DesignCon® 2024
WHERE THE CHIP MEETS THE BOARD

Are 1.0 mm Precision RF Connectors Really Required for 224 Gbps PAM4 Verification?

Brandon Gore, Samtec Corporation
[Brandon.Gore@samtec.com]

Richard Mellitz, Samtec Corporation
[Richard.Mellitz@samtec.com]

Andrew Josephson, Samtec Corporation
[Andrew.Josephson@samtec.com]

Francesco de Paulis, University of L'Aquila
[francesco.depaulis@univaq.it]

Luis Boluna, Keysight Technologies
[luis.boluna@keysight.com]

John Calvin, Keysight Technologies
[john.calvin@keysight.com]

Rick Rabinovich, Keysight Technologies
[rick.rabinovich@keysight.com]

Mike Resso, Keysight Technologies
[mike_resso@keysight.com]

Abstract

As Ethernet and OIF Standards groups begin to wrestle with the technical feasibility of 224 Gb/s (Gbps) PAM4 signaling in order to standardize electro-optical requirements of data center communication, a particular challenge of measurement verification is beginning to emerge. Previous generations of standardization have set a bandwidth minimum target as ¾ of the baud rate. This paper will explore what is meant by bandwidth during the standardization process, the implications of test and verification attached to certain bandwidth requirements, as well as differences between acquisition range, band limited filters, and s-parameters for time domain processing. Sensitivities of 1.0 mm RF connector mechanical tolerances on measurement to simulation comparison are presented. Channel level performance metrics with the trending of current channel compliance methods are compared. The summary includes the challenges and potential mitigations ahead.

Author(s) Biography

Brandon T. Gore is presently a Principal Technologist at Samtec managing both the Signal Integrity R&D and Electronic Industry Standards teams. His research focuses are advanced interconnect materials, glass packaging, direct drive optics, and general signal integrity bottlenecks for 200Gbps data rates. He is an active contributor to both IEEE 802.3 and OIF Common Electrical I/O projects at 112 Gbps and 224 Gbps. Brandon received the PhD degree in electrical engineering from the University of South Carolina under Dr. Paul G. Huray.

Richard Mellitz is presently a Distinguished Engineer at Samtec, supporting interconnect signal integrity and industry standards. Richard has been a key contributor to IEEE802.3 electrical standards for many years. He led efforts to develop radically new IEEE and OIF time domain specification methods called COM (Channel Operating Margin) and ERL (Effective Return Loss). Early in his career he founded and chaired an IPC committee authoring the industry's first TDR standard. Richard holds many patents in interconnect, signal integrity, design, and test. Richard received the IEEE Standards Association Medallion and the Intel Achievement Award (IAA) for spearheading the industry's first graduate signal integrity programs at the University of South Carolina. Recently, Richard was honored with the DesignCon 2022 Engineer of the Year Award.

Andrew Josephson is a Technologist with Samtec focusing on emerging data rate technology development, maturation and standardization. Prior to joining Samtec he was a Distinguished Member of the Technical Staff with General Dynamics where he held various roles as an SI/PI SME, HW Designer and HW Systems Engineer focusing on niche high performance, airborne and strategic embedded computing applications. In the wireless space, he has worked on shipboard missile defense radar systems, environmental electromagnetic effects (E3) and co-site analysis and tactical spectrum management. He began his career as a signal integrity instrumentation HW designer with Wavecrest Corporation. He holds a B.S.E.E from the University of Minnesota.

Francesco de Paulis was born in L'Aquila, Italy in 1981. He received the M.S. degree in Electrical Engineering in May 2008 from Missouri University of Science and Technology (formerly University of Missouri-Rolla), USA, and the PhD degree in Electrical and Information Engineering in 2012 from the University of L'Aquila, L'Aquila, Italy. He is currently a Research Professor at the University of L'Aquila
His main research interests are in SI/PI design on PCB, packages, interposers and chips, high speed channel characterization and design, RF interference in mixed-signal system, EMI and EMC.

Luis Boluña is a Senior Application Engineer for Keysight Technologies. He has extensive experience in both the measurement and simulation of high speed SerDes architectures and backplane designs. His background is Signal Integrity and Mixed Signal Circuit Design. He has worked in Silicon Valley almost 29 years with Agilent, Cisco Systems, Rambus, Microsoft, and National Semiconductor. Luis has two US patents and has a Pioneer Award from Cisco Systems. His research interests are in system design, testability, simulation, and validation of high speed designs. Luis holds a BSEE from UC Santa Barbara with an emphasis in Solid State Physics.

John Calvin is a strategic planner and datacom technology lead for Keysight Technologies. John has been bridging the measurement science gaps of emerging Telecom and DataCom development efforts for 20 years. He has chaired multiple physical layer interoperability working groups and is currently serving as a contributing member to IEEE 802.3, OIF-CEI, InfiniBand, and PCIe development efforts. John holds a BSEE from Washington State University, and his graduate level studies are in signal processing from Stanford University.

Rick Rabinovich, IEEE802.3 Ethernet voter member, is a Distinguished Engineer at Keysight Technologies, specializing in 3D modeling of electromagnetic structures and PCB stackup optimization for 10G/25G/50G/100G/200G/400 GbE. Former IEEE Communication Society member, Rick was a Senior Principal Design Engineer at Alcatel-Lucent and an Alcatel-Lucent Bell Labs Distinguished Member of the Technical Staff. He has authored several technical articles in the IEEE Communications and EDN magazines and holds two US patents in communications. Rick's previous positions include Hardware Technology Director and Fellow Associate at Spirent Communications, and senior position at Northrop. Rick holds a BS from the Buenos Aires University Engineering College and attended computer post-graduate courses at Cal State Los Angeles, UCLA, and UCI.

Mike Resso is the Signal Integrity Application Scientist in the Internet Infrastructure Solution Group of Keysight Technologies and has over thirty years of experience in the test and measurement industry. His background includes the design and development of electro-optic test instrumentation for aerospace and commercial applications. His most recent activity has focused on the complete multiport characterization of high-speed digital interconnects using Time Domain Reflectometry and Vector Network Analysis. He has authored over 30 professional publications including two books on signal integrity. Mike has been awarded one US patent and has twice received the Agilent "Spark of Insight" Award for his contribution to the company. He received a Bachelor of Science degree in Electrical and Computer Engineering from University of California.

1 Overview

Two case studies are investigated where a DUT (device under test) is configured with two RF connector types (1.0 mm and 1.85 mm) so that comparisons can be made in the frequency domain and time domain. Additionally, in the case of the second DUT, 224 Gbps traffic is passed to determine if any impact is noticed between the RF connector types. The details of each case study are described in Section 4 to include case no.1 as a stripline DUT populated with either a 1.0 mm or 1.85 mm vertical, compression mount RF connector, as well as case no. 2 with a ganged, cabled RF connector attached to a PCB that emulates a 20 dB channel. Frequency domain (FD) data is collected using a 4 port 110 GHz vector network analyzer (VNA). This is preceded by a discussion around a key FD figure of merit for RF connectors and coaxial cables which is the cutoff frequency (f_c) of propagating higher order modes.

The time domain (TD) metrics to compare performance between 1.0 mm and 1.85 mm choices are vertical eye closure (VEC) and 12 Edge Jitter. TD performance data is captured using a high-speed oscilloscope with receiver processing capability. A 224 Gb/s (Gbps) PAM4 capable arbitrary pattern generator produces a PRBS13Q data stream which is sent through the device under test (DUT) to the oscilloscope. The oscilloscope was configured with equalization and filtering which are expected to be similar to what is deployed in 224 Gbps PAM4 devices.

High-speed digital electrical data transmission has its basis in electromagnetic wave theory. At 224 Gbps PAM4, the baud rate is 112 GBd and the Nyquist frequency is 56 GHz. The pulse width is 9 ps which in a material with permittivity of approximately 3 means that 1 pulse is 1.5 mm. What new phenomena needs to be addressed? How does this affect VEC and jitter measurement?

2 Historical Context

As Ethernet and OIF Standards groups begin to wrestle with the technical feasibility of 224 Gbps PAM4 signaling in order to standardize electro-optical requirements of data center communication, a particular challenge of measurement verification is beginning to arise. Previous generations of standardization have set a bandwidth minimum target as ¾ of the baud rate by specifying a 4th order Butterworth filter with a cutoff (fr) set to this frequency to represent the receiver bandwidth needed. For 112 Gbps PAM4 signaling, the baud rate is 56 GBd so that the bandwidth requirement target is 42 GHz. This can be affirmed by referencing the IEEE 802.3 Annex 93A.1.4.3 [1].

By setting the filter cutoff frequency to 42 GHz, the data captured during channel component verification is implicitly limited. Searching through 112 Gbps PAM4 frequency domain specifications yield a maximum frequency of 43.25 GHz [2]. This aligns to the maximum frequency of certain vector network analyzer (VNA) models that are limited by the cutoff mode of the connector used; in this case the 2.4 mm Q-band connectors have a cutoff of 50 GHz [3]. Following this historical trend, the bandwidth for 224 Gbps PAM4 would be set to 84 GHz which bypasses 67 GHz max frequency VNAs and the 1.85 mm V-band connectors (limited to 72 GHz [3,9]) and require a 110 GHz VNA and 1.0 mm W-band connectors.

2.1 Filtering Discussion for 224 Gbps PAM4 Signaling

The identification of filtering needs for testing links at 224 Gbps PAM4 is also important for evaluating the impact of device noise on the channel performance. The bandwidth limitation for the characterization of the channel will mitigate the impact of the noise on the received waveform, if assuming a random white noise. Theoretically, several noise contributions will sum up at the receiver, the random noise (SNDR at the transmitter side and η_0 at the receiver side), the crosstalk, the jitter (both random σ_{RJ} and EOJ or A_{DD}) as well residual post-DFE inter-symbol interference (ISI). From a frequency domain perspective such contributions may have an impact at different frequencies, depending on the specific physical/electrical mechanisms that are generating them. However, from a practical point of view, with the objective to evaluate the impact of bandwidth on the noise, a first approximation suggests employing a simple Gaussian process for a white noise to model the overall noise contributions. An attempt to quantify the impact of the bandwidth on the noise was carried out in [4] by applying a Raised Cosine (RCos) filter at the receiver (RX) side.

One of the questions still open in the standard forums is what should the s-parameter frequency range be? The answer will depend on the receive filter 3 dB frequency specified. A study was presented at the IEEE P802.3dj Task Force [5] that indicates that if fr equals ¾ of the baud rate, the s-parameters range needs to be as high as 90 GHz, but if fr is 0.55 of the baud rate, the maximum frequency is reduced between 70 to 75 GHz.

Ultimately channel performance is judged by its ability to transmit information. Standards like IEEE 802.3 require minimum receiver filter capability. This determines what is important for verification and measurements. The following sections discuss that relationship. VEC (Vertical Eye Closure) is used as the litmus test figure of merit for channel performance. In addition, the transmitter performance and how it relates to jitter is explored.

3 Modal and Mechanical Considerations for RF Connectors

One of the main ratings of RF connectors and coaxial cables is the cutoff frequency below which the connector operates without supporting the undesirable propagation of higher order modes. This section will discuss the modal theory, RF connector geometries of pertinent connector types, flexible coaxial cable geometry, and potential tradeoffs made to avoid RF connectors and cables with reduced modal Inter Symbol Interference (ISI).

3.1 Modal Theory

The typical signal propagation along a coaxial line is expected to be supported by the TEM mode, since voltages and currents can be uniquely determined by the expressions (1), (2) starting from the electric and magnetic fields existing on the coaxial cross-section.

$$V = -\int_l E \cdot dl \quad (1)$$

$$I = \oint_C H \cdot dl \quad (2)$$

The equations (1)-(2) allow to simplify the 3D Maxwell equations into the 1D Telegraphers' equations based on voltage and current quantities instead of electric and magnetic field quantities. However, coaxial lines, beside supporting the TEM mode propagation from DC, are able to support higher-order modes of TE or TM type. Such modes are usually kept quiet as long as the bandwidth of the propagating signal being below the cut-off frequency of the first higher-order mode. Even if higher-order modes can be excited by impedance discontinuities along the line, their amplitude decays quickly being considered evanescent modes below the cut-off. However, in the case of a signal having a bandwidth exceeding the cut-off, the first (or more) higher-order modes can propagate. Since such modes are characterized by a different propagation constant, and thus speed, signal distortion may appear at line end.
The theory for analyzing the higher-order mode propagation is quite complicated and does not add much to the discussion and objective of this paper. However, the cut-off frequency f_c of the first higher-order mode TE11 can be computed from the cut-off wave number k_c according to (3).

$$f_c = \frac{c_0 k_c}{2\pi\sqrt{\varepsilon_r}} \quad (3)$$

However, the derivation of k_c comes from the solution of (4)

$$J_n'(k_c a) Y_n'(k_c b) = J_n'(k_c b) Y_n'(k_c a) \quad (4)$$

Where $J_n(x)$ is the Bessel's function of the first kind or order n, and $Y_n(x)$ is the Bessel's function of the second kind or order n. Moreover, a and b are the diameter of the inner and outer conductor of the coaxial line, respectively. In (4) the first derivative of $J_n(x)$ and $Y_n(x)$ is present. Eq. (3) should be solved numerically; alternatively, the approximated expression (5) can be used:

$$k_c = \frac{2}{a+b} \quad (5)$$

The maximum frequency suggested by the IEEE Standard 287.1-2021 [6] are slightly lower than the cut-off, just to be safe in all applications, and allow for manufacturing uncertainties that may move the first higher-order mode within the measurement band. The numerical solution of (4) is obtained for the 5 types of coaxial connectors used in microwaves from 3.5 mm up to the 1.0 mm connectors, and they are reported in Table I.

Table I. Cut-off frequencies and maximum usable bandwidth for coaxial connectors

Type	Outer Diam (mm)	Inner Diam (mm)	f_c TE_{11} (GHz)	f_{max} (GHz)
3.5	3.5	1.5199	38.8	33
2.92	2.92	1.27	46.5	40
2.4	2.4	1.0423	56.5	50
1.85	1.85	0.8036	73.3	65
1.00	1.00	0.434	135.7	110

Moreover, the data are also plotted in Fig. 3.1 by including the cut-off values computed using the approximate expression (5). Fig. 3.1 includes the coaxial cross-section in scale for all 5 types of air lines.

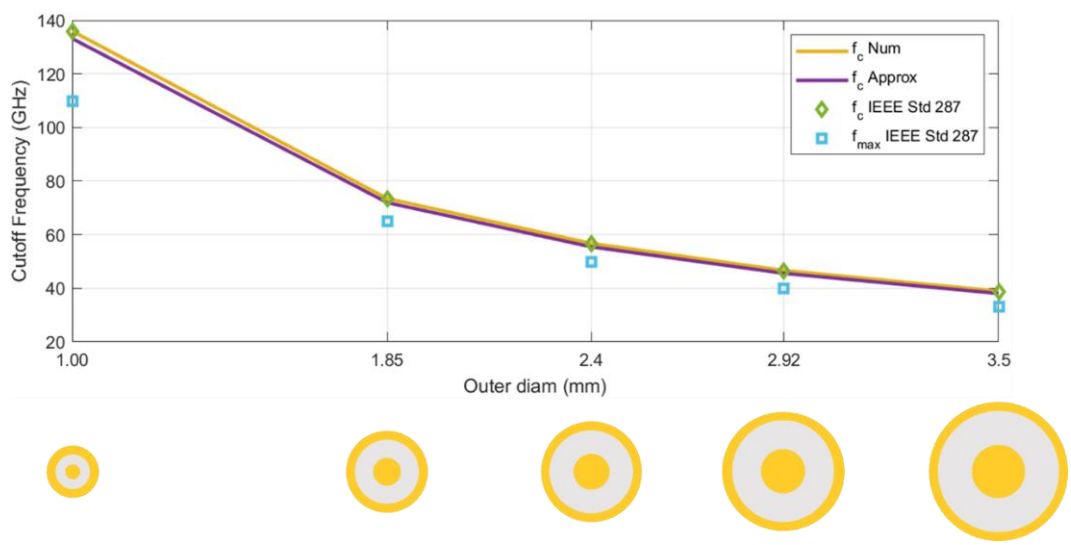


Fig. 3.1. Cut-off frequency of the first higher-order mode (TE11), and sketch of the standardized coaxial connectors (the thickness of the outer shield, not being relevant for the numerical calculation, is set equal for all).

3.2 RF Connector and Cable Geometry

There are two well established RF connector types with a maximum suggested operation frequency above the fundamental of 56 GHz for 224 Gbps PAM 4 signaling which are the 1.85 mm and 1.0 mm. Thus, these two types are considered pertinent to this paper.

Two distinct areas of interconnect are considered as having the potential to support the propagation of higher order modes. The first is the RF connector region itself which has industry defined mating interfaces and internal airline region geometry requirements. The second is the coaxial cable that mates to the RF connector when required for increased device hookup flexibility. While industry specified cross-sections for the cables do not exist, other than general measured thickness, the cable construction general consist of an inner conductor wire gauge surrounded by foamed fluoropolymer dielectric and a flat wire wrap. The cross-section geometry and material choices are designed with modal cutoff frequency targets as a figure of merit.

Fig. 3.2 shows these two regions for a 1.85 mm RF086 cable assembly with an outer cable diameter of 0.086 in (2.18 mm) next to a 1.0 mm RF047 cable assembly with an outer cable diameter of 0.047 in (1.19 mm) [7,8].

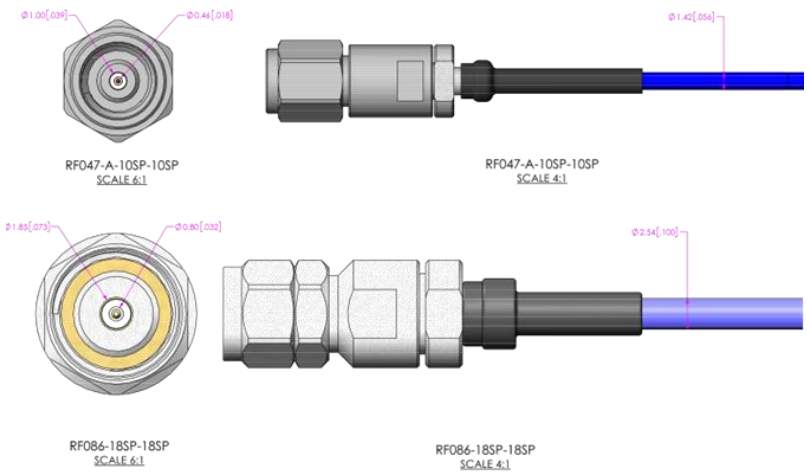
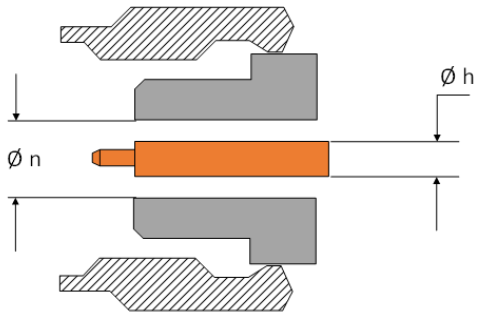


Fig. 3.2. Scaled drawing of a RF086 1.85 mm and RF047 1.0 mm cable assembly

3.2.1 1.85 mm vs 1.0 mm RF Connectors

The specification for the connectors of interest can be found in IEC 61169-32 [9] and IEC 61169-31 [10]. Fig. 3.3 provides two of the main standardized features in the airline region for instrumentation (Grade 1) grade connectors.



Connector Type	n (mm)		h (mm)	
	nominal	tolerance	nominal	tolerance
1.85	1.85	+/- 0.0127	0.8036	+/- 0.0127
1.0	1.0	+/- 0.0050	0.4340	+/- 0.0025

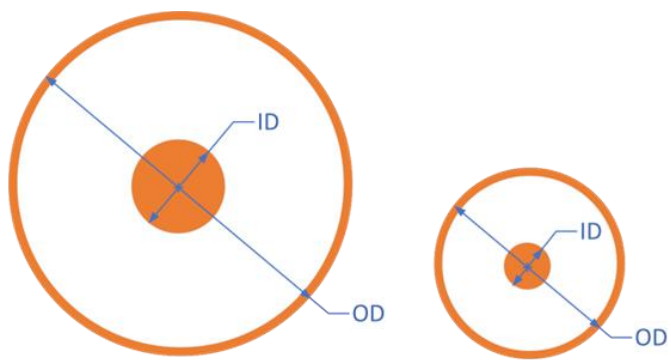
Fig. 3.3. Cut view cross-section illustration showing the geometries and tolerances of the airline regions of the 1.85 mm and 1.0 mm connector.

To extend the cutoff frequency within the RF connector, the cross-section geometry must shrink and therefore the sensitivity to the physical geometries increases. As indicated in Fig 3.3, tighter tolerances are required which for the 1.0 mm dimensions selected are 2.5 to 5 times tighter than 1.85 mm [9,10]. Though not pictured, these types of connectors require a dielectric bead to support the inner conductor. Managing the internal reflections that this bead as well as the increased electrical significance to features of geometry transition within the connector makes delivering 110 GHz of maximum frequency a complex engineering endeavor. It is also easy to sense tradeoffs around the robustness as the interface becomes more delicate in addition to potential susceptibility to part-to-part variation.

3.2.2 086 vs 047 Flexible Coaxial Cable Assemblies

The next region of an RF cable assembly to consider is the coaxial cable. Two choices of flexible cable geometry are common for both 1.85 mm and 1.0 mm RF connectors. These are 086 and 047 coaxial cables.

In constructing flexible coaxial cables, a likely practice is to use a foamed a fluoropolymer dielectric that lowers the effective dielectric constant in the range of 70% to 80% compared to velocity of propagation in air [11]. The inner conductor wire gauge diameter (ID) and dielectric thickness then determine expected cutoff frequency. Example calculations and comparison of cross-section at scale are shown in Fig. 3.4.



Cable Type	ID (in)	OD (in)	Cutoff (GHz)
086	0.023	0.086	72
047	0.011	0.047	112

Fig. 3.4. Side view cross-section illustration showing 086 coaxial cables as 4x the cross-section area as 047 coaxial cables.

A perhaps non-intuitive trade off in the shrinking of the coaxial cross-section to extend the cutoff frequency for higher order modes is the increase in cable attenuation (insertion loss) due to the increase in AC resistance of the smaller coaxial cross-section. Fig. 3.5 contains an insertion loss plot of example RF047 and RF086 cable assembly samples at a length of 152 mm [7,8]. Even over this short distance, the extra loss accumulated is 1.75 dB at 56 GHz. Admittedly, these constructions are slightly different so that all the degradation cannot be attributed to AC resistance; however, it does illustrate a potential

reduction in reach and thus cabling usability in a testing environment where high modal bandwidth is needed.

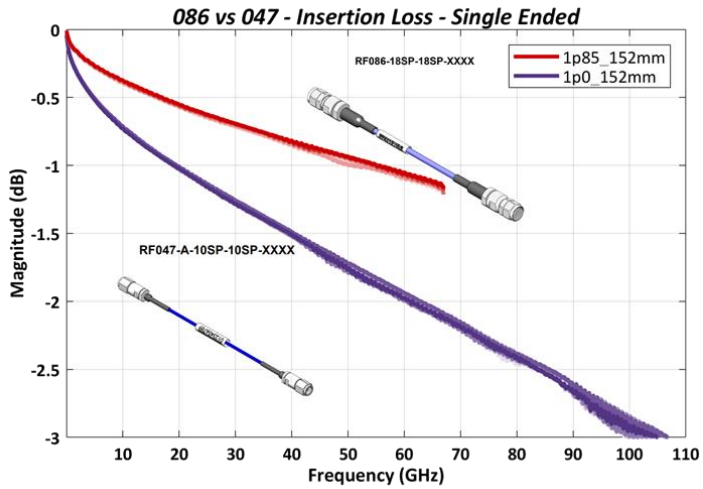


Fig. 3.5. Insertion loss comparison of Samtec RF047 (purple) and RF086 cable assembly (red) samples at a length of 152 mm

Both conductor and insulator geometry and material characteristics determine the electrical insertion loss performance of the coaxial cable behind the connector. As with trends in electronic packaging, dimensional shrinking increases AC conductor losses due to finite conductivities in conductor metals and surface plating, presenting a natural tradeoff between insertion loss from DC up to Nyquist versus modal bandwidth the cable can support up to the connector with defined and specified f_c. In addition to the AC signal losses from conductor cross section reduction, state of the art coaxial cable manufacturing methods are challenged to achieve the higher velocity factors in the smaller diameter higher modal bandwidth cross sections that are achievable, mature and cost effective in the larger diameters. Thus in practice, larger diameter coaxial cable for terminating larger diameter 1.85 mm connectors often benefits from higher velocity factors, lower dielectric constants, and fewer wavelengths per unit length as a result further differentiating the coax cable/connector model bandwidth tradeoff.

3.2.3 3D EM Modeling of Modal ISI Signature

A 3D EM field solver modeling exercise is presented with a simple geometry of 086 coaxial cable at a length of 20 mm to intuit the signature of higher order modes on the insertion loss. A uniform 3D EM modeling produces no ISI signature or visible modal field behavior of higher order mode propagation. However, a slight perturbation created in the uniform cross-section by introducing a slight offset in the center conductors at the midpoint in the model excites the expected TE11 mode propagation.

Fig. 3.6 contains a vector plot of the electric field in the coaxial cross-section at the predicted modal cutoff frequency. This figure also contains the longitudinal electric field strength magnitude along the length of the coaxial model which displays the propagation of TE11 modes. With the 3D EM field solver model containing a modal impedance

discontinuity, the insertion loss magnitude shows a signature of modal ISI. The signature appears as a narrow band resonance at the cutoff frequency.

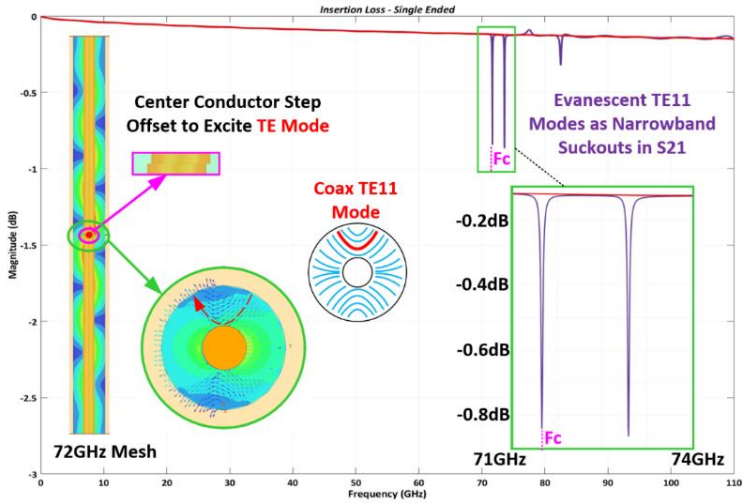


Fig. 3.6. 3D EM field solver investigation of

While it is out of scope of this paper to attempt a 3D EM field solver modeling treatment on properly simulating higher order modes, the authors were satisfied with the intuition provided by this simplistic modeling exercise.

4 Case Studies on the Impact of 1.0 mm Connectors in the Frequency Domain

Two case studies are presented that compare 1.0 mm RF connectors in similar interconnect environments as 1.85 mm RF connectors. The first is a simplistic DUT consisting of two lengths of stripline measured with both connector types that are compression attached directly to the PCB. The second case study is conducted on a graduated loss platform that is PCB based and uses ganged RF cable assemblies to achieve enough density to emulate a x4 channel to enable testing of a SERDES quad. Passive measurements are discussed in this section. Active measurements passing 224 Gbps PAM4 signaling are analyzed in Section 5.

4.1 Case Study No. 1: Compression RF Connectors on a Printed Circuit Board

An effort to characterize PCB materials to 100 GHz for 224 Gbps PAM4 readiness [12,13], was initially completed with 1.0 mm compression mount connectors mounted on a test board containing stripline DUTs of several lengths and routing angles. The internal layers were accessed with single layer microvias as illustrated in Fig. 4.1. The vertical transition region, sometimes referred to as the Break Out Region (BOR), was designed in a 3D EM Field Solver for bandwidth of 100 GHz. This coaxial-to-planar transition region is of great interest to determine high bandwidth performance of high-speed digital interconnect. Minimizing any reflections in the BOR by designing a controlled impedance environment is critical to achieving high performance channels.

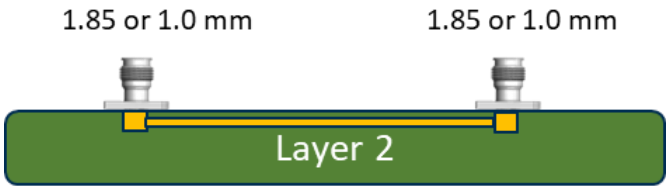


Fig. 4.1. Simple illustration of the printed circuit board configuration

During the measurement phase, it was noted that the 1.0 mm compression connectors did not perform as well as expected. As an experiment, the measurements were re-taken by swapping out the with 1.85 mm compression mount connector and measured using a 67 GHz VNA. The change in RF connectors allowed for a change in the device measurement setup.

Fig. 4.2 contains the 110 GHz VNA, 1.0 mm test setup which shows the measurement heads are supported perpendicular to the board and 152 mm, 1.0 mm test cables attach to the 1.0 mm compression mount connectors. All single ended traces are populated with 1.0 mm compression connectors.

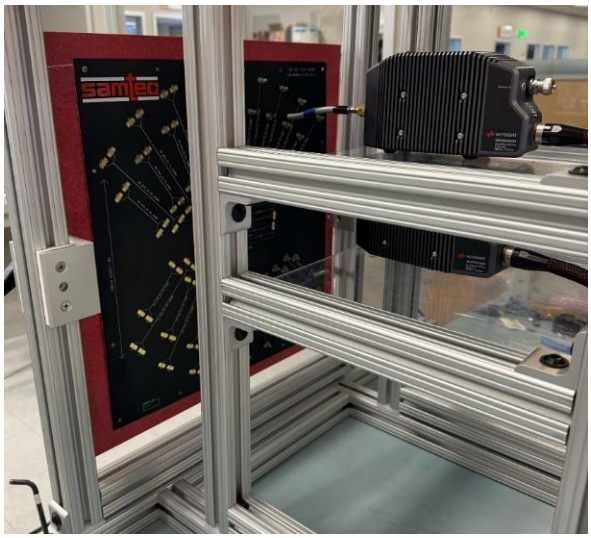


Fig. 4.2. 110 GHz VNA, 1.0 mm test setup

Figure 4.3 displays the 67 GHz, 1.85 mm test setup where the 1.0 mm connectors have been replaced with 1.85 mm connectors. Note that the 67 GHz VNA does not require remote measurement heads, and that 1.85 mm measurement cables are used which are longer.

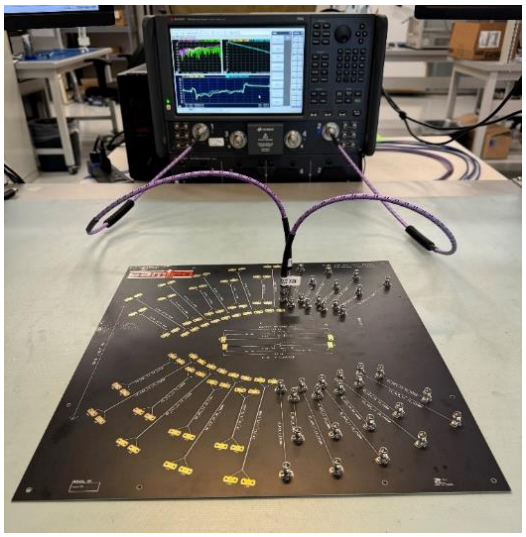


Fig. 4.3. 67 GHz, 1.85 mm test setup

When the short, 50 mm stripline DUTs are compared as in Fig. 4.4, the 1.0 mm populated striplines have a degraded return loss especially in the frequency range of 40 GHz to 55 GHz.

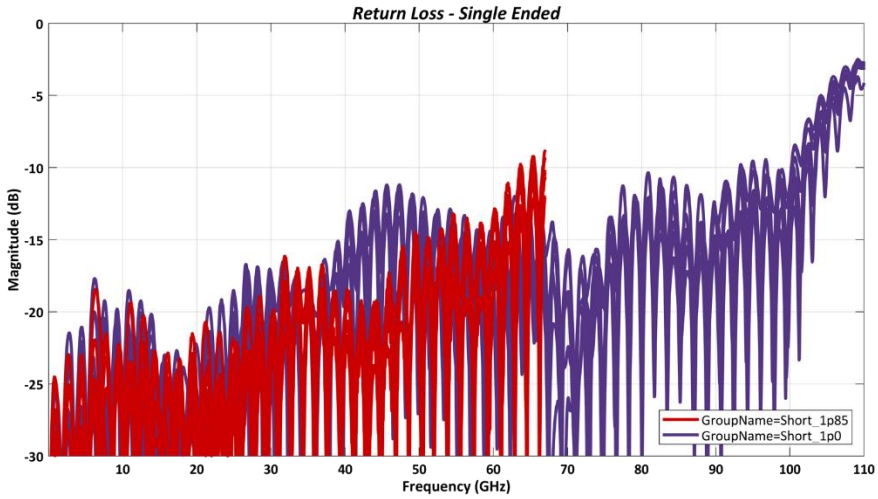


Fig. 4.4. Comparison of 1.0 mm populated stripline return loss (purple) and 1.85 mm populated stripline return loss (red)

One of the regions identified as the culprit of the degraded performance was the interior features of the 1.0 mm connector itself. The remaining reflection miscorrelation were attributed to not having a high enough fidelity 1.0 mm 3D EM model from the vendor to allow for adequate design of the BOR.

Fig. 4.5 contains the TDR response converted from the frequency domain and excited with a 6 ps risetime. The time scale is plotted to highlight the RF connector body and BOR region which ends at 0.08 ns. The more complex internal structure of the 1.0 mm connector as well as the greater part to part variation is apparent even within a small sample size of 80 captured responses.

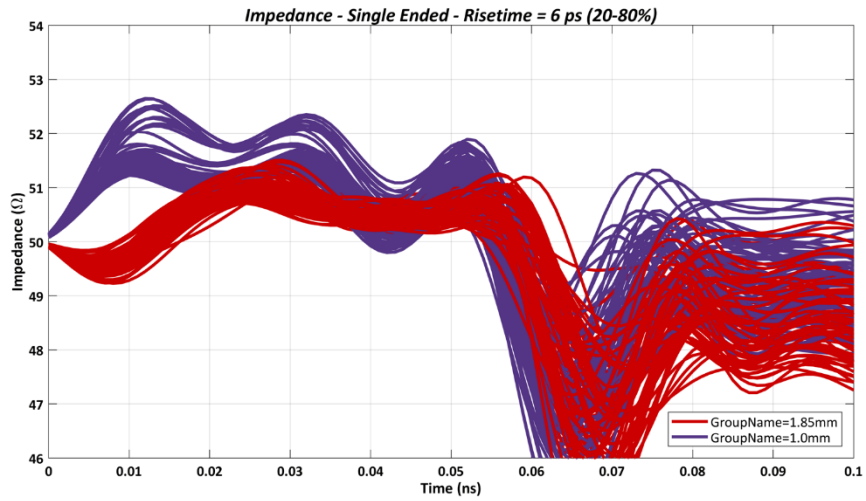


Fig. 4.5. Comparison of 1.0 mm populated stripline (purple) and 1.85 mm populated stripline (red) TDR responses

4.2 Case Study No. 2: Graduated Loss Platform for Multi Lane Channel Emulation

This case study uses a DUT that is more complex than the previous, but it more closely emulates a channel response. This graduated loss platform was designed as a reference for Serializer-Deserializer (SERDES) characterization at 112 Gbps and 224 Gbps using ganged and cabled RF connectors to achieve greater density over vertical, compression RF connectors. A desirable feature of the test platform is to allow for the selection of precision insertion loss targets by moving Samtec BE70A [14] cable assemblies around a PCB that is mounted in an integrated frame and protective shroud. Scaling from single lane characterization to x4 Serdes quads, often sharing common reference clocks, is possible by using the x8 coaxial cable counts from Samtec Bulls Eye® [14] connector body configurations in one, scalable platform.

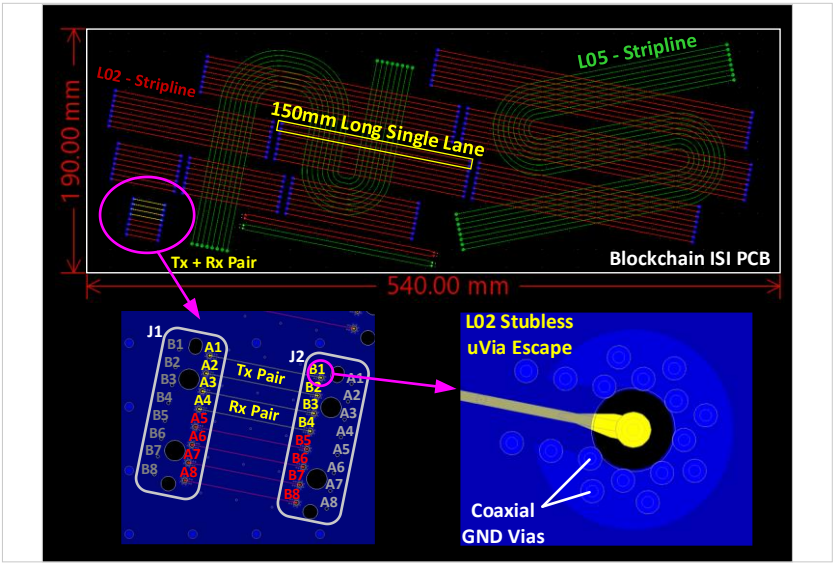


Fig. 4.6 Features of Variable Loss Platform

The PCB stackup utilizes single layer microvias for precision stubless stripline escape routing from the Bullseye connector bodies. Breakout region impedance match bandwidth to 80 GHz is obtained through 3D EM field solver optimization, GND via hole pattern, via to plane clearances, and via to trace launch geometries [15]. PCB laminate materials and copper foils are chosen to provide a simple relationship for 224 Gbps testing. Each step in PCB length represents an additional 2.5 dB of insertion loss around 56 GHz. For 224 Gbps testing, there are 10 available PCB DUT wire lengths ranging from 25 mm to 250 mm in 25 mm steps providing for 25 dB of dynamic range.

A complete BE70 cable to PCB to BE70 Cable channel loss of 20 dB at 56GHz was chosen corresponding to 150 mm BE70 cable lengths and a 150 mm long PCB transmission line loss DUT. Cable sets were skew matched to < 500 fs.

For the purposes of this paper, the BE70 cables were terminated with a non-standard configuration of 1.0 mm RF connector terminated to the 150 mm length of RF086 coaxial cable. This custom configuration was compared with the standard configuration using 1.85 mm RF connectors terminated to 150 mm length of RF086 coaxial cable. As picture along with the test setup in Fig. 5.8, metrology grade 1.85 mm to 1.0 mm adapters are required to measure the 1.85 mm configuration up to 110 GHz.

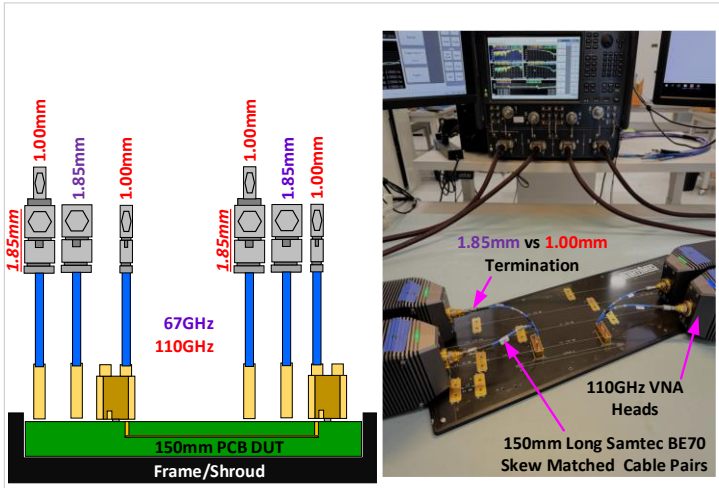


Figure 4.8: 1.85mm Vs. 1.00mm Characterization Test Setup

In total, three different cable configurations were characterized mated to the same 150 mm PCB trace length.

- Standard 1.85 mm termination on RF06 Coax to 67 GHz
- Standard 1.85 mm termination on RF06 Coax to 67 GHz BW extended to 110 GHz with metrology grade adapters
- Standard 1.0 mm termination on RF086 Coax to 110 GHz (connector exceeds coax BW)

4-port s-parameter data was collected allowing comparisons of frequency domain.

4.2.1 DUT Frequency Domain Measurements

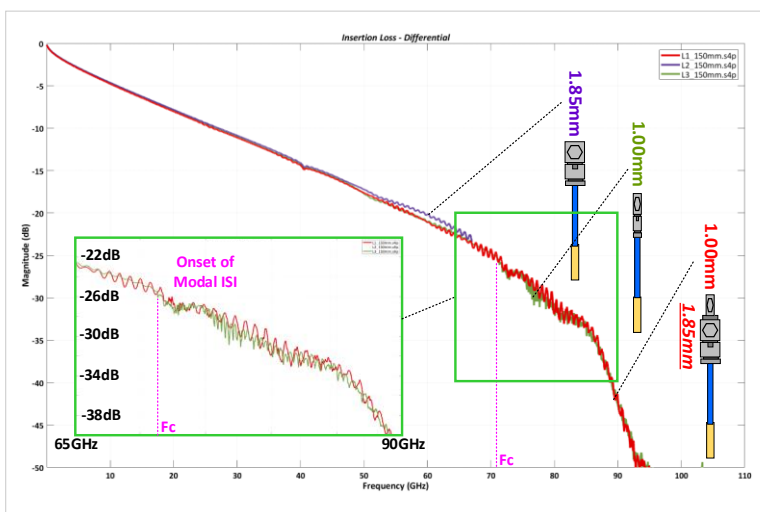


Figure 4.9: Insertion loss of the various cable terminations into 150 mm PCB trace length

The frequency domain insertion loss comparison of the three cable test configurations mated to the 150 mm PCB trace length is given in Fig. 4.9. The channels show clean

insertion loss performance up to, and exceeding, the modal cutoff frequency (f_c) of the RF086 coaxial cables. This allows for the existence of potentially resolvable time domain ISI that is higher order modal in nature. The signature presents itself as higher order insertion loss ripple above f_c being more identifiable in the 1.0 mm terminated cable assemblies. The 1.85 mm configuration introduces adapters which likely introduce additional return loss in addition to the termination impedance matching achievable in a 1.85 mm used beyond the designed frequency range. These reflections appear to dominate the modal ISI present in the measurement.

Since the higher order modes are themselves evanescent, the modal loss mechanisms are anticipated to be near reflectionless penalties and the effect may be limited in the instrumentation grade setup.

The impact of the modal ISI introduced by coaxial cable cross-section which is present in both the 1.0 mm and 1.85 mm RF connector configuration is hard to quantify. The signature is certainly apparent in the insertion loss. If considering an entire interconnect, it is reasonable that the modal ISI would be dominated by TEM impedance mismatches.

Additionally, frequency domain parameters are themselves out of context of actual device perception. Fig. 4.10 contains the various filters applied during channel compliance methodologies of IEEE 802.3 [16] using the trending setting from recent P802.3df taskforce meeting [17].

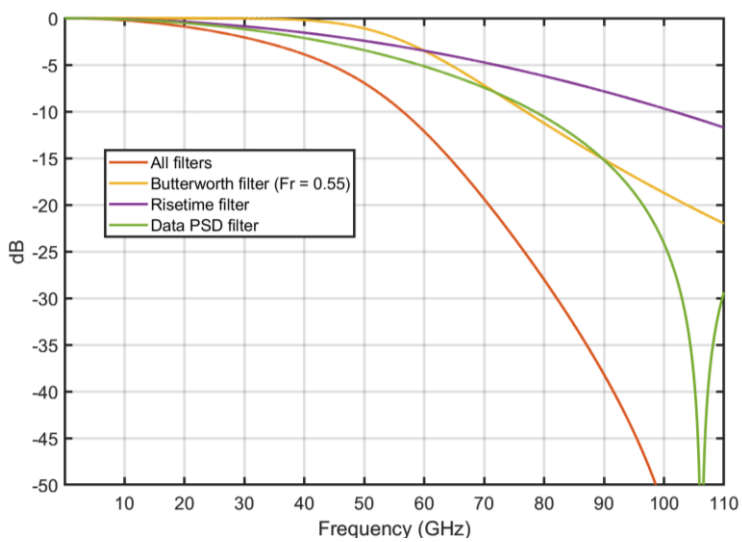


Figure 4.10: Filters responses applied during channel compliance methodologies in IEEE 802.3

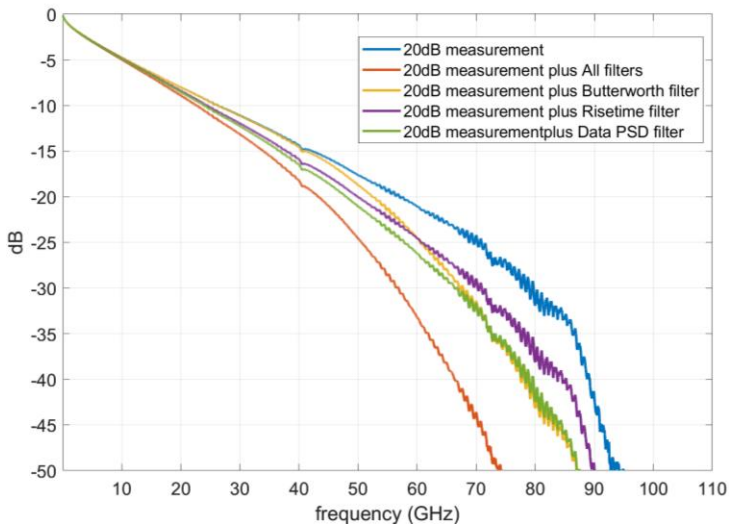


Figure 4.11: Filters responses applied to the measured 1.85 mm, 20 dB measurement.

In Fig. 4.11, these filters are progressively applied to the measured insertion loss of the 1.85 mm configuration. The insertion loss is greatly increased and therefore the signature of model ISI is no longer observable.

5 Passing 224 Gbps PAM4 Signaling over the Channel Emulation Platform

5.1 Measurement Setup Description

Time domain measurements were performed on the Channel Emulation Platform to determine VEC and Output Jitter (also known as 12 Edge Jitter) across different frequency response bandwidths. A description of the test setup is as follows, a Keysight M8050A 120 GBd Bit Error Rate Tester with a 1.0 mm M8059A remote head was used to deliver a 224 Gbps PAM 4 PRBS13Q signal at 800 mV differential amplitude to the variation of cables and adapters to the board under test. The output of the board under test is connected to a pair of 110 GHz Keysight N1046A Remote Heads plugged into a Keysight N1000A DCA Oscilloscope. Fig. 5.1 shows a picture of this setup.

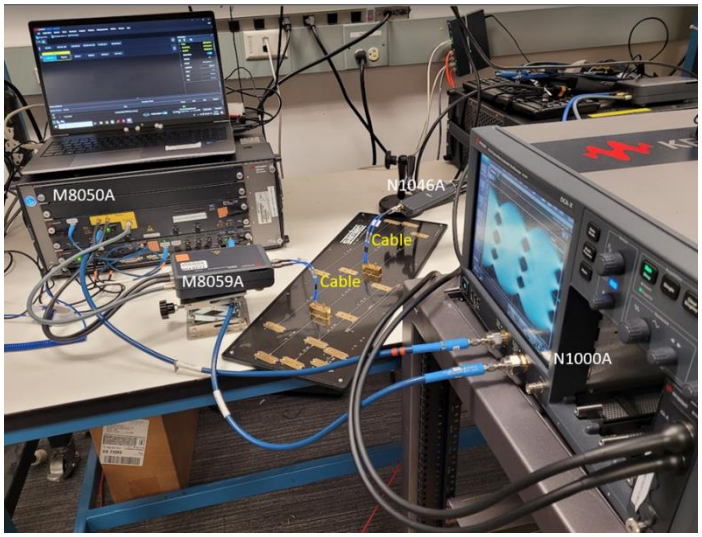


Fig. 5.1 Lab setup for time domain meausrements

In the paper we have discussed three cable configurations (see section 4.2). Given the fact that the N1046A remote heads are natively 1.0 mm an adapter (Keysight 11901 1.0 mm Male to 1.85 mm Female, see Fig. 5.2) would be needed to make it interface with the 1.85 mm RF06 Coax Cables. And the second configuration connects natively with no need for adapters to the 1.0 mm RF06 Coax Cables. To summarize, only two physical setups are needed.

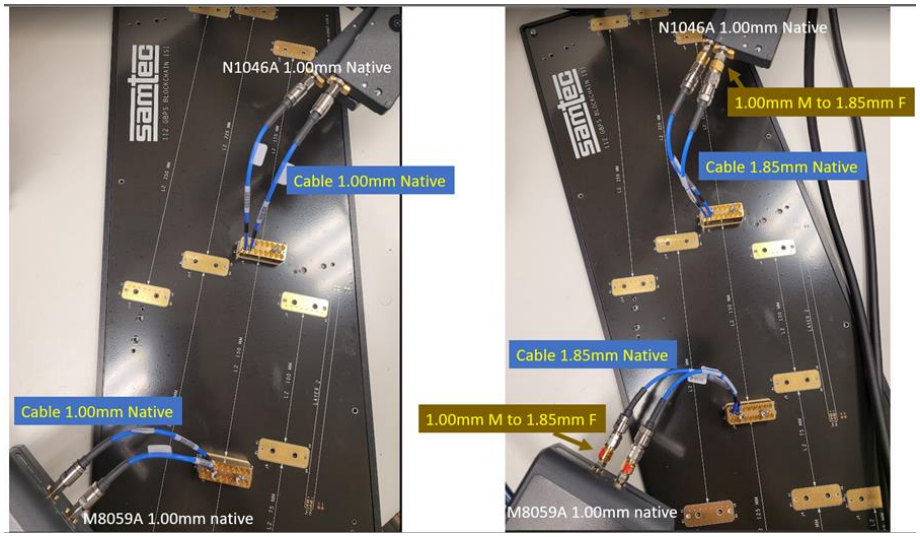


Figure 5.2 Both physical test cases including adapter 1.0 mm M to 1.85 mm F

5.2 VEC and Jitter measurement at 224 Gbps

An illustration between 1.0 mm and 1.85 mm instrumented connectors is depicted in Fig 5.3 as two setup cases. The measurement comparison using VEC and jitter was determined with the Keysight N1000A DCA oscilloscope. The 112 GBd (224 Gbps) PAM 4 PRBS13Q signal with a 800 mV differential amplitude was driven using the M8050A pattern generator and remote head M8059A. The Keysight N1000A DCA oscilloscope was used as a receiver using a cascade of a 4th order Butterworth filter, CTF

(continuous time filter) in Fig 5.4, a feed forward equalizer with 6 pre cursor taps and 23 post cursor taps, and finally a 1 tap DFE. In addition, the differential channels were de-skewed within the oscilloscope. The 3 dB loss frequency (fr) of the Butterworth filter was set to 56 GHz, 67 GHz, and 85 GHz.

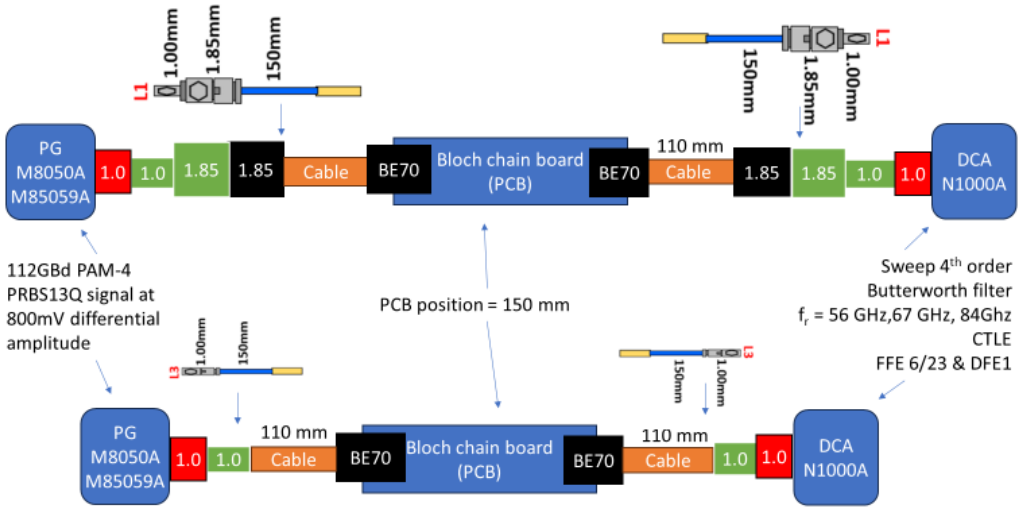


Figure 5.3: Two case studies

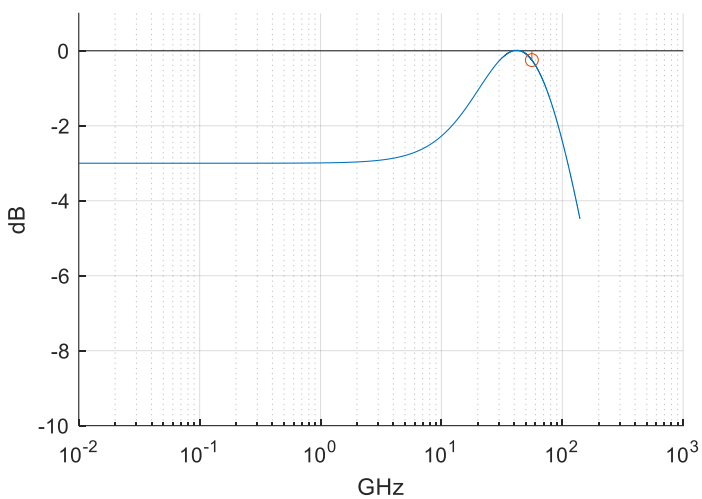


Figure 5.4: Receiver CTF Response

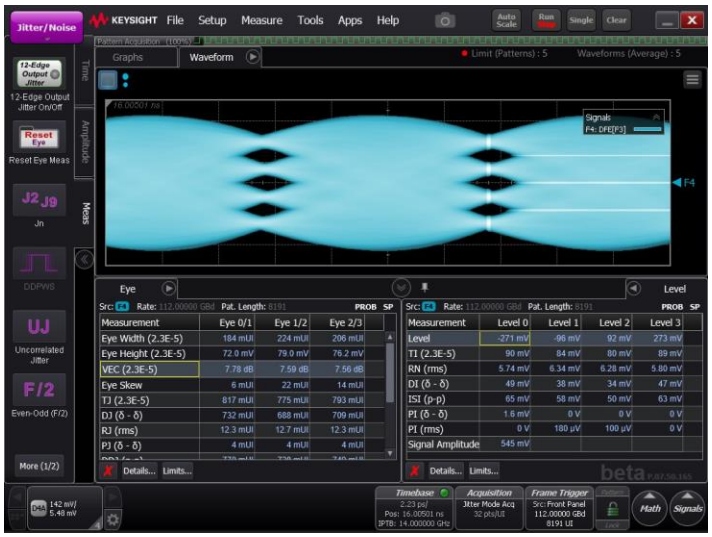


Fig. 5.5 Oscilloscope screen capture of 1.0 mm RF connector type and fr of 84 GHz

The result for the two connector setups and the fr settings are showing in Table II and example of the oscilloscope screen is shown in Fig 5.5 of the 1.0 mm RF connector type (line 3 of Table II).

Table II: VEC and jitter measurement summary

Connector Type	fr (GHz)	VEC01 (dB)	VEC12 (dB)	VEC23 (dB)	J3u (mUI)	Jrms (mUI)	EOJ (mUI)
1.0	56	7.43	6.58	6.93	178	22.7	22
1.0	67	7.19	6.24	6.59	203	25.3	22
1.0	84	7.78	7.59	7.56	198	25	20
1.85	56	7.91	7.1	7.49	187	23.6	22
1.85	67	7.97	7.07	7.96	211	26.1	20
1.85	84	9.39	9.04	10.4	200	25.2	20

The 1.0 mm and 1.85 mm connector cases show little VEC difference for fr of 56 GHz and 67 GHz (Table II) while there is much larger delta with a fr of 84 G. It is encouraging to note that all cases are below the historical 112 Gbps PAM4 VEC limit of 12 dB. Interestingly enough, the IEEE P802.3dj and OIF 224 Gbps PAM4 projects are trending to 0.5 times the baud rate which would be 53.125 GHz and 56 GHz respectively. Figure 5.6 is another way of viewing Table II. Lower fr seems to mitigate the higher frequency impacts on VEC caused by the 1.85 mm connector.

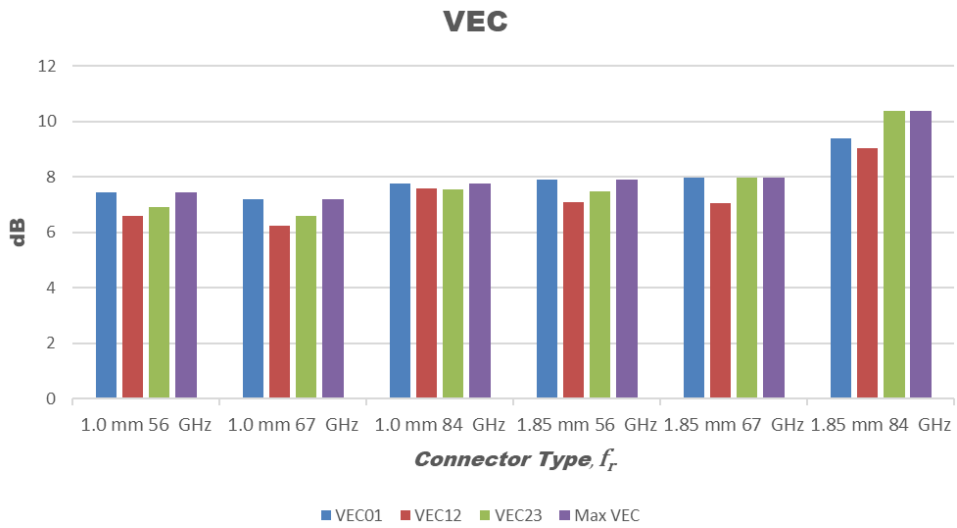


Fig. 5.6 VEC measurements

Jitter measurement insight can be inferred from the delta in jitter between 1.0 mm and 1.85 mm connectors tempered by fr. It is interesting to note that jitter measurements, as opposed to VEC measurements, have not used the Butterworth receiver filter in the past. Instead, a 4th Order Bessel Thomson filter [1] is applied to these phase sensitive measurements.

Conclusions and Future Work

The conclusions about 1.0 mm applicability seem to be somewhat mixed. Since 224 Gbps PAM4 standards are trending toward using receiver filters with a cutoff filter around Nyquist frequency either 1.0 mm or 1.85 mm connector seem OK for receiver compliance or VEC testing. Since the requirements for bandwidth when measuring jitter has been higher in the past, 1.0 mm connectors may be appropriate for those device characterizations.

A note of caution has emerged from this work. Although 1.0 mm RF connectors promise higher frequency domain fidelity in the range beyond the rating of 1.85 mm RF connectors, there is likely to be additional usability constraints, increased costs, as well as manufacturing penalties as described in Section 3.2.

There are several items left for future work. As noted in Sections 4.2 and 5.2, both the VNA and Oscilloscope mating required a metrology grade adapter to mate with the 1.85 mm RF connectors. We did not investigate the impact of a lower bandwidth pattern generator or oscilloscope that would remove the need for these adapters.

The channel emulation DUT was populated with RF086 coaxial cable for both the 1.0 mm and 1.85 mm configurations which means that the propagation of higher order modes was still possible. A comparison at the same DUT loss on the smaller diameter RF047 coaxial cable would have been interesting to pursue. But, due to time constraints and reduction in cable reach to achieve similar insertion loss, this is left as future work.

This paper focused on the time domain metrics as would be employed for host or receiver compliance. The impact of RF connector type and receiver filter cutoff for transmitter devices would be an interesting future consideration.

Advanced receiver techniques such as maximum likelihood sequence estimation (MLSE) are becoming more common and are potentially required for 224 Gbps PAM4 long reach electrical channels. Exploring the impact on reducing the receiver bandwidth filter cutoff when these techniques are employed could further help inform the sufficiency of channel bandwidth requirements.

References

[1] "IEEE Standard for Ethernet," in IEEE Std 802.3-2022 (Revision of IEEE Std 802.3-2018), vol., no., pp.1-7025, 29 July 2022, doi: 10.1109/IEEESTD.2022.9844436.

[2] "Common Electrical I/O (CEI) – Electrical and Jitter Interoperability agreements for 6G+ bps, 11G+ bps, 25G+ bps, 56G+ bps and 112G+ bps I/O", OIF Implementation Agreement OIF-CEI-05.0, Website OIF_doc_FM12.book (oiforum.com)

[3] "Radio-Frequency Connectors - Part 40: Sectional Specification For 2.4 Series RF Connectors" IEC 61169-40 Ed. 1.0 b:2010

[4] L. Boluna, J. Calvin, F. de Paulis, R. Mellitz, R. Rabinovich, M. Resso, "A novel approach to 224 Gb/s reference receiver design using raised cosine response for noise mitigation", *DesignCon 2023*

[5] R. Rabinovich, "212 Gb/s PAM4 per Lane C2M Channels Frequency Range and Rx Filter" IEEE P802.3df 200 Gb/s, 400 Gb/s, 800 Gb/s, and 1.6 Tb/s Ethernet Task Force, May 2023, online: https://www.ieee802.org/3/dj/public/adhoc/electrical/23_0504/rabinovich_3dj_elec_01b_230504.pdf.

[6] "IEEE Standard for Precision Coaxial Connectors at RF, Microwave, and Millimeter-Wave Frequencies--Part 1: General Requirements, Definitions, and Detailed Specifications," in *IEEE Std 287.1-2021*, pp.1-136, 16 Sept. 2022, doi: 10.1109/IEEESTD.2022.9889249.

[7] "RF086-18SP-18SP-XXXX RF Characterization Report", Feb. 2023, online: 20230309_t-3314602_rfc_rf086-18sp-18sp-xxxx.pdf (samtec.com)

[8] "RF047-A-10SP-10SP-XXXX RF Characterization Report", Feb. 2023, online: 20230914_t-3829369_rfc_rf047-a-10xx-10xx-xxxx.pdf (samtec.com)

[9] "Radio-frequency connectors - Part 32: RF coaxial connector with inner diameter of outer conductor 1,85 mm (0,072 in) with screw coupling – Characeristic impedance 50 ohms (type 1,85)", IEC 61169-32 Ed. 1.0 b:1999

[10] "Radio-frequency connectors - Part 31: RF coaxial connector with inner diameter of outer conductor 1,0 mm (0,039 in) with screw coupling – Characeristic impedance 50 ohms (type 1,0)", IEC 61169-32 Ed. 1.0 b:1999

[11] "Precision RF Micowave/Millimeter Wave"Precision RF Micowave/Millimeter Waves Interconnect Solutions Guide", Samtec Inc., pg.12, Aug. 2020 online: https://suddendocs.samtec.com/literature/samtec_precision_rf_design_guide.pdf"

[12] Gore, "Measurement Evaluation of PCB Electrical Performance 200 Gb/s PAM4" IEEE Project P802.3df - IEEE P802.3df 200 Gb/s, 400 Gb/s, 800 Gb/s, and 1.6 Tb/s

Ethernet Task Force, Sept 2022, online: Measurement Evaluation of PCB Electrical Performance 200 Gb/s PAM 4 (ieee802.org)
[13] B. Gore, “Next-Generation PCB Loss Analysis”, Signal Integrity Journal, June 12, 2023, online: Next-Generation PCB Loss Analysis | Signal Integrity Journal
[14] “BE70A-S-18SP-2-02-0152 High Speed Characterization Report”, Samtec Inc. Feb. 2023, Online: 20230228_t1567406_hsc_be70a-s-18xx-2-02-xxxx_rev01.pdf (samtec.com)
[15] S. Tucker, “PCB Stackup and Launch Optimization in High-Speed Designs”, *DesignCon 2022*
[16] "Annex 93A," in IEEE Std 802.3-2022 (Revision of IEEE Std 802.3-2018) , pp.6703-6711, 29 July 2022, doi: 10.1109/IEEESTD.2022.9844436.
[17] R. Mellitz, “200 Gb/s per lane KR Backplane Objective Proposal” IEEE P802.3df 200 Gb/s, 400 Gb/s, 800 Gb/s, and 1.6 Tb/s Ethernet Task Force, May 2023, online: 200 Gb/s per lane KR Backplane Objective Proposal (ieee802.org)