White Paper

Measuring Channel Operating Margin

1. Introduction to Channel Operating Margin

Channel Operating Margin, COM, has emerged at high data rates as a single measure of channel performance that includes the effects of both signal impairments and the techniques used to compensate for those impairments.

COM measures the performance margin of an interconnect, but it can also be used to examine the interoperability margin of a high speed serial system. Since COM includes the calculation of the ISI (inter-symbol interference) that remains after equalization and the effects of noise, jitter, and crosstalk, the derivation of COM itself can offer insights into the strengths and weaknesses of a design.

COM was developed for channel characterization in 100, 200, and 400 Gigabit Ethernet (GbE) standards: IEEE 802.3bj and IEEE 802.3bs. It's specified for both NRZ (non-return to zero) and PAM4 (4-level pulse amplitude modulation) standards and is derived nearly the same way in both cases.

In the next section we'll see how COM evolved from channel requirements at lower data rates and how it spans a multi-dimensional space of design variables, a design space. In Section 3 we work through the details of how COM is calculated. We'll see that COM is a difficult measurement whose accuracy depends on the accuracy of all the measurements and models that go into it. Due to its complexity, comparing independent COM measurements of a given interconnect can be challenging; in Section 4 we'll discuss ways to determine if disparate measurements are inconsistent and how to determine which measurement is right. We conclude in Section 5 with a discussion of how COM is likely to evolve and the key equipment features necessary for accurate and reproducible COM measurements.

2. Channel Operating Margin and Design Space

The primary design issues for serial data technology as it approached gigabit data rates (Gb/s) centered on impedance matching and timing problem concerning the trace length synchronization needed to meet setup and hold requirements

At several Gb/s rates, we could evaluate channel performance by measuring eye closure in terms of jitter and the ISI (inter-symbol interference) caused by the frequency dependence of the channel insertion loss. Except for rare extraordinary problems, differential signaling adequately suppressed crosstalk.

As we approached 10 Gb/s, simple equalization techniques like transmitter pre- and de-emphasis along with simple receiver DFE (decision feedback equalization) were needed to open eye diagrams.

As we passed 10 Gb/s, channel loss and reflections started to become serious problems so channel performance was specified by IL(*f*) (insertion loss or SD2D1), RL(*f*) (return loss or SD1D1), and ILD(*f*) (insertion loss deviation) masks, like those in Figure 1. At this point, crosstalk started to emerge as a problem. Standards addressed it in terms of ICR(*f*) (insertion loss to crosstalk ratio) masks and ICN vs IL (integrated crosstalk noise) compliance template, Figure 2. Instead of putting strict limits on crosstalk, ICR(*f*) masks allow designers to balance crosstalk and loss impairments. In addition, the ICN compliance template balanced crosstalk against insertion loss at the Nyquist frequency.

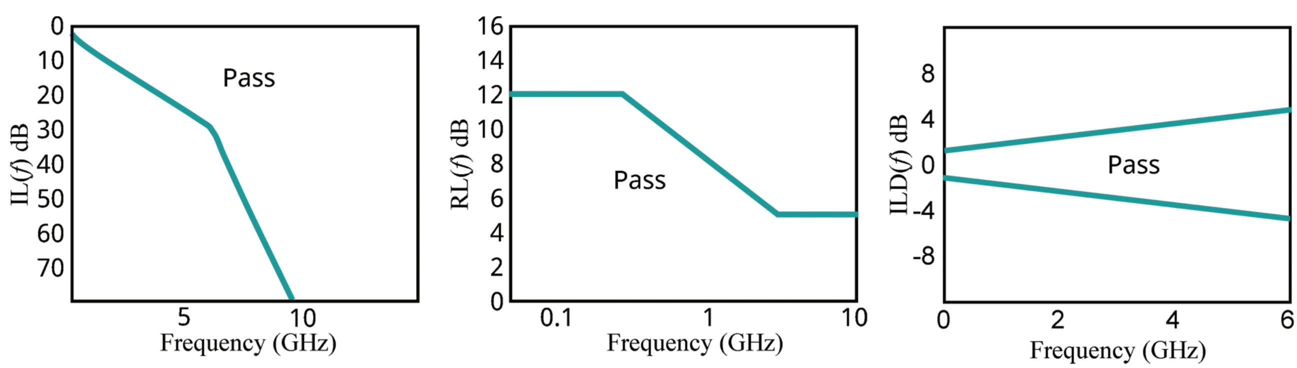

Figure 1: S-parameter masks: IL(*f*), RL(*f*), and ILD(*f*).

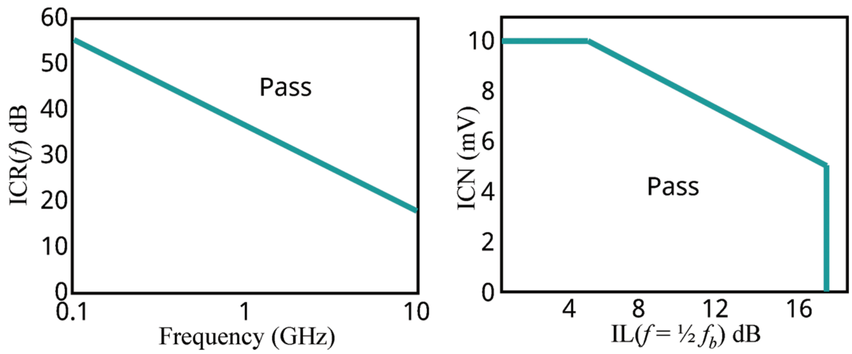

Figure 2: Insertion loss-Crosstalk masks: ICF(*f*) and ICN vs IL(*f* = ½*f*b).

At data rates of 25 Gb/s and higher, eye diagrams tend to be closed by ISI and increasing crosstalk.

COM was introduced at the 25 Gb/s channel rates of 100 GbE and will be included in the emerging 53 Gb/s channel rates of 400 GbE standards for both NRZ and PAM4 signaling. By prescribing minimum values for COM, the standards allow designers to choose how they optimize signal impairments and equalization schemes while meeting BER (bit error rate) specifications.

Since COM is a function of many signal integrity variables, it introduces the concept of a "design space." The best design should correspond to the maximum value of COM in this multidimensional design space.

Figure 3 gives the concept of COM. The eye diagram used to measure COM is calculated from measurements of S-parameters, jitter, and noise along with models of SERDES (serializer-deserializer) response. The calculated eye diagram includes the insertion loss and return loss of the channel, crosstalk, and random jitter and noise with equalization schemes at both the transmitter and receiver. COM is given by the ratio of the signal amplitude, A_{signal}, to the vertical eye closure, $A_{NoiseXtalk}$, defined with respect to the system error rate,

$$\mathrm{COM} = 20\log\frac{A_{Signal}}{A_{NoiseXtalk}} \tag{1}$$

Most standards require

$$\mathrm{COM} > 3\ \mathrm{dB}.$$

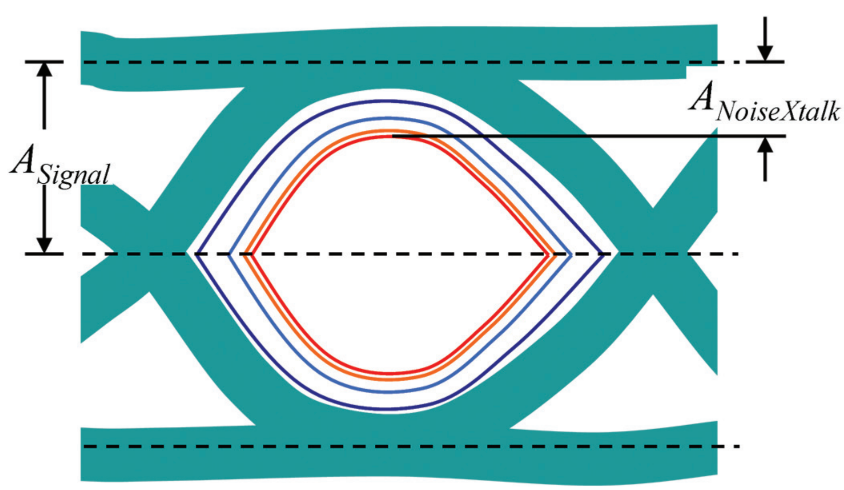

Figure 3: Graphical depiction of COM. The eye diagram and BER contours include the effects of both transmitter and receiver equalization.

3. Deriving COM from S-parameters

The derivation of COM involves analysis of every aspect of a high speed serial data system.

Table 1 shows the many parameters that go into the derivation of COM required for an NRZ implementation of IEEE 802.3bj 100 GbE and a PAM4 implementation of IEEE 802.3bs PAM4.

	Symbol	NRZ Value	PAM4 Value
Standard		100 GbE	400 GbE
Number of signal levels	L	2	4
Signaling rate	f_b	25.78125 Gb/s	25.78125 GBd
Max Frequency step	Δf	0.01 GHz	
Transmitter differential peak output Victim Far-end aggressor Near-end aggressor	 A_V A_{FEXT} A_{FEXT}	 0.4 V 0.4 V 0.6 V	
Package model Single-ended device capacitance Single-ended package-to-board capacitance	 C_d C_p	 0.25 pF 0.18 pF	
Single-ended reference resistance	R_0	50 Ω	
Single-ended termination resistance	R_d	55 Ω	
Receiver 3 dB bandwidth	fr	$0.75 \times f_b$	
Transmitter equalizer min cursor tap	$C(0)$	0.62	0.60
Transmitter equalizer, pre-cursor tap Minimum Maximum Step size	$C(-1)$	 –0.18 0 0.02	 –0.15 0 0.05
Transmitter equalizer, post-cursor tap Minimum Maximum Step size	$C(1)$	 –0.38 0 0.02	 –25 0 0.05
CTLE DC gain Minimum Maximum Step size	g_{DC}	 –12 dB 0 dB 1 dB	 –15 dB 0 dB 1 dB
CTLE zero frequency	f_z	$f_b/4$	
CTLE pole frequencies	f_{p1} f_{p2}	$f_b/4$ f_b	
Transmitter differential peak voltage Victim FEXT aggressor NEXT aggressor	 A_v A_{Fext} A_{Next}	 0.4 V 0.4 V 0.6 V	
Level separation mismatch ratio	R_{LM}	1	0.92
Transmitter SNR	SNR_{Tx}	27 dB	31 dB
Number of samples per UI	M	32	
DFE length	N_{DFE}	14 UI	16 UI
DFE tap limit for $n = 1$ for $n = 2$ to N_{DFE}	$b_{max}(n)$	 1 1	 1 0.2
RMS RJ	σ_{RJ}	0.01 UI	0.015 UI
Dual-Dirac peak-peak jitter	A_{dd}	0.05 UI	0.025 UI
One-sided spectral noise density	η_0	5.2×10^{-8} V²/GHz	
Target detector error rate	DER_0	10^{-5}	3×10^{-4}

Table 1: COM settings from 802.3bj, 100GBASE-KR4 (NRZ) and 100GBASE-KP4 (PAM4)—some of these values may be preliminary so be sure to check the standards document before making any compliance measurement.

COM cannot be measured directly. It is derived from a combination of measurements and assumptions.

The transfer functions of the channel, its crosstalk aggressors, the transmitter output, and receiver input are calculated under the assumption that they are LTI (linear time invariant). The system transfer function is then used to calculate the response of a single pulse, the so-called single bit response, SBR(t). In NRZ terms, think of a pulse as a single logic 1 in a long string of 0s; in PAM4, it's a single S3 in a long string of S0s.

The equalized SBR and all the signal impairments are used to calculate the vertical slice of the eye diagram centered at the time-delay sampling point where the DER (detector error rate) is a minimum. DER is a generalized term that is equivalent to the BER for NRZ systems and the SER (symbol error rate) for PAM4 systems.

The peak signal amplitude, Asignal, is the signal level and the noise-crosstalk amplitude, $A_{noise\text{-}Xtalk}$, is the vertical eye closure defined with respect to the DER prescribed by the standard as in Figure 3 and Table 1.

As we dive into the COM derivation, pay close attention to the different measured quantities and assumptions that go into it—especially how measurement uncertainties might propagate into the final result.

Let's derive COM for the differential channel with input labeled port 1 and output labeled port 2 in the four channel 200 GbE system depicted in Figure 4.

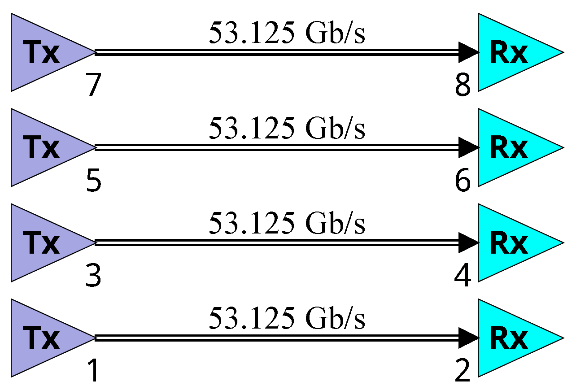

Figure 4: A four channel 200 GbE system.

3.1 Calculate the System Transfer Function

The transfer function of a circuit element represents its loss properties and its frequency and phase response. It is also the frequency domain representation of the impulse response. We build the system transfer function, *H(f)*, from the transfer functions of each element from the transmitter through the channel to the receiver, including transmitter FFE (feed-forward equalization) and receiver CTLE (continuous time linear equalizer) but not (yet) receiver DFE (decision feedback equalizer), Figure 5. The system transfer function is given by the product of the individual transfer functions.

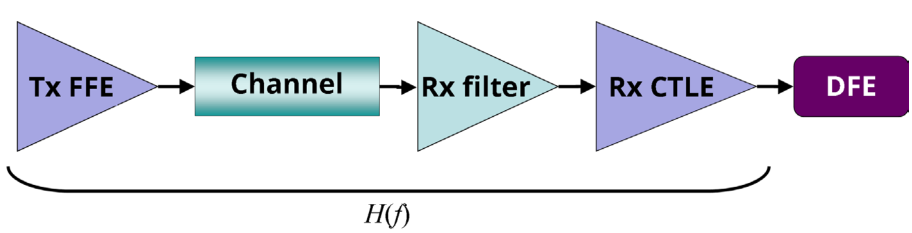

Figure 5: Block diagram of the system elements.

The transmitter output transfer function includes two terms: First, the package response, $H_{Tx}(f)$, comes from either a minimal model specified by the standard or the output response of the transmitter you're considering and, second, the transfer function of the transmitter FFE scheme, $H_{TxFFE}(f)$. The FFE taps are free parameters whose values will be optimized in the next step, Section 3.2.

The transfer function of the channel, $H_{channel}(f)$, is derived from the S-parameters of everything from the transmitter to the receiver. Since current COM implementations ignore common noise and differential-common node conversion, we only need the differential terms: the through term, SD2D1, the reflection term, SD1D1, the FEXT terms, SD2D3, SD2D5, SD2D7, and the NEXT terms, SD2D4, SD2D6, and SD2D8.

The receiver transfer function includes the receiver termination impedance, $H_{Rx}(f)$, usually specified as a 4[th] order Bessel-Thompson filter with bandwidth at $0.75f_b$, where fb is the baud rate, and the CTLE response, $H_{RxCTLE}(f)$. The CTLE gain is a free parameter whose value will be optimized in the next step. Since the receiver DFE is a nonlinear device (i.e., it is not LTI) its effects are included separately.

The resulting transfer function,

$$H(f) = H_{Tx}(f) \times H_{TxFFE}(f) \times H_{channel}(f) \times H_{Rx}(f) \times H_{RxCTLE}(f),$$

is a function of frequency that depends on the still-to-be-determined values of the FFE taps and CTLE gain, Figure 6.

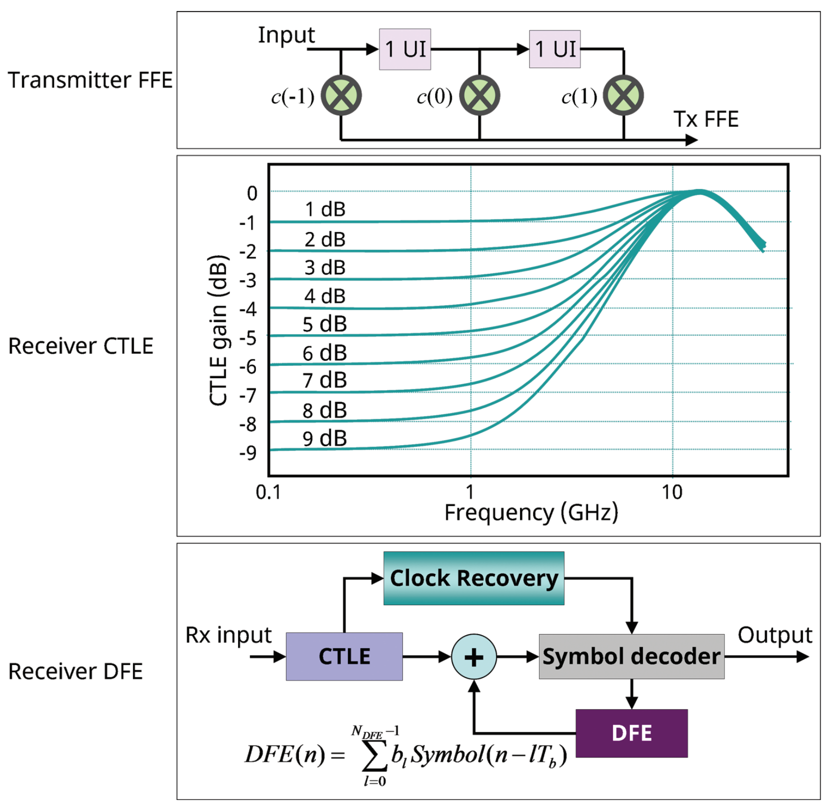

Figure 6: Diagrams of FFE, CTLE, and DFE.

3.2 Optimize Transmitter FFE and Receiver CTLE

The next step is to optimize the equalization parameters. Think of the equalization parameters geometrically as though they span a configuration space. For example, if we have three transmitter FFE taps {c (–1), c (0), c(1)}, one CTLE gain g_{DC}, and 14 DFE taps {b(0), b(1), b(2), ..., b(13)}, then it's an 18 dimensional space. To find the best combination of the 18 parameters we need a function of those 18 parameters that reaches a maximum when the equalization schemes are optimized. That function is a figure of merit: FoM(c (–1), c (0), c(1), gDC, b(0), b(2), ..., b(13)). To find the optimal set of parameters we can either perform a grid search that steps through their possible values or use an optimization algorithm. There are many optimizers available; several are included in MATLAB.

The FoM specified by current COM implementations is like an incomplete, intermediate value of COM focused on ISI

$$\mathrm{FoM} = 10\log\frac{A_{Signal}^2}{\sigma_{Tx}^2 + \sigma_{ISI}^2 + \sigma_J^2 + \sigma_{Xtalk}^2 + \sigma_{SpecNoise}^2} \tag{2}$$

where A_{signal} is the peak signal amplitude and the denominator includes five terms: transmitter noise, σ_{Tx}, residual ISI (i.e., the ISI left over after equalization), σ_{ISI}, the jitter contribution to amplitude noise, σ_J, peak crosstalk, σ_{Xtalk}, and the spectral noise at the output of the CTLE (i.e., the input to the DFE), $\sigma_{SpecNoise}$. The way that the noise terms appear in Eq. (2) indicates that we're assuming all of these noise sources are characterized by their standard deviations and that they all follow Gaussian distributions—an assumption made to simplify the calculation.

We calculate FoM from the single bit response, SBR(t). SBR(t), is given by the convolution of an ideal pulse and the impulse response, Figure 7:

$$\mathrm{SBR}(t) = \int pulse(t')h(t-t')dt'$$

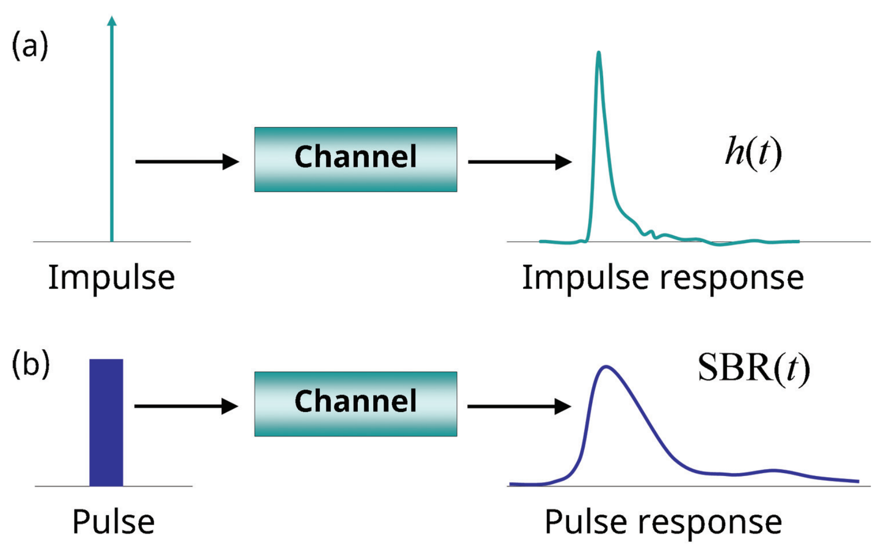

Figure 7: (a) Impulse response, $h(t)$, and (b) SBR(t) (single bit response).

The impulse response, *h(t)*, is the time domain version of the transfer function. Since *H(f)* and, hence *h(t)* are discrete functions, the transformation from the frequency to the time domain is an IFFT (inverse fast Fourier transform) and the Bessel-Thompson filter at the receiver serves as an anti-aliasing filter,

$$\text{Impulse response} = h(t_n) = \text{IFFT}[H(f)]$$

and so we get

$$\text{SBR}(t_n) = \sum_p pulse(t_p)h(t_n - t_p)$$

where h is sampled M times per UI (unit interval) and the sum is carried out over every sample and UI; Table 1 indicates that a typical value is $M = 32$.

We define the sampling point, tsp, one UI after the initial rise of SBR(t), Figure 8.

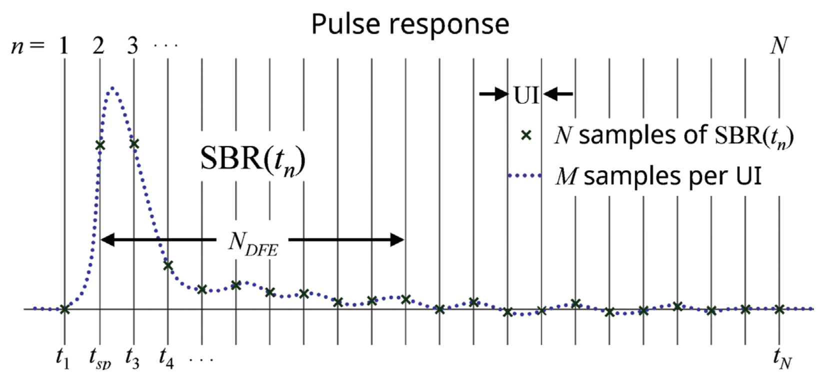

Figure 8: SBR(t) sampled M times per UI.

The signal amplitude is the amplitude of the pulse response at the sampling point:

$$A_{signal} = \text{SBR}(t_{sp}) \qquad \text{for NRZ and}$$
$$A_{signal} = 1/3\ \text{R}_{\text{LM}} \times \text{SBR}(t_{sp}) \quad \text{for PAM4}$$

The factor of 1/3 for PAM4 accounts for the inter-symbol voltage swing of the four voltage levels and R_{LM} is the level separation mismatch ratio.

The transmitter noise term, σ_{Tx}, in Eq. (2) is given by the product of the signal amplitude and the reciprocal of the signal-to-noise-ratio, that is, the product of the signal amplitude and the ratio of the noise to the signal:

$$\sigma_{Tx} = \frac{SBR(t_{sp})}{SNR}$$

where SNR is expressed as a simple fraction, not in dB.

To get the residual ISI term, σ_{ISI}, in Eq. (2) we need to include receiver DFE. The differences between SBR(t_n), which already includes transmitter FFE and receiver CTLE, and the DFE corrections:

$$\Delta_{ISI}(n) = \begin{cases} 0 \quad n = 0 \\ SBR(t_{sp} + nT_b) - b(n) \times SBR(t_{sp}) & 1 \le n \le N_{DFE} \\ SBR(t_{sp} + nT_b) & N_{DFE} < n \end{cases}$$

The first term indicates that SBR and ideal pulse have the same amplitude at the sampling point. If equalization were perfect, there would be no ISI and all the terms would be zero. Notice how DFE is included explicitly through its taps, $b(n)$, of which there are N_{DFE}. The residual ISI, at least for calculating the FoM, is given by the sum of the squares of the differences:

$$\sigma_{ISI}^2 = \sum_{n=1}^{N} \Delta_{ISI}^2(n) \quad \text{for NRZ and}$$

$$\sigma_{ISI}^2 = \tfrac{5}{9} \sum_{n=1}^{N} \Delta_{ISI}^2(n) \quad \text{for PAM4}$$

where the sum is over the N UI extent of SBR(t_n) as shown in Figure 8 and the factor of 5/9 for PAM4 accounts for the density of possible PAM4 symbol transitions.

The jitter-to-amplitude noise conversion, σ_J, in Eq. (2) is given by the sum of the squares of the jitter-to-noise conversions at every unit interval of the SBR:

$$\sigma_J^2 = \sum_{n=1}^{N} (A_{DD}^2 + \sigma_{RJ}^2)\mu_n^2 \quad \text{for NRZ and}$$

$$\sigma_J^2 = \tfrac{5}{9} \sum_{n=1}^{N} (A_{DD}^2 + \sigma_{RJ}^2)\mu_n^2 \quad \text{for PAM4}$$

where A_{DD} is the peak-to-peak dual-Dirac model DJ (deterministic jitter), σ_{RJ} is the RMS RJ (random jitter), and μ_n is the slope of SBR(t_n) at the center of the nth UI. The sum is over the extent of SBR(t_n) as shown in Figure 8. It's worth pointing out that A_{DD} and σ_{RJ} are combined in an unconventional way purely to simplify calculation of FoM. Jitter is handled more accurately in the final COM calculation.

The peak crosstalk, σ_{Xtalk}, in Eq. (2) is given by the sum of the maximum levels of crosstalk in each unit interval of SBR(t_n) for every aggressor; for the kth aggressor, the maximum crosstalk amplitude is $A_{k\text{-}max\ Xtalk}(n)$. The peak crosstalk is the sum of the squares

$$\sigma_{Xtalk}^2 = \sum_{k}^{K-1} \sum_{n}^{N} A_{k-\max Xtalk}^2(n) \quad \text{for NRZ and}$$

$$\sigma_{Xtalk}^2 = \tfrac{5}{9} \sum_{k}^{K-1} \sum_{n}^{N} A_{k-\max Xtalk}^2(n) \quad \text{for PAM4}$$

where the sum includes the contributions of all K-1 aggressors and N unit intervals of SBR(t_n).

Finally, the RMS spectral noise at the output of the CTLE (i.e., the input to the DFE), $\sigma_{SpecNoise}$, in Eq(2) is given by

$$\sigma_{SpecNoise}^2 = \eta_0 \int \left| H_{Rx}(f) H_{RxCTLE}(f) \right|^2 df$$

where η_0 is the one-sided spectral density (in V^2/GHz), $H_{Rx}(f)$ is the receiver input filter transfer function, and $H_{RxCTLE}(f)$ is the CTLE transfer function.

The FoM we've derived is now a function of the FFE taps, the CTLE gain, and the DFE taps. The optimal set of equalization parameters is obtained by finding the maximum value of FoM either through a grid search or by using an optimization algorithm.

We can now calculate the signal amplitude, A_{signal}, from SBR(t_n) with optimized equalization:

$$A_{signal} = \mathrm{SBR}_{opt}(t_{sp}) \quad \text{for NRZ and}$$

$$A_{signal} = 1/3\ R_{LM} \times SBR_{opt}(t_{sp}) \quad \text{for PAM4}$$

3.3 Calculate the Noise and Crosstalk Interference

With the equalization scheme set and the numerator, A_{signal}, in Eq. (1) determined, we turn to the noise and interference-crosstalk terms to get $A_{NoiseXtalk}$ and then complete the calculation of COM.

We acquire the noise and crosstalk PDF (probability density function) by modeling a vertical slice of the eye diagram.

Eye diagrams consist of overlaid waveforms of individual symbols. The key ingredients to each symbol-waveform can be separated into deterministic and random components. The deterministic components consist of discrete shifts in the level of each symbol-waveform and the random components cause statistical fluctuations about those levels.

We configure a vertical slice of the eye by assembling the deterministic variations of SBR(t_n) into a PDF and convolve it with a smearing PDF built from the random components.

The discrete levels are determined by every combination of:

- The 2 possible symbols, (–1, +1), for NRZ or the 4 possible symbols, (–1, –1/3, +1/3, +1), for PAM4
- The ISI variations of each symbol level, given by the N values of $\Delta_{ISI}(n)$
- The DJ component of jitter-to-amplitude noise conversion, given by the N values of $A_{DD}\mu_n$.
- The crosstalk variations of each symbol level, given by the set of $N\times(K-1)$ values of $A_{k\text{-}maxXtalk}(n)$: the peak crosstalk at every UI for each of the K–1 aggressors. Notice that this assumes that the largest crosstalk impairments are always phase-aligned with the sampling point.

These deterministic contributions combine into a PDF of discrete levels $\{y_q\}$. If we let L= the number of symbol levels, 2 for NRZ and 4 for PAM4, then $\{y_q\}$ consists of the set of $LN^3(K-1)$ combinations of deterministic effects, just as they would in an eye diagram. The resulting PDF of discrete levels is

$$\mathrm{PDF}_{levels}(y) = \frac{1}{LN^3(K-1)} \sum_{q=1}^{LN^3(K-1)} \delta\left(y - y_q\right)$$

Random noise causes each level to fluctuate according to the combined PDF of the noise sources. Each noise source follows a Gaussian distribution described by it’s own RMS voltage. Since the convolution of a set of Gaussians is also a Gaussian, the PDF for the random components is given by

$$\mathrm{PDF}_{Gaussian}(y) = \frac{1}{\sqrt{2\pi\sigma_G^2}} \exp\left(-\frac{y^2}{2\sigma_G^2}\right)$$

where

$$\sigma_G^2 = \sigma_{Tx}^2 + \sum_{n=1}^{N} \sigma_{RJ}^2 \mu_n^2 + \sigma_{SpecNoise}^2$$

Convolving the two gives the COM noise-interference $PDF_{NoiseXtalk}(y)$:

$$\mathrm{PDF}_{Noise-Xtalk}(y) = \frac{1}{LN^3(K-1)\sqrt{2\pi\sigma_G^2}} \sum_{q=1}^{LN^3(K-1)} \exp\left(-\frac{\left(y - y_q\right)^2}{2\sigma_G^2}\right)$$

Since COM is defined with respect to a specified detector error rate, DER_0, as in Table 1, we need to convert the PDF into a CDF (cumulative distribution function)

$$\mathrm{CDF}_{NoiseXtalk}(y) = \int_{-\infty}^{y} \mathrm{PDF}_{NoiseXtalk}(y')\,dy'$$

We calculate $A_{NoiseXtalk}$ by solving $\mathrm{CDF}_{NoiseXtalk}(A_{NoiseXtalk}) = \mathrm{DER}_0$ for $A_{NoiseXtalk}$. And, finally, we get

$$\mathrm{COM} = 20\log\frac{A_{Signal}}{A_{NoiseXtalk}}$$

4. How to compare separate COM measurements

With so many steps in the process, it's difficult to estimate how the uncertainties of the measured inputs propagate to the uncertainty of COM. When we're confronted with COM measurements of a given system that are performed by different people who have used different equipment, it can be difficult to determine whether the two values are consistent or which is more accurate.

The measured quantities that go into COM include:

- Channel S-parameters, including crosstalk terms
- Transmitter voltage noise, σ_{Tx}
- RMS RJ, σ_{RJ}, and dual-Dirac DJ, A_{DD}
- The one-sided spectral density, η_0

If the measurements have significant disagreement, then compare the measured inputs. Pay close attention to the S-parameters because they have a large impact on every step of the COM derivation except the spectral noise. By measuring S-parameters with a high performance VNA (vector network analyzer) like an Anritsu VectorStar® or Shockline™ model, you can be confident that your S-parameters won't introduce large uncertainties.

The only way to be certain that two measurements disagree is to calculate their total uncertainties. With the COM calculation set up, the measurement uncertainties can be propagated through the COM calculation by repeating the calculation many times each with the measured quantities varied within the limits of their individual uncertainties. Each calculation yields a different value of COM and the total uncertainty can be calculated from the resulting distribution.

If the measured inputs are consistent but the COM results are not, then there must be a difference in implementation. Look for a discrepancy by comparing intermediate results of the calculation like values of the residual ISI, the jitter-to-amplitude noise conversion, and crosstalk.

If the two calculations seem identical and the measured inputs are consistent, then the problem probably comes from the optimization of the equalization parameters. Determining the global maximum in a multi-dimensional space is a notoriously difficult, long-standing problem in computer science.

5. The Utility of COM and Equipment Considerations

Once the tools are in place, calculating COM should be fast and efficient compared to a simulation.

The purpose of COM is to characterize a channel in a system with a minimally-specified SERDES, but COM can also help determine the functionality of a SERDES with different channels. You can specify the transmitter and receiver models however you like. COM can help determine the best equalization schemes: the number of FFE or DFE taps and whether you need both transmitter FFE and receiver CTLE. You can estimate the interoperability of different parts, but be careful, COM is not a replacement for IBIS-AMI (Input/output Buffer Information Specification-Algorithmic Modeling Interface) models; it cannot verify a design the way that a simulation can.

The derivation of COM provides diagnostic information through intermediate quantities used in its calculation: residual ISI and the interplay of specific transmitter FFE, receiver CTLE and DFE equalization techniques, the impact of spectral noise inside the receiver, and, of course, the worst-case impact of crosstalk.

Our derivation of COM followed the techniques specified by 100 GbE IEEE802.3bj and 400 GbE IEEE802.3bs (which is still preliminary as this paper goes to press). The derivation made several expedient approximations. Future implementations of COM will modify these techniques. By aligning the phases of the peaks of all crosstalk signals, COM assumes worst-case crosstalk in every instance. Since the equalization scheme was optimized with a FoM that assumes ISI, DJ, and Xtalk were all Gaussian, it's possible that a more effective equalization scheme could exist that would raise the value of COM. In addition to correcting some of these approximations, future COM specifications are likely to include more complete package models, more accurate jitter effects including the effect of clock recovery, common mode noise, differential-to-common conversion, and common mode rejection.

5.1 Necessary Equipment for accurate and reproducible COM measurements

At data rates of 25+ Gbaud, the accuracy of multiport S-parameter measurements takes on terrific importance. Since COM involves so many intermediate calculations, even small S-parameter inaccuracies can lead to large COM discrepancies and days wasted trying to find the problem.

S-parameters estimated from TDT/TDR measurements lack the bandwidth, dynamic range and noise floor necessary to calculate reliable values of COM. COM measurements require noise floors of –90 dBm, dynamic ranges over 90 dB, and ample bandwidth: for NRZ at 28 Gb/s signaling, 42 GHz to cover the 3rd harmonic and 70 GHz for the 5th harmonic and, at 56 Gb/s, 84 GHz for the 3rd harmonic and 140 GHz for the 5th. For PAM4 signaling, 35 GHz at 14 Gbaud and 70 GHz at 28 Gbaud to reach the 5th harmonics.

As we've seen, comparing two inconsistent COM values is a difficult, time-consuming exercise whose end result merely indicates whether or not the calculation was done correctly the first time. The easiest way to get it right is to base your calculations on measurements that you can trust.

5.2 Selecting a VNA for COM measurements

In selecting a VNA there are several important parameters. Anritsu Company has a complete line of VNAs from the ultra-high performance 4-port VectorStar VNAs with bandwidths up to 145 GHz to more economical ShockLine VNAs.

Table 2 shows typical performance abilities for two models. We already mentioned the bandwidth, dynamic range, and noise floor requirements. Since VNAs measure the frequency response one frequency at a time in steps from the low frequency limit up to the high frequency limit. The low frequency limit is of terrific importance in assuring that frequency domain measurements can be accurately transformed to the time domain. Since VNA S-parameters are sets of discrete measurements, it's also important to keep the frequency step size as small as possible; which corresponds to a large number of measurement points.

	VECTORSTAR ME7838D	**SHOCKLINE MS46524B**
Bandwidth	145 GHz	43.5 GHz
Low frequency limit	70 kHz	50 kHz
Typical Dynamic Range	110 dB	125 dB
Typical Noise Floor	–100 dBm	–115 dBm
Number of measurement points	100,000	20,000

Table 2: Features of a VectorStar and a ShockLine configuration that are important for PAM4 signaling.

The combination of unique low-frequency coverage with up to 100,000 measurement points make VectorStar VNAs all but immune to aliasing and causality problems. The more economical ShockLine VNAs also have excellent low frequency coverage and up to 20,000 measurement points.

Precise S-parameter measurements also requires accurate VNA calibration and the ability to de-embed the test fixture. Anritsu VNAs offer the industry's best calibration and measurement stability and come with advanced de-embedding network extraction software.

VectorStar VNAs achieve the most accurate time domain measurements available—and COM is ultimately a time domain quantity. COM is a complicated, difficult measurement. Don't let inaccurate measurements waste your time.

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Specifications are subject to change without notice.

• United States
Anritsu Company
1155 East Collins Boulevard, Suite 100,
Richardson, TX, 75081 U.S.A.
Toll Free: 1-800-267-4878
Phone: +1-972-644-1777
Fax: +1-972-671-1877

• Canada
Anritsu Electronics Ltd.
700 Silver Seven Road, Suite 120,
Kanata, Ontario K2V 1C3, Canada
Phone: +1-613-591-2003
Fax: +1-613-591-1006

• Brazil
Anritsu Electrônica Ltda.
Praça Amadeu Amaral, 27 - 1 Andar
01327-010 - Bela Vista - Sao Paulo - SP - Brazil
Phone: +55-11-3283-2511
Fax: +55-11-3288-6940

• Mexico
Anritsu Company, S.A. de C.V.
Av. Ejército Nacional No. 579 Piso 9, Col. Granada
11520 México, D.F., México
Phone: +52-55-1101-2370
Fax: +52-55-5254-3147

• United Kingdom
Anritsu EMEA Ltd.
200 Capability Green, Luton, Bedfordshire LU1 3LU, U.K.
Phone: +44-1582-433280
Fax: +44-1582-731303

• France
Anritsu S.A.
12 avenue du Québec, Batiment Iris 1-Silic 612,
91140 Villebon-sur-Yvette, France
Phone: +33-1-60-92-15-50
Fax: +33-1-64-46-10-65

• Germany
Anritsu GmbH
Nemetschek Haus, Konrad-Zuse-Platz 1
81829 München, Germany
Phone: +49-89-442308-0
Fax: +49-89-442308-55

• Italy
Anritsu S.r.l.
Via Elio Vittorini 129, 00144 Roma Italy
Phone: +39-06-509-9711
Fax: +39-06-502-2425

• Sweden
Anritsu AB
Kistagången 20B, 164 40 KISTA, Sweden
Phone: +46-8-534-707-00
Fax: +46-8-534-707-30

• Finland
Anritsu AB
Teknobulevardi 3-5, FI-01530 VANTAA, Finland
Phone: +358-20-741-8100
Fax: +358-20-741-8111

• Denmark
Anritsu A/S
Kay Fiskers Plads 9, 2300 Copenhagen S, Denmark
Phone: +45-7211-2200
Fax: +45-7211-2210

• Russia
Anritsu EMEA Ltd.
Representation Office in Russia
Tverskaya str. 16/2, bld. 1, 7th floor.
Moscow, 125009, Russia
Phone: +7-495-363-1694
Fax: +7-495-935-8962

• Spain
Anritsu EMEA Ltd.
Representation Office in Spain
Edificio Cuzco IV, Po. de la Castellana, 141, Pta. 5
28046, Madrid, Spain
Phone: +34-915-726-761
Fax: +34-915-726-621

• United Arab Emirates
Anritsu EMEA Ltd.
Dubai Liaison Office
P O Box 500413 - Dubai Internet City
Al Thuraya Building, Tower 1, Suite 701, 7th floor
Dubai, United Arab Emirates
Phone: +971-4-3670352
Fax: +971-4-3688460

• India
Anritsu India Pvt Ltd.
2nd & 3rd Floor, #837/1, Binnamangla 1st Stage,
Indiranagar, 100ft Road, Bangalore - 560038, India
Phone: +91-80-4058-1300
Fax: +91-80-4058-1301

• Singapore
Anritsu Pte. Ltd.
11 Chang Charn Road, #04-01, Shriro House
Singapore 159640
Phone: +65-6282-2400
Fax: +65-6282-2533

• P. R. China (Shanghai)
Anritsu (China) Co., Ltd.
27th Floor, Tower A,
New Caohejing International Business Center
No. 391 Gui Ping Road Shanghai, Xu Hui Di District,
Shanghai 200233, P.R. China
Phone: +86-21-6237-0898
Fax: +86-21-6237-0899

• P. R. China (Hong Kong)
Anritsu Company Ltd.
Unit 1006-7, 10/F., Greenfield Tower, Concordia Plaza,
No. 1 Science Museum Road, Tsim Sha Tsui East,
Kowloon, Hong Kong, P. R. China
Phone: +852-2301-4980
Fax: +852-2301-3545

• Japan
Anritsu Corporation
8-5, Tamura-cho, Atsugi-shi,
Kanagawa, 243-0016 Japan
Phone: +81-46-296-6509
Fax: +81-46-225-8359

• Korea
Anritsu Corporation, Ltd.
5FL, 235 Pangyoyeok-ro, Bundang-gu, Seongnam-si,
Gyeonggi-do, 13494 Korea
Phone: +82-31-696-7750
Fax: +82-31-696-7751

• Australia
Anritsu Pty Ltd.
Unit 20, 21-35 Ricketts Road,
Mount Waverley, Victoria 3149, Australia
Phone: +61-3-9558-8177
Fax: +61-3-9558-8255

• Taiwan
Anritsu Company Inc.
7F, No. 316, Sec. 1, Neihu Rd., Taipei 114, Taiwan
Phone: +886-2-8751-1816
Fax: +886-2-8751-1817

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