

Advancing beyond **Anritsu**

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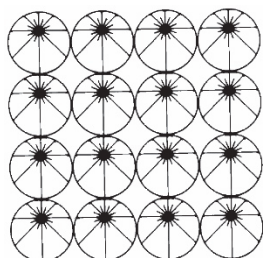
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Foreword



A Technology Compass for the Next 100 Years

Director, CTO
Hanako Noda



Anritsu Technical Review (in Japanese*) was first published in October 1957. It was an era when Japan was heading toward a period of full-fledged economic growth with the momentum of post-war recovery in the background. President and Representative Director Masakazu Taomoto at that time contributed the Foreword to the inaugural issue writing passionately, "Electronics is the darling of all industrial technology fields. In our industry, technology is paramount, and again, technology is key. Technology is not built overnight, but is built by layering new techniques on the foundations established by our predecessors." In the inaugural issue, cutting-edge technology at that time was shared widely through various technical papers.

The shape of society and industry has changed dramatically in the more than half a century since then, and communications technology has expanded continuously from analog to digital, to wireless and optical communications, and further to our data society brought about by IoT and AI. And now, Anritsu has transitioned to a communication measurement business, leveraging the rich experience and technology cultivated in the wireless communication equipment business when the journal was first published, providing innovative and high-quality services to the market with cutting-edge measurement technology. Furthermore, Anritsu measurement technologies support society in various other ways, covering a wide range of fields, including food and pharmaceutical inspection, devices supporting communications and communication measurement, and environmental measurement.

*: The English version has been published since 1982.



Business Visions: Beyond testing, beyond limits, for a sustainable future together

Anritsu is operating under these business visions. While mastering current measurement technologies cultivated over many years, we are creating added value beyond the traditional "measurement" and "inspection" frameworks by fusion with other different ideas and technologies. Anritsu is adopting a new approach to enhance people's quality of life by combining high-precision measurement technologies from the communications field with sensing and image/information processing technologies for evaluating food and pharmaceutical safety, as well as integration with various other fields, such as energy management. Acquiring accurate and reliable data, in other words — precise measurement, is especially essential now, as AI applications and big data analysis advance. Anritsu's strengths in accurate measurement help further social developments. Furthermore, combining communications and AI technologies with food and pharmaceutical inspection will help solve social challenges, such as predicting the causes of

defects, reducing food waste, and responding to the declining working population through automation and labor-saving.

In the communications world, rollout of 6G services is coming into view with efforts to develop communication infrastructure that integrates wired and wireless technologies. This fusion of wired and wireless communications aims to build infrastructure that truly meets societal demands, such as low latency, high capacity, and low power consumption. Additionally, AI-based autonomous networks offer opportunities to provide new solutions that we are confident will lead to innovations transcending the limits of current business operations. Furthermore, Anritsu is committed to its mission of assuring a sustainable and attractive future for the next generations of employees, customers, and all stakeholders. While social challenges, such as climate change and energy issues, still remain unsolved, we believe our technologies can help find answers. In a world where advanced technologies such as AI and AGI (Artificial General Intelligence) have the potential to significantly transform society, Anritsu is meeting the challenge of assuring a sustainable society where people and technology are in harmonious collaboration achieved through activities that accurately measure and convert measurement results into valuable information for society. One example is our collaboration with Takasago Manufacturing, which recently joined our group.



Takasago Manufacturing's mission is to contribute to sustainable social prosperity and growth for all by providing energy control solutions. Specifically, the company's strengths are high-capacity energy control to evaluate the performance of key components for next-generation vehicles

(including EVs and PHEVs), such as batteries, inverters, and motors. By combining Takasago Manufacturing's technologies with Anritsu's strengths in communications measurement and inspection technologies, we have dramatically expanded our approaches and capabilities in achieving a future sustainable decarbonized society, as well as contributing to communications as social infrastructure and to food and pharmaceutical safety. By integrating our expertise with partners, we are beginning to develop a more comprehensive solution-oriented business going beyond our traditional "measurement" domain.

Additionally, in order to expand our measurement abilities, our Advanced Technology Research Laboratory has acquired atomic-scale precision processing technology for graphene — an interesting new material. While focusing on its physical properties, we are exploring applications for next-generation devices. Meanwhile, our research on 6G is advancing development of fundamental technologies for handling high frequencies such as the millimeter-wave and sub-terahertz bands. We expect that results in these cutting-edge research fields will serve as an important foundation for innovation in next-generation communications and electronics devices.

Towards Next 100 Years

Since its inception, Anritsu Technical Review has served as a knowledge platform for sharing insights with engineers both inside and outside the company. In the first issue published in 1957, the pages were packed with cutting-edge information of that time, such as HF (high frequency) receivers for coastal stations and direction finders. Looking back, the research achievements of that era became the foundation of our current communication and measurement technologies, establishing a solid first step toward further evolution. The various themes featured in the first issue were all innovations that pioneered that era, and also represented challenges to technologies anticipating the future of society.

This issue marks a significant milestone — the 100th edition of Anritsu Technical Review (in Japanese). At the same time, this is also a milestone year marking Anritsu's 130th anniversary. 6G and IOWN are revolutionizing communications, AGI will bring structural social changes,

while achieving decarbonization and sustainable economies remain serious global challenges. Facing these trends requires greater cooperation not only between engineers but between all stakeholders. Anritsu Technical Review aims to continue being a "compass" that both deepens these cooperative relationships and guides a path to the future.

Even technologies and ideas that seem at first glance to belong to different fields actually have great potential to connect through the common foundation of "measurement." By bringing together communications and inspection, we might dramatically improve food safety. By combining industrial equipment control technology with communications infrastructure, we might build resilient networks that can withstand disasters. We aim to break through barriers and pioneer the next 100 years with an approach that transcends conventional wisdom and frameworks, going beyond all boundaries.

By moving ahead one step at a time without stopping we shall expand our "measurement" technology into even more fields, bringing new value to society and industry. My hope is that this journal will continue sharing the results of meeting these challenges with all of you to serve as a technology and wisdom guide as we head toward the next 100 years.

We look forward to your continuing support and expectations for both Anritsu Technical Review and our Anritsu Group.





Special Articles

- Application of Graphene Towards Thermal Management
- Application of Graphene Towards Hydrogen-Gas Sensor
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- Wideband Millimeter-Wave Complex Permittivity Measurement Technology
- Pushing the limits: The Future of Wideband Vector Signal Generators
- Sub-THz MMIC Switches Key Enablers for Sub-THz Wireless Systems

Application of Graphene Towards Thermal Management

Takamoto Yokosawa, Tomohiro Matsui

[Summary]

A novel thermal properties different from the classical understanding are expected when wave nature of phonon, a quasi-particle of lattice-vibration which is responsible for the heat conduction, becomes significant by miniaturizing devices relative to the mean-free-path of phonons. In this study, unconventional thermal conductivity, which increases with decreasing the nanoribbon width, were discovered by fabricating graphene with long mean-free-paths into nanoscale mesh structures with atomically ordered zigzag edges. The maximum thermal conductivity of the mesh structure is equivalent to or greater than the highest thermal conductivity reported for carbon nanotubes, and is nearly an order of magnitude larger than the heat dissipation sheets characterized by high thermal conductivity. Moreover, the thermal conductivity can be varied by a factor of nearly 8 simply by changing the nanoribbon width. Since these features are obtained regardless of a substrate, this feature could be applied to heat dissipation and rectification sheets.

1 Introduction

We, human beings, have acquired techniques for controlling electricity and handling light. Recently, the importance of technologies for handling and controlling heat, such as predicting temperature, cooling materials, and dissipating heat efficiently, is increasing. The size ranges widely from the μm scale of electronic devices to the cm scale of devices like smartphones, and even to the meter scale of automobiles and factories. For example, in electronic devices, not only the dissipation of heat generated by electronic components, but also the heat-dissipation pathways must be designed carefully to prevent heat adversely affecting adjacent components. In fact, heat generation by electronic devices has become a barrier to integration of components and performance improvement. Such heat control — thermal management — technologies, can be essential for a future IoT and energy-efficient society.

Graphene, which is a one-atom-thick carbon film, is the focus of thermal management research, too, because it is extremely thin, lightweight, transparent, mechanically strong yet flexible, and has excellent electrical and thermal conductivity. As shown in Table 1 listing the thermal conductivity of major materials, graphene has a thermal conductivity of nearly an order of magnitude greater than common metals, such as copper.

As quasi-particles called electrons and photons conduct electricity and light, heat is conducted by quasi-particles

called lattice vibrations (phonons). The average distance these phonons travel without colliding with each other is called the mean-free-path, and the mean-free-path is important for evaluating conduction properties. When the system size is longer than the mean-free-path (or when the mean-free-path is short compared to the system size), the quasi-particles collide with each other before they collide with the system boundary. In this case, the ‘particle’ nature of the quasi-particles becomes dominant, and conduction becomes diffusive. In contrast, in nanoscale devices or at low temperatures where the mean-free-path is long, quasi-particles are scattered at the system boundaries before colliding with each other. In this case, conduction becomes ballistic, and a wave-like nature emerges, in which information intrinsic to quasi-particles, such as the phase, is preserved. Generally, diffusive conduction, expressed by the Boltzmann equation, dominates, and ballistic conduction is only possible in nanostructures at ultra-low temperatures. However, since the mean-free-path of graphene is long due to its high crystallinity, unique thermal conduction properties, that cannot be obtained with ordinary materials, can be achieved even at room temperature.

Table 1 Thermal Conductivity of Major Materials

Material	Thermal Conductivity (W/(m · K))
Graphene	3000 – 5300 ¹²⁾
Carbon nano tubes	3000 – 3500 ^{3), 4)}
Diamond	1000 – 2200 ¹⁾
Graphite	72 – 2100 ⁵⁾
Copper	403 ⁶⁾
Si	168 ⁶⁾
SiO ₂	1.4 ⁶⁾
Water	0.561 ⁶⁾
Air	0.024 ⁶⁾

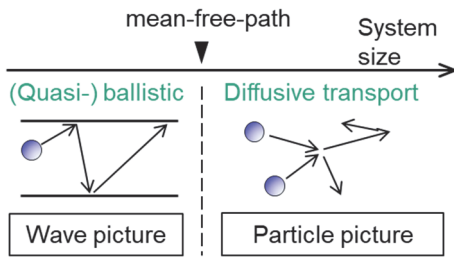


Figure 1 Schematics of ballistic and diffusive transports.

On the other hand, considering the device structure, new functions can be achieved by designing artificial nano- or micro-meter scale structures. It is called “meta-materials”, which enables characteristics that were previously thought impossible with conventional materials, such as negative refractive indices and invisibility cloaks. In such way, novel properties can be expected for graphene by nano-structuring and by arranging nano-structures periodically.

This paper introduces the thermal conductivity of graphene which is nano-fabricated into mesh and ribbon structures using our unique hydrogen plasma etching technique. General understandings of the thermal conductivity were described in section 1. Section 2 introduces previous research on thermal conductivity of microfabricated graphene, and section 3 introduces samples and measurement methods in this study, and discusses the measured results.

2 Thermal Conductivity of Microfabricated Graphene

Thermal conductivity is generally decreased by nano-structuring, since phonon scattering at the edges is increased, and this is also the case for graphene. For example, it is observed for an array of nanoscale graphene ribbons (Graphene Nano Ribbons: GNR) with widths (W) of 45 to 130

nm that the thermal conductivity decreases by decreasing W ⁷⁾. Here, GNR is fabricated by oxygen plasma etching of graphene obtained by exfoliating graphite on SiO₂ substrates (exfoliated graphene). This research measured the thermal conductivity by applying heat from a heater to one end of the GNR and measuring the temperature change at the opposite end. Since the GNRs are supported by SiO₂ substrate in this study, the thermal conductivity is determined by the thermal conduction with and without GNRs.

Smaller thermal conductivity has also been reported⁸⁾ as the neck width becomes narrower in graphene meshes with circular holes, where the average distance between holes (neck width) is about 8 to 21 nm. Here, graphene itself was obtained by chemical vapor deposition (CVD) method. The meshed graphene in this research were fabricated by reactive ion etching (RIE) using a polymer mask, and then transferred to a SiN membrane with holes to prepare a meshed graphene suspended from the substrate. Thermal conductivity was determined by Raman spectroscopy.

Thermal rectification, in which heat flows only in the desired direction, can be achieved by controlling thermal conductivity with nanostructured graphene. Asymmetric thermal conduction has been reported⁹⁾ for suspended graphene where a region with 5 to 6 nm diameter holes is arranged in a square lattice with 30 nm spacing adjacent to a region without holes. It was observed that the thermal conductivity is higher when heat flows from a region with holes to a region without holes than in the reverse direction. This is thought to occur because high-energy phonons localize in the region with holes, resulting in an asymmetric temperature distribution. Applications for thermal rectification devices (thermal diodes) allowing heat to flow only in one direction can be expected using this effect.

3 Thermal Conductivity of Graphene with Zig-Zag Edges

3.1 Sample Preparation

We have developed a technique to process graphene at the atomic scale using hydrogen plasma etching¹⁰⁾. Since the edges of the processed graphene have atomically-ordered zigzag structures^{11, 12)} terminated with single hydrogen atoms¹³⁾, ideal zigzag edges that maintain graphene’s sp² bonding all the way to the edge can be obtained with this technique. This

contrasts with edges obtained by conventional processing methods such as oxygen plasma etching and RIE that have random structures at the atomic scale and, in some cases, may be disordered in an amorphous state.

Hydrogen plasma etching forms nanoscale hexagonal holes (nanopits) in graphene, centered around defects that serve as seeds for etching. These hexagonal nanopits can be used as basic elements for processing graphene. If the defects, seeds for nanopits, are processed into an elliptical shape, elongated hexagonal nanopits are formed, and GNRs can be obtained between the hexagonal nanopits arranged in parallel. Unlike previous studies⁷⁾, the edges of these GNRs are ordered atomically in a zigzag pattern, so they are called as zigzag GNRs (zGNRs). Additionally, if hexagonal nanopits are arranged in a triangular lattice, a graphene mesh structure can be created. This could be considered as an “antidot lattice” focusing on the periodic arrangement of holes, or it could also be considered as a “zGNR network” focusing on how the zGNRs are connected in a mesh-like pattern. In this article, this structure is described as a zigzag graphene nanomesh (zGNM). By etching away the substrate after attaching electrodes to the fabricated graphene, a suspended structure bridged between the electrodes are obtained. Since graphene is extremely thin, it is susceptible to influences such as substrate roughness and charge transfer from the substrate. However, these effects can be eliminated by creating a suspended structure.

3.2 Thermal Conductivity Measurement by Raman Spectroscopy

There are two general methods for measuring thermal conductivity: the steady-state method and the transient method. The steady-state method applies a thermal gradient to a device by setting different temperatures at both ends of the device, and thermal conductivity is determined by measuring the temperature at various points within the device. In contrast, the transient method applies a known amount of heat to the device and the thermal conductivity is determined from the resulting temperature change of the device. This study measured the thermal conductivity of graphene devices using the transient method known as opto-thermal Raman measurement employing Raman spectroscopy. In this method, the laser light for Raman spectroscopy serves as the heat source, and temperature

changes are measured using shifts in the Raman spectrum. In case of graphene, G, 2D, and D bands appear in the Raman spectrum, and the G and 2D bands are known to red-shift with increasing temperature¹⁴⁾⁻¹⁶⁾. Among these two G and 2D bands, the 2D band arises from second-order phonon scattering and changes its shape depending on the number of graphene layers, while the shape of the G band originating from in-plane vibrations of graphene honeycomb lattice does not depend on the number of graphene layers. Therefore, the temperature of graphene was estimated from the shift of the G band in this study.

The G-band shift $\Delta\omega_G$ (cm^{-1}) is reported^{14), 15), 17)} to be proportional to the graphene temperature change ΔT as shown in the Eq. (1).

$$\Delta\omega_G = \chi\Delta T \quad (\chi \cong -1.6 \times 10^{-2} \text{ cm}^{-1}/\text{K}) \quad (1)$$

The amount of heat absorbed by graphene P_{abs} (W) is expressed by Eq. (2) as a function of the excitation laser power P (W) irradiated to graphene, the absorptivity of graphene α ($\sim 2.3\%$ ¹⁸⁾), the number of layers N (thickness $t \cong N \times 0.335 \text{ nm}$), and the ratio of the area occupied by graphene inside the laser spot S .

$$P_{\text{abs}} = P(1 - \exp(-\alpha N)) S \quad (2)$$

Figure 2 shows the G-band peaks measured for a zGNM at different laser powers. As the power increases, the temperature of graphene rises, causing the G-band to shift toward the lower frequency (longer wavelength).

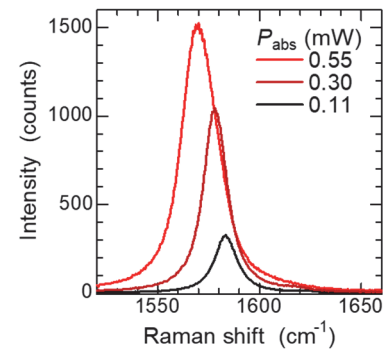


Figure 2 Changes of G-band peak when varying laser power irradiating graphene. As the amount of heat absorbed by graphene increases, the signal intensity becomes larger, while the peak center shifts toward the lower frequency.

Since the thermal conductance K (W/K) is the coefficient of temperature change when heat is loaded to a substance, that is

$$P_{\text{abs}} = K\Delta T. \quad (3)$$

K can be described with P_{abs} and $\Delta\omega_G$ as

$$K = \chi \left(\frac{\partial \Delta \omega_G}{\partial P_{\text{abs}}} \right)^{-1} \quad (4)$$

In this study, zGNM, graphene processed into a mesh pattern (Figure 3(a)), and zGNR, graphene processed into a ribbon pattern (Figure 3(b)), were used as samples. We used exfoliated graphene with 2 to 10 layers thick and varied the width of the nanoribbon (W) from 20 to 160 nm. The arrangement periodicity of hexagonal nanopits is 350 to 540 nm for zGNM. Electrodes were attached to the zGNM and zGNR devices as shown in Figure 3, and the excitation laser irradiated the central part of a device. As a result, the electrodes act as a heat bath at room temperature, and the heat is expected to flow from the central part of the device toward the electrodes. Equation (5) was used to determine the thermal conductivity κ ($\text{W/m} \cdot \text{K}$) from K taking into account the size of the device.

$$\kappa = \frac{L_{\text{eff}}/2}{tW} K \quad (5)$$

Here, L_{eff} is the length of the thermal conduction path indicated by the red lines in Figure 3, and W is the width of the nanoribbons making up the ribbon or mesh. Although a quasi-one-dimensional thermal conduction path was assumed for mesh-structured devices, heat can also flow parallel to the electrodes, in reality. It possibly results in overestimation of κ . Additionally, among two types of devices used in this study: suspended (floating above substrate) and supported (held on substrate), it should be noted that the effects of the substrate are included in κ of supported devices.

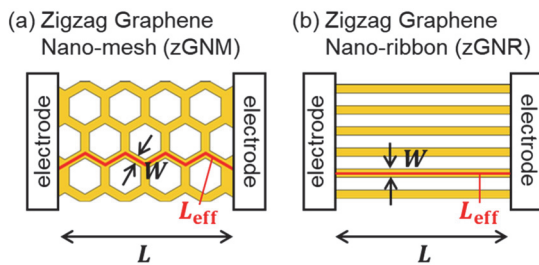


Figure 3 Schematics of (a) zGNM and (b) zGNR devices with electrodes attached. A quasi-one-dimensional path indicated by the red line was used to calculate κ for zGNM.

To prevent degrading the graphene device by continuous irradiation of the excitation laser, the laser irradiation time should be as short as possible. However, this results in a poor spectrum signal-to-noise (S/N) ratio. To overcome these contrary effects, short time (~ 0.02 s) measurement was repeated. On the other hand, the laser focus becomes

misaligned due to thermal drift as measurement progressed. Therefore, each spectrum was obtained by changing the height of the objective lens, and the G-band wave number when the laser was focused at the sample position was extracted. Figure 4 shows the dependence of the G-band wavenumber (ω_G) and intensity (I_G) on the objective lens height (z) for each measurement. The results show how z , where ω_G becomes minimum and I_G becomes maximum, changes gradually as measurement was repeated. The measured minimum values of ω_G obtained in this way were averaged.

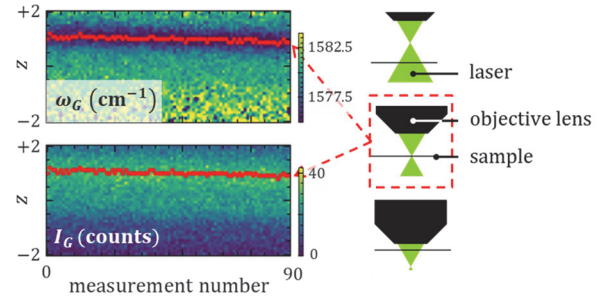


Figure 4 Changes in G-band wavenumber (ω_G) and intensity (I_G) at each measurement while repeatedly scanning lens height (z), and schematic of relationship between spread of laser irradiated through objective lens and sample position. The objective length height, z , where ω_G becomes minimum (and I_G becomes maximum) for each measurement is indicated by a red dot. The laser is focused at this height.

Figure 5 shows the G-band shift ($\Delta \omega_G$) normalized by device dimensions ($(2tW/L_{\text{eff}})\Delta \omega_G$) relative to the heat absorbed by graphene (P_{abs}) for devices with various ribbon widths (W). It shows that the G-band shifts largely as P_{abs} increases. According to Eqs. (4) and (5), the reciprocal of this slope corresponds to the thermal conductivity κ . Contrary to the classical picture, the slope of the graph becomes more gradual as W becomes narrower, meaning that κ is increasing.

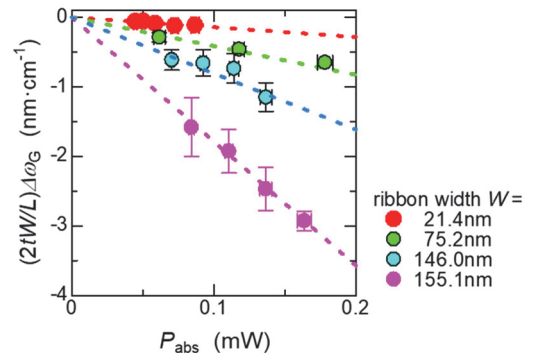


Figure 5 P_{abs} dependence of G-band shift normalized by device dimensions for zGNMs with different ribbon widths (W). Linear fits are shown as dotted lines.

3.3 Thermal Conductivity of Suspended Sample

Figure 6 shows the dependence of thermal conductivity (κ) on ribbon width (W) for (a) suspended zGNM and (b) suspended zGNR. The graphene thickness (t) dependence of κ suggests that, κ does not depend on t for 4 or fewer graphene layers (thin) and 5 or more graphene layers (thick). Therefore, in Figure 6, samples with 4 or fewer layers are shown in blue, and samples with 5 to 10 layers are shown in red. Only the κ for zGNRs with 5 to 10 layers is shown in red, since zGNRs with 4 or fewer layers could not be created. Surprisingly, κ is found to be inversely proportional to W for suspended zGNMs (Figure 6(a)) with 4 or fewer layers (thin). In addition, for 5 to 10-layer (thick) zGNMs, κ remains nearly constant for W down to ~ 30 nm. This behavior contradicts the classical picture where κ decreases as the structure becomes smaller due to increased phonon scattering.

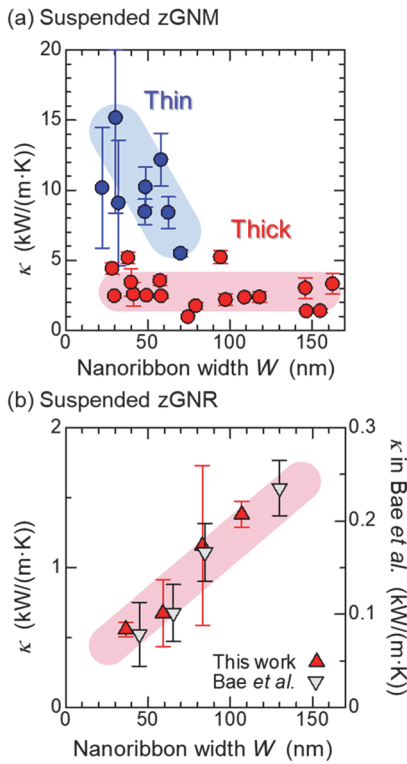


Figure 6 Dependence of thermal conductivity (κ) on ribbon width (W) for (a) suspended zGNM (●) and (b) suspended zGNR (▲/▽). Thin samples with 4 or fewer layers are shown in blue, and thick samples with 5 to 10 layers are shown in red. ▲ in (b) represents the results of this measurement (left axis), and ▽ represents the results of previous research⁷⁾ (right axis).

On the other hand, for suspended zGNR, κ decreases as W becomes narrower, similar to previous research. As superimposed the results of previous research⁷⁾ in Figure

6(b), our result shows the similar dependence of κ on W . Accordingly, it can be considered that the inverse proportionality of κ to W in suspended zGNM is due not to the zigzag edges but to the mesh structure. Although the detailed examination of this mechanism is the subject of future research, it is possible that κ is enhanced due to the interference of phonons in the nano-scale mesh structure as a result of the wave nature of phonons becoming non-negligible. In fact, enhancement of the thermal conductance as the spacing between circle holes decreases (phononic boost) has been calculated for an 80 nm silicon thin film with periodically placed circular holes (phononic crystal) at ultra-low temperature of 1K¹⁹⁾. This is because of the increased phonon density caused by phonon interference.

Although the dependence of κ on W for suspended zGNRs is similar to previous study, the value of κ is 6 to 7 times larger than the prior one. Compared to this study, the GNRs in previous study were supported on SiO₂ substrates²⁰⁾, and were made from single-layer graphene²¹⁾, while this study used graphene with 5 to 10 layers thick. Nevertheless, even after taking these differences into account, the zGNRs fabricated in this study still show approximately four times larger values of κ than GNRs from previous study. This suggests that the atomically ordered zigzag edges of zGNRs fabricated in this study increase the κ compared to previous devices with random or amorphous edges. In fact, several theoretical studies^{22), 23)} have proposed that GNRs terminated with zigzag edges (zGNRs) show larger κ than GNRs terminated with armchair edges (aGNRs). On the other hand, since κ can be overestimated for zGNMs as discussed before, the fact that zGNMs show several times larger κ than zGNRs might be due to such mathematical effects.

3.4 Thermal Conductivity of Supported Sample

Figure 7 shows the dependence of κ on nanoribbon width (W) for (a) supported zGNM and (b) supported zGNR. Since the κ is calculated only from the device dimensions and the effect of the substrate is not considered, the κ here are nominal values. Since the κ of supported samples do not depend on the graphene thickness (t) unlike suspended samples, the data for all samples with 2 to 10 layers thick are shown without categorizing. Here, non-classical behavior where κ increases as W decreases are observed for both supported zGNMs and zGNRs.

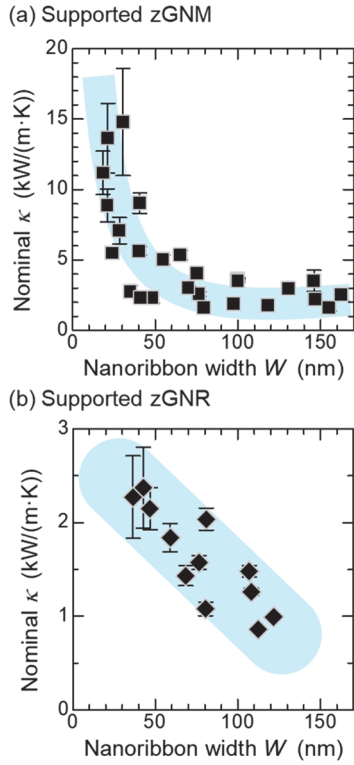


Figure 7 Dependence of thermal conductivity (κ) on nanoribbon width (W) for (a) Supported zGNM and (b) Supported zGNR.

Considering the results of suspended devices, the fact that the κ does not suppressed in supported samples despite nanostructuring can be attributed to the substrate effects rather than device structure or zigzag edge effects. This suggests that, despite the SiO_2 being an electrical insulator and poor thermal conductor, heat flow to the SiO_2 substrate cannot be ignored, and the substrate effect becomes more significant as device size decreases. This can be explained by three-dimensional (3D) heat spreading from the device to the substrate²⁴⁾. As shown in Figure 8, heat from the device surface flows perpendicular to the surface into the substrate. However, it flows radially at the device edges. While heat flow in the perpendicular direction is dominant when the device is large, the effect of 3D flow into the substrate becomes more significant as the device becomes smaller, increasing the nominal κ of a device.

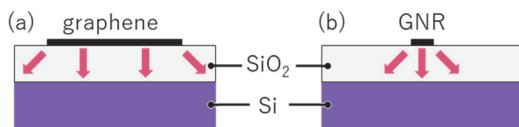


Figure 8 Schematics of heat dissipation from (a) graphene and (b) GNR to substrate.

4 Conclusions

Similar to electrical conductivity, thermal conductivity (κ) generally decreases as devices become smaller. However, under conditions where the wave nature of quasi-particles responsible for heat conduction becomes significant, κ can possibly increase with device miniaturization. This study demonstrated that the thermal conductivity (κ) increases as nanoribbon width (W) decreases by processing graphene with high crystallinity and long mean-free-path processed into a mesh structure with atomically ordered zigzag edges. The increase in κ becomes significant at W of <100 nm and reaches $\sim 15,000$ W/(m·K) at W of ~ 30 nm. This is the world's first experimental demonstration that κ increases due to the wave nature of the lattice vibration and its interference.

Graphene is expected to be used as a key material for heat dissipation and thermal control in high-frequency circuits, power-supply circuits, high-sensitivity sensors, etc., due to its higher thermal conductivity (κ) compared to other materials. It is useful from an application perspective that its κ can be further enhanced by fabricating it into a mesh structure as demonstrated here. For carbon nanotubes (CNTs), which are attracting attention as heat-dissipation materials due to high κ , the ideal value of κ is reported²⁵⁾ to be 6,600 W/(m·K). There are also reports²⁶⁾ of high thermal conductivity reaching 13,300 W/(m·K) by employing specialized sample preparation techniques. Although κ for suspended zGNMs can be overestimated when it is calculated from thermal conductance (K), it is worth noting that we obtained a κ that is equivalent to or higher than that of CNTs with high thermal conductivity. Commercially, heat-dissipation components based on graphene with high thermal conductivity reaching 1,800 W/(m·K) have been reported^{27), 28)}, but the κ value of zGNMs fabricated in this study exceeds that by almost an order of magnitude, suggesting promising future applications.

It is also an advantage of the zGNM fabricated in this study that the value of κ can be controlled by a factor of about 8 simply by changing the W . This feature enables to design graphene sheets with desired κ depending on the applications.

Another important discovery is that this unconventional behaviour can be obtained in the presence of a substrate. It can overcome the problem of single layer graphene sheets

curling up due to the Mermin-Wagner theorem. Additionally, in general, multi-layer graphene is easier to handle than single layer graphene. The fact that multi-layer graphene supported on substrate can have a high value of κ that is inversely proportional to W strongly suggests possibilities for practical applications.

Acknowledgements

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Application of Graphene Towards Hydrogen-Gas Sensor

Masahiro Kamada, Tomohiro Matsui

[Summary]

Since graphene is exposed to the external environment, it is sensitive to surrounding conditions, such as ambient gases. Therefore, it can be developed to high-sensitivity molecular sensors by decorating with receptors that capture target molecules. This article introduces application of graphene to hydrogen-gas sensing devices which have been drawing attention from various fields including energy and medicine. By fabricating graphene into nano-ribbons/meshes and decorating them with palladium as a hydrogen receptor, we successfully developed a high-sensitivity hydrogen-gas sensor capable of detecting hydrogen concentrations below 1 ppm in vacuum. The response of the device electrical resistance to hydrogen gas was different from the results of previous studies, suggesting a possibility of unknown physical phenomenon. Future work to clarify the sensing mechanism and to develop practical device structure and test systems are expected to lead to small, low-power hydrogen-gas sensors.

1 Introduction

Two-dimensional electron systems used for semiconductor devices are inherently located hundreds of nanometers deep from the surface because they are created at the interface of different materials. In contrast, graphene exists on the substrate surface and is exposed to the outside world. Therefore, graphene is sensitive to the external environment, and its physical properties can easily change by altering the surrounding environment, such as ambient gases. Conversely, these changes in graphene's physical properties can be used to evaluate the surrounding environment, which is why graphene is considered to be a promising sensor¹⁾. However, target selectivity is difficult to achieve by using graphene as is, and the sensitivity is not necessarily adequate. Selectively detecting molecules rather than physical quantities such as magnetic or electric fields can be achieved by covering graphene with membranes that only allow passage of the target molecules, or by decorating graphene with receptor substances that capture target molecules. For example, high-sensitivity odor sensors have been proposed by decorating graphene with peptides that have amino acid sequences added to interact with odor molecules²⁾. There are also reports of making biosensors for COVID-19 that are more accurate than antigen tests and faster than PCR tests by decorating graphene with antibodies to the SARS-CoV-2 spike protein³⁾. In principle, it is possible to use graphene in sensors targeting various molecules by changing the receptors.

On the other hand, hydrogen gas is attracting attention as a next-generation energy source for achieving a carbon-neutral society, as well as having applications in medical and disease diagnosis settings. However, hydrogen gas is explosive in air at a wide range of concentrations from 4% to 75%, and also degrades steel materials (hydrogen embrittlement) due to its high material permeability. Being colorless, tasteless and odorless, it is undetectable by human senses. Therefore, detection of hydrogen leaks and measurement of hydrogen concentrations are important at each stage of hydrogen generation, storage, transport, and use. Conventionally, sensors using mass spectrometry are used to measure concentrations accurately, but the equipment becomes large. Small sensors primarily for leak detection either detect changes in the electrical resistance of metal oxide semiconductors due to redox reactions with hydrogen, or detect changes in the electrical resistance of platinum (Pt) wires due to hydrogen gas combustion, or detect changes in the thermal conductivity of gases. However, all operate at high temperatures of several hundred degrees Celsius, which not only increases power consumption but also raises issues about hydrogen explosions, depending on the gas concentration⁴⁾. Therefore, development of hydrogen-gas sensors that are not only fast and sensitive, but are also small, low power and can operate at room temperature is anticipated.

This article introduces the possible applications of graphene, which is nano-fabricated with our unique hydrogen

plasma etching technique, towards hydrogen-gas sensors. Section 1 outlined the potential applications of graphene for molecular sensors and the current state of hydrogen-gas sensing. Section 2 introduces previous research on hydrogen-gas sensing using graphene, and Section 3 discusses our unique methods of preparing samples and evaluating sensor characteristics, as well as experimental results.

2 Hydrogen Gas Sensing Using Palladium-Decorated Graphene

While platinum (Pt) and palladium (Pd) are both known hydrogen receptors, this study only used Pd, which is the more common metal that stores hydrogen even at room temperature. When Pd is exposed to hydrogen gas, hydrogen molecules (H_2) adsorbed on the Pd surface decompose into hydrogen atoms (H), which penetrate into the gaps of the Pd crystals to expand the crystals, forming hydrogenated palladium (PdH_x), and changing the Pd physical properties, such as volume and electrical resistance. The palladium-decorated graphene used in this study detects changes in the Pd physical properties as changes in the graphene electrical resistance. While the presence or concentration of hydrogen gas can be obtained from measuring the physical properties of Pd itself, using Pd-decorated graphene allows for detection of even more subtle changes in Pd properties, meaning even smaller amounts of hydrogen gas can be detected. In addition, it keeps the amount of precious Pd metal low. Furthermore, being able to operate at room temperature without heating for sensing is another major advantage.

It is reported⁵⁾⁻¹¹⁾ that, when Pd-decorated graphene is exposed to mixed gases containing hydrogen at room temperature under atmospheric pressure, the electrical resistance of the graphene device increases according to the concentration of hydrogen gas, and when the hydrogen gas is removed, the electrical resistance returns to its original value. This resistance change is explained by the differences in work functions between graphene, Pd, and PdH_x . In other words, electrons are transferred from graphene to Pd because graphene's work function ($\phi_G = 4.5$ eV) is smaller than Pd's ($\phi_{Pd} = 5.6$ eV) and, therefore, holes are doped into graphene. However, when Pd is hydrogenated, the work function of PdH_x ($\phi_{PdH} = 3.2$ eV) is smaller than the work function of graphene, so the charge flow is reversed, with electrons transferring from PdH_x to

graphene, and, therefore, electrons are doped into graphene. As a result, the electrical resistance of the originally hole-doped graphene devices decreases with Pd adsorption, and increases with exposure to hydrogen.

On the other hand, the sensor functional improvements are expected by nano-structuring of graphene and the processed graphene edges may have functionality in themselves¹²⁾. The length of edges per surface area increases in nanostructures, which improves the sensitivity of detection using edges^{9), 13)}. While devices with graphene nanoribbons (GNRs) processed into nanoscale ribbon shapes arranged in parallel, have lower sensitivity compared to unprocessed cases because the graphene area is smaller and the Pd amount is reduced, there are reports⁸⁾ of faster reaction when hydrogen gas is introduced and recovery when hydrogen gas is removed as well as higher sensor reproducibility.

3 Nanoscale Graphene Hydrogen-Gas Sensor

3.1 Fabricating Nanoscale Graphene

A key feature of this study is nanoscale fabrication of graphene using hydrogen plasma^{14), 15)}. Etching graphene with hydrogen plasma creates hexagonal holes (nano-pits) centered on defects and it is known that the edges are atomically aligned in a zigzag structure^{16), 17)}. With common processing methods, such as oxygen plasma etching or mechanical cutting, the graphene edges are random structures at the atomic scale, and in most cases, the hexagonal rings of carbon atoms are broken into pentagonal or heptagonal rings, or become amorphous^{14), 15)}. In contrast, edges obtained by hydrogen plasma etching have an ideal zigzag shape that maintains the honeycomb structure of graphene¹⁶⁾⁻¹⁸⁾. In nanoscale devices where edge-to-area ratio is larger, it is important to create controlled edges, which is possible using this method with good reproducibility. Additionally, graphene's zigzag edges have electronic states at the energy of the charge neutrality point (Dirac point) due to the breaking of crystal symmetry^{14), 19)}. This is called a zigzag edge state. Therefore, there are hopes that new functionality can be acquired through zigzag edge states while maintaining the graphene properties¹³⁾. This study prepared zigzag graphene nanoribbons (zGNR), which are nanoscale ribbon-shaped graphene terminated with zigzag edges, and zigzag graphene nanomesh (zGNM), which is processed into a mesh by arranging hexagonal nanopits in a

triangular lattice. The zGNM can be interpreted either as a network structure of zGNRs or as an antidot lattice, depending on the perspective or the physical properties being considered. Figure 1(a) shows an atomic force microscope (AFM) image of a zGNR device, and (b) shows an AFM image of a zGNM device. zGNM is believed to have higher industrial value because it is structurally more robust than zGNR as it is connected in a network pattern. Devices used in this study had a total width of 2 to 5 μm , a length of 2 to 3 μm , and a ribbon width (W) of 100 to 200 nm.

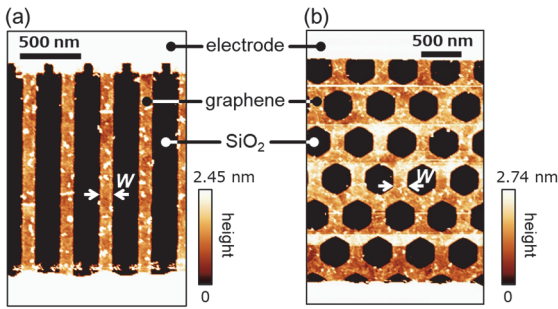


Figure 1 AFM Images of (a) zGNR (approx. 2-nm thick, $W \sim 120$ nm) and (b) zGNM (approx. 2.5-nm thick, $W \sim 200$ nm)

3.2 Pd Decoration of Graphene

Several methods are known for decorating graphene with Pd, and the most common is the thermal evaporation in a vacuum, with which Pd is adsorbed on to the graphene as nanoparticles⁵⁾. When the deposited amount is small, the Pd nanoparticles are isolated. As the deposited amount increases, the Pd nanoparticles begin to cover the entire surface of the graphene, and eventually the Pd begins to exhibit electrical conductivity as a film as the bonds between nanoparticles strengthen. In this study, Pd was vacuum-deposited on zGNR and zGNM by heating Pd with an electric current. Figures 2(a)-(c) show AFM images of the Pd nanoparticle layer for several deposition film thicknesses. Here, the Pd-film thickness was measured using a film thickness meter with a quartz micro-balance placed alongside the graphene sample. Since this corresponds to the film thickness when Pd is attached uniformly to the surface, we express it as the ‘nominal’ film thickness (d_{nom}). In reality, as shown in Figures 2(a)-(c), Pd is deposited as nanoparticles with diameters of 10 to 20 nm, so the surface has irregularities, and the film thickness actually measured by AFM is thicker than d_{nom} . It is known that actual film thickness depends on the Pd deposition rate, too. In this study, Pd was deposited with the deposition rate of approx. 0.9 $\text{\AA}/\text{s}$ with respect to d_{nom} .

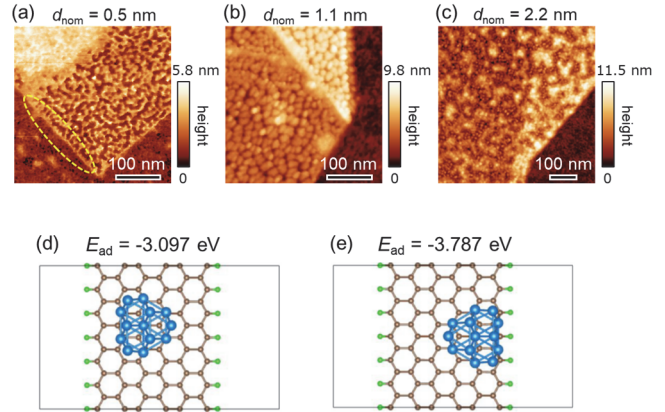


Figure 2 (a)-(c) AFM images of Pd deposition films with nominal thicknesses $d_{nom} =$ (a) 0.5 nm, (b) 1.1 nm, (c) 2.2 nm. As shown by the yellow dotted line in (a), Pd is adsorbed selectively at the edges of graphene flakes. (d)(e) Adsorption structures of Pd clusters (Pd_{13}) placed at center (d) and edge (e) of zGNR, along with their respective adsorption energies (E_{ad}).

As shown by the dotted line in Figure 2(a), the Pd adsorption state at the graphene-flake edges differs from that on the graphene surface. Pd is adsorbed on the surface as nanoparticles, but Pd clumps at edges. This occurs regardless of whether the edge structure is zigzag or armchair-type¹⁴⁾. This selective adsorption of Pd at the edges has also been confirmed by first-principle calculations using the *Fugaku* supercomputer (Figures 2(d)(e)). When calculating the adsorption energy (E_{ad}) of a cluster of 13 Pd atoms (Pd_{13}) adsorbed on zGNR with a width of 0.99 nm, E_{ad} was -3.07 eV on the graphene surface (Figure 2(d)), while -3.78 eV on the zigzag edge (Figure 2(e)). This indicates that Pd clusters are energetically more stable when adsorbed at the edge, meaning they are more likely to adsorb at edges. Since the physical properties of Pd are being measured in hydrogen sensors, the more Pd, the higher the sensor sensitivity in principle. Therefore, the presence of graphene zigzag edges may result in improved sensor sensitivity.

When considering hydrogen sensor applications, it is thought that Pd nanoparticle is better to increase the surface area of Pd exposed to hydrogen gas. Conversely, the Pd film thickness should be as thin as possible to transmit the changes in Pd’s physical properties to graphene more quickly and efficiently. Therefore, in this study, Pd nanoparticles were controlled to have a nominal thickness (d_{nom}) of approx. 0.7 nm to cover the entire surface of the graphene with a single layer.

3.3 The Way to Evaluate Hydrogen-Gas Sensor Characteristics

Figure 3(a) shows the configuration of a hydrogen-gas sensor using Pd-decorated graphene, and Figures 3(b)(c) show the schematic illustrations of the system for evaluating its features. Electrodes are attached to graphene processed into zGNM or zGNR before decoration with Pd. The substrate holding the graphene is highly doped conductive silicon (Si), covered with an insulating surface of SiO₂. The graphene electronic state can be changed by the field effect by applying a gate voltage (V_g) to the Si (Figure 3(a)). This study mainly used a vacuum measurement system (Figure 3(b)). In this setup, only hydrogen gas is introduced into the measurement chamber, which is at a base pressure of ≈ 0.01 Pa, and the introduced amount of hydrogen gas is controlled by the partial pressure (P_{H_2}) of hydrogen gas. On the other hand, most previous studies measured the electrical resistance of the device while flowing hydrogen gas diluted with inert gases such as dry air, nitrogen, or argon into the measurement chamber (Figure 3(c)). In other words, the measurement chamber is at atmospheric pressure. This latter method is closer to actual conditions for detecting leaks to atmosphere, but, since non-hydrogen gas molecules are also present, it is unsuitable for examining the pure reaction between the device and hydrogen.

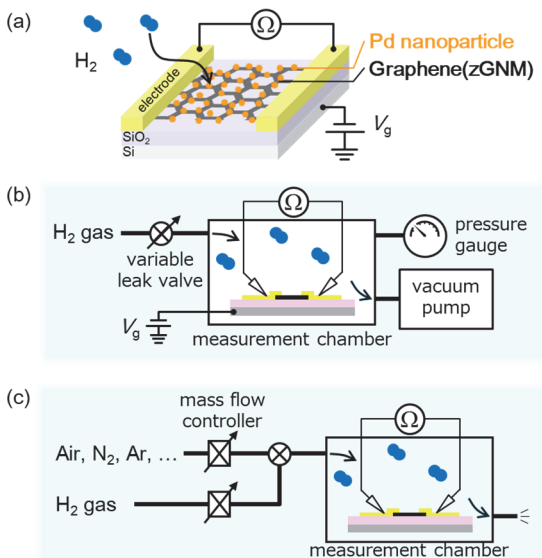


Figure 3 (a) Schematic diagram of Pd-decorated graphene hydrogen gas sensor. (b)(c) Schematic diagrams of hydrogen sensor characterization systems mainly used in this study. Configuration for measurement in vacuum (b), and configuration for measurement at atmospheric pressure (c) used widely in previous research.

3.4 Hydrogen Gas Sensing with Pd-Decorated zGNR/zGNM

Figure 4(a) shows the response by Pd-decorated zGNR (device A) to hydrogen gas. This was measured using the system shown in Figure 3(b). As shown, the electrical resistance (R) of graphene decreases with the introduction of hydrogen gas, and the magnitude of this change depends on the partial pressure (P_{H_2}) of the hydrogen gas. Figure 4(b) shows the ratio of change in R relative to the initial electrical resistance of the device (R_0) before introducing hydrogen gas ($\Delta R/R_0 = (R - R_0)/R_0$) versus P_{H_2} . Although there is some variation, the ratio of change increases as the amount of hydrogen gas increases. This dependency suggests that device A can measure down to $P_{H_2} \leq 0.1$ Pa, corresponding to a concentration of 1 ppm — a sensitivity capable of detecting hydrogen gas in human breath²⁰). Consequently, we found that a hydrogen-gas sensor with high sensitivity even at room temperature can be realized using our fabrication technique.

In the case of device A (Figure 4(a)), R continued changing even at 2 minutes after hydrogen gas was introduced, and had not returned to the initial value even at 5 minutes after stopping introduction of hydrogen gas. However, since R starts changing within 0.5 s after introducing hydrogen gas into the measurement chamber, the continuing sensor response is not because the chemical reaction rate is slow. The continuing change in R is possibly because the amount of hydrogen gas contributing to the reaction per unit time is small, requiring time for the entire device to change. Decreasing the device size or improving its structure might lead to a sensor where R saturates within a practical measurement time. In fact, a device, R of which saturates within 50 s after introducing hydrogen gas, had also been obtained.

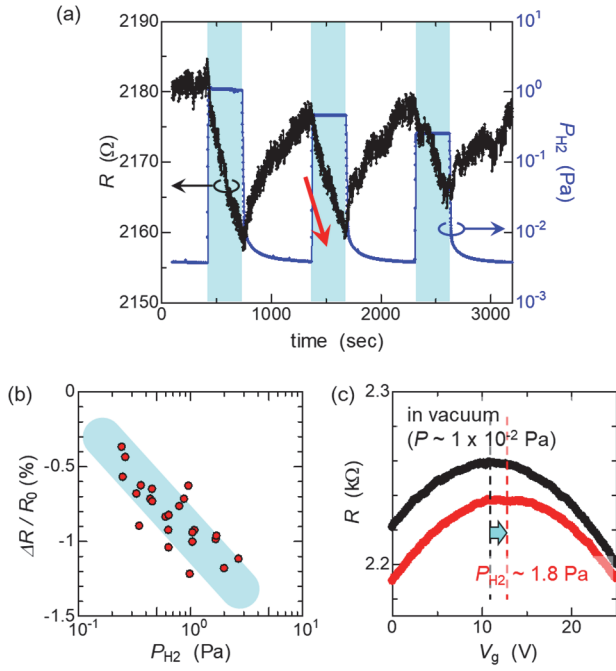


Figure 4 (a) Changes in R of Pd-decorated zGNR (device A) when hydrogen gas is introduced into measurement chamber in vacuum. The electrical resistance of device A (black) is on the left axis, and P_{H_2} is on the right axis. (b) $\Delta R/R_0$ of device A versus P_{H_2} . (c) V_g dependency of R for device A. The black line shows measurement in vacuum ($P \sim 1 \times 10^{-2}$ Pa), and the red line shows measurement at $P_{H_2} \sim 1.8$ Pa.

Figure 4(c) shows the measurement of the V_g dependency of R for the same device A to investigate the origin of these electrical resistance changes. Graphene has linear energy dispersion with the charge neutrality point (Dirac point) as its intersection. Therefore, the V_g dependency of R reaches its maximum at $V_g = V_{\text{Dirac}}$ corresponding to the Dirac point energy. Figure 4(c) shows that as hydrogen gas is introduced, not only does R decrease, but V_{Dirac} also shifts in the positive direction. This behavior was reproducible with the introduction and removal of hydrogen gas, and was similarly observed in all measured samples regardless of the graphene shape. The fact that V_{Dirac} is positive even before hydrogen-gas introduction indicates that device A was originally hole-doped, and the increase in V_{Dirac} in the positive direction at hydrogen-gas introduction suggests that hydrogen gas causes hole-doping in graphene. In other words, this study found that the Pd-decorated graphene used in this study is hole-doped by hydrogen-gas introduction, resulting in decreased electrical resistance at $V_g = 0$.

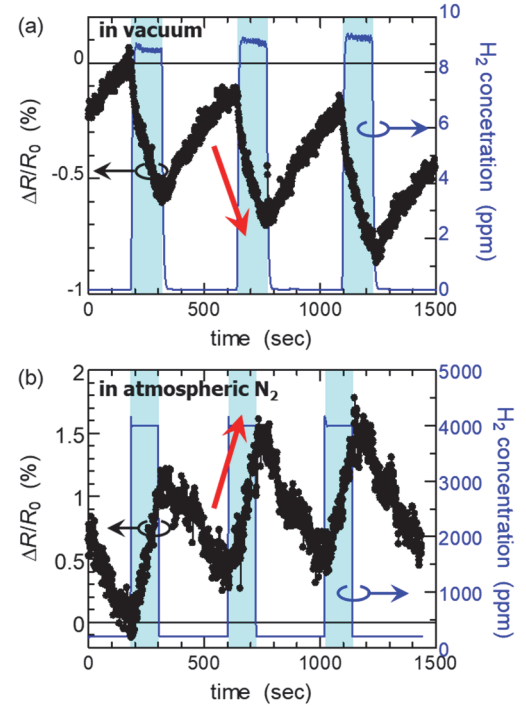


Figure 5 Comparison of hydrogen-gas response of Pd-decorated zGNR (device B). (a) Measurement in vacuum, (b) Measurement in nitrogen atmosphere at atmospheric pressure.

As a result, the Pd-decorated graphene devices fabricated in this study can be expected as highly sensitive hydrogen-gas sensors operating at room temperature. However, the result where R decreases due to hydrogen gas is opposite to that reported in previous research. Differences between this and previous studies lie in the graphene processing method and the measurement environment explained in section 3.3. While this study made measurements in vacuum, previous studies measured at atmospheric pressure. Therefore, we compared the reaction for the same device in vacuum and using hydrogen gas diluted with dry nitrogen gas at atmospheric pressure. Figures 5(a) and (b) show the results for Pd-decorated zGNR (device B) in vacuum and at atmospheric pressure, respectively. These figures show that, similar to Figure 4(a), R decreases with the introduction of hydrogen gas in vacuum, but increases in nitrogen gas at atmospheric pressure, which is consistent with previous research. Additionally, the hydrogen-gas concentration needed to obtain approximately 0.5% of $\Delta R/R_0$ was about 9 ppm (0.9 Pa) in vacuum, but approximately 2000 ppm (0.2%) at atmospheric pressure. These results show a decreased sensitivity of more than 2 orders of magnitude. Since the dependency of V_g on R was not measured at atmospheric pressure, it is impossible to clearly demonstrate the doping effect shown in Figure 4(c). However, since we

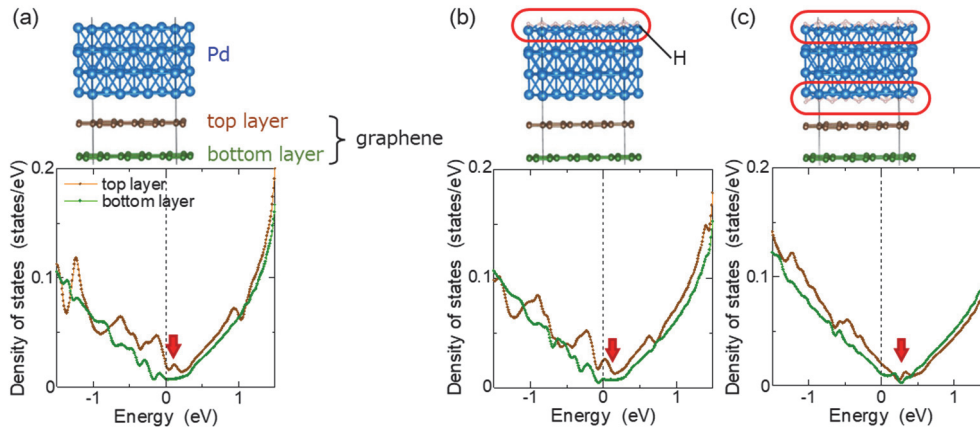


Figure 6 Cross-sections of the system where Pd clusters adsorbed on bilayer graphene (top panel) and electron density of states of graphene (bottom panel). (a) Without hydrogen atoms, (b) Hydrogen atoms adsorbed on top surface of Pd cluster, (c) Hydrogen atoms adsorbed on both top and bottom of Pd cluster.

confirmed that the device is hole-doped even at atmospheric pressure, the increase in R caused by hydrogen-containing gas is likely because electrons are doped into the graphene.

Consequently, we think the different results between our study and previous studies is not due to the device itself, but is rather due to differences in the measurement environment, specifically whether measurement is in vacuum or at atmospheric pressure. Generally, a vacuum environment increases adhesion between graphene and substrate, or removes impurities adsorbed between graphene and substrate or on the graphene surface. However, it is difficult to imagine that such effects would change the chemical reaction. In this case, the significant effect can be considered as that only hydrogen gas is introduced at in-vacuum measurement, whereas gas molecules other than hydrogen are also present at atmospheric pressure measurements. Very recent measurements introducing hydrogen gas diluted with dry air in vacuum also resulted in increased electrical resistance of the device. Considering these factors, it appears that the presence of gas molecules other than hydrogen is related significantly to the behavior of the device resistance. When hydrogen gas comes into contact with Pd-decorated graphene, holes are doped into the graphene and the electrical resistance of the device decreases, but this reaction is reversed when other gas molecules are involved. Although previous studies have used dry air, nitrogen, and argon to dilute hydrogen, the direction of the reaction is the same in all cases. Therefore, the identity and role of gas molecules other than hydrogen that contribute to increased electrical resistance of the device remain unknown and are a subject for future research.

To verify the possibility of hole doping by hydrogen molecules in Pd-decorated graphene, first-principle calculations were performed using *Fugaku* to investigate the charge transfer to graphene when hydrogen molecules that decomposed into atoms on the Pd surface are adsorbed onto Pd. *Fugaku* simulated a system where Pd atoms are adsorbed in layers on bilayer graphene. The electron density of states of graphene was calculated when hydrogen atoms are adsorbed on the layered Pd cluster surface, while simultaneously calculating the charge transfer from hydrogen-adsorbed Pd to graphene using the Bader method. Figure 6 shows the calculated model (top panel) and the corresponding electron density of states (bottom panel). Since the Dirac point is the intersection of linear energy dispersion, the energy at which the electron density of states is minimized corresponds to the Dirac point (arrow in Figure 6). Figure 6(a) shows the case without hydrogen, Figure 6(b) shows the case when hydrogen atoms are adsorbed on the top surface of the layered Pd cluster, and Figure 6(c) shows the case when hydrogen atoms are adsorbed on both the top and bottom surfaces of the layered Pd cluster, the latter of which is in contact with graphene. Although the fine structure of the electron density of states differs in each case, when hydrogen atoms are adsorbed on the top surface of the Pd cluster, the energy of the Dirac point does not change much compared to the case without hydrogen. However, when hydrogen atoms are intercalated between the Pd cluster and graphene and adsorbed on the bottom surface of the Pd cluster, the results clearly show that the energy of the Dirac point shifts toward the higher energy side. This corresponds to hole-doping in graphene, which is consistent with our experimental results.

Although graphene hole-doping also occurs according to the height of the Pd cluster from the graphene surface, the size of the charge transfer calculated by the Bader method is more than an order of magnitude larger when hydrogen is adsorbed on the bottom surface of the Pd cluster than when there are no hydrogen atoms, even at the same Pd-to-graphene distance. Moreover, an experiment showed there was no change in the device electrical resistance when introducing nitrogen gas without hydrogen. Considering all these observations, the observed hole-doping is likely a reaction specific to hydrogen gas and is possibly an effect of the hydrogenation of the bottom surface of the Pd cluster.

4 Conclusions

This study discussed the potential use of Pd-decorated graphene as a hydrogen-gas sensor using Pd as a hydrogen-gas receptor. By depositing Pd on graphene as zGNR and zGNM nano-fabricated by hydrogen plasma etching technique, we successfully created a highly sensitive hydrogen-gas sensor capable of detecting hydrogen concentrations below 1 ppm even at room temperature. Future work requires reducing variations in the electrical resistance (R) of the sensor device and investigating and improving sensitivity and response-speed limiting factors. On the other hand, in the aspect of basic research, we discovered that the R direction of change differs according to the hydrogen gas introduction environment. This discovery challenges previously established theory about the hydrogen-gas detection mechanism of Pd-decorated graphene based on the work functions of graphene, Pd, and PdH_x. Future research is required to clarify this mechanism.

Although this study targeted development of a hydrogen-gas sensor using Pd as a hydrogen receptor, sensors for other molecules could be created by changing the receptor. This research offers various sensor applications such as biosensors and odor sensors.

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in Japan (ARIM) of the Ministry of Education, Culture, Sports, Science and Technology (MEXT), Grant Number JPMXP1223UT1029, JPMXP1224UT1080 and was conducted in the Takeda Cleanroom with help from the Nanofabrication Platform Center of the School of Engineering, the University of Tokyo. Additionally, first-principle calculations were outsourced to ASMS Co., Ltd. and carried out using computational resources provided by the *Fugaku* supercomputer (Project Number: hp240309).

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- 20) The concentration of hydrogen gas in human breath is typically around 5 ppm.

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Elliptical Chokes Operating in Full WR-3 Band

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[Summary]

The WR-3 band (220 to 330 GHz) in the sub-THz wave band is expected to be used for wireless communication because it is relatively unaffected by atmospheric attenuation. Switches with mechanically sliding waveguides are used for low-loss signal switching in these high-frequency bands but electromagnetic waves leak from the mechanical switch parts. Chokes covering the fully operating band of standard waveguides would be useful for suppressing this leakage. Four elliptical choke designs (simple, non-split, split, and flat type), which operate across the full WR-3 band, were investigated and their effectiveness was demonstrated by simulation and actual measurements. Simulations compared changes in transmission loss due to waveguide misalignment, showing that the split elliptical choke in particular is less susceptible to the effects of waveguide misalignment. However, the split elliptical choke suffers larger leakage loss of ≤ 14 dB compared to the other three types of elliptical chokes, and finding a more effective solution remains a challenge for future research.

1 Introduction

There are high hopes for the 300 GHz band, which can achieve transmission speeds of tens of Gbps to support future increases in mobile data traffic. The WR-3 band (220 to 330 GHz) within these frequencies is expected to be used for wireless communication because atmospheric attenuation due to oxygen molecules and water vapor in the atmosphere absorbing radio waves is less than 10 dB/km¹⁾ making WR-3 best suited to wireless applications. In response, basic R&D into spectrum measurements is examining use of the WR-3 band for wireless communications²⁾. This research has resulted in the development of a waveguide filter bank by switching between multiple bandpass filters covering part of the WR-3 band (255 to 315 GHz) as a preselector for a spectrum measurement system³⁾. This filter bank is configured to switch between filters using 8 movable sliding waveguide filters arranged between the input and output waveguides. To prevent mechanical wear, the input and output waveguides and waveguide filters are not in contact and rectangular chokes are installed on opposing surfaces across the gap. Although chokes can effectively suppress transmission losses by reducing electromagnetic wave leakage from the gap at the connection point⁴⁾, chokes narrow the operating bandwidth according to the width of the gap (g)⁵⁾. While the referenced³⁾ chokes secured transmission losses of less than 0.1 dB and isolation of more than 50 dB even at gap width of 0.1 mm, the usable bandwidth was limited to 255 to 325 GHz,

from a practical standpoint it is better to use chokes supporting the full operating bandwidth of standard waveguides. It is difficult to reduce the gap in mechanical waveguide switches to less than 20 or 30 μm due to constraints imposed by machining precision, sliding friction, and wear. However, the ratio of the gap to the wavelength becomes larger as the frequency increases, making it more difficult to fabricate chokes operating over a wide bandwidth. Moreover, keeping any misalignment between waveguides below a few μm is difficult and the impact of misalignment becomes greater at higher frequencies. Additionally, the referenced³⁾ waveguide filter bank has chokes at both sides of the connection, but it is better to install chokes on only one side from the manufacturing cost perspective. Based on these ideas, this study examined chokes for the WR-3 band meeting the following conditions: (1) full-band operation, (2) high tolerance to waveguide misalignment, and (3) one-sided choke configuration.

Rahiminejad *et al.* has described⁶⁾ a non-contact gap adapter installed on one side of the waveguide connection and satisfying the requirements of (1) and (3) above. It is a circular waveguide connection structure that suppresses leakage between flange surfaces using a periodic structure developed by Kildal *et al.*⁷⁾ and is manufactured using SU8 polymer. While the design has good transmission characteristics in the 220 to 325 GHz operation range with a gap ($g = 0.05$ mm), the manufacturing process is complicated, and there are concerns about aging deterioration of the polymer material.

With respect to requirement (2), the egg-shaped choke flange described by Pyne *et al.*⁽⁸⁾ is reported to have no significant change in operating bandwidth due to parallel and angular misalignment of the connected waveguides. While the effect of misalignment is small, the operating bandwidth for $g = 0.5$ mm (approximately $1/50$ wavelength) is 8.5 to 11 GHz, which covers only about 75% of the full WR-90 band (8.2 to 12.4 GHz). Therefore, using the referenced choke⁽⁸⁾ for the WR-3 band would probably not cover the full bandwidth even with a gap ($g = 0.02$ mm).

This article reports our investigation into four new elliptical chokes that operate across the entire WR-3 band when installed only on one side of the connection with $g = 0.05$ mm (approximately $1/20$ wavelength). We focus on the WR-3 band where gap effects are significant and describe numerical analysis results in actual dimensions and real frequencies for intuitive understanding. The results can also be discussed as general theory by normalizing with the waveguide cutoff wavelength or cutoff frequency. Unless stated otherwise, the numerical analysis ignores conduction losses.

2 Elliptical Choke Basics

2.1 Proposed Simple Elliptical Choke

Figure 1(a) shows a general model with a gap ($g = 0.05$ mm) between flat flanges connecting two waveguides. Electromagnetic waves leak between the flanges as shown in Figure 1(b) when there is no choke groove. This example shows the results of harmonic analysis using an electromagnetic field simulator (Femtet, Murata Software Co., Ltd.) to model the opposing flange faces of a standard WR-3 waveguide (0.864×0.432 mm). Harmonic analysis drives a sine wave from a set oscillation source and analyzes the frequency response characteristics of the model to that sine wave. Figure 1(b) shows the z -direction electric field intensity at a frequency of 280 GHz. The red solid line indicates where the z -direction electric field intensity is zero, representing lines of equal phase. The power of leaking electromagnetic waves is almost circular at distances more than one wavelength from the waveguide opening, but the shape is closer to an ellipse near the waveguide. Based on the results in Figure 1, it seems natural to use an elliptical choke groove as the prototype. The leakage power from the gap is defined by Eq. (1), and the leakage loss is defined by Eq. (2). Figure 1(c) shows

the transmission loss $|S_{21}|$ and leakage in the WR-3 band, where the leakage loss is ≤ 10 dB across the entire WR-3 range, meaning $\geq 10\%$ of the power leaks out.

$$\text{Leakage} = 10 \log_{10}(1 - |S_{11}|^2 - |S_{21}|^2) \quad (\text{dB}) \quad (1)$$

$$\text{Leakage loss} = -\text{leakage} \quad (\text{dB}) \quad (2)$$

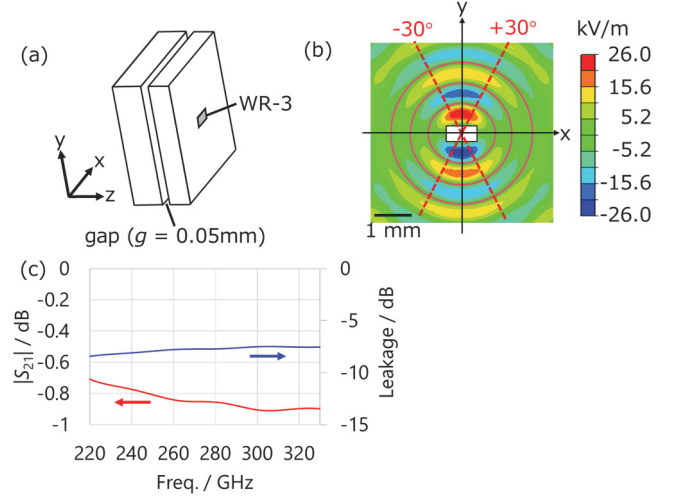


Figure 1 Leakage of electromagnetic field between flat flanges with gap ($g = 0.05$ mm) (a) Shape of harmonic analysis model, (b) Z -direction electric field strength without choke (WR-3, 280 GHz) (red solid line: Z -direction electric field strength = 0, red dotted line: position at $\pm 30^\circ$ from y -axis), (c) $|S_{21}|$ and leakage

2.2 Characteristics of Simple Elliptical Choke

This section outlines the choke principle (Figure 2). A groove with a length of one quarter of the wavelength perpendicular to the incident wave acts as a choke. Reflection at the short-circuited bottom of the groove reverses the phase of the electric field. Viewed from the incident side at the branch, the phase of wave 1 reflected at the bottom of the groove and returning to the branch, becomes in-phase with the incident wave. Conversely, viewed from the direction of the electric field, reflected wave 2 on the emission side becomes out-of-phase with the incident wave, so the two cancel each other out, suppressing leakage to the emission side.

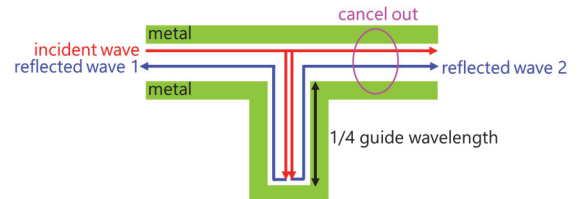


Figure 2 Choke principle

Chokes are inherently narrow bands because they prevent transmission through interference, so achieving wideband

operation requires investigation of conditions by the simulation. Using harmonic analysis of simple elliptical choke groove parameters for inner major radius (a), minor radius (b), groove width (w_c), and depth (h), and setting h to approximately $1/4$ of the guided wavelength suppresses power leakage to the outside from the choke groove over the entire WR-3 operating band. Moreover, the resonance frequency spacing (described later) becomes wider as the space surrounded by the waveguide opening and the inner wall of the choke groove becomes narrower. However, for machining, it is necessary to maintain a width of at least 0.1 mm. In practice, a transmission loss of ≤ 0.2 dB and leakage loss ≥ 15 dB were obtained at frequencies below 320 GHz as shown in Figure 3(b) when harmonic analysis of a simple elliptical choke was performed for $g = 0.05$ mm with the dimensions shown in Figure 3(a). This result shows that an elliptical choke can suppress electromagnetic wave leakage over a wide bandwidth. However, the transmission loss and leakage loss deteriorate near 324 GHz, which is insufficient for operation across the entire WR-3 band. The cause of the deterioration is described below. Since the distance from the waveguide opening to the near end of the choke groove is set to less than $1/4$ of the guided wavelength, the electric field entering the gap has almost uniform intensity in the radial direction and can be considered to propagate along the waveguide opening. In other words, since the gap near the waveguide opening of the simple elliptical choke can be considered as a ring-shaped transmission line surrounding the opening, resonance may occur in this ring. In fact, the calculated localized electric field intensity distribution around the opening suggests the possibility of resonance in the gap area at 324 GHz (Figure 3(c)).

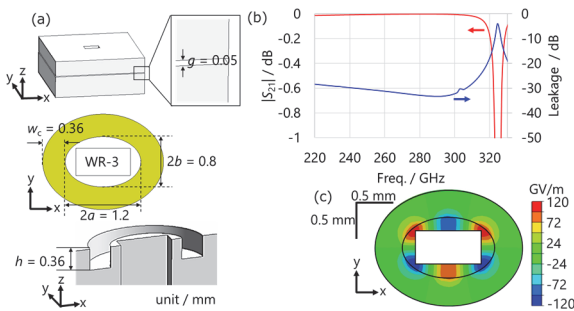


Figure 3 Harmonic analysis results for simple elliptical choke (a) Shape of analysis model, (b) $|S_{21}|$ and leakage, (c) Z-direction electric field intensity distribution at 324 GHz

2.3 Simple Resonance Analysis Model

To analyze resonance in the simple elliptical choke discussed in the previous section, we modeled the choke structure space, including the gap between the waveguide opening and the choke groove. Resonance analysis analyzes the model natural frequencies and natural vibration modes. Since the far end of the choke groove can be considered an electric wall and the extension plane of the waveguide inner wall can be considered a magnetic wall, we used a simplified resonance analysis model around the gap portion (Figure 4(a)). The electric field entering the gap is at opposite phases on both sides of the waveguide cross-sectional long edge (+y side and -y side) due to TE₁₀-mode propagation in the waveguide. Therefore, resonance in the gap is limited to cases where the direction of the electric field is antisymmetric with respect to the xz plane, and symmetric with respect to the yz plane. Only the mode at 317 GHz within the WR-3 band satisfies this condition, showing good agreement with the 324 GHz harmonic analysis result. Figure 4(b) shows the z-direction electric field intensity distribution at 317 GHz.

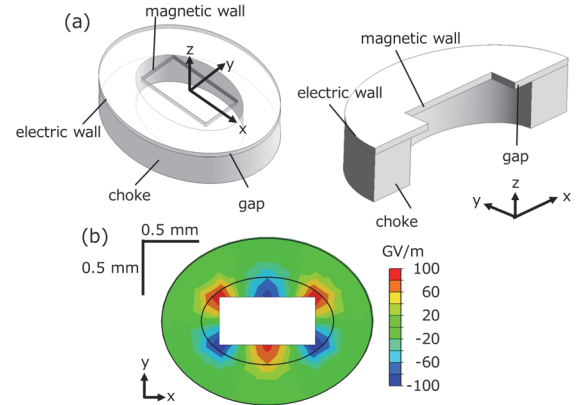


Figure 4 Simple elliptical choke resonance analysis results (a) Analysis model, (b) Z-axis (direction) electric field intensity distribution at 317 GHz

3 Structure of Suppressed-Resonance Elliptical Choke

As described in the previous section, resonance within the WR-3 operating band is a limiting factor for wideband operation. Three approaches to avoid the impact of resonance are:

- (1) Non-split elliptical choke: Provides open stub protruding into choke groove from near end of choke groove at waveguide opening, moving resonance frequency to lower range than operating band

- (2) Split elliptical choke: Creates ridge waveguide to divide choke, suppressing resonance by opening space where resonance occurs
- (3) Flattened elliptical choke: Narrows space resonance space to move resonance frequency to higher range than operating band

3.1 Non-Split Elliptical Choke

Short branch lines called stubs are added to a transmission line to add parallel impedance to the transmission line for load matching, among other applications. Installing stubs (Figures 5 to 7) in the gap of a simple elliptical choke might change the propagation conditions in the circumferential direction to significantly shift the resonance frequency. Figure 5 shows the resonance frequencies of each mode when a stub with width $w_s = 0.1$ mm and length $d = 0.155$ mm is placed in the waveguide opening plane in the direction angle θ . Distance d is the distance between A and B in Figure 5(a), showing the case where it is about half the width of the groove (w_c). However, the stub width (w_s) is set to be thin and within the mechanical processing constraints mentioned in section 2.2. The red, blue, and green circles in Figure 5(b) are calculated values and the solid lines are lines that complement these circles. There are only three resonances below 400 GHz excited in the gap by the waveguide TE₁₀ propagation mode: red: mode 1, blue: mode 2, and green: mode 3, and only mode 2 is within the WR-3 band. Resonance analysis shows that the resonance frequencies represented by the red, blue, and green graphs are consistent with the hollow circles and squares obtained by harmonic analysis. When a stub is placed along the diagonal direction of the waveguide ($\theta = 26.5^\circ$), the difference in resonance frequency between mode 2 and mode 3 is maximum at 138 GHz. While mode 2 is within the WR-3 band, the interval between the resonance frequencies of modes 2 and 3 is sufficiently wider than the WR-3 band's operating bandwidth (110 GHz), so it may be possible to cover the full WR-3 band by changing other shape parameters.

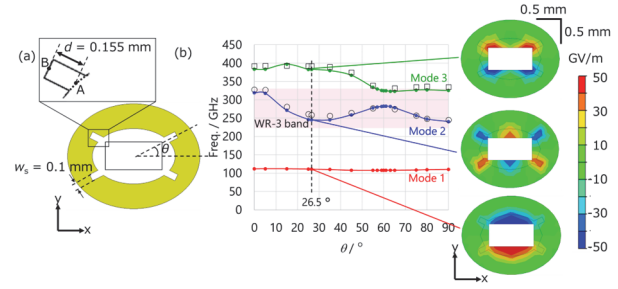


Figure 5 Resonance frequency vs stub angle ($d = 0.155$ mm) (a) Resonance analysis model, (b) θ dependence of resonance frequency and z -direction electric field strength of modes 1 to 3 ($\theta = 26.5^\circ$) (red: mode 1, blue: mode 2, green: mode 3; open circles and squares: resonance frequencies obtained from harmonic analysis)

Figure 6 shows the results of analyzing the resonance frequencies of each mode with respect to d when θ is fixed to 26.5° . Figure 6 confirms that the mode-2 resonance frequency decreases as d increases. In particular, when $d > 0.24$ mm, neither mode 2 nor mode 3 resonates within the WR-3 band.

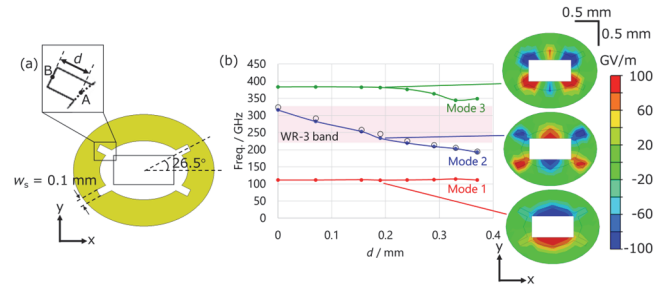


Figure 6 Resonance analysis results with stub ($\theta = 26.5^\circ$) (a) Resonance analysis model, (b) d dependence of resonance frequency and z -direction electric field strength of modes 1 to 3 ($d = 0.19$ mm) (red: mode 1, blue: mode 2, green: mode 3; open circles: resonance frequencies from harmonic analysis)

Therefore, as shown in Figure 7(a), we devised a structure where the outer surface of the choke groove protrudes near the stub, and the stub is further extended. Figure 7 shows the search results including this configuration for the shape parameters of the non-split elliptical choke to maximize the frequency range in which electromagnetic wave leakage can be suppressed. Specifically, we searched for the values of a , b , w_c , w_s , the width of the protruding wall section in the choke (w_{sc}), d , and the distance (d_{sc}) between A and C in Figure 7(a) of the non-split elliptical choke groove when $g = 0.05$ mm, $h = 0.36$ mm, and $\theta = 26.5^\circ$. The harmonic analysis results for the shape in Figure 7(a) showed transmission loss of ≤ 0.1 dB and leakage loss of ≥ 22 dB at 220 to 330 GHz (Figure 7(b)).

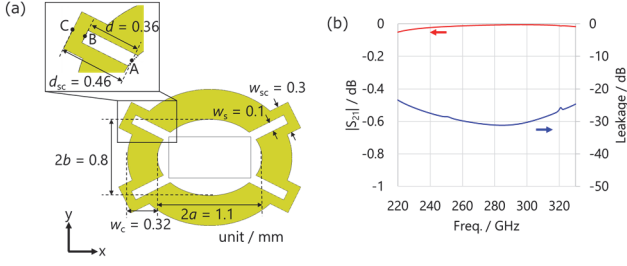


Figure 7 Non-split elliptical choke harmonic analysis results (a) Shape of analysis model, (b) $|S_{21}|$ and leakage

3.2 Split Elliptical Choke

When the stub is extended outside the choke groove, the formed ridge waveguide divides the choke groove. As a result, even if resonance occurs in the gap near the waveguide, resonance should be suppressed by leakage to the outside the choke groove. Therefore, while split elliptical chokes increase leakage compared to non-split elliptical chokes, they are less likely to produce strong resonances that would inhibit wideband operation. Here it is called a split elliptical choke because the choke groove is divided into four parts. We searched for shape parameters a , b , w_c , and w_s that would maximize the frequency range in which electromagnetic wave leakage can be suppressed when $g = 0.05$ mm, $h = 0.36$ mm, and $\theta = 26.5^\circ$. For the shape shown in Figure 8(a), harmonic analysis results with transmission loss of ≤ 0.2 dB and leakage loss of ≥ 14 dB were obtained in the 220 to 330 GHz range (Figure 8(b)).

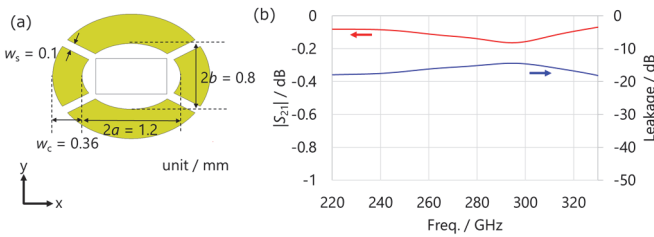


Figure 8 Split elliptical choke harmonic analysis results (a) Shape of analysis model, (b) $|S_{21}|$ and leakage

Despite the expectation that the split portion of the opposing flange surfaces would function as a transmission line, the following points explain why electromagnetic wave leakage loss was suppressed to 14 dB or more across the entire WR-3 range. As shown by the red dotted line in Figure 1(b), the electromagnetic waves leaking into the gap radiate with strong electric fields in the range of approximately $\pm 30^\circ$ from the y-axis when viewed from the center of the waveguide. Therefore, we think that electromagnetic wave leakage is suppressed effectively by having the split elliptical choke

cover the y-axis region where the electric field radiates strongly.

3.3 Flat Elliptical Choke

Obviously as the volume of the resonator becomes smaller, the resonance frequency becomes larger. Therefore, by making the space from the waveguide opening to the choke groove smaller than the simple elliptical choke, the resonance at 324 GHz should shift to the higher frequency side. To achieve this, we designed a flat elliptical choke that shortened the short side of the waveguide cross-section and made the elliptical choke flatter. A shortened waveguide with reduced short sides can be connected to a WR-3 standard waveguide with suppressed reflection using an adapter. The conversion loss at the connection increases as the size difference between the waveguides becomes larger. On the other hand, the resonance frequency increases as the space from the waveguide opening to the inside of the choke groove is narrowed. Considering this tradeoff, we simultaneously explored the shape parameters of the waveguide short side and the flat elliptical choke to design a resonance frequency of ≥ 330 GHz when $g = 0.05$ mm, with a minimum machining width of 0.1 mm. As a result, we obtained a waveguide short side of 0.2 mm and the values of a , b , and w_c shown in Figure 9(a). Analysis results for this shape, gave a transmission loss of ≤ 0.1 dB and a leakage loss of ≥ 25 dB or more in the 220 to 330 GHz range (Figure 9(b)).

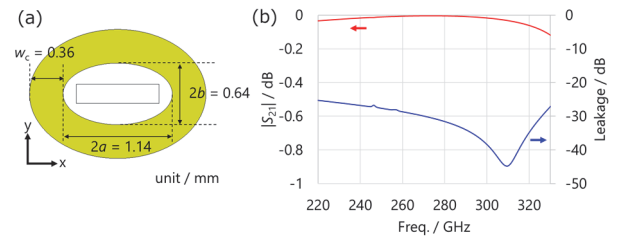


Figure 9 Flat elliptical choke analysis results (a) Shape of analysis model, (b) $|S_{21}|$ and leakage

4 Multiplexing Elliptical Chokes

Figure 10 shows the leakage harmonic analysis results for single, double, triple, and quadruple elliptical chokes. The four types of multiple elliptical chokes are based on the single choke shown in Figures 3 and 7 to 9, and are made into multiple structures by providing choke grooves with the same aspect ratio. The non-split elliptical choke shown in Figure 10(b) has stubs only in the innermost elliptical choke

groove, while the double to quadruple parts have simple elliptical choke grooves. For multiple chokes, the thickness of the wall (w_w) separating the choke grooves was set at 0.12 mm, considering the technical constraints of machining and the strength of the material. It was confirmed that the simple, non-split, and flat quadruple elliptical chokes shown in Figures 10(a), (b), and (d), respectively, have decreased leakage as the number of choke grooves increased. However, in the multiple split-type elliptical choke shown in Figure 10(c), the stubs dividing the choke grooves function as ridge waveguides, and no dependence of leakage on the number of choke grooves was observed, especially above 310 GHz. Additionally, complex resonances occur with multiplication of choke structure.

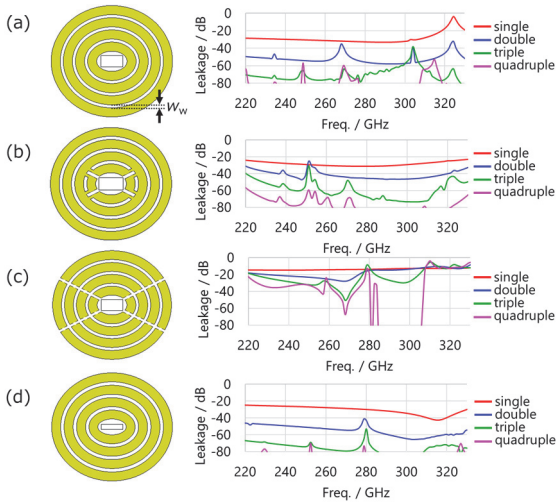


Figure 10 Leakage harmonic analysis results for single, double, triple, and quadruple elliptical chokes: (a) Simple elliptical quadruple choke, (b) Non-split elliptical quadruple choke, (c) Split elliptical quadruple choke, (d) Flat elliptical quadruple choke

5 Elliptical Choke Test Results

This section presents the measured results for simple, non-split, split, and flat elliptical chokes. Based on the multiple-choke harmonic analysis results in the previous chapter, tested (DUT) simple, non-split, and flat elliptical chokes used a quadruple choke structure that secures sufficient leakage loss of 50 dB with only a single-sided choke. On the other hand, the split elliptical choke was made as a single structure. For comparison, a waveguide without a choke was also prepared as a reference DUT.

Each choke was fabricated by machining aluminum alloy

(A6061). Measurements were performed using a DUT formed by connecting a waveguide with a choke and a waveguide with a protrusion, as shown in Figure 11(a). Each corner of the flange surface of the waveguide with the protrusion had a 0.05 mm high protrusion (right side of Figure 11(a)), so the g equals 0.05 mm when connected to a waveguide with or without a choke. The length of each waveguide was set to 59 mm (Figure 11(b)). Measurements were made using a vector network analyzer connected to a WR-3 band frequency extender. The DUT with a flat quadruple elliptical choke was connected to the WR-3 flat flange of the frequency extender through a 4-stage Chebyshev type adapter. DUTs with simple quadruple elliptical choke, non-split quadruple elliptical choke, and split elliptical choke were connected directly.

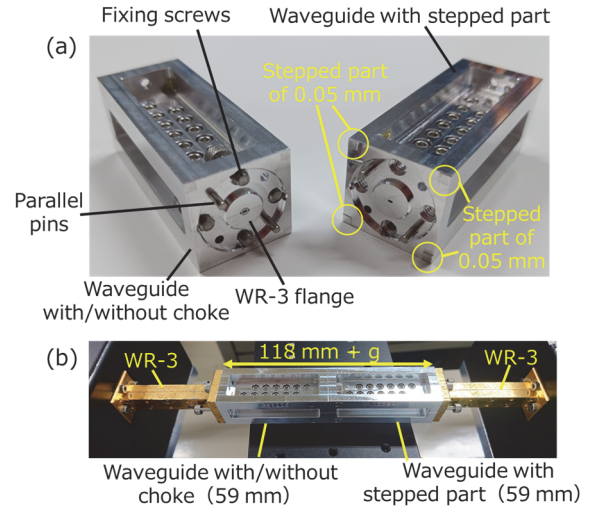


Figure 11 Test Setup (a) External appearance of waveguide with/without choke (left) and waveguide with protrusion (right), (b) Connection of waveguide with/without choke and waveguide with protrusion, and connection to WR-3 flange of frequency extender

Figure 12 shows the $|S_{21}|$ harmonic analysis and measured results for four types of elliptical choke with waveguide connection where $g = 0.05$ mm. The simulation values consider the aluminum alloy conductor loss, and the calculations assume a perfectly smooth surface. The difference between simulated and measured values was 0.5 to 1.2 dB for all elliptical choke types. For example, the difference was 0.8 dB at 275 GHz in the middle of the WR-3 band for the non-split elliptical choke. This difference is thought to be because the analysis did not account for surface roughness.

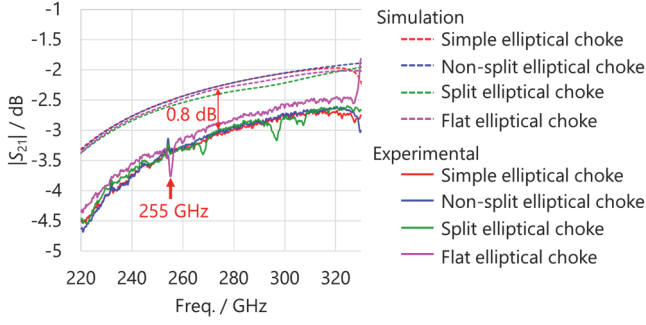


Figure 12 $|S_{21}|$ harmonic analysis and measured results for simple quadruple elliptical choke, non-split quadruple elliptical choke, split elliptical choke (single), and flat quadruple elliptical choke ($g = 0.05$ mm)

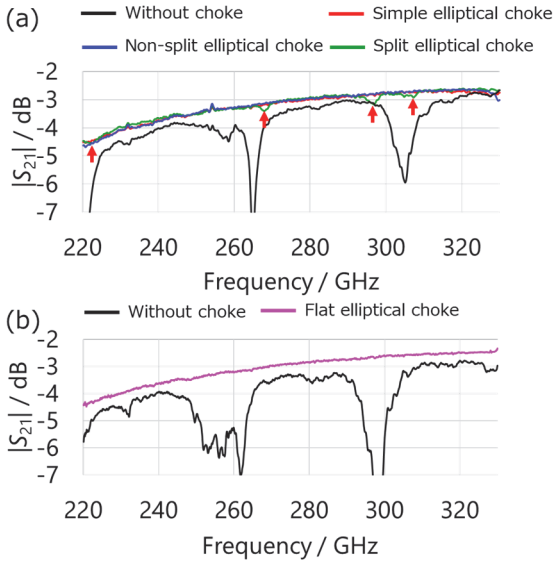


Figure 13 $|S_{21}|$ measured results for DUT and reference DUT ($g = 0.05$ mm) (a) Simple quadruple elliptical choke, non-split quadruple elliptical choke, split elliptical choke (single), (b) Flat quadruple elliptical choke

Figure 13 shows the measured $|S_{21}|$ values when connecting the DUT (simple, non-split, split, flat) to the WR-3 flat flange and elliptical choke waveguide and for the reference DUT connected to the WR-3 flange. An increase in $|S_{21}|$ of about 0.5 to 5 dB compared to the reference DUT (solid black line) was observed across the full WR-3 range in all choke structures, indicating that electromagnetic wave leakage was successfully suppressed.

6 Effect of Displacement Between Opposing Waveguides

Referencing the state when the flange surfaces of two waveguides with propagation axes on the same straight line are directly opposed with a gap of $g = 0.05$ mm, the

displacement (offset) from this position is expressed as $(\Delta x, \Delta y)$ [mm]. Here, Δx and Δy follow the coordinate system shown in Figures 1(a) and (b). The waveguide propagation axes are kept parallel, and angular misalignment is not considered. A displacement of 0.05 mm was applied in each direction, and $|S_{21}|$ was evaluated by harmonic analysis. Figure 14 shows the harmonic analysis results for simple, non-split, flat elliptical quadruple, and split elliptical chokes with intentional displacement in the x and y directions. Table 1 summarizes the resonance frequencies determined from the analysis results in Figure 14. A peak is observed (Figure 12 (solid pink line)) at 255 GHz from the measurement results of the flat elliptical quadruple choke. Since this frequency matches the peak that occurs when a displacement of $\Delta x = 0.05$ mm is applied, this is believed to be due to offset in the x direction.

Despite the harmonic analysis results for the split elliptical choke not exhibiting (Figure 14(c)) clear resonances, the measured results (Figure 13(a) (solid green line)) for the split elliptical choke have four peaks at locations indicated by red arrows. To verify this difference, a harmonic analysis was performed by incorporating the actual flange shape shown in Figure 15(a) into the model. Figure 15(b) shows the result. Figure 15(b) replots the actual measured results from Figure 13(a) (green solid line) for comparison. As shown by the green dotted line, there are peaks at the four places indicated by the red arrows, and these match the actual measured results (green solid line). Therefore, the resonance observed in the split choke is not due to misalignment between the waveguides, but is thought to be caused by reflections at the WR-3 flange end due to the ridge waveguide.

Based on the above, although peaks appear when there is intentional displacement in the x- or y-direction in any of the simple, non-split, and flat chokes, no clear peaks appear in the split choke with any displacement, confirming that the split elliptical choke is less susceptible to misalignment than other elliptical chokes.

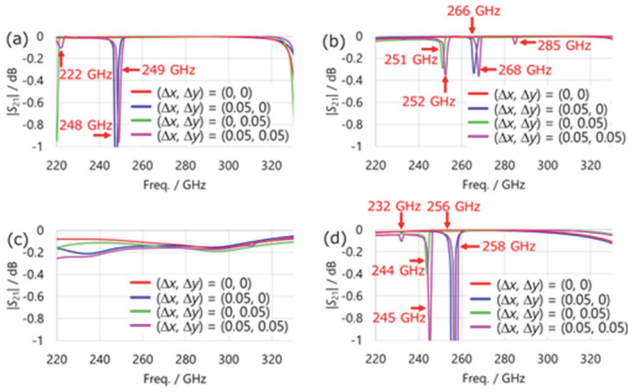


Figure 14 Harmonic analysis results when Δx , Δy displacements of (0.05, 0), (0, 0.05), (0.05, 0.05) mm applied to opposing standard waveguide ($g = 0.05$ mm) (a) Simple elliptical quadruple choke, (b) Non-split elliptical quadruple choke, (c) Split elliptical choke, (d) Flat elliptical quadruple choke

Table 1 Resonance frequencies when Δx , Δy displacements of (0.05, 0), (0, 0.05), (0.05, 0.05) mm applied to opposing standard waveguide ($g = 0.05$ mm) (a) Simple elliptical quadruple choke, (b) Non-split elliptical quadruple choke, (c) Split elliptical choke, (d) Flat elliptical quadruple choke

		unit / GHz	
Δx / mm	Δy / mm	0	0.05
0	0	248	
0.05	0	—	222, 249

		unit / GHz	
Δx / mm	Δy / mm	0	0.05
0	0	266	
0.05	0	251	252, 268, 285

		unit / GHz	
Δx / mm	Δy / mm	0	0.05
0	0	—	—
0.05	0	—	—

		unit / GHz	
Δx / mm	Δy / mm	0	0.05
0	0	256	
0.05	0	244	232, 245, 258

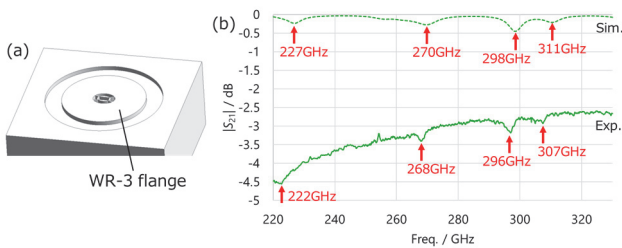


Figure 15 Results of actual measurements of split elliptical choke and harmonic analysis results considering WR-3 flange (a) Shape of analysis model, (b) $|S_{21}|$ measured and harmonic analysis results

7 Conclusions

Chokes are designed originally to perform in narrow bandwidths, and broadening their bandwidth is technically difficult. This article considers four elliptical choke designs (simple, flat, split, and non-split) that operate across the full WR-3 band, and demonstrates their effectiveness through

simulations and actual measurements. Additionally, harmonic analysis comparing changes in transmission loss due to waveguide misalignment confirmed that the split elliptical choke is less susceptible to the effects of waveguide misalignment. However, the split elliptical choke has a larger leakage loss of ≤ 14 dB compared to the other three types of chokes and finding a more effective solution is a task for the future.

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Wideband Millimeter-Wave Complex Permittivity Measurement Technology

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[Summary]

Circuit designs for wideband millimeter-wave signals must properly handle the dispersion properties of dielectric materials comprising the circuit, and technologies to easily and accurately measure the complex permittivity of these materials is urgently needed. Generally, resonance-based systems for measuring complex permittivity are very accurate, but can be cumbersome, are limited to resonance modes suitable for measurement, and require multiple resonators covering a wide bandwidth. This article introduces a complex permittivity measurement system based on the balanced circular disk resonator (BCDR) method facilitating wideband measurement covering 10 to 145 GHz without changing equipment. In addition, we discuss the impact of resonator shape inaccuracies on measurements accuracy.

1 Introduction

Use of the millimeter-wave (mmWave) band is being promoted to secure adequate bandwidth for mobile network communications suffering from limited available frequency resources, and the 300-GHz frequency range is a promising candidate. Circuit design for such high-frequency bands requires basic information on complex permittivity, specifically the relative permittivity (hereafter referred to simply as permittivity) and the dielectric loss tangent. On the other hand, the 60-GHz band is being used for the development of automobile millimeter-wave radar and the complex permittivity of resin materials is being measured to quantify reflection characteristics. High-precision complex permittivity measurements at high frequencies over wide bandwidths are required for various applications.

Figure 1 shows typical permittivity measurement methods. The capacitance method determines the permittivity by measuring the capacitance of a parallel-plate capacitor with the test dielectric material (MUT) sandwiched between electrodes^{1), 2)}. This method is used mainly for low-frequency bands because the resonance between parallel-plate electrodes at higher frequencies cannot be ignored. The transmission method calculates the permittivity by loading the test material on part of a transmission line, such as a waveguide, and measuring its scattering matrix (S-parameters)^{1) to 3)}. Although the transmission method can measure wideband permittivity, the accuracy depends on the transmission line dimensional precision, and gaps between the line and test material cannot be ignored. In particular, dimensional

accuracy requirements approach the limits of manufacturing precision in the millimeter-wave band at wavelengths less than a millimeter. The resonance method calculates the permittivity by loading the test dielectric into a resonator and measuring resonance frequency. The dielectric loss tangent can also be determined from the Q value of the resonance. The measurement frequency of the permittivity is limited to the resonance frequency. Therefore, measuring permittivity over a wide bandwidth requires either making sequential measurements using multiple resonators or using different resonant modes. The resonator requires high mechanical precision to measure the permittivity accurately, and the cavity resonator method is used to accurately evaluate the permittivity of dielectric substrates, etc., at limited frequencies⁴⁾.

However, the Balanced Circular Disk Resonator (BCDR) method can calculate the permittivity from resonant frequencies measured using a symmetrical resonator configured as a sandwich of two outer identical dielectric plates of the test material and an inner conductive circular disk positioned between sufficiently wide parallel conducting plates. As explained in section 2, this method is often used by selectively exciting and utilizing resonant modes where the electric field becomes maximum at the central axis of the disk. Although measurement is discrete, the permittivity can be measured over a wide frequency range by using resonant modes up to about the 10th order. Moreover, while many other methods measure the permittivity for electric fields parallel to the surface of plate-shaped dielectrics, the BCDR

method also measures the permittivity for electric fields perpendicular to the surface. The BCDR method was proposed by Kobayashi in the 1970s and is used widely to measure the permittivity of substrates used by the millimeter-wave band^{5) to 14)}. Recently, it is being used for higher frequency bands and permittivity measurements up to 170 GHz using waveguide coupling have also been conducted¹⁵⁾.

This paper introduces practical examples of permittivity measurements using the BCDR method. Section 2 covers the measurement principle and section 3 describes the constructed measurement system. Section 4 explains some measurement examples and section 5 considers the impact of resonator shape inaccuracies on measured values.

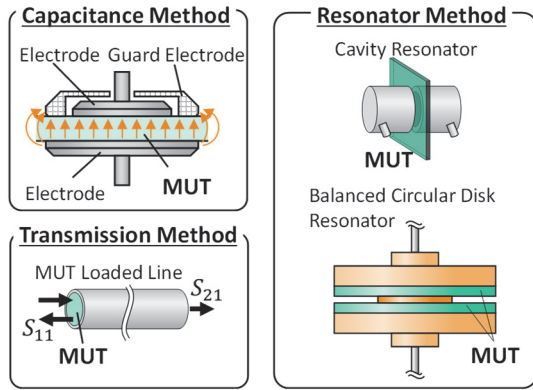


Figure 1 Typical permittivity measurement methods

2 Balanced Circular Disk Resonator Method

2.1 Resonator Structure and Resonance Mode

The (BCDR) method was described by Kobayashi *et al.* in 1974^{5), 6)}. Figure 2 shows the structure. The resonator is configured by layering a conducting flange, the test dielectric material (MUT: material under test), and conducting circular disk. It is excited from an excitation port that is fed by a coaxial line or from the side by a strip line (Figure 2(b)). The crosshatched region sandwiched between the conducting wall and conducting circular disk area (Figure 2(a)) can be considered as a cylindrical resonator. If the thickness of the resonator is limited to less than half the wavelength, the electric field becomes uniform in the thickness direction, and the resonance is limited to TM modes^{5), 13)}. TM modes are identified by the number of nodes in each direction, and are expressed as TM_{nm0} , with subscripts representing the circumferential direction, radial direction, and thickness direction (always 0 since there are no nodes), respectively. The TM_{nm0} resonance frequency is expressed as,

$$f_r = \frac{c}{\sqrt{\epsilon_r}} \frac{x_{nm}}{2\pi R} \quad (1)$$

where, c is the velocity of light in vacuum, ϵ_r is the permittivity, and R is the resonator radius. Also x_{nm} is the eigenvalue corresponding to TM_{nm0} and is determined by the resonators structural parameters and permittivity of the MUT. Therefore, the permittivity of the MUT at that frequency can be derived by substituting the resonator dimensions, resonance mode, and measured resonance frequency into Eq. (1). This requires identifying the resonance mode, which is achieved by selectively exciting only TM_{0m0} modes as described in the next section. Figure 3 shows a calculation example for resonance frequency versus permittivity, when conducting circular disk diameter = 12 mm, dielectric thickness = 0.2 mm, excitation port diameter = 0.9 mm, excitation port depth = 0.3 mm, and conducting disk thickness = 50 μm ¹⁵⁾. The multiple solid lines in the figure represent TM_{0m0} modes ($m = 1, 2, \dots$) from the left. For example, if the permittivity is 3 regardless of frequency, resonance occurs at multiple frequencies indicated by the dots.

As described above, the eigenvalue x_{nm} in Eq. (1) is calculated from the dimensions of each resonator part and the permittivity. The simplest model for the BCDR is the cylindrical region above or below the conducting disk, as shown by the crosshatched region in Figure 2(a), and the eigenvalue is expressed by the root of the derivative of the Bessel function. However, in reality, the electromagnetic field extends outside this cylindrical region and to the excitation port, so the eigenvalue differs from that of a simple cylinder. Since this eigenvalue cannot be obtained analytically, the mode analysis method described in section 2.3 has been reported^{5) to 10), 15)}.

Figure 4(a) shows an example calculation for the BCDR reflection and transmission spectrum using finite element analysis, and Figure 4(b) shows the electric field distribution of each resonant mode. Figure 4(a) shows the case when the excitation port is intentionally set at a position offset from the center of the conducting disk, and there are numerous TM_{nm0} modes other than $n = 0$. The TM_{0m0} modes among these are modes where the electric field becomes maximum at the center of the cylindrical resonator and can be excited by applying an electric field in the thickness direction at the center of the resonator. The circular waveguide described in the next section is used to perform this selective excitation^{11) to 14)}.

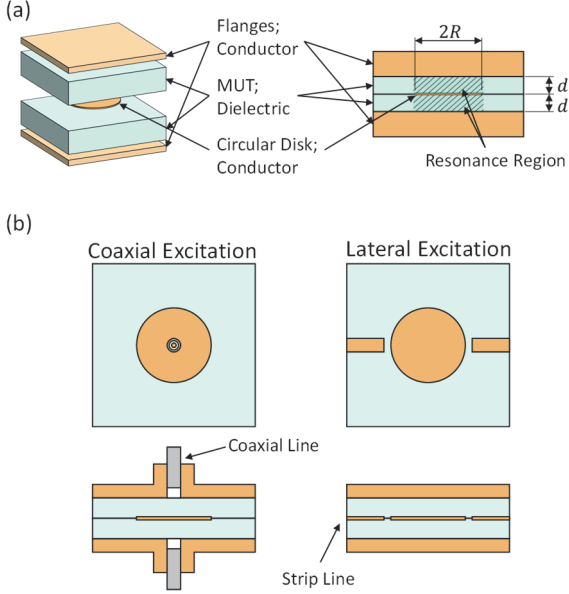


Figure 2 (a) Balanced circular disk resonator, (b) Two types of balanced circular disk resonators with different excitation structures

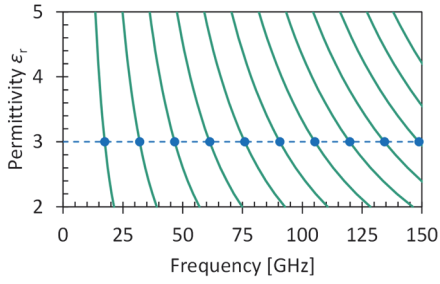


Figure 3 Relationship between resonance frequency and permittivity

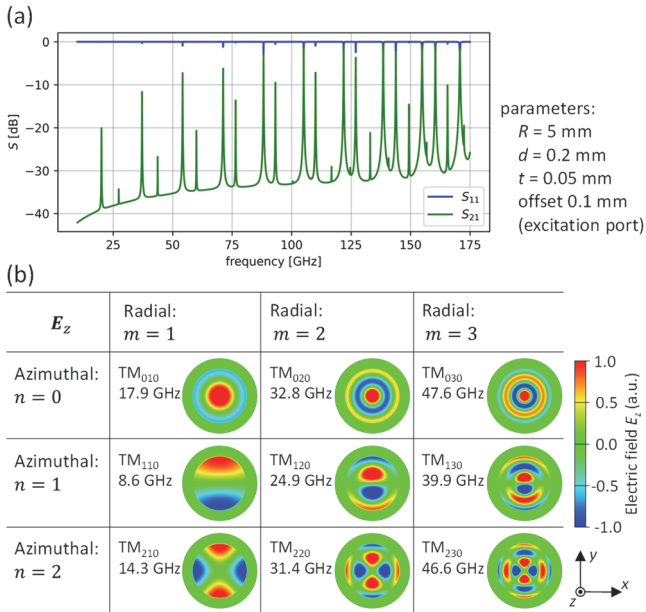


Figure 4 (a) Resonator transmission spectrum S_{21} and reflection spectrum S_{11} , (b) Electric field distribution in thickness direction of TM_{nm0} mode. Here, the thickness direction of the measurement target is defined as the z -axis.

2.2 Excitation Port Shape

An electric field must be applied in the thickness direction of the dielectric at the center of the conductive disk to selectively excite the TM_{0m0} mode. This is achieved by coupling with a thin circular waveguide with a cutoff frequency sufficiently higher than the measurement frequency band¹³⁾. Furthermore, the lateral excitation shown in Figure 2(b) makes it difficult to read the resonance because the Q value decreases compared to coaxial excitation¹⁴⁾. Figure 5(a) shows the coaxial excitation structure. The excitation port is a circular waveguide with a diameter of $2a$ such that the following relationship is satisfied:

$$f_{c, TM_{01}} = \frac{p_{01}c}{2\pi a} \gg f_{\max} \quad (2)$$

where, $f_{c, TM_{01}}$ is the cutoff frequency of the lowest order propagating mode TM_{01} , and $f_{\max}$ is the maximum value of the measurement frequency. Also, p_{01} is the first root of the zeroth-order Bessel function of the first kind, $J_0(x)$ such that $J_0(p_{01}) = 0$. Figure 5(b) shows the geometry of the excitation port and the electric field. A non-propagating field is generated in the cylindrical excitation port, and the BCDR TM_{0m0} mode is excited in this way.

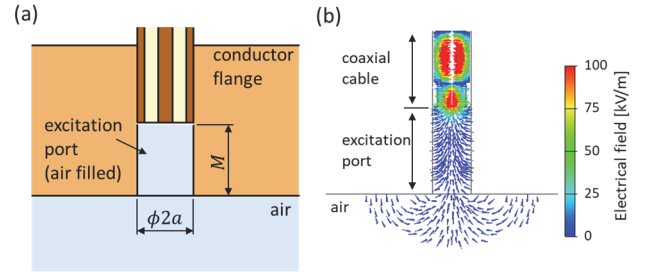


Figure 5 Excitation port and the electric field behavior (a) Port geometry, (b) Electric field

2.3 Relationship between Resonance Frequency and Permittivity

Figure 6(a) shows the BCDR structure. As described in section 2.1, the BCDR can be represented most simply by a cylindrical model like Figure 6(b), with the upper and lower surfaces being conductors and the outer circumferential surface being a magnetic wall. At this time, the eigenvalue in Eq. (1) becomes $x_{nm} = p'_{nm}$. Here, p'_{nm} denotes the m th root of the derivative of the Bessel function of the first-kind and order n , i.e., $J'_n(p'_{nm}) = 0$. However, in reality, the true eigenvalue differs from the simple cylindrical model due to the spread of the electromagnetic field beyond the outer circumferential surface and the spread of the electromagnetic field into the spatial

part of the excitation port. To incorporate these effects, we developed calculation methods for eigenvalues using the models shown in Figures 6(c), (d), and (e). Since these derivations are complex, we leave the details to original papers^{5), 6), 8), 9), 15)} and only briefly describe the general concept here.

As is widely known, the solution to the Helmholtz equation with symmetry around the central axis can be expanded into modes expressed as the product of a Bessel function in the radial direction, a trigonometric function in the circumferential direction, and a trigonometric function in the axial direction. Figure 6(c) is a model considering the spread of the electromagnetic field beyond the cylinder^{5), 6)}. Taking this as an example, in the cylindrical region within radius R (called region I), the upper and lower surfaces are electrical walls; in the outside region (called region II), the lower surface is an electrical wall, and the upper surface is a magnetic wall. These boundary conditions must be satisfied and the electromagnetic field that is continuous across regions I and II becomes the solution for the entire region. Following the general procedure of the mode matching method, for example, in the case of TM_{0m0} , if the electric field in the height (thickness) direction is represented as $\mathbf{E}_z$, then $\mathbf{E}_z$ expressed as a linear combination of modes in region I and $\mathbf{E}_z$ expressed as a linear combination of modes in region II are equated. Similarly, if the magnetic field in the circumferential direction is represented as $\mathbf{H}_\theta$, equations for this can also be obtained in the same way as for $\mathbf{E}_z$. These can be organized with the unknown coefficients of the linear combination (vector) $\mathbf{a}$ and coefficient matrix $\mathbf{M}$, which can be written in the form $\mathbf{M}\mathbf{a} = \mathbf{0}$. For the unknown coefficient vector $\mathbf{a}$ to have solutions other than zero vector $\mathbf{0}$,

$$\det \mathbf{M}(f_{ra}; \varepsilon_{ra}, R, d) = 0 \quad (3)$$

must be satisfied, where f_{ra} is the resonance frequency, R is the radius of the circular conduction disk, and d is the sample thickness. The permittivity ε_{ra} can be found numerically by determining f_{ra} through measurement and substituting R and d into Eq. (3).

Although even more complex, eigenvalues can also be determined using the mode matching method when using the models in Figures 6(d) and (e)^{8), 9), 15)}.

On the other hand, the dielectric loss tangent is calculated from the Q -value of the measured resonance. The unloaded Q -value Q_u , the Q -value of conductor loss Q_c , and the Q -value of dielectric loss Q_d have the following relationship

$$\frac{1}{Q_u} = \frac{1}{Q_c} + \frac{1}{Q_d} \quad (4)$$

The dielectric loss tangent $\tan \delta$ and Q_d have an inverse relationship (Eq. 5))

$$Q_d = \frac{1}{\tan \delta} \quad (5)$$

Conductor loss, Q_c can be calculated from the electromagnetic field distribution^{7), 15)}. Generally, the energy and conductor losses of the electromagnetic field in the spatial region outside the cylinder and in the excitation port are small compared to those inside the cylinder. From this, if we consider a cylindrical model like Figure 6(b) for the TM_{nm0} mode, Q_c can be approximated as follows¹⁶⁾:

$$Q_c \approx d \sqrt{\pi \mu_0 f_r \sigma} = \frac{d}{\delta_s} \quad (6)$$

Here, μ_0 is the permeability of vacuum ($1.26 \times 10^{-6} \text{ N/A}^2$), σ is the conductivity of the conductor surrounding the resonator, and δ_s is the skin depth. Finally, the dielectric loss tangent $\tan \delta$ is determined as follows:

$$\tan \delta = \frac{1}{Q_u} - \frac{1}{d \sqrt{\pi \mu_0 f_r \sigma}} \quad (7)$$

In actual measurements, the coupling between the resonator and excitation port is made sufficiently loose, and the Q -value read from the width of the resonance peak is used as Q_u .

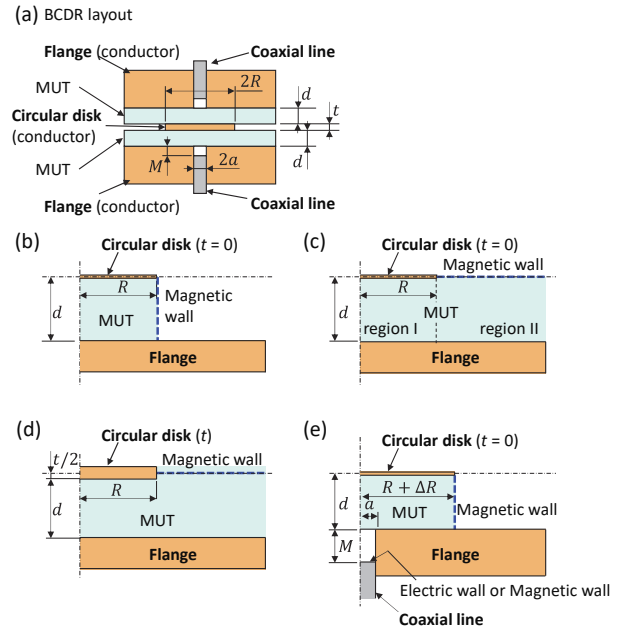


Figure 6 Resonator analysis models (a) BCDR structure, (b) Cylindrical model surrounded by magnetic walls, (c) Model considering spread from edges, (d) Model considering thickness of conductor disk, (e) Model considering influence of excitation port.
MUT: Material Under Test

3 Measurement System Configuration

This study constructed a device to make batch permittivity measurements in the millimeter waveband with an upper limit of 145 GHz. Figures 7(a) and (b) show the configuration and external appearance of the device. The BCDR has 0.8-mm coaxial connectors as input and output ports, which are connected to a wideband vector network analyzer (VNA, ME7838D) with a maximum frequency of 145 GHz. This measurement system eliminates the need to reconfigure or reconnect equipment for each frequency band, making it possible to read a series of resonance peaks at once.

Figure 7(c) shows the resonator configuration. A copper disk is sandwiched between two pieces of MUT that are held between copper flanges. Excitation is via a semi-rigid coaxial cable (155 GHz max. operating frequency, 0.89-mm outer diameter) inserted into two circular excitation ports (0.92-mm inner diameter) in the copper flange, and is loosely coupled to the resonator with a gap, M , of 0.3 mm. The two excitation ports are aligned with the center axis with an accuracy of $\pm 5 \mu\text{m}$. On the other hand, the center position of the copper disk relative to the excitation ports is adjusted along with the two MUT plates sandwiching the copper disk.

The measurement procedure is as follows: ① Calibrate VNA at 0.8-mm coaxial connector at resonator input. ② Measure MUT thickness. ③ Install MUT and copper disk. ④ Apply pressure to conductor flange. ⑤ Capture transmission spectrum (S_{21}). ⑥ Check for higher-order modes and if present, adjust position of copper disk. ⑦ Acquire and analyze spectrum.

Spectrum measurement calibration is performed at the 0.8-mm coaxial connector terminal at the resonator input. To assure accuracy across a wide bandwidth, calibration is divided into two frequency bands. SOLT calibration is used for frequencies below 70 GHz, while offset short calibration was performed and synthesized for frequencies above 70 GHz where load termination matching is difficult.

First-peak detection is performed at spectrum analysis and the resonance frequency and Q -value are determined by fitting a Lorentz function to each peak. The permittivity ϵ_r is obtained by solving Eq. (3) from the obtained resonance frequencies. On the other hand, the dielectric loss tangent δ is obtained by substituting the Q -value into Eq. (4).

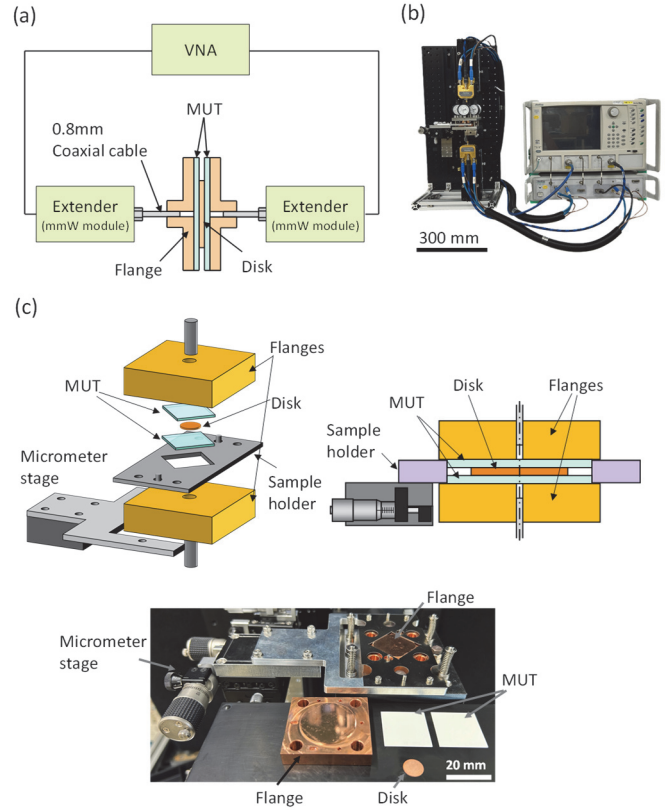


Figure 7 Measurement system configuration (a) Block diagram, (b) External appearance, (c) Resonator structure

4 Measurement Results

4.1 Measurement of Complex Permittivity of Typical MUT

Complex permittivities of typical MUT were measured to verify the operation of the prototype measurement system. The measured MUT samples were cyclo-olefin polymer sheet (CoP, Zeon Corporation, ZF-14, 0.188-mm thick), synthetic quartz plate (Covalent Inc., T-4000, 0.124-mm thick), and millimeter wave composite substrate (Megtron 7, R-5775N, Panasonic Industry Co., Ltd., 0.200-mm thick). A 9.902-mm diameter, 50- μm thick copper conductor disk (cut from 99.99% pure copper foil manufactured by Nilaco Corporation) was used for the CoP sheet and synthetic quartz plate measurements. A 11.994-mm diameter, 12- μm thick copper disk formed from the laminated copper foil on the substrate was used for the millimeter wave composite substrate measurement. Figure 8 shows transmission spectra obtained by measuring each MUT and the complex permittivity analysis results. Almost periodic TM_{0m0} mode resonances are observed up to about 140 GHz in each transmission spectrum. Permittivities shown in Figure 8 represent the mean and the standard deviation measured

after five cycles of mounting and dismounting for each MUT. In addition, results measured at the Kanagawa Institute of Industrial Science and Technology (KISTEC) are also shown for comparison. The standard deviation of the permittivity was within $\pm 0.3\%$ of the mean value across the entire frequency band, and matching KISTEC measured values were within this repeatability range.

4.2 Changes in Resonance Peaks Depending on Conductor Disk Diameter

The measurement frequency upper limit is restricted by the signal-to-noise ratio (SNR) of the transmission spectrum near the resonance peak¹⁴⁾. On the other hand, increasing the BCDR conductor disk diameter can increase both the density of the spacing between resonance peaks and the number of measurement points, but the resonance peaks decrease because higher-order resonances suffer greater losses. Therefore, the conductor disk diameter should be set to the appropriate size to obtain sufficient resonance peak values even at high frequencies, as well as an adequate number of resonance modes across the entire bandwidth. The change in resonance peaks when changing the resonator diameter was confirmed using this measurement setup. The millimeter wave composite substrate was used as the MUT sample, and copper disks (8.990, 11.994, and 14.998-mm diameter, 12- μm thick) were formed using the laminated copper foil on one side of the substrate. Figure 9 shows the measured results. As shown by Eq. (1), the diameter of the conductor disk determines the resonator size, and the spacing between resonance peaks becomes narrower as the resonator size increases. The measured results show a similar trend. The permittivity and dielectric loss tangent derived from each peak frequency are 3.19 ± 0.02 and 0.005 ± 0.002 , respectively, near 75 GHz. These values are consistent with the measured results in section 4.1. When comparing results by diameter, although modes up to $m = 6$ can be read for a resonator diameter of 9 mm, the peak values of the transmission spectrum at mode $m > 6$ decrease for diameters of 12 and 15 mm, suggesting the possibility that peak estimation cannot be performed normally due to the impact of noise. Accordingly, when evaluating samples with unknown complex permittivities, it is necessary to use a moderately small resonator diameter so that each resonance peak value does not decrease too much.

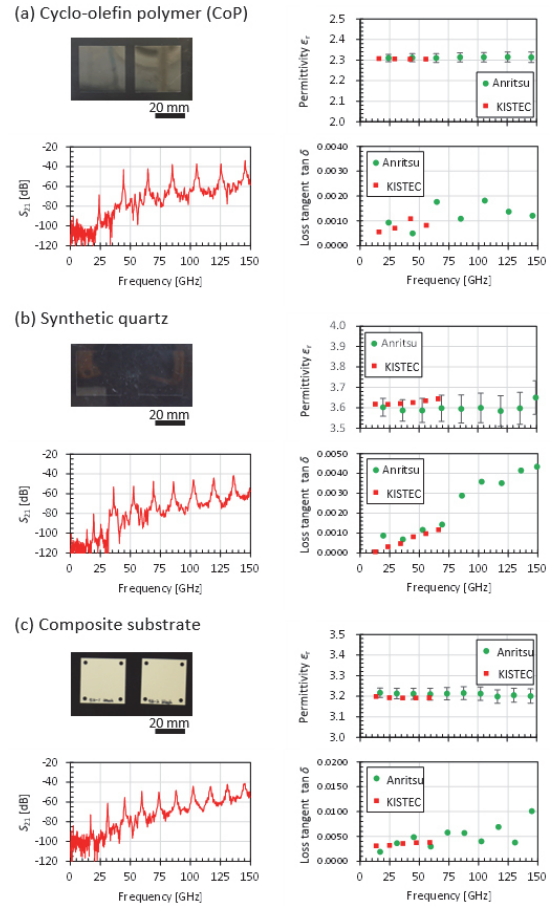


Figure 8 Measured complex permittivity results. (a) Cyclo-olefin polymer (CoP) sheet, (b) Synthetic quartz plate, (c) Millimeter wave composite substrate

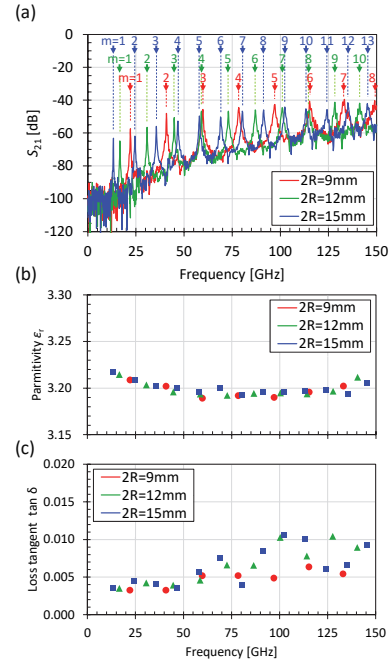


Figure 9 Measured complex permittivity results using circular conductor disks of different diameters for same MUT (a) Resonator transmission spectrum, (b) Estimated permittivity values, (c) Estimated dielectric loss tangent values

5 Impact of Feed Port Position Accuracy on Resonance Modes

The BCDR method determines permittivities from resonance frequencies based on a rigorous model. However, if the model differs from the actual resonator shape, the accuracy of permittivity measurements is limited. Consequently, the resonator mechanical dimensions must be precisely controlled. While several potential dimensional errors can be anticipated, this study examined the positioning accuracy of the excitation ports.

As described in section 2.3, avoiding BCDR excitation modes other than TM_{0m0} requires aligning the center of the excitation port with the center of the conductor disk. This positioning must be done carefully while maintaining accuracy by establishing an external mechanical reference. If the center is misaligned, TM_{nm0} modes (hereafter called high-order TM_{nm0} modes) will be excited, and the magnitude of the excitation port correction described in section 2.3 will probably change. To consider this effect, we verified the impact of positioning accuracy on the BCDR resonance modes experimentally.

Figure 10(a) models BCDR with an eccentric excitation port by reusing the MUT used in section 4.2 (Panasonic Megtron7, 0.200-mm thick). An 11.994-mm diameter, 12- μ m thick copper foil circular pattern formed on the MUT was used as the conductor disk. The center was offset by fine-tuning the moving stage holding the MUT (Figure 7(c)), making the excitation port center eccentric relative to the conductor disk center. We scanned the eccentricity from -250 to 250 μ m in 50 - μ m steps using the mechanical center position calculated from the external dimensions of the resonator as a reference. The mechanical center was set by installing the conductor disk within a tolerance of ± 100 μ m relative to the conductor disk center.

Figure 10(b) shows the transmission spectra when changing the eccentricity. The TM_{0m0} mode resonances (black arrows in figure) appeared when the eccentricity was within 100 μ m. On the other hand, above 100 μ m, resonance peaks due to higher-order TM_{nm0} modes appeared at the positions indicated by blue triangles in the figure. Mode identification is impossible if the higher-order TM_{nm0} mode peaks cannot be distinguished from the TM_{0m0} modes. Furthermore, the intensity decreases for TM_{0m0} modes, and the resonance

frequency shifts for the 5th order and above. If the peak intensity decreases, the measured Q-value is lower, and the measured dielectric tangent is higher. Also, the resonance frequency shift causes an offset in the measured permittivity. From the characteristics of the above spectra, precise positioning of the excitation port center is important not only to avoid exciting unintended higher-order TM_{nm0} modes, but also to ensure that the resonance conditions of the fundamental vibration mode TM_{0m0} match the strict model.

Figure 10(c) shows the measured resonance frequencies for TM_{020} , TM_{050} , and TM_{080} . The top panel in Figure 10(c) shows the changes in resonance frequency with respect to the eccentricity of the conductor disk; the bottom panel shows the permittivity calculated from the resonance frequency. The resonance frequency peaked at an eccentricity of -50 μ m from the center, and the characteristic frequency shift toward the lower frequency side occurred as the absolute eccentricity increased from that position. In this case, as shown in the bottom panel of Figure 10(c), the calculated permittivity increases as the eccentricity moves away from -50 μ m. The measured permittivity is lowest at all resonance peaks when the eccentricity is -50 μ m, and it is thought that this position corresponds to the electrical center of the resonator. Also, when comparing TM_{020} , TM_{050} , and TM_{080} , the higher the resonance frequency, the greater the impact on the measured permittivity with increasing eccentricity. Achieving an accuracy of $\pm 0.5\%$ under the current permittivity measurement conditions requires positioning the excitation port within ± 0.15 mm of the resonator center.

Based on above-observed changes in resonance high-resolution, positioning adjustment of the excitation port is possible using the following procedure. First, adjust so that the higher-order TM_{nm0} modes are sufficiently small and only the TM_{0m0} modes are excited. Next, monitor the resonance frequency of the higher-order TM_{0m0} modes and adjust to maximize the resonance frequency. By using this procedure, positioning can achieve an accuracy of $\geq \pm 50$ μ m for the MUT and resonator used in this study, and the resulting uncertainty in the permittivity can be expected to be $\leq \pm 0.1\%$.

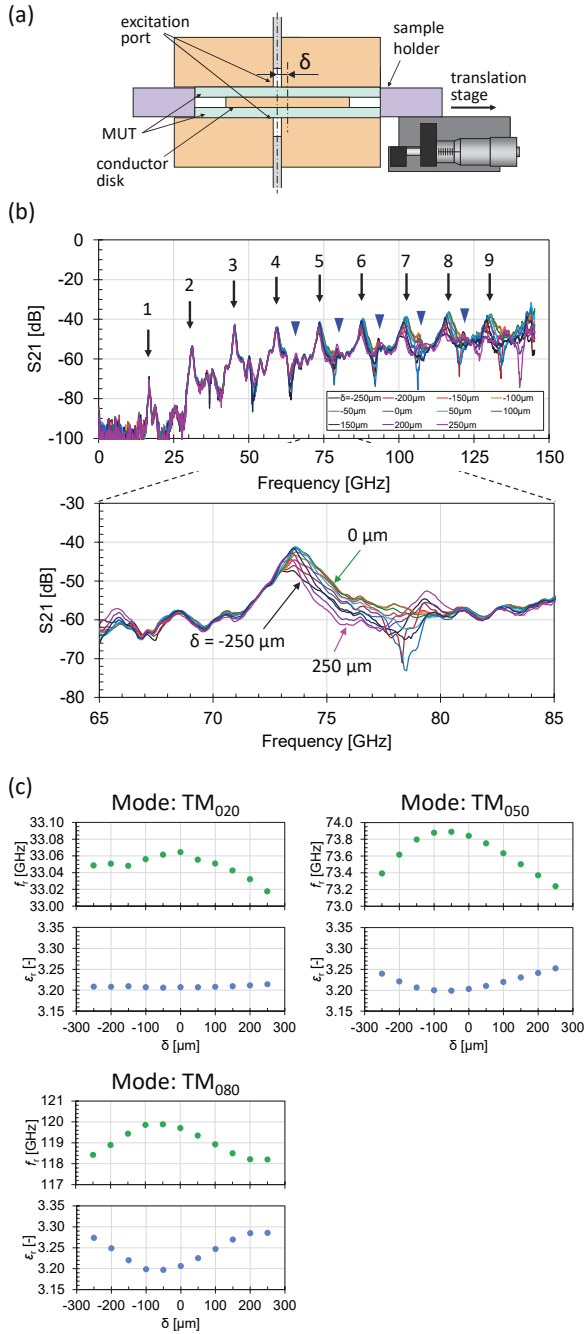


Figure 10 (a) Excitation of modes other than basic resonance mode due to eccentricity of excitation port, (b) Transition of resonance frequencies of basic resonance modes, where blue triangles indicate resonance peaks of higher-order TM_{nm0} modes, (c) Changes in resonance frequency and estimated permittivity due to eccentricity of conductor disk center relative to excitation port. TM_{020} (fr ~32 GHz), TM_{050} (~76 GHz), and TM_{080} (~120 GHz) modes shown as examples.

6 Conclusions

We built a system to measure complex permittivity using Anritsu's wideband VNA to take advantage of the inherent

wideband nature of the BCDR method. Generally, the measurement frequency range of the BCDR method is determined by the resonator size, excitation port size, and VNA measurement bandwidth, and this study measured permittivity up to 145 GHz. To our knowledge, this is the widest bandwidth implementation of permittivity measurement that does not require reassembly or reconnection of equipment for each frequency band. Furthermore, we used this measurement system to experimentally demonstrate the effect of the relative position of the excitation port center axis and the conductor disk on measurement accuracy, and proposed a method for adjusting their relative positions. In addition to describing wide bandwidth, accuracy, high maximum measurable frequency, and simplicity, we describe general requirements for a permittivity measurement system. Future research on the BCDR method is expected to extend measurement to higher frequency bands beyond 300 GHz by scaling down the excitation port etc., to develop more accurate and simpler measurement system.

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Pushing the limits: The Future of Wideband Vector Signal Generators

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[Summary]

As the wireless technology evolves, so does the need for more powerful and precise testing tools in the world of communications. That demands higher data rates, sensing and complex processing. Enter the future of wideband vector signal generators (VSGs) – the unsung heroes of signal generation that are pushing the boundaries of processing speed, bandwidth, and accuracy. With advancements in digital signal processing, VSGs are not only handling wider frequency ranges but also accommodating the increasingly complex modulation schemes of modern wireless systems. From 5G and beyond to radar and satellite communications, superior frequency hopping, these cutting-edge devices are redefining the possibilities, enabling engineers to stay ahead in an ever-demanding tech landscape.

1 Introduction

The demand for wideband vector signal generators is indisputable. This is because of unique demands, like generating multiple communication systems simultaneously, simulating complex environments, enabling high-bandwidth telecommunication, superior frequency hopping patterns, and higher-resolution radars.

The main challenge lies in developing new fundamental technologies to enable such an instrument.

In this paper, we present technologies and techniques including Digital communication links, DSP, Processing speed, Calibration techniques, Digital communication standards, Powerful FPGAs, high-speed and high-resolution DACs, small node CMOS semiconductors, synchronization techniques, and phase and magnitude stability.

2 Technology

A system that can generate 10 GSymbol/sec with a 16-bit resolution with real-time DSP processing power for filtering, equalization, and power control is demonstrated in this article.

At the core of every advanced wideband vector signal generator lies a high-speed DAC architecture¹⁾. There are multiple ways to achieve high-speed DACs with high resolution (16 bits).

- (1) Single DACs, require smaller transistor technology to reduce the parasitic capacitance effect. This comes also with lower power.
- (2) Multichannel DACs, using external I/Q modulators which feature more than double the speed using the

same DAC technology. This method requires complex synchronization and calibration.

- (3) Hybrid DAC architectures combining different DAC technologies better suit different frequencies and power requirements.

A high-speed digital circuit is required to stream the generated data. JESD204C can achieve speeds up to 32 Gbps/lane. PCIe 5 and PCIe 6 support 32 GTransfer/s and 64 GT/s²⁾. PCIe 5 can reach up to 64 GBytes/s for an x16 connection. Although PCIe is not directly used with DACs it can serve as a backend for streaming. Multichannel architecture is scalable and provides more bandwidth than single channel.

FPGAs are widely used for high-speed DAC interfaces, but the target applications of wideband vector signal generators spike the demand for powerful DSPs. This demand puts significant pressure on offloading well-defined tasks such as interpolation from FPGA to ASIC chips often integrated into the DAC chip. Another opportunity in the future is to integrate an AI model to regenerate samples instead of interpolation to improve signal purity.

It must be mentioned that for streaming signals using OFDM or similar complex modulation schemes, the IFFT functionality must be offloaded from the OS to meet timing criteria. For Beam steering and MIMO applications, a multichannel phase coherent VSG is required. From the RF point of view, a wideband vector signal generator with a wide tuning range from a few MHz to mm-wave frequencies poses a challenge in filtering and maintaining signal purity. New techniques must be developed to reduce unwanted spurs,

images, or harmonics. To this end, active calibration and adjustment can be implemented with feedback to keep the device running in an optimal condition. All this must be done while keeping magnitude and phase stable with temperature drift. Filtering technology must be also pushed to the limits for mmwave frequencies.

Developing hardware that can work on mm-wave frequencies at a lower cost that meets requirements can be challenging.

An example of a 60 GHz filter mounted on a PCB in Figure 1 demonstrates a BPF from 50-60 GHz mounted on a PCB using chip-on-board technology.

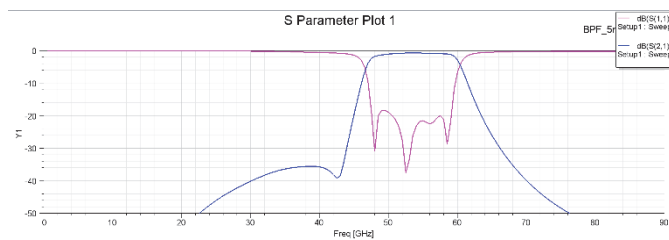
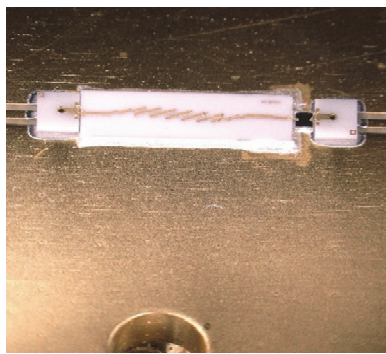


Figure 1 Anritsu BPF 50-60 GHz on a 5 mil Substrate on PCB

3 Calibration

Starting with baseband, multichannel high-speed DACs must be synchronized and calibrated for magnitude and phase.

This is also a key parameter for multi-channel VSGs to test multiple applications including MIMO.

I/Q imbalance is a key parameter that must be minimized, especially for wide bandwidth communications. An accurate and efficient method to measure I/Q Imbalance calibration across a wide band is to generate a single sideband multitone signal and evaluate the image rejection. Figure 2 shows performance before self-correction. It must be noted that this kind of correction is done internally in the instrument to maintain repeatability and preserve performance. This technique relies fundamentally on calibrated traceable power sensors internal to the instrument.



Figure 2 Single-side multitone generated signal showing 20 dBc image rejection before Calibration.

After correction, a multitone image is improved by more than 15 dB. It is also demonstrated that at 130 GHz a 40 dB of image rejection can be maintained for wideband signals.

This measurement has a strong correlation with EVM on a single carrier.

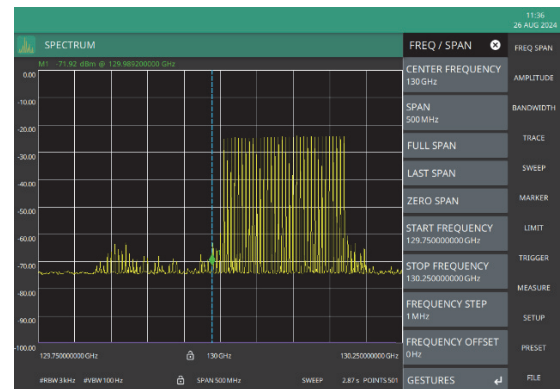
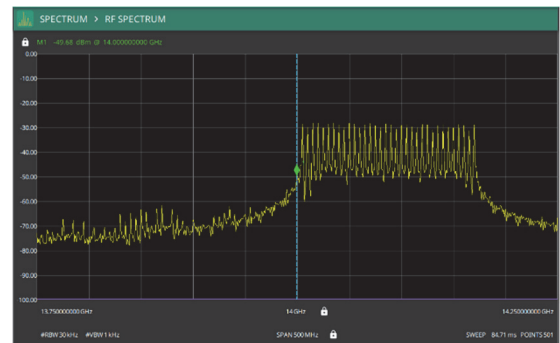


Figure 3 a) Image rejection improvement at 14 GHz, b) Image rejection at 130 GHz

Amplitude flatness is another critical factor in generating vector-modulated signals. It has a direct impact on a single-carrier modulated signals. Ideally, signals at different frequencies must be at the same power levels to reduce the DSP power and improve dynamic range. Figure 4 demonstrates a 10 GHz signal before and after flatness correction.

It must be noted that adjusting power levels requires a continuous adjustment of flatness correction, this makes it

challenging to do preprocessing ahead of time. Hence, the demand for DSP flatness correction for instant adjustment is necessary.

The calibration techniques we utilized enable utilizing VSGs to evaluate high-end receivers. Specifically at multiple TX planes and frequencies the VSG is calibrated to test the receiver at different stages. This provides improved overall calibration and performance understanding of the receiver.

Receiver testing is frequently limited by the quality of the transmitter, even loopback tests are limited as it may not be clear whether the TX or RX is the one limiting the performance. This TX plane calibration approach separates the TX from the RX performance.

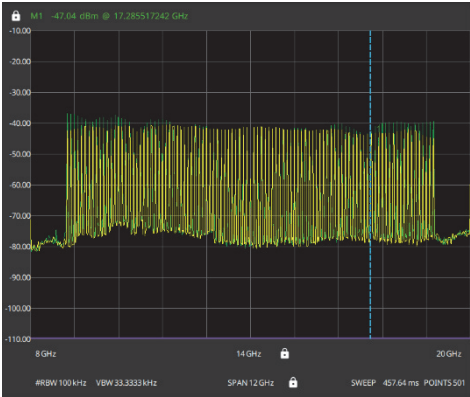


Figure 4 10 GHz signal before correction (green) and after correction (yellow)

4 Measurement

The quality of EVM on a single carrier signal demonstrates the performance of the hardware of the device and its calibration³⁾. Testing using OFDM signals will not show the flaws of magnitude distortion of the instrument. This is because OFDM is immune to this distortion as it relies on multiple narrow-band carriers. On the other hand, OFDM can test the dynamic range.

Figure 5 demonstrates a 64 QAM single carrier EVM performance at 42 GHz with 5 GSps occupying 6.7 GHz of analog BW.

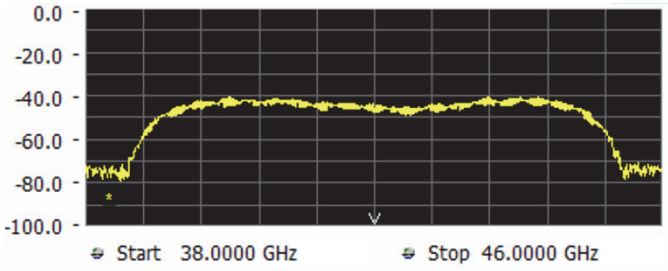
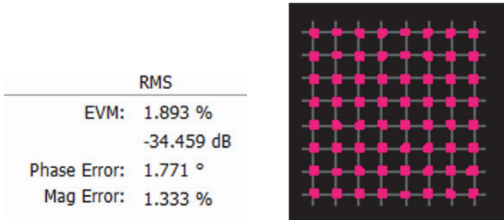


Figure 5 single carrier 64 QAM, 5 GSps at 42 GHz.

At 145 GHz, a signal quality of EVM=3.4% is demonstrated using a loopback measurement. From the constellation, it is evident that the degradation is dominated by the signal-to-noise ratio in Figure 6. This measurement includes both transmitter and receiver performance.

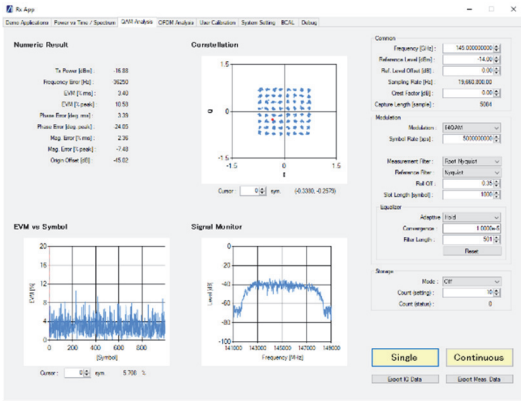


Figure 6 EVM measurement for multiple BWs and carrier frequencies

Upon measuring signal quality at different BWs and carrier frequencies it is more evident that pushing the limits of multi-GHz of modulation BW can be achieved at higher-order modulation schemes, Table 1 demonstrates true device performance using single carrier test signals at frequencies across multiple frequency points.

Table 1 EVM measurement for multiple BWs and carrier frequencies.

Configuration	EVM
64 QAM, 2.5 GSps @ 14 GHz	0.9%
64 QAM, 5 GSps @ 14 GHz	1.7%
64 QAM, 10 GSps @ 14 GHz	2.2%
64 QAM, 5 GSps @ 42 GHz	1.9%
64 QAM, 5 GSps @ 145 GHz	3.4%

5 Conclusions

This paper demonstrates the world-class performance of a wideband vector signal generator. The data shows

exceptional accuracy (16 bit), low EVM, and seamless operation across broad frequency ranges (from MHz to SubTHz).

By achieving such high levels of accuracy and fidelity, we are not only advancing the capabilities of current communication systems but also setting the stage for future innovations. This represents a significant leap forward in technology, showcasing the potential to drive further breakthroughs in a wide range of high-performance applications.

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Sub-THz MMIC Switches Key Enablers for Sub-THz Wireless Systems

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[Summary]

The development of wideband MMIC switches that operate from DC to Sub-THz frequencies represents a significant leap forward in maximizing the potential of the electromagnetic spectrum. The ability to seamlessly cover such a broad frequency range is crucial for a variety of advanced applications, as it allows for greater flexibility and efficiency in Sub-THz communication systems: imaging, sensing, and instrumentation. Miniaturizing and integrating multiple components, including the switch, onto a single compact chip enhances the overall system performance, enabling more versatile and reliable Sub-THz solutions. DC to Sub-THz switches are central to overcoming the challenges of broad-spectrum operation, in particular improvements in insertion loss, isolation, and bandwidth are key. These advancements are set to unlock new possibilities and propel the next generation of RF technologies forward.

1 Introduction

This report presents detailed measurements of a Sub-THz MMIC switch operating from DC to 145 GHz. The ability to achieve such broad-bandwidth operation is crucial for a variety of advanced applications, including high-speed communication systems, precision imaging, sensing, and sophisticated instrumentation. As the demand for faster data transmission and higher-resolution imaging continues to grow, the development of switches that can seamlessly operate from DC to Sub-THz spectrum becomes increasingly important. These wideband switches enable more versatile, compact, and efficient solutions, paving the way for the next generation of RF and microwave technologies. The following measurements demonstrate the key performance parameters of this Sub-THz switch and underscore its potential impact on the future of RF applications.

2 Technology

Wideband switch design faces trade-offs in technology selection, with GaAs PIN diodes offering low insertion loss but limited integration capabilities, and InP HBT-based switches providing high frequency and linearity advantages at the cost of isolation and power consumption. pHEMT technologies also offer integration options, but not for Sub-THz frequencies. This report focuses on using InP DHBT-based switches for integration with amplifiers and multipliers, utilizing a multi-metal layer structure to improve isolation and manage RF and DC components.

3 Design

A Single-pole-single-throw (SPST) switch from DC to 145 GHz is designed in InP HBT-based technology. The switch features either reflective or absorptive off-state.

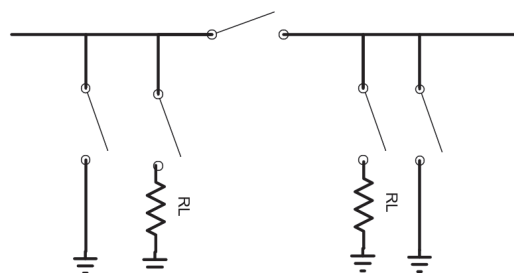


Figure 1 Representation of an SPST reflective, absorptive Sub-THz switch.

4 Measurements

Upon visual inspection of the chip layout, it was found that the top layer doesn't have visual defects. However, the very bottom metal layer had some yield issues that were noticed on multiple chips. This caused some switching states to either not function at all or poorly connect. This matter is still under investigation but the diverse design of multiple switches on the same chip shown in Figure 2 was able to point out some manufacturing flaws. It was found that the same layout implementation showed different results on the same chip.

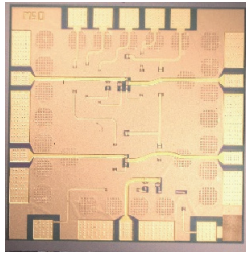


Figure 2 1x1 mm multi-switch chip module.

Some poor connections were visible from the top layer, those sections were discarded from this measurement.

Another section of the switch had minimal reliance on the bottom layer which resulted in consistent results. Dies with manufacturing flaws were not tested.

Figure 3 shows the section of the absorptive and reflective terminations using the same size transistors. Each transistor in Figure 3 was biased and tested individually to compare current dissipation with simulated values.

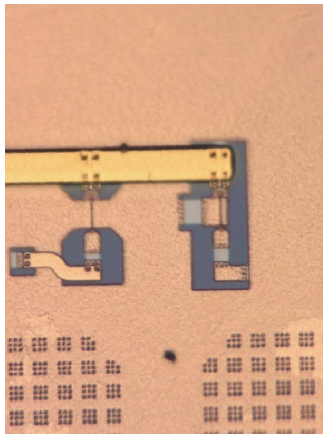


Figure 3 Absorptive and reflective termination.

Biasing the switch to the designed operation points showed the correct current and power consumption.

A TRL calibration was used to calibrate the probes across the band of measurement. The switch is probed for s-parameters in Figure 4 minimizing insertion loss in Sub-THz switches is crucial for maximizing the performance and range of these systems, particularly in applications where signal strength is a key factor for successful data transmission and reception. Two different switches were tested for insertion and return loss. In Figure 5 we have a consistent result for insertion loss for the Thru state when comparing two different switches. The Isolation measurement for both reflective and absorptive states is shown to maintain isolation of 15 dB below across the board. It was evident that the reflective state had 1 dB better isolation for the same bias conditions.

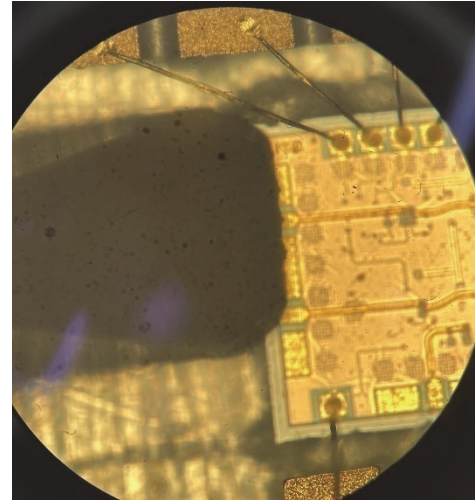


Figure 4 Biasing and probing the switch from DC to 145 GHz.

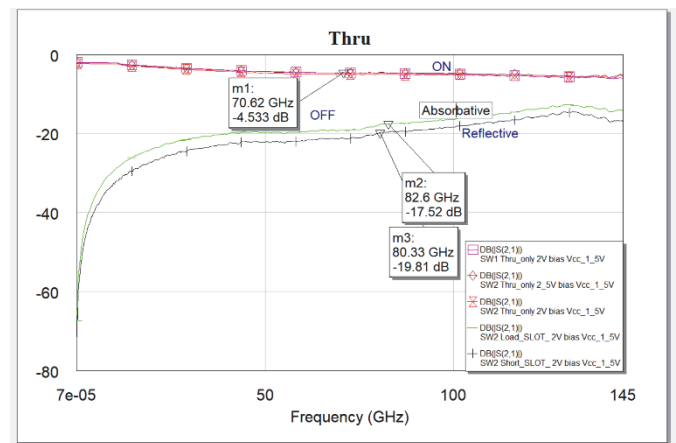


Figure 5 Insertion loss measurement (cyan, brown for 2 different switches) and Isolation (green and black for absorptive and reflective states).

Figure 6 demonstrates Thru states for multiple switches and different bias conditions. increasing the bias voltage by 0.5 volts improved the insertion loss by 0.5 dB.

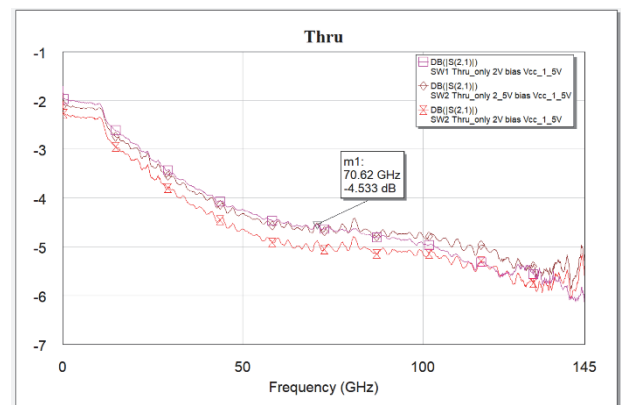


Figure 6 Insertion loss measurement at different bias conditions and different switches.

Finally, In the context of switches, especially at high frequencies like Sub-THz or microwave bands, good return loss ensures that the switch's ports are well-matched to the system's characteristic impedance (often 50 ohms), minimizing reflections and preventing power from being reflected into the source. Return loss is maintained lower than 14 dB across the band and better than 20 dB above 80 GHz Figure 7.

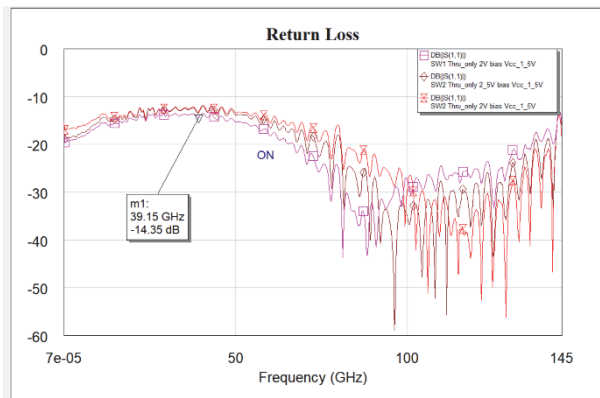


Figure 7 Return loss for ON state.

Importantly, this approach was successful in maintaining a steady insertion loss without ripples or resonances from DC to 145 GHz.

In general, designing a switch using HBT is challenging because of leakage and challenging bias conditions. This report sets the stage for the future of integrating Sub-THz switches on a chip using InP HBT technology.

5 Conclusions

This article presents the design and fabrication of an RF Sub-THz MMIC switch using InP DHBT, which produced promising results. This approach showcased significant potential for miniaturized, advanced RF systems across the DC to Sub-THz spectrum. The unique properties of InP DHBT, such as high electron mobility and thermal conductivity, enhance performance, making it a strong candidate for advanced telecommunications. This approach demonstrated that having an ultrawideband switch is possible using InP DHBT technology. While InP DHBT may not be the optimal material for isolation, this limitation adds complexity to the design process and may affect the switch's maximum operational frequency. However, the increased power handling capability of DHBT, compared to HEMT, strengthens its potential for use in front-end ultrawideband transmitters.

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General Articles

- Development of DX Solution AccelVision
- Mobile Device Field Tests with Excellent Reproducibility
- Development of MT8862A IEEE 802.11be 320-MHz Bandwidth Option
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- Development of High-Performance OFDR System with Phase Integration Coherence Compensation
- Development of World-First 100% Tablet Inspector using NIR Transmission Spectroscopy
- Electrical Performance Enhancements of Fixed Waveguide Attenuators
- Millimeter-wave Optical Network Analysis

Development of DX Solution AccelVision

Tomohiro Wakita, Yuta Shinchi

[Summary]

Declining productivity due to labor shortages caused by falling birth rates and an aging population has become a serious problem in industrial sectors, especially manufacturing. The AccelVision industrial DX solution developed to address these challenges aims to improve production efficiency by advancing labor-saving measures through two-way communications, external system integration, and flexible adaptation to existing equipment. Furthermore, integration with video management systems visualizes events to provide intuitive understanding and support fast and accurate responses. Additionally, collaboration with partner companies is expanding solutions tailored to customers' requirements, in addition to safety management and network monitoring. This article introduces the features of AccelVision, and development of Estinargy, the edge platform device playing a key role in the solution.

1 Introduction

Recent falling birth rates and an aging population have led to serious productivity issues due to labor shortages in various industries. Industrial digital transformation (DX) using digital technology to transform business operations has become a focus of attention to solve these challenges. DX aims to improve production efficiency by saving labor through automation of data collection and analysis, replacing traditional manual processes. However, DX implementation remains slow with causative factors including the difficulty of introduction into existing facilities with mixed equipment types and the lack of resources and appropriate tools for quickly analyzing vast amounts of data.

We have developed the AccelVision industrial DX solution to increase productivity by saving labor through reduced time losses at situation assessment and on-site instructions when resolving problems, as well as cutting data analysis workloads, particularly in manufacturing and other industries. As described in section 2, AccelVision addresses these challenges by enhancing two-way communications and facilitating rapid response to production equipment troubles as well as efficient data collection and analysis. Additionally, it helps build flexible systems that can adapt to existing facilities, reducing implementation barriers by offering optimal proposals tailored to customers' environments. Furthermore, we aim to provide sustainable support for DX promotion by offering solutions that leverage our long experience in network bandwidth control technology and expertise across

product development, maintenance, and general support.

The AccelVision design adopts a structure combining partner components to complement core key components, enabling easy implementation and versatile expandability (Figure 1). Furthermore, using the Estinargy edge platform device described later configured with packaged software offers optimal solutions tailored to each customer's unique challenges (Figure 2).

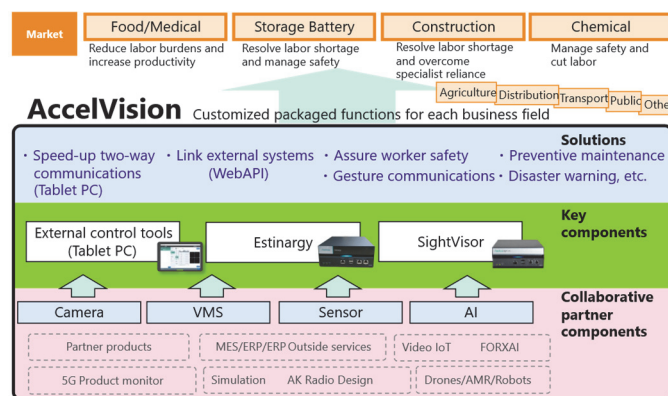


Figure 1 AccelVision configuration

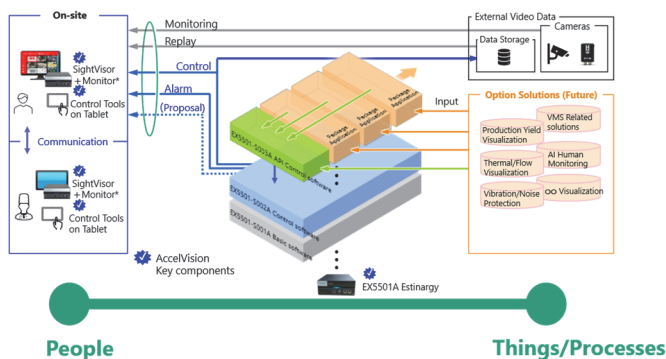


Figure 2 Estinargy configuration

2 AccelVision Features

2.1 Faster Two-Way Communications

Supervisors in production management offices on factory production lines plan, monitor, and adjust production for efficient production and controlled delivery times, and costs. However, when a line problem occurs, supervisors go directly to the site to understand the situation and provide advice, which slows response and line recovery times.

By using AccelVision to share on-site video and production status between supervisors and line operators, instructions and completion notifications can be controlled remotely. This two-way communication improves operation efficiency and cuts line downtime (Figure 3).

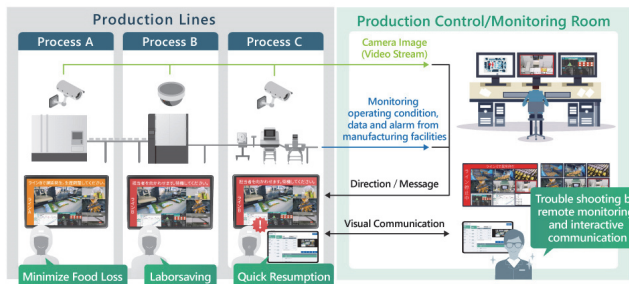


Figure 3 Two-way communications configuration

2.2 Faster Troubleshooting Response to Production Line Equipment Problems

When line equipment has a problem, contact control (DI/DO) is used commonly to notify external systems (such as SCADA) monitoring control and collecting data, as well as to display warnings alerts to workers. However, such equipment does not help understand the failure details, and there are issues with visually confirming events that have occurred already.

AccelVision uses a distributed remote monitoring system to remotely monitor and control equipment such as emergency power supplies and power distribution equipment. It can receive contact signals, add details such as problem descriptions and occurrence times, and display warning information, and then transmit this information to external systems through a Web API. Not only does it control warning displays similar to existing systems, but it also integrates with a Video Management System (VMS) to monitor video from before and after an event occurs. Visualizing events helps more intuitive understanding and enables quick and accurate responses (Figure 4).

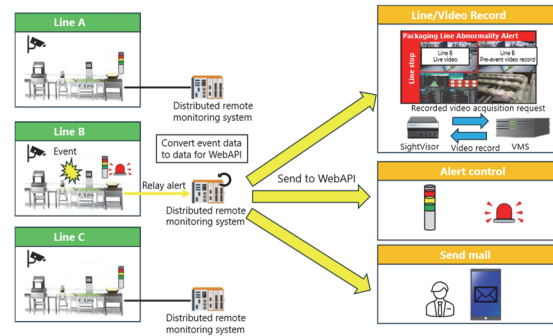


Figure 4 External system integration

2.3 Efficient Data Collection and Analysis

Brief “micro-stops” occur routinely on production lines and are often not recorded by the system because they are deemed unlikely to lead to serious issues. Additionally, when the amount of information about the problem is extensive, background searches are time-consuming and labor-intensive, and it’s difficult to understand the situation details from the trouble description alone.

AccelVision collect all information from both simple problems like micro-stops to major issues and this information can be grouped by production line, equipment, or importance level for reviewing.

Furthermore, AccelVision maps events, allow quick location identification, and because it can replay video from before and after the event to confirm details at the time, quick and accurate cause analysis is possible (Figure 5).

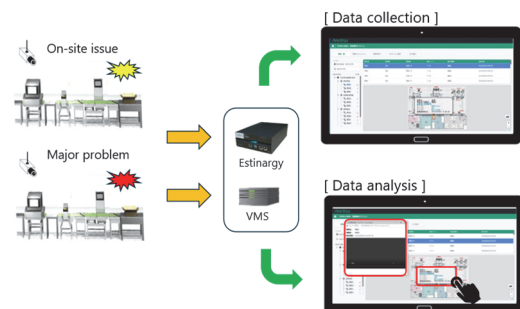


Figure 5 Data collection and analysis

3 Estinergy Development

3.1 Objectives

Implementing the features described in sections 2.1, 2.2, and 2.3 requires a relay function to receive notifications from external devices and perform corresponding controls. Additionally, continuous business improvement requires centralized management of event notifications with links to video data that administrators can easily check and analyze at any time.

We developed the ‘Estinargy’ edge platform device forming the core of AccelVision to implement the required functions (Table 1).

Estinargy is an all-in-one hardware device that collaborates with line equipment and external devices, centralizes management of event data, and links with recorded video. It plays a key role in supporting efficient management of event data combined with on-site video to enable a rapid response.



Figure 6 Estinargy Unit

3.2 Software Package Management and Expandability

To respond flexibly to individual customer’s challenges, we adopted a package management system for software delivery that separates object structures for each function. This supports provision of software combinations with the necessary functions matching customer needs while functionality can be expanded easily by adding features via the WebUI (Table 2).

3.3 Flexible API Provisioning

Achieving integration with external systems requires an API that is both easy to use and versatile. Estinargy is based on REST API, which is commonly used for data transfer between different applications.

The REST API is a form of Web API, which is a simple and scalable architecture based on the HTTP protocol. Resource operations use HTTP technologies (GET, POST, PUT, DELETE).

Since data from external devices varies widely, it is necessary to standardize data formats and consider reserve fields to minimize data types. Estinargy facilitates integration by limiting HTTP methods to POST and restricting data formats to one JSON (JavaScript Object Notation) type. JSON is a lightweight data exchange format mainly used by web applications. It offers excellent extensibility and readability, enabling a versatile data format that includes event, video, and device control information, as well as reserve areas for future expansion.

By disclosing the data format to partner companies and system integrators (SI), we’ve created a system that can receive arbitrary data from other makers’ line equipment and

sensors. This supports flexible and extensible collaboration combining AccelVision with other companies’ systems, expanding the range of solution offerings.

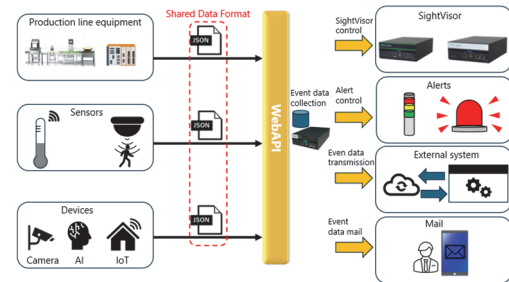


Figure 7 Integration using Web API

3.4 WebUI Provision using Blazor

Package software consists of a web application (front-end) providing a Web user interface (WebUI) and applications that operate inside the device (backend) for data processing and other functions.

Development of the Estinargy WebUI implemented the front-end using Blazor, an open-source web framework for developing web applications using C#, HTML, and CSS based on .NET, which is described later.

Traditional web application development implements the client side primarily in JavaScript and the server side in other languages (PHP, Java, etc.), which presents challenges in terms of learning costs and development efficiency.

In contrast, Blazor allows unified implementation of both client-side and server-side processing to a certain extent in C#. Furthermore, developers can build rich and interactive WebUIs efficiently by using the extensive libraries and tools available through .NET. This enables development of web applications with advanced functions while maintaining code consistency across different technology stacks.

3.5 UI Ingenuity

Event data collected from external devices is aggregated, allowing past occurrences to be viewed via the WebUI. However, simply displaying information means users must search vast amounts of data to find necessary information, which requires considerable effort. Therefore, the following three enhancements were implemented to improve usability.

(1) Hierarchical Event Data

Grouped display is achieved by organizing event data hierarchically (Figure 8), enabling quick identification of the event line to verify the target among vast amounts of data efficiently. This hierarchical organization classifies

information and allows for high-relevance events to be understood at a glance, resulting in significantly improved operability and visibility.

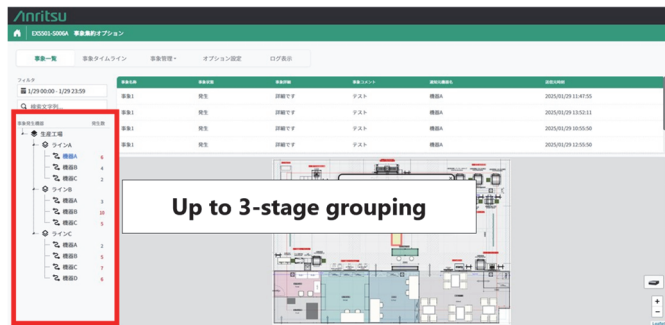


Figure 8 Hierarchical event display

(2) Event Occurrence Location Map

Event locations are plotted on a pre-prepared map.

Additionally, detailed data are checked easily via popup displays (Figure 9). This approach significantly reduces the effort required to identify event locations, enabling quick data comprehension.



Figure 9 Event map location display

(3) Event Timeline Display

Event data is displayed on a timeline (Figure 10), allowing users to see the sequence and interrelationships at a glance. This makes it easier to organize information along a time axis even during multiple events, enabling efficient tracing of occurrence situations.

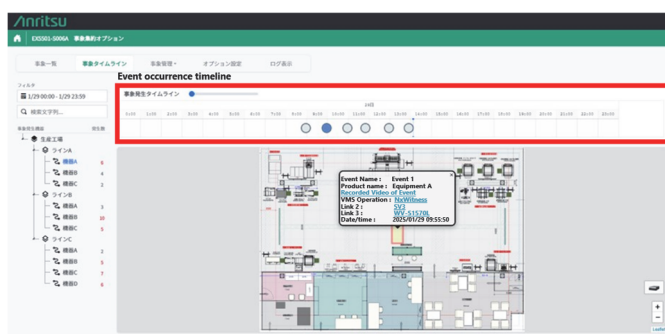


Figure 10 Event data timeline display

3.6 Inter-service Communication using gRPC

Internal, efficient, and fast data transfer between the frontend and backend uses gRPC, an open-source Remote Procedure Call (RPC) that can process multiple requests simultaneously by using HTTP/2.0.

Furthermore, gRPC is ideal for inter-microservice communication, making it very effective for Estinargy, which consists of multiple software packages. Additionally, it offers high scalability, allowing flexible adaptation to future migration to cloud servers depending on the provided functions.

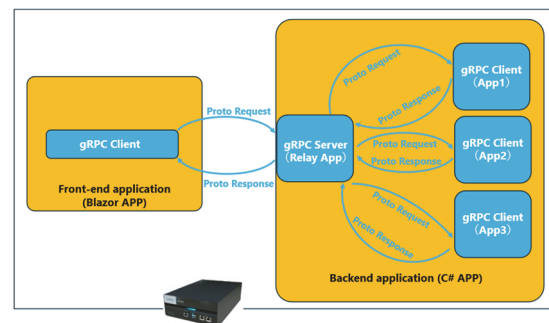


Figure 11 Internal communications architecture using gRPC

3.7 .NET Based Multiplatform Support

Estinargy uses .NET as its application platform. .NET serves as a development environment for multi-platform applications that support various operating systems (OS). It was selected for this development after considering the goal of deployment across a wide range of OS and hardware environments, and taking into account the availability of LTS (Long-Term-Support).

4 Video Technology

4.1 VMS Integration

As described in section 2.2, Video Management System (VMS) integration enables replay of recorded video in a web browser from a specified time range before event occurrence, making it possible to analyze events visually with captured video.

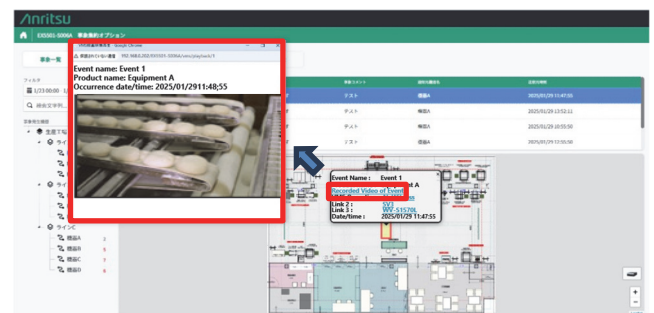


Figure 12 Replaying video record of event

4.2 SightVisor Integration

SightVisor is a multi-decoder that can display video or web content on up to 9 split screens. SightVisor and switching display control at an event displays relevant camera footage and event information automatically. Additionally, live video and recorded video can be played simultaneously by retrieving recorded footage from the VMS. This allows users to seamlessly recognize alerts, assess the current situation, and identify and analyze the cause to efficiently understand and respond to incidents.



Figure 13 Replaying video record of event

4.3 Multiple/Consolidated Video Recording using HDMI Encoder

Multiple camera videos and Web screens can be consolidated as one video record by using a commercial HDMI encoder to encode the video output from SightVisor and record the redistributed video with VMS.

Using this method, multiple locations can be consolidated into one video to cut search times for target videos, allowing quick confirmation of the video data relevance.

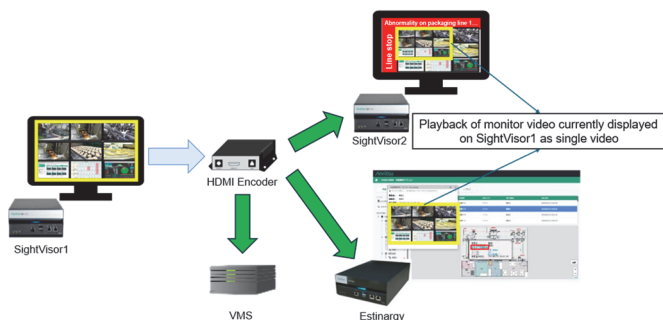


Figure 14 Consolidating multiple video records using HDMI encoder

5 Collaboration with Partners

5.1 On-Site Innovative uSIGMA® Collaboration Solution

Manufacturing and logistics processes involving people and objects (assembly, inventory management, etc.) are not well visualized (easily visible), which is a bottleneck for DX and business management/improvement.

To meet this challenge, Adsol Nissin Co., Ltd. offers the uSIGMA® solution supporting workplace transformation by visualizing, analyzing, and using location and movement data for people and inventory, etc., in factories and warehouses to reduce unnecessary work and improve and transform field operations. By linking with AccelVision to further strengthen the uSIGMA® visualization and analysis capabilities, video can be used in operations management and work analysis so managers can intuitively understand ‘what people were doing’ and ‘where objects are located.’

Specifically, it is possible to play live or recorded video from a specified time to trace the movements of people and objects by requesting camera data from Estinargy to analyze using uSIGMA®.

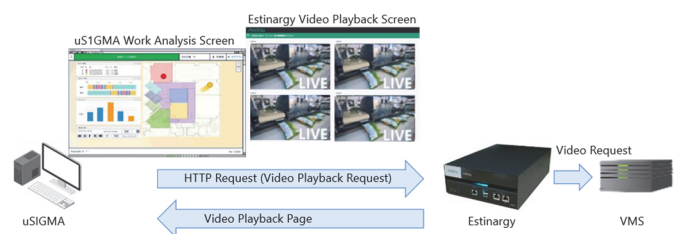


Figure 15 uSIGMA-AccelVision Integration

5.2 Collaboration with Factory and Building OT/IoT Security Services

In today's era when companies, local governments, and various industries are promoting DX, Internet-connected terminals are not limited to IT equipment but are increasing in number to include control devices in factories and buildings, raising concerns about cyberattack risks.

Against this background, NTT Advanced Technology Corporation (NTT-AT) is developing an ‘OT/IoT Security Services for Factories and Buildings’ solution that visualizes communications within internal networks and monitors devices connected to control networks and their communication status. Adding video to this proposed NTT-AT solution facilitates fast and accurate situation assessment at an event. Integration with AccelVision, creates a unified monitoring solution for both cyber (network) and physical (video) aspects to reduce on-site personnel workloads.

Specifically, when NTT-AT's proprietary software (INI: Internal Network Inspector), which automatically analyzes communication logs collected by the Fortinet ‘Fortigate’ network security appliance sold by NTT-AT, detects unauthorized terminals or connections to unusual destinations, it

notifies Estinargy via the WebAPI. Upon receiving a notification, administrators can efficiently confirm the situation visually by recording the event and playing the footage at the event occurrence at sites such as data centers to quickly identify signs of cyberattacks and strengthen responses.

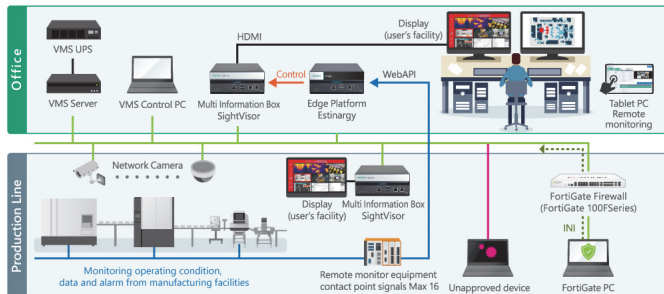


Figure 16 Integrated cyber-physical solution

5.3 Collaboration with ADDSAFE Workplace Safety Support Package

Japanese manufacturing industries, which account for 20% of all industries in the country, suffer the greatest number of workplace related accidents. Furthermore, recent advances in manufacturing automation and product diversification has an increased demand for rapid responses, leading to increased burden on workers and higher risks of occupational accidents. In this context, Konica Minolta Corporation uses its 'ADDSAFE' advanced image processing and AI technology to recognize people and objects from camera footage as a solution to detect abnormal events, such as falls and unauthorized entry in real-time using AI video analysis.

By integration with AccelVision, ADDSAFE detects falls and unauthorized entry and notifies Estinargy, enabling immediate notification to relevant parties and allowing rapid response using two-way communications for better on-site safety management.

Future long-term collection of case and analysis data may help prevent incidents before they occur.

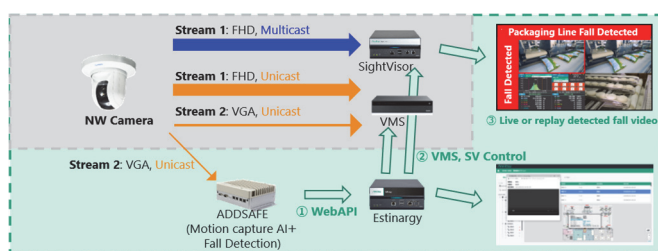


Figure 17 ADDSAFE-AccelVision integration

6 Anritsu Application Examples

6.1 Tohoku Anritsu No. 2 Factory

Tohoku Anritsu, the key factory manufacturing Anritsu's products, has implemented a 'Temperature and Process Monitoring' solution as part of DX initiatives to visualize work processes. This solution uses AccelVision to collect events that occur during the work process and visualize the situation using video.

For example, when an abnormality occurs in a collaborative robot, the event is notified to Estinargy, and the situation at the abnormality is recorded and visualized as video, facilitating analysis of the work process to support decisions about work improvement and optimization.

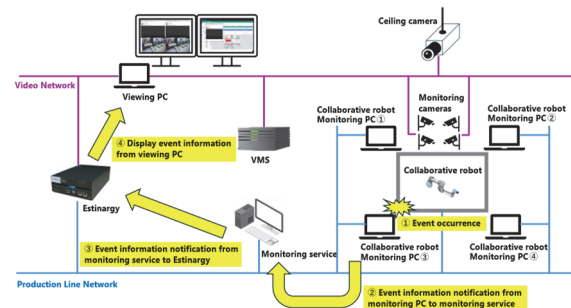


Figure 18 Temperature process monitoring-AccelVision integration

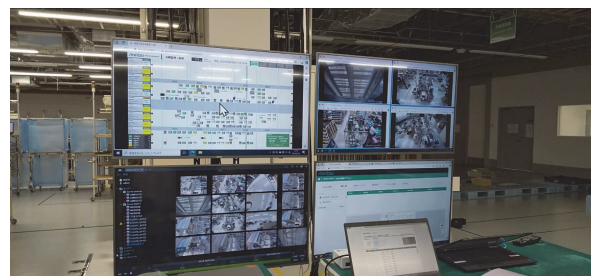


Figure 19 AccelVision at Tohoku Anritsu No. 2 Factory

6.2 Anritsu 5G Laboratory

The Anritsu 5G Lab started operation in April 2021 as a site for technical verification support, introduction of use cases utilizing Local 5G, and collaborative innovation.

AccelVision uses the Anritsu 5G Lab as a site for collaborating with partner companies as described in section 5, and is exploring use cases, verifying feasibility, and conducting demonstrations with invited customers.

AccelVision is also on permanent show, allowing visitors to experience Anritsu's industrial DX solutions. The laboratory is developing products with higher customer value by collecting feedback from customers who have actually experienced AccelVision, thoroughly investigating necessary functions,

and implementing a cycle of additions.

For example, intrusion detection is being demonstrated using ADDSAFE. Figure 21 shows the display when the intrusion detection is notified to AccelVision, allowing visitors to experience how they can detect entry into dangerous areas in real-time, confirm the situation through video, and respond quickly.



Figure 20 ADDSAFE Intrusion detection



Figure 21 ADDSAFE - AccelVision collaboration demonstration

7 Conclusions

We have developed AccelVision, a solution to promote industrial DX primarily in manufacturing industry, as well as Estinargy, an edge platform device forming the AccelVision core. AccelVision enables visual analysis using video data through two-way communication and coordination with external systems, making it possible to improve production efficiency by advancing labor-saving measures.

Furthermore, by collaborating with partner companies, we will continue to expand solutions tailored to customers' needs, not only for on-site safety management and network monitoring, but also to accelerate DX across various industries.

Its effectiveness has been confirmed through demonstrations at Tohoku Anritsu's No. 2 Factory and the Anritsu 5G Lab, and further development is expected through future technological innovation.

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Publicly available

Table 1 Specifications

Item		EX5501A
Main interfaces	LAN interface	1000 BASE-T×2
	Video output	Digital video (HDMI×2connectors) *However, uses only one HDMI in normal mode
	Audio output	Analog (2 ch) 3-pole stereo mini jack×1
	Output impedance	Approx. 150Ω
Home	Software list	Installed software list for added functions/WebUI link component names
Server operational data	System data	CPU usage, memory usage free disk space
	Network data	LAN1 and LAN2 MAC address, IP addresses
Maintenance	System data	Equipment name/serial number/installed software/license details/system data
	System settings	Software updates/license/date and time auto settings/software log/database initialization/equipment internal log initialization/equipment restart/equipment shutdown
	Network settings	LAN1 and LAN2 IP address settings/Static routing setting/Ping
	Backup/restore	Database backup/restore
Control	Remote control	Power-on/off using infrared remote controller
	Settings	WEB browser

Table 2 Option Software List

Model	Name	Remarks
EX5501-S001A	Basic software	Standard install in edge platform device
EX5501-S002A	SightVisor Integrated management software	Web application supporting SightVisor control functions
EX5501-S003A	WebAPI Option	Software to receive SightVisor control and external device data in REST format as HTTP Request
EX5501-S004A	Data integration option	Software to display camera icons on maps or floor plans based on camera position data, enabling visual selection of cameras
EX5501-S005A	Camera PTZ option	Software providing conversion and relay functionality from JBEE (Japan Building Electrical and Electronics Engineers Association) commands to ONVIF format control, to control ONVIF format cameras that cannot be directly controlled from SightVisor
EX5501-S006A	Event aggregation option	Software to store and view event information notified from external devices, and to play video footage linked to event information
EX5501-S007A	Event aggregation option for partners	Software to store and view event information notified from external devices, control video related to event information (including playback, fast-forward, rewind, etc.), externally transfer event information, and forward email

Mobile Device Field Tests with Excellent Reproducibility

Hirofumi Suganuma, Keisuke Takizawa, Genta Oki, Kota Kuramitsu

[Summary]

Ensuring the performance and functionality of mobile devices requires testing in actual on-site field environments as well as laboratory testing based on specific test cases. However, field test results may not necessarily be reproducible due to changes in base station operating conditions and radio propagation channels. This article proposes a Field Simulation Test (FST) solution to overcome these field-test problems. This solution tests devices in the laboratory based on radio propagation channels captured on-site supporting repeated laboratory tests under the same conditions as actual field conditions. Additionally, we introduce examples of how this solution is applied in on-site field tests.

1 Introduction

Following commercial rollout in 2020, the fifth-generation mobile communication system (5G) has spread rapidly and 5G subscriptions are predicted to account for half of all mobile subscriptions by 2028¹⁾. Moreover, commercial deployment of the sixth-generation mobile communication system (6G) is expected around 2030, and discussions have started on requirements, use cases, and fundamental technologies^{2), 3)}.

Before user equipment (UE), such as smartphones and tablets, reaches users, it must be verified and evaluated to ensure stable, high quality. For example, during R&D, various tests are conducted on radio frequencies (RF), protocols, data throughput, power consumption, etc., to evaluate device performance and functionality. Additionally, conformance tests^{4), 5)} and carrier acceptance tests (CAT)⁶⁾ are also conducted before deployment to verify interconnectivity between devices and mobile networks. Conformance tests evaluate whether a device complies with the technical standards specified by the Third Generation Partnership Project (3GPP). CAT verify whether the device has the unique functions defined by the carriers. Generally, these tests are conducted in laboratory environments based on test cases that assume specific conditions.

On the other hand, device testing is also conducted in actual on-site field environments to complement the test cases specified in laboratory testing⁴⁾ and to evaluate device quality as experienced by users⁷⁾. In the future, mobile communication services are expected to expand from best-effort type to quality-assured type to support demanding emerging applications¹²⁾, such as autonomous vehicles like

drones⁸⁾ and self-driving cars⁹⁾, industrial Internet of Things (IoT)¹⁰⁾, and telemedicine¹¹⁾. Therefore, the importance of device field testing is expected to continue increasing. However, in actual field environments, there are issues, such as the difficulty of repeatedly conducting device tests under identical conditions due to changes in the operational status of base transceiver stations (BTS) and radio propagation environments, meaning that test results may not always be reproducible^{13), 14)}. Consequently, device testing based on actual field environments but with controllable conditions is needed.

In these circumstances, this article proposes a Field Simulation Test (FST) solution to analyze BTS signals measured in actual field environments, model the radio propagation environment, and use that model to test devices in the laboratory. This solution can repeatedly test devices under identical conditions matching actual field environments. Additionally, we evaluate the model generation accuracy in known radio-propagation environments to demonstrate the effectiveness of this solution, and introduce an actual field application example.

2 Proposed Solution

2.1 Overview of Field Simulation Test (FST)

Figure 1 shows the concept of the FST solution. Although 3GPP specifies test cases including radio-propagation conditions for device testing, it is difficult to cover all environmental conditions. Therefore, device field tests aim to complement test cases⁴⁾. At conventional field tests, staff from device manufacturers travel to actual field environments to conduct tests using the device under test

(DUT). This requires enormous time and costs because field tests are repeated for each new device development as well as to verify improvements. Furthermore, field-test results may not necessarily be reproducible due to changes in BTS operational status and radio propagation environments. In response to these efficiency and reproducibility issues with device field tests, this FST solution test devices in a laboratory while simulating the radio propagation environment acquired in actual field environments, making it possible to repeatedly test devices under identical conditions in a laboratory based on actual field environments.

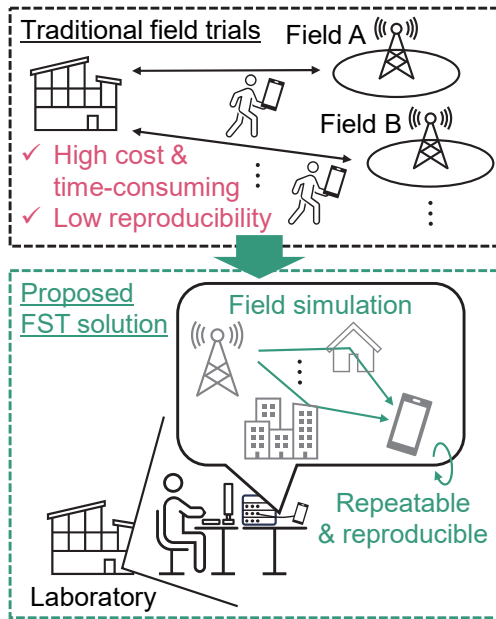


Figure 1 FST solution concept

Figure 2 shows the overall configuration of the FST solution, consisting of three steps: capturing, analysis and modeling, and testing. At the first capturing step, UE logs containing communication protocol sequence data are captured in the actual field environment, and downlink (DL) signals transmitted from BTS are captured as baseband in-phase and quadrature (IQ) data. The next analysis and modeling step generates fading models characterizing the radio propagation environment by analyzing the recorded IQ data while using the UE logs. The last testing step simulates the actual field environment using a BTS simulator and fading simulator based on the generated model to evaluate device connectivity.

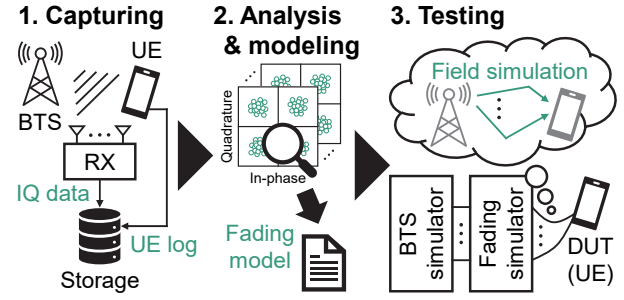


Figure 2 Overall configuration of FST solution

The following sections describe each of these three steps in detail.

2.2 Capturing

At the capturing step, DL signals transmitted from 5G New Radio (NR) or 4G Long Term Evolution (LTE) BTS are captured as baseband IQ data simultaneously with UE logs for any required time period. Figure 3 shows the configuration of the RF Air Monitor System for recording IQ data. As shown in this figure, this solution supports either a standard configuration for long-term multi-band recording, or a lite configuration that fits in a backpack, depending on requirements.

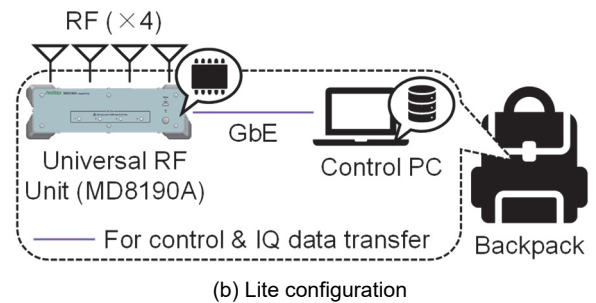
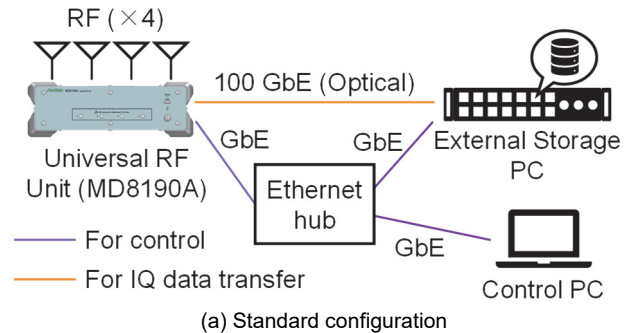


Figure 3 RF Air Monitor system configuration

The standard configuration in Figure 3(a) consists of a Universal RF Unit (MD8190A), PC for external storage, and control PC. The Universal RF Unit is a general-purpose RF interface device for converting analog RF signals to digital baseband IQ data. Using this solution, the Universal RF Unit with connected antennas receives RF signals transmitted from

BTS and converts them to IQ data. The external storage PC with large-capacity solid state drive (SSD) is connected to the Universal RF Unit via 100GbE optical fiber. IQ data converted by the Universal RF Unit are transferred in real-time to the external storage PC for storage on the SSD. The control PC is connected to both the Universal RF Unit and external storage PC via a GbE cable and controls both using the Log Capture Control Software (MX879000PC). Additionally, the RF Air Monitor System supports simultaneous capture using two Universal RF Units. This enables simultaneous recording of two NR standalone (SA) or LTE bands with carrier aggregation, as well as simultaneous recording of both NR and LTE NR non-standalone (NSA) bands. This configuration requires a rack-mounted external storage PC for long-term recording and is intended primarily for use in vehicles.

On the other hand, the lite configuration in Figure 3(b) eliminates the external storage PC so the entire equipment is small enough to fit in a backpack. In this configuration, the received RF signals are first converted to IQ data and saved temporarily in the Universal RF Unit internal random access memory (RAM). The IQ data saved in RAM is then transferred to and saved in the control PC via GbE cable. Although the lite configuration has limited recording time compared to the standard configuration, it cuts the workload when transporting equipment and recording.

Table 1 shows the RF Air Monitor System specifications supporting capture of RF signals from 0.4 to 6 GHz for NR frequency range 1 (FR1) and LTE systems. Bandwidths up to 100 MHz are supported and data can be captured for up to 4 RF ports per Universal RF Unit. The continuous recording time is 120 seconds for the standard configuration and 8 seconds for the lite configuration when capturing 100-MHz bandwidth signals for 4 RF ports.

Table 1 RF Air Monitor System Specifications

Number of RF ports	4
Frequency	0.4 to 6 GHz
Bandwidth (BW)	≤100 MHz
Sampling rate	30.72 Msps for ≤20 MHz BW 61.44 Msps for ≤50 MHz BW 122.88 Msps for ≤100 MHz BW
Recording time	120 s for 100 MHz BW × 4@Standard 8 s for 100 MHz BW × 4@Lite

Figure 4 shows the Log Capture Control Software for capturing IQ data in real-time while checking the power and

frequency spectrum of signals being received by the Universal RF Unit.

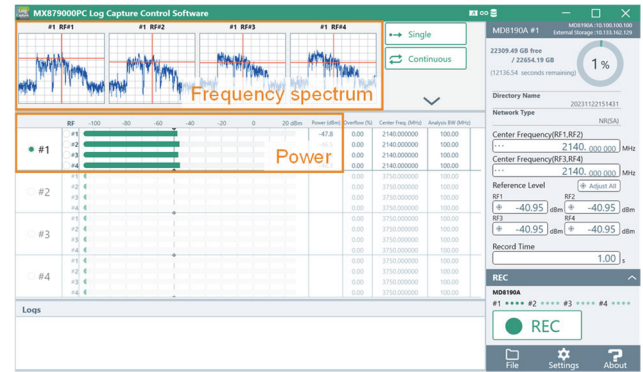


Figure 4 Log Capture Control Software (MX879000PC)

2.3 Analysis and Modeling

At the analysis and modeling step, the IQ data recorded using the RF Air Monitor System described in the previous section is analyzed to generate fading models that characterize the radio propagation environment. This step is carried out through the software FST Studio (MX879020PC) installed on a PC.

Table 2 shows the recorded target signals for analysis and modeling. As shown, FST Studio supports analysis of NR NSA, NR SA, and LTE signals. Specifically, the NR analysis targets are synchronization signal/physical broadcast channel (SS/PBCH) block, physical downlink shared channel (PDSCH), and channel-state information reference signal (CSI-RS)¹⁵. Additionally, the LTE analysis target is cell-specific reference signal (CRS)¹⁶. These target signals are used to estimate the instantaneous responses of the radio propagation channels and are processed while referencing radio resource control (RRC) messages in the UE logs acquired simultaneously with these signals as needed. FST Studio defines the open UE Log Open Interface file format in order to read RRC messages. UE logs captured in the field by users can be used for analysis after conversion to this format.

Table 2 Captured Data for Analysis and Modeling

Network type	NR SA/NR NSA (Option3x)/LTE
Bandwidth	10/15/20/25/30/35/40/45/50/60/70/80/90/100 MHz for NR, 1.4/3/5/10/15/20 MHz for LTE
Channel type	SS/PBCH block/PDSCH/CSI-RS for NR CRS for LTE
SCS	15/30 kHz
Number of layers	1 layer for NR SS/PBCH block 1/2/3/4 layer(s) for NR PDSCH-DMRS/CSI-RS and LTE CRS

Figure 5 shows an example of radio propagation channel analysis results using FST Studio. As shown, the instantaneous impulse response of the radio propagation channel is displayed in the top part of the screen, and the received power is displayed in the lower part. Based on this result, the user chooses the time range of the instantaneous response to generate the fading model. FST Studio adopts the well-known tapped delay line (TDL) model¹⁷⁾ as the fading model. The following parameters are primarily calculated based on the selected instantaneous response time range.

- Power delay profile (PDP): Relationship between multipath delay time and average power in radio propagation channel response
- K factor: Power ratio between line-of-sight (LOS) component and non-line-of-sight (NLOS) component
- Antenna correlation: Correlation of radio propagation channel response between either Tx or Rx antennas that significantly affects quality of multiple-input and multiple-output (MIMO) transmission

Furthermore, FST Studio also has a simple analysis function to quickly check IQ data captured in the field, allowing users to quickly spot abnormalities and reduce the need and costs of recording do-overs.

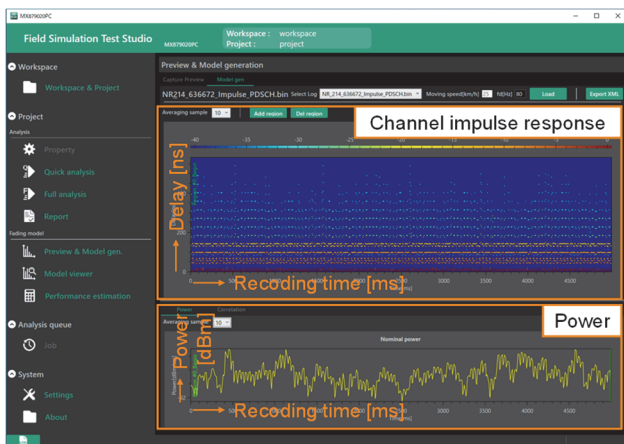


Figure 5 Example of FST Studio (MX879020PC) analysis results

2.4 Testing

Testing using the Radio Communication Test Station (MT8000A) evaluates the device connectivity in accordance with the actual field environment based on the fading model described in the previous section. Figure 6 shows typical device test system configurations. The MT8000A simulates NR and LTE BTS behavior and fading. Figure 6(a) shows an NR SA configuration for 4×4 MIMO NR device testing.

Figure 6(b) shows an NR NSA configuration for both 4×4 MIMO in both NR and LTE device testing. The test system is controlled using the Rapid Test Designer (RTD) installed in the control PC.

By using the above-described test configuration, devices can be tested repeatedly under the same conditions based on actual field-environment characteristics.

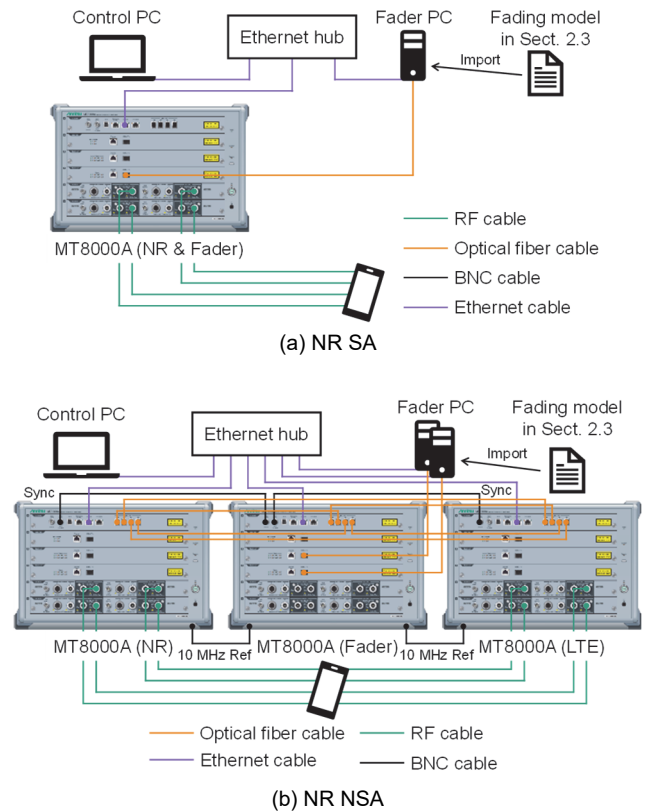


Figure 6 Example of device test systems

3 Demonstration

3.1 Evaluating Fading Model Generation Accuracy

This section describes using an experiment to evaluate the validity of modeling radio propagation characteristics, which is the key to achieving device tests matching actual field conditions.

Figure 7 shows the experimental configuration. The evaluation generates a fading model from signals captured in known fading environments and evaluates its accuracy. Specifically, first, a DL signal that has passed through a known fading channel is output using BTS and fading simulators. Then, that DL signal is recorded as IQ data using the FST solution and a fading model is generated by analyzing the data. Finally, the parameters of the generated model are compared with those of the known fading channel.

This example uses the 3GPP TDL-A model¹⁸⁾ for the NLOS environment known fading with delay spread set to 30.0 ns and antenna correlation set to high (Tx correlation coefficient $\alpha = 0.90$, Rx correlation coefficient $\beta = 0.90$)^{19), 20)}. In addition, the model assumes the BTS simulator band is set to n78 (3.5-GHz center frequency), the bandwidth is 100 MHz, and 4×4 MIMO NR PDSCH signals are output.

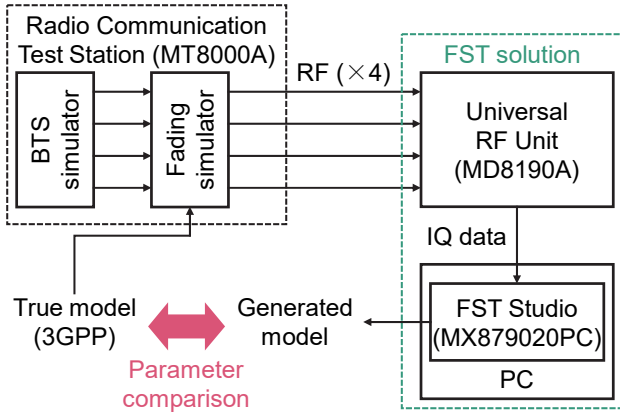


Figure 7 Experimental configuration for evaluating generated fading model accuracy

Table 3 shows the results of fading model generation. In this table, the K factor value of -99 dB represents an NLOS environment. This table shows that the FST-generated fading model K factor, Root Mean Square (RMS) delay spread, and antenna correlation parameters generally match those of the 3GPP TDL-A model.

Table 3 Generated Fading Model Results (TDL-A)

	Generated model (FST solution)	True model (3GPP)
K factor	-99 dB	-99 dB
RMS delay spread	27.2 ns	30.0 ns
Tx correlation α	0.88	0.90
Rx correlation β	0.89	0.90

3.2 On-site Field Usage Example

This section introduces an example FST application for on-site field usage. First, we captured the DL signal from a sub-6 GHz band 5G NR BTS as IQ data in an NLOS environment in a suburban area. Next, the PDSCH signal in that IQ data was analyzed to generate a fading model.

Figure 8 shows the generated PDP based on the captured signal as well as the K factor. This figure shows that the generated-model environment matches NLOS because the K factor value is -99 dB, and the power of the second or third wave is larger than the power of the first wave with zero

delay. A fading model generated in this way can be used in the testing step described in section 2.4, facilitating device testing matching actual on-site field environments. Note that the RMS delay spread is 38.0 ns when calculated from the PDP. This is consistent with the RMS delay spread from 32 to 153 ns defined in the rural macro (RMA) scenario for the sub-6 GHz band in 3GPP TR38.901¹⁸⁾.

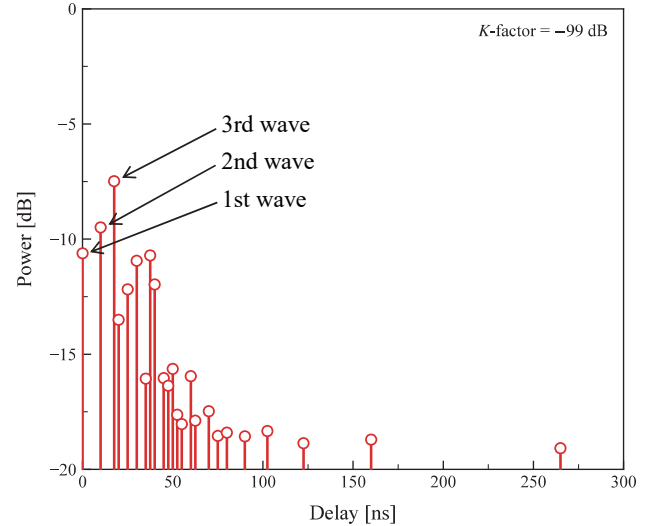


Figure 8 Generated PDP based on signal captured in suburban area

4 Conclusions

This article proposes an FST solution to solve issues such as efficiency and reproducibility at on-site field testing of mobile devices. The FST generates a fading model that characterizes the radio propagation environment by analyzing BTS signals measured on-site, and conducts laboratory device tests while simulating the actual on-site field environment using that model. This facilitates repeated device tests under the same conditions matching actual field environments. Evaluation of the validity of the model generated by this solution in known fading environments gave generally accurate values for K factor, RMS delay spread, and antenna correlation. By applying this solution to an actual on-site sub-6 GHz band 5G NR BTS we could generate an NLOS fading model matching the captured signal environment.

Anritsu is continuing to develop solutions that contribute to easier and accurate field testing, helping widespread adoption and development of mobile communication systems.

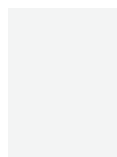
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Development of MT8862A IEEE 802.11be 320-MHz Bandwidth Option

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[Summary]

The number of WLAN devices is increasing yearly with demands for more stable, faster, and higher-capacity communications. To address growing data-traffic demands, the IEEE 802.11be standard added the 320-MHz bandwidth and 4096QAM modulation schemes³⁾. Anritsu's Wireless Connectivity Test Set MT8862A is a popular solution for evaluating RF performance^{4), 5)}. Consequently, we have developed new hardware and software update options for the MT8862A to support the IEEE 802.11be 320-MHz bandwidth as an industry-first^{*1} instrument for RF tests using the Network mode.

*1 Anritsu survey in August 2023

1 Introduction

Use cases for Wireless LAN-equipped (WLAN) devices continue to expand, including playback of 4K/8K videos using smartphones and tablet devices, enhancement of services using Augmented Reality (AR)/Virtual Reality (VR) technology on wearable devices, and control/remote monitoring applications in factories and medical equipment. Additionally, to meet increasing data traffic demands, the Institute of Electrical and Electronics Engineers (IEEE) is increasing the maximum communication speed of WLAN devices by widening the channel bandwidth to 320 MHz and upgrading the modulation scheme to 4096QAM in the latest IEEE 802.11be standard^{1) to 3)}. Anritsu markets its Wireless Connectivity Test Set MT8862A solution for measuring the RF transmitting and receiving (Tx/Rx) characteristics of IEEE 802.11a/b/g/n/ac/ax (2.4-GHz, 5-GHz, and 6-GHz band) WLAN devices^{4), 5)}. It supports both the Network mode, enabling Tx/RX measurements in the actual operating mode of the device under test (DUT), and the Direct mode, which uses chip vendor-specific test modes used on production lines, etc. Therefore, it is used for both Over The Air (OTA) measurements and Conducted measurements.



Figure 1 Wireless Connectivity Test Set MT8862A

We have enhanced the MT8862A functions to meet future mainstream IEEE 802.11be RF measurement needs by developing the Extended Bandwidth Hardware MT8862A-011 option, as well as the 320 MHz Bandwidth MX886200A-031 and WLAN 802.11be MX886200A-003 software options. This article describes the key IEEE 802.11be technologies, the MT8862A development policy supporting IEEE 802.11be, and the option development details and key specifications.

2 IEEE 802.11be

IEEE 802.11be adds the following functions as key technologies³⁾.

- 320-MHz bandwidth
- 4096QAM
- MRU (Multiple Resource Unit)
- MLO (Multi-Link Operation)

This section discusses the 320-MHz bandwidth, 4096QAM, and MRU technologies that were deeply involved in this development.

2.1 320-MHz Bandwidth

To increase the maximum communication speed, the maximum channel bandwidth in the 6-GHz band has been expanded from 160 to 320 MHz. The 320-MHz channel consists of two adjacent 160-MHz channels, and two types of channel configurations, 320 MHz-1 and 320 MHz-2 are defined. Figure 2 shows this channel arrangement.

2.2 4096QAM

The maximum value for the QAM modulation method has been increased from the existing 1024 to 4096, increasing the

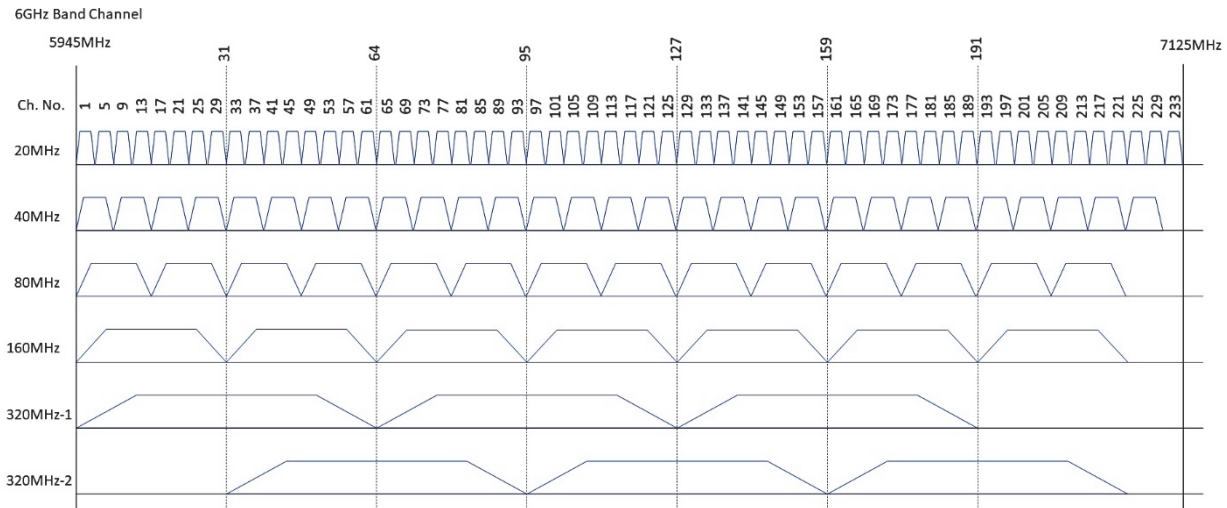


Figure 2 6-GHz band channel distribution

number of bits that can be transmitted per modulation symbol from 10 bits to 12 bits, and enabling a 20% speed improvement. The specified permissible relative constellation error for 4096QAM is -38 dB.

2.3 MRU

To use frequency resources efficiently, a Multiple Resource Unit (MRU) is defined, which allocates multiple Resource Units (RUs) to a single user. RUs are allocated in subcarrier quantities of 26, 52, 106, 242, 484, 996, 2x996, and 4x996 tones. The MRU is a group of subcarriers combining multiple RUs defined as 52+26-tone, 106+26-tone, 484+242-tone, 996+484-tone, 2x996+484-tone, 3x996-tone, and 3x996+484-tone combinations.

3 Development Concept

3.1 Developing High-Performance Hardware

Our aim was to develop hardware to measure a 320-MHz bandwidth and 4096QAM, which are the main IEEE 802.11be performance features. The design residual Error Vector Magnitude (EVM) for both the signal generator and signal analyzer was set at -43 dB based on market performance requirements, commercial release timing, and product cost.

3.2 Compatibility with Existing Hardware

To keep upgrade costs down for existing users, the design allows upgrades to 320-MHz bandwidth support by replacing the MT8862A built-in module. Additionally, by compensating for the differences from the current hardware of MT8862A by software and FPGA circuits, we provide an environment

where users can perform RF testing using exactly the same procedures and operations as the current MT8862A, without being conscious of hardware differences.

3.3 Advance Compatibility Testing with 802.11be Terminals

To cut time to market (TTM), development started early in the IEEE 802.11be formulation stage. While complementing as yet undefined parts of the standard using connection verification by chip vendors that develop standards, we also focused on enhancing the completeness of the signaling mode, which is a key MT8862A feature.

4 Hardware

There were two performance goals in meeting the technical challenges of this hardware development as follows:

- ① Expanding maximum bandwidth (160 → 320 MHz)
- ② Upgrading EVM at maximum bandwidth to reach -43 dB target

EVM (point ②) is one of the most important performance items generally required for communications measuring instruments, and this development achieved both points ① and ② for the MT8862A SG and SA functions. Additionally, repurposing in-house design assets help cut both costs and TTM, meeting instrument-market requirements.

The MT8862A hardware is divided into three units based on their roles (Figure 3). The roles of each hardware unit and the main changes from the conventional MT8862A hardware to achieve the above-mentioned performance goals ① and ② are described in following sections organized by SG/SA functions.

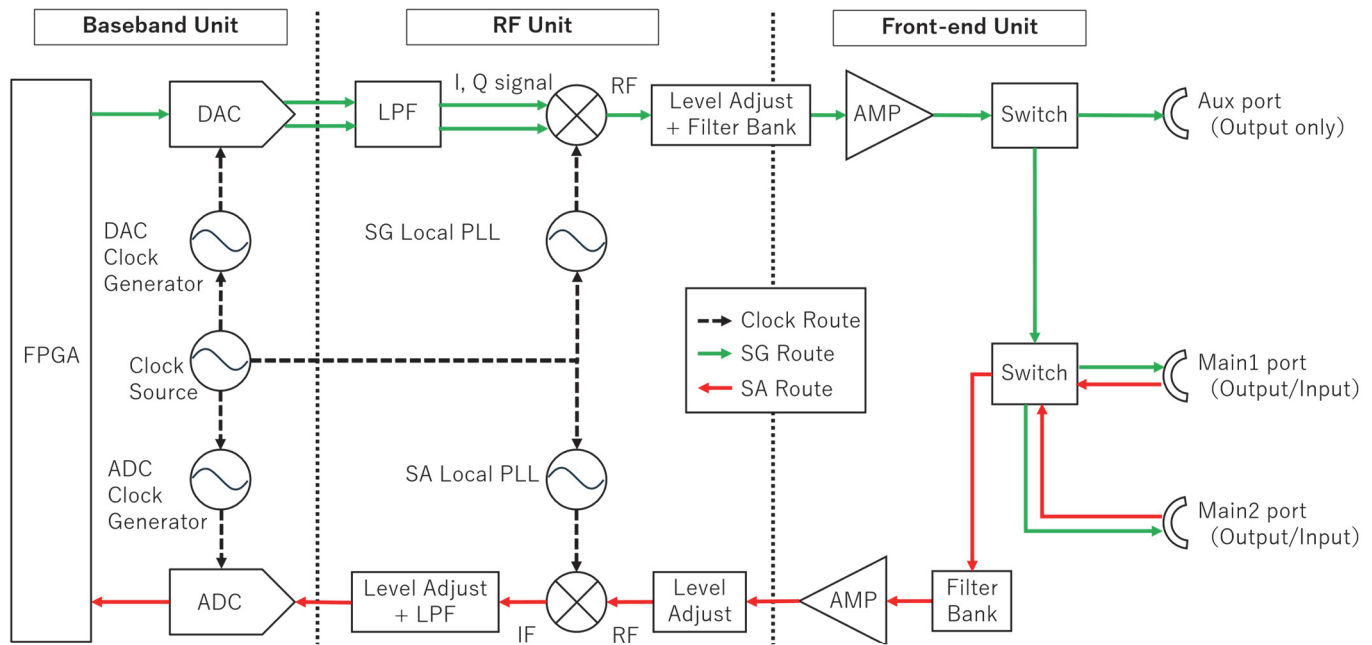


Figure 3 MT8862A Hardware block diagram supporting IEEE 802.11be 320-MHz bandwidth

4.1 MT8862A Signal Generator Functions

The green arrows in Figure 3 show the SG function signal path. Details of each unit are described below.

4.1.1 Baseband Unit

The waveform data output from the FPGA in the baseband unit is converted to IQ data by the Digital-to-Analog Converter (DAC) and output to the RF unit.

The DAC in the baseband unit is the main change to the SG function. The DAC in the conventional MT8862A cannot handle the digital-waveform data rate required for the 320-MHz bandwidth, so it was upgraded to a DAC that can accept input at a higher data rate. This required increasing the sampling clock frequency, so a new frequency multiplier was added. To ensure performance, consideration was given to preventing phase noise deterioration when selecting the frequency multiplier.

4.1.2 RF Unit

The IQ signals received by the RF unit are frequency-converted to RF signals by the quadrature modulator. The RF signal output from the quadrature modulator has unwanted waveforms removed by the filter, and is output to the Front-end unit after the signal level is adjusted by the digital attenuator.

Doubling the maximum signal bandwidth from 160 to 320 MHz could cause harmonic problems and multiple responses, resulting in an unwanted wider bandwidth at the RF unit. Consequently, the frequency plan was reviewed, and the

mixer local frequency was changed along with selection of filters with appropriate cutoff frequencies.

Additionally, in general, setting an SG to high output levels can cause distortion components to become the dominant factor in degraded EVM performance. Degraded EVM performance at high output level settings was suppressed by using an amplifier with good distortion performance at the latter level-adjustment stage. Moreover, improved EVM and lower cost increases were achieved by reviewing the level plan including digital-stage level adjustment, and optimizing the attenuator configuration of the previous MT8862A.

4.1.3 Front-end Unit

The received RF is output to the Aux, Main1, or Main2 ports specified by the user.

The Front-end unit challenge is that the distortion performance of the final amplifier contributes significantly to EVM performance at high output levels. An upgraded amplifier with better distortion performance than the one in the previous MT8862A was used to solve this issue.

4.2 MT8862A Signal Analyzer Functions

The red arrows in Figure 3 show the SA function signal path. Details of each unit are described below.

4.2.1 Front-end Unit

The RF signal input either to the MT8862A Main1 or Main2 port has unwanted waveforms removed by a filter in the Front-end unit and is amplified by the Low Noise Amplifier (LNA). The amplified RF signal then passes to the RF unit.

When the RF signal input level is low, the signal is amplified by the LNA, so the LNA characteristics affect EVM performance. Adopting an LNA with better frequency characteristics than the previous MT8862A improved the EVM performance compared to using conventional components. Cost was also considered when choosing the better LNA, resulting in a cost saving of about 40% compared to conventional components.

4.2.2 RF Unit

RF signals received from the Front-end are frequency-converted to Intermediate Frequency (IF) signals in the RF Unit. The RF signal level is adjusted at the stage prior to the mixer to achieve the appropriate mixer input level; unwanted waveforms are removed from the IF signal after the mixer and the signal level is adjusted.

Expanding the signal's maximum bandwidth required changing the frequency dial. Like the SG function, the second harmonics overlap became more stringent compared to the previous MT8862A model as the signal bandwidth doubled. Additionally, special attention was required for handling unwanted waveforms during mixing. Therefore, the local mixer frequency was changed, and filters in the path were modified to achieve appropriate pass and blocking characteristics. Additionally, Figure 4 shows the conditions for preventing overlap between the wanted wave and its second harmonic. The sum of half the bandwidth of the wanted wave (labeled "a" in figure) and half the bandwidth of the second harmonic (labeled "b" in figure) must be smaller than the IF (labeled "IF Center" in figure). The IF is adopted according to the above conditions.

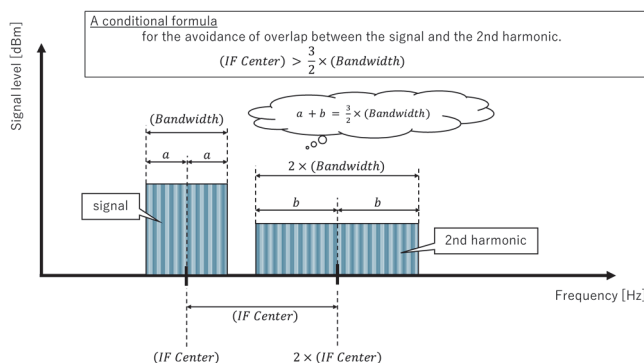


Figure 4 Conditions preventing overlap between wanted wave and second harmonic

4.2.3 Baseband Unit

The IF signal output from the RF unit is digitized by the

Analog-to-Digital Converter (ADC) in the baseband unit. The digital data output by the ADC is analyzed by FPGA and software. The main change in the SA function of the baseband unit is the ADC. The ADC in the previous MT8862A has a sampling rate of 400 MSPS, which cannot handle signals with a 320-MHz bandwidth. Therefore, it was replaced with an ADC capable of sampling at a higher rate. Although it has fewer effective bits (ENOB) than the ADC in the previous MT8862A, EVM deterioration is prevented by oversampling at a higher sampling rate.

4.2.4 ADC Clock Generation Phase-Locked Loop IC

Unlike the previous MT8862A, since oversampling is performed at a sufficiently high rate by the ADC, a faster sampling clock was necessary. Moreover, a new Phase-Locked Loop (PLL) IC with low phase noise was used to achieve the target EVM value of -43 dB at the 320-MHz maximum bandwidth, taking clock jitter requirements into account.

4.3 Changes to Shared SG/SA Functions

The PLL IC used as the local oscillator for the mixer for both SG and SA functions was changed. Phase noise, which is one component of EVM performance, is highly dependent on the phase noise performance of the local signal during the mixing stage. Therefore, changing to a PLL IC with lower phase noise contributed to improved EVM performance.

5 FPGA

The main FPGA modifications were to the interface and expansions to accommodate hardware component changes. Additionally, parallel processing was implemented in the digital signal processing block to support higher DAC and ADC sampling rates.

The signal processing block configuration was designed to handle compatibility with existing operations, such as selecting the sampling rate of received data passed to software, minimizing the impact of changes.

6 Software

This section describes the MT8862A software modifications to support 320 MHz. Figure 5 shows the software functional blocks.

6.1 Analysis Section

6.1.1 Modulation Analysis Sampling Rate

The biggest challenge in the analysis section was the

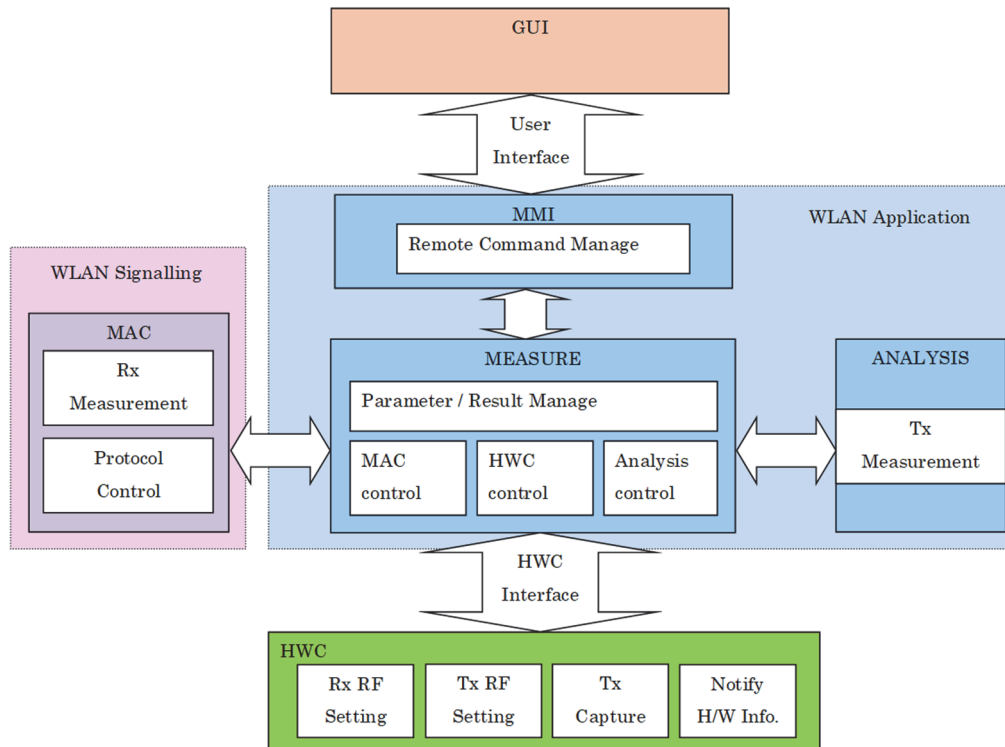


Figure 5 MT8862A Software function blocks

change in sampling rate. Since the IEEE 802.11ax standard maximum bandwidth is 160 MHz, the prior software was designed with a fixed sampling rate of 200 MHz. However, it was necessary to increase the sampling rate to 400 MHz because IEEE 802.11be extends the maximum bandwidth to 320 MHz. Also, to use the software in common with the previous MT8862A, it was necessary to ensure that IEEE 802.11be devices could be analyzed even using the 200-MHz sampling rate system.

Therefore, we designed the implementation for the newly developed IEEE 802.11be processing to use variable sampling rates, while existing IEEE 802.11ax and earlier processing is resampled to 200 MHz, allowing reuse of the existing implementation. Figure 6 shows a flowchart focusing only on the transition of sampling rates in the analysis section. This implementation style made the software compatible with both the previous and updated MT8862A.

6.1.2 IEEE 802.11be Analysis Section

The IEEE 802.11be analysis section requires a variable sampling rate. However, resampling according to bandwidth was expected to cause problems such as waveform degradation due to filtering and strain on computational and memory resources. Therefore, since the subcarrier spacing in the PPDU-format data region is fixed at 78.125 kHz, we designed

the analysis section to perform Discrete Fourier Transform (DFT) so subcarrier spacing becomes 78.125 kHz. This approach suppress waveform degradation compared to Fast Fourier Transform (FFT), enabling high-quality analysis. There were concerns about increased measurement time, but the increase in computation time was suppressed by reducing the number of resampling operations.

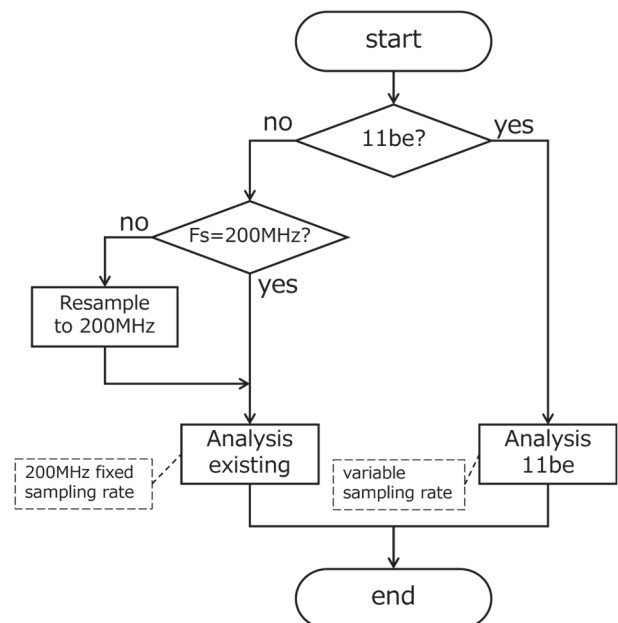


Figure 6 Flowchart of sampling rate transitions in analysis section

6.2 WLAN Signaling Section

In the WLAN signaling section, we mostly implemented support for adding IEEE 802.11be parameters, such as 4096QAM, 320-MHz bandwidth, and MRU to control packets. Software development proceeded by verifying network connections in advance with chip vendors using existing hardware. The occurrence of defects in the actual instrument was suppressed by preverifying the signaling section more than 6 months before integrated testing of the new hardware.

6.3 GUI Section

In addition to IEEE 802.11be settings and measurement results, we added a new configuration screen for the newly standardized MRU to the web GUI WLAN measurement software screen. The MRU configuration screen displays RUs up to 320 MHz as icons, helping understand the layout and

combination of RUs at a glance. Additionally, providing a function to switch between Single Resource Unit and Multiple Resource Unit facilitates switching within the same screen. Figure 7 shows the Single Resource Unit configuration screen when the bandwidth is 20 MHz, and Figure 8 shows the Multiple Resource Unit configuration screen.

Furthermore, due to GUI screen size constraints, displaying the entire view for the 320-MHz bandwidth made the 26-tone icons too small, creating an operation issue. To prevent poor usability, we added zoom and scroll functions, allowing display of the 26-tone icons at the normal size, balancing overall visibility and operability. Figure 9 shows the Single Resource Unit configuration screen in full view when the bandwidth is 320 MHz, and Figure 10 shows the configuration screen when zoomed-in.

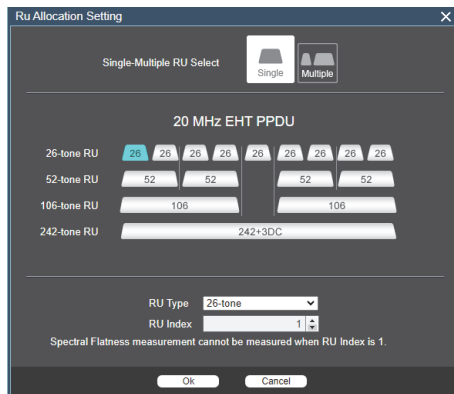


Figure 7 20-MHz Single Resource Unit Setting screen example

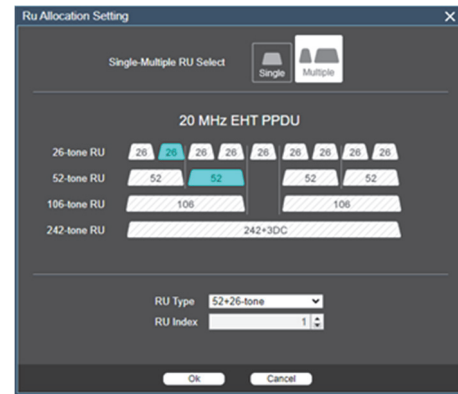


Figure 8 20-MHz Multiple Resource Unit Setting screen example

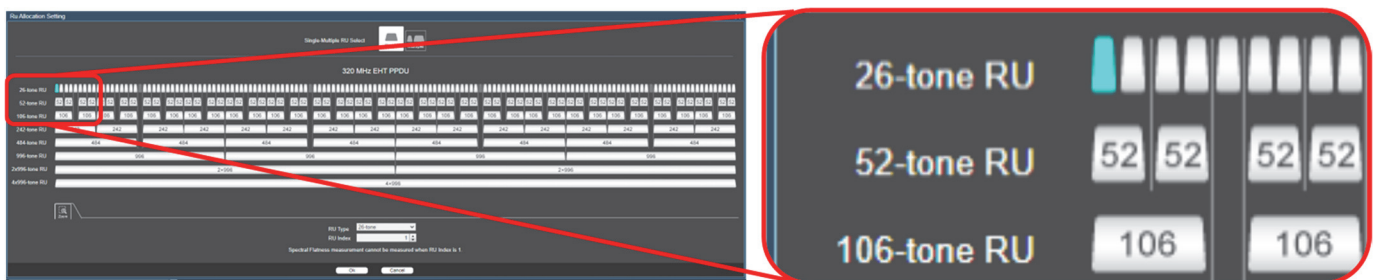


Figure 9 320-MHz Resource Unit setting screen full view example



Figure 10 320-MHz Resource Unit setting screen zoomed-in example

7 Key Specifications

Tables 1 and 2 list the main specifications of the MT8862A with 320-MHz bandwidth support.

8 Measurement Result Examples

Figure 11 shows a Constellation from EVM measurement as an example. This Constellation is the result of measuring an IEEE 802.11be MCS13 signal with 4096QAM modulation at a center frequency of 6105 MHz and an input level of -10 dBm. The EVM measurement result is -45.31 dB, meeting the development target of -43 dB. Figure 12 shows the Spectrum Mask measurement result. This Spectrum Mask is the result of measuring an IEEE 802.11be 320-MHz bandwidth signal at a center frequency of 6105 MHz and an input level of -10 dB. As shown in Figure 12, it supports measurements up to ± 480 MHz.

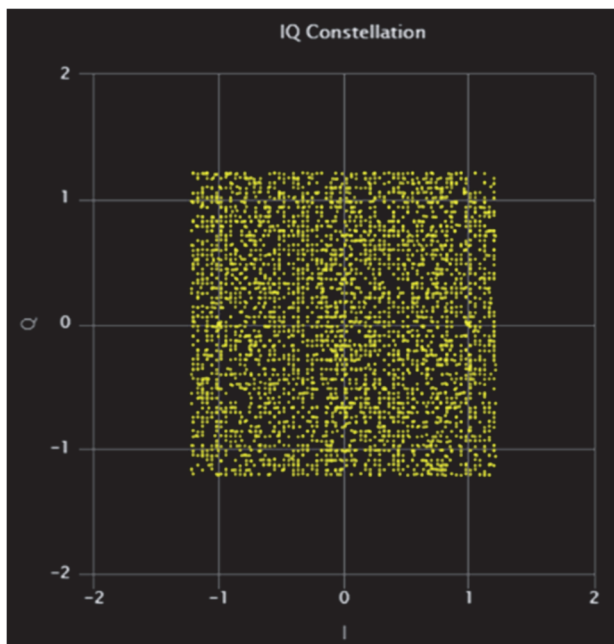


Figure 11 Example of 4096QAM Constellation for 320-MHz bandwidth signal

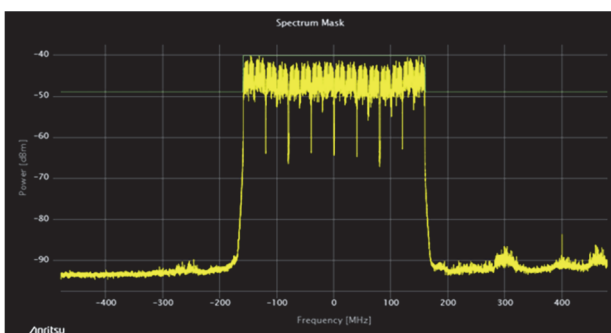


Figure 12 Example of Spectrum Measurement for 320-MHz bandwidth signal

9 Conclusions

We developed an update option for the MT8862A to support the IEEE 802.11be 320-MHz bandwidth for future mainstream RF measurement requirements. This MT8862A update is an industry-first supporting IEEE 802.11be RF measurements using the Network mode. With this update, the MT8862A provides a seamless RF measurement environment for 320-MHz bandwidth signals.

By providing measurement solutions supporting future new standards and features, Anritsu is continuing its efforts to meet the demands of users developing and manufacturing WLAN devices, helping widespread adoption and development of WLAN devices.

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Table 1 MT8862A Wireless Connectivity Test Set key specifications with 802.11be 320-MHz Bandwidth Option installed

Receiver	Frequency range	2.4 to 2.5 GHz, 5.0 to 6.0 GHz, 6.0 to 7.3 GHz
	Frequency setting resolution	1 Hz
	Level setting range	−65 to +25 dBm
	Level setting resolution	0.1 dB
	Level accuracy	Measurement conditions: CW, Measurement Bandwidth: 320 MHz, 20° to 30°C, Input signal below setting level and excluding linearity error effects, After calibration $5.785 \text{ GHz} \leq \text{Frequency} \leq 7.3 \text{ GHz}$ $\pm 0.7 \text{ dB}$ ($-30 \text{ dBm} \leq \text{Setting level} \leq +25 \text{ dBm}$) $\pm 1.0 \text{ dB}$ ($-50 \text{ dBm} \leq \text{Setting level} < -30 \text{ dBm}$)
Transmitter	Frequency range	2.4 to 2.5 GHz, 5.0 to 6.0 GHz, 6.0 to 7.3 GHz
	Frequency setting resolution	1 Hz
	Level setting range	−120 to 0 dBm
	Level setting resolution	0.1 dB
	Level accuracy	Output setting: CW 20° to 30°C Output level: $\geq -110 \text{ dBm}$, After calibration $\pm 1.0 \text{ dB}$, $\pm 0.7 \text{ dB}$ (typ.) ($2.4 \text{ GHz} \leq \text{Frequency} \leq 2.5 \text{ GHz}$) $\pm 1.3 \text{ dB}$, $\pm 1.0 \text{ dB}$ (typ.) ($5.0 \text{ GHz} \leq \text{Frequency} \leq 6.0 \text{ GHz}$) $\pm 1.3 \text{ dB}$, $\pm 1.0 \text{ dB}$ (typ.) ($6.0 \text{ GHz} < \text{Frequency} \leq 7.3 \text{ GHz}$)
	Signal purity	Harmonics: $\leq -25 \text{ dBc}$

Table 2 MX886200A WLAN Measurement Software key specifications with 802.11be 320-MHz Bandwidth Option installed

Frequency range	Frequency range	2.4-GHz band: 2412 to 2484 MHz 5-GHz band: 5180 to 5885 MHz 6-GHz band: 5955 to 7115 MHz
Amplitude setting	Input level range	−50 to +25 dBm
	Input level accuracy	After calibration, 20° to 30°C $\pm 0.7 \text{ dB}$ ($-30 \text{ dBm} \leq \text{Input level} \leq +25 \text{ dBm}$) $\pm 1.0 \text{ dB}$ ($-50 \text{ dBm} \leq \text{Input level} < -30 \text{ dBm}$)
	Bandwidth	320/160/80/40/20 MHz (802.11be)
EVM	Setting range	−20 to +25 dBm
	Residual EVM	OFDM (802.11be 320-MHz bandwidth) $< -44 \text{ dB}$ (nom.) ($-10 \text{ dBm} \leq \text{Input level}$, Average of 20 packets, Channel estimate: Multi packet, MCS13)
RF Signal generator	Level setting range	−120 to 0 dBm (Aux Out Connector) −120 to 0 dBm (Main 1/2 Connector, Frequency $\leq 6 \text{ GHz}$ and Channel Band 2.4 GHz/5 GHz) −120 to −5 dBm (Main 1/2 connector, Frequency $> 6 \text{ GHz}$ or Channel Band 6 GHz)
	EVM	802.11be $\leq -40 \text{ dB rms}$ (nom.) (5180 to 5885 MHz, 0.8 μs GI, Channel bandwidth 80 MHz, 20° to 30°C) $\leq -41 \text{ dB rms}$ (nom.) (5250 to 5815 MHz, 0.8 μs GI, MCS13, Channel bandwidth 160 MHz, 20° to 30°C)

Publicly available

Development of MU104014B for DCI Coherent Transmission Quality Measurement

Tomohiro Ito, Ittetsu Kaji, Eizo Hiraiwa, Atsushi Saegusa

[Summary]

Computationally intensive services such as generative AI are driving the need for larger data centers but limited building space in urban areas as well as the need to spread risk from large-scale disaster favors regionally distributed data center locations. Distributed Data Center Interconnects (DCI) are switching increasingly from conventional high-cost WDM lines to lower-cost 400ZR and OpenZR+ connections. Consequently, we have developed a network tester supporting 400ZR and OpenZR+ optical transmission quality measurements.

1 Introduction

Construction of data centers and metro networks is progressing rapidly, driven by factors such as generative AI, fast adoption of cloud services, and Digital Transformation (DX) promotion. In particular, extant data centers face difficulties in significantly increasing data processing capacity due to constraints on power supply and air-conditioning systems. Consequently, adding distributed data centers is an effective means to scale-up processing capacity, leading to increased demand for Data Center Interconnects (DCI) between different sites (Figure 1).

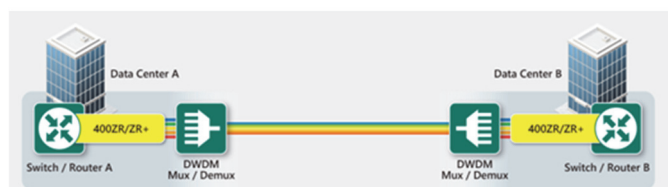


Figure 1 DCI Network

WDM lines owned by major telecommunications carriers have been used widely for DCI, but the high line usage fees are an issue. To cut costs, the OIF standards organization specifies installation of 400ZR-OIF pluggable coherent transceivers in routers and switches. Although 400ZR was intended for applications with transmission distances of less than 120 km, transmission beyond 120 km is achieved using the OpenZR+ standard resulting from combination with OpenROADM specifications (Table 1).

Table 1 400ZR vs OpenZR+

	400ZR	OpenZR+
Application	Edge DCI	Regional/Long-haul DCI
Standard	OIF 400ZR	OpenZR+ MSA
Client interface	400GbE	100 to 400GbE
Reach@400G	<120 km	>120 km
Line interface	400G	100/200/300/400G
SD-FEC	CFEC	OFEC
Maximum power consumption	<20 W	>20 W

Against this background, we developed the updated 400G (QSFP-DD) Multirate Module MU104014B for the MT1040A (Figure 2), which is a measurement instrument for evaluating various networks, including Ethernet, OTN, and CPRI. This module supports 400ZR and OpenZR+ transceivers and can evaluate optical transmission quality and 400 Gbps Ethernet transmissions. The MU104014B is based on the existing MU104014A supporting 400G Ethernet, and adds the following features:

- Effective cooling

Optical transceivers compatible with 400ZR and OpenZR+ have high power consumption and generate significant heat. Therefore, we enhanced the air cooling system and implemented measures to prevent failures caused by the heat generated by the optical transceivers themselves.

- Support for OpenZR+ optical transceivers

We enabled frequency-grid, output-wavelength, and output-power settings. The wavelength can be set by channel, frequency, or wavelength, according to the user's needs.

- Measurement of coherent transmission quality

We enabled reading of optical transmission quality results measured by optical transceivers, allowing time-series

display and storage.

- Evaluation of multiclient transmission qualities

OpenZR+ not only supports 400G Ethernet, but also enables 100G Ethernet multiplexed transmission. Therefore, we developed applications supporting 1×100, 2×100, and 4×100 Gbps.



Figure 2 MT1040A External view

2 MU104014B 400G (QSFP-DD) Multirate Module

In addition to an RJ45 interface, the MU104014B has electrical interfaces for various transceivers meeting SFP28, QSFP28, and QSFP-DD standards. It supports protocol tests for Ethernet (400GbE, 100GbE, 40GbE, 25GbE, 10GbE, 1GbE, 1000MbE, 100MbE, 10MbE), as well as for OTN, FC, and CPRI transport networks. Figure 3 shows the external view.

It also has a Sync Clock Out connector outputting a trigger signal for an external optical waveform sampling oscilloscope to monitor optical waveforms output from optical transceivers connected to various interfaces.

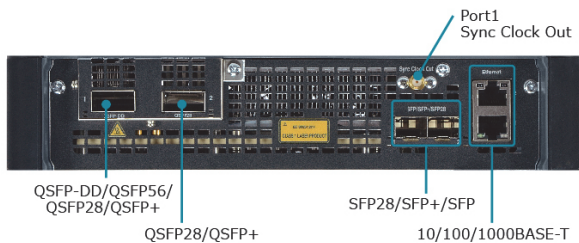


Figure 3 MU104014B External view

Two MU104014B modules can be installed in the MT1040A (Figure 4), for evaluating end-to-end bidirectional communications, and testing Network equipment's performance by connecting client side and line side to the module.



Figure 4 MU104014B Dual-module configuration

3 Hardware

As described earlier, 400GZR and OpenZR+ optical transceivers require high cooling performance due to their significant power consumption and heat generation. At the same time, since the MU104014B is used for on-site measurement targets, it must also be compact and quiet.

3.1 Target Performance

While OpenZR+ optical transceivers can generate more than 20 W of heat as mentioned in the previous section, they are still in the development stage, and the maximum power consumption has not been specified. Therefore, we set 23 W as the target based on datasheets for many OpenZR+ transceiver models and aimed for cooling performance that would not trigger temperature alarms even at an ambient temperature of 40°C.

The operating temperature in environments using OpenZR+, such as data centers, is expected to be 25°C or lower. However, some locations can become hot locally due to heat exhaust from equipment racks and other sources. To ensure sufficient margin in cooling performance even when used in such locations, we designed the system assuming an operating temperature environment of 40°C.

3.2 Increasing Cooling Efficiency

Optical transceiver cooling capacity depends on the QSFP-DD connector housing and the fan airflow volume. Since the MU104014B uses a stacked module configuration, fans cannot be installed on the back side. Therefore, we enlarged the heat sink on the housing upper surface and fan, and increased the fan rotation speed. These improvements, almost tripled the airflow compared to existing products, making it possible to sufficiently cool hotter Open ZR+ module across a wide ambient temperature range. Figure 5 shows the MU104014B cooling configuration.

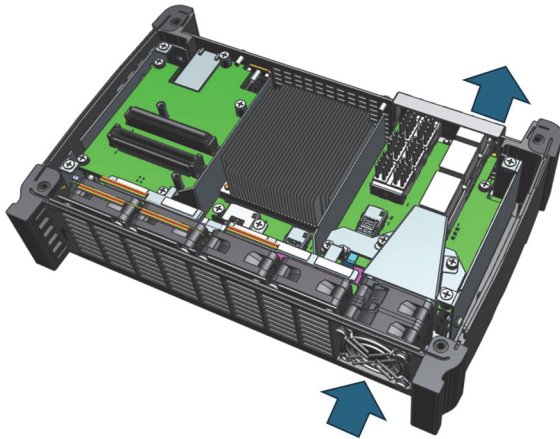


Figure 5 MU104014B Cooling configuration

3.3 Optimizing Fan Control

Increasing the fan maximum rotation speed as described in the previous section, increased the noise compared to existing products. Therefore, it was necessary to control and optimize the fan speed according to the optical transceiver temperature. However, feeding-back the readings from the optical-transceiver built-in temperature sensor to the fan control could destabilize temperature control.

Therefore, we added a dedicated sensor to monitor the MU104014B intake air temperature and achieved stable transceiver temperature and noise reduction by controlling the fan speed based on that temperature.

3.4 Testing Fan rpm and Transceiver Temperature

We investigated OpenZR+ temperature requirements and established the target temperature to 70°C. Controlling the fan speed based on ambient air temperature required investigating ambient air temperature and optical transceiver heat generation. For this purpose, we placed a QSFP-DD heat-generating module in MU104014B and monitored the temperature of the thermometer inside the heat-generating transceiver while in a 25°C constant-temperature chamber (Figure 6).

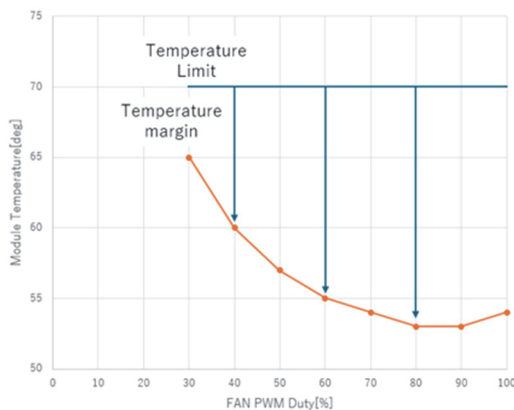


Figure 6 Fan rpm vs temperature margin

When the ambient air temperature rises by $\Delta^{\circ}\text{C}$, the above-described temperature margin decreases by $\Delta^{\circ}\text{C}$. We designed the fan control table so the fan speed is not increased more than necessary while ensuring the temperature margin does not reach zero (Figure 7).

The control table was designed with four fan speed modes: Low, Middle, High, and Extra. Each mode was controlled to change according to an external temperature threshold. We implemented hysteresis for the thresholds, so the higher-speed transition occurs at the temperature threshold $+2^{\circ}\text{C}$ and the lower-speed transition occurs at the temperature threshold -2°C . This prevents oscillations caused by small temperature changes.

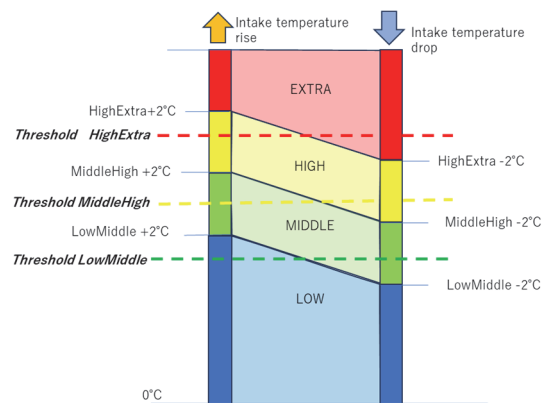


Figure 7 Fan control table

3.5 Safety Measures and Optimum Cooling Settings

Any deviations from the temperature model during the above-described verification risks damaging the optical transceiver due to insufficient cooling. Therefore, we designed a system to: monitor optical transceiver temperature alarms; power-down and shut-down the optical output at an alarm; and display a log of the temperature alarm occurrence.

Furthermore, we designed the system so that excess or deficient cooling caused by differences from the temperature model can be optimized to match the transceiver by providing a mode where users can switch the above-described fan speed modes manually.

3.6 Competitor Comparison

We inserted the QSFP-DD heat-generating transceiver, varied the ambient temperature, and compared the cooling performance using a competitor's measuring instrument. Since the competitor's instrument does not control the fan, the module temperature increased linearly versus outside

air temperature. In comparison, the MU104014B maintained a constant module temperature even in high-temperature environments due to the improved cooling efficiency and optimized fan control.

While current optical transceivers operate at appropriate temperatures using competing test instruments, the MU104014B demonstrate sufficient cooling capacity even with higher-power products.

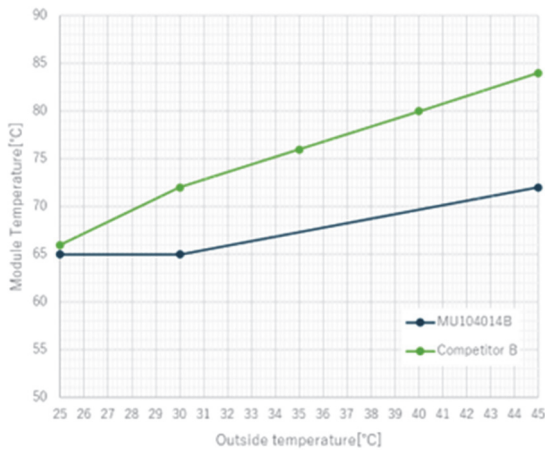


Figure 8 Module temperature vs ambient temperature

4 Additional MU104014B Functions

This section explains functions added to the MU104014B module to support OpenZR+.

4.1 Coherent Mode Setting

OpenZR+ transceivers support multiple applications, including 400ZR DWDM DP16QAM ZR applications, and ZR+ ZR400-OFEC-16QAM applications. When inserting and starting an optical transceiver, a number assigned to each application (APPSEL number) must be assigned to the transceiver. As a result, users must find the APPSEL number to activate up to 15 applications. For user convenience, we added a coherent mode setting allowing direct selection of 400ZR and OpenZR+ applications without knowing the APPSEL number (Figure 9).

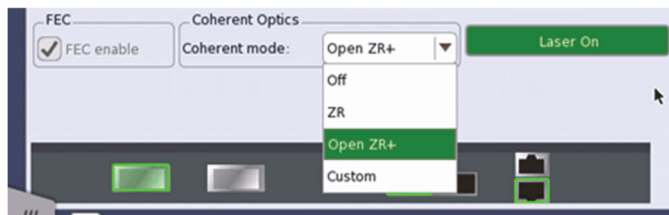


Figure 9 Coherent mode setting screen

Currently, 400ZR and OpenZR+ are the mainstream applications in use, and we have made it possible to select these

directly from a menu. Coherent transmission standards are still in the development stage, and vendor-specific implementations and new standard applications are also emerging, so we added a Custom mode setting to the menu for future-proof support. This Custom mode is designed to allow users to set the application to use while checking the contents of the transceiver applications in a list (Figure 10).

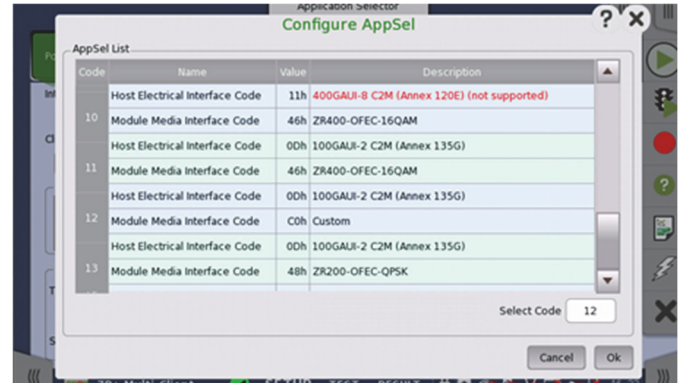


Figure 10 Custom application dialog

4.2 Tunable Settings

400ZR and OpenZR+ are used by setting wavelength and output power for connection to the WDM transmission equipment. The setting method is defined in the Common Management Interface Specification (CMIS), and can be changed by specifying the channel (offset) and grid from the 193.1-THz reference channel.

<Calculation Example (Frequency)>

When Grid = 75 GHz and Channel = 24

$$193.1 + (24 (\text{Channel}) \times 0.025 (\text{Grid})) = 193.7 \text{ THz}$$

On the other hand, since users often specify frequency or wavelength for transmission equipment, we added a screen (Figure 11) for three setting types: frequency, wavelength, and channel:

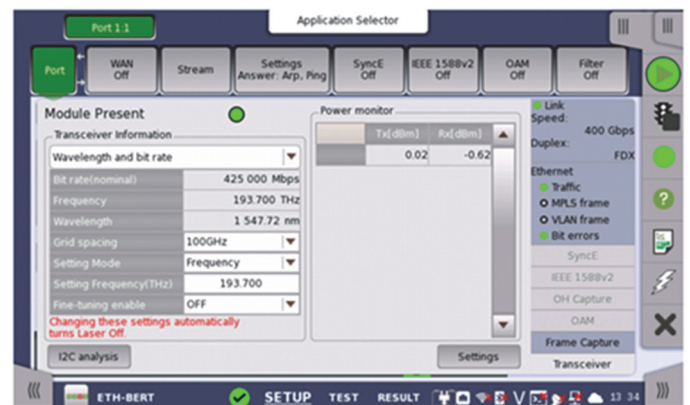


Figure 11 Output frequency (wavelength) setting screen

4.3 Measuring Coherent Transmission Quality

The MU104014B captures the following internal data from 400ZR and OpenZR+ transceivers for on-screen display.

- Rx bits, Rx frames
- Tx power, Rx power
- OSNR (Optical Signal to Noise Ratio)
- PreFEC BER (Uncorr. Frames)

Capturing information from the transceiver at minimum intervals of 1 second and logging it as measurement results confirms the relationship between the throughput of the 400GbE client signal and coherent-signal polarization or power variations. Additionally, exporting results to a CSV file facilitates detailed offline analysis, such as trend analysis.

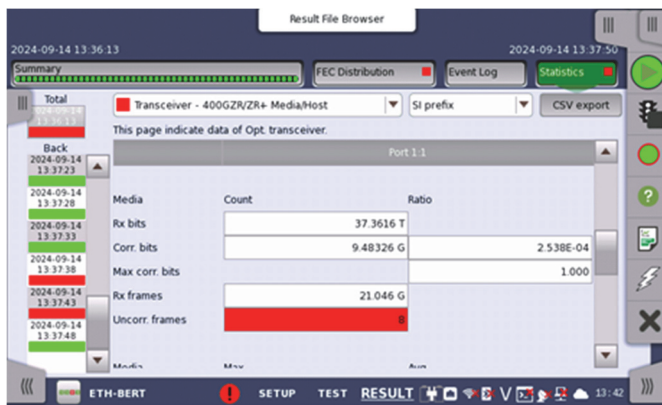


Figure 12 Measurement results screen

4.4 Technical Challenges Supporting 400ZR and OpenZR+

Multiple register operations must be performed in a specific sequence for 400ZR and OpenZR+ startup and tunable settings. Although these should work if configured according to the procedures described in the CMIS specification, the following problems have occurred due to differences between manufacturers and CMIS versions.

- Differences in waiting time after register operations
- Manufacturer-specific additional settings
- Differences in startup procedures depending on CMIS versions

We confirmed the detailed specifications for manufacturer-specific operation with the manufacturers and incorporated them so as not to cause problems with the standard CMIS procedures. For different processes depending on versions, we established a system to switch configuration procedures after reading the transceiver supported version. This

enabled consistent operation even with transceivers from different manufacturers or with different versions.

5 Multi-Client Application Support

In addition to the MT1040A 400G Ethernet application, we developed the OpenZR+ Multi-Client BERT MU104014B-036 option, which evaluates multiplexed 100G Ethernet transmissions by adding a software license. In addition to 400GbE measurements, this application adds client mode settings for selecting 100GbE multiplexing.

While 400ZR transceivers only transmit 400GbE, OpenZR+ transceivers can not only transmit 400GbE but can also multiplex multiple 100GbE signals. Each of these transmitted signals is called a client, and multiple 100GbE clients are independent of each other. The OpenZR+ Multi-Client BERT application can select 4×100 GbE, 2×100 GbE, or 1×100 GbE client modes, and can configure and measure each 100GbE client individually. Displaying the number of client ports corresponding to the selected client mode makes it easier to identify the current configuration.

Also, since the types of supported client modes vary depending on the OpenZR+ transceiver, on-screen indications display which client modes are supported by the currently installed transceiver, making client mode configuration easier.

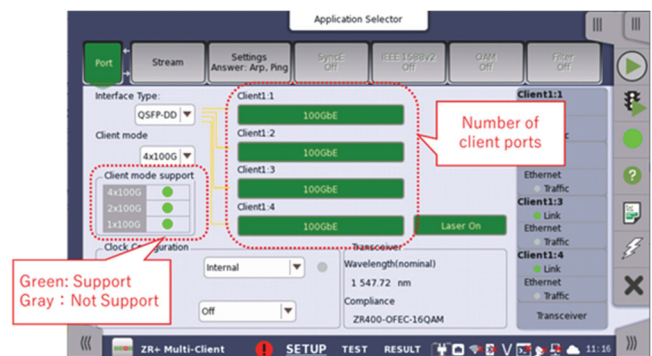


Figure 13 Multi-Client application

6 Automated rOSNR Test System using MU104014B Module and MS9740B

Development of the MU104014B module facilitated configuration of a system to automatically measure the required Optical Signal-to-Noise Ratio (rOSNR) of 400ZR and OpenZR+ transceivers.

This system uses an Amplified Spontaneous Emission (ASE) light source as the noise source, and the amount of

noise is adjusted by a variable optical attenuator. An ASE light source amplifies and emits spontaneously emitted light, and can be used to evaluate how systems and components perform in the presence of noise because it has broadband, low-coherence characteristics.

The rOSNR is measured by monitoring the Pre-FEC BER values of the 400ZR and OpenZR+ transceivers inserted into the MU104014B, and the OSNR values received by the Optical Spectrum Analyzer MS9740. To determine the OSNR precisely, the optical attenuator is adjusted while monitoring the measured values from the MU104014B and MS9740B.

This can be utilized by users examining multi-vendor connections, as well as by users with single-vendor setups, for verifying connectivity between transceivers with different revisions and for quantitative evaluation of transmission resilience for each transmission channel.

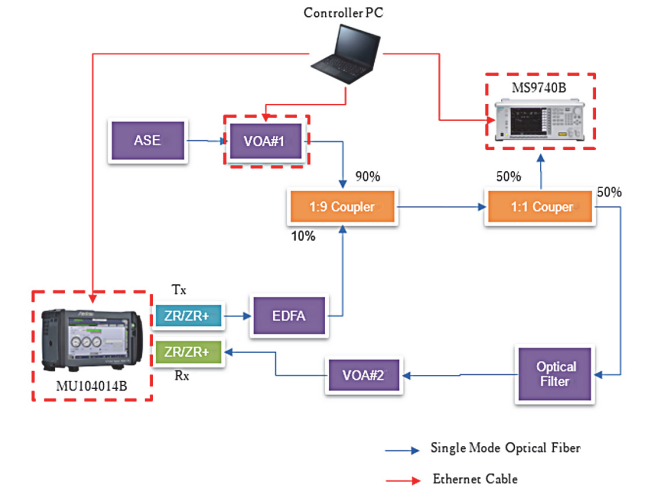


Figure 14 rOSNR Test system setup

7 Conclusions

We have developed test equipment for evaluating 400ZR and OpenZR+ transceivers being adopted for connections between distributed data centers. 400ZR and OpenZR+ are transitioning from the verification to the operational phase, and this test equipment is being utilized by various companies for network deployment and verification.

Anritsu is continuing to contribute to improving the quality of future network evolution.

References

1) OIF: “Implementation Agreement OIF-400ZR-02.0”, Nov, 2022
2) Open ZR+ MSA: “OpenZR+ MSA Specification, version 1.0” Sep, 2020
3) OIF: “Implementation Agreement OIF-CMIS-05.2” Apr, 2022
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Publicly available

Development of High-Performance OFDR System with Phase Integration Coherence Compensation

Takanori Saitoh, Akiko Tada

[Summary]

We devised a Phase Integration Coherence Compensation (PICC) method as an essential technology for OFDR systems and developed an OFDR system using this technique. This unique Anritsu PICC OFDR system can generate virtual reference interferometer signals with arbitrary delay lengths to help maximize the coherence of interference signals reflected from any location. Furthermore, high-quality OFDR spectra with coherence compensation across the entire measurement range can be captured by dividing the measurement distance (length) and implementing coherence compensation by generating reference interferometer signals with optimal delay lengths for each divided region. Additionally, we developed a strain and temperature distribution measurement system that calculates optical-fiber elongation and compression using this OFDR spectrum, enabling high-precision, real-time measurement of strain and temperature distribution over a 400-m optical fiber.

1 Introduction

Progressive aging of older large infrastructure, such as bridges, tunnels, and dams reaching the end of their design life, as well as social infrastructure, such as roads and railways, makes monitoring structural health increasingly important. Traditionally, strain is measured using a Fiber Bragg Grating (FBG)¹⁾ to monitor changes in the reflection profile of an optical fiber core caused by changes in the fiber refractive index resulting from fiber elongation and compression. As a result, attaching an FBG to the measurement target can monitor the target strain. Since an FBG is just several millimeters long, they are attached to monitor the structure only at key locations, and structural strain is not measured at locations with no attached FBG.

However, the monitored area could be increased greatly if, instead of making ‘point’ measurements using an FBG, an optical fiber itself could be used a strain sensor to measure the strain distribution along the full fiber length. Therefore, methods such as Distributed Acoustic Sensing (DAS)²⁾, Brillouin Optical Time Domain Reflectometry (B-OTDR)³⁾, Brillouin Optical Correlation Domain Analysis (B-OCDA)⁴⁾, and Optical Frequency Domain Reflectometry (OFDR)⁵⁾ have been developed to measure strain distribution along the full length of optical fibers. While DAS can detect impact-type strain over a range of several tens of kilometers, it cannot measure static strain, and its distance resolution is low at several meters. B-OTDR and B-OCDA can measure

strain over a range of several kilometers, but their strain resolution is approximately $100 \mu\epsilon^*$, which is low sensitivity.

OFDR has a measurement range of about 100 m, which is shorter than the other methods, but it is superior in terms of strain and distance resolutions. Anritsu is experienced in developing a 3D shape measurement system^{6), 7)} based on the OFDR method. Consequently, we leveraged our existing technological assets to develop an OFDR system to measure strain and temperature distribution, which is expected to see growing future demand.

When measuring the strain and temperature distribution using the OFDR method, Rayleigh scattered light generated in the optical fiber is detected by an interferometer, but its polarization differs depending on the position within the optical fiber. Therefore, polarization compensation is required. Also, coherence compensation technology is required to detect interference signals beyond the coherence length of the light source. This article discusses OFDR principles, polarization-compensation methods, Anritsu’s proprietary phase integration coherence compensation method (patent pending), and methods for calculating strain and temperature distribution.

* $1 \mu\epsilon$ is the amount of strain when an object 1 m in length stretches by $1 \mu\text{m}$.

2 OFDR Principles

2.1 OFDR Optical System

Figure 1 shows the OFDR basic optical system.

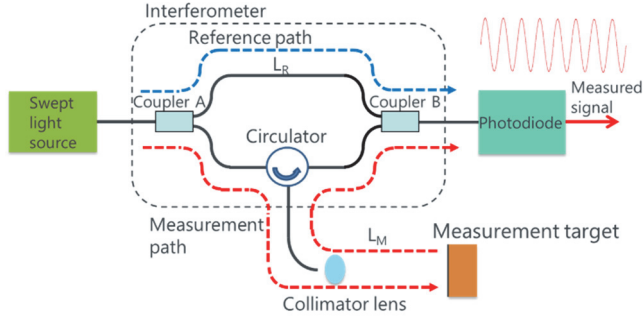


Figure 1 OFDR Optical system

The swept light emitted from the swept light source is split into two paths at coupler A; one path goes directly to coupler B (reference path), while the other path passes through a circulator, is emitted into free space through a collimator lens, and after being reflected by the target object, returns through the collimator lens and circulator to coupler B (measurement path). The swept light propagating through the reference path and the measurement path is combined at coupler B, and then converted to an electrical signal by a photodiode.

Assuming that the optical frequency of the swept light changes linearly with time, the optical frequency $\nu(t)$ at time t is expressed as:

$$\nu(t) = \nu_0 + k \cdot t \quad (1)$$

where ν_0 is the optical frequency as time 0 and k is the sweep speed (Hz/s). If we define the difference between the optical path length L_R of the reference path and the optical path length L_M of the measurement path as $\Delta L = |L_M - L_R|$, the difference in propagation time τ from the swept light source to the photodiode is:

$$\tau = \frac{\Delta L}{c} \quad (2)$$

where c is the velocity of light in vacuum.

If we denote the electric field intensities of the swept light passing through the reference path and the measurement path as E_R and E_M , respectively, the interference intensity P detected by the photodiode is:

$$\begin{aligned} P &= |\vec{E}_R + \vec{E}_M|^2 \\ &= |E_R e^{i2\pi\nu(t) \cdot t} + E_M e^{i2\pi\nu(t-\tau) \cdot t}|^2 \quad (3) \\ &= |E_R|^2 + |E_M|^2 + 2E_R E_M \cos\left(2\pi \frac{k\Delta L}{c} t\right) \end{aligned}$$

If the DC component of the interference signal is cut off, only

the third term in the final expression of Eq. (3) is observed.

In other words, an electrical signal with a frequency of

$$f_I = \frac{k\Delta L}{c} \quad (4)$$

is output from the photodiode. Therefore, after sampling the electrical signal, if the OFDR spectrum is calculated by FFT (Fast Fourier Transform), a peak is detected at the position of f_I on the OFDR spectrum and the optical path difference ΔL can be calculated from Eq. (4).

The above assumes that the optical frequency of the swept light changes linearly. However, in reality, the optical frequency is swept nonlinearly. Therefore, linearization correction must be applied to the measurement signal either through sampling methods or by software processing. This process is called linearization or resampling.

2.2 Linearization

Figure 2 shows the OFDR optical system including linearization and polarization compensation systems. The swept light emitted from the swept light source is split into two paths at coupler C; one path enters the measurement interferometer like that shown in Figure 1, and the other path enters the reference interferometer for linearization.

The measurement interferometer has a built-in polarization switch. The half-frequency signal of the sweep signal output from the swept light source is input to the polarization switching port of the polarization switch, causing the polarization of the reference path to switch between 0° (P polarization) and 90° (S polarization) with each sweep. Polarization is compensated by adding the calculation results from the measurement interferometer signals of each polarization state. Also, to ensure that the swept light input the polarization switch has linear polarization, the connection between the swept light source and the polarization switch uses polarization-maintaining fiber.

The reference interferometer is a Michelson interferometer design using Faraday mirrors (FR) as reflectors; the light from each FR returns to coupler D in the same polarization state regardless of the delay length ΔL_{AUX} with the advantage of eliminating variations in the interference signal due to polarization fluctuations.

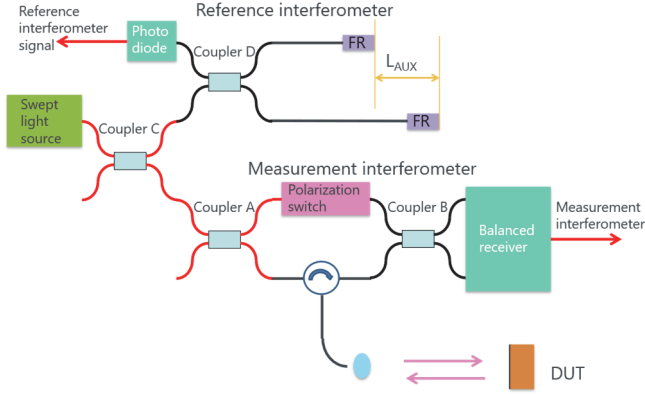


Figure 2 OFDR optical system including linearization system

From Eq. (4), the reference interferometer signal has the frequency

$$f_{AUX} = \frac{2k\Delta L_{AUX}}{c} \quad (5)$$

The frequency is doubled compared to Eq. (4) due to the round trip through ΔL_{AUX} . Also, since the FSR (Free Spectral Range) of the interferometer, that is, the optical frequency required for the interference signal to change by one cycle of a sine wave, can be expressed as:

$$FSR = \frac{c}{2\Delta L_{AUX}} \quad (6)$$

Eq. (5) can be expressed as:

$$f_{AUX} = \frac{k}{FSR} \quad (7)$$

The right-hand side represents 'how many times the sweep speed k corresponds to the FSR' or in other words, f_{AUX} can be described as the frequency of 'the interference signal that reaches a maximum each time the optical frequency is swept by FSR'.

Equation (4), which represents the frequency of the measurement interferometer signal, and Eq. (7), which represents the frequency of the reference interferometer signal, are both proportional to the sweep speed k . Therefore, it is possible to linearize the measurement interferometer signal from the change in frequency of the reference interferometer signal.

There are two linearization methods using either software or hardware^(6), 7). The biggest difference between the two is the interference signal sampling method where software linearization samples at equal time intervals, while hardware linearization samples at equal optical frequency intervals.

As described later, strain measurement requires multiple linearization cycles while shifting the measurement interferometer signal and reference interferometer signal

along the time axis, and this process can be simplified by sampling each interferometer signal at equal time intervals. Consequently, software linearization is better for strain measurement.

2.3 Software Linearization

Software linearization samples the reference interferometer signal and measurement interferometer signal simultaneously at equal time intervals using an A/D converter. Then, based on the reference interferometer signal, which serves as a ruler for optical frequency, the measurement interferometer signal is linearized so that it becomes linear with respect to optical frequency. The specific method is described below.

First, the phase of the reference interferometer signal is calculated. Next, FFT is performed on the reference interferometer signal, and the latter half of the FFT calculation result is replaced with zeros before performing inverse FFT on this. The phase angle of the complex number calculated by inverse FFT is the phase of the reference interferometer signal.

Since the calculated phase of the reference interferometer changes repeatedly between $-\pi$ and $+\pi$, unwrapping is performed by adding 2π to the phase at points where it jumps from $+\pi$ to $-\pi$, and conversely, by adding -2π to the phase at points where it jumps from $-\pi$ to $+\pi$, to calculate a continuous phase χ (Figure 3).

The continuous phase quantity increases by 2π each time the optical frequency is swept by the FSR of the reference interferometer, and is proportional to the optical frequency.

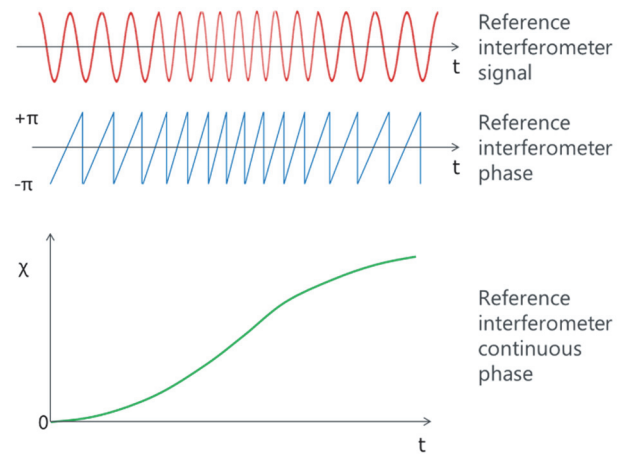


Figure 3 Reference interferometer signal phase calculation

Next, the continuous phase χ is divided equally, and the strength of the measurement interferometer signal at the

division points is extracted (green circles in center of Figure 4). The extracted data is not at equal intervals on the time axis, but becomes data sampled at equal intervals on the optical frequency axis. Sharp peaks are observed at frequency positions corresponding to reflection points on the OFDR spectrum, which is the FFT calculation result for this data.

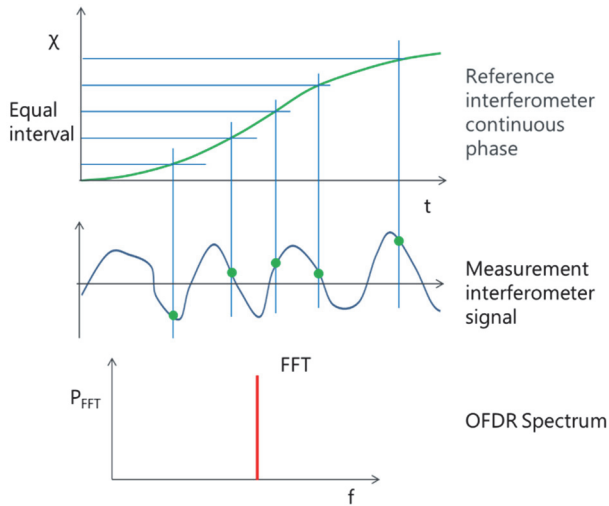


Figure 4 Data extraction

2.4 Anritsu Swept Light Source

Anritsu has developed and released a high-coherence optical frequency swept light source⁸⁾. Figure 5 shows the external appearance of the light source, which can sweep the optical frequency continuously in phase without mode hops over almost the entire sweep range. Table 1 lists the main specification of this light source.



Figure 5 AQB5500P External view

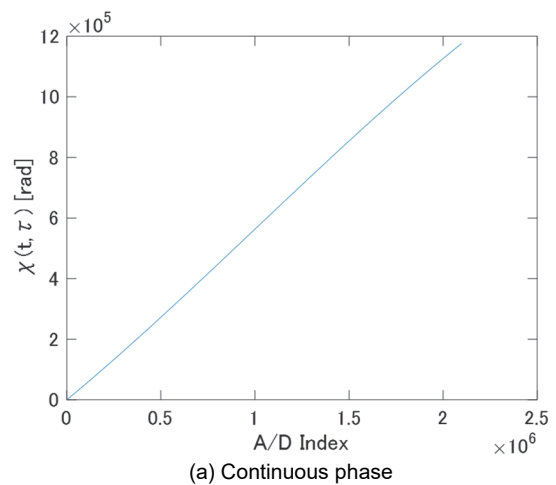
Table 1 AQB5500P Specifications

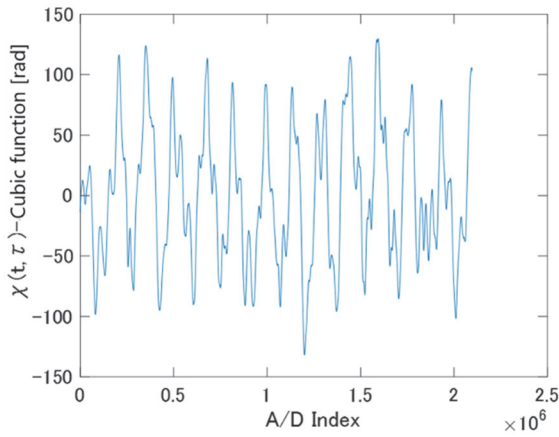
Center wavelength λ_c	1550 $\pm$ 5 nm (Fixed factory setting)
Wavelength sweep width W	10 to 50 nm
Sweep frequency f_R	150 $\pm$ 20 Hz (Fixed factory setting)
Optical output power	>10 dBm
Coherence length	>100 m
Trigger setting resolution	1 pm

The trigger signal output from this light source allows the timing between High and Low switching to be specified as wavelength. While sweeping from short to long wavelength, the trigger becomes High when the swept light wavelength reaches an arbitrarily set wavelength; while sweeping from long to short wavelength, the trigger becomes Low when the swept light wavelength reaches another arbitrarily set wavelength. It is possible to always sample in the same wavelength range by using this trigger signal to start sampling of the A/D converter.

The optical frequency sweep waveform of this light source is sinusoidal. Therefore, the optical frequency near the center of the sweep changes according to a cubic function.

Figure 6(a) shows the continuous phase measured using an Anritsu swept light source and a reference interferometer with a delay length of 83.5 m. It changes according to a cubic function, similar to the sweep waveform. Figure 6(b) shows the value obtained by subtracting the cubic function component from this continuous phase. The fluctuation amplitude of about 100 rad is typical of individual light sources and is reproducible.





(b) Continuous phase with cubic function component subtracted

Figure 6 Continuous phase of Anritsu light source

3 Phase Integration Coherence Compensation

The effect of linearization at the DUT position is maximum when the distance to the DUT matches the delay length of the reference interferometer. The effect decreases as the distance from this position increases due to the effect of the light-source coherence. When the distance to the DUT is limited, such as in OCT measurements, it is sufficient to prepare a reference interferometer with a delay length matching the position of the DUT. However, this becomes a significant problem in distance measurement systems where the measurement point position is undefined, or in strain-distribution measurement systems where measurement points span the entire optical fiber.

NTT has commercialized a software-linearization Phase Noise Compensation (PNC)⁽⁹⁾ method in which the linearization effect is maximum at integer-multiple positions of the reference interferometer delay length. While this method significantly expands the measurable strain-distribution range, the coherence decreases at half-integer multiple positions of the reference interferometer length, which are intermediate positions.

Anritsu has also developed an OFDR distortion distribution measurement system⁽¹⁰⁾ that compensates for coherence using hardware linearization, but requires addition of one more interferometer.

Therefore, this time, while maintaining the conventional optical system, we devised a PICC method that artificially generates reference interferometer signals with arbitrary delay lengths. By linearizing the measurement interferometer signal through software linearization from

the generated reference interferometer signal, it is possible to maximize coherence at any DUT position, which is especially useful for distortion distribution measurements.

The principle of the phase integration method is described below.

If the light phase is $\phi(t)$, the velocity of light is c , and the delay time of the reference interferometer with delay length L_{ref} is $\tau = 2nL_{ref}/c$, the reference interferometer signal phase $\chi(t, \tau)$ can be expressed as:

$$\chi(t, \tau) = \phi(t) - \phi(t - \tau) \quad (8)$$

Approximating Eq. (8) gives Eq. (9)

$$\chi(t, \tau) \simeq \frac{d\phi}{dt} \tau \quad (9)$$

And integrating Eq. (9) gives Eq. (10)

$$\phi(t) \simeq \frac{1}{\tau} \int \chi(t, \tau) dt = \Phi(t) \quad (10)$$

where the middle and right-hand terms are the phase integration.

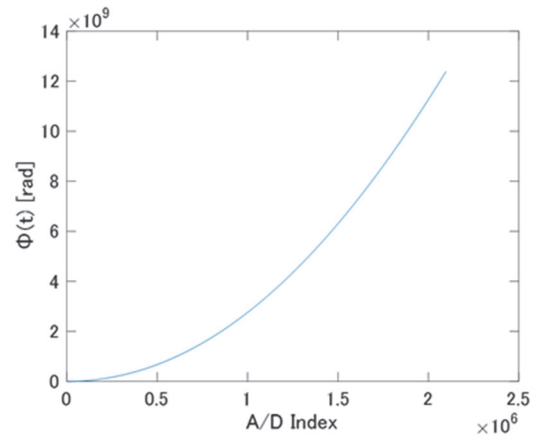


Figure 7 Phase integration

Figure 7 shows the phase integration $\phi(t)$ calculated from $\chi(t, \tau)$ shown in Figure 6. Since $\chi(t, \tau)$ is nearly linear, the phase integration takes the form of a quadratic curve.

The phase $\chi(t, \tau')$ of the interferometer signal that would be output by an interferometer with delay length L' , and delay time $t' = 2nL'/c$ can be calculated using Eq. (11)

$$\begin{aligned} \chi(t, \tau') &= \phi(t) - \phi(t - \tau') \\ &= \frac{1}{\tau} \int_{t_0}^t \chi(t, \tau) dt - \frac{1}{\tau} \int_{t_0}^{t-\tau'} \chi(t, \tau) dt \\ &= \frac{1}{\tau} \int_{t-\tau'}^t \chi(t, \tau) dt \end{aligned} \quad (11)$$

In other words, based on the reference interferometer signal with delay length L_{ref} , the reference interferometer signal with any arbitrary delay length can be calculated from the phase integration.

If FFT is performed after software linearization of the measurement interferometer signal based on the reference interferometer signal calculated by Eq. (11), an OFDR spectrum can be calculated where coherence is maximized at any arbitrary distance L' . As shown in Figure 8, at that time it is necessary to shift the linearization timing by the time $\delta t = 2n(L' - L)/c$ corresponding to the delay length difference between L' and L .

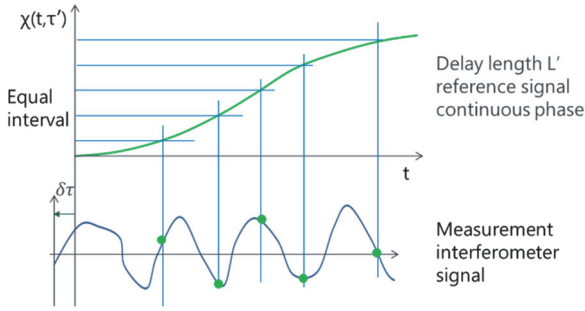


Figure 8 Delay length difference correction

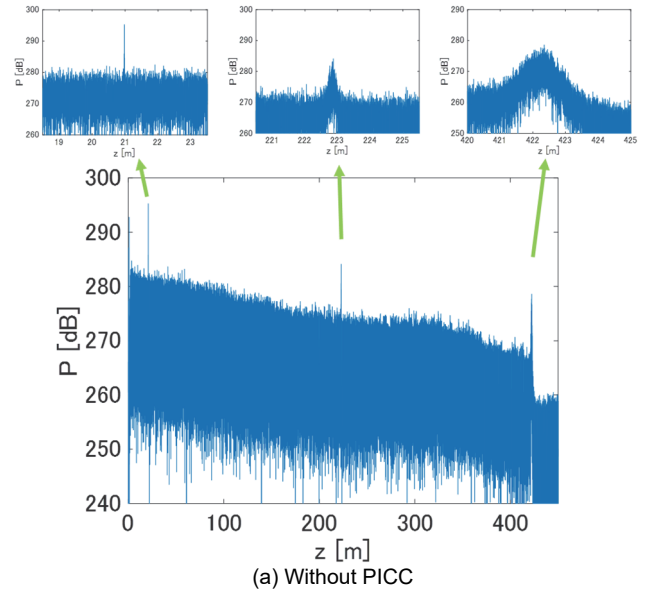
4 Measuring Reflection Peak using PICC OFDR

We configured an OFDR system using an Anritsu wavelength-swept light source with a sweep center wavelength of $1.55 \mu\text{m}$ and a sweep width 10 nm, a reference interferometer with a delay length of 83.5 m, and a high-speed A/D converter (TELEDYNE ADQ35) with a sampling frequency of 5 GS/s. This OFDR system configuration can monitor the reflection distribution from optical fibers as an OFDR spectrum over a range of approximately 450 m.

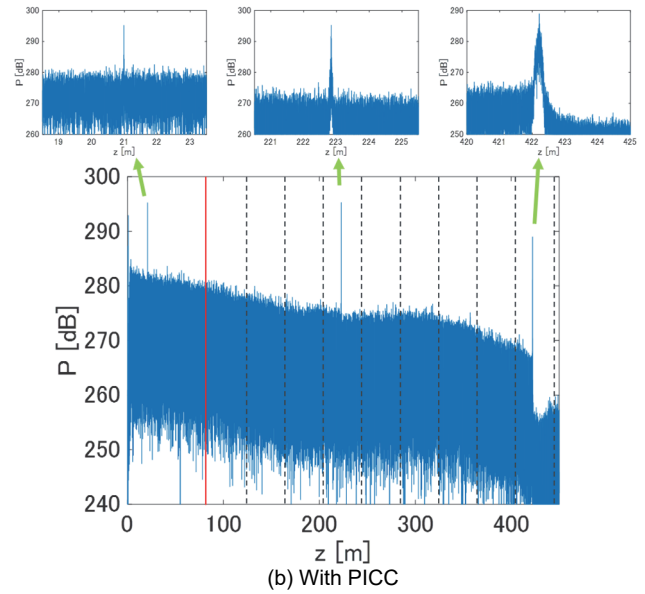
The reflection distribution and reflection peaks from the optical fiber connection points were observed when three optical fibers with lengths of 20 m, 200 m, and 200 m were connected in series. Figure 9(a) shows the OFDR spectrum without coherence compensation. Rayleigh scattering distribution in the optical fiber is observed in the range of 0 to 420 m, and strong reflection peaks are observed at the optical fiber connection points. The coherence length of the wavelength swept light source in air is about 100 m, and the round-trip distance that can be traveled without degraded coherence in an optical fiber is 30 m. Therefore, while the reflection peak at 20 m has a sharp shape, the reflection peaks at 220 m and 420 m are spread out by about 0.5 to 1 m due to degraded coherence.

Next, Figure 9(b) shows the OFDR spectrum with PICC, which was implemented at points every 40 m from the position of the reference interferometer delay length (red

line in figure). The OFDR spectrum in the figure is an integrated OFDR spectrum created by connecting the 0 to 103.5-m region of the OFDR spectrum measured without PICC and the regions of ± 20 m around the implementation points of the OFDR spectra measured with PICC performed every 40 m. All three reflection peaks were sharper, and the peak widths became several centimeters.



(a) Without PICC



(b) With PICC

Figure 9 OFDR Spectrum

5 Strain Distribution Measurement Method

The OFDR spectrum in Figure 9 shows the Rayleigh scattering light intensity distribution from the connected SM fiber between 0 m and 420 m at the base of the reflection peak. This Rayleigh scattering light spectral shape, which

appears to be white noise at first glance, reflects the refractive index distribution of the SM fiber. Since the refractive index distribution is determined during SM fiber manufacturing and is virtually unchangeable, this OFDR spectrum intensity is reproducible. While strain distribution measurement is possible using changes in the OFDR spectrum intensity, if, for example, shape changes in a 100-mm region are measured with 100- μm precision, the strain resolution is 1000 $\mu\epsilon$.

Therefore, to measure strain, the phase of the OFDR spectrum is observed rather than its intensity. The phase of the measurement interferometer signal observed by OFDR changes depending on the optical path from the light source to the reflection point. If the wavelength of the swept light is λ and the refractive index of the optical fiber is n , the phase of the measurement interferometer changes by 2π [rad] when the distance from the light source changes by $\lambda/2n$ (≈ 0.5 μm). Since phase can be measured with an accuracy of about 0.1 [rad], strain can be measured with an accuracy of 0.1 $\mu\epsilon$.

The actual calculation method is described below.

The measurement interferometer signal, which has been linearized after coherence compensation by the phase integration method described in the previous section, is subjected to FFT to calculate the OFDR spectrum. As an example, Figure 10 shows the real part (blue line) and imaginary part (red line) of the OFDR spectrum for a 10-cm region from 50 to 50.1 m. Unlike the OFDR spectrum intensity, this shape changes with slight external disturbances.

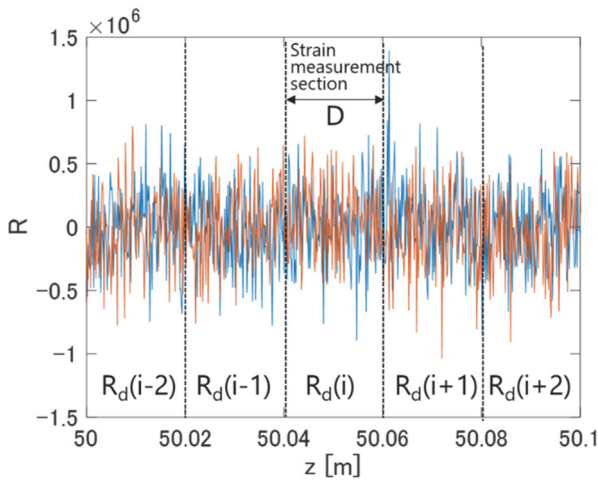


Figure 10 Real and imaginary parts of OFDR spectrum

As shown in Figure 10, the OFDR spectrum is divided into sections of desired length D . This section length becomes the distance resolution of the strain distribution measurement.

For example, when the section length $D = 20$ mm and the OFDR resolution $\Delta z = 0.446$ mm, there are 44 data points in each section.

Next, zero padding is applied to the data within the section length to make the number of data points a power of 2. This is done to implement the subsequent FFT effectively, but also to increase the number of data points and improve the resolution after FFT. If the number of data points after zero padding is N_{RDz} , then $Dz = \Delta z \times N_{RDz}$ becomes the calculated section length and is the parameter that determines the strain resolution as described later (Figure 11).

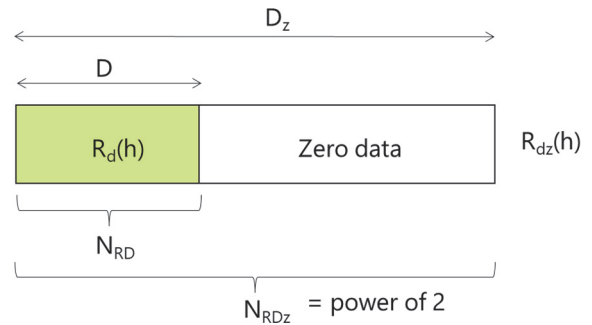


Figure 11 Zero padding

The zero-padded data is inverse-FFT transformed, and its real part is calculated as the frequency component of the phase change within the section.

The polarization-compensated M_d^{PS} is obtained by adding the frequency component intensity M_d^P calculated by the above method from the measurement interferometer signal measured in the P-polarization state, and the frequency component intensity M_d^S calculated similarly from the measurement interferometer signal measured in the S-polarization state.

If the frequency component intensity phase change (M_d^{PS}) is gradual, the low-frequency components become larger; if the change is rapid, the high-frequency components become larger. Additionally, if a section is compressed, the phase changes become denser, shifting the overall spectrum towards the high-frequency side; if a section is stretched, the phase changes become sparser, shifting the spectrum towards the low-frequency side.

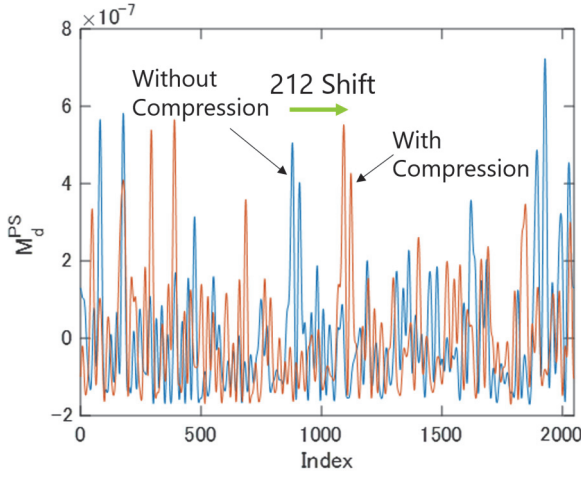
Figure 12 M_d^{PS} Shift

Figure 12 shows M_d^{PS} measured with no external force applied to the optical fiber (blue line) and with compression applied (orange line). There are 2048 data points after zero padding. Multiple peaks are seen in each state, but in this example, M_d^{PS} has shifted by 212 toward the high-frequency side due to compression.

A commonly used method for calculating the amount of shift in M_d^{PS} calculates the cross-correlation function S_{xc} and determines the position showing the maximum value as the shift amount. In other words, the value of m for which $S_{xc}(m)$ becomes maximum is taken as the shift amount.

$$S_{xc}(m) = \int g(t)f(t-m)dt \quad (12)$$

One method for calculating Eq. (12) is to perform FFT on the M_d^{PS} (Eq. (13)) for each state, multiply the results and perform inverse-FFT on that result.

$$S_{xc}(m) = FFT^{-1}\{FFT(g(t)) \times (FFT(f(t)))^*\} \quad (13)$$

This method calculates the cross-correlation rapidly. Figure 13 shows the cross-correlation during elongation and compression. The peak position of the cross-correlation shifts to the negative side during elongation and to the positive side during compression.

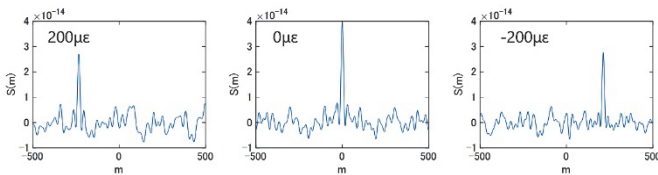


Figure 13 Cross-correlation function

When a section is compressed and the phase increases by 2π it means the phase rotation has increased by one complete rotation, meaning that its frequency component has increased by 1. Additionally, an increase in phase by 2π

means that the optical path length within the section has increased by $\lambda/(2n)$. During compression, not only does the fiber length become physically shorter, but also the refractive index increases. If the coefficient indicating this photo-elastic effect is denoted as ρ (≈ 0.78), the relationship between the shift amount m and the strain amount s is given by Eq. (14).

$$s = \frac{\delta D}{D} = \frac{m\lambda}{2nD_z\rho} = \frac{m\lambda}{2nN_{RDz}\Delta z\rho} \quad (14)$$

Additionally, since m is an integer, the strain resolution δs becomes

$$\delta s = \frac{\lambda}{2nN_{RDz}\Delta z\rho} \quad (15)$$

and the strain resolution improves as N_{RDz} is increased by zero padding.

Due to the nature of the cross-correlation function calculation, $-N_{RDz} < m < N_{RDz}$. Therefore, the maximum measurable strain s_{max} is:

$$s_{max} = \frac{\lambda}{2n\Delta z\rho} \quad (16)$$

and is about $1500 \mu\epsilon$ in this system. However, this is when the M_d^{PS} S/N is sufficiently large, and the typical maximum strain is 20% to 50% of this value.

The strain distribution applied to the optical fiber can be measured by performing the above calculations for each section.

The peak value in Figure 13 becomes smaller when the elongation and compression within the section length are non-uniform, or when there are many noise components in the frequency components, resulting in poor reproducibility. Therefore, the correlation coefficient, which is normalized by dividing the peak value by the square root of the sum of squares of M_d^{PS} , becomes an index of the quality of the OFDR spectrum.

The correlation coefficient was measured when a 400-m long SM optical fiber was connected. Figure 14 shows the difference in correlation coefficient with and without PICC. Without coherence compensation, the correlation coefficient deteriorates rapidly beyond the position of the reference interferometer delay length (85.3 m) where coherence is maximum. However, with PICC, the correlation coefficient is maintained above 90% through all regions of the connected optical fiber.

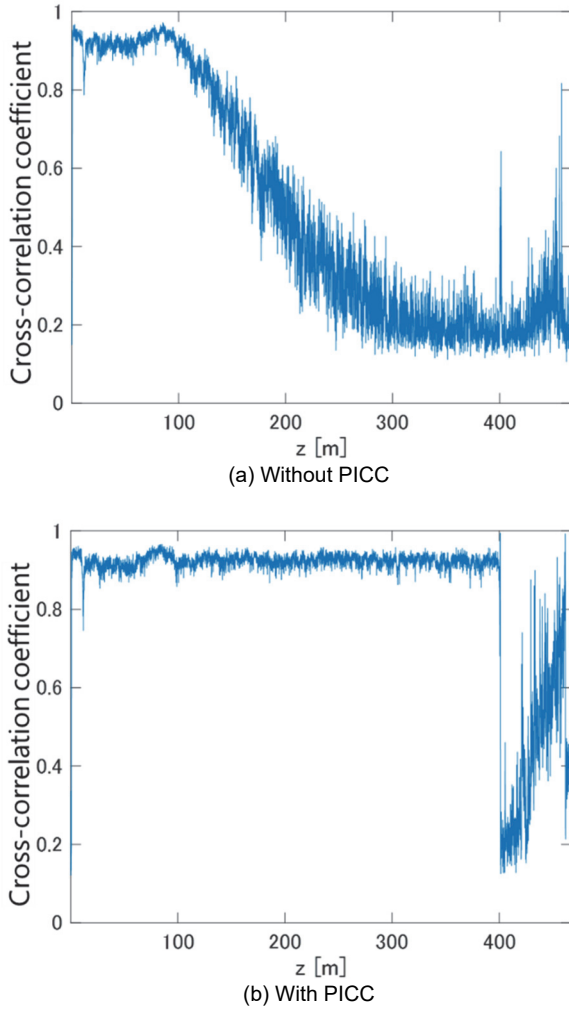


Figure 14 Cross-correlation function

6 Strain Measurement using PICC OFDR

6.1 Static Strain Distribution Measurement

Figure 15 shows the external view of the developed PICC OFDR-based strain distribution tester. The laptop workstation on the top of the tester has a built-in Graphics Processing Unit (GPU). The main unit under the workstation contains an Anritsu wavelength-swept light source (10-nm wavelength sweep width, 144-Hz sweep frequency), an interferometer (83.5-m reference interferometer delay length), and an A/D converter (2 channels, 5 GS/s sampling frequency). The number of A/D samples was set to 2 million and A/D sampling is performed over 1.7 nm near the approximate center of the 10-nm sweep width. The OFDR resolution is 0.446 mm, and the OFDR spectrum observation range is 450 m.

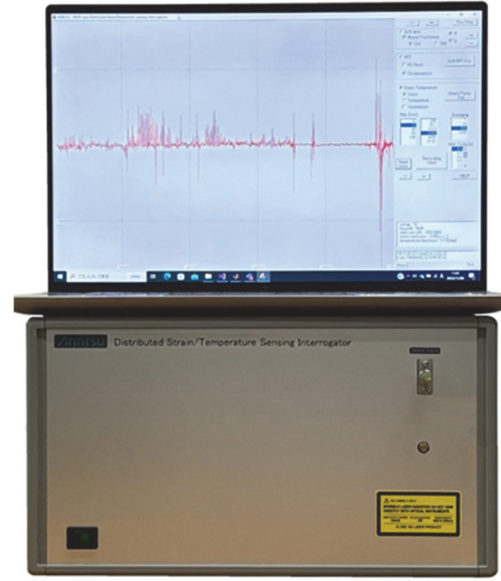


Figure 15 External view of strain distribution tester with PICC OFDR

To evaluate the static strain distribution measurement performance of the PICC OFDR strain distribution tester, the relationship between the strain value calculated by Eq. (14) and the actually applied value was measured using the optical fiber elongation/compression test rig shown in Figure 16. This test rig allows elongation of an optical fiber by fixing the SC connector parts of a 510-mm long optical fiber with SC connectors at both ends to a micrometer-equipped stage on an optical rail.

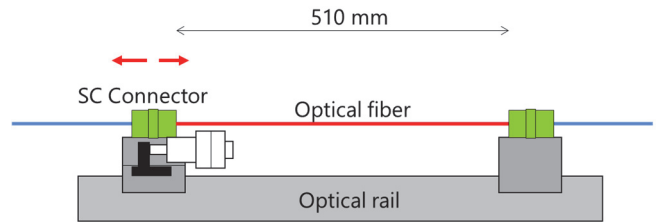


Figure 16 Optical fiber elongation/compression test rig

The actually applied strain value s_{App} is calculated from the micrometer value δl and the length l of the elongated section as:

$$s_{App} = \frac{\delta l}{l} \quad (17)$$

Figure 17 shows the strain distribution with this test rig connected to the ends of a 400-m long optical fiber while strain is applied measured with a spatial resolution of 100 mm and a strain resolution of 0.34 $\mu\epsilon$. Approximately 1500 $\mu\epsilon$ of tension was pre-applied as the reference state, and negative strain distribution was also measured from there by loosening the tension. Measurements were possible up to

+900 $\mu\epsilon$ on the elongation side and up to -800 $\mu\epsilon$ on the compression side, but when greater elongation or compression was applied, the cross-correlation deteriorated and normal strain distribution measurement was impossible. However, it reached 60% of the maximum measurable strain of 1500 $\mu\epsilon$ calculated from Eq. (17). This is because the PICC maintained the cross-correlation at a sufficiently large value even at a position of 400 m. Furthermore, without coherence compensation, the strain distribution could not be measured at all because it was buried in noise.

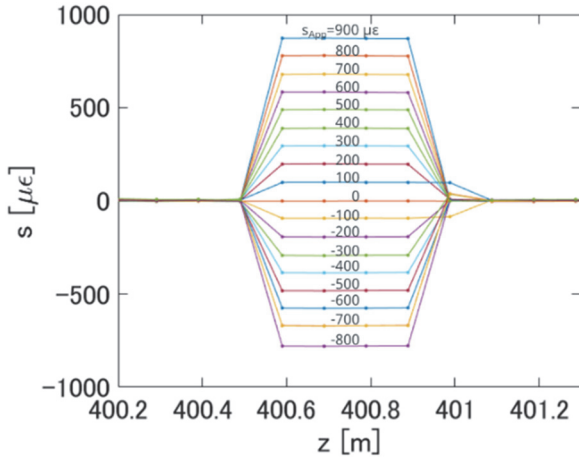


Figure 17 Strain distribution during strain application

Figure 18 shows the relationship between the applied strain s_{App} and the measured strain s . The two have a linear relationship, and measurement achieves high accuracy with an error of 1.3%.

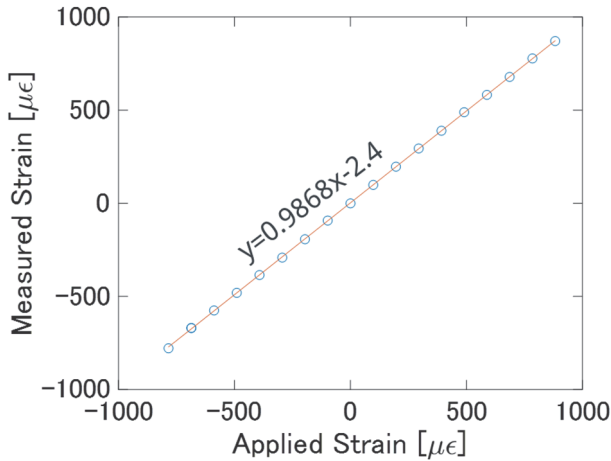


Figure 18 Relationship between applied strain amount and measured value

The relationship between spatial resolution and strain distribution was measured. Figure 19 shows the strain distributions measured with spatial resolutions set to 20, 50, 100, and 200 mm. As the spatial resolution becomes smaller,

the steep strain rise sections can be measured more accurately.

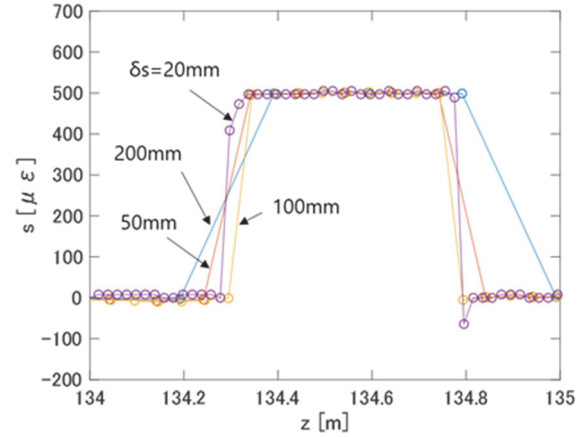


Figure 19 Spatial resolution dependency

Next, the relationship between strain resolution and strain distribution was measured. A fiber reel with a diameter of about 20 cm was placed in a location with a temperature gradient and the strain distribution due to thermal expansion was measured with a spatial resolution of 20 mm. Figure 20 shows the results measured with strain resolutions of 1.4, 5.3, and 23 $\mu\epsilon$. Naturally, the higher the measurement resolution, the more precisely strain changes can be measured.

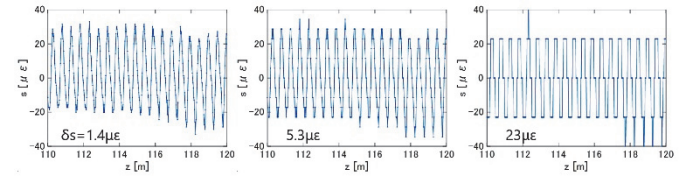


Figure 20 Strain resolution dependency

6.2 Dynamic Strain Distribution Measurement

The dynamic strain distribution was measured using the optical fiber elongation/compression test rig shown in Figure 16 operated with an automatic stage, which was programmed to repeatedly increase the strain linearly to 600 $\mu\epsilon$ over 2.4 s, pause for 0.2 s, and then decrease the strain to -600 $\mu\epsilon$ over 2.4 s. The strain distribution measurement settings were spatial resolution of 100 mm, strain resolution of 1.37 $\mu\epsilon$, phase integration coherence compensation interval of 40 m, and measurement frequency of 10 Hz.

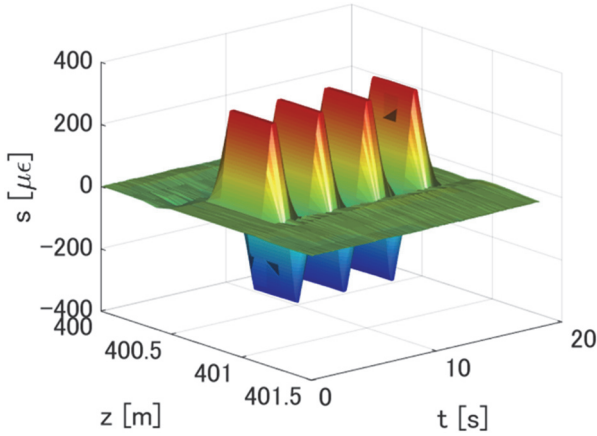


Figure 21 Dynamic strain measurement results

Figure 21 shows the dynamic changes in the measured strain distribution, indicating the applied dynamic strain is faithfully measured.

The blue line in Figure 22 indicates the temporal change in strain at a distance of 400.8 m, that is, the cross-section in the distance direction shown in Figure 21; the red line, indicates the strain applied from the automatic stage. The measured and applied strains match closely.

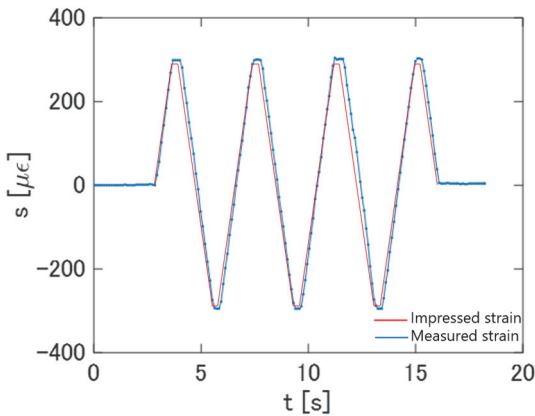


Figure 22 Temporal change in strain value at 400.8 m

Strain distribution measurement performs many FFT calculation cycles. Table 2 lists the number of FFT calculations for one strain distribution measurement when the number of A/D samples = 2 million, the OFDR resolution = 0.446 mm, the spatial resolution = 100 mm, the strain resolution = 1 $\mu\epsilon$ ($N_{RDz} = 2048$), and the PICC interval = 40 m. Ten real-time measurements per second are possible using the laptop built-in GPU built to process the enormous calculations.

Table 2 FFT Calculation Times

FFT Data Points	Calculation Times
2097152	58
2048	9000
4096	4500

The above calculation implements optimal linearization in all regions, but if the PICC implementation point is limited to one location, coherence compensation occurs only in its vicinity (about ± 20 m), and coherence deteriorates elsewhere. However, the measurement speed increases if the amount of computational processing is reduced by implementing limited PICC. Figure 23 shows the maximum value of the correlation function when PICC is implemented only at 400 m. The maximum cross-correlation value near 400 m exceeds 90%, but it decreases to 80% in the region from 100 to 300 m. However, the measurement speed was 20 Hz, which is twice as fast as when PICC is performed at 40-m intervals across the entire range.

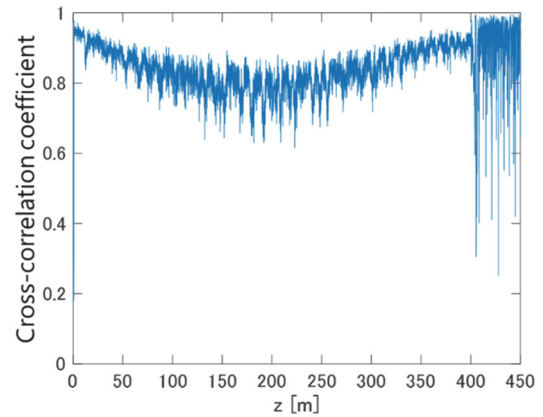


Figure 23 Cross-correlation coefficient when PICC applied only at 400 m

We created a cantilever (Figure 24) system capable of applying strain more rapidly than the test rig shown in Figure 16. This cantilever is a plastic plate with one end fixed, and optical fibers attached to the plate front and back surfaces, allowing strain to be applied to the fibers by flexing the plate.

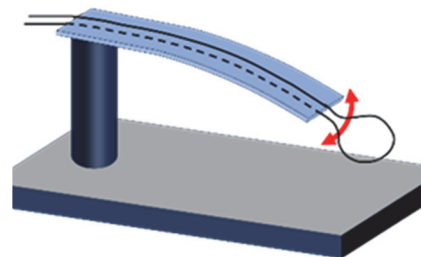


Figure 24 Cantilever

The strain distribution at the cantilever surfaces was measured by manually bending and then releasing the cantilever, causing it to vibrate. The strain distribution measurement conditions were set to a spatial resolution of 50 mm and a strain resolution of 1 $\mu\epsilon$. Figure 25 shows the temporal change in strain distribution, and Figure 26 shows the temporal change in strain at the center of the cantilever front and back surfaces. The detailed measurement captures how the strain at the front and back surfaces changes in a counter pattern during damping.

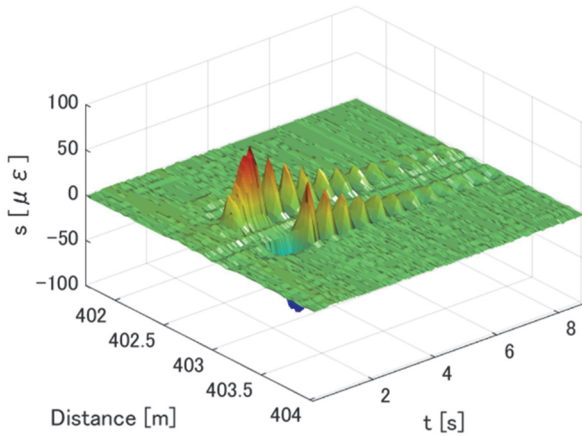


Figure 25 High-speed dynamic strain distribution measurement results

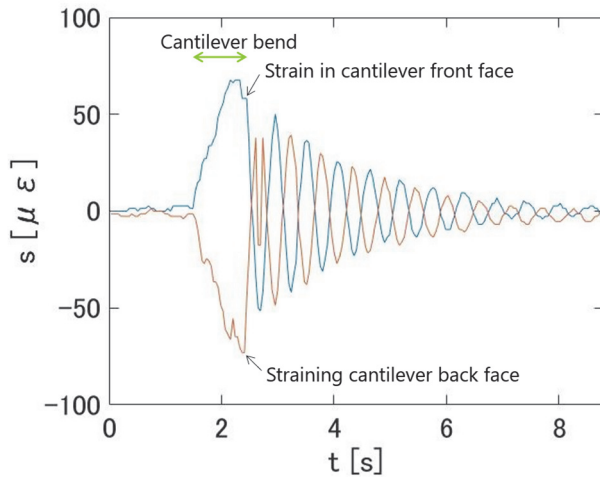


Figure 26 Strain at cantilever center front and back

7 Temperature Distribution Measurement using PICC OFDR

While the degree of optical-fiber elongation and compression can be measured at strain distribution measurement, it is impossible to distinguish whether the cause is due to external physical factors or thermal expansion.

Therefore, the measured strain assumes there is no thermal expansion. Conversely, if measurement assumes there is no strain, thermal expansion is causing elongation and compression and the temperature distribution can be measured from that strain. If the photo-elastic coefficient ρ for strain in Eq. (14) is changed to the thermoelastic coefficient ξ , the temperature distribution can be measured as:

$$T = T_{ref} + \delta T = T_{ref} + \frac{m\lambda}{2nD_z\xi} \quad (18)$$

where, T_{ref} is the temperature of the optical fiber when the frequency component reference value M_{dref}^{PS} was acquired.

Three fiber reels with optical-fiber lengths of 50, 30, and 20 m, each with a winding diameter of 12 cm were connected in series, and the temperature distribution was measured using this as a temperature-distribution sensor. As shown in Figure 27, device X with a heat source inside was placed on a wooden test table in the laboratory. The 50-m fiber reel was placed on the wooden test table (measurement point A), and the 30-m and 20-m fiber reels (measurement points B and C, respectively) were both placed on device X. The heat source inside device X was directly under measurement point C.

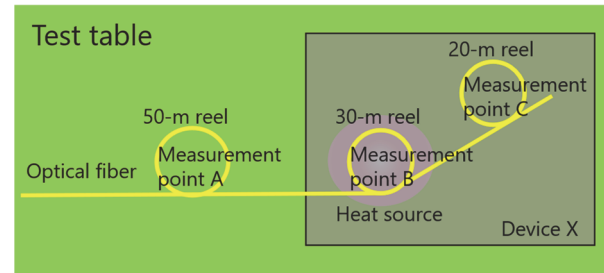


Figure 27 Temperature distribution sensor arrangement

Measurement of the temperature distribution was started at 20:00 in winter, and the power to device X was turned on at the same time. The overall temperature at that time was 22°C. The laboratory air conditioning stopped at 20:00 PM and restarted next morning at 07:00. The temperature resolution was set to 0.07°C, the spatial resolution to 50 mm, and the measurement time interval to 1 minute; measurements were continued for about 12 hours. Figure 28 shows the results.

Figure 28(a) shows the temperature distribution over 12 hours for all regions; (b) shows the temperature distribution at 1 minute and 5 hours after the measurement start; and (c) shows the temperature changes at the 25, 75, and 90 m points. The temperature at measurement point A (0 to 50 m) remained almost constant, but there was a temperature

increase of 3.5°C at measurement point B (50 to 80 m) and 3°C at measurement point C (80 to 100 m). The temperature increase at points B and C was completed in about 15 minutes after the start, and remained stable thereafter.

When the laboratory air conditioning started at 07:00 AM (11 hours after the measurement start), there was a temporary temperature drop of 0.6°C at all points. This is because cooler air rather than warm air is blown more strongly immediately after the air conditioning starts. The subsequent gradual cooling is thought to be due to cooling of device X by the air flow.

This test shows how the PICC OFDR system enables high-precision temperature distribution measurement over a wide range.

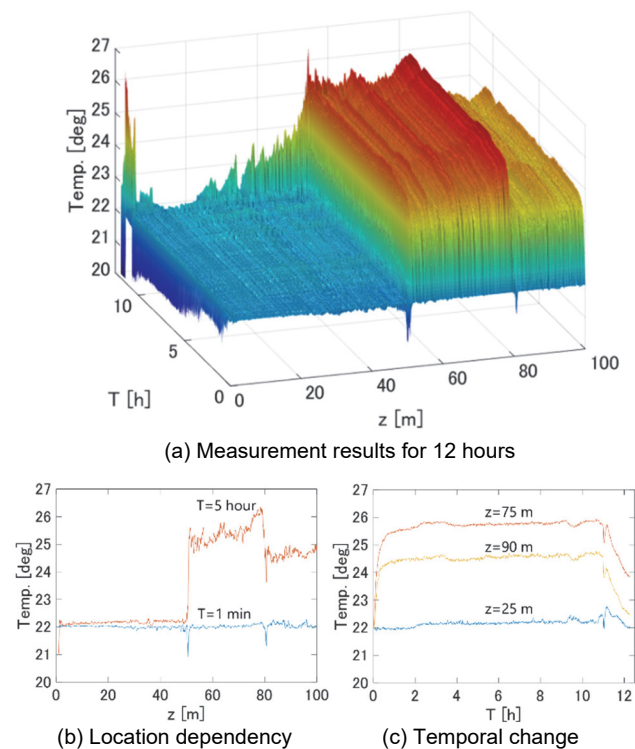


Figure 28 Measured temperature distribution results

While temperature distribution over a wide area can be measured relatively easily using thermal cameras, etc., the PICC OFDR method of measuring temperature distribution is useful for measurement targets that cannot be photographed from outside, such as batteries embedded in electric vehicles or underground power cables.

8 Conclusions

We devised a PICC method as an essential coherence compensation method for OFDR systems, and developed an OFDR system implementing this technique. This PICC OFDR system facilitates coherence compensation along the full fiber length, supporting high-quality measurement of OFDR spectra. Additionally, we developed a program to rapidly calculate the strain distribution of optical fibers from the obtained OFDR spectra, enabling real-time measurement of strain and temperature distributions.

Presented high-precision strain and temperature distributions measured using this PICC OFDR system demonstrate that it significantly improves OFDR performance.

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Development of World-First 100% Tablet Inspector using NIR Transmission Spectroscopy

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[Summary]

The use of inspection equipment for 100% product measurement is becoming increasingly widespread in pharmaceutical manufacturing. This trend supports the collection of production status data for active process control and helps eliminate accidental out-of-specification products. This article details the features, operating principles, basic configuration, and development points of an inspection system for 100% non-destructive inspection of tablet component quantities, cross-contamination, and foreign matter inclusion using transmission-type near-infrared absorption spectroscopy (NIR) that Anritsu is currently developing. It also introduces the technical approach to meet quality improvement requirements in pharmaceutical manufacturing.

1 Introduction

1.1 Market Background – Necessity for 100% Tablet Inspection¹⁾

Pharmaceuticals require strict quality control due to their direct impact on people's lives. They are manufactured in accordance with manufacturing management and Good Manufacturing Practice (GMP) standards established by each country.

Component analysis using High Performance Liquid Chromatography (HPLC) is utilized widely in pharmaceutical quality control. However, since it is a destructive test and requires long measurement time, some situations require sampling tests. Additionally, developing appropriate sampling plans that consider various conditions requires significant effort.

Due to these circumstances, there is a recent growing trend in introducing Process Analytical Technology (PAT), which monitors quality in real-time during the manufacturing process to improve production quality. NIR spectroscopy is an example of PAT that is widely known as a non-destructive analytical method to qualitatively and quantitatively evaluate active ingredients in tablets by utilizing light absorption spectra. It was included as a general test method in the first supplement to the 18th edition of the Japanese Pharmacopoeia promulgated in 2021. Additionally, the analytical method validation guidelines (ICH-Q2(R2)) of the International Council for Harmonization of Technical Requirements for Pharmaceuticals for Human Use (ICH) revised in 2024, added validation information for multivariate analysis including NIR. It is increasingly recognized as a standard

test method for pharmaceutical process control and Real-Time Release Testing (RTRT).

While test methods using NIR are becoming more widespread, component quantity testing remains limited to sampling applications, like HPLC. Measurement times to analyze each individual tablet range from several seconds to several tens of seconds. Considering future continuous production lines, this approach is unsuitable for inline use. Since it is difficult to capture and visualize time trend information, the possibility of occasional out-of-specification products cannot be eliminated. Out-of-specification products may occur due to unstable powder handling at the start or end of production in various manufacturing equipment, or from the mixing state of active ingredients and excipients. Additionally, the current reality is that it is difficult to control unexpected quality risks, such as pharmaceutical residue during product changeovers, mix-ups and other human errors, or the risk of foreign component contamination due to external factors. These risks could be reduced if 100% tablet inspection using NIR could be performed inline.

To solve the above-described challenges and meet the demand for quality improvement in the pharmaceutical market, we developed the world's first^{*1} NIR 100% inspection equipment using transmission (Figure 1), applying dynamic measurement and optical sensing technologies that our company has developed over more than 50 years. This article introduces our technical approach.

^{*1} based on Anritsu survey results



Figure 1 External view of NIR 100% Tablet Inspector

2 Measurement and Analysis Methods

NIR^{*2)} is a type of spectroscopic method for qualitative and quantitative evaluation of test substances by measuring and analyzing light absorption spectra in the near-infrared light region^{*2)} of the test substance.

Light absorption in the near-infrared region is weaker than absorption due to fundamental vibrations in the infrared region. Therefore, when near-infrared light is irradiated on a sample containing the test substance, light reflected from the sample surface or light transmitted through the sample can be captured. These light signals are observed in a state where some light wavelengths in specific regions have been absorbed due to the physical properties of the test substance. Non-destructive inspection of the test substance in solid samples becomes possible by analyzing and interpreting the spectrum of transmitted light containing absorption information due to the physical properties of the test substance.

^{*2)} Light generally in the 750 to 2500 nm wavelength band between visible light and infrared light

2.1 NIR Measurement Method

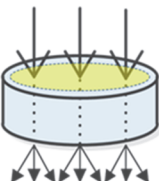
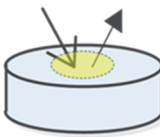
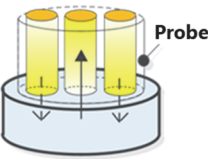
Table 1 lists NIR measurement methods. Diffuse reflection method can easily measure a large amount of light and can also yield high signal-to-noise ratio results. However, the inspection area is limited to near the surface, making it impossible to obtain information about components inside a tablet. Using interactance method, light can penetrate and inspect up to several millimeters below the surface depending on the probe configuration. However, since the probe must touch the surface of the test substance, it is difficult to apply this method to dynamic measurements, necessary for 100% inspection. This method has less measurable light compared to the diffuse reflection method, resulting in a lower signal-to-noise ratio and relatively higher technical difficulty for measurement.

The NIR 100% tablet inspector uses the transmission method to analyze components inside test tablets. This method evaluates the absorbance spectrum by comparing the intensity of the transmitted light—which carries information about the tablet’s interior—with that of the irradiated light or a reference medium, focusing on the degree of attenuation at each wavelength.

2.2 Component Analysis using NIR Transmission Spectroscopy

Analysis used measured values from NIR spectra and HPLC, etc. NIR spectra undergo preprocessing such as absorbance calculation, normalization, derivation, etc., combined with machine learning methods such as PLS and PCR, and are then compared and calibrated with HPLC measured values. This process facilitates quantitative evaluation of components in solid samples and detection of foreign contaminants in tablets.

Table 1 Comparison of NIR Measurement Methods

NIR Measurement Method	Transmission	Diffuse Reflectance	Interactance
			
Inspection Range (individual test)	Internal including surface	Only surface	Near surface
Dynamic measurement	⊙	⊙	× (Contact with test probe)

For quantitative analysis, multiple samples with known concentrations are prepared to create calibration curves. The transmission spectra obtained from each sample are analyzed and studied to create a calibration-curve model that outputs quantitative values. The concentration of the test substance is determined after feeding the absorbance spectrum into the model.

Foreign contaminants in tablets are detected using good-product control limits as a threshold. The measured absorbance spectrum containing foreign contaminants varies significantly from the thresholds established from acceptable products, which ultimately identify defective tablets.

3 Features, Basic Structure, Development Points and Implementation Method

3.1 Features

This tablet inspector evaluates the interior quality of tablets by transmitting near-infrared light through the tablet and conducting spectroscopic analysis of the light reduction. Analyzing the measured spectrum using the spectroscopic system facilitates inspection of component quantity, cross-contamination, and foreign matter contamination.

Designed for inline use on production lines, it inspects at high speeds of up to 250,000 tablets per hour. Randomly orientated tablets are fed into the machine, separated one-by-one through an alignment mechanism and then irradiated with NIR light. Transmitted light passing through tablets is measured using a spectrometer, and sorting is performed through pass/fail evaluation for each tablet using the established calibration curve.

3.1.1 Detecting Different Tablets with Similar Characteristics

The system can also differentiate tablets with similar external characteristics, which are commonly difficult to identify by traditional visual or appearance inspection. Defective products are then segregated through the machine's sorting/rejection system. This guarantees consistency in the quality of every tablet, reducing risks of leakage and cross-contamination.

3.1.2 Detecting Contaminants (Hair, etc.) in Tablets

NIR analysis is based on the principle that light is absorbed when it interacts with molecular components like CH groups in organic compounds. This method can non-

destructively detect and segregate tablets which contain foreign organic contaminants - like hair. These organic contaminants are unidentifiable using conventional devices like inspection cameras, or even metal detectors.

3.2 Basic Inspector Structure

The NIR 100% Tablet Inspector (Figure 2) consists of three main components: the transport system, the measurement system, and the operation control system.

The tablet transport system consists of supply, alignment, suction, and sorting units.

The measurement system includes a light source that emits near-infrared light, a projection unit that accurately directs this light onto the tablet, and a detector that captures the light transmitted through the tablet with efficiency.

The operation control system includes a display and functions for setting device parameters, controlling tablet transport equipment and sensors, evaluating and sorting tablets based on estimated active ingredient content, displaying statistical data, and recording operational history.

Tablet path and measurement flow are described below. Tablets are fed from the top-left hopper unit and are supplied to a succeeding vibration feeder. Broken tablets are screened through a section on the said feeder unit. Unrejected tablets are then handed to a rotary feeder and are aligned to a single file by a regulating guide. After alignment, the flat sides of the tablets stick to a negative-pressure suction disc, and their orientation is then adjusted with additional guides. Tablets sticking to the disc have their NIR spectra captured and analyzed as they pass through the measurement system. During this stage, component concentration calculations and product quality analysis are performed through signal processing. Abnormal tablets are rejected while acceptable ones are transferred from suction discs to a conveyor and discharged from the front right side of the inspector (Figure 3).

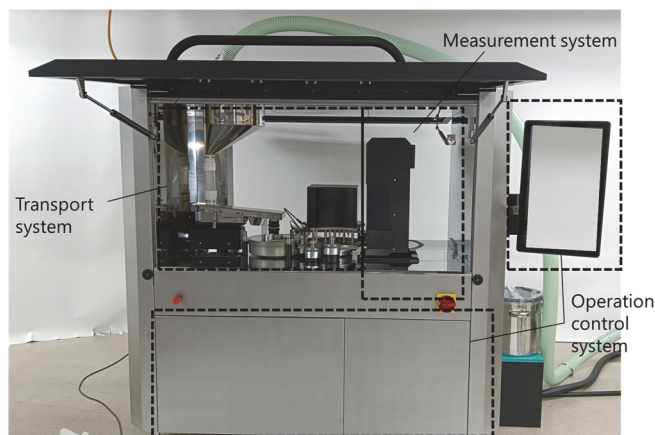


Figure 2 NIR Inspector configuration

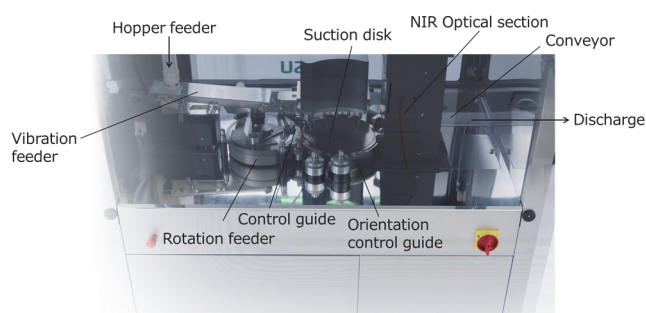


Figure 3 Tablet path

3.3 Development Points and Implementation Methods

The transmission measurement method has had challenges with poor signal-to-noise ratio due to weak measurable light. However, by developing our own optical unit, we achieved a method that can analyze tablet components non-destructively at a higher speed. (Figure 4).

Assuming inline use on production lines, the high-speed inspection system targets a maximum capacity of 250,000 tablets per hour. We achieved a processing time of ≤ 11 ms per tablet by using high-speed measurement, evaluation, handling, and sorting technologies.

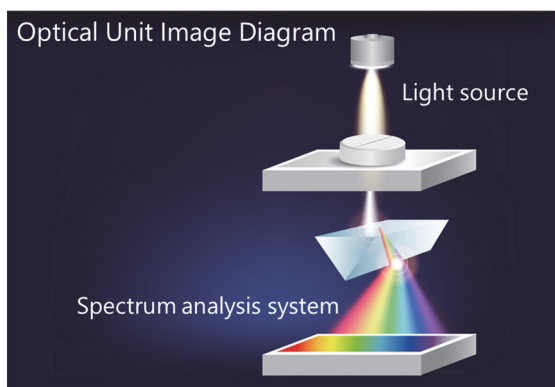


Figure 4 Optical unit

3.3.1 High-Speed Measurement and Evaluation Processing

Achieving a processing capacity of 250,000 tablets per hour requires processing each tablet within 11 ms. Furthermore, since tablet orientation must be detected in < 1 ms, precise timing detection required development of an FPGA-based dedicated detection mechanism. Timing-based measurement captures the near-infrared spectra, and signal processing is performed according to pre-registered methods to perform pass/fail evaluation and sorting. Each measurement, evaluation, and sorting process are designed to use a separate processing mechanism so as not to cause interfere to each other.

3.3.2 Handling

Stable and accurate measurement of transported tablets requires controlling tablet orientation, position, vibration errors, and speed fluctuations. This handling system uses guides to minimize variations that occur to tablets when they stick to the suction disc. This enables measurement accuracy, achieving circumferential deviation within ± 0.2 mm, orientation angle variation within 3° , and speed fluctuation rate through the suction disc mechanism within $\pm 0.5\%$.

3.3.3 Sorting/Rejection

The inspector requires sorting/rejection equipment to ensure that both uninspected and defective tablets are reliably rejected from the line, while simultaneously preventing anything else other than acceptable tablets from transporting to the next stage. Figure 5 shows the sorting/rejector arrangement.

Air-jet rejectors are used to sort out defective tablets from about 90 tablets transported per second based on the pass/fail evaluation results, achieving tablet sorting within 11 ms by using fast-response solenoids and optimizing the flow path.

Additionally, the sorting unit has air-jet rejectors on the downstream side of the defective-tablet rejection, which ejects uninspected tablets. Furthermore, a physical flipper-style gate downstream of the complete discharge section sorts improperly sorted defective tablets, ensuring that defective or uninspected tablets never flow to the acceptable tablet side.

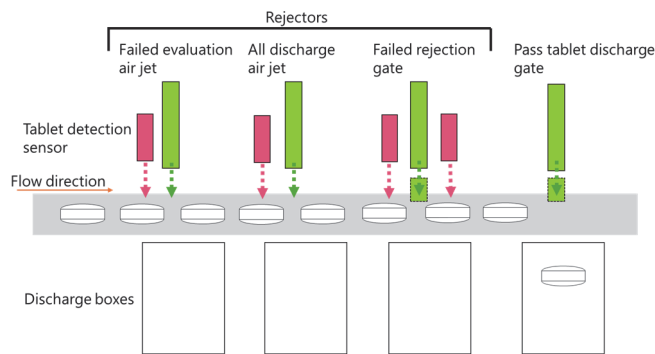


Figure 5 Sorting/Rejector configuration

3.3.4 Efficient Productivity

In addition to the 100% inspection function, there are the following: a function to register settings according to tablet types as product varieties; a NAVI function to support product registration in a guided format; an adjustment function to limit variation in results due to operator skill levels, enabling high-reproducibility measurements; and calibration curve creation software for converting NIR spectra to component concentrations.

3.3.4.1 Product Inspection Records NAVI

Tablet registration is completed by following the work process guided by the navigation-style screen (Figure 6). Time requirements are reduced, and work efficiency is improved by automatically calculating the optimal values for various parameters.

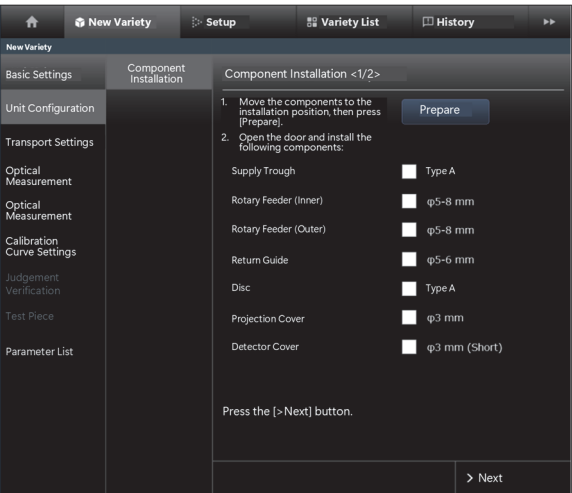


Figure 6 Navigation-style operation screen

3.3.4.2 Tablet Orientation Adjustment Control Function

To improve measurement reproducibility, tablets must be transported in a stable orientation. However, since tablet shape and thickness vary, the orientation of each tablet must be adjusted individually. This inspector uses guides to adjust tablet orientation, and a camera confirms the position

adjustment accuracy. This high-speed camera positioned beside the suction disc visually monitors the orientation of tablets during transport. The orientation can be easily and intuitively adjusted by aligning the tablets within the marker lines shown at the top and bottom of the camera image, using guide adjustments (Figure 7).

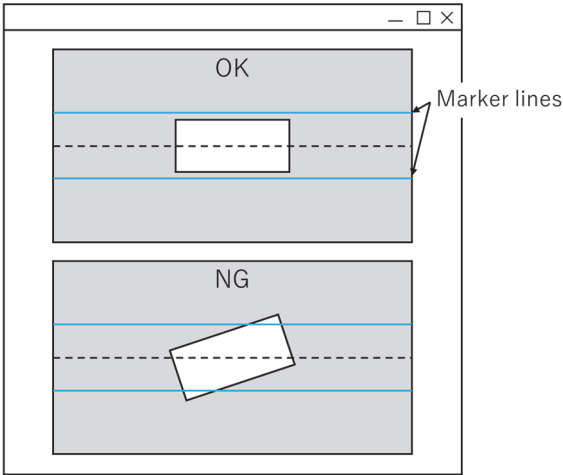


Figure 7 Tablet orientation control adjustment function screen

3.3.4.3 FDA 21CFR PART11 Support

The FDA 21CFR Part 11 regulation established by the Food and Drug Administration (FDA) concerns ‘electronic records and electronic signatures,’ and requires audit trail records and user authority management, among other things, as requirements for electronic records to secure reliability equivalent to paper records. This inspector has an audit trail and qualified person authentication functions in compliance with FDA 21CFR Part 11.

The audit trail function records the history of operations and errors related to production and documentation within the equipment, which can be used to monitor deviations from manufacturing procedures and analyze causes of abnormalities. Additionally, statistical data can be saved as production records linked to production lot numbers.

The qualified person authentication (user authentication) function requires a user code and password for operation. Unauthorized operations are prevented by setting permissions for each user. Additionally, operation permissions can be customized according to customers’ management standards.

3.3.4.4 Calibration Curve Software

Calculation of component amounts from NIR measured results requires preparing multiple tablets and creating a

calibration curve (Figure 8) establishing the correlation for estimating actual component amounts from NIR spectra.

Obtaining an appropriate calibration curve for each target tablet requires expertise in preprocessing (Figure 9) multiple parameters, such as filtering, normalization, and differentiation processing. This inspection machine has analysis software to create calibration curves (processing to calculate component amounts) from NIR spectra - reducing the work burden on users by parameter adjustment assistance features.

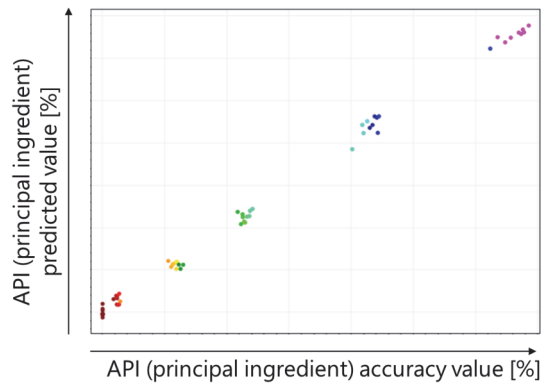


Figure 8 Calibration curve

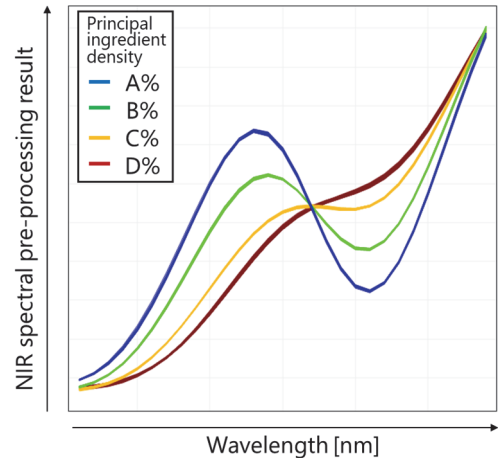


Figure 9 NIR Spectral preprocessing results

3.3.5 Sanitation

The main base is mirror-finished so potential blind spots are visible from reflections. Parts that mate with the main base are tapered, creating a shape (Figure 10) that prevents powder residue accumulation and is easy to clean.

Components, such as frames, are designed with sanitation in mind by making them solid rather than hollow. Parts that touch tablets and require daily cleaning can, with few exceptions, be detached and reattached without tools, allowing easy cleaning.

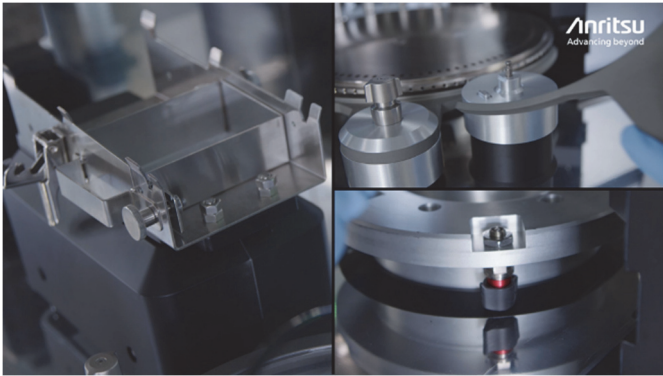


Figure 10 Main base and parts with powder buildup resistant design

4 Key Specifications

Table 2 Key Specifications

Item	
Supported tablets	Type: Uncoated Shape: Flat disk, R type Diameter: ≤5 to 12 ±0.2 mm Thickness: ≤2 to 5 ±0.2 mm Hardness: ≥39 N
Maximum inspection speed	250,000 tablets per hour (Depends on tablets)
Main unit dimensions	1600 W × 1000 D × 1800 H mm
Mass	1200 kg
Display section	23.8-inch IPS LCD (Full HD)
Operation section	Touch panel (Capacitive)
External frame	SUS304, A5052, PET
Settings	1000 product types
Operating environment temperature	20° to 26°C (<1°C per hour temperature change)
Operating environment relative humidity	40% to 60% (No condensation)
Power source and consumption	3-phase 200 Vac ±10% 50/60 Hz 80 A
Waterproof standard (IEC 60529)	IP30 (Including operating section)

5 Conclusion

Inspection of all tablet interiors is now realizable through this inspector which uses high-speed NIR transmission technology. Until now, 100% inspection of tablets before packaging has primarily been surface-appearance inspection, but going forward, it will be possible to inspect the ‘interior’ of all tablets (constituting most medication tablets), thus enabling supply of higher-quality pharmaceuticals. On the other hand, recent labor shortages do not help prevent careless

mistakes using meticulous quality assurance by human hands.

Going forward, we aim to expand the range of tablets supported by this inspector, as well as improve usability for practical implementation by using automation to solve labor shortage issues while still ensuring pharmaceutical quality.

Appendix

The results of the calibration curve obtained using this equipment are recorded here for reference. We created a calibration curve model using one-half of 116 tablets from six types of acetaminophen formulation as training data under the calibration curve creation conditions listed in Table 3 and then tested the created model with the remaining tablets. The error in quantitative value prediction (RMSEP) was within ±1.5% (Figure 11).

The ‘Sakura Bloom Tablets P2 Mock’ model for creating Common Technical Documents (CTD) for pharmaceutical applications using RTRT uses RMSEP ≤1.5% as the accuracy criterion for validating calibration models created using NIR. We confirmed that actual measurement data using real drugs with this inspector satisfies the above criteria.

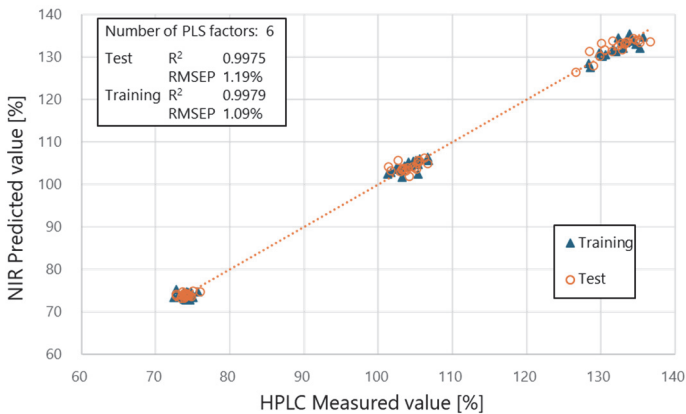


Figure 11 Plot of quantitative values obtained by NIR and HPLC measurements

Table 3 Calibration Curve Conditions

	Training Data	Test Data
R ²	0.9979	0.9975
RMSEP	1.09%	1.19%
Sample size	58	58
Preprocessing	SNV+2 Sub differential	
Wavelength range	1005 to 1178 nm	
Frequency domain	8488 to 9950 cm ⁻¹	
Repeatability (3σ)	1.40%	

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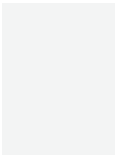
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Electrical Performance Enhancements of Fixed Waveguide Attenuators

Tom Roberts

[Summary]

Attenuators have many uses in RF, Microwave, Millimeter-wave circuits and instrumentation. The quality of an attenuator is typically described by its impedance match, attenuation flatness and attenuation accuracy over the frequency band the transmission line medium supports. This paper describes impedance match improvements and examines attenuation flatness of full-band, fixed waveguide attenuators. These improvements can be scaled and applied to any waveguide size.

1 Introduction

Attenuators are critical components at RF, Microwave and Millimeter-wave frequencies to adjust the signal path loss and gain settings of electronic systems. Adjustable waveguide attenuators at receiver front ends can increase system dynamic range¹⁾ but are often bulky and expensive, increasing the overall system costs. Attenuators at the front of sensitive receiving front ends can allow for a greater input signal levels without driving the receiver into compression at the expense of increased receiver noise floor. A mechanically adjustable attenuator could prevent the receiver from compressing and maintain a low noise floor level. However, repeatability of mechanical attenuators would increase the receiver calibration uncertainty²⁾.

A recent paper³⁾ describes a 15-dB fixed waveguide attenuator in WR6. Simulation and measurement results indicate a worst-case VSWR of 1.25 (19-dB Return Loss). In this paper, a similar waveguide attenuator with fixed attenuation levels of 5-dB and 25-dB is studied with a goal of improving VSWR and examining the impact on attenuation flatness.

2 Mechanical Considerations

2.1 Waveguide Housing Split Plane

Consideration of how the waveguide housing would be machined is critical to avoid potential electrical performance issues. Typically, waveguide housings are machined as two symmetrical pieces along the E-plane or broad-wall (see Figure 1). The E-plane is the preferred split plane because it is less sensitive to seam imperfections to that of a split along the H-plane or narrow wall. Waveguide housings split along the H-plane with seam imperfections may not present sufficient conduction along the narrow wall and can introduce in-

band resonances⁴⁾. If the H-plane is located at the top-most portion of the inside cavity, the sensitivity to seam imperfections remains. Investigating and comparing the impacts to housing seam imperfections along the E-plane and H-plane can be examined using a 3D EM simulator⁵⁾.

Many waveguide housing designs incorporate an H-plane split for convenience. Reasons for choosing an E or H split plane are varied: ease of machining interior details, placement of interior substrates and components, etc. Described in this paper, the enhanced D-Band (WR-6) attenuator design utilizes an E-plane housing split which benefits not only from a low-sensitive seam placement, but also the mechanical ease in which the attenuator performance is enhanced.

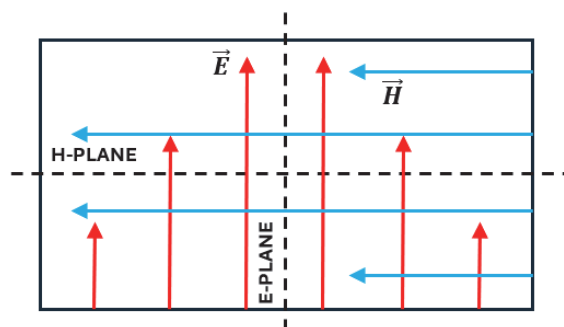


Figure 1 Waveguide housing E and H split planes

3 Measurements and Measurement Limitations

3.1 Attenuator Characteristics

An ideal attenuator has characteristics of perfect match and flat attenuation over its operating frequency band. In practice, measurement of an attenuator using a VNA will have imperfect match and attenuation flatness. The reflection characteristics of the attenuator should approach the VNA's corrected directivity value within 10 to 15-dB for a worst-case corrected directivity value of 45-dB as a practical

minimum design goal. The measurement instruments corrected directivity can be viewed as the value where the DUT return loss measurement uncertainty is large and would be in the range of -6 dB to $+\infty$ dB.

Many fixed waveguide attenuators commercially available and described in a recent paper³⁾, have a worst-case match of about 20-dB and their structures are likely similar. Relative to our stated goal, it seems this worst-case match of 20-dB could be improved by identifying and correcting the cause of discontinuity mismatches caused by the inclusion of the attenuator elements which is the focus of this paper.

4 Treating the attenuation element as a transition

4.1 Influence of substrate without resist

Assuming the waveguide housing incorporates an E-Plane housing split, and the substrate lies flat against either one or both broad-walls, the substrate must be divided into two pieces to not obstruct the housing split location as shown in Figure 2. First, the substrates without Tantalum Nitride (TaN) resist material are examined in an EM simulator for their impact on the waveguide's reflection properties. Including the substrate within the waveguide forms two transitions within the waveguide structure. Minimizing the reflections caused by the inclusion of the substrate is the goal.

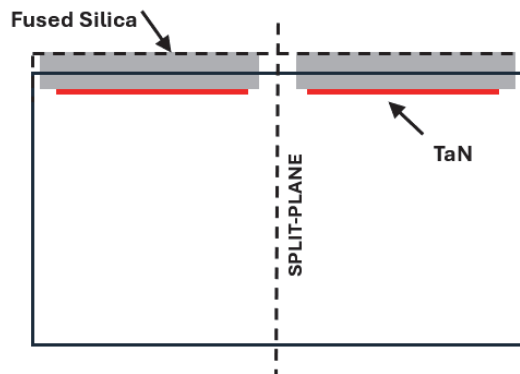


Figure 2 Broadwall placement of embedded substrates (end-view).

4.2 Compensating the transition

It was found that embedding a 250-mil $\times$ 61-mil $\times$ 5-mil thick Fused Silica substrate within the WR6 waveguide on the broad wall by 60% (as shown in Figure 3) improves the reflection response significantly vs. not embedding the substrate at all. Figure 4 shows the S11 reflection response. Embedding the substrate shows approximately a 10 to 15-dB improvement in S11 over the full WR6 waveguide band. This

gives confidence that the attenuators match can be improved with a minor waveguide housing change. Due to structure reciprocity, S12 and S22 are not included in the results.

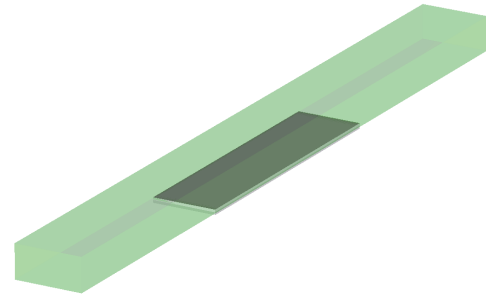


Figure 3 WR6 Embedded fused silica substrate without the TaN resist layer.

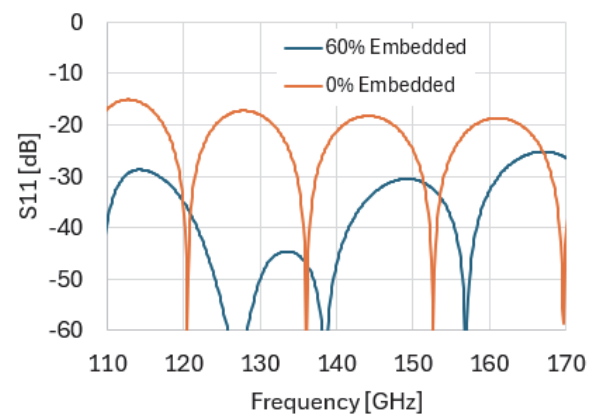


Figure 4 WR6 reflection response of embedded and non-embedded substrate without the TaN resist layer.

An end-mill machine tool is used to machine the compensation feature into the broad wall. Typical end-mill tools have a flute length-to-diameter ratio of 3:1. The required end-mill flute length is 32.5 mils so its diameter would be approximately 11-mils. Figure 5 shows the radius formed by the end-mill tool just beyond the substrate edge. This housing feature was incorporated into the simulation models and didn't show significant impact on the results.

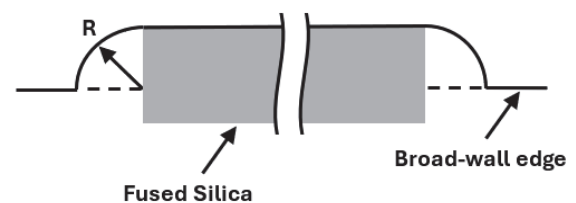


Figure 5 End-mill tool radius detail at substrate ends

5 Attenuator Designs: 5-dB and 25-dB

5.1 Thin film resist taper

A TaN sheet resistivity was chosen as 50 ohms/square pulled back from each substrate edge by 2-mils. Instead of

using an elliptical taper of the thin film resist material, a simple linear taper was examined for its impact on S11 and S21. Linear taper angles of 0°, 22.5° and 45° at both substrate ends were examined using the simulators parametric sweep option. Since the thin film resist is near the embedded substrate transition, it is expected the resist material will have some impact on match. Sections 5.1 and 5.2 below discuss the impact on match with thin film resist linear taper values.

5.2 5-dB attenuator design

A 5-dB attenuator was designed with two 5-mil thick Fused Silica substrates each having a size of 150-mil × 30.5-mil centrally positioned on the broad wall of a WR6 waveguide as shown in Figure 6. Due to structural symmetry, a magnetic plane boundary condition along the waveguides E-plane was used to simplify model computation.

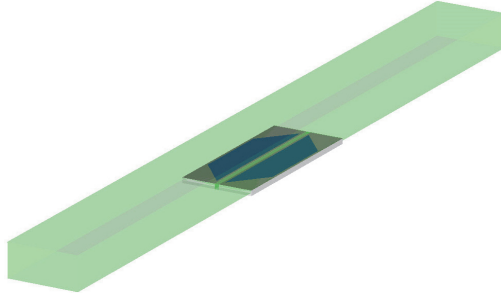


Figure 6 WR6 5-dB split-substrate attenuator w/tapered TaN resist layer.

The influence of taper angles with S11 and S21 was found to be modest for the 5-dB design. A small trade-off between taper angle, match and attenuation flatness as shown in Figure 7. A taper angle of 0° exhibits the best match but slightly worse attenuation flatness. Conversely, A taper angle of 45° exhibits the best attenuation flatness but worse match relative to smaller taper angles. A similar trade-off was found for negative taper angles.

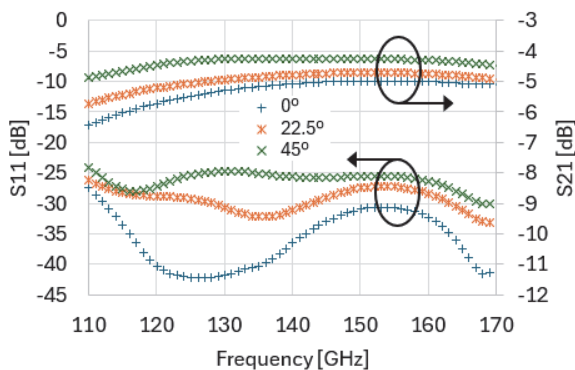


Figure 7 WR6 5-dB attenuator reflection and transmission response vs. TaN resist taper angle.

5.3 25-dB attenuator design

A 25-dB attenuator was designed with four 5-mil thick Fused Silica substrates each having a size of 375-mil × 30.5-mil centrally positioned along the length of a WR6 waveguide as shown in Figure 8. To increase attenuation, substrates were attached to both broad walls. Structural symmetry allowed use of a magnetic boundary condition along the waveguides E-plane and an electric boundary along the waveguides H-plane to simplify model computation.

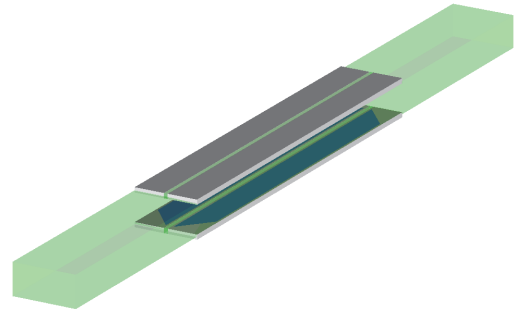


Figure 8 WR6 25-dB split-substrate attenuator w/tapered TaN resist layer.

As in the 5-dB attenuator design, the 25-dB attenuator shows only modest improvements in S11 with taper angle. However, there was no observable difference in attenuation flatness. Like the 5-dB attenuator design, attenuation value decreases with increasing taper angle since an increasing amount of TaN resist material is removed. S11 and S21 results are shown on Figure 9.

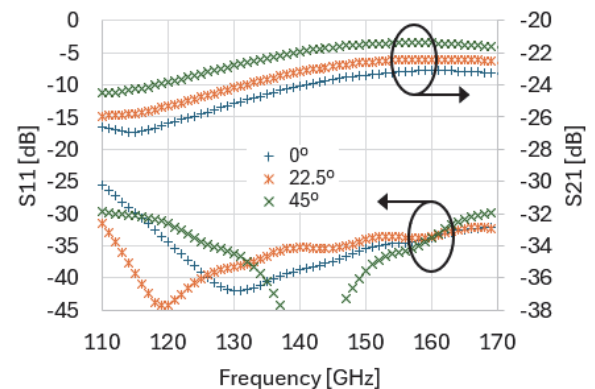


Figure 9 WR6 25-dB attenuator reflection and transmission response vs. TaN resist taper angle.

5.4 Manufacturing considerations for high attenuation value designs

The 25-dB attenuator design described in section 5.3 utilized both broad walls and an increased substrate length to increase the attenuation value. Fused Silica substrates are relatively fragile depending on their thickness and length-

to-width aspect ratio. Substrates with large length-to-width aspect ratios may exhibit smaller thin film processing yields. For the 25-dB attenuator design, the length-to-width aspect ratio is about 12.3:1 which may exceed thin film substrate manufacturing design guidelines and result in poor yields due to cracking or breakage. This problem can be addressed by sub-sectioned substrates while adhering to a maximum substrate length-to-width ratio.

The attenuator substrate design only needs to be processed with a TaN resist layer on the substrate topside. The backside of the substrate does not need a metal layer, so this reduces thin film processing costs. A low dielectric constant epoxy can be used to attach each substrate to the housing inside channels.

6 Conclusions

RF, Microwave and Millimeter wave components can often be broken down into their constituent parts to examine and improve their electrical performance parameters. In this paper, two examples of WR6 fixed attenuators were examined by treating the waveguide transmission line structure as a transition optimization problem with the TaN resist layer removed. Once the electrical parameters (i.e. S11) of this low-loss transmission line structure are optimized, the TaN resist material is introduced to gauge its impact on the attenuator's electrical parameters such as match, absolute attenuation and attenuation flatness.

Unfortunately, time constraints did not allow the completion of the built 5-dB and 25-dB attenuator assemblies and compilation of test results. Results of the built and measured WR6 attenuators will be provided in a future paper.

7 References

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Millimeter-wave Optical Network Analysis

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[Summary]

The desire for higher single-wavelength modulation bandwidths in optical networks requires broader band RF characterization of components including electrical-to-optical (EO) and optical-to-electrical (OE) converters and this is often done with microwave techniques. Extending these measurements to the mm-wave regime, may require added attention to the measurement techniques, uncertainties, and characterization processes.

1 Introduction

There has been no reduction in the desire for increasing aggregate bandwidth in optical communications links. For many years, the increase was mainly accomplished by multiplexing additional optical wavelengths (or other optical parameters) or changing modulation techniques but perhaps not increasing the per-carrier modulation bandwidth. Recently, there has also been a desire to increase each of these modulation bandwidths which means increasing RF bandwidths on the electrical side of the system into the 100+ GHz range¹⁾⁻²⁾. In turn, this pushes the RF characterization tools for the optical conversion and other components into the higher mm-wave regimes. This characterization may also include modulation domain measurements of purely optical components as the bandwidths can start to interact with traditionally optical mechanisms.

Measurements of optical converters to mm-wave frequencies will be explored here with a focus on changes happening at higher frequencies. Subtopics include the calibration and de-embedding steps, primary device characterization (to use as transfer standards) and composite uncertainties. Examples will be considered, and the effects of changing device media (in an RF connectivity sense) will also be presented.

2 Measurement approach

The main aim of network analysis with optical conversion components is to evaluate the ‘channel’ impairments that an information signal would experience during modulation and demodulation steps. Usually, this is just done with the standard swept CW tone of a vector network analyzer (VNA) but, if there are significant nonlinearities in play, an actual modulated signal could be used as well³⁾.

With 100+ GHz of available modulation range in some conversion components, a broadband VNA becomes a requirement and instruments reaching to 220+ GHz are available⁴⁾. The broadband aspect can be important as digital signals normally form the information payload in these systems, so it is a continuous bandwidth requirement from near DC and not a banded one. Structure in the frequency response function, even at intermediate frequencies, could introduce bit error rate or eye diagram issues. Because of the importance of the response form, the calibration of the VNA can take on added importance at the higher frequencies. This can be particularly relevant when the physical media are unconventional (e.g., the modulator is excited with wafer probes, or the detector is in a microstrip fixture).

Eye diagrams are important to assessing the effect of the functional response and it is common to use the mm-wave channel data (showing loss, phase slope and structure) to compute the effect on a digital signal and this is also commonly done as part of the network analysis procedure. An example eye diagram for a 56 Gbaud/s PAM4 signal traversing a 145 GHz EO/OE path is shown in Figure 1 along with a native response without the EO/OE system. Note that the vertical scales in this figure are not the same, and net loss was not compensated. The roll-off of the optical path was not extreme to 145 GHz so the eye distortion was modest for this signal.

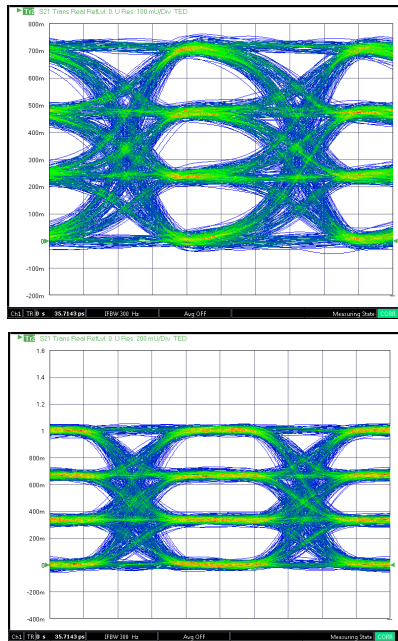


Figure 1 Eye diagrams of a 56 Gbaud/s PAM4 signal passing through an EO-fiber-OE chain (measured to 145 GHz; top) and the same signal just measured directly by the same 145 GHz receiver (bottom). There is a change of the eye shape with the optical path as the converters had loss slope with frequency. No pre-emphasis or gain correction was used for these measurements.

Typically, one wants to identify the behavior of a particular component (e.g., a modulator) and the use of characterized devices and de-embedding allows extract the response of one component from that of the chain. The conversion behavior of one device (a photodiode for the discussion here) is characterized in a traceable fashion and then this information is propagated to other devices by de-embedding from the composite measurement as suggested in Figure 2. In that figure, the base VNA calibration puts the reference planes at the RF inputs of the EO and OE devices so a S_{21} measurement gives a picture of conversion losses and the optical path between the converters. By de-embedding the now ‘known’ characteristics of one of the converters, the measurement reveals the characteristics of the other converter and those of the optical path⁵⁾. Typically, only the transmission characteristics of the OE are used so the loss and phase change of that device are removed from the measurement. Since the EO-OE chain is rarely bilateral, multiple match interactions between the two ports are not a problem and more complete de-embedding methods offer little benefit.

That optical path between the converters may be negligible in a loss sense but other characteristics may be relevant as will

be discussed in a later section. An additional de-embedding step can be used to reveal (a limited number of) optical path characteristics (loss and group delay) once the converters are known.

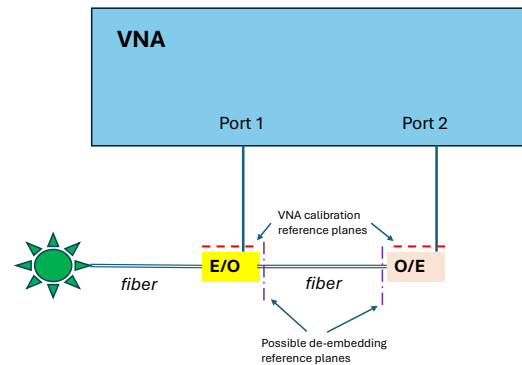


Figure 2 A typical measurement setup where an electrical VNA calibration is performed to set initial reference planes, and the de-embedding is used to remove EO and/or OE converter effects (in a conversion loss sense).

3 Characterized OE devices

A high-quality characterization of one of the conversion devices is needed to be able to solve for the other. This process can start on either side but here we use the photodiode as the basis as its stability over time is easier to quantify and the bias dependencies are often weaker. The base characterization is often done using electro-optic sampling (EOS)⁶⁾ where a very narrow optical pulse (on the femtosecond scale, compared to the expected picosecond-scale impulse response of the photodiode) is used to excite the photodiode and the photodiode's output pulse is analyzed temporally. From a transform of that output pulse, the frequency response of the photodiode is obtained. An example pulse and frequency response curve for a 145 GHz photodiode obtained with an EOS process⁷⁾ is shown in Figure 3. It is interesting to note that the frequency response curve is not entirely monotonic. Internally, the photodiode has multiple mismatch centers on the RF side (including the RF connector) and within the active region itself that can lead to minor ripple. Importantly, this frequency response structure is stable in time so the device can act as a transfer standard. Also of note is that the phase of the frequency response is also extracted, and this can be useful for evaluating group delay distortions through the path or the device of interest. The full phase function would have a steep slope corresponding to the electrical length of the device and it is deviations from a linear slope that are of primary interest and this deviation is plotted in Figure 3.

This primary electrical response can then be used to characterize a modulator or other parts of the system, through de-embedding, as was discussed in the previous section. This process can then be extended by using a now-known modulator/EO device to evaluate a new photodiode.

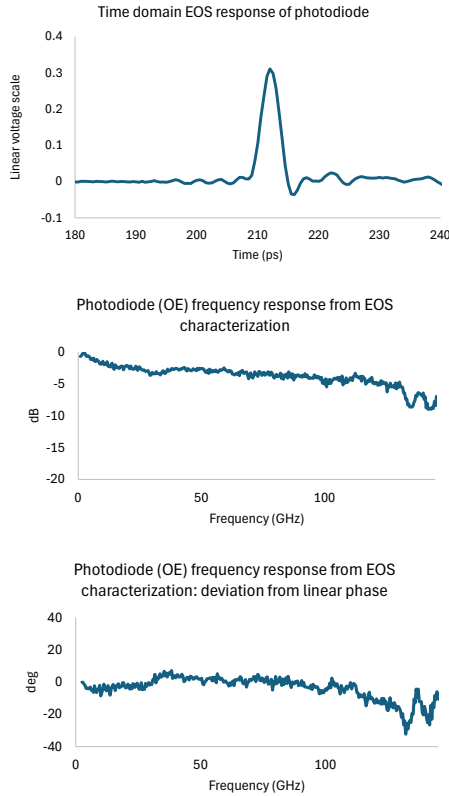


Figure 3 The EOS impulse response of an example 145 GHz photodiode is shown along with the frequency domain transformed responses (magnitude and phase).

The EOS fundamental process is extensible in bandwidth so characterizing higher frequency diodes is possible. The laser pulse widths available readily are not a limit (until at least 500 GHz). The transmission structure for temporal analysis is also readily doable to at least 220 GHz but, as this is usually done on-wafer, it does bring in additional considerations for the probe design and the calibration procedure.

4 Uncertainty topics

The measurements of any of the components in the EO-OE chain have a variety of uncertainty contributors. The most direct is the VNA measurement itself and this purely electrical measurement has been studied in the literature extensively⁸⁻¹¹. The calibration process contributes some uncertainty and receives much attention. Since many broadband

modulators and detectors have significant RF mismatch (e.g., -5 dB above 100 GHz is not uncommon as of this writing), the residual mismatch terms can be noticeable. Electrical connection repeatability and drift may be important. Network analyzer linearity can sometime play a role but typically the setup can be constructed (keeping RF signal levels bounded at certain locations) so this is of minimal importance. Since the conversion losses in many optical devices are high, one may run into signal-to-noise limits in the VNA measurement, particularly at low optical power levels. The latter follows if one looks at the modulated optical signal as in the sketch of Figure 4. If the optical power drops, the modulation sidebands will also generally drop in amplitude, and this will directly reduce the power to the VNA receivers.

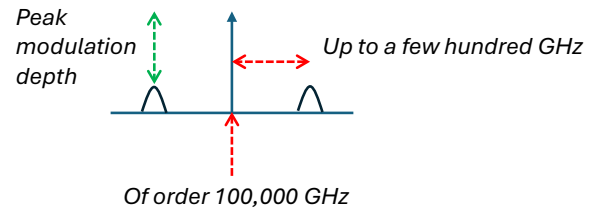


Figure 4 Spectrum of the optical signal incident on the photodiode when measuring a conventional modulator.

One can see the result of this in composite EO-OE measurements at three optical power levels as in Figure 5. Even though the curve shapes are unchanged, they get closer to the VNA noise floor (nominally -110 dB on this scale) so dB-scale impacts are possible at -25 dBm optical power in this example. Reducing the IF bandwidth of the VNA measurement, using access loops (to reduce loss to the receiver) or increasing RF power (up until linearity of the VNA or the EO system are challenged) are all means to mitigate this. Changes in the highest frequency structure are due to non-linearities in this setup.

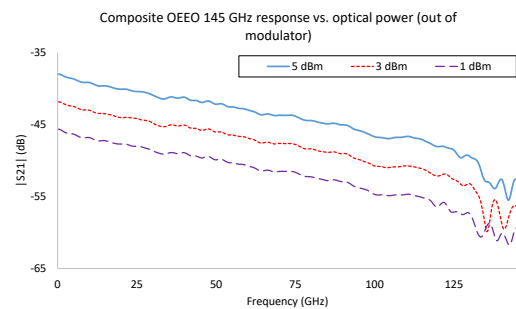


Figure 5 Optical power impacts on measured EO-OE responses.

Compared to lower frequencies (<70 GHz), the broadband

frequencies discussed do increase measurement challenges:

- A modest degradation of noise floor specifications (e.g., -100 to -110 dBm instead of -120 or -130 dBm)
- A worsening of connector repeatability and cable drift
- An increase in calibration-related uncertainties (which are compounded in mixed media scenarios to be discussed)

The optical system itself may introduce additional uncertainty terms including optical power stability, repeatability and drift of the optical connections, and bias stability (of either EO or OE components).

There is also the uncertainty in the characterization of the primary converter used for the de-embedding process (usually traceable to a National Metrology Institute). Often, this begins with the EOS process discussed in the previous section and uncertainties include residuals from mismatch correction, finite bandwidths, and imperfections in de-embedding steps⁶⁾. An example characterization curve along with uncertainty limits is shown in Figure 6.

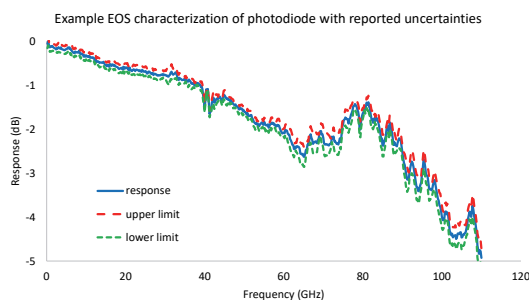


Figure 6 An example of a primary (EOS-based) characterization of an OE (photodiode) device along with the upper and lower uncertainty limits.

The combination of these uncertainties can follow one of the several established practices^{8), 11)} and can lead to a better understanding of when different terms are dominant.

5 Measurement examples

A simple example would be the characterization of an EO modulator to 145 GHz. A photodiode characterized by EOS to this frequency range was available and that EOS data will be used for de-embedding. First, the VNA (Anritsu ME7838D) was calibrated using a 0.8 mm coaxial calibration kit. We would expect transmission uncertainties of 0.1-0.2 dB for higher signal levels and the noise floor of this instrument is -100 to -110 dBm over most of the frequency range¹²⁾. The photodiode characterization uncertainties, in

this case and unlike Figure 6, were highest below 30 GHz (increasing to 1 dB at the low end) and averaging 0.4 dB over the higher frequencies.

Combining these uncertainty terms, one can end up with a plot as shown in Figure 7. Because of the influence of the VNA noise floor and the dependence on optical power, the x-axis is in terms of detector signal level. For these measurements with 3 dBm optical power into the detector, and given the responsivities involved and the -10 dBm RF drive, the received level will not be below -50 dBm. At high frequencies, repeatability and VNA calibration residuals tend to dominate while, at lower frequencies, initial photodiode characterization is more important.

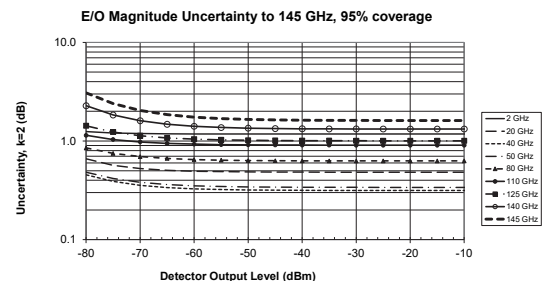


Figure 7 A calculated uncertainty curve for the extraction of an EO response to 145 GHz. This includes uncertainty terms from VNA measurement, OE characterization, aspects of the optical setup and repeatability.

The composite response and the result after de-embedding are shown in Figure 8. Both responses are normalized to 0 dB at low frequencies. In this case, the de-embedding had only modest effects below about 130 GHz indicating the EO roll-off was dominating the composite measurement. At the higher frequencies, the OE photodiode had more roll-off on its own (per the characterization), so the extracted EO response shows more of a difference. The corresponding eye (with a similar setup to Figure 1 but with a different normalization) is in Figure 9 for the EO alone. One can see an eye width reduction from that in Figure 1-lower as would be expected for such a loss profile.

At the higher data rates, there are cases where the bandwidth characteristics of the optical components can contribute¹³⁾. One example is a polarization-maintaining fiber where the two RF modulation sidebands on the optical carrier may see different phase velocities with the difference increasing with modulation rate. When these sidebands recombine in the (nonlinear) photodetection process, the phase difference can convert to an amplitude difference. The

change in dispersion of the fiber can come with a connection process, a change in mechanical stress applied or a seemingly innocuous fiber substitution. An example is shown in Figure 10 and has the appearance of a change in RF bandwidth.

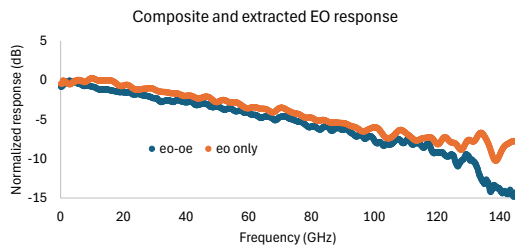


Figure 8 An example composite EO-OE response along with the de-embedded EO result to 145 GHz.

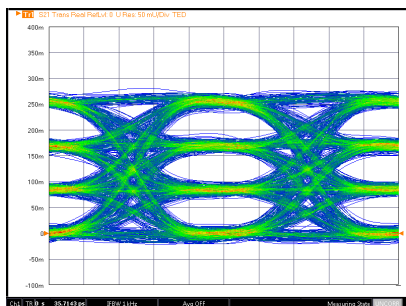


Figure 9 The computed eye diagram response for the EO of Figure 8 with a 56 Gbaud/s PAM4 signal into a 145 GHz system.

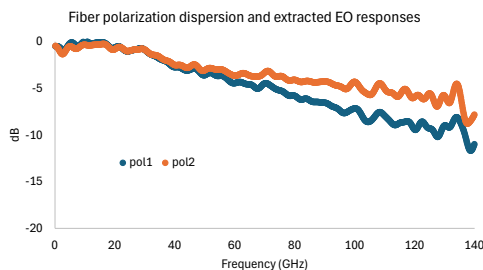


Figure 10 An example showing how a change in fiber polarization dispersion results in a perceived bandwidth change in an extracted EO device response.

6 Mixed-media complications

With higher levels of optical system integration (to include the popular area of Silicon Photonics¹⁴⁾), this type of measurement more often includes mixed media (e.g., one port is probed on-wafer and another port may be fixtured or coaxial). A sketch of such a setup is in Figure 11a. Not only is the optical engineering more complex¹⁵⁾ but the mm-wave calibration and de-embedding is more challenging and perhaps elevates uncertainties.

There are several options on how to perform the network analysis calibration:

- (1) Perform the VNA calibration on-wafer and then embed a probe's or fixture arm's characteristics (Figure 11b)

This can make use of standard on-wafer calibration methods that are relatively well-understood into the mm-wave range. There are issues (e.g., higher order mode propagation on the wafer/substrate, coupling to nearby metallized structures...) but the geometry is at least controlled. Characterizing the probe/fixture arm is critical and this network extraction problem has been discussed extensively¹⁷⁾. The uncertainties usually increase with frequency.

- (2) Perform the VNA calibration coaxially and de-embed probes (see Figure 11c)

At modest mm-wave frequencies (e.g., to 150 GHz), this may be simpler as well-characterized coaxial calibration kits exist. The probe characterization challenge remains.

- (3) Perform a hybrid calibration where two base calibrations (e.g., an on-wafer calibration and a coaxial calibration) are combined with the use of a single reciprocal device¹⁸⁾ (often a probe). See Figure 11d.

Here a portion of the error coefficients from two base calibrations are combined with a reciprocal processing step¹⁸⁾ to form the calibration at the desired reference planes. This is usually a lengthier procedure and can be mechanically complex to execute. It does avoid the need for separate probe characterization (or knowledge of the probe's characteristics from another source).

All of these approaches add to VNA uncertainties, and the ideal choice may depend on mechanical details of the setup (e.g., how difficult to demount each probe or fixture arm). Above 100 GHz, these additional steps may add a few tenths of a dB or more of net uncertainty. From an electrical point of view, the repeatability of the different connections is a factor as may be the mismatch of the EO and OE devices (i.e., for the more mismatched device, repeatability and calibration uncertainties may have greater impact so more care on that port may be warranted).

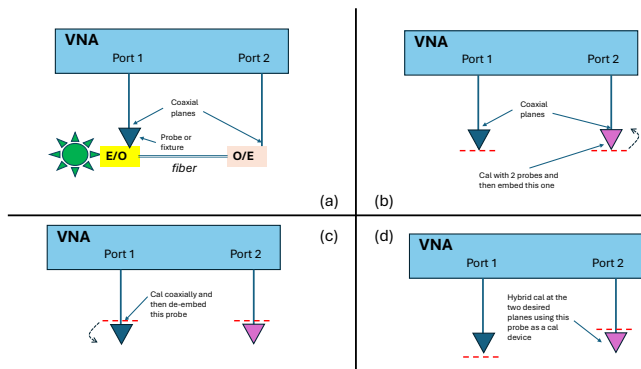


Figure 11 A situation where one optical converter uses a different RF media connection than the other (Figure 11a). There are a number of different VNA calibration approaches to handle this (Figures. 11b-d).

7 Conclusions

The desire for increasing direct modulation bandwidths on optical links has helped drive higher frequency characterization methods for optical converters (and related devices) using broadband network analyzers and broadband primary device measurements. At these higher frequencies, uncertainties increase, the dominant terms may change, and new ones may appear. As integration is also increasingly needed at these frequencies, the measurement media also change and can necessitate modified calibration and de-embedding strategies.

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Contribution to other Publications

Contribution to other Publications (April 2024 to March 2025)

Author	Affiliation	Title	Publication
Akiko Tada Takanori Saitoh	D D	High accuracy OFDR system using phase integral coherence compensation and interferometer base point shift	71st Meeting on Lightwave Sensing Technology LST-11 (2024)
Akiko Tada Masaru Koshihara Takanori Saitoh	D D D	OFDR-type distributed strain measurement using phase integration coherence compensation method and interferometer base point shift	IEICE General Conference 2025, B10A-10, Mar., 2025
Masahiro Kamada Kenichi Sasaki Tomohiro Matsui	R *1 R	Comprehensive study of optical contrast, reflectance, and Raman spectroscopy of multilayer graphene	Carbon Trends 16 , 100389 (2024).
Hiroshi Shibata Kohei Watanabe Hiroshi Yamamoto Ken Ito Motohiro Banno Shinji Morimoto Yosho Kanda	*2 *2 *2 *2 M M M	Network Sections Using BER Measured in DCO Modules	IEICE Technical Report NS2024-209 (in Japanese)
Yasuhito Kajiware	I	Mass inspection and contaminants inspection technology in the pharmaceutical manufacturing process	Journal of the Society of Powder Technology, Japan, 2024 Volume 61 Issue 7 Pages 427-430
Eiji Taniguchi Junichi Sano Takashi Suzuki Hirotaka Shono Tsunoda Kotomi	I I I I I	Transmission Near Infrared Spectroscopy-Based Inspection Machine, Capable to Inspect Interior of Tablets at a Rate of 250,000 Tablets per Hour	Journal of pharmaceutical machinery and engineering 33 (4), 302-306, 2024
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Naoki Yamagishi Takahiro Matsuda Hirofumi Suganuma Tsutomu Mitsui	*3 *3 M M	Single-Reflection Point Distribution Estimation for IRS Placement Problem	2024 15th International Conference on Ubiquitous and Future Networks (ICUFN), pp. 212-217, July 2024.
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R : Advanced Research Laboratory
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