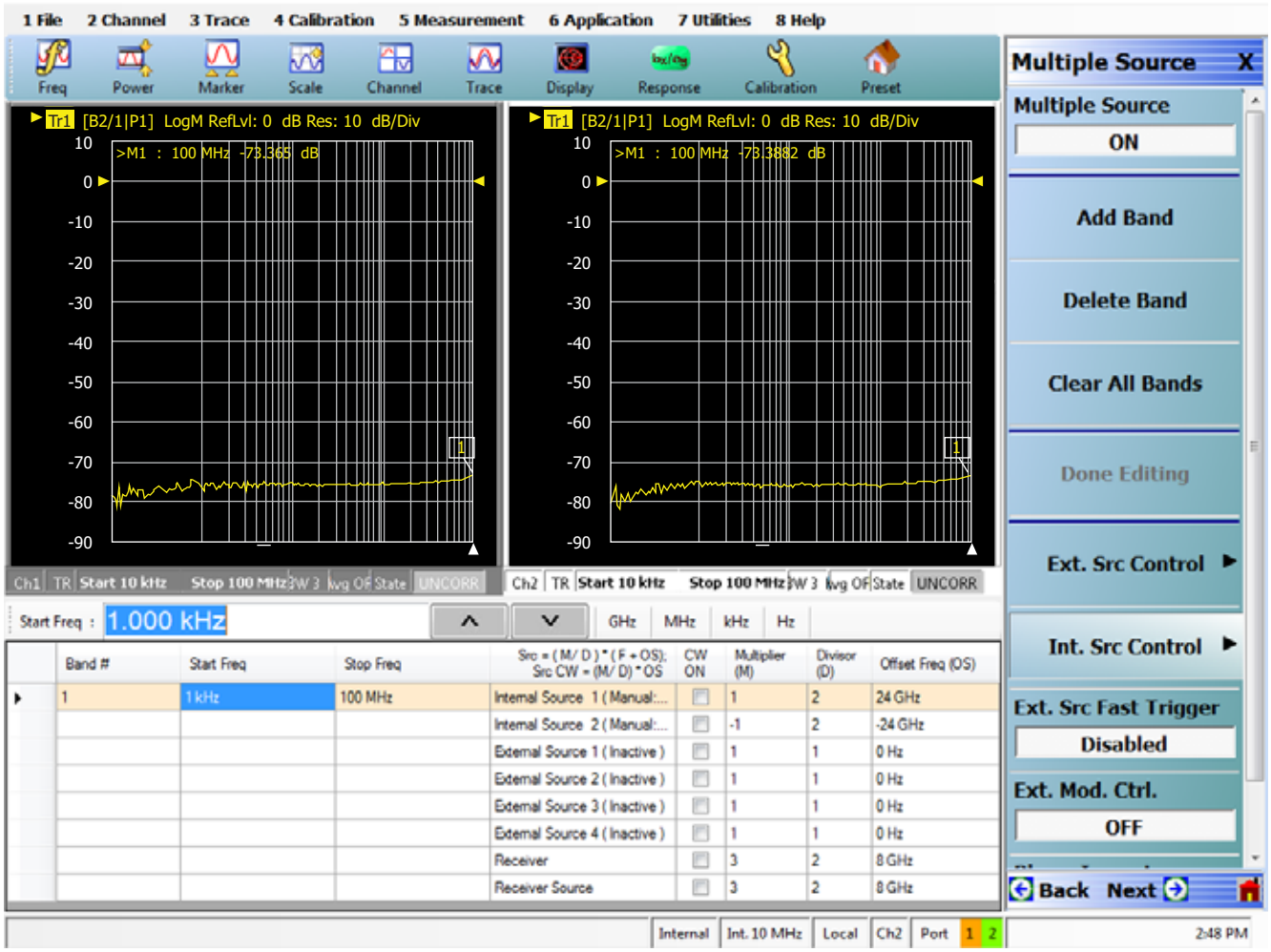


An example power sweep measurement of tones and an IMD product for a mmWave configuration is shown here. The manual multiple source control method was used for this sweep type.

Figure 28-44. Example Power Sweep Measurement of Tones and IMD Product—mmWave Configuration

The equations used in the measurement of Figure 28-45 may appear unusual. Since the main tones are defined to be ±Delta/2 relative to f_c and Delta is now the runner variable, the divisor of 2 is required. The third order product is 3*Delta/2 above f_c so the 3/2 pre-factor is required. The offset values in the equations are adjusted to the Multiplier*Offset/Divisor is equal to the desired f_c for all equations. For this measurement, two internal sources were used so both sources were made ACTIVE using the Int. Src. Control button visible on the lower-middle right of Figure 28-45. A low IFBW (3 Hz) was used for these measurements to accommodate effects of the small Delta values during part of the sweep. Segmented sweep can also be used to more dynamically change IFBW with Delta.

This particular DUT shows very little asymmetry between products. There are some differences at the lower Delta end but that is from noise floor. There is a small deviation at 10 MHz Delta related to the system reference spur. In this manual multiple source method, the spur avoidance functions related to Delta and IF and/or system reference are not available.



An example Delta-sweep measurement is shown here, orchestrated with manual multiple source control, to look at upper and lower 3rd order products of an amplifier vs. Delta.

Figure 28-45. Example Delta-sweep Measurement—Multiple Source Control

Of course, the active channel mixer setup feature can also be used in conjunction with IMDView but, since that is designed for the Transmission/Reflection (S-parameter) mode rather than the IMDView mode, not all operations make sense. For example, a conversion calibration such as Receiver/Norm cannot be applied or performed while making IMD measurements. Still, the mixer setup tool can be used to set up multiple source equations that will be used in overlay with IMDView.

For IMDView to be able to process the equations correctly, the source equations must be in their default state. This means the X-axis variable selected in mixer setup must correspond to the driving port.

The situation will get complicated and some manual Multiple Source editing may be needed if an external source is used as the driving signal for the mixer measurement (it can certainly be an LO signal, just not the RF or IF driver). Port swaps may also introduce issues.

As an example, a downconverter will be configured with a 10–12 GHz input range, a fixed external 9.5 GHz LO (not under VNA control) and an IF of 0.5–2.5 GHz. The active channel mixer setup is shown in Figure 28-46. After this is setup, IMDView was entered and a collection of parameters were selected for measurement. Note that the frequency range for IMDView is the input range for the DUT (and this is enforced by how mixer mode works). The IMD configuration dialog is shown in Figure 28-47.

IMDView does not alter the multiple source equations generated by the mixer setup process but, as discussed, just superimposes its local frequency plan onto those equations. Thus if one looks at the multiple source equations (see Figure 28-48), one sees the mixer plan. One might be disturbed by the –9.5 GHz offset on Source 2 (this is a dual source system), but this is only used during enhanced match calibrations which are not in play when IMDView is active.

A power calibration and a receiver calibration are all that would normally be used with such a measurement. The receiver calibration requires some care since, in its current state, the source and receiver would not be aligned for that calibration. It is recommended that a receiver calibration be performed with multiple source turned off (IMDView still on) and the frequency range changed to encompass the DUT output range. The range can then be reset for the actual measurements.

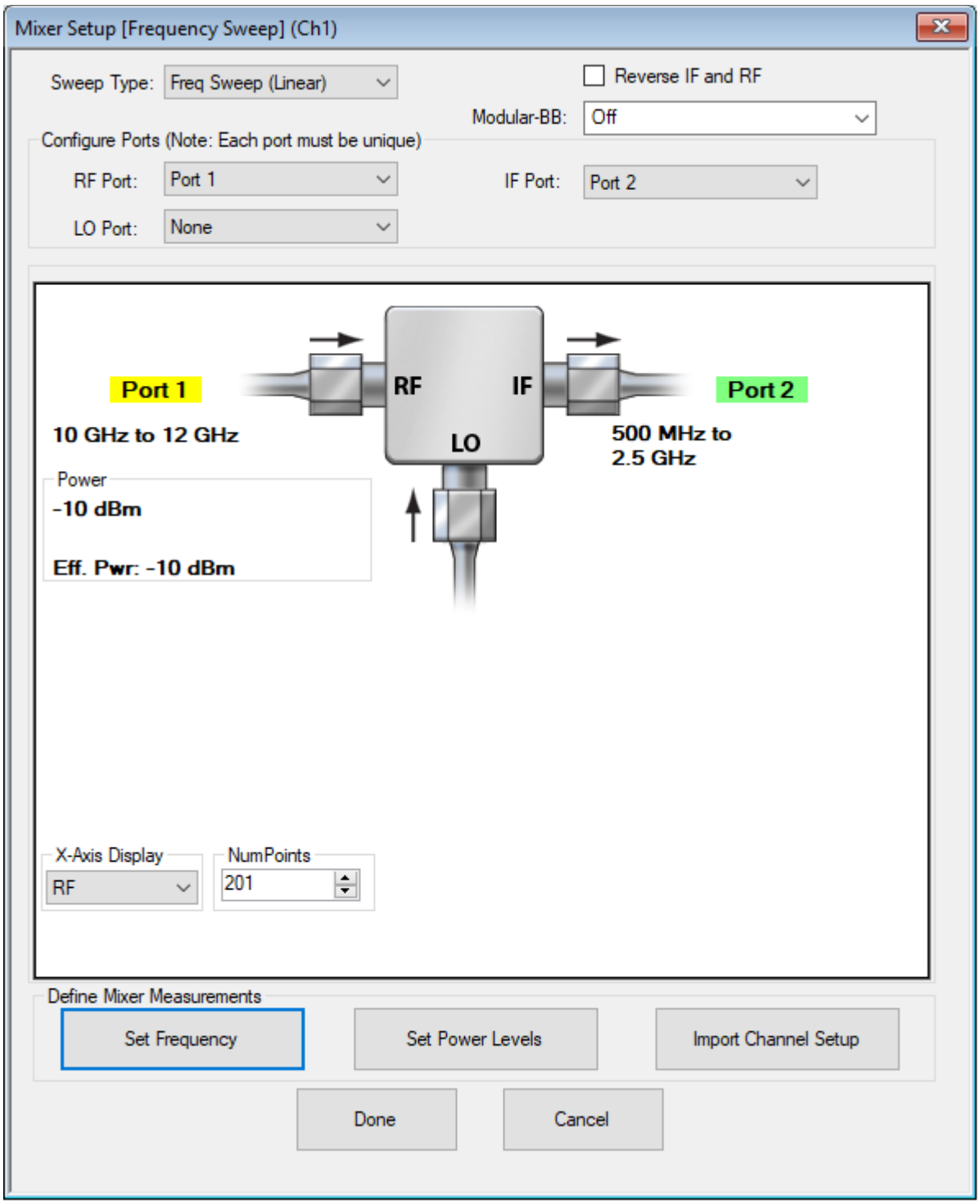


Figure 28-46. Mixer setup dialog for the active-channel-dialog-based IMD measurement.

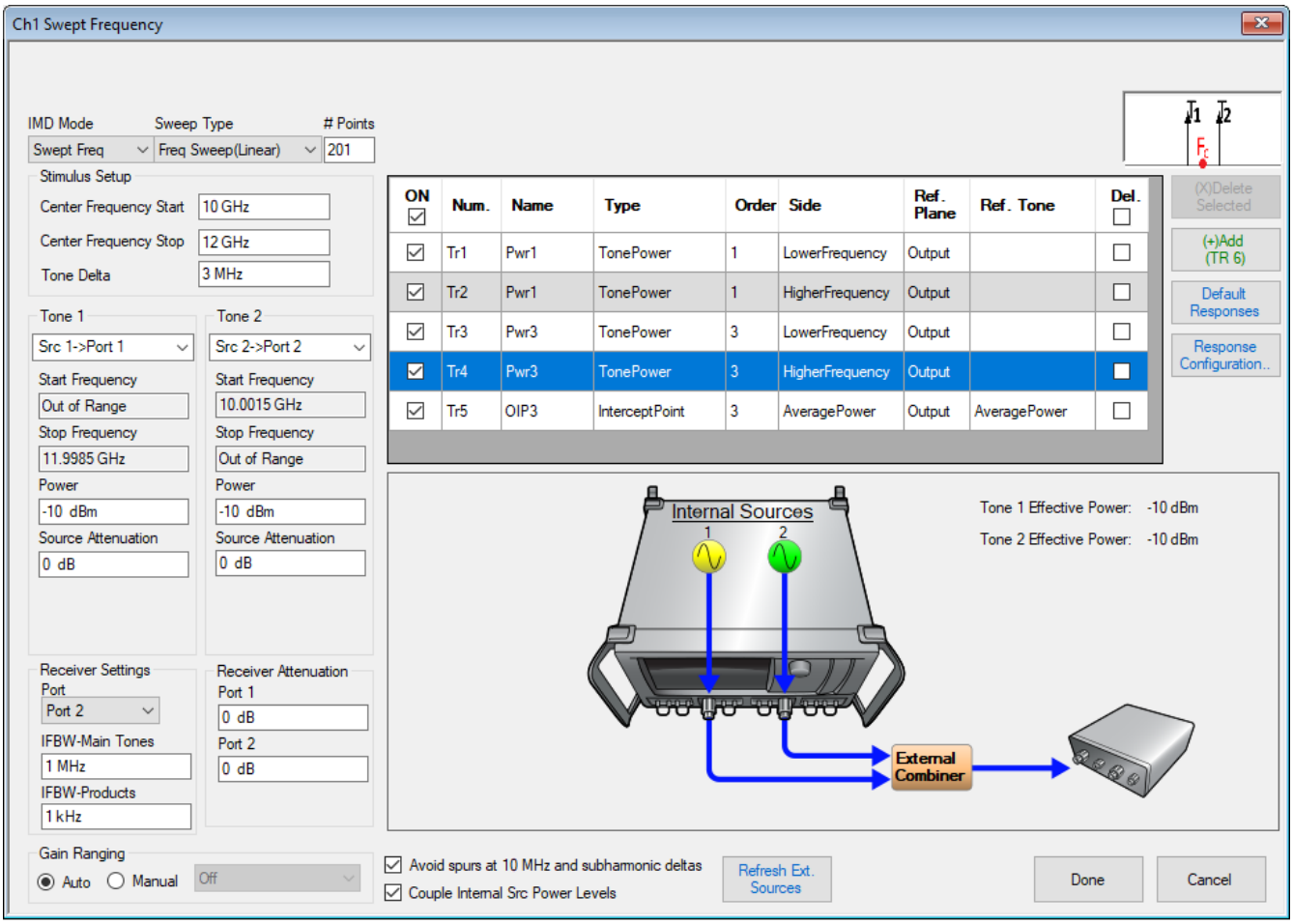


Figure 28-47. The IMD configuration for the example mixer measurement.

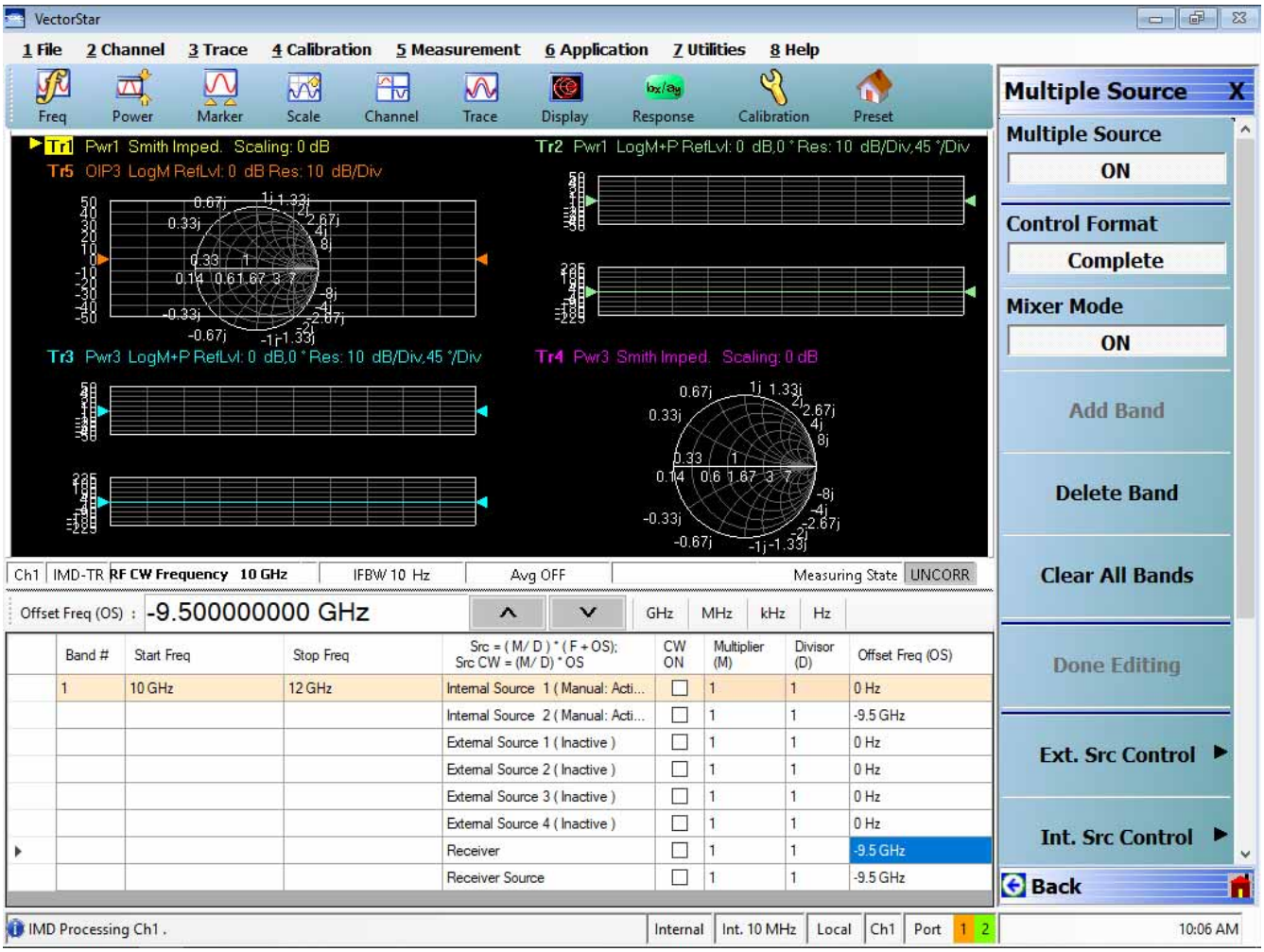


Figure 28-48. The multiple source equations after the settings in the previous two figures have been activated.

28-14 Summary

The IMDView application (Option 44) in the MS464xB provides tools to orchestrate and simplify a variety of intermodulation distortion measurements including CW-like SpectrumView analysis and swept frequency measurements. Many different configurations are possible and some information has been provided in this chapter to help optimally set up the measurements, get the desired responses, perform any needed calibrations and minimize uncertainties. Considerable flexibility is also available for mmWave measurements and more manual measurements (including for frequency converting devices, higher mmWave bands, Delta-sweeps, etc.) can be set up with the help of multiple source control.

28-15 References

1. "Intermodulation distortion", *Anritsu Application Note 11410-00213*, Apr. 2000 and "Intermodulation distortion measurements," *Anritsu Application Note 11410-00257*, Sept. 2000.

2. W. J. Warren and W. R. Hewlett, "An analysis of the intermodulation method of distortion measurement," *IRE Proc.*, vol. 36, Apr. 1948, pp. 457–466.

3. F. Assadourian, "Intermodulation distortion and efficiency analysis of multicarrier repeaters," IRE *Trans. Comm. Systems,* Vol. 8, Jan. 1960, pp. 68–71.

4. J. S. Kenney and P. Fedorenko, "*I*dentification of RF power amplifier memory effect origins using third-order intermodulation distortion amplitude and phase asymmetry," *2006 Int. Micr. Symp. Dig.*, June 2006, pp. 1121–1124.

5. *J. C. Pedr*o, J. P. Martins, and P. M. Cabral, "New method for phase characterization of nonlinear distortion products," *2005 Int. Micr. Symp. Dig.*, June 2005, pp. 971–974.

6. J. Martens and K. Noujeim, "IMD phase analysis and mmWave frequencies," *2012 IEEE MTT-S Int. Micr. Symp.* Dig., June 2012, pp. 1–3.

7. "Millimeter-wave intermodulation distortion measurements," *Anritsu Application Note 11410-00684*, Oct. 2012.

Chapter 29 — Fast CW (Option 46)

29-1 Introduction

There are a number of measurement applications where large amounts of data (often measured as fast as possible) at a single frequency/power level are needed. These applications usually involve collecting the data as some state of the DUT is changing (angle of an antenna in a range, state of a phase array module, bias state of a device, etc.) and that state can be changed fairly rapidly by an (often) external controller. As one might expect, synchronizing this DUT state with the data stream can be an important topic. Option 46 provides two separate Fast CW types that differ in terms of acquisition rates and how the data is output and processed:

Standard Fast CW

- This uses the standard MS464XX IF system which allows maximum acquisition rates of ~200,000 measurements/sec (and as slow as 1 measurement/sec). This mode is only available remotely to maximize speed and the display does not update in this mode.
- There are two main sub-modes: real-time streaming over the remote interface with unlimited data collection capabilities, and a buffered collection with a maximum buffer size of ~40 million measurements for a single S-parameter. (Actual size allowable may vary based on memory available on the system)

Advanced Fast CW (Requires Option 35)

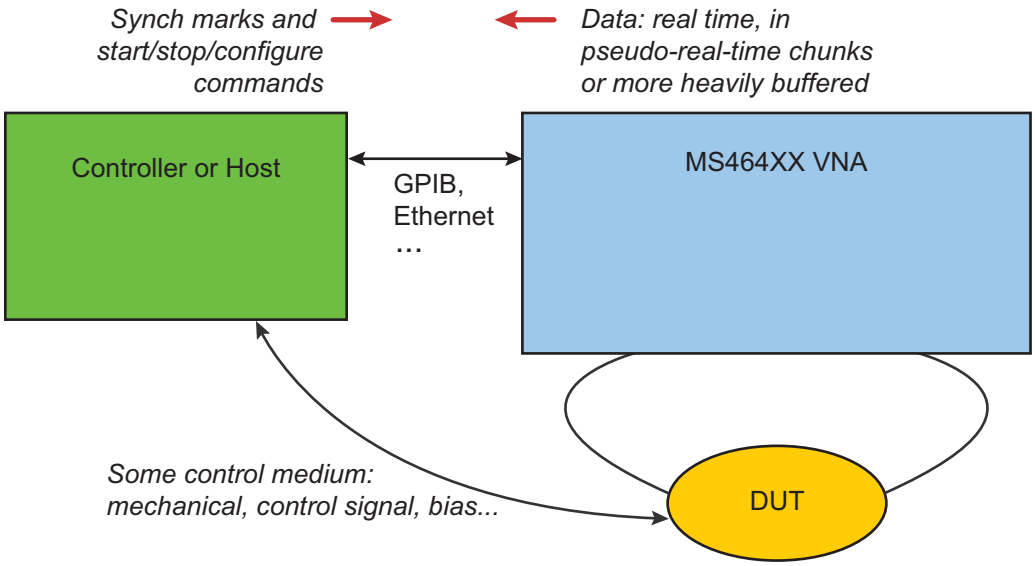
- This uses the high speed digitizer option to reach maximum acquisition rates of 100 million measurements/sec (adjustable to as slow as <4 million samples/sec). This mode is always buffered and is accessible from both the user interface and remotely.
- Raw data exporting is available as is pre-processing over variable measurement widths and time ranges.
- The maximum buffer size is 50 million measurements in each of four receiver channels (200 million measurements total) without the memory addition of Option 36 and 200 million measurements per receiver (800 million measurements total) with Option 36.

This chapter will discuss how these modes (and their various sub-modes work) along with some applications information.

29-2 Standard Fast CW

Note	Standard Fast CW functionality only requires Option 46, and is only available remotely.

The standard Fast CW mode uses the basic instrument acquisition engine with the data paths simplified to increase speed and a number of additional commands introduced to control data flow and offer some configuration choices. The block diagram of the measurement is shown in Figure 29-1 on page 29-2. The sources and receivers are programmed before the Fast CW process begins and triggering can be used to sequence between different Fast CW epochs (i.e., to move between frequencies, power levels, switch positions in the instrument, etc.). Once the Fast CW process begins, the measurement acquisition runs pretty much as fast as possible for a given bandwidth and either streams data out over the remote interface or dumps the data into an internal buffer for simplified access and handling. Synchronization within the data streams (which in some external literature is called 'triggering') is done with a data marking process that is similar in concept to that discussed in the pulse measurements chapter of this guide (Chapter 25). Conventional per-point triggering is permitted but will reduce the maximum speed to on the order of 10000 measurements/second. There is, however, a process for marking time within the data stream that will support the higher data rates which is discussed later in this section (a flag is placed in the data stream based on a separate command).



The basic measurement configuration of the standard Fast CW mode is shown here. This is a remote-only mode and allows streaming or buffered data delivery.

Figure 29-1. Standard Fast CW Mode Measurement Configuration

Streaming Sub-mode

In this mode, data is made available as fast as possible over the remote interface with the speed usually limited by the remote interface itself. The 'chunk' size of data made available for transfer is adjustable from 1 to 500 measurements (with the definition of that term to be made clear shortly) to allow the user to optimize to their remote application in terms of transfer overhead, processing needs, etc. Although slightly unconventional, the data is shoved out onto the bus as fast as it is available and it is the responsibility of the user's controller to grab the data (often with something like an fscanf command in C or various commercial control products) as it is available or samples may be missed. It is for this reason that the 'chunk' size is variable within a range. The size can be optimized for the practical speed and overhead requirements of the user's interface.

This sub-mode is particularly useful when processing or control decisions have to be made in the middle of a measurement sequence. An example would be curtailment of a bias excursion once the DUT gain reached a certain level. Another example would be the halting of an antenna pattern measurement when the received power dropped below a certain level.

In this sub-mode, only one channel is used so all other channels are made inactive during the Fast CW measurement. Those channels will be restored and made active (if they were active before) when one exits the Fast CW mode. Most GPIB commands are blocked when in Fast CW because they interfere with the data transfer process (a detailed list is available in the programming manual) but a general rule is that all setup tasks should be performed prior to activating Fast CW and, if further setup changes are needed, one should exit Fast CW, make the changes, and then return.

Note	There is no firm limit on how much data can be collected in the sub-mode since it is streaming data. The practical limit will generally be set by memory or capability of the remote controller.

There are two further types within the streaming sub-mode; Type 1: One response parameter and Type 2: Base unratioed parameters

Type 1: One Response Parameter

In this case, the output is a single complex value representing the response parameter (which can be an S-parameter or a user-defined parameter) in the first active trace ('active' in the sense that it is turned on in Trace Management). In the preset setup (four traces: S11, S12, S21 and S22), S11 would be the parameter that is output. The response parameter is always transmitted to the remote interface in the form of a complex float in binary form so the transferred block would be of the form:

```
#18<4 bytes for real><4 bytes for imaginary><EOL>
```

when a single measurement was requested. The format here is the standard binblock format discussed in the *MS464XX Programming Manual*—10410-00322:

- It leads with a #.
- The next digit indicates how many digits will follow next (the next part describes the chunk size)
- The next digits (length specified in the previous bullet) indicate how many bytes will be in the block. In the above example, there are 8 bytes.
- The end-of-line indicator or other selected terminator.

If, for example, 3 measurements were selected for the 'chunk', then the data would come as the form:

```
#224<4 bytes for real1><4 bytes for imaginary1><4 bytes for real2><4 bytes for
imaginary2><4 bytes for real3><4 bytes for imaginary3><EOL>
```

Note	There is a separate command that can configure this header to be fixed length, see the MS4640X Programming Manual for more information. **Also, the byte ordering by default is LSB first within each four byte value.**

Type 2: Base Unratioed Parameters

In this case, the output consists of three complex values representing reference parameter for the current driving port (a1 or a2) and the two test parameters (b1 and b2). The current driving port is set by the first active trace ('active' in the sense of being turned on in trace management). The term 'measurement' in the earlier discussion still refers to one acquisition-set but now that is represented by three complex values. The parameters are always transmitted to the remote interface in the form of three complex floats in binary form so the transferred block would be of the form:

```
#224<4 bytes for real_a><4 bytes for imaginary_a><4 bytes for real_b1><4 bytes for
imaginary_b1 ><4 bytes for real_b2><4 bytes for imaginary_b2 ><EOL>
```

when a single 'measurement' was requested. The format here is the standard binblock format discussed in the *MS464XX Programming Manual*—10410-00322:

- It leads with a #.
- The next digit indicates how many digits will follow next (the next part describes the chunk size)
- The next digits (length specified in the previous bullet) indicate how many bytes will be in the block. In the above example, there are 24 bytes.
- The end-of-line indicator or other selected terminator.

If, for example, 2 ‘measurements’ were selected, then the data would come as the form

#248<4 bytes for real_a_1><4 bytes for imaginary_a_1><4 bytes for real_b1_1><4 bytes for imaginary_b1_1 ><4 bytes for real_b2_1><4 bytes for imaginary_b2_1><4 bytes for real_a_2><4 bytes for imaginary_a_2><4 bytes for real_b1_2><4 bytes for imaginary_b1_2 ><4 bytes for real_b2_2><4 bytes for imaginary_b2_2><EOL>

In the 4-port case, if ports 1 or 2 are driving, a1=a2 will be provided (shared reference coupler). If ports 3 or 4 are driving, a3=a4 will be provided.

More details are available in the programming manual, but here is a synopsis of commands relevant to the streaming sub-mode.

:CALCulate:FCW[:STATe](?)

[command]

Turns on the streaming Fast CW mode on the selected channel (ON or OFF).
Note that there is some delay between turning the mode ON and actual collection to allow for buffers to be set up. This can be as long as 100 ms.

[query]

Returns the state of streaming Fast CW mode on the selected channel.

:CALCulate:FCW:MODE(?)

[command]

Sets whether Fast CW will return the S-parameter data (SPAR), or the receiver data (RCVR). These correspond to type 1 and type 2 sub-modes discussed above, respectively.

[query]

Returns the current Fast CW mode

:CALCulate:FCW:DATA?

[command]

NONE

[query]

Outputs the data collected thus far in the form set by the current mode. Note that fscanf can be used for a no-command version of data retrieval in the streaming sub-mode.

:CALCulate:FCW:MARK

[command]

Inserts a marker in the collected data. This is discussed more in the next sub-section. The command includes a 32 bit number that will be placed in the real field of the data (and the imaginary field will be set to 0) as soon as latency allows.

[query]

NONE

:CALCulate:FCW:STReam:POINts(?)

[command]

Sets the selected no. of points for streaming mode of FastCW in the given channel.

[query]

Outputs the selected no. of points for streaming mode of FastCW in the given channel.

:CALCulate:FCW:DCOLlect(?)

[command]

Sets the type of collection. Choices are STREAM (for this section), HOLD, CONT, and STOP (HOLD and CONT are for the buffered sub-mode, STOP halts all collection and releases the buffers.

[query]

Outputs the current type of collection for Fast CW.

Buffered Sub-mode

The buffered sub-mode is a bit different in that all of the data will flow into a potentially sizable memory buffer and the user can request data from that buffer when convenient or when the data is ready. The user can specify the size of the buffer and, when the data reaches that point, collection will cease and the CFB bit of the extended status register will be set (that can be polled by a remote application and there is a wait-based command WFFB which works like the WFS (wait for sweep) command used for regular T/R measurements). The query command **:CALCulate:FCW:IBUF:POINts?** can also be used to determine how many measurements are currently in the buffer (and the two types discussed under 'streaming' apply here as well so a 'measurement' can be 8 bytes (one parameter) or 24 bytes (3 unratioed parameters)).

The maximum size of the buffer is set by available memory which is constrained by the VectorStar application, the operating system and any other applications that may be running. The maximum is ~40 million points for type 1 (8 bytes per point) or ~20 million points for type 2. (Actual size allowable may vary based on memory available on the system). If other applications are running on the instrument's PC or the PC memory is otherwise occupied, the effective maximum numbers may be reduced. The query command **:CALCulate:FCW:MPCount?** can be used to find the maximum allowable number of points at a given instant in time (but note the Fast CW must be ON for this command to produce an accurate value; a default maximum value will be provided otherwise). Note that the amount of time that this number of points represents is determined by the acquisition time per sample which is set by IFBW and point-by-point averaging. The minimum amount of time per sample (1 MHz IFBW and no averaging) is ~5 μs.

While the data format is fixed (see the streaming section) for transmission over the remote interface, the user does have a choice of where in the post-processing chain the data is retrieved from.

- Raw data (user RF calibrations applied but any receiver calibrations will be applied)
- Corrected data (after user RF calibration is applied…typically only reflection only, 1-path–2-port or frequency response calibrations are used in fast CW to avoid changes in the driving port; reference plane extensions are also included))
- Final data (corrected and after all post-processing steps such as DataMemMath, graph formatting, etc.; note that post-processing tasks that rely on groups of points for analysis, such as smoothing, group delay, time domain, etc. are not available in this mode).

The command is defined by

:CALCulate:FCW:DTYPe(?)

[command]

Sets the type of data to be collected in the internal buffer: Corrected, Final, or Raw.

[query]

Outputs the type of data that will be collected.

Some caution should be exercised with the last two choices in a very fast measurement scenario as the maximum speed can decrease and, if a 2-port calibration is involved, it can decrease dramatically as driving port switching is required).

One basic command controls the operation of the buffer.

:CALCulate:FCW:DCOLlect(?)

[command]

This command sets whether the system should start collecting Fast CW data, stop collection, or pause (hold) collecting data (CONT, HOLD or STOP...the value STREAM is used for the streaming sub-mode).

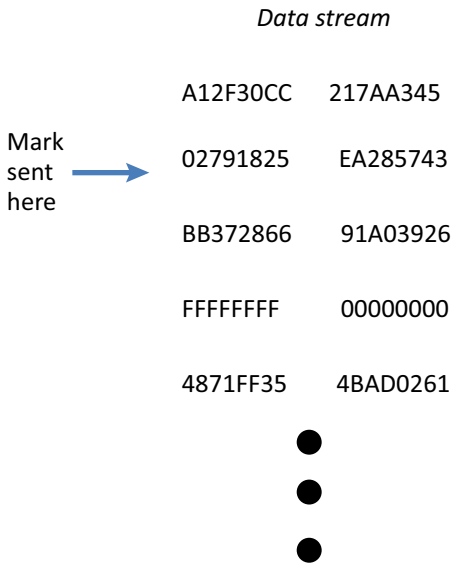
[query]

Outputs the current data collection setting.

Note	If it is attempted to retrieve data before the buffered selection is complete, the reading commands will produce an error.

The command **:CALCulate:FCW:MARK** (sent as MARK x where x is a float (32 bit)) is used to mark the data stream. This is intended as a synchronization tool to indicate when the DUT is in a particular state (particular rotating antenna angle, different bias state, different switch state, etc.) without slowing down the process like a conventional triggering mechanism might. When the command is sent, the mark is inserted in the data stream as soon as possible (but it may take 20 µs or more for the mark to be inserted due to bus latencies) so it can be identified later. Any float can be used as the mark and, since data is stored and sent as real-imaginary parts, the mark is put in the real position while 0 is put in the imaginary position. Using 0 for the mark is often convenient as then a 0,0 pair will appear in the data stream and this is usually very easy to identify since an exact 0 almost never will occur in a live data stream due to the residual noise floor. An example of this process is shown schematically in Figure 29-2 on page 29-8 when the mark value (in hex terms) was FFFFFFFF. Conceptually, the reader may notice similarities to External Sync Marking discussed in the Pulse measurements chapter of this guide (Chapter 25) and indeed they involve the same concept. The MARK process is not as direct since it must go through the remote interface layer so the latency is much higher than the ~5 ns for the sync marking process in Advanced Fast CW and in PulseView.

The MARK command also applies to the streaming sub-mode discussed earlier but, because of the more rapid transfers, the mark will be placed in the next available sub-chunk. The latencies are similar for the two sub-modes otherwise.



The MARK marking process in the data stream (in hex form) is shown here. Part way into the stream, a mark with hex value FFFFFFFF was sent and placed in the real part of the next available point entry (the imaginary part of the mark is always set to 0).

Figure 29-2. MARK Marking Process in Data Stream

As discussed earlier, there is some delay associated with turning on Fast CW (**`:CALC:FCW 1 or :CALC:FCW ON`**), particularly for the buffered sub-mode, as time is required to set up the buffers. This time can exceed 300 ms. To get a more accurate control of the timing (and allow for almost instantaneous MARKs at the beginning of the record), the following command sequence is suggested:

`:CALC:FCW:DCOL HOLD` (put the acquisition in hold while Fast CW is Off)

`:CALC:FCW1` (Now turn on Fast CW)

`…wait 300ms or more` (give time for the buffers to be set up)

`:CALC:FCW:DCOL CONT` (now start acquisition with minimal latency)

Much as in the PulseView measurements chapter (Chapter 25), we use the term 'triggering' here to refer to an event to start a macroscopic measurement cycle. In the context of Fast CW, this usually means the initiation of the collection or the stopping and re-initiation of the collection (to change a frequency or power level for example). Thus, the external, GPIB and manual per-sweep trigger events have no effect when collection is in process. When the Fast CW mode is stopped, any of the above triggering modes can be used to re-initiate collection once the Fast CW process is armed. Per-point triggering is permitted but since the speed of the triggering interface in CW mode is limited to about 10 kHz, this will significantly limit the maximum data rates. Again, the MARK command is used to indicate events on a finer time scale within the collection period.

The acquisition speed in this mode is set by the IF bandwidth selection modified by point-by-point averaging with the maximum being with a 1 MHz IFBW and no averaging. The actual ADC-level acquisition rate in this case is ~1.5 MHz. Due to overhead in moving data around, the maximum rate that data can be put into the buffer (or otherwise made available) is approximately 200 kHz. This rate scales with decreases in IFBW or increases in averaging (roughly).

Other commands of relevance to the buffered sub-mode are shown below. More information is available in the programming manual. OCS is a native command to transfer the entire buffer of data.

:CALCulate:FCW:CPCount?

[command]

NONE

[query]

Returns the count of the number of points that have been collected so far.

:CALCulate:FCW:MPCount?

[command]

NONE

[query]

Returns the maximum amount of points that can be collected in buffered Fast CW mode (when FCW is ON).

:CALCulate:FCW:IBUF:POINts(?)

[command]

Sets the selected no. of points for internal buffer mode of FastCW .

[query]

Outputs the selected no. of points for internal buffer mode of FastCW .

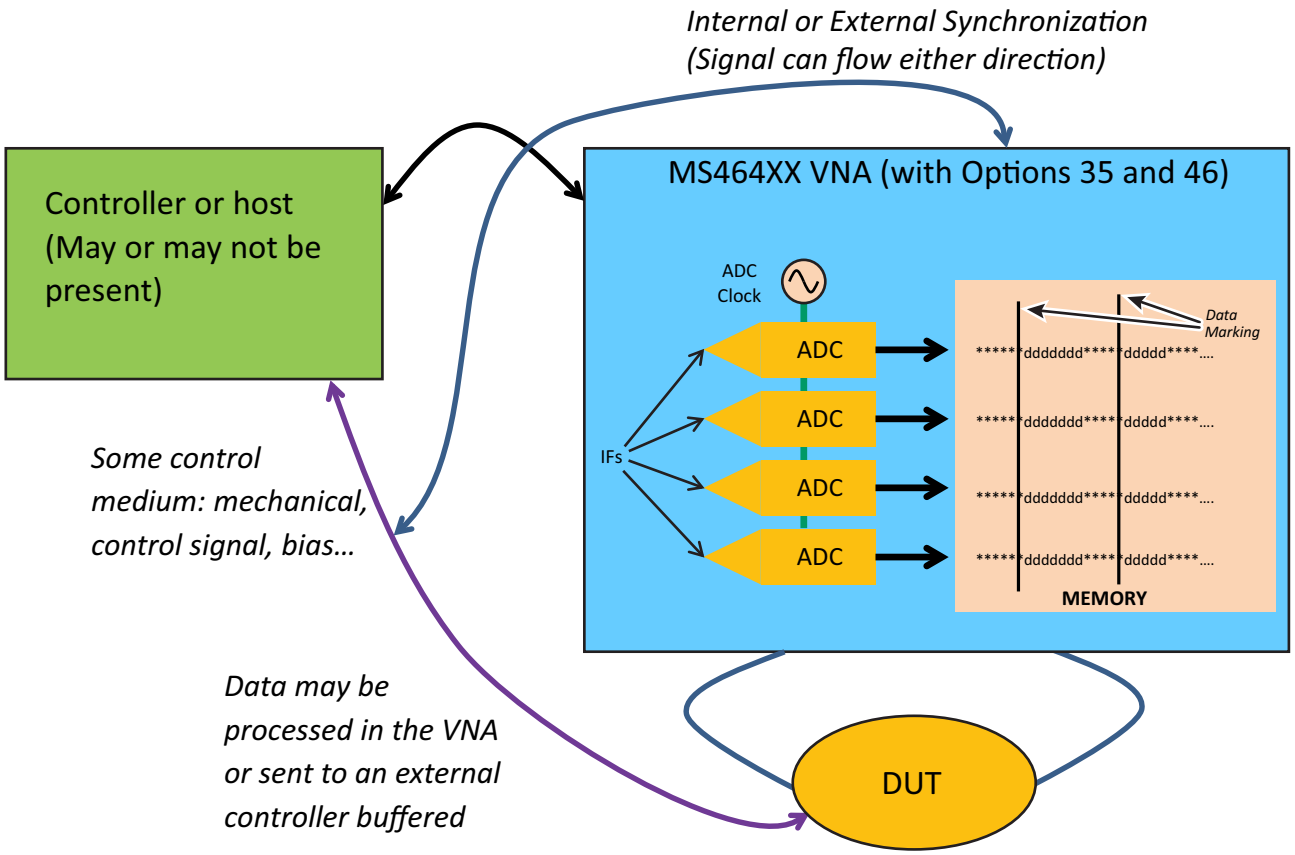
Note	Since there can be GPIB buffer issues with very large buffer sizes, the maximum data transfer size is 5 million points for type 1 and 2 million for type 2. If the buffer is set for a larger size, multiple data transfers must be requested.

29-3 Advanced Fast CW

Note	Advanced Fast CW *requires* Option 46 and Option 35.

The advanced CW measurement uses different IF and digital hardware internally that allows this mode to support much higher measurement rates and larger buffer sizes. The choice between these two modes may often be dictated by the rate-of-change of the system being tested. If changes of interest occur only on the scale of milliseconds, then the standard mode will probably be adequate. If changes occur on the scale a few microseconds or faster, the advanced fast CW method would be a likely choice. For intermediate cases, both modes can work and the choice may end of being one of synchronization flexibility, data handling details and convenience. The Advanced Fast CW mode does have a UI component so it does not need to be run remotely and there is a post-processing layer that can be used to isolate data of likely interest.

The measurement scheme is shown in Figure 29-3 and the process is very similar, on a conceptual level, to a pulse profiling measurement discussed in Chapter 25 (only without the pulses). When the measurement starts (which can be externally triggered), the data starts flowing into the digitizer memory. Periodic (or pseudo-periodic) marks can be placed in the data using an internal time reference or using external sync marking. Much as with the marking discussion in the previous section, this can be used to highlight a state-change on the DUT or the mechanical position of some part of the system under test (among other things). The latency of these marks is under 5 ns so the timing can be fairly precise. When internal sync is used, the marks are placed at a precise time interval specified by the user. When external sync is used, the marks are placed when a rising edge is sent to the Ext Synch In port on the instrument's rear panel. The estimated period of sync marks should be entered (in the UI or remotely) when using external sync marking in order to simplify the plotting of data versus time but this is not required.



The acquisition plan for Advanced Fast CW is shown here.

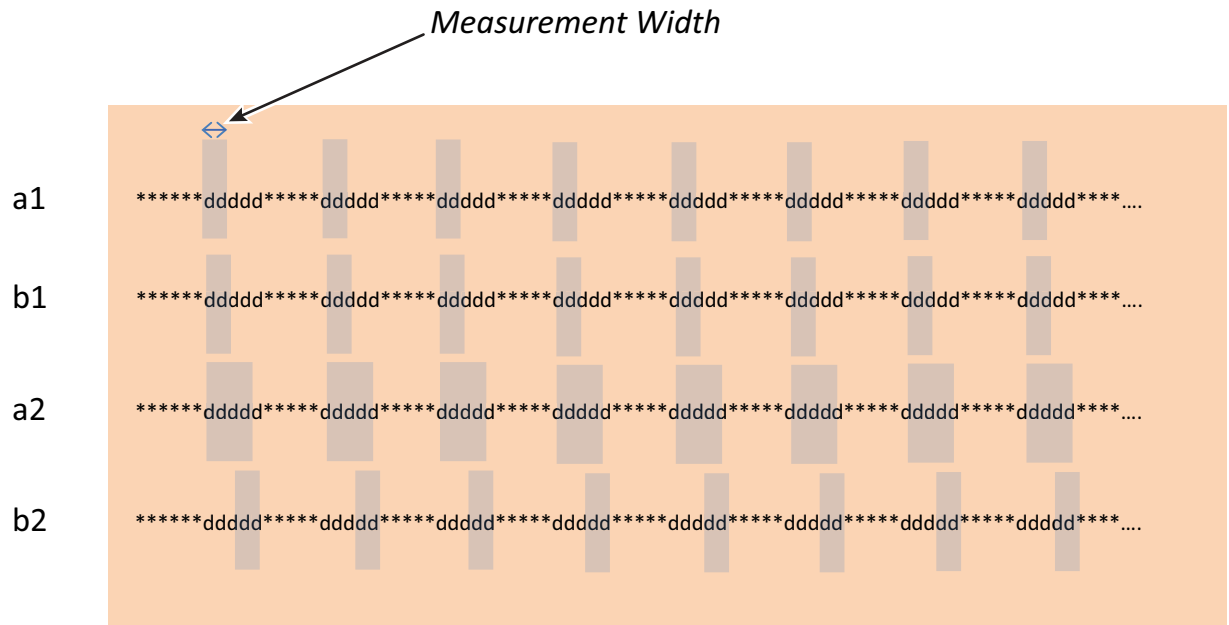
Figure 29-3. Advanced Fast CW Acquisition Plan

Some examples of different portions of the data that may be of interest:

- A sample near the synchronization events (e.g., each represents a discrete DUT state and it is only required to capture a value once that state has settled and the intervening time, when the DUT is settling for example, is not of interest.)
- Continuous time information
- Some sub-sampling of continuous time data that is not associated with a synchronization event

Many of these types of data reduction are available from the instrument and the results plotted (and, of course, are also available remotely). In all of these analysis types, the entered IF Bandwidth and Point-by-point averaging values are not used. 'Measurement width' (which will be defined) controls effective bandwidth and sweep-by-sweep averaging can be used for trace noise reduction on a longer time scale.

The simplest post processing to discuss is 'full time record' where samples can be analyzed with almost any density over the full set of data that was acquired. The concept of the 'measurement width', which may be familiar to readers of the pulse measurements chapter, is important. The user can specify how large a window should be analyzed to create a 'point' and this window width is expressed in terms of time. That amount of data in the time record will be blocked off and something similar to a discrete Fourier transform applied to create a 'point' with an effective bandwidth of 1/(measurement width). The available values depend on the acquisition rate (1/sampling interval) but for the default high speed rate, the minimum measurement width is 5 ns and the maximum is the depth of memory collected. The user also specifies a time range in total to be analyzed and how many points to be computed. The relationship of these variables is shown pictorially in Figure 29-4. Note that the widths and delays can be independently selected for each receiver. If a large number of points with a large measurement width are selected, the processing time will increase beyond seconds. In this method, the synchronization marks are not used except to denote the start of the measurement and, in this way, can function as a secondary triggering path that is much more closely tied to acquisition timing than is traditional external/GPIB/manual triggering so latencies are much lower and more predictable.



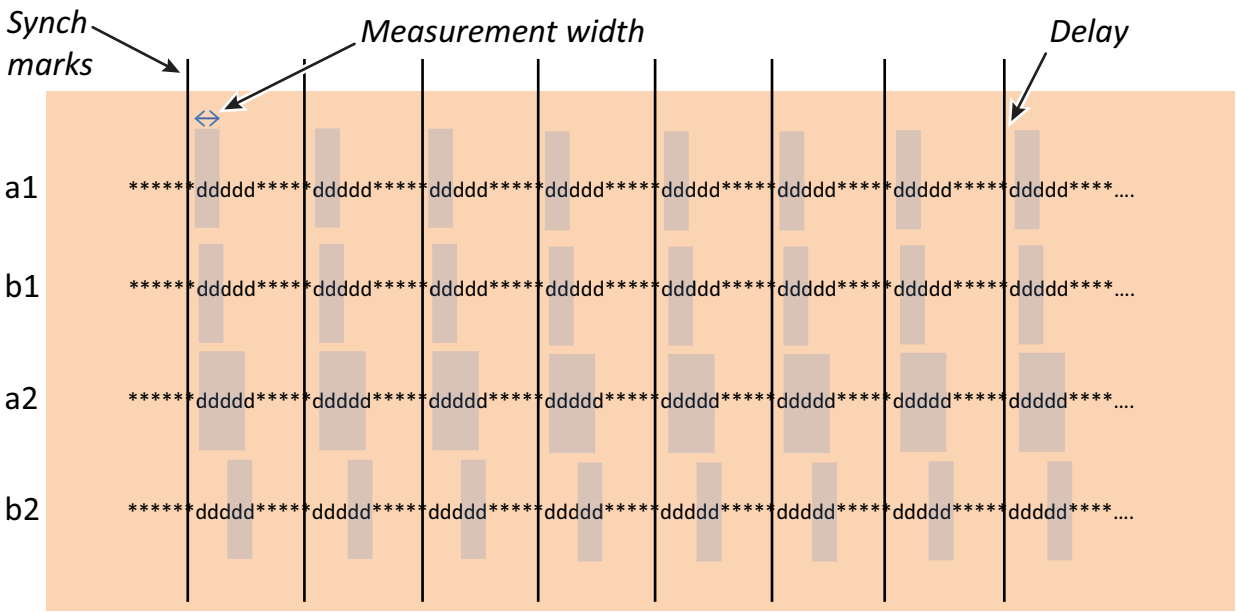
Data is analyzed in each measurement width shown. The position is specified by a start time and a stop time for each receiver. The number of widths analyzed must be the same for all receivers.

The processing scheme for the 'full time record' sub-mode is shown here. The analyzed measurement 'widths' are uniformly spaced in time for each receiver with the start and stop times independently selectable for each receiver.

Figure 29-4. Advanced Fast CW Processing Scheme—Full Time Record Sub-mode

The full time record approach can be useful when transient behavior of the DUT is of interest (as it moves from one state to another for example) or when an analog variable is being changed (such as bias or mechanical angle of a receiving antenna) and there is a less discrete sense to the measurement.

The other processing approach is point-by-point and is similar, in a calculation sense, to the pulse-to-pulse method in pulse measurements. The synchronization marks (whether they are internal or external) are central to this method since each analysis 'point' is associated with a mark. Much like the 'full time record' method, the measurement width defines the individual chunk of data to be analyzed but now the location of that chunk is defined by a time delay relative to the synchronization mark instead of in terms of absolute time. This processing approach is illustrated in Figure 29-5 and this method is useful when the various DUT states have a discrete nature to them and synchronization is possible (in either direction). It can also be used with no synchronization in which case the interval timing set on the instrument becomes the source of the timing marks.



Data is analyzed in each measurement width shown. The position of each 'width' is specified by a delay relative to the synch marks. The delay and width can be independent for all receivers.

The processing scheme for the 'point-by-point' sub-mode is shown here.

Figure 29-5. The processing scheme for the 'point-by-point' sub-mode is shown here.

Unlike in PulseView measurements, the concept of averaging between intervals does not apply so the IF bandwidth and point-by-point averaging values are not used. The measurement width controls the effective DFT bandwidth and the noise floor and trace noise. Sweep-by-sweep averaging is still available as this involves multiple acquisitions that can be combined and can, in particular, reduce trace noise. Note that there can be some triggering complications when using sweep-by-sweep averaging depending on the measurement since the DUT state will have to repeat on the multiple sweeps used.

Of course, the raw data (samples in time domain) can also be dumped for off-line processing. Because of the potential size of this data, it can only be saved in a binary format. To allow transportability, the format saves data into a sequence of 16 MB files. A MATLAB script for accessing the data is below and an executable version with some plotting and processing capabilities is available from Anritsu.

```
File = fullfile(pathname,filename);
FileLoc=pathname;
if isempty(FileLoc)
  FileLoc = 'current';
  error_flag=1;
  set(handles.status_field,'string','Directory field is empty.');
end

if exist(File,'file') == 2
            error_flag=0;
  try
       sA1 = 'Ch1_fastcw_'; %different preamble is used for different channels
       nSamplesPerFileMax = 2097152; %16MB and 8 bytes per sample
       TimeStep=2.5252e-9; %time step in seconds...NOTE: this is for the default
       resolution

       thisMax = 0;
       StartSample = 1;

       for ifile=1:nFiles    %nFiles= Number of files for the acquisition of
       interest
            fileID = (ifile-1)*nSamplesPerFileMax;
            str = sprintf('%d',fileID);

            % get file info
            fid = fopen([FileLoc, sA1, str, '.dat']);
            fileInfo = dir([FileLoc, sA1,str,'.dat']);
            fileSize = fileInfo.bytes;

            Samples_to_read = fileSize/8;

            basedata = fread(fid, [4,Samples_to_read], '*uint16');
            fclose(fid);
            basedata = basedata';

            iY1_a1 = basedata (:,1);   % There are four acquisition channels (a1...)
            iY2_b1 = basedata (:,2);
            iY3_a2 = basedata (:,3);
            iY4_b2 = basedata (:,4);
            iYSync1 = bitshift(bitand(iY1_a1 ,32768),-15);   % this is the synch bit
            for a1...it is the same for all channels
```

```
iYOR1 = bitshift(bitand(iY1_a1,16384),-14);    % this is the overrange
bit for a1
iY1_a1 = bitand(iY1_a1 ,16383);   % strip off the information bits


[NumPts,d] = size(iY1_a1);


ynormalized_1=(single(iY1_a1)-8192)/8192;
[chunk_size, k]=size(y_chunk1);


a1((ifile-1)*nSamplesPerFileMax+1:(ifile-1)*nSamplesPerFileMax+chunk_s
ize)=ynormalized_1;
%and absolute time
t((ifile-1)*nSamplesPerFileMax+1:(ifile-1)*nSamplesPerFileMax+chunk_si
ze) = (thisMax+(0:NumPts-1)) * TimeStep;



%absolute samples
thisMax = thisMax + NumPts;



y1(StartSample:thisMax) = y_chunk1(1:NumPts);
y_original1(StartSample:thisMax) = iY1_a1(1:NumPts);
ySync1(StartSample:thisMax) = iYSync1(1:NumPts);
yOR1(StartSample:thisMax) = iYOR1(1:NumPts);



% now do b1
iYSync2 = bitshift(bitand(iY2_b1,32768),-15);
iYOR2 = bitshift(bitand(iY2_b1,16384),-14);
iY2_b1 = bitand(iY2_b1 ,16383);
ynormalized_2=(single(iY2_b1)-8192)/8192;
b1((ifile-1)*nSamplesPerFileMax+1:(ifile-1)*nSamplesPerFileMax+chunk_s
ize)=ynormalized_2;


%now do a2
iYSync3 = bitshift(bitand(iY3_a2,32768),-15);
iYOR3 = bitshift(bitand(iY3_a2,16384),-14);
iY3_a2 = bitand(iY3_a2 ,16383);
ynormalized_3=(single(iY3_a2)-8192)/8192;
a2((ifile-1)*nSamplesPerFileMax+1:(ifile-1)*nSamplesPerFileMax+chunk_s
ize)=ynormalized_3;
```

```
            %now do b2
            iYSync4 = bitshift(bitand(iY4_b2 ,32768),-15);
            iYOR4 = bitshift(bitand(iY4_b2,16384),-14);
            iY4_b2 = bitand(iY4_b2 ,16383);
            ynormalized_4=(single(iY4_b2)-8192)/8192;
            b2((ifile-1)*nSamplesPerFileMax+1:(ifile-1)*nSamplesPerFileMax+chunk_s
            ize)=ynormalized_4;
            isync((ifile-1)*nSamplesPerFileMax+1:(ifile-1)*nSamplesPerFileMax+chun
            k_size)=iYSync4; %Using the b2 sync as the sync output (arbitrary)

            %next loop value
            StartSample = StartSample + NumPts;

        end
      catch Main
        error_flag=1;
      end
    end
```

The format of the file names is:

```
Ch1_fastcw_nnn.dat
```

Where nnn is an index number (first file is always 0, second file is generally 2097152, etc.). The file name preamble changes for different channels. For the code above and for the executables provided by Anritsu, it is important to not modify the file names or, if needed, to only add characters at the beginning of the file name (even then, the variable sA1 in the above code will need to be generalized).

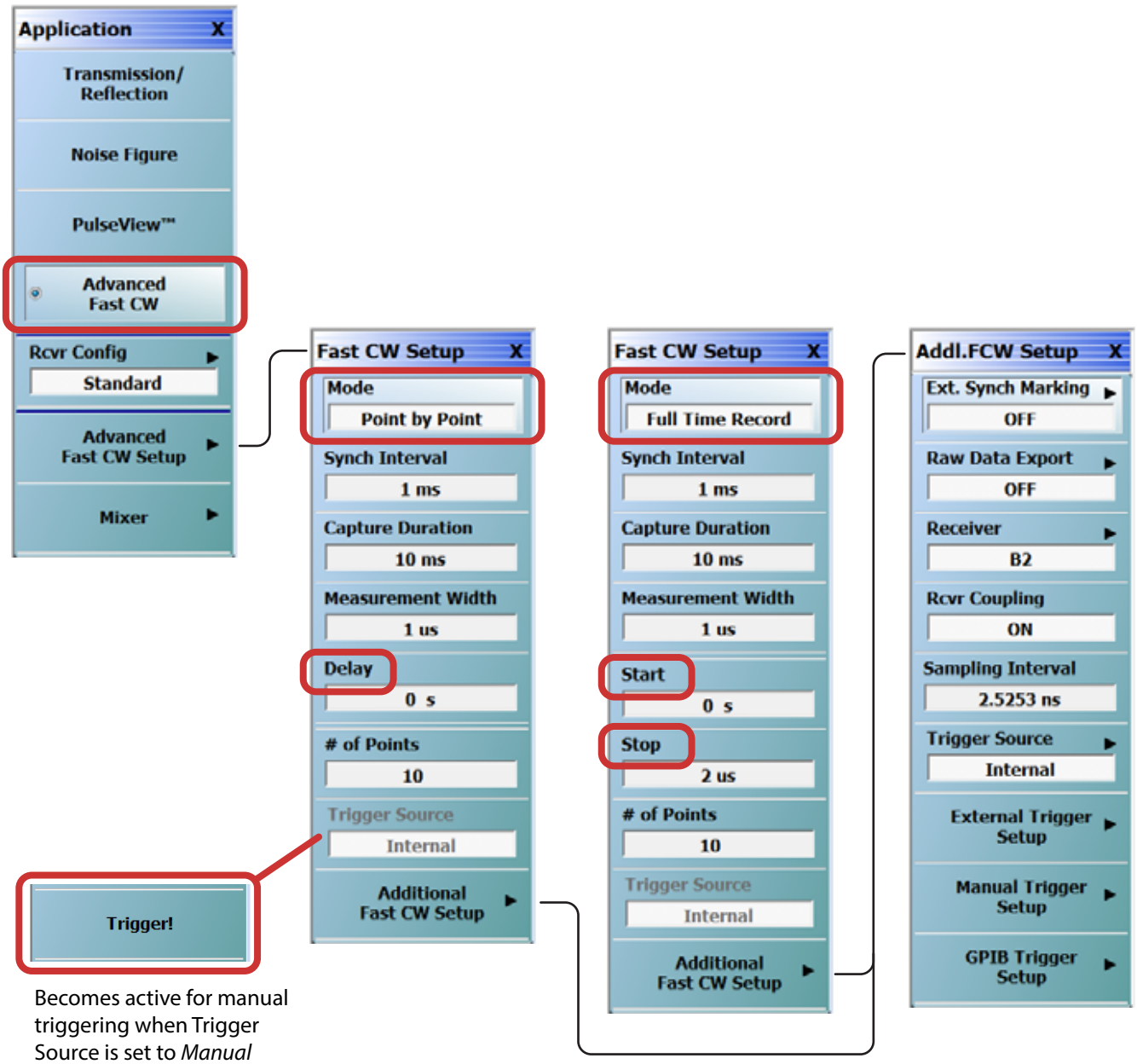
An important point about the data is that each receiver has a field consisting of 14 data bits, one overrange bit and one synch bit forming 2 bytes per receiver. Data from all four receivers are always saved. The synch bit is common to all receivers as those paths are locked but the bit is repeated in the 2-byte-field for each receiver for convenience.

Advanced Fast CW Setup Menus

The parameters are all configured on the Advanced Fast CW setup menus shown in Figure 29-6 on page 29-16 and Figure 29-7 on page 29-17. The Enable Raw Data Save button brings up a file save dialog for the formats just described above. The Sampling Interval (equivalent to 'Resolution' in PulseView) and Capture Duration entries are analogous to concepts in PulseView and are described in more detail shortly.

Most of the remaining parameters have already been discussed as they configure the data analysis for the full time record and point-by-point subtypes. The Synch Interval relates to the synchronization marking interval (which could be internal or external as discussed) and is akin to Pulse Repetition Interval in PulseView. In the internal synch mode, this Synch Interval value is explicitly used for the time between synch marks. In external synch mode, it is the external signal that establishes the mark timing but the entry in the menu is used to help with labeling the x-axes on plots and for creating the independent variable vector for text-based data saving. The External Synch selection has three choices: Off (which implies internal synch), Ext. rising (external synch based on the rising edge) and Ext. falling (external synch based on the falling edge).

Note that the Trigger menu items have been replicated under the Fast CW menu for convenience. Changes made here will appear under the Trigger menu and vice-versa (they are the same control variables).



The APPLICATION menu with Advanced Fast CW selected is shown here. The Fast CW Setup menus provide parameter entries for this application. Note the Fast CW Setup menu items change slightly depending on the selected mode: When Point by Point mode is changed to Full Time Record, the Delay time field is replaced by Start Time and Stop time fields. The Additional Fast CW Setup sub-menus are shown in Figure 29-7 on page 29-17.

When *Manual* is selected as the Trigger Source (selected from the Additional Fast CW Setup menu), the Trigger Source status on the Fast CW Setup menu becomes active for manual triggering.

Figure 29-6. ADVANCED FAST CW APPLICATION Menus

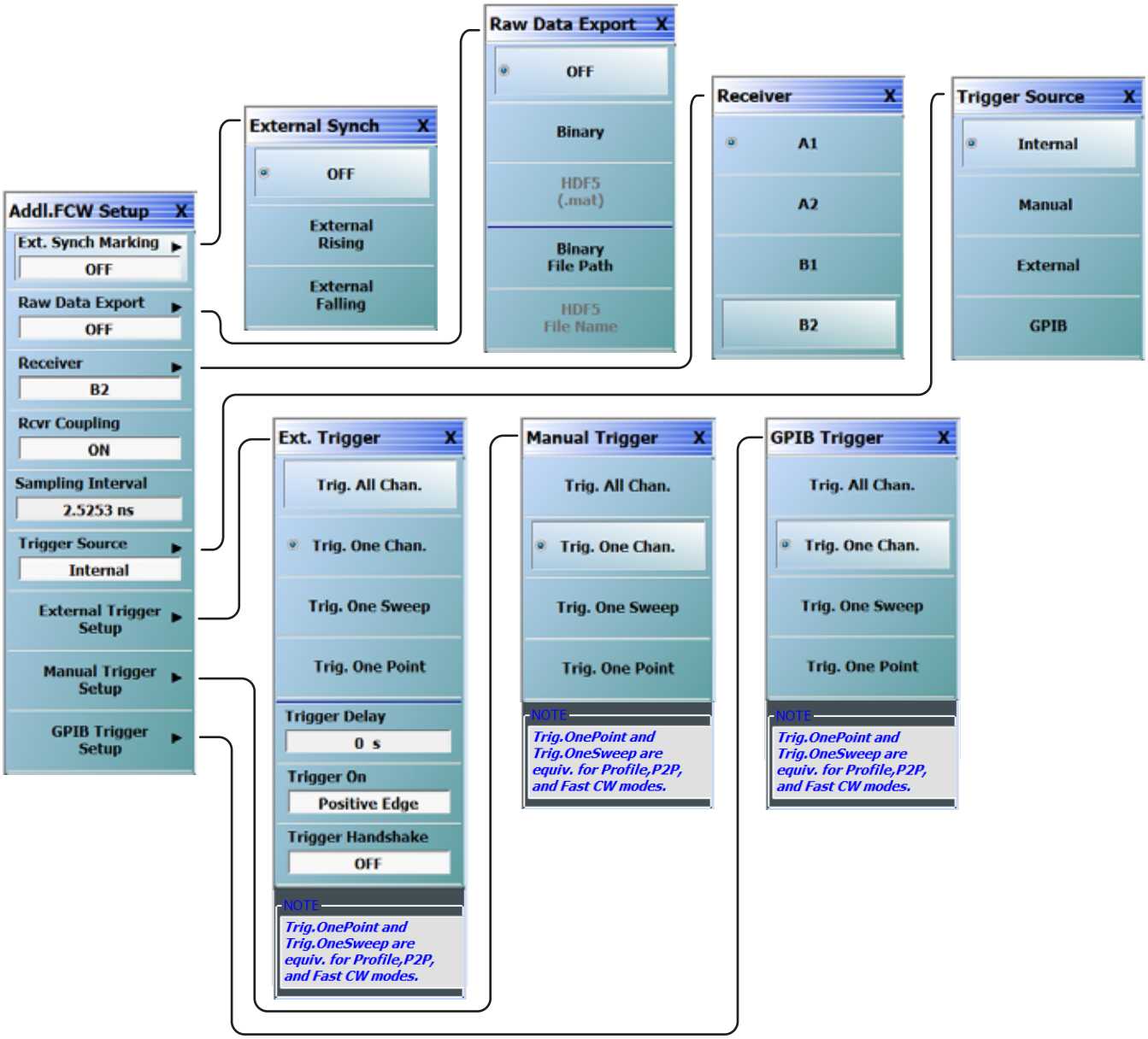


Figure 29-7. ADDITIONAL FAST CW SETUP Sub-menus

The Sampling Interval (1/ (acquisition rate)) is variable and this affects how much data in terms of time can be stored. At the default Sampling Interval near 2.5ns, the maximum depth is 0.5 seconds without the expanded memory option (Option 36) and 2.5 seconds with Option 36. The Sampling Interval can be increased to a maximum of 70 ns which leads to a maximum depth of 14 seconds or 70 seconds (without or with Option 36, respectively). In terms of useful data points that can be stored, the Sampling Interval does not matter but how the data is analyzed will have an impact on the maximum point count (here 'point' refers to a realized frequency domain measurement value). Since the raw time domain waveform is at the system IF, normally data is analyzed in terms of a short-time DFT as discussed earlier in this chapter and the classical minimum would be four time samples to form a data point. Under that rule, the maximum point count is 50 million or

200 million (without or with Option 36) for EACH of the four receiver channels (a1, a2, b1 and b2). Internally, fitting options are available that allow analysis on 2 or 3 time samples which allows the point count to increase to 100 million/400 million with a corresponding increase in trace/data noise. The time-shifting single sample approach that allows 2.5 ns resolution/Sampling Interval in the PulseView mode is not available in Fast CW due to slightly different data handling procedures. Of course, if larger amounts of effective averaging (more samples in the DFT), then the number of available 'points' will correspondingly decrease unless one starts to overlap time intervals. The latter is allowed (and can always be done offline) but introduces an effective smoothing function so that the time-resolution-of-data is not increasing as much as the point count would indicate.

Capture Duration (also termed Acquisition Length) has also been discussed in Chapter 25 and simply informs the system how much data to collect in the buffer in terms of time. As listed in the previous paragraph, the maximum allowed value in this field is a function of the Sampling Interval. Note that the number of plotted frequency domain points in 'point-by-point' mode is capped at ~Capture Duration/Synch Interval since that ratio determines how many sync events can be stored (it may be one less depending on the delay value). In Full Time Record more, the Stop Time must be less than the Capture Duration.

Much as in the PulseView measurements chapter (Chapter 25), we use the term 'triggering' here to refer to an event to start a macroscopic measurement cycle. In the context of Fast CW, this means the initiation of the collection or the stopping and re-initiation of the collection (to change a frequency or power level for example). Thus, the external, GPIB and certainly manual trigger events have no effect when collection is in process. Once an acquisition is complete, the above triggering modes can be used to start another one. The per-point and per-sweep variants have the same effect since it is a CW mode. Again, the external sync marking concept is used to indicate timing events on the finer time scale within the acquisition.

29-4 Application Examples

A first example shown below uses the Standard Fast CW mode and the internal buffering sub-mode. This particular measurement is of an antenna pattern where signals are available in the controller when the rotating antenna passes certain preset angles. These will be treated as sequential polling events in the below code for configuration-specific reasons but there are many other ways of performing this task. It is desired to mark the data at those points for later correlation analysis. As standard Fast CW is remote-only, this example will be presented in terms of control code. This code here is from MATLAB and other environments will do something similar although the command formatting will be different. It is assumed that the connection to the remote VNA has already been established (as object 'obj1') when the below commands are executed. In particular, it is important to allocate the Input and Output Buffer sizes appropriately. If transferring a large amount of data from the VNA (which we will be in this example), the Input Buffer Size (expressed in Bytes) must be able to accommodate it.

Standard Fast CW Example

```
% Set size of the buffer
fprintf(obj1, ':SENS:FCW:IBUF:POIN 100000’);

% Start the collection and start the rotation hardware (a user
% function call does this.
fprintf(obj1, ':SENS:FCW:DCOL CONT');
StartAntennaRotation(one_cycle);

% Mark the data during acquisition.
% The rotation is known to be slow relative to acquisition
% so a ‘while’ loop can work. When a given angle is passed in this
% setup, the bit goes high and stays high.
while Angle_90 == 0
end
% Mark the data with a 0 real (and 0 imaginary)
fprintf(obj1,':SENS:FCW:MARK 0');
while Angle_180 == 0
end
fprintf(obj1,':SENS:FCW:MARK 0');
while Angle_270 == 0
end
fprintf(obj1,':SENS:FCW:MARK 0');

% Now just wait for the internal buffer to fill up
fprintf(obj1,'WFFB');

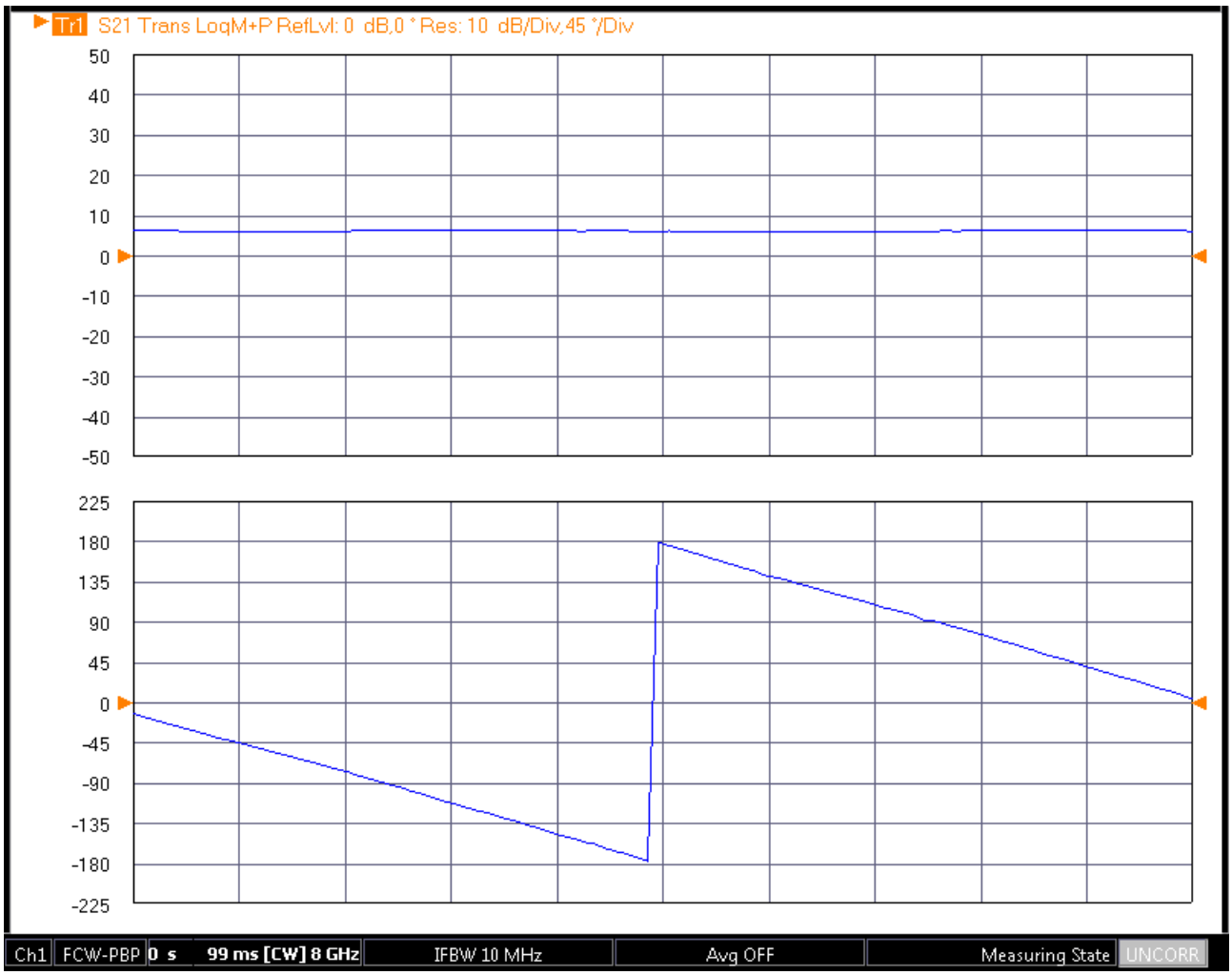
% Get the number of points and then transfer the data from the
% VNA internal buffer
data7 = query(obj1, ':SENS:FCW:CPC?', '%s\n' ,'%s');
points=str2num(data7);
fprintf(obj1,'OCS');

% The binblockread command automatically strips the header and reads
% the size. Fread can also be used but the header must be stripped
% manually. The data is read in by bytes instead of 32 bit float
% size chunks to make byte swapping simpler (LSB is first by default)
[Data, count]=binblockread(obj1,'uint8');
```

The variable Data would then next be processed into complex floats typically while looking for the mark variables 0+j0 that would indicate the cardinal angle positions.

Advanced Fast CW Example

For a first Advanced Fast CW example, consider an externally synchronized measurement where a DUT has a large number of variable phase states and it is desired to step through them relatively quickly and acquire a complex S21 measurement at each step. An external control module will step the DUT state every millisecond and also send a sync pulse to the VNA (slightly delayed so it is guaranteed that the DUT is settled in time; if this was not the case, the Delay variable could be used to compensate on the VNA). Point-by-point measurements will be used since we wish to associate a measurement with each DUT state. On the Advanced Fast CW menu, the sync interval would be set for 1 ms and, since we happen to need to study 100 DUT states, the capture duration will be set to 100 ms and the number of points to 100. The measurement width was selected to be 500 µs to bring trace noise down to an improved level (the practical maximum may be on the order of 800–900 µs based on the Sync Interval). The result is shown in Figure 29-8 along with the DUT gain although this was not as much of a focus as was the phase.

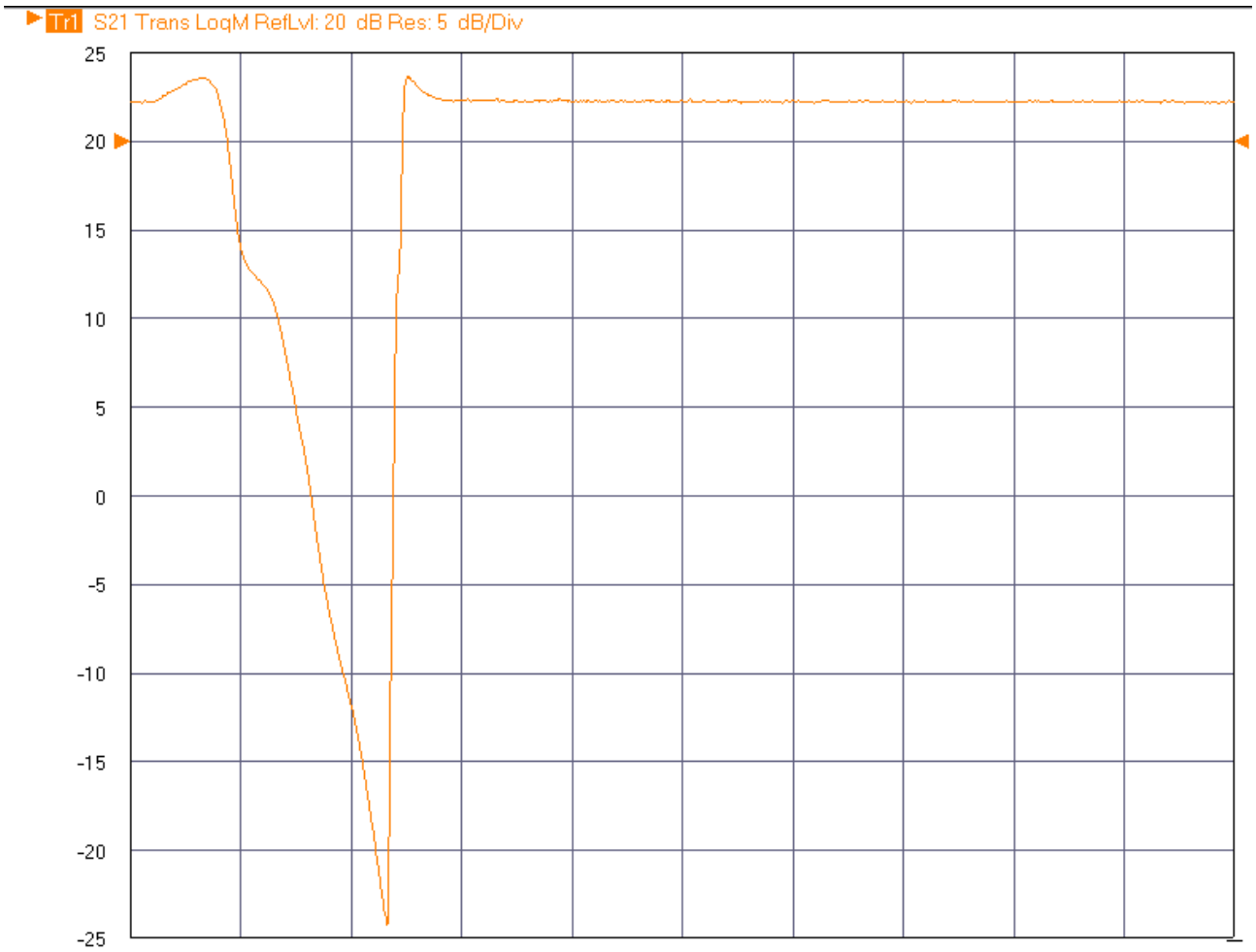


An example point-by-point Advanced Fast CW measurement result is shown here. A DUT was stepped through 100 phase states while the instrument received an External Synch signal to note each step.

Figure 29-8. Advanced Fast CW Example—Point by Point

Because the speed was not extremely high in this measurement, Standard Fast CW could have been used but the timing synchronization was simpler using Advanced Fast CW and the data processing of each measurement window could be performed by the instrument. Note that this particular measurement was performed uncalibrated and un-normalized. The user, in this case, was performing calibration off-line so the instrument was being used to collect synchronized, pre-processed raw data.

As a ‘full time record’ example, consider a module that is subject to occasional bias system transients (due to an internal overload condition). It is desired to understand how the DUT gain (at 18 GHz in this example) responds to that transient…both in terms of the duration of gain suppression and the extent of any gain overshoot on recovery or pre-shoot. The capture duration was set to 1 ms and 1000 points covering that duration were requested. The measurement width was set to 100 ns. The resulting plot is shown in Figure 29-8. One can see a ~ 1 dB pre-shoot and a ~1 dB overshoot in gain around the transient. The user in this case would then look to see if that transient gain change would cause issues further downstream in their system implementation. Also of interest is the very long duration of the transient (nearly 150 µs) which was unexpected since the actual bias transient at the DC source was less than half that time. The user was then investigating time constants within the bias system.



A full-time record application example attempting to capture the shape of the gain effect of a bias transient on a DUT is shown here. The x-axis ranges from 0 to 1 ms (the capture duration and the stop time).

Figure 29-9. Advanced Fast CW Example—Full Time Record

In this example, the timing resolution desired was faster than could be supported by standard Fast CW so the use of Advanced Fast CW made more sense. The UI level processing in the full time record mode also helped in speeding the analysis of the data.

29-5 Summary

The *Fast CW* option provides a number of different modes for collecting data at rates ranging from slow values to 100 million measurements per second in everything from a streaming fashion (for unlimited data flow) to a fully buffered approach for collecting up to 8 GB of data. In choosing the correct mode, there are several factors to consider

- What timing resolution and event-marking time accuracy is needed? If the resolution need gets below the ~10 microsecond range and/or if marking accuracy of better than ~10 microseconds is needed, the Advanced Fast CW method is probably preferred.
- If buffering, what is the maximum length needed? If greater than ~20–40 million points, the Advanced Fast CW method is preferred. (Actual size allowable may vary based on memory available on the system)
- Is real-time (or pseudo-real time with small chunks of data) streaming needed? If so, the remote-only standard Fast CW is preferred.
- Is interval processing of the data or UI control desired? If so, Advanced Fast CW is preferred.

Chapter 30 — RF Options

30-1 Introduction

This chapter covers several RF-deck-specific options and related features that can be used in a number of different applications. Several of these options are touched on in other chapters of this guide but this chapter is to serve as a central information point on items directly affecting signal levels into and out of the instrument. The relevant options are

- **Option 32:** Source 2 re-direct to Port 1 (requires Option 31—Dual Sources). This is used for signal combining typically and can be particularly useful for IMD measurements.
- **Option 51:** Access loops to source paths and coupled test and reference paths. Options 51, 61, and 62 are mutually exclusive.
- **Option 61/62:** Step attenuators added to the source and coupled test paths as well as access loops.

Note	Options 51, 61, and 62 are mutually exclusive.

- **Option 53:** External ALC access to allow for leveling in different places in the setup (e.g., after a pre-amplifier added into a source loop path).

As one can gather from the above descriptions, the source loops, the source-side attenuators and the amplifier access can all be geared toward increasing or decreasing the drive power of the VNA's internal source. This may be useful for:

- Inserting a driver amplifier for power amplifier measurements (where the DUT test conditions require a higher drive level than the VNA normal port power specifications would support).
- Increasing power for very high loss measurements (in an antenna setup, for example) where signal-to-noise ratio may be a challenge.
- Decreasing the power below normal ALC range for extremely sensitive devices (e.g., very high gain amplifiers or devices with very low compression points).

The reference and test loops and attenuators (and the amplifier) correspondingly alter the levels going into the instrument's downconverters. Commonly, the level must be reduced to avoid instrument compression when measuring high power DUTs. The level may also need to be increased in very high loss situations (the antenna example again) which could be done with the amplifier in the loop or by directly injecting the signal into the downconverter (when reflectometry on that port is not needed).

30-2 Option 32 and the Re-direct Switch/Combiner

Option 32 was covered somewhat in Chapter 28 of this guide on IMD and elsewhere but it has uses for other measurements and definitely affects RF power distribution so the salient points will be revisited here.

As discussed previously, Option 32 is only available with dual source systems (Option 31) and provides a means for coupling second source power onto the primary source signal coming out Port 1. This is done with a switch and coupler arrangement shown schematically in Figure 30-1 (in reality, there are separate switches and couplers for above and below 2.5 GHz which is not shown here).

The switch/combiner are placed after their respective source loops when those options are installed (Option 51/61/62) so per-tone amplifiers/pads can be inserted in the source loops and they will not be exposed to the combined signal in IMD applications. For the multiple source case, this switch activation is done from the Other Setup | Power Setup submenu under the Power menu as shown in Figure 30-2 on page 30-3

Note	The **Source Out Redirection** button will not be available if in IMD mode. The source selection in that case must be made from the IMD configuration dialog.

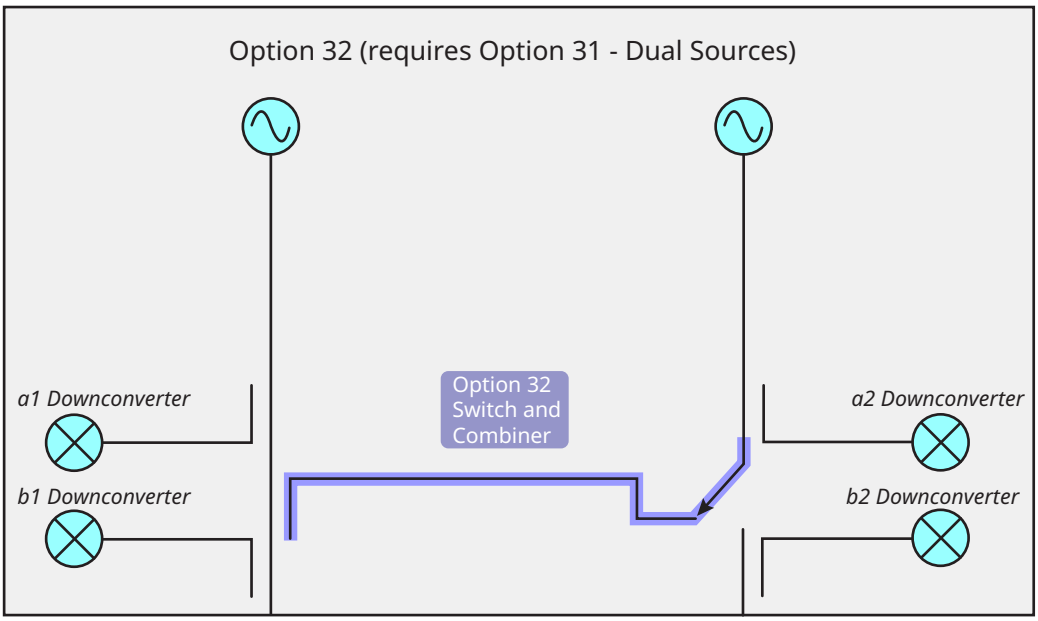


Figure 30-1. Schematic of Option 32 Switch and Coupling Components

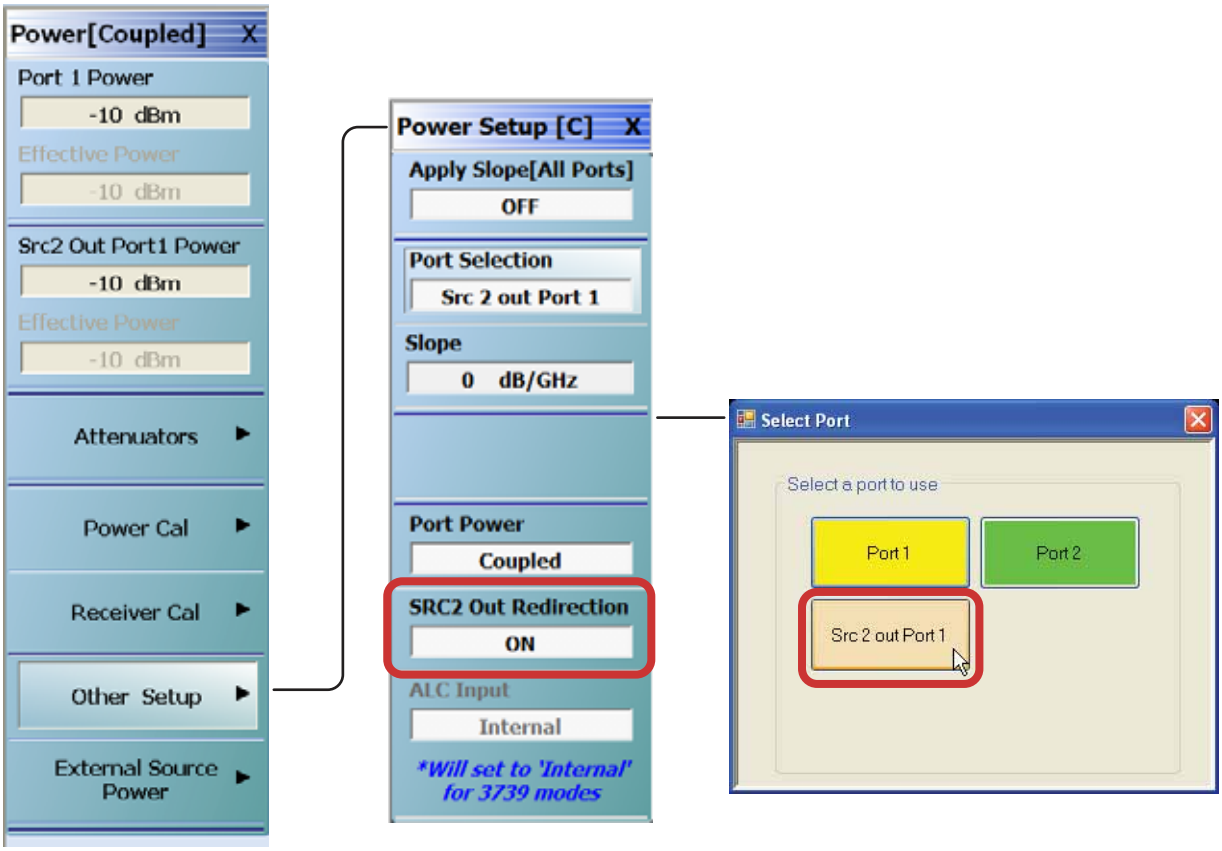
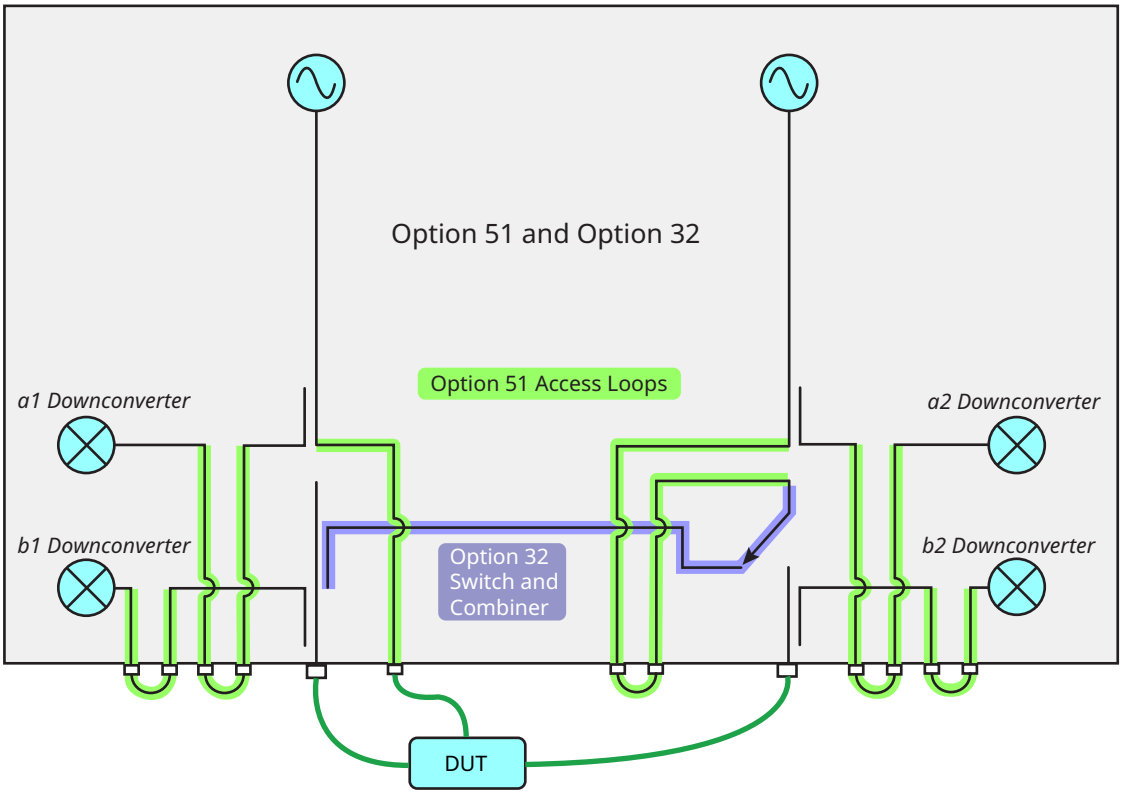


Figure 30-2. Power and Power Setup Menus: Option 32 Source 2 Out Redirection Control

The obvious signal combining application is two-tone Intermodulation Distortion which was covered in Chapter 28, "IMDView™ (Option 44)". There are times when a manual IMD measurement must be orchestrated using multiple source control (e.g., a complex frequency conversion path with multiple multiplications/divisions and multiple frequency offsets) but the approach would be the same. There may be other complex test setups where Source 1 is used through its access loop and it is desired to drive Port 1 with Source 2 as suggested by Figure 30-3 on page 30-4 and the Option 32 switching can be used to accommodate this configuration.

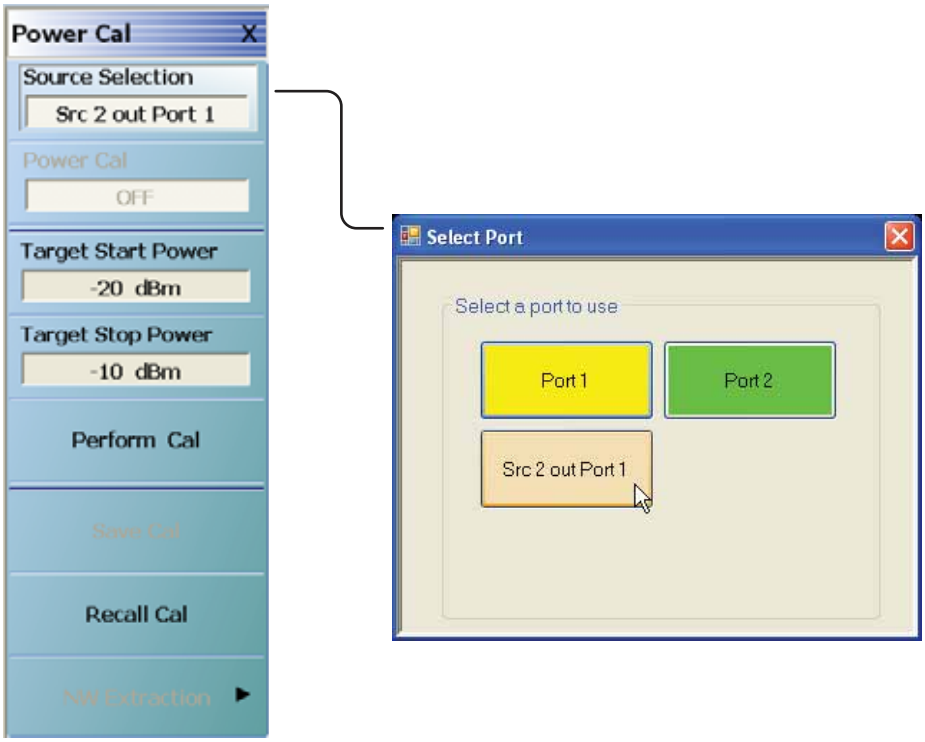
Because of the coupling and switch loss, the available power from Source 2 out Port 1 is reduced relative to that out Port 2 (by ~15 dB above 2.5 GHz and ~20–23 dB below 2.5 GHz with the difference being due to different combining structures). The isolation of the combining structures exceeds 20 dB which helps reduce source-to-source intermodulation residuals.



An application example for Option 32 is shown here where src 2 out Port 1 is used more for redirection than for signal combining and may be suitable for certain complex switched test setups.

Figure 30-3. Application Example for Option 32

Because of the power difference in Src2->P1 (as it is labeled in parts of the user interface), user power calibrations are commonly desired although there is a factory ALC calibration to account for those path differences in this mode to first order. When the redirection switch is activated, whether within IMD or not, a third choice appears on the power calibration selection as shown in Figure 30-4. The power calibrations work the same as with the regular configurations but keep in mind the loss levels when selecting the Target Power. If too high a target is set, the power calibration will likely fail or will pass with a rather high error (should normally be << 1 dB).



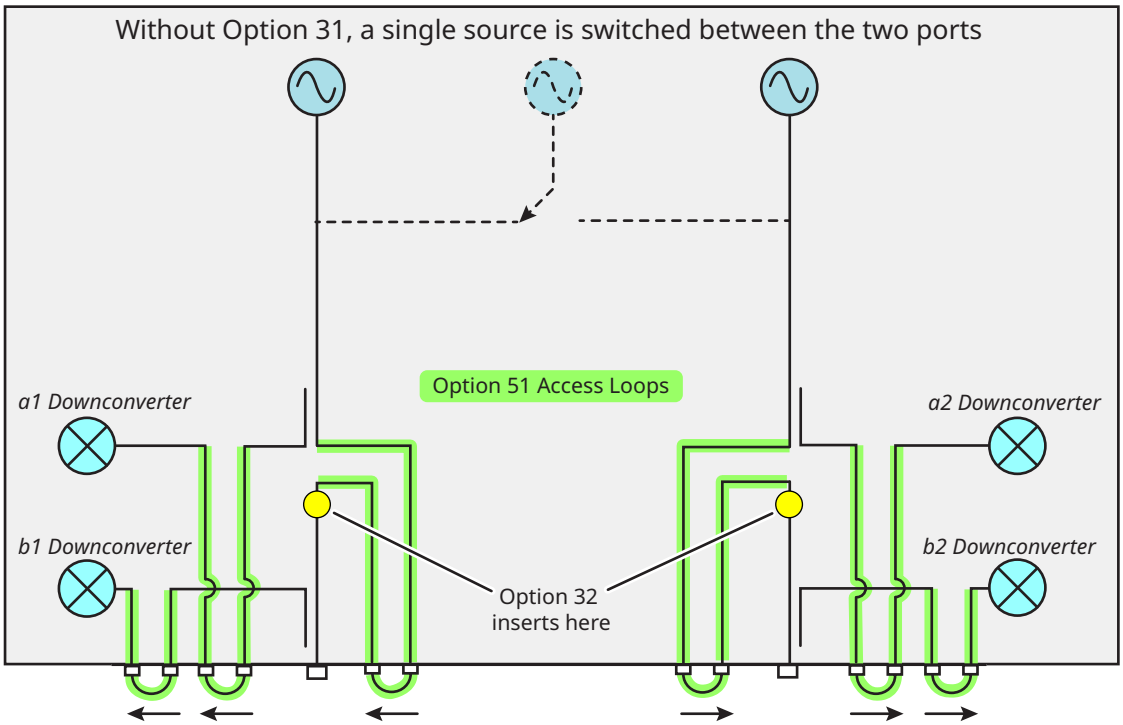
Access to the user power calibrations for the Src2->P1 path is provided by the power calibration selection button which opens the Select Port dialog. The redirection switch must be 'ON' for this extra choice to be available.

Figure 30-4. Source 2 out Port 1 Selection Dialog

The Source 2–>P1 path is not useful in broadband mmWave modes as the routing to the multipliers is different and one would not normally want to combine signals prior to multiplication anyway (as the residual intermodulation products will be, by definition, extremely high). For mmWave applications, an external combiner instead of Option 32 is recommended (contact the factory for details).

30-3 Option 51 and Access Loops

In some sense, Option 51 is the simplest of the RF options: it is to provide access to source and receiver paths. The isolated block diagram is shown in Figure 30-5. One important point is that the source loops are between reference and test couplers. For the labeling of the source loops, the base of the arrow corresponds to the source side and the pointed end corresponds to the side closer to the test port.



The block diagram showing the loops of Option 51 is presented here. Only front panel loops are shown in this figure which are for frequencies >=2.5 GHz. The loops for <2.5 GHz frequency access are on the rear panel. The arrows shown match the markings on the instrument.

Figure 30-5. Option 51 and Access Loops

Power leveling is done prior to the source loops so the ALC dynamics will not be affected by device added into the loop path but, of course, the power delivered to the port will be affected. User power calibrations are used to improve the accuracy of the power at the user reference plane and are particularly helpful when these source loop access points are used. An important point about the power calibrations when using a driver amplifier or other network in the source loops is the relationship between Entered power, Effective Power and Target Power (the former being on the top level of the Power menu and the last being entered as part of a user power calibration). Note that the leveling detection point is within the a1 and a2 coupling systems and hence precedes the source loop. For a leveling point that can include, for example, a pre-amplifier in the source loop, see Option 53 (external ALC) discussed later in Section 30-5.

Entered Power: This is the ALC system setting. When no user power calibration is applied, the system uses the factory ALC calibration to setup the hardware to provide close to the indicated value to the test port. When a user power calibration is applied, the power delivered may now be only loosely related to the ALC entry value.

Target Power: This is the value that the user power calibration will attempt to achieve. If a value here is unrealistic based on the instrument power range, devices added to the source loop, and networks between the instrument port and the reference plane, the calibration will fail or produce a large error.

Effective Power: This is a calculated value of the power available at the reference plane based on knowledge of the user power calibration and step attenuator settings (see Option 61 or 62). Immediately after a user power calibration, assuming it was successful, the Target Power will be available at the reference plane and the Effective Power will match the target power. If the Entered Power is changed after the user power calibration, the Effective Power will change by an equivalent amount.

Example:

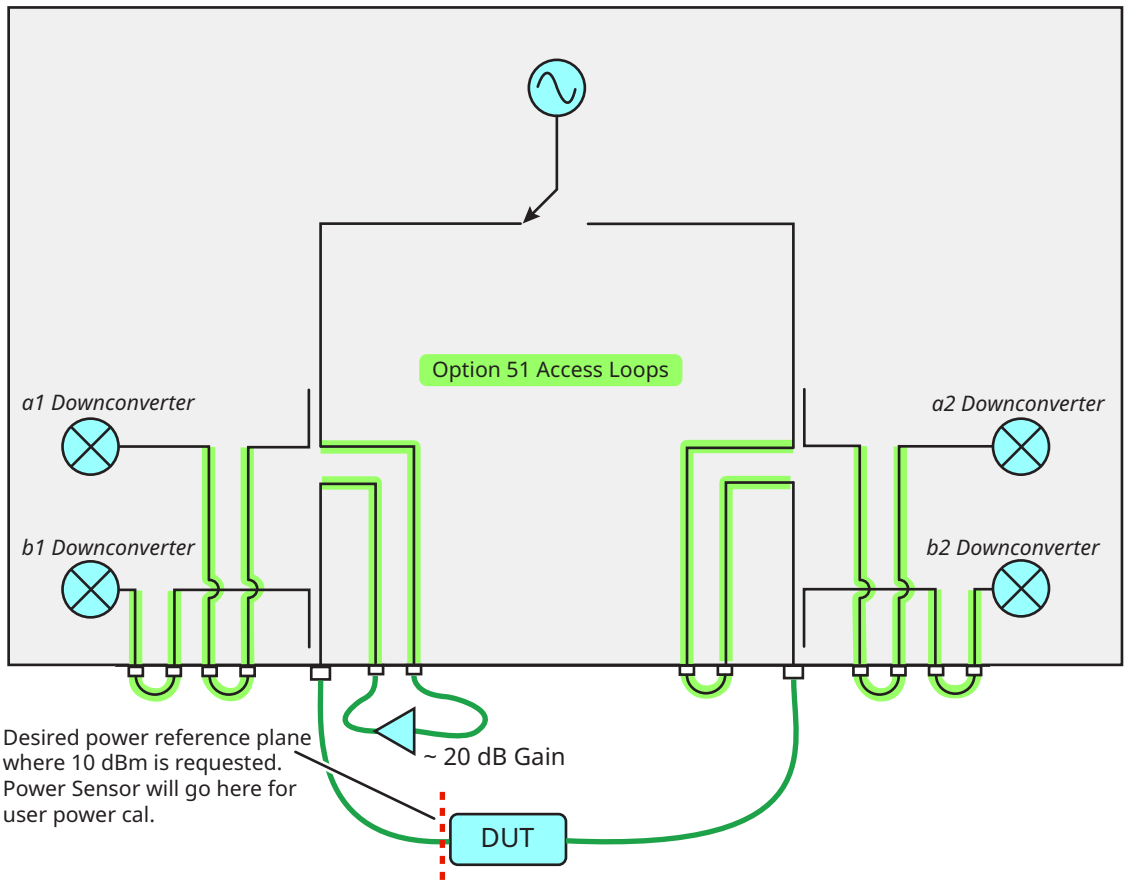
The power menu Entered Power (ALC setting) is –10 dBm.

A user power calibration is to be performed with a Target Power of –15 dBm (which will be achievable unless ~>15 dB of gain or > 15 dB of loss (depending on frequency and instrument model) has been added to the source loop path or between the port and the reference plane).

After the calibration is complete, the Effective Power will read –15 dBm.

If the Entered Power on the power menu is now changed from –10 to –7 dBm, the Effective Power will change to –12 dBm. At the reference plane, approximately –12 dBm of power will be available (unless considerable loss has been added either to the source loop or between the port and the reference plane).

When a large gain or loss is added, more careful consideration of the Entered and Target power may be warranted. Consider the example of Figure 30-6 where a relatively high gain amplifier has been added to the source loop (~20 dB) and a target power of 10 dBm is desired and the frequency range is 20 GHz to 30 GHz.



An example is illustrated here of performing a user power calibration with a relatively high gain amplifier added to a source loop.

Figure 30-6. User Power Calibration Example

If one used 10 dBm for the Entered Power and the Target Power, there may be unleveled messages displayed prior to the calibration (depending on the instrument model) and the driver amplifier will be initially exposed to power levels in that +10 dBm range. Not only may this be an issue for the amplifier, but it may also damage the power sensor that will be connected at the reference plane and it may hamper the ability of the calibration to complete (even if the sensor was not damaged) due to convergence issues. A reasonable practice is to deduct the gain from the Entered Power first (so –10 dBm would be entered). This need not be precise. It is only telling the system where it should approximately start the power search.

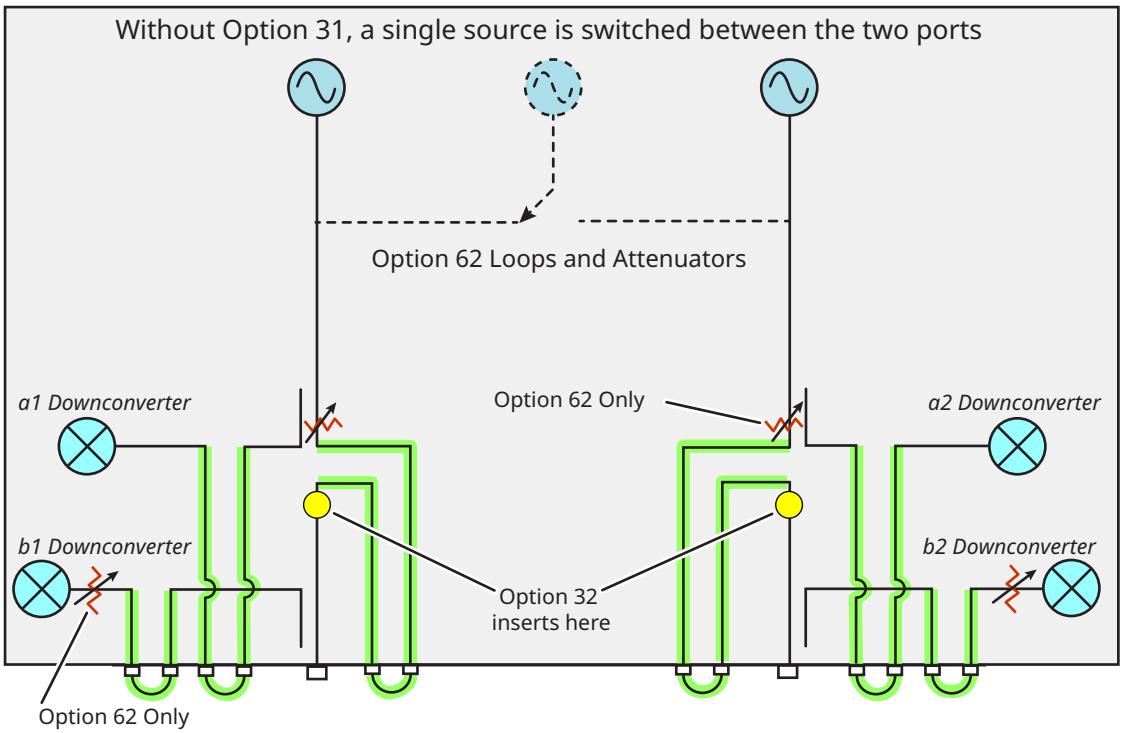
The source loops are labeled as having a damage limit of +20 dBm and this is somewhat conservative but pertains mainly to the loop side closer to the front panel (pointed end of the arrow).

The access loops for the receivers are positioned between the coupled arms of the couplers and the actual downconverters. One can directly access the receivers here for 13 dB to 20 dB higher sensitivity but a correspondingly lower compression point. Note the maximum power levels posted on the instrument; these primarily apply to the receiver inputs (the pointed end of the arrows). Amplifiers or pads can also be inserted into these loops to bias the receiver response towards lower or higher expected signal levels, respectively. Again, the noise floor (in port-referred dBm terms) and the compression point will move accordingly as networks are added here. Generally the 0.1 dB compression point at these receiver inputs will be at least 0 dBm.

30-4 Option 61/62 and Step Attenuators

The final VNA option in this chapter involves the addition of step attenuators (as well as loops since Options 51, 61, and 62 are mutually exclusive and all add loops). This addition is useful for achieving substantially lower port powers than the normal ALC range allows and to facilitate the testing of higher power DUTs that might compress the instrument receivers natively. The block diagram is shown in Figure 30-7 on page 30-10 and there are several important things to note:

- Option 61 includes only two step attenuators (Port 1 source and b2 test), as might be commonly needed in amplifier test. Option 62 includes four step attenuators (both source paths and both test paths), as would be needed when gross power and level control is needed in both directions.
- The attenuators are positioned after the loops on the receive (test) paths, in the sense of normal signal flow, so they can be used to alter signal levels of whatever devices are placed in the loop paths. The attenuators are before the loops in the source paths to allow control of exposed source power at the loop ports.
- The attenuators have 10 dB steps and a maximum attenuation of 70 dB (for the MS4642B and the MS4644B) or 60 dB (for the MS4645B or the MS4647B). These values are not precise and the factory ALC calibration does not take the source side attenuator values into account (so the Effective Power is approximate). If precise drive levels are needed, a power calibration is recommended. For absolute levels below –25 to –30 dBm, a diode-based power sensor will be required. For levels below ~ –45 dBm, adding test attenuation will help (on the power being power-calibrated) as there is some LO leakage power at the port that can interfere with low level integrated power measurements like a power sensor will perform.
- The receive-side attenuation settings are accounted for in the factory receiver calibration so the unratioed parameter readings will still be roughly correct if the attenuators are engaged.
- These attenuators are electromechanical devices and hence have a repeatability of only about 0.1 dB. If the attenuator state has been changed after a calibration has been performed (such as a 2 port SOLT calibration), the calibration should probably be redone even if the attenuator is later returned to its original value (unless the calibration is only being used for coarse measurements).
- Also, since these devices are electromechanical, they have a finite switch lifetime of ~3 million cycles. Because of this, the attenuator settings are handled in a unique way by the instrument: the settings are per-system (meaning that they cannot be changed between channels and are only saved as part of the .cha setup file and not as part of the .chx per-channel setup file). To allow the settings to change between channels could mean the attenuators switching at a very rapid rate (since channel change time can be as short as a few ms). If attenuators are used as part of the setup, saving and recalling the .cha file is recommended.
- Also since the attenuators are electromechanical, they do have a finite settling time (scale ms to a 0.1 dB level). In certain automated test environments with high resolution needed on insertion loss, additional settling time may need to be incorporated into the test sequence.
- The attenuators can handle 2W of power dissipation so may limit power handling even when engaged.



The block diagram showing Option 62 is illustrated here. Option 61 includes step attenuators in two positions while Option 62 has step attenuators in four positions as shown. Only front panel loops are shown in this figure, which are for frequencies ≥ 2.5 GHz. The loops for < 2.5 GHz frequency access are located on the rear panel.

Figure 30-7. Option 62 and Step Attenuators

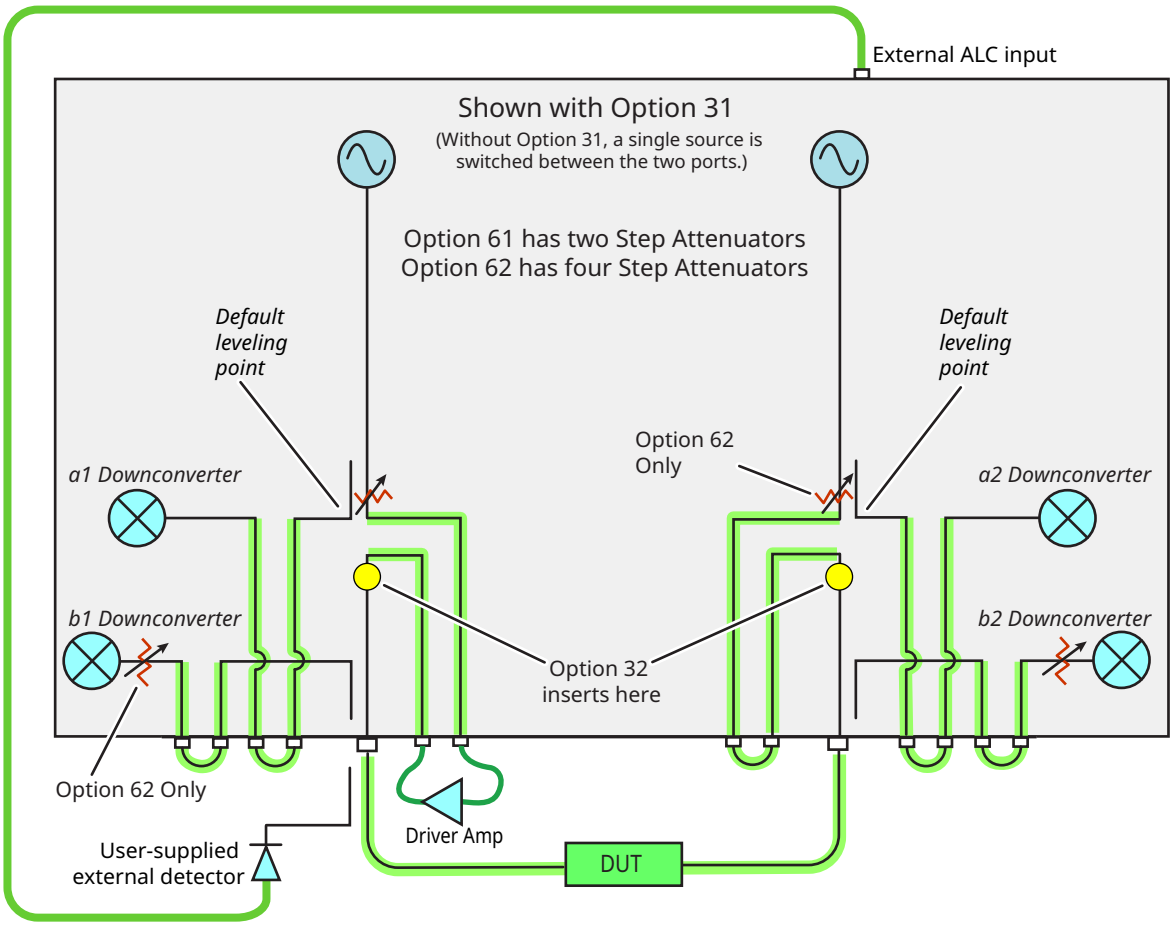
30-5 Option 53/8x and External ALC Access

In normal operation, leveling is controlled by a detector situated in the coupled reference arm (or at the source itself for frequencies under 2.5 GHz). There are cases when it might be useful to have the detection point be somewhere else:

- If an external pre-amplifier is being used (in the source loop or at the VNA test port), it may be useful to level at the output of that amplifier to account for the drift of that amplifier. In this case, usually a coupler+detector pair would be placed somewhere after the amplifier although some candidate external amplifiers have built-in power detection.
- The DUT drive chain may involve a frequency converting stage and it may be desired to level at that converted reference plane.
- There may be a very long cable run to the DUT and maintaining a specific drive to the DUT is important. In this case, it may be desirable to place a remote detector to level at a point closer to the DUT (and since the detected return signal is at much lower frequencies, it may be less impacted by the distance but one does still have to be aware of noise ingress and IR losses).

The external ALC port on the rear panel is also used for broadband/millimeter wave applications (see Chapter 18 of this guide) but is activated globally with Option 53 (note that when in 3739 broadband/mmWave modes, the external ALC connector takes on the mmWave leveling role). As suggested by Figure 30-8, the default leveling point is before the source loops. This is to prevent the power from railing to a high level when the loop is opened under normal operating conditions. However, it may be desirable to level after any driver amplifier that may be inserted into the loop. An example setup is shown in Figure 30-8. Note that either an external coupler and detector arrangement is needed or an external detector must be connected to the reference loop path via a splitter, hybrid, or similar device. While many detectors can be used, the following should be kept in mind:

- The detector must be negative going with an allowed range of about 0 to –1.3 V (–1.3 V corresponding to highest power observed). The damage level is larger (> +15 V and < –15 V) but the internal DAC comparison circuit will no longer respond for voltages more negative than about –1.3 V or for positive voltages.
- The external ALC input has a floating ground with impedance to chassis ground of ~3 kΩ. The effective input impedance is ~2 kΩ.
- When activated, the power level may be very low or very high since the system has no a priori knowledge of the detector's characteristics. Measurement or user power calibration of the port power should be done before connecting a DUT.
- If the detector is not connected and external ALC is enabled, the system will have no way of knowing of the absence and that power will rail to a high level. Again, it is especially important when using external ALC to verify power delivery and control before connecting a sensitive DUT.
- Since external leveling is most commonly needed with active applications, external ALC access requires the presence of Option 61 or 62 (loops with step attenuators) discussed earlier in Section 30-4.

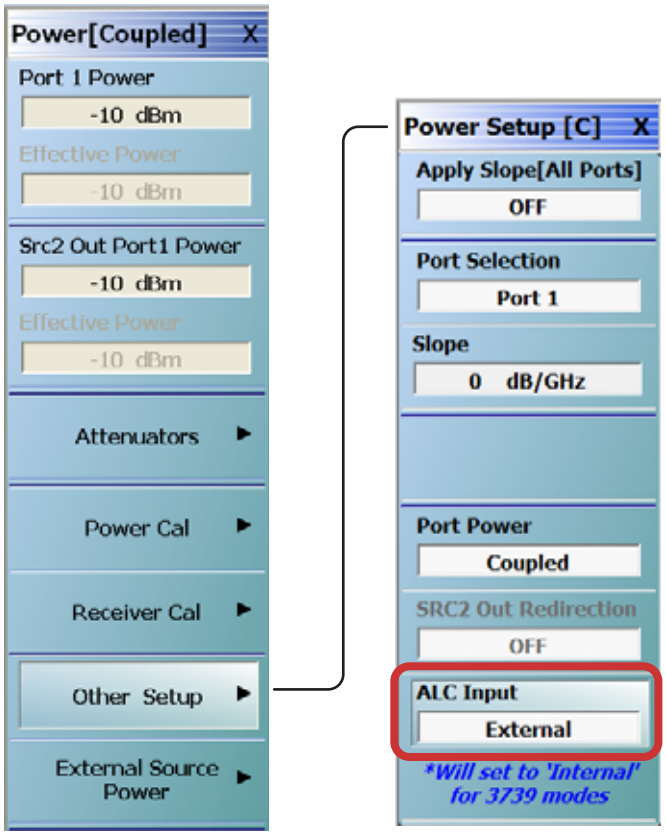


An example setup using an external ALC input is shown here.

Figure 30-8. Option 53, External ALC Access

As might be expected, user power calibrations are quite important when using external ALC. As discussed with Option 51, use the Entered Power value and the Target Power value to not only get the desired power but to give the instrument a hint about how much drive is going to be needed to get to the target level. Depending on the sensitivity and range of the detector employed, the power range that can be calibrated may be limited so some experimentation may be required.

The External ALC use is activated from the Other Setup submenu under the Power menu as shown in Figure 30-9 on page 30-13. Again, this behavior is overridden when in the broadband/mmWave modes that use the external ALC port for mmWave power control. When in one of those 3739 modes, the ext. ALC connector on the VNA should be reconnected to the 3739X test set.



External ALC control is activated under a power submenu. This menu button is only available with Option 53 or Option 8x installed.

Figure 30-9. ALC Input Control Selection

30-6 Summary

A series of RF deck options have been presented in this chapter that are useful for a variety of applications where drive and receive signal levels must be manipulated. Power amplifier measurements (and for amplifiers in general) can often benefit from a number of these options to increase pre-DUT drive power and reduce levels into the receivers to avoid compression. Mixer LO power needs, high loss measurement scenarios (including on antenna ranges), modulated measurements and other uses can also benefit. In all of these cases when levels are being modified, additional flexibility in how power leveling is performed augments the other features.

Chapter 31 — Spectrum Analysis (Option 49)

31-1 Introduction

The Spectrum Analysis option (Option 49) in the MS464XB allows for convenient spectral domain measurements of quantities such as harmonics, spurs, other distortion products, and general frequency content of a variety of a devices. The user interface is configured in a way closer to that of classical spectrum analyzers, but a number of VNA-like behaviors are available, including many different sweep modes, source drive, power and receiver calibrations, multiple receiver measurements, and multiple source control for complex frequency plans. This chapter describes the operational characteristics of this application along with some example measurements.

31-2 Principles

In a fundamental sense, a network analyzer always was capable of spectrum analysis assuming one could decouple source behavior from receiver behavior. Indeed, this was the intent behind features such as Multiple Source Control (Chapter 18 — Multiple Source Control (Option 7)). Still, the execution of typical spectrum-analyzer-like measurements within a VNA could be inconvenient and lead to unexpected results and/or slow execution time. The Spectrum Analysis option codifies some of the classical setup requirements for common spectral measurements, optimizes sweep and measurement behavior, and provides data reduction behavior that is more appropriate for this class of measurements. The hardware is essentially unchanged from non-optioned units so the inherent front-end characteristics of a VNA will be retained (very good linearity but higher noise figure, less spurious and image rejection) as will the measurement flexibility (internal sources that can be programmed where desired, multiple receivers, correction capability) so the most favorable applications will perhaps be slightly different than for a classical, dedicated spectrum analyzer. A comparison of the VNA front end and that of a ‘classical’ spectrum analyzer is shown in Figure 31-1. This is not a global statement since some VNAs have RF-side gain control, etc. but rather it is intended to illustrate some common differences in the receiver designs, at least historically.

Common measurements will include harmonics, tone-based distortion measurements, some spurious measurements, and spectral occupancy tests on devices that may include multiple outputs or input-output comparisons. Internal (and controlled external) sources can be used for stimulus in a flexible manner so measurements requiring a complex stimulus and or frequency conversion analysis are also possible candidates. Some more difficult measurements may include detailed spurious hunts (as the internal sources and LOs do not have the level of spectral purity of some spectrally-focused analyzers), wide-bandwidth modulations tests or detailed analysis of non-stationary modulated signals. Note also that harmonic conversion is used by the VNA receivers at some frequencies and there is no pre-selection, so there will be harmonically-related responses observed.

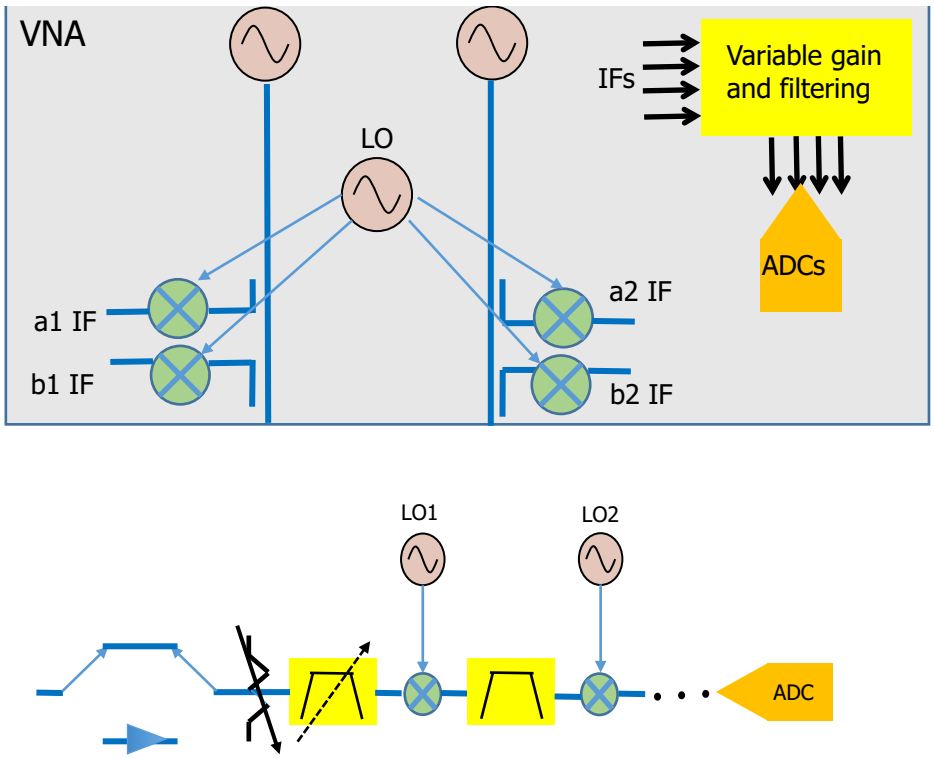


Figure 31-1. Simplified comparative block diagrams for a representative VNA (top; multiple sources and receivers, simpler front end) and an example spectrum analyzer (bottom; more RF-side gain and filtering, single channel usually).

31-3 General Operation

Note that Spectrum Analysis is treated like an application in the MS464XB (like Noise Figure and Pulse) so setups are not carried over from the other applications when entering. The application will preset to a single trace display (Power Out graph type with units of dBm, b2/1 as the response parameter) but previous spectrum analysis .chx files can be recalled. All applications are per-channel so one can configure other channels as Transmission/Reflection (regular S-parameter measurements), Noise Figure, etc.

The basic menu upon entering the application is very simple and mainly consists of shortcuts to the most commonly used other menus for this application: frequency, scale, averaging and marker (see Figure 31-2). These menus have some changes from what sees in other applications (e.g., RBW is displayed instead of IFBW on the averaging menu, some processing choices are not available, etc.). The first button opens a configuration dialog that is discussed in detail below. The Sweep Mode button toggles between Classical (spectrum analyzer-like) and VNA-like. The difference between the two modes relates to the measurement point density and is also discussed below.



Classical Mode



VNA-Like Mode

Figure 31-2. Main Spectrum Analysis Setup Menu

The setup for the spectrum analysis application is a bit different from that in the other VNA applications and the nomenclature has shifted in places to be closer to that of a classical spectrum analyzer, but the operational aspects are similar. The first tab of the configuration dialog (Spectrum Analysis Setup) is shown in Figure 31-3.

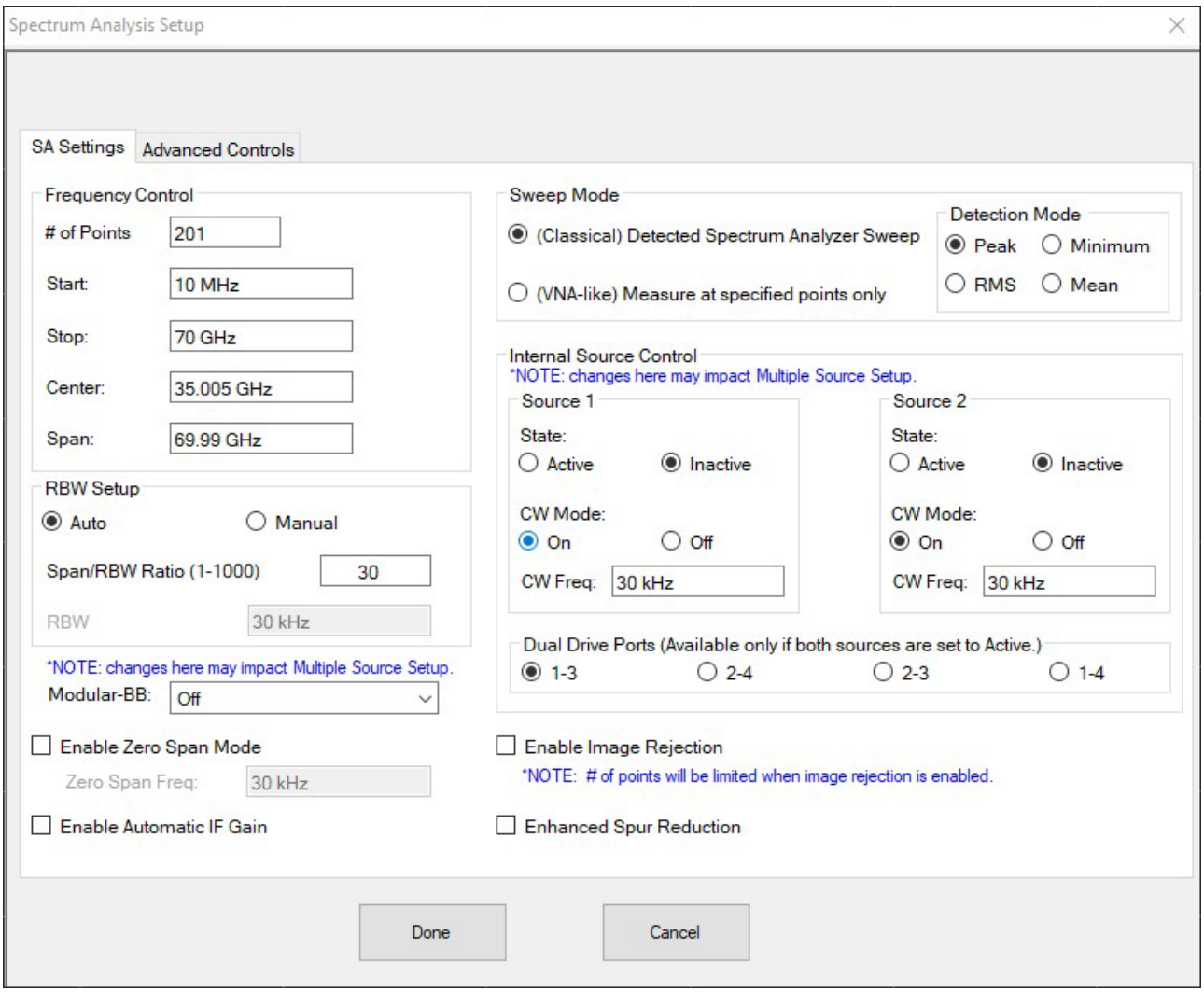


Figure 31-3. First Tab Of the Spectrum Analysis Setup Dialog

This happens to be for a four-port system, but the only addition relative to that for a two-port system is the ‘Dual Drive Ports’ section (that selection will also not be available for a single-source four-port system).

Sweep Mode: This selects how the instrument will actually sweep over the requested frequency range. With the 'VNA-like' mode, the instrument will measure only at the points in the frequency list (and segmented sweep is available). Thus, signals not precisely at those points (or within a RBW or so) will not be seen. This mode is fast and ideal when the frequencies of interest are known (or when the step size is on the order of a RBW, so that more of the complete spectrum can be captured). The 'Classical' mode is more like a traditional spectrum analyzer in that all frequencies in the range are scanned with adjacent RBW-sized chunks. Thus signals will not be missed, but it can be a much slower measurement. More total points are allowed than in TR and other applications since multiple sweeps are employed behind-the-scenes. There is a cap on the number of measurements in a sub-sweep (those measurement associated with a single display point) of 25000 or 100000 (and half that if image rejection is enabled). If the sub-sweep would require more measurements than that, the measurement count will be truncated. Note that sub-sweeps require more time for frequency table recalculation so the sweep rate is more irregular and a message will appear in the status bar to note the altered sweep behavior.

When in this 'Classical' sweep, the detection mode is relevant in that it determines how the measurements between display points will be reduced to a single value for that display point.

Peak: The measurement with the maximum linear magnitude of the measurements in the subrange of the given parameter will be displayed. This is the default selection.

Minimum: The measurement with the minimum linear magnitude of the measurements in the subrange of the given parameter will be displayed. This is useful for very dense signals in the sweep range of interest.

RMS: The root-mean-square value of the linear magnitudes of the given parameter will be used as the result for the subrange. This can be useful for noise-like signals.

Mean: The arithmetic mean of the linear magnitudes of the given parameter will be used as the result for the subrange. This can also be useful for noise-like or stochastic signals or those that are otherwise very dense.

To summarize the sweep modes:

VNA-like: Fast and measures only at the frequency points specified. All frequency-domain sweep types are permitted. For complete spectral coverage, the step size needs to be on the order of a RBW, but this is not always needed.

Classical: Measures over the entire range specified with actual measurements dictated by Span/RBW ratio. This is a slower measurement process but will not miss signal content. Only linear frequency sweep and log sweep types are available.

As an example, consider the four spectra shown in Figure 31-4. In all cases, a signal at 28 GHz was applied to the VNA and a displayed set of 51 points were used with a step size of 2 MHz and the RBW was 1 MHz. Channels 1 and 2 are in the VNA-like point mode while channels 3 and 4 are in the classical mode with the peak detection selected. In channels 1 and 3, the center was set at 28 GHz so a display point corresponded to the input signal and the signal is clearly visible. The noise floor looks a bit different in the classical mode since the peak values in each subinterval are displayed. In channels 2 and 4, the center was set to 28.003 GHz so 28 GHz is not a display point frequency. One can then see the signal is distorted in the VNA-like mode (channel 2) since no measurements are made within more than an RBW of the input signal. In channel 4, the signal is still observable with the correct magnitude since the subinterval measurements capture the tone.

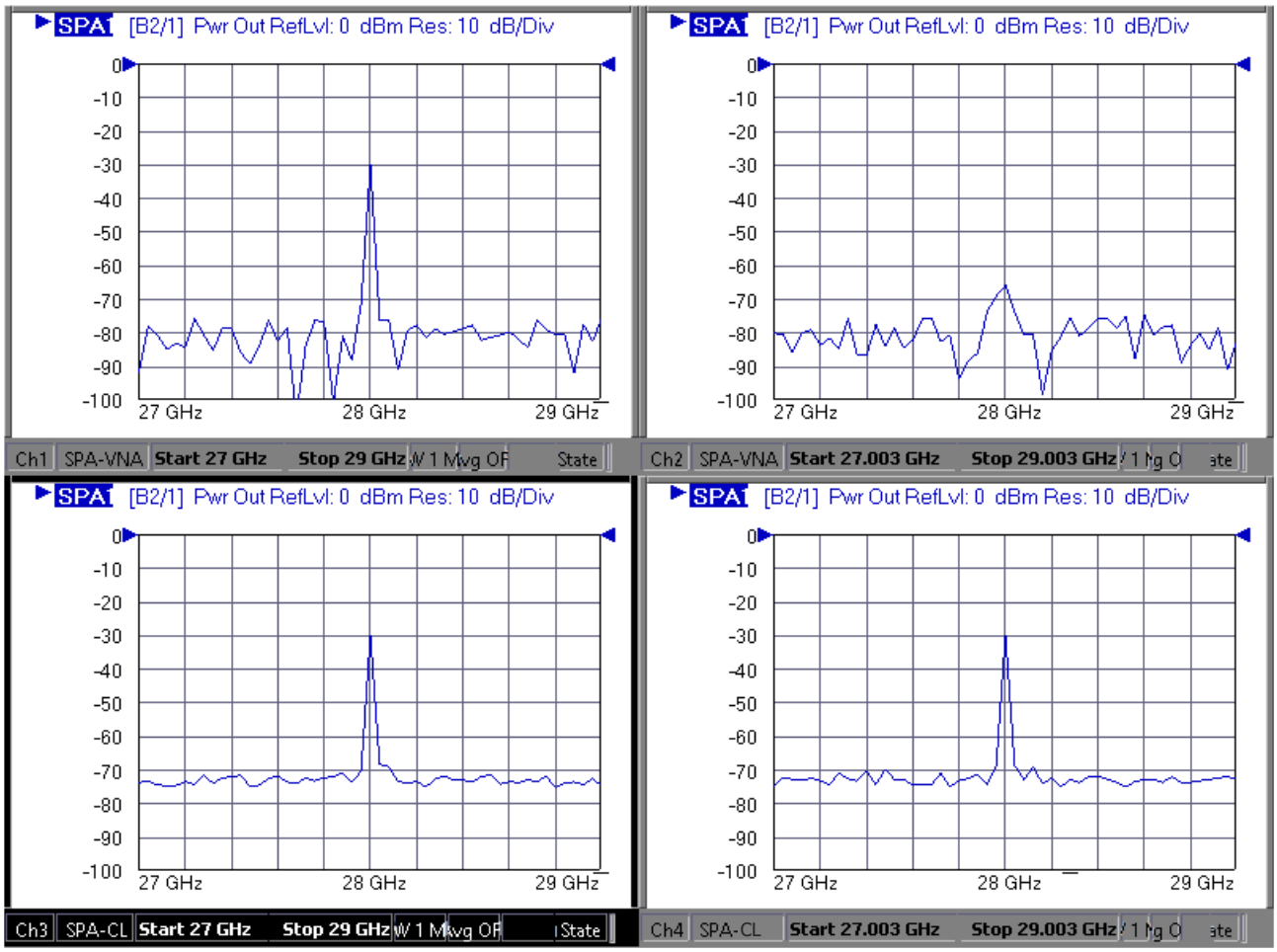


Figure 31-4. Spectrum Analysis Modes: VNA-like and Classical

Frequency Control: These entries mimic those on the main frequency menu and are placed in this composite dialog for convenience. The number of points represents the data points to be plotted (or saved, etc.) which, in the case of a Detected Spectrum Analyzer Sweep (see Figure 31-4), is smaller than the number of measurement points.

RBW Setup: This is the analyzer IF bandwidth to be used in the measurement. If Auto is selected, the bandwidth will scale with the span using the entered ratio (the default value of this ratio is 100; if the span is very large at all, the RBW will max out to 1 MHz). If that computation results in a RBW < 1 Hz or > 1 MHz, the RBW will be limited to the endpoint values. If Manual is selected, the RBW is entered directly in the range of 1 Hz to 1 MHz. Note that, at this time, there is no direct equivalent of video bandwidth (VBW) but that behavior can be emulated to some degree with sweep-by-sweep averaging.

Internal Source Control: By default, the internal source(s) of the VNA is/are turned off when in the spectrum analysis application but it/they can be activated to serve as a stimulus. If generically activated (CW mode OFF), the source(s) will track the receiver much like a tracking generator would. The power level of the source(s) can be set on the usual Power menu. If CW mode is ON, the source will be parked at the indicated frequency for the entire measurement cycle. If a dual source four-port instrument is being used and both sources are activated, the dual-drive ports selection will be relevant and will dictate which ports are sourcing (this overrides any parameter-related driving port selection).

If a different frequency plan is desired for the internal source(s) or if external synthesizers are to be used, the multiple source control system can be used. See Chapter 18 — Multiple Source Control (Option 7) for more information.

Modular BB: This enables selection of mm-wave banded and broadband operation (ME7838X systems) and will change the allowed sweep range limits accordingly. If variations on these modes are needed (or other, 3rd-party mm-wave modules are in use), multiple source control can also be used. See Chapter 17 — Broadband/mmWave Measurements (Option 7, Option 8x) and Chapter 18 — Multiple Source Control (Option 7) along with "Additional Comments On Spectrum Analysis With mm-wave Modules" on page 31-16 for more information on both the broadband systems and the use of multiple source control. The basic selections on this pull down menu are shown below and include all of the ME7838 broadband systems as well as a number of the banded variants.

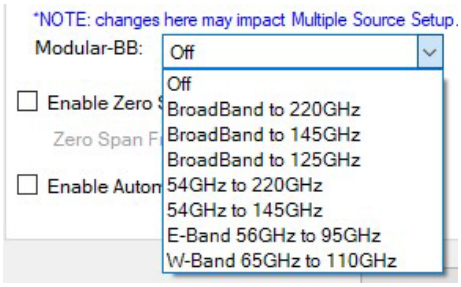


Figure 31-5. Modular BB Selection Pull-down

Different broadband and banded mm-wave system choices can be enabled here, which will set the receiver (and sources if used) appropriately.

When in these modular BB modes, the receiver will be programmed according to the plan for the module selected (including the use of the VNA receivers below a nominal 30 GHz for the broadband modules). If the source(s) is/are active, it/they will also use the frequency plan for the module selected. If a more complex relationship is desired, then multiple source control must again be used.

Note that factory receiver calibrations are not available for the modular BB configurations so performing a user receiver calibration is normally recommended.

Enable Zero-span Mode: This is the equivalent of VNA CW mode where the receiver is parked at a specific frequency but multiple measurements are displayed at that frequency (the x-axis labels can be in units of time; this can be changed to just a measurement count index under the Display menu). Note that unlike in the Transmission/Reflection and other applications, the zero-span frequency will be enforced (at the expense of start/stop on the frequency menu) and will not be coerced to the nearest frequency in the existing start/stop list.

Enable Automatic IF Gain: The automatic IF gain system changes gain levels based on the received IF signal so it may be undesirable for spectrally-broad or complex signals (can lead to the IF gain being set too high and compressing) so this is defaulted to OFF in the spectrum analysis application. If measuring a spectrally tight signal in-band or if measuring noise-like signals or if using the VNA source as a tracking source, this gain function can be activated which will improve the noise floor.

Enable Image Rejection: While the front-end of a VNA receiver is simple, a level of image rejection can be achieved by scanning both upper and lower response locations which is what this checkbox enables. This function does lengthen sweep time but can reject the image response by over 80 dB for simple signals. It is not as appropriate for non-stationary or signals with bandwidths of more than about 50 MHz. When in VNA-like mode, selection of this function reduces the maximum point count to 12500 (50000 if 100K mode is active). In classical mode, it doubles the number of sub-sweeps and increases sweep time.

For most frequencies, the instrument IF is 12.35 MHz so an image response will appear 24.7 MHz away if this box is not checked. By acquiring additional information with another measurement at an offset LO, it is possible to separate out desired from image responses. An example, with a single sinusoidal tone at 32 GHz as input, of the results with and without image rejection enabled is shown in Figure 31-6.

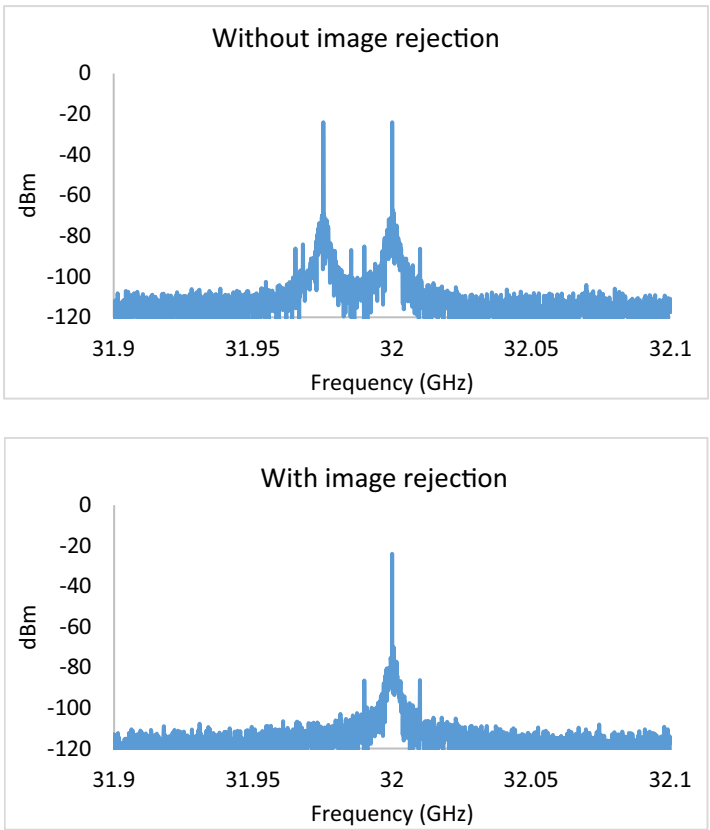


Figure 31-6. The effect of image rejection is shown here for a simple single-tone input. Image rejection does slow down sweep speed and places some constraints on point count.

Because of the additional measurements required for image rejection, the maximum measurement point count is limited to 12500 or 50000 instead of 25000 or 100000 for the VNA-like mode. The display point limit for any given configuration of the classical detected operating mode will similarly be reduced by a factor of two. For the same reason, image rejection will also slow the sweep time.

Enable Spur Reduction: When enabled, spur reduction will use a more elaborate system of checks to reduce (particularly) self-conversion spurs from the LO and spurs generated between the source(s) and LO when the former is/are active. There will be no significant improvement in spurs created by external signals interacting with the receiver. Enabling this function does increase the frequency table computation time so it will take a bit longer for the sweep to start after a change is made to the frequency list. There is no impact to per-point measurement time after the sweeps start.

Note also that the front-end of the VNA is not preselected, so harmonic responses of the converters will appear on-screen. As an example, suppose the user has input a 7 GHz input so the internal LO will be at (7GHz +IF). If the receiver is sweeping at higher frequencies where the same physical LO is used (e.g., displayed frequency of 21 GHz + 2 IF) then another receiver response will appear. This is another case where the VNA-like sweep mode can be useful in identifying true harmonics.

The advanced controls tab is shown in Figure 31-7. These controls pertain only to the case when internal source(s) is/are activated.

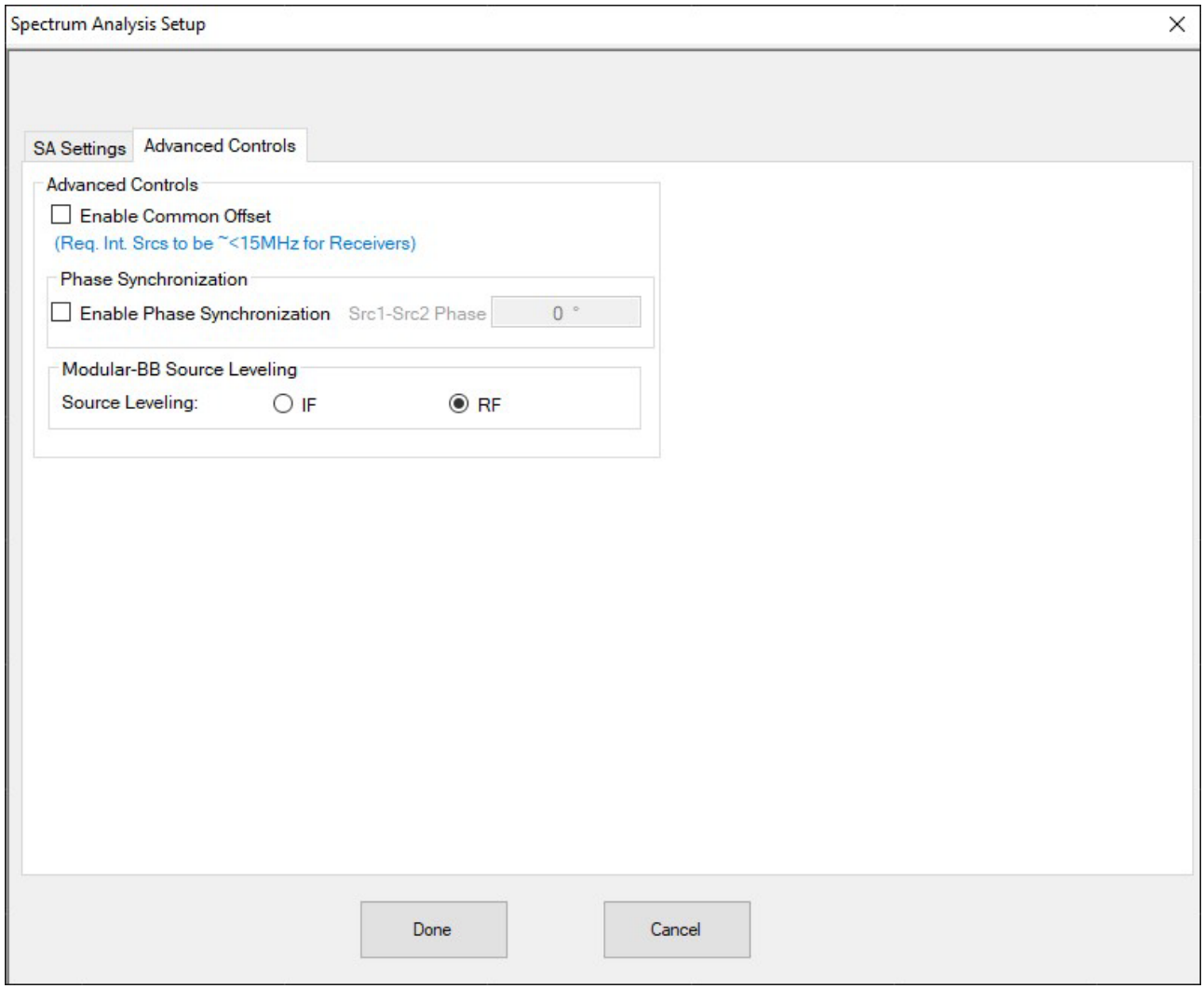


Figure 31-7. The second, and more advanced, tab of the Spectrum Analysis Setup Dialog

Enable Common Offset: This is a control that alters the detailed programming instructions sent to the internal sources and LO. When enabled, it can reduce net IF noise levels, but can only work when the source(s) and LO are close in frequency (~<15 MHz separation before multiplication); that is, close to a tracking application.

The main benefits accrue when working at higher mm-wave frequencies using external modules. If the internal source(s) and LO are not close in frequency, lock errors may be observed and there may be sweep disruptions.

Phase Synchronization: This only applies in dual source systems when both sources are active. This selection changes the synthesis configuration so the sources will have a controlled and repeatable phase relationship. This may be useful when the sources are being used to stimulate an active differential device (see Chapter 26 for general information on the topic of true mode drive). Another application may be to compare phase relationships in cross-polarization or leakage experiments (note that phase measurements are possible in the spectrum analysis application but generally the sources will be put in a tracking relationship to the receiver). The phase relationship is driven by a factory phase calibration at the test ports (see Chapter 26 for more details).

Modular BB Source Leveling: If a modular BB mode was selected on the previous screen (and a source is made active), this control defines how power control leveling is performed at mm-wave frequencies (generally above 54 GHz). IF leveling requires the source be tracking the receiver (otherwise it will go unleveled and rail to high power). RF leveling can be used in general but has less control range than IF. The Power menu is used to enter the requested source power and user power calibrations are available.

31-4 Receiver and Power Calibrations

The base VNA factory receiver calibration will be on by default so the unratioed response parameters can read-out correctly in dBm (and units will be displayed as such with markers on Power Out and Power In graph types) referenced to the test port of the instrument. For power reference planes at some other point (e.g., at the end of a cable), a user receiver calibration is required as discussed elsewhere in this Guide. As of the writing of this chapter (mid 2022), factory receiver calibrations do not exist for ME7838X systems or four port systems so user receiver calibrations are always needed in those cases for 'valid' dBm readings.

As with the other applications, the receiver calibrations proceed by transferring power accuracy (either from the factory ALC calibrations, which always exist, or a user power calibration). The sources will be turned on for receiver calibrations but will automatically use the full frequency list (not the source frequency only).

While it is something of a tautological measurement, the use of the internal source (acting as a tracking generator) after a receiver calibration can give some evidence of calibration completion and that there is internal consistency. A plot showing such a measurement is in Figure 31-8.

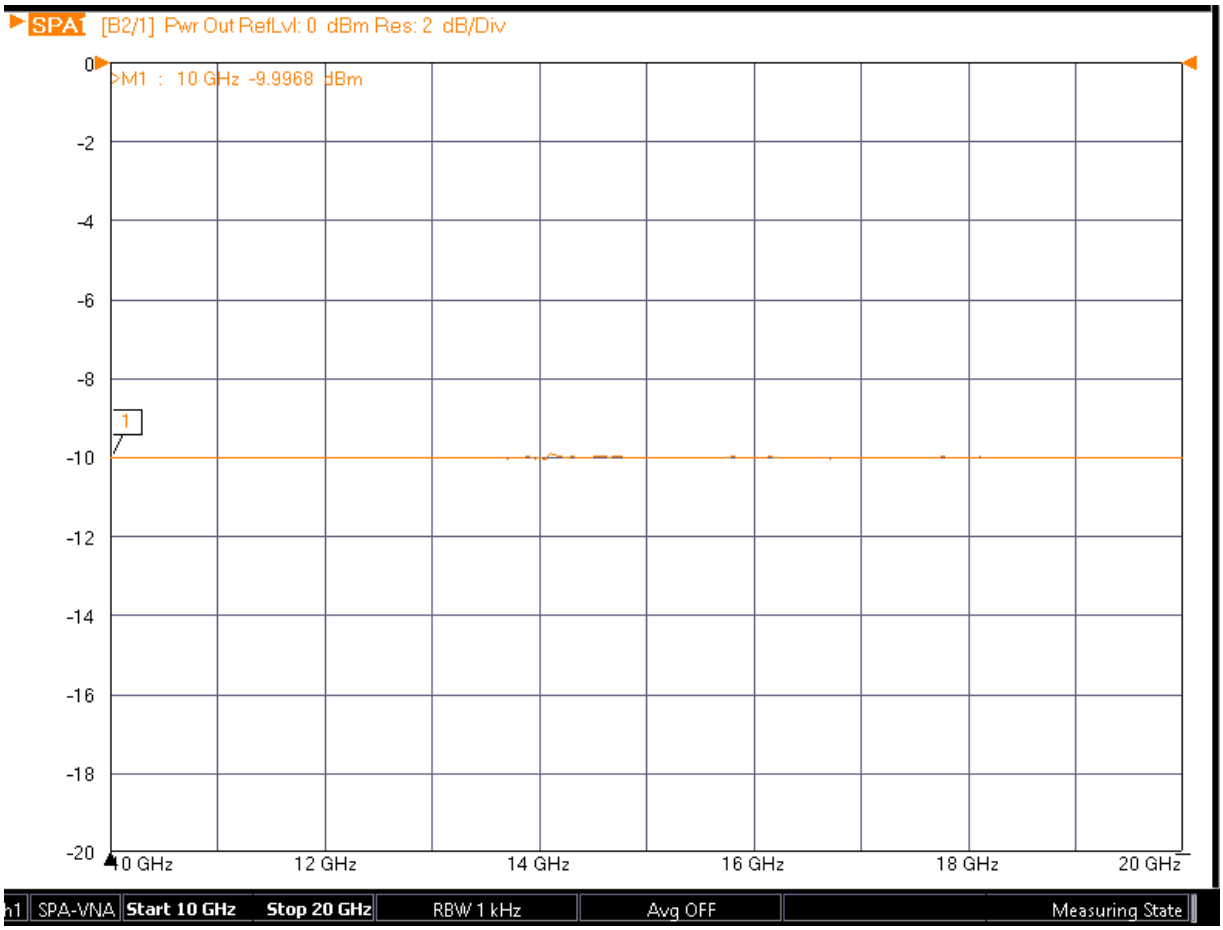


Figure 31-8. Example: Measurement with the tracking source still connected to the receiver port just after a receiver calibration.

The power accuracy of the source has been transferred to the receiver with the reference plane defined by the source power calibration.

31-5 Example Measurements and Extended Cases

A common measurement process is to look at harmonics or IMD content at the output of a DUT and the spectrum analysis application can make this very easy visually when the DUT is driven at a single frequency. External sources can be used as the stimulus or the internal source(s) can be employed. Whichever source is used, the harmonic content of that source can limit measureable DUT harmonic content. The internal source harmonics (2nd and 3rd) are generally below -20 to -30 dBc but fractional harmonics may also be present (and can be significant at mm-wave frequencies). These levels can be checked using a thru line measurement or using the reference receivers.

Similarly, source-side IMD products are possible but are generally not an issue except when measuring very low distortion products. More information is contained in Chapter 28 — IMDView™ (Option 44). For both harmonic and IMD measurements, it is, of course, always important to keep the receiver levels well below the compression point (nominally 0.1 dB compression at 10 dBm but this varies somewhat with frequency, see the Technical Data Sheet for more information).

Consider a reasonably complex measurement example using the ME7838 series of broadband/mm-wave systems to make harmonic measurements over an extended frequency range. The DUT in this case was an amplifier with two outputs and both were monitored using two VNA ports and a third port was used as the stimulus source (on a ME7838A4X four port system). The a3/1 (driving port) signal was also monitored to allow visualization of source harmonics. In this particular example, the stimulus was at 35 GHz and receiver calibrations were performed on the two test receivers of interest (b2 and b4) and the sourcing-side reference receiver (a3). The VNA-like sweep mode was used to save measurement time (since we know exactly where we are looking) and a 1 GHz step size was employed at a resolution bandwidth of 100 Hz. Note that in the spectrum analysis setup dialog, the 'Broadband to 125 GHz' selection was chosen for this ME7838A4X system.

Trace 3 (a3/1) in Figure 31-9 shows the reference (sourcing side) response and one can see the 2nd harmonic is at approximately -71 dBc (3rd harmonic below -100 dBc) and is probably not affecting the measurement. The DUT output harmonics are visible on traces 1 and 2 and one can see the 2nd harmonic is on the order of -33 dBc and the 3rd harmonic is at about -67 dBc. The DUT outputs are reasonably symmetric, but are certainly not at identical levels.

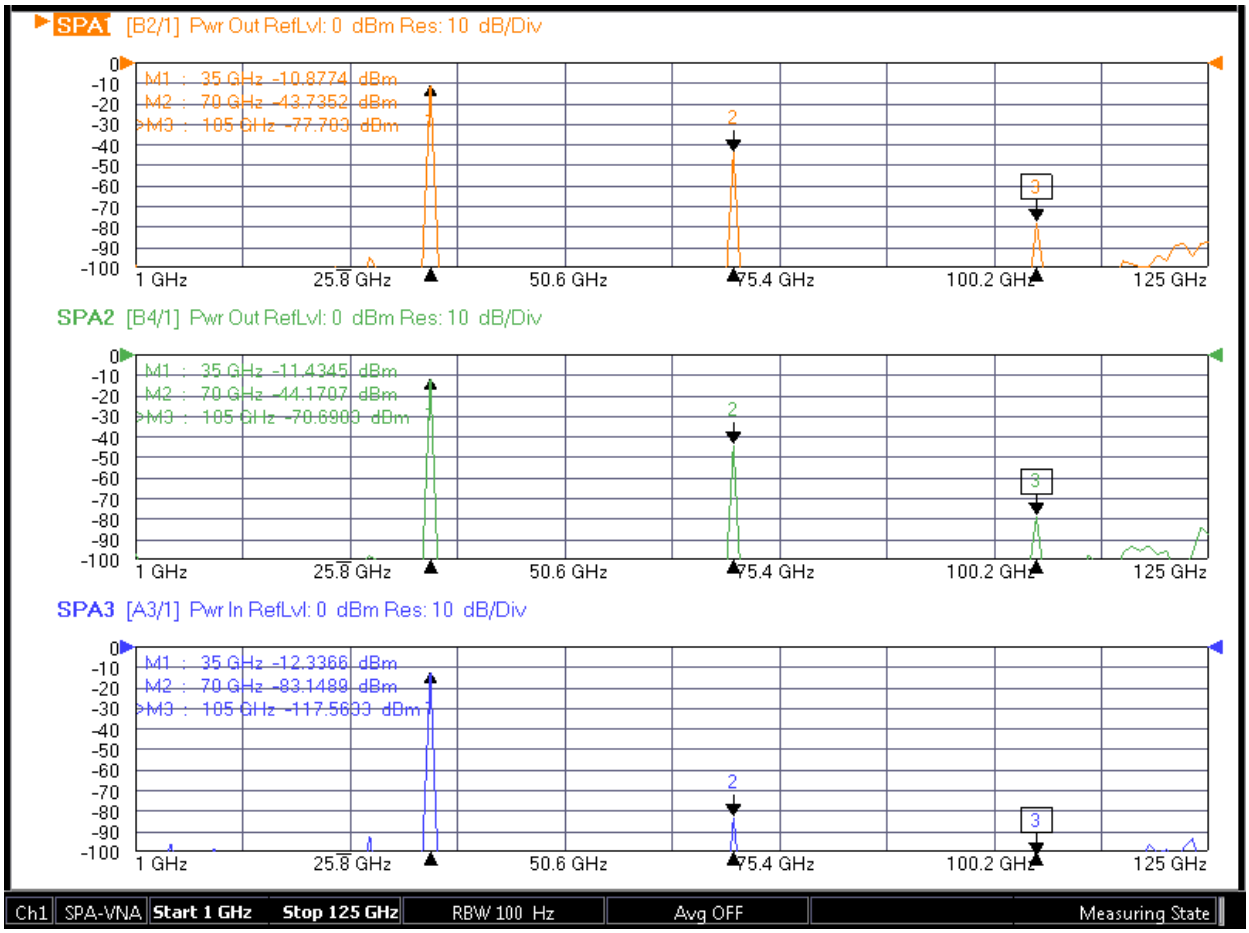


Figure 31-9. Example: Harmonic measurement for a two-output DUT.

The reference channel was used to observe the source-side input harmonics.

Another spectrum analysis use case is that of Max Hold (and, sometimes, Min Hold) where the instrument monitors a particular frequency range and the largest amplitudes of an external signal are captured across frequency. Examples include capturing the power envelope of a free-sweeping source, trying to quickly gather maximum power output over frequency of a transmitter, identifying the larger transient signals that appear vs. frequency, etc. In this VNA option, this capability is implemented using the Equation Editor and the functions MAXHOLD() and MINHOLD(). The argument for these functions is the trace containing the response variable of interest (e.g., b2/1). The sweep range of the instrument is set to anything that may be of interest for the signal source being analyzed (within the range of the hardware, of course).

As an example, a free-sweeping source, nominally sweeping over 21-29 GHz, was connected to the VNA port 2. We wished to evaluate its maximum power flatness over this range (the variation at a particular frequency was known to be small). For simplicity, the VNA was set to sweep over 20-30 GHz and Tr2 was set up with the equation editor function MAXHOLD. The result is shown in Figure 31-10. One can see the output values had a spread of about 2 dB over the 21-29 GHz range. At the extreme edges of the receiver sweep, one can see the MAXHOLD values of the noise floor. A relatively wide RBW was used for this measurement as the in-band values were of primary interest.

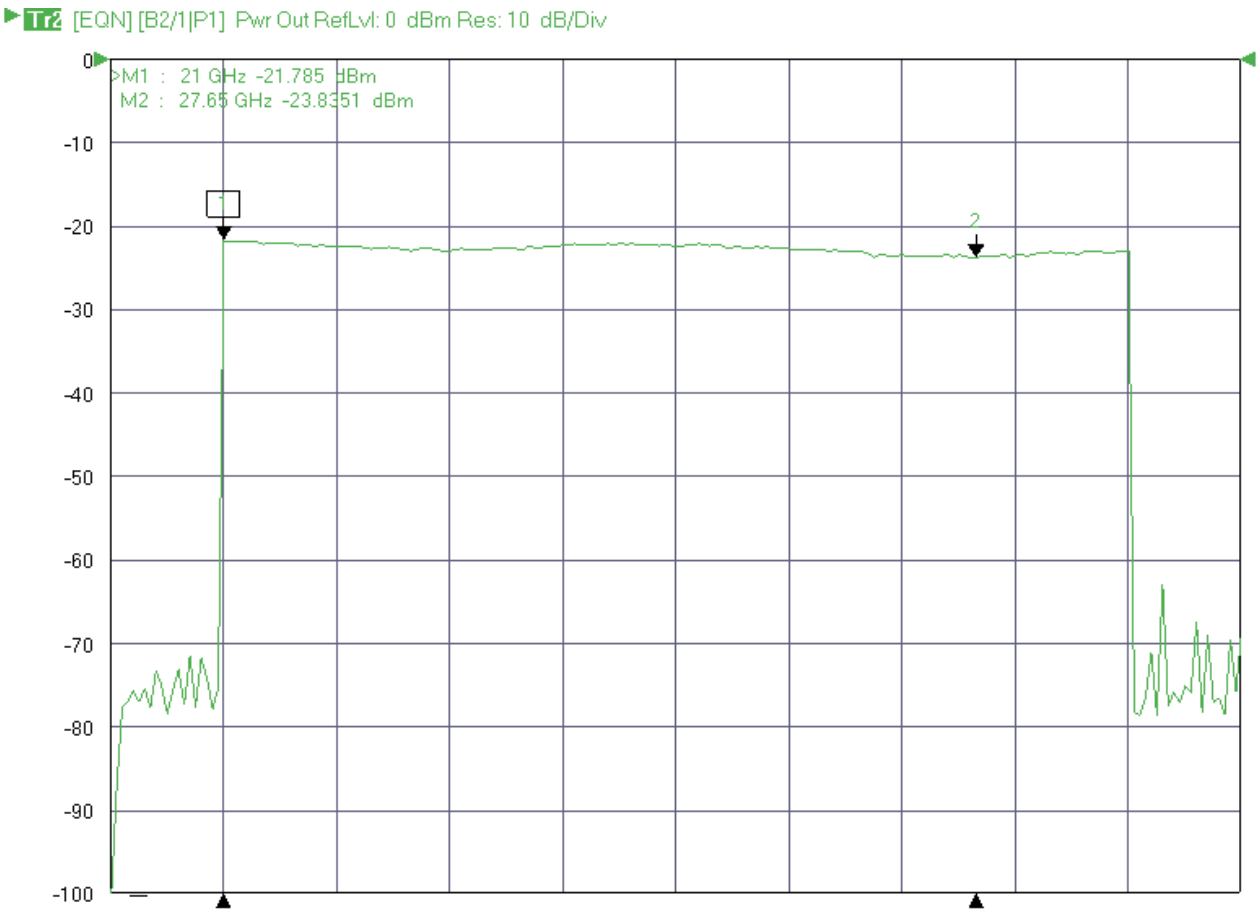


Figure 31-10. The MAXHOLD plot (note the [EQN] annunciator for the equation editor) of the output of a free-sweeping signal source.

Modulated measurements are also possible but, due to the nature of the image rejection process, modulation bandwidths wider than about 50 MHz will result in some image corruption. A simple FM example is shown in Figure 31-11 where the modulation index was 4 leading to the usual Bessel-like sideband dependence. Manual RBW control happened to be used for this measurement and sweep-by-sweep averaging was used in lieu of video bandwidth (VBW) control. The effective filtering bandwidth of sweep-by-sweep averaging is related to sweep time so will end up being sub-kHz for almost all practical setups. As always when measuring a modulated signal, consider the integrated power and power statistics when evaluating the level against the receiver compression point.

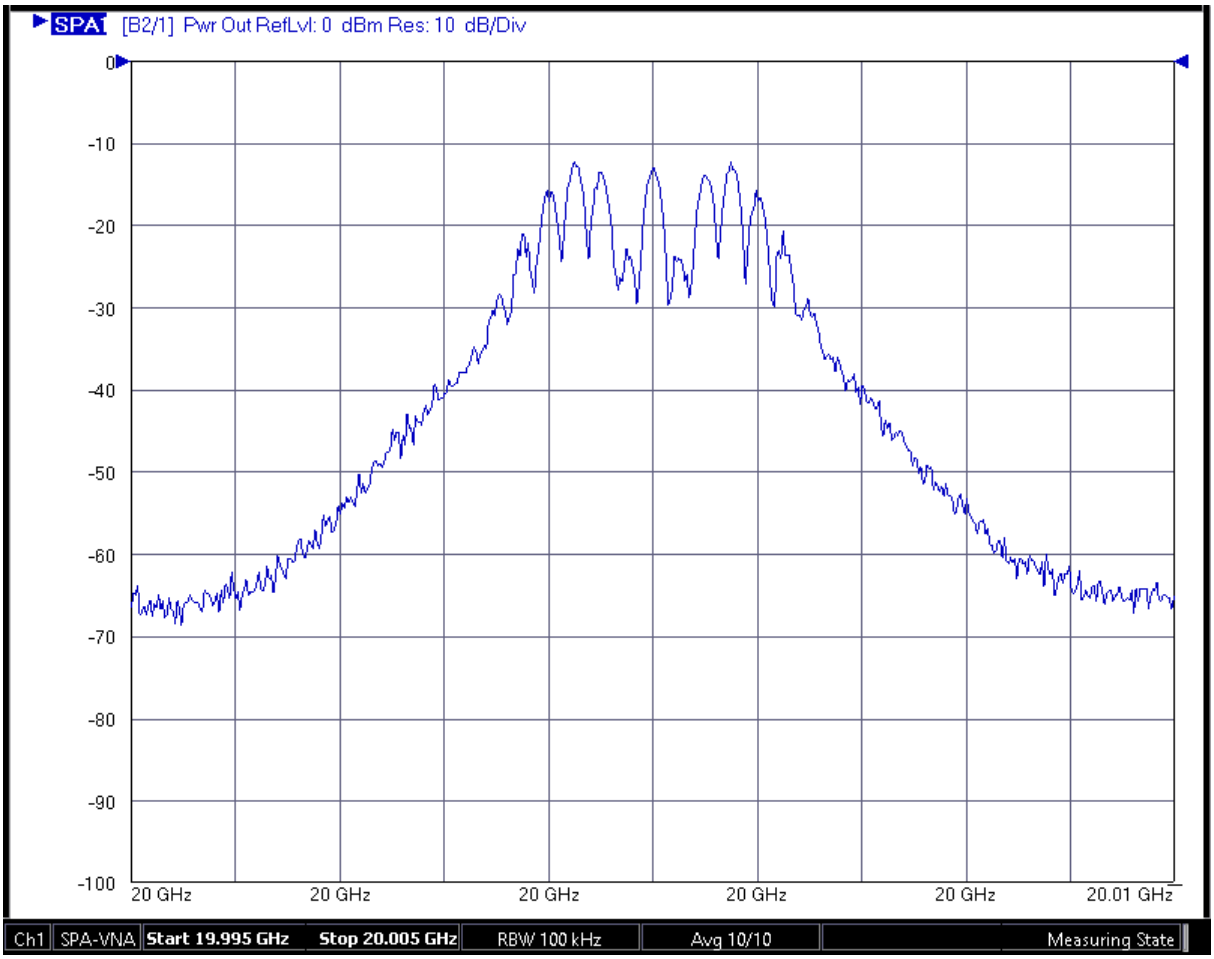


Figure 31-11. A wideband (in the sense of modulation index, not actual bandwidth) frequency modulated signal was measured.

Sweep-by-sweep averaging was used as a form of video bandwidth modification and is available in all spectrum analysis modes.

Note that the above example was with a very simple stationary signal (in the statistical sense). If a non-stationary modulation is used (i.e., most practical communications traffic) and the symbol rate is rapid compared to the measurement rate, the image rejection will degrade.

Another example reaches into the higher mm-wave frequencies where 3rd party mm-wave modules may be employed. This requires the direct editing of multiple source control entries (see Chapter 18 — Multiple Source Control (Option 7)) after entering the spectrum analyzer application. It was desired to look at the spectral characteristics around a 400 GHz signal (locked to the 10 MHz time base of the VectorStar) using mm-wave modules employing a 24 multiplication factor. As shown in Figure 31-12, the receiver and receiver source equations are set up with a 24 divisor and the receiver offset variable should be set to -(N-1)*12.35 MHz, where N is the divisor. The source was also programmed with its divisor so it could be used for receiver calibration. Note that the multiple source band definition had to be set manually for the module being used. Again, more information on the general setup and use of these modules is available in Chapter 18 — Multiple Source Control (Option 7).

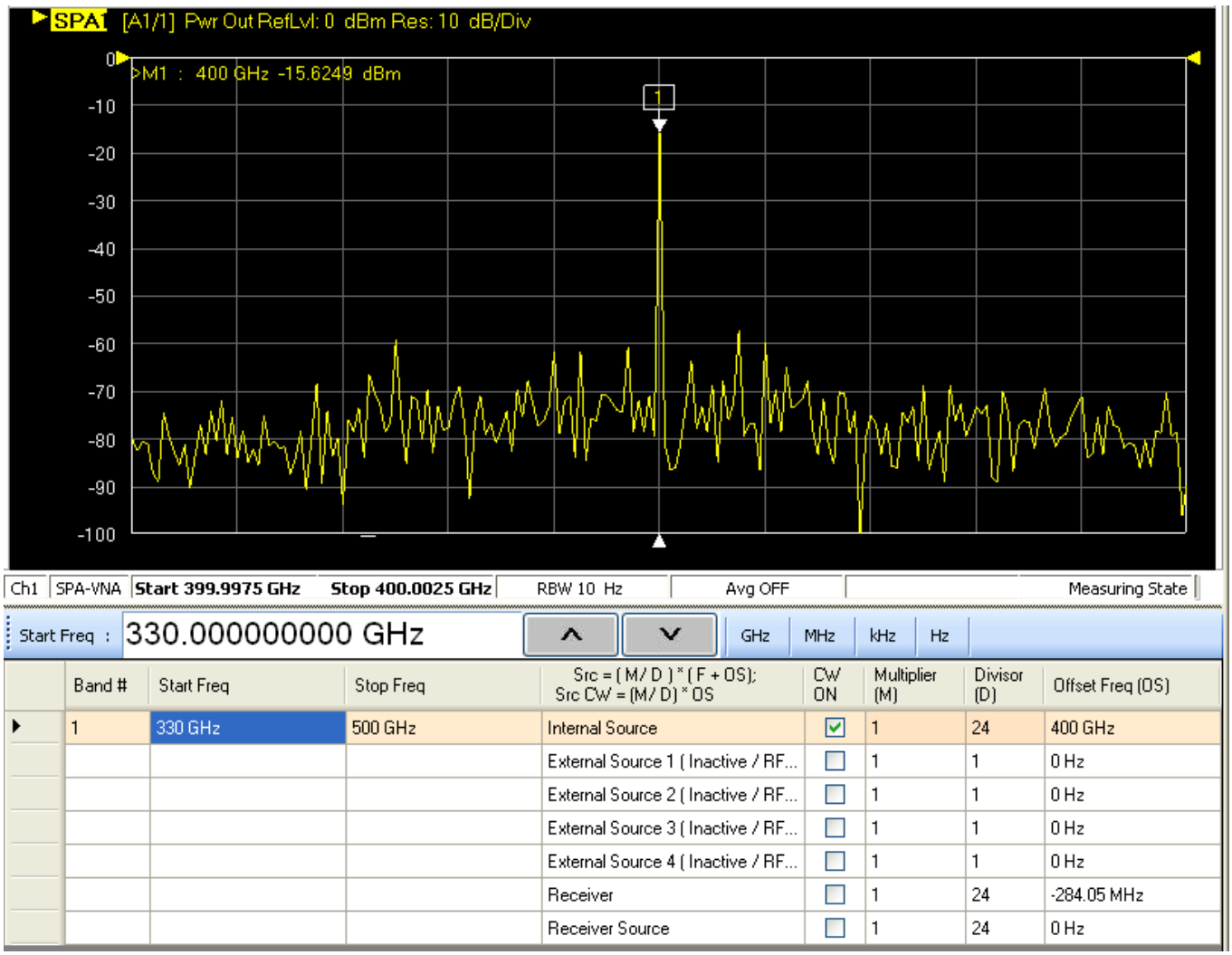


Figure 31-12. Example: Spectral Content Measurement of a Signal at 400 GHz

31-6 Additional Comments On Spectrum Analysis With mm-wave Modules

While a 3rd party mm-wave module measurement example was shown in the previous section, the details of this measurement can be nuanced so this section will explore the setup in additional detail.

As the general external module selection is not available in the spectrum analysis configuration dialog, the setup is always going to be somewhat manual in multiple source control. Considerable information on the multiple source system is provided in Chapter 18 but below are some more spectrum-analysis-specific hints.

Following the example above, here are some comments on the table entries. Of particular importance is how the receiver equation is configured. If an external source is being used as stimulus, it would have to have the appropriate equation and that source would be made active on the menu.

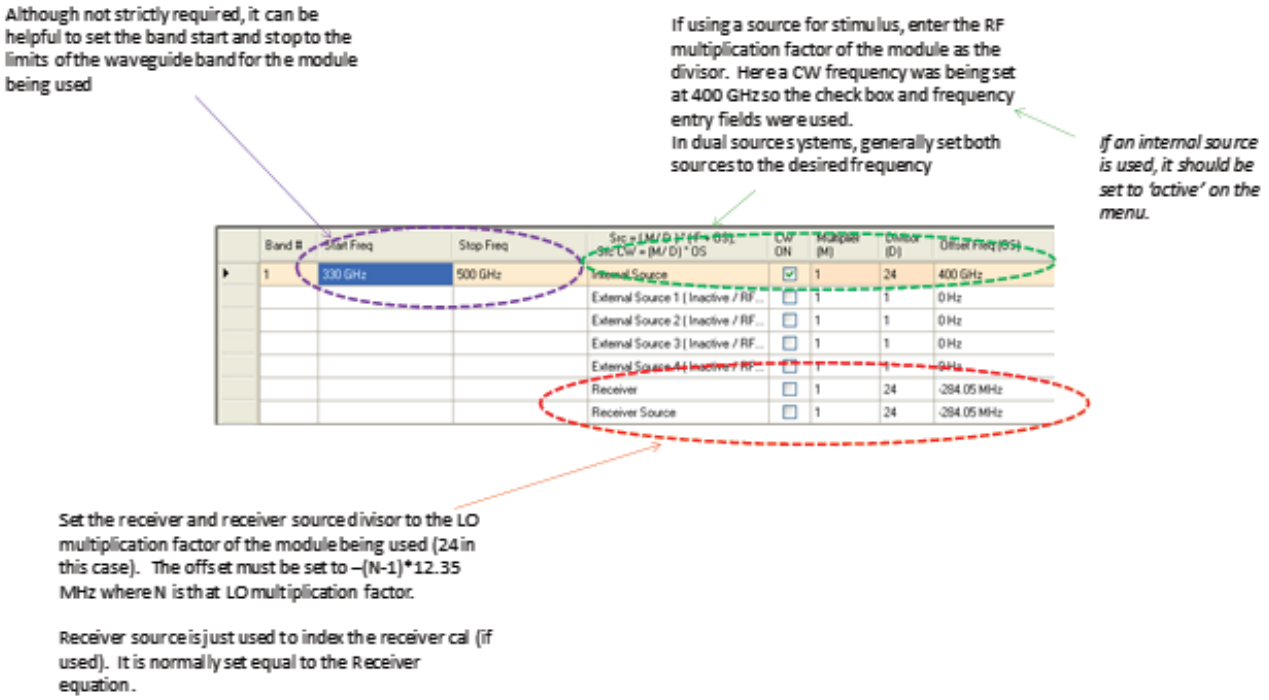


Figure 31-13. Receiver Equation Configuration

Note that external module control shown in Figure 31-14 must be used to properly setup the 3739x/3736x test set(s). This dialog is accessible from the multiple source menu. If one goes to the Spectrum Analysis configuration dialog, the settings in the external module control dialog are reset and may need to be re-edited.

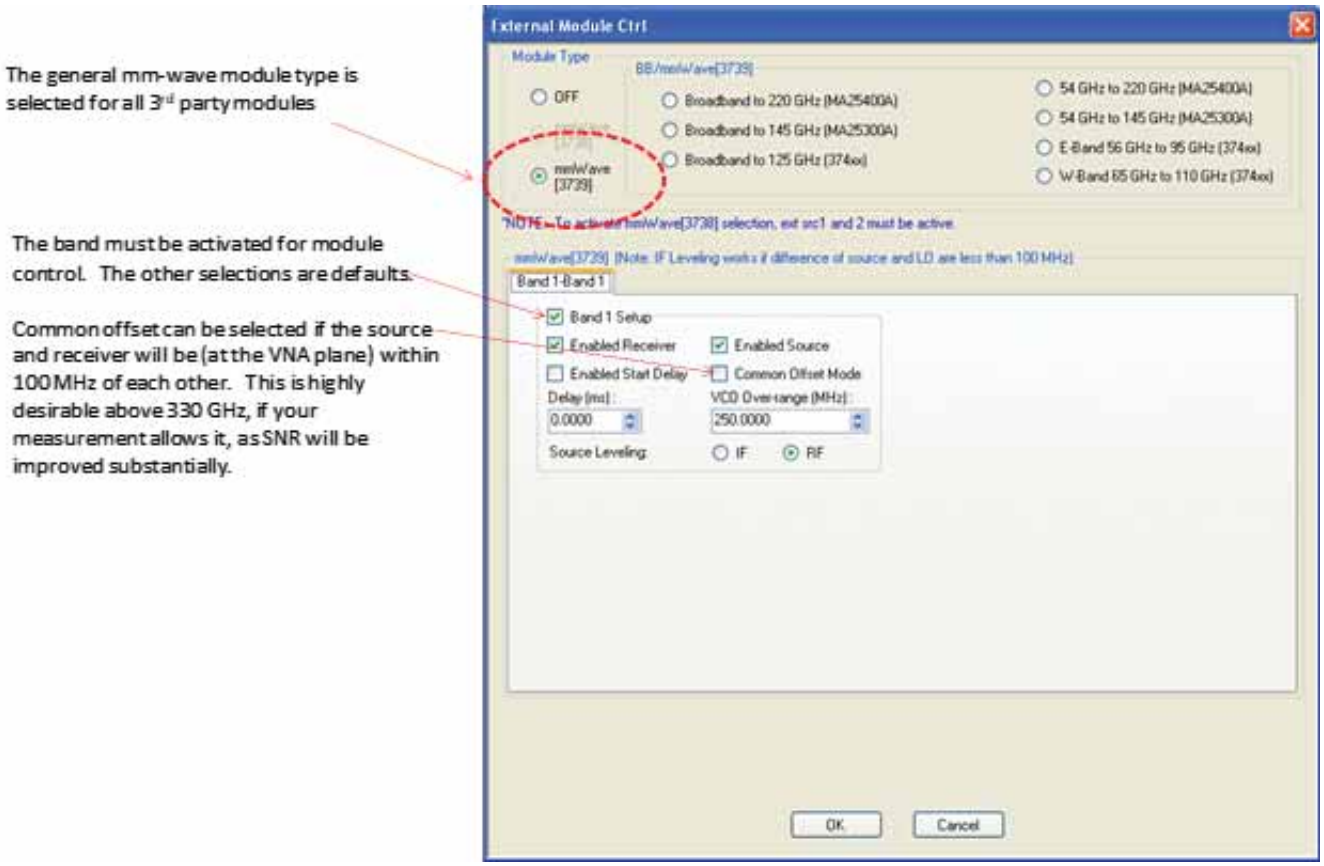


Figure 31-14. External Module Control

After the above dialog is complete, the external module control menu button should show the proper mode: mmWave[3739].



Figure 31-15. Select Done Editing

On the multiple source menu, don't forget to select 'Done Editing' when the configuration is complete.

Because of the large multiplication factors that may be involved, the spectral width of the source signal (if used) may be wide. Working with the RBW setting may be useful to get more resolvable spectral lines. Too narrow and power will be excluded. Too wide and neighboring measurement points can obscure the peak. Values in the 100s of Hz to 10s of kHz are commonly used.

With wide RBWs and high multiplication factors, the image rejection algorithm can sometimes get confused. This largely happens because the desired tone responses can fluctuate wildly from point to point with an overly-wide RBW so the identification of the 'real' tones becomes complicated. A narrower RBW can help or image rejection can be disabled.

31-7 Tips and Hints

- User power calibrations can be used to improve the accuracy of receiver calibrations as in the other applications on the instrument.
- The factory receiver calibration is referenced to the test port. User receiver calibrations can, as usual, be used to shift the absolute power reference to a different plane. Factory receiver calibrations only exist for the base two-port VNA and not for mm-wave configurations or four-port systems.
- Reference plane extensions can be applied (particularly the loss functions are useful).
- Any graph type and any response parameter can be used. The traditional S-parameters and complex graph types are most useful when a tracking measurement is being performed but they can also be helpful for multiple receiver ratios derived from a common external source (e.g., multiple antenna testing).
- Spur reduction cannot avoid all spurs, particularly when using the internal sources.
- The VNA-like point-based sweep mode can be very fast (with very good noise floor) for harmonic, IMD and other measurements where the tones of interest have known locations. Using segmented sweep can make this even faster by skipping more frequencies that are not of interest.

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