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Don't Judge a Bit Just by Its Fourier: 112 Gbps PAM4 Component Optimization and Selection

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Abstract
With a new generation of "56-112 Gbps components" of varying performance levels coming to the market, distinction and selection is increasingly important yet very difficult. Categorizing components using traditional frequency domain responses does not always correlate to channel performance, as the channel takes a holistic time-domain simulation approach to compliance. This paper explores data-rate and spectral content dependent component evaluation that is sensitive to the nuances important to channel performance. Using component level Effective Return Loss (ERL), Integrated Crosstalk Noise (ICN) and their interaction, a diverse set of components are evaluated and quantitatively correlated to the respective 56-112 Gbps channel performances.

Author(s) Biography
Steve Krooswyk works as a New Product SI Engineer for high-speed connectors at Samtec. His 15 years of signal integrity experience has focused on the design, simulation, and test of high-speed interconnect and I/O, with an interest in correlation. Previously, Steve was the tech lead for PCI Express SI at Intel's data center division. He is an author of the book titled, “High Speed Digital Design: Design of High-Speed Interconnects and Signaling” and holds a Master’s degree from the University of South Carolina.

Madhumitha Rengarajan is a New Product SI Engineer at Samtec where she designs high speed connectors. Some of her specific interests include determining the characteristics that enable a passive channel to successfully transmit data at a given data rate, designing products that meet the characteristics, monitoring standards organizations for new passive channel evaluation techniques, formulating simulation techniques for high speed connector and footprint designs, and improving correlation between simulation and measurement. Previously, she led the team responsible for high speed connector design at FCI/Amphenol ICC. She has a Master’s degree in Electrical Engineering from Penn State University and has 11 years of signal integrity experience. She has many patents issued and several pending in high speed connector design realm.

1. Introduction
High-speed PHYs will necessitate accompaniment of higher performing interconnect, much of which will still be copper, as a new generation of data center equipment emerges. Driven by up-coming Ethernet standards, data rates will trickle to a downstream need of 56 Gbps to 112 Gbps per lane, and system manufacturers will face challenges to select sufficient interconnect. Component manufacturers will encounter similar difficulty as they work to design products that meet meaningful criteria.

Typically, interconnect design and selection is completed via frequency domain response analysis. Although certain parameters (such as Insertion Loss, Return Loss, Crosstalk, etc.) are traditionally considered independently during design and assigned a “set of limits,” there are multiple disadvantages to such an approach. Weaknesses include: limits for each parameter are set independent of each other (uncoupled); limits also do not correlate to the spectral density corresponding to the data rate they are meant to represent; in the worst case, a single value is applied across all frequencies, even beyond Nyquist; and violations to channel limits may have little-to-no impact on channel performance leading to excessive false negatives.

During the design cycle, choosing between two concepts with opposing good qualities is always a challenge. In addition, a component with a well-matched impedance profile (and hence RL) does not gain any corresponding crosstalk requirement relief during selection. The uncoupled metrics therefore limit the solution space, impede innovation, increase cost, and limit interconnect density.

When establishing frequency domain limits, experienced engineers already know to de-rate the magnitude significance as frequency increases; however, noise and return loss magnitudes tend to increase with frequency. It is not decisively clear how important a high return loss, or a high crosstalk, is at the end of the spectrum of interest, therefore spectrally relevant metrics are necessary to provide meaningful pass/fail criteria.

Failing “limits” do not preclude the poor function of parts within a system. For example, crosstalk resonances created in interconnects are a common source of frequency limit violation that may have little-to-no impact on system transmission. The correlation of uncoupled frequency domain limits to channel performance metrics (such as eye height, eye width and COM) is questionable at best. Interconnect evaluation and selection using uncoupled frequency domain responses leads to excessive false failures and component over-design, making it both the most conservative and the most expensive method.

2. History of Alternate Metrics

Many passive component specifications have realized the adequacy of frequency domain limits in establishing good correlation-to-channel performance. As the data rates increased from 10 Gbps to 25 Gbps and beyond, many industry-leading specifications have introduced new metrics that support or supplant the use of frequency domain limits in component evaluations. This section, while not an attempt to create a complete list, highlights a selection of alternative non-frequency domain and/or spectrally relevant metrics that are/were under consideration for qualifying various passive interconnect form factors.

2.1 IEEE802.3 ba

First introduced in 2007 at IEEE 802.3, ICN is a single-value, area under the curve, spectral dependent crosstalk parameter that is weighted per the operating frequency and gives more significance to relevant frequencies [1]. ICN eliminated one of the major drawbacks of the frequency domain limits: uncorrelated specification excursions and channel performance. Typically, ICN limits for channels are set as a function of the loss of the system (shown in Figure 1). The simultaneous consideration of system loss and ICN as a metric allows more loss for low noise components. Section 3.1 of this paper describes this metric in further detail.

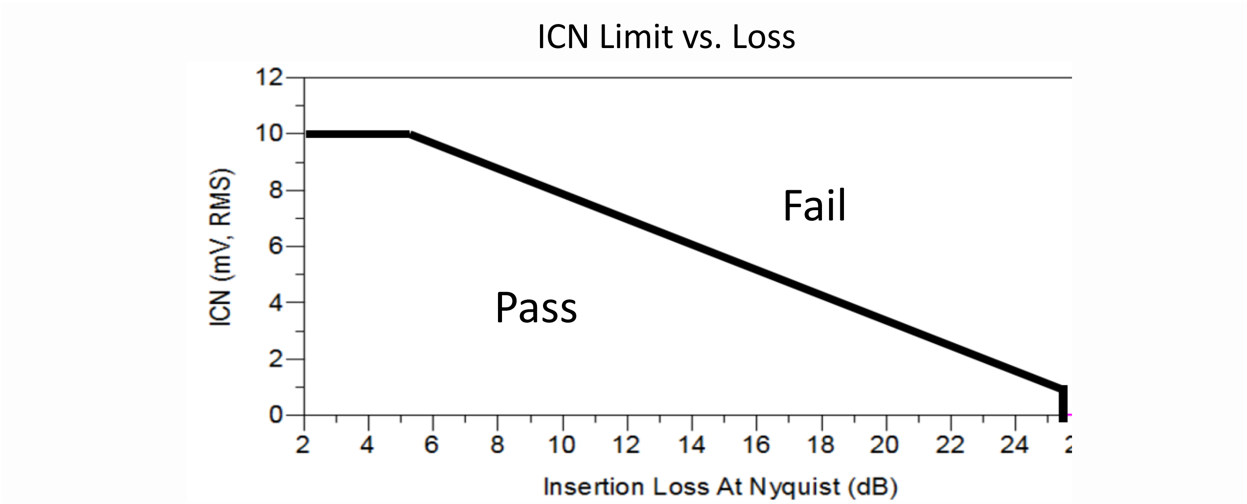

Figure 1: IEEE802.3ba ICN vs IL at Nyquist Limit [1]

2.2 Universal Series Bus (USB Cable and Connector Specification)

UCB Type-C specification [2] for the SuperSpeed Passive Cable assemblies specifies an area under the curve value for all the frequency domain metrics. Metrics such as Integrated Multi-reflection (IMR), Integrated Return Loss (IRL) and Integrated Crosstalks (INEXT and IFEXT) were introduced as the normative metrics in place of traditional Insertion Loss, Return Loss and Crosstalk metrics. These metrics quantitatively specify the impact of reflection on a single value specification that plots the area under the curve against the loss in the system. As an example, Figure 2 shows the IMR and IRL limits set forth by the USB-C specification, defined as a function of the fitted loss at Nyquist.

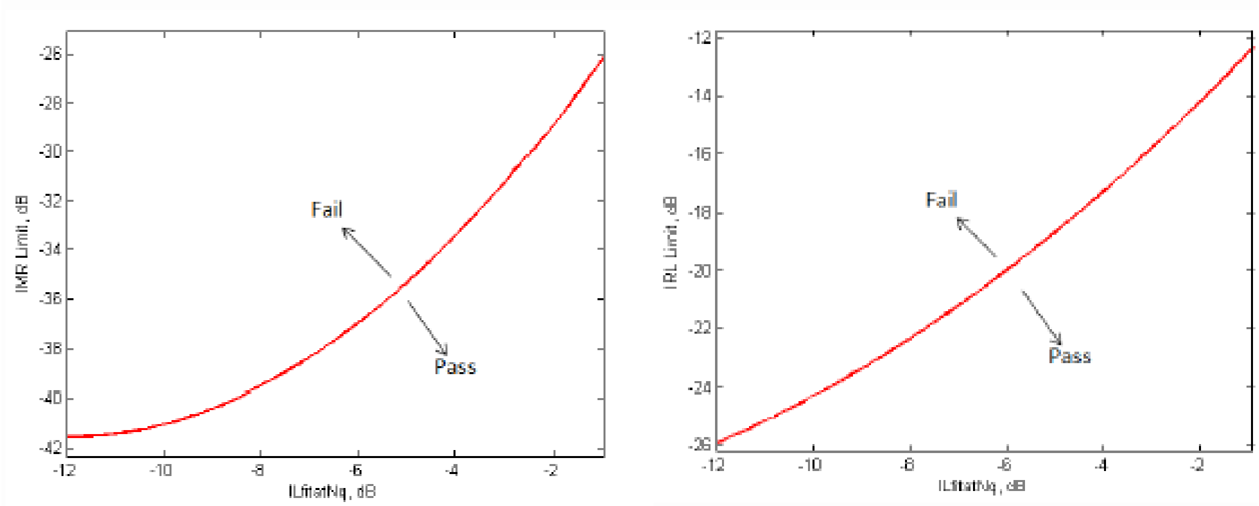

Figure 2 IMR and IRL limits from USB-C Specification [2]

2.4 PCISIG Interconnect Specification

Recently, PCISIG events have facilitated the discussion of spectrally relevant connector metrics. In the work presented at PCI-SIG Developers Conference [3], it is shown that components with and without excursions to frequency domain limits. Figure 3 shows no difference in channel performance when crosstalk was analyzed using the Seasim channel evaluation tool.

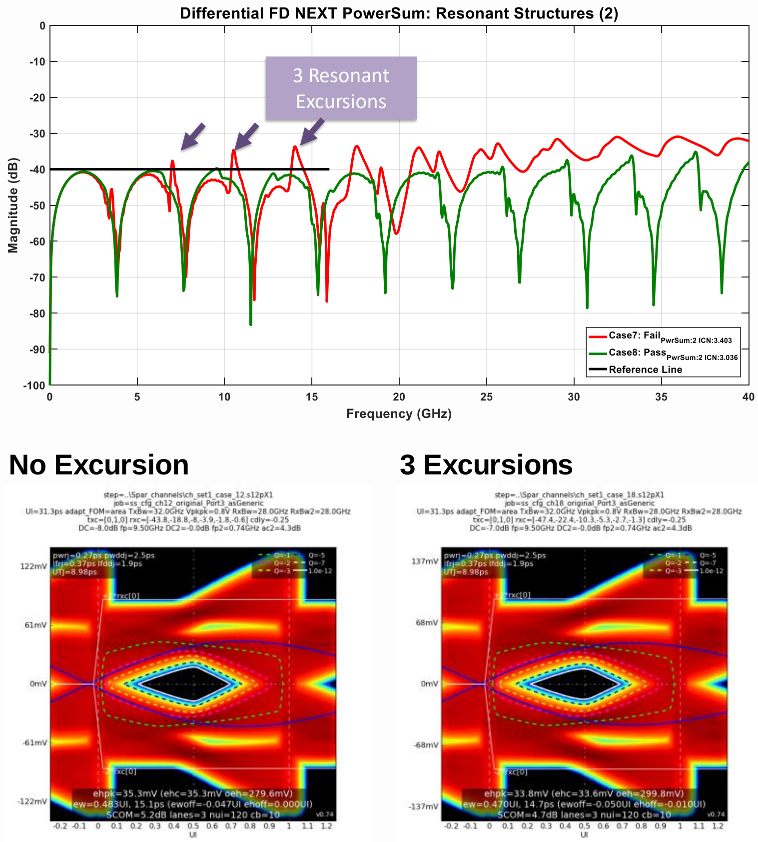

Figure 3 Excerpt from PCIe DEVCON - Excursion of XT limits have no impact on channel [3]

2.5 IEEE802.3ba and Optical Internetworking Forum (OIF) Channel Specifications (25+ Gbps per lane)

Many industry-leading specifications, such as IEEE's 802.3 and OIF, have moved to holistic approaches to channel evaluation, including Channel Operating Margin (COM). Starting with 25 Gb/s specifications, the IEEE and OIF have, as shown in the equation below, used COM as the channel metric for most of their channel specifications. IEEE and OIF are also the leading specifications in defining the channel metrics for alternate signaling technology such as PAM4.

$$COM = 20 * log10(As/N)$$
$$N = Peak\ BER\ noise \text{ and } As = Peak\ BER\ Height$$

Peak BER noise in COM includes, ISI, crosstalk, TX/RX noise sources (jitter, duty cycle distortion); peak BER height includes TX/RX equalization and signal conditioning. COM calculation also includes package impact on channels analyzed. While COM is not a component specification, others have argued crosstalk excursions/resonances do not impact channel COM performance [4].

As a crosstalk component, ICN has been historically calculated and included in COM; the effects of reflection while included in the form of "noise," however, were not quantified as a metric in the very first COM calculations. This has caused specification users to estimate the return loss requirements at the channel level especially for PAM4 signaling, a signaling type more sensitive to return loss impacts than NRZ. "Effective Return Loss" (ERL) has been recently introduced at IEEE; intended to capture the effect of channel return loss while simultaneously including the insertion loss and DFE effects [5], ERL is a single number metric (like COM). Since its introduction, ERL has been the momentum in the IEEE and OIF standards bodies.

3. Channel and Component Metrics Under Consideration

Our industry recognizes the inadequacy of frequency domain metrics as tools for channel and component evaluation. Effort has been made to define spectrally relevant channel metrics that consider the signal conditioning present at the transceiver. The authors of this paper would like to extend the approach to components and present a component evaluation and selection metric that is holistic, spectrally relevant, and correlated to channel level performance. In addition, the requirements for NRZ vs. PAM4 signaling at the component level are also reviewed.

The channel metric used for correlation is COM (both 56G NRZ and 112G PAM4 versions). Using component level ERL, ICN, and their interaction, a diverse set of components are evaluated and quantitatively correlated to their respective 56-112 Gbps channel performances. Optimization trade-offs are possible by establishing an ERL-ICN to COM sensitivity relationship. The analysis is done for NRZ vs. PAM4 signaling schemes and the differences in requirements are detailed. The component metrics used for correlation to the channel metric are component level ICN and ERL. Note that the component level metrics are specifically called out as "DUT xxx" or subscripted with term "conn" to differentiate that the metric is calculated on the component level only (no channel).

3.1 Integrated Crosstalk Noise (ICN):

ICN is a spectral dependent parameter that weights the power for a specified operating frequency, giving more significance to relevant frequencies. The spectral weighting is created using a power weighting filter (PWF). The PWF includes a sinc function set to the operating frequency to represent the relevant spectral content. The function is then modified by filters representing the transmitter and receiver I/O bandwidths. The function is not modified for any encoding scheme. However, the final integration may exclude low frequency points. The below equations show the calculation used to find the ICN values from S-parameters.

$$PWF(f) = UI \cdot sinc(UI \cdot f)^2 \cdot \left(\frac{1}{1 + \left(f \frac{T_r}{0.2365} \right)^4} \right) \cdot \left(\frac{1}{1 + \left(\frac{f}{F_{RX}} \right)^8} \right)$$

$$ICN = \sqrt{2 \cdot \Delta f \cdot \sum_n PWF(f_n) \cdot \sum_x A_x^2 \cdot |SDD21XTK_x(f_n)|^2}$$

Where:
UI – Unit interval (s)
f – Nyquist frequency (Hz)
Tr – Transmitted (20-80)% Risetime (s)
F_{RX} – 3 dB receiver bandwidth (Hz)
A_x – Amplitude of the crosstalk aggressors (V)
Δf – Frequency step (Hz)

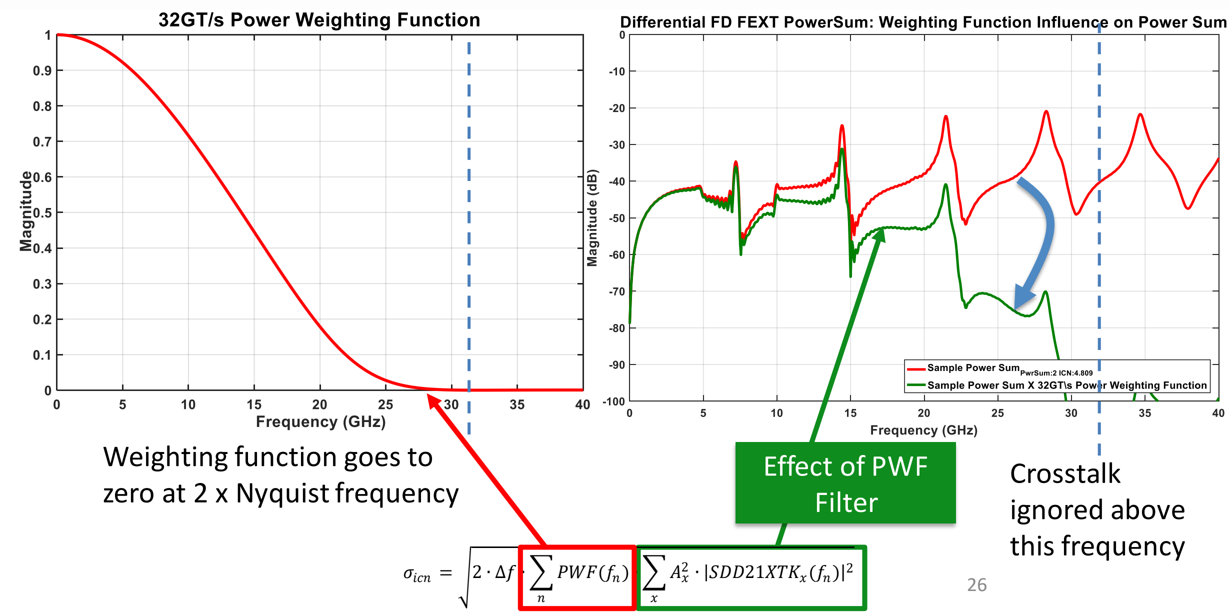

Figure 4 ICN Calculation: Impact of PWF on Crosstalk [3]

Figure 4 shows how the PWF makes the crosstalk spectrally relevant for a given data rate. As frequency is increasing, relevant magnitude is decreasing until no energy exists at twice Nyquist.

3.2 Effective Return Loss (ERL)

ERL is a return loss metric which is adjusted for the insertion loss inherent in return loss and adjusted for the presence of a DFE. Like COM, ERL is a single value metric which can be derived from return loss. A Pulse Time-Domain Reflectometry (PTDR) waveform is the basis for ERL determination and can be extracted either from s-parameters using the return loss terms or from a step TDR waveform. As described in [5], the steps to calculate ERL is as follows:

Effective reflection waveform, $\boldsymbol{R_{eff}(t) = PTDR(t)\ TG_{rr}(t)\, TG_{loss}(t)}$

$\boldsymbol{TG_{rr}(t)}$ and $\boldsymbol{TG_{loss}(t)}$ are time gated weighting functions that respectively account for losses in the channel. They are defined as:

$$\boldsymbol{TG_{rr}(t) = \begin{cases} 0, & t < T_{fx} \\ \rho_x(1+\rho_x)e^{-\frac{\left(\frac{t-T_{fx}}{T_b}-(N_b+1)\right)^2}{(N_b+1)^2}}, & T_{fx} \le t \le T_b(N_b+1) + T_{fx} \\ 1, & otherwise \end{cases}}$$

$$TG_{loss}(t) = \begin{cases} 0, & t > T_{fx} \\ 10^{\frac{\beta_x\left(t - T_{fx} - T_b(N_b+1)\right)}{20}}, & T_{fx} \le t \le T_b(N_b+1) + T_{fx} \\ 1, & otherwise \end{cases}$$

Where:
N_b is the number of DFE taps in the signaling architecture
T_b is the time for one symbol (aka UI)
t is time in seconds
T_{fx} is the time when a signal emerges from the test fixture
ρ_x is the effect average system reflection outside the DUT
β_x incremental available signal loss factor

$\boldsymbol{R_{eff}(t)}$ is super-positioned from many bits/symbols (PRBS) and a probability density function (PDF) for aggregate reflections is calculated. From the PDF, a cumulative density function (CDF) is generated. Finally, the probability of error as dictated by signaling and error correction schema is used to select the ERL value along the CDF curve.

There are multiple advantages to using ERL: It is a single value specification that lends itself well for grading designs; it removes the penalty for short packages; it incorporates the effects of signaling architecture; it directly correlates to COM; it is a parameter driven specification that can unify the specification across multiple sections such as packages, TX/RX, channel etc. [5].

3.3 Channel Operating Margin (COM)

Detailed presentations of COM have been given by [6]. [1] also shows extensive technical description of the COM calculation. [7] provides an excellent tutorial on how the COM calculations are carried out in today's IEEE standards.

COM, a time domain system budget for channels and devices, was introduced in the 2015 100Gb/s IEEE802.3 standard for Ethernet backplanes and copper cables. It provides a unifying budget for the whole system. COM specifies a minimum signal to noise ratio computed in the time domain from a pulse response. This signal-to-noise metric is closely tied to the base band nature of the high-speed differential digital signaling. The following steps as outlined in [8], explains how COM is computed.

1. Compute total channel gain $H_{tot}(f)$ from linear filters and channels s-parameters
    a. Combine channel S-parameters with reference Tx and Rx package S-parameters to get S-parameters of complete channel, $H_{21}(f)$.
    b. Add Tx/Rx low pass filters
2. Compute Single Bit Responses (SBR) from convolution of:
    a. 1 UI wide source of appropriate amplitude
    b. Tx FFE filter $H_{TxFFE}(f)$
    c. The pole/zero CTLE filter $H_{CTLE}(f)$
    d. Through channel $H_{tot}(f)$
3. Compute available single amplitude (S_x), search for optimal Tx FFE setting, and CTLE setting (NRZ/wo FEC example)
4. Determine interference signals (without symbol gain)
    a. Create SBRs for Thru, NEXT, FEXT

b. For all channels, perform a linear transformation on H_{tot} (step 1) with $H_{CTLE}(f)$ CTLE setting found in step 3.
c. For Thru and FEXT, additionally perform a linear transformation with $H_{TxFFE}(f)$ FEXT, and NEXT interference signals are the above SBRs
d. The Thru interference signal is the Thru SBR accounting for "Ref Rx Equalizer"

5. Compute interference PDFs for some number of sampling phases (32) and select the PDF for the worst variance and then compute a CDF.
   a. Create PDFs for the Thru, all NEXT, and all FEXT channels using convolution with the PDF of a symbol for the port type and sampled interference signals.
   b. Join allowance for other effects with the joint PDF created from all channel PDFs
   c. Compute a CDF computed using the cumulative sum of the joint PDF.
6. Compute COM as the dB ratio of the available signal amplitude (S_x) to the CDF voltage at the specified raw BER probability.

$$\boldsymbol{COM = 20 * log10(As/N)}$$
$$N = Peak\ BER\ noise = As - Peak\ BER\ Height$$

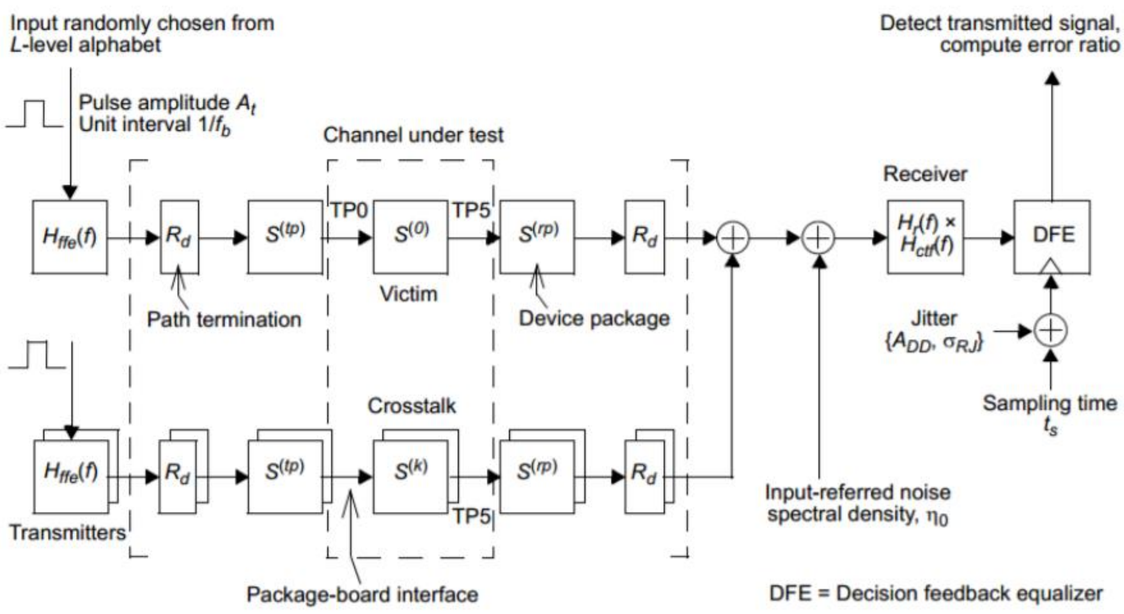

Figure 93A–1—COM reference model

Figure 5: IEEE COM Model from Annex93A [1]

Figure 5 shows the IEEE block diagram of the above-mentioned process. All the above computations can be accomplished using the MATLAB® code provided for download at [9]. The accompanying configuration file enables the customization of multiple parameters and what if scenarios. More details on how to use the scripts and parameters is also outlined in [7].

4. Channel Description

A block diagram of channels used for evaluation are presented in Figure 6. The components of the channel evaluated are:

1. Traces on PCB1
2. Connector like structure
3. Traces on PCB2

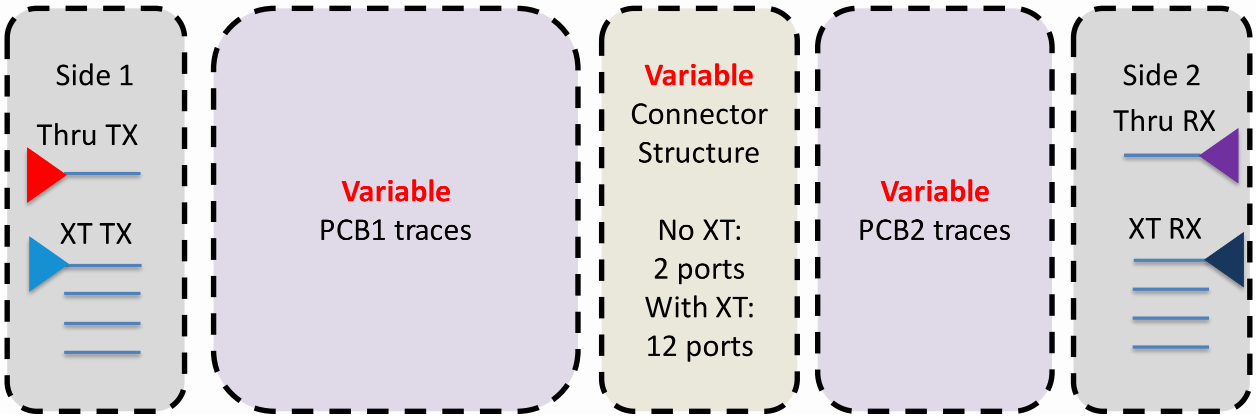

Figure 6: Channel Topology Evaluated

4.1 "Connector-like" structure

In this paper, channels were evaluated with and without crosstalk to isolate the impact of component return loss and crosstalk. The approaches taken to get the interconnect models needed to get the diversity in thru and crosstalk performances are described in this section.

4.1.1 Modeling Throughs

A diverse range of connector through responses were desired for evaluation. Connector modeling is traditionally completed in 3D EM software, where solve time is lengthy. Alternatively, the authors chose to represent a connector's "through path" with a series of transmission lines. Impedances are alternated to represent the discontinuities observed through a connector and the PCB breakout region. These alternating impedances may represent for example a via barrel that could be inductive or capacitive, a low impedance from BGA or edge card attachment, a high impedance through the mating beam, and other discontinuities existing in an interconnect and its associated PCB attach.

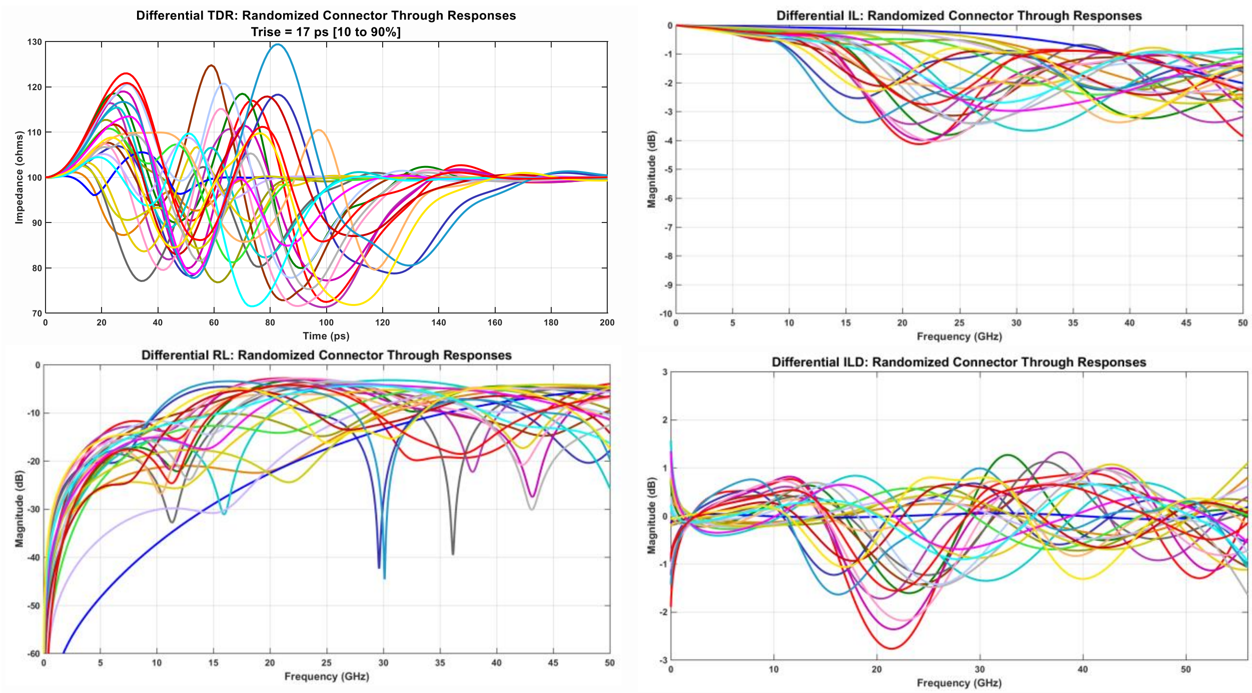

Figure 7: From top left – Impedance, Insertion Loss, Return loss and ILD of Chosen Through Responses

As shown in Figure 7, twenty-five different transmission line through-structures with randomized impedance magnitudes were created with discontinuities in the range of 100+/- 20 ohms at a 17ps (10-90%) rise time. Further, lengths of the transmission lines were varied creating structures varying from 18 to 67ps propagation delay. These variations lead to RL magnitudes from mild to severe below the Nyquist frequency of 28 GHz (corresponding to 56 Gbps NRZ or 112 Gbps PAM4). The variety in RL leads to a good variety in IL and ILD of the models. Significant reflections appear as ripple on the insertion loss profiles. The deviations can be measured as ILD with most responses within +/- 1.5dB from the fitted IL.

4.1.2 Modeling Component Crosstalk

A "connector-like" 3D model is created with varying lengths and dielectrics to create the crosstalk variety needed. Note that one such geometry is shown in Figure 8.

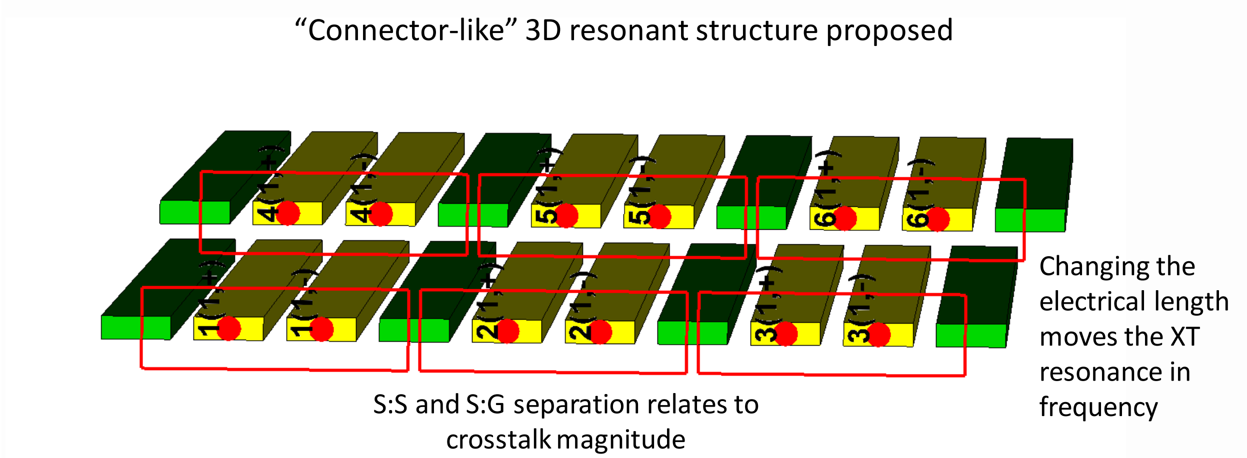

Figure 8: "Connector-like" Structure Used to Generate the Crosstalk Models Required

The crosstalk component used for analysis in this paper is FEXT. Power-summed FEXT of the components chosen for analysis is shown in Figure 9, and include the corresponding ICN calculations.

Significant crosstalk sources, exceeding 30dB, where included in the analysis to evaluate the sensitivity to far-end aggressors – those which may be tolerated in a high loss environment. The test channel in this paper includes 17.5 dB of PCB losses at 28 GHz. The crosstalk power sums are combined with the PCB losses to characterize the crosstalk entering the receiver package on the right side of Figure 9.

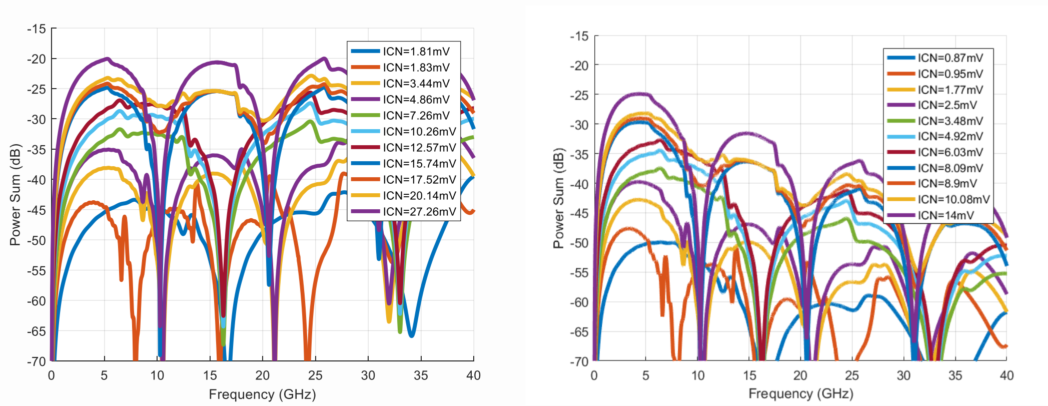

Figure 9: Crosstalk Power Sums (left) Connector Only (right) Connector with PCB Channel

4.2 PCB Traces – PCB1 and PCB2

Total trace length to be used in the channel established as the maximum loss at Nyquist under consideration at the IEEE standards [IEEE] for 400 Gbps is near 28 dB. Additional consideration that was used in setting the maximum loss is a channel length with a COM value near 3.25 dB without any connector included. The maximum allowable loss at Nyquist frequency was split between the traces of PCB1 and PCB2 as shown in

.

Some of the channel ILs are shown in Figure 10. Note that the DUT has an impact on channel Insertion Loss Deviation (ILD). Also, as seen from the IL plot with packages, the packages add a substantial amount of loss and ILD to the channel loss. Notice that the IL for a variety of RL/ERL is given here as shown in the RL plot in Figure 11.

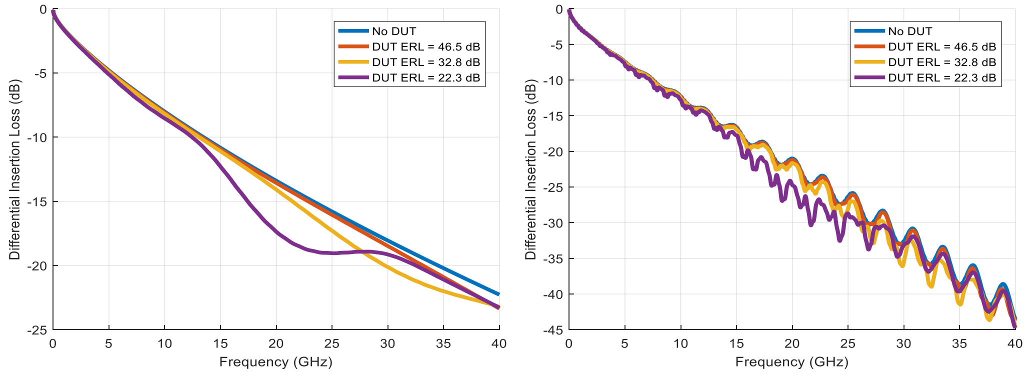

Figure 10: Insertion Loss of Total Channel Without and With Packages (Various Connectors Included)

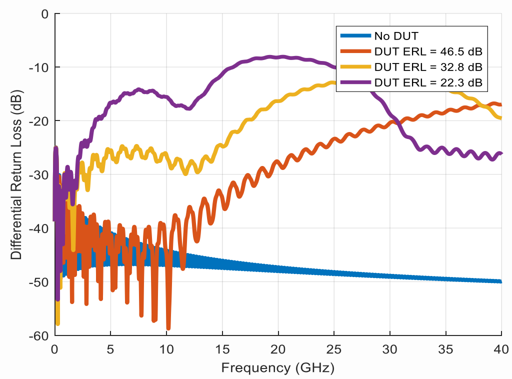

Figure 11: Channel RL Without Packages

5. Channel Evaluation Method

The two metrics under consideration for a component specification are: ERL and ICN. The channel metric used for correlation is COM. This section of the paper provides data that correlate component level ERL and ICN to their respective channel level counter parts. In addition, this section also lists the various parameter values used for ICN, ERL and COM calculations.

5.1 Validating ERL as a Component Metric

The correlation of component ERL to channel level ERL is yet to be established. The authors, for each of 25 through-connectors, computed the ERL value of the connector only (DUT ERL) and compared it to the corresponding channel level ERL.

As shown in the Figure 12, the channel metric follows or correlates to the component metric except for an offset due to the channel loss. Note that the packages are excluded from ERL_{Chan}. This correlation provides the confidence to derive a meaningful component metric from component level ERL (ERL_{conn}).

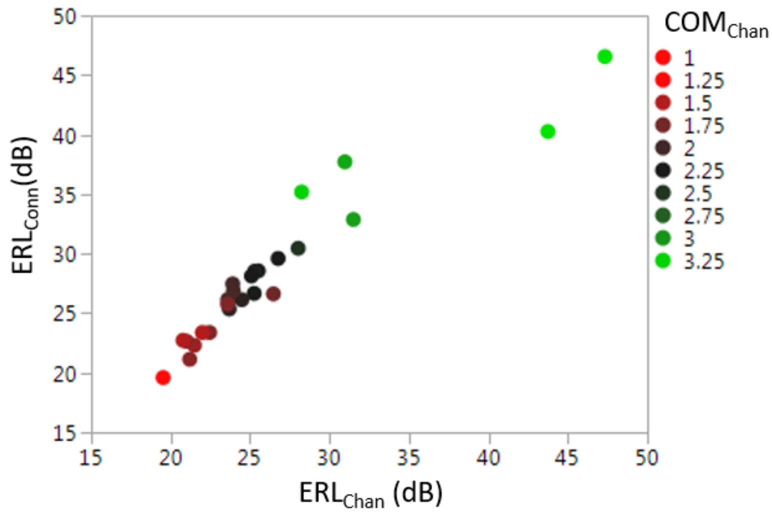

Figure 12: ERL_{conn} vs. ERL_{chan} at 112G PAM4

5.2 Validating ICN as a Component Metric

The only crosstalk contributing structure in the channel is a "connector-like" component. Without any crosstalk in the PCB, it can be expected that the magnitude of tolerable ICN_{Conn} levels will be high. This shows the channel is reducing the crosstalk entering the receiver by 60%. The ICN_{Chan} calculation does not include the package losses and would otherwise be further attenuated.

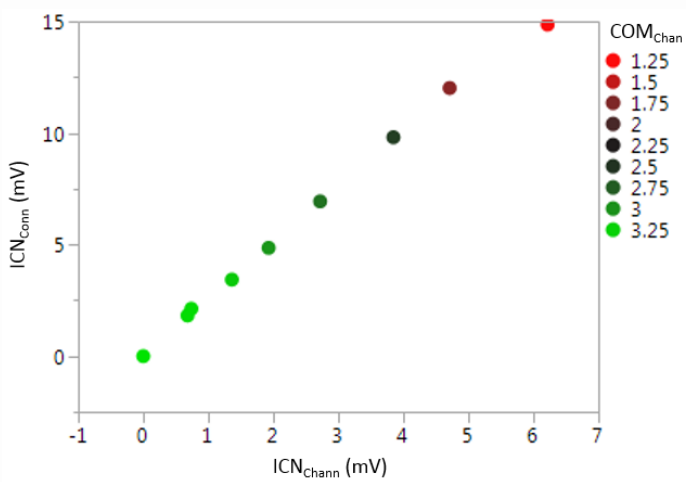

Figure 13: ICN_{conn} vs. ICN_{chan} at 112G-PAM4

5.3 ICN, ERL and COM Settings

Standards bodies have not been consistent in specifying ICN input parameters and applications, namely the IEEE and OIF. The settings used for our analysis can be found in Table 3. Also, it may reasonable to scale ICN_{Conn} for PAM4 signaling by 75% due to the maximum signal transition possible, although ICN_{Conn} is not scaled here. IEEE and OIF specifications differences for ICN channel evaluation include:

1. Max integration frequency: percentage of Nyquist frequency

2. Rise time used in calculations
3. Voltage swing definition: IEEE differential vs OIF single-ended
4. Channel or test fixture application

Table 3. ICN Input Parameters

Description	**Parameter**	**Value**
Start integration	F_{min}	50 MHz
End integration	F_{rx}	0.75 * 58 GHz
Transmit magnitude	A_v	0.6 V
Far-end aggressor magnitude	A_{fe}	0.41 V
Rise Time	T_r	6.16 ps

Where:
UI – Unit interval (s)
f – Nyquist frequency (Hz)
Tr – Transmitted (20-80)% Risetime (s)
F_{RX} – 3 dB receiver bandwidth (Hz)
A_x – Amplitude of the crosstalk aggressors (V)
Δf – Frequency step (Hz)

The simulation inputs for ERL calculation are shown in Table 2. In the application of ERL for connectors, either simulated or de-embedded measurements, requires that ERL fixture removal features are disabled.

Table 2 ERL Input Parameters

Description	Parameter	Value
Diff reference impedance		100 Ohms
Fixture removal		0
Loss compensation	B_x	$1.7e^9$
Re-reflection	ρ_x	0.18
DFE compensation	N_b	12 UI
Rise Time	T_r	18.3 ps

To evaluate the impact of each connector at the system level, full link COM simulations needed to be completed at 112 GB/s PAM4 and 56 GB/s NRZ. However, at the time of this paper no specifications for the proceeding data rates has been completed, and therefore the simulation speculates scaled transceiver configurations derived from 28 GB/s NRZ and 56 GB/s PAM4 specifications.

Previous work has demonstrated that PAM4 signaling experiences at 9dB penalty over NRZ signaling. System budgets must absorb this penalty into the silicon or channel allowances by improving jitter and equalization or reducing channel loss. For the purposes of this study, the penalty was assigned to silicon and the same channel budget was maintained. Equal channel losses for both data rates created an environment for assessing the change in reflection and crosstalk tolerance by the date rate change alone.

The COM configuration for 112G-PAM4 is a speculative solution for LR channel type. Notably, the configuration applies 12-taps of DFE and three (3) TXLE precursors. The configuration for 56G-NRZ has higher noise levels on TX_SNR, eta0, and DER as well as reduced TXLE and CTLE equalization magnitude. Both configurations have the same ERL calculation parameters and number of DFE taps and therefore compute the same ERL regardless of signal level count. Table 3 of COM settings used is shown below.

Table 3. COM Settings - 56G-NRZ and 112G-PAM4

Description	**Parameter**	**112G-PAM4 Values**	**56G-NRZ Values**	**Units**
Bandwidth Frequency	f_b	58	58	GBs
Start frequency	f_min	0.05	0.05	GHz
Frequency step	Delta_f	0.01	0.01	GHz
Package die load	C_d	[1.3e-4 1.3e-4]	[2.0e-4 2.0e-4]	nF
Test packages	z_p select	[2]	[2]	
TX package length	z_p (TX)	[12 30]	[12 30]	mm
NEXT package length	z_p (NEXT)	[12 30]	[12 30]	mm
NEXT package length	z_p (FEXT)	[12 30]	[12 30]	mm
RX package length	z_p (RX)	[12 30]	[12 30]	mm
Package pin load	C_p	[1.1e-4 1.1e-4]	[1.1e-4 1.1e-4]	nF
Reference	R_0	50	50	Ohms
Termination	R_d	[50 50]	[50 50]	Ohms
Receiver bandwidth	f_r	0.75	0.75	*f_b
Minimum cursor	c(0)	0.6	0.6	
1st Pre-cursor [min:step:max]	c(-1)	[-0.28:0.025:0]	[-0.15:0.02:0]	
2nd Pre-cursor [min:step:max]	c(-2)	[0:0.05:0.1]	[0:0.05:0.1]	
3rd Pre-cursor [min:step:max]	c(-3)	[-0.1:0.025:0]	[-0.1:0.025:0]	
4th Pre-cursor [min:step:max]	c(-4)	0	0	
1st Post-cursor [min:step:max]	c(1)	[-0.05:.025:0]	[-0.05:.025:0]	
CTLE DC gain [min:step:max]	g_DC	[-20:1:10]	[-12:1:10]	dB
CTLE zero frequency	f_z	23.2	23.2	GHz
CTLE 1st pole frequency	f_p1	23.2	23.2	GHz
CTLE 2nd pole frequency	f_p2	58	58	GHz
Transmitter swing	A_v	0.41	0.41	V
FEXT aggressor swing	A_fe	0.41	0.41	V
NEXT aggressor swing	A_ne	0.6	0.6	V
Signaling Levels	L	4	2	
Samples per UI	M	32	32	

DFE length	N_b	12	12	
	N_b_step	0.0115	0.0115	
DFE magnitude limit 1^{st} tap	b_max(1)	0.7	0.7	
DFE magnitude limit 2^{nd} tap+	b_max(2..N_b)	0.2	0.2	
Random jitter RMS	sigma_RJ	0.01	0.01	UI
Dual-Dirac Jitter meak-peak	A_DD	0.02	0.05	UI
Single sided noise	eta_0	8.20E-09	9.02E-08	V^2/GHz
Transmitter SNR	SNR_TX	32.5	20	dB
Leve separation mismatch ratio	R_LM	0.95	1	
Detector Error Ratio	DER_0	1.00E-04	1.00E-05	

5.4 Establishing Traces Lengths on Either Side of Connector

For the given loss of 28 dB at 28 GHz, the chosen material properties resulted in a physical trace length of 11.5”. Full link simulations were first performed to understand the sensitivity of the connector position within the channel and the implication of the position selection for any subsequent simulations. Six (6) connectors with varied ERL magnitudes from benign to severe were selected and the trace length on either side of the connector (PCB1 and PCB2) were swept in 0.025 fractions of the total 11.5” length, or 0.28” increments.

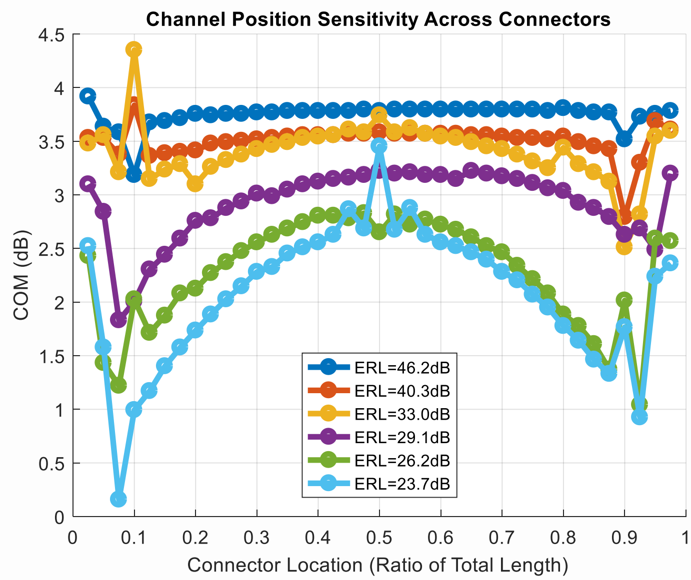

Figure 14: Connector Position in Channel (as a ratio of total length) vs. COM

Figure 14, shows the plot of connector position as a ratio of total length vs. COM. The variety of ERL is distinguished through the different colors. The trend lines look like a frown with the optimal performance located in the middle of the channel and worst performance near the transmitter or the receiver.

Although the ends of the channel are always worse, the impact of connector position near the end is highly variable. Connectors with an ERL >= 33dB may observe a relative penalty of 0.25dB COM applied to the end of a channel compared to center, while worse ERL performances may have a relative penalty in excess of 1.0dB COM. In order words, worse ERL performances have an increasing potential for degraded performance which may only be realized in an adverse channel position.

Near the ends of the channel, another observation can be made about the increased performance and the receiver equalization reach. Connectors located very close to the transmitter or receiver provide re-reflections that occur closely in time and can be equalized by the receiver DFE. Consistently, the two points at each end of the channel are improved and correspond to a length of 0.57" or less. This reach is subject to the number of DFE taps available; 12-taps are exercised in these simulations. Notable variation is also observed at the 3^{rd} and 4^{th} position in the channel and the investigation of this anomaly is beyond the scope of this paper.

All further simulations in this paper will place the connector at an 80% channel position of the total length to ensure an approximate worst-case impact of any connector evaluated. In choosing this worst-case position, we absolve the designer from questioning whether the connector position within the channel could invalidate the methodology proposed in this paper and break the system.

5.5 NRZ vs. PAM4 Trace Lengths

As previously discussed, the 112G-PAM4 channel is more sensitive to noise and is simulated with a smaller silicon noise budget compared to 56G-NRZ. For the purposes of comparison, the silicon budget is calibrated to achieve approximately the same COM performance before introducing a connector. By equalizing the COM value of the two reference channels, the relative change in performance can be reviewed as a connector is included and it's ERL increases.

Alternatively, calibration to the same COM performance could have been made with longer 56G-NRZ channels and proportionally smaller silicon budgets. However, equal loss channel budgets were chosen for this paper to create equivalent ICN_{conn} and ERL_{conn} channel attenuation, regardless of signaling level.

6. Results and Discussion

6.1 ERL vs. COM

The Twenty-five connectors with various ERL_{conn} levels were evaluated in full-link 112G-PAM4 and 56G-NRZ channels. The results of COM_{chan} vs ERL_{conn} is shown in Figure 15. At either data rate of 56G-NRZ and 112G-PAM4, a steady COM level near 3.25dB is observed for high ERL levels and hence can be concluded as point of diminishing returns. It is also observed that channel performance begins to degrade significant below 32dB of ERL, although some slight change or dithering of results can be observed above 32dB. Near an ERL of 25dB or below, channel performance is detrimentally impacted with a reduction upwards of ~2 dB of COM from the connector alone.

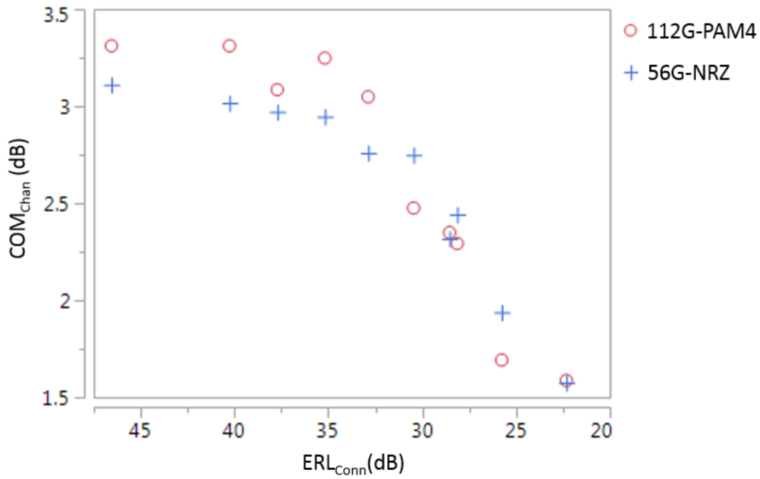

Figure 15: ERL_{conn} vs. COM_{chan}

There were no significant differences found between 112G-PAM4 and 56G-NRZ after a comparison of the rate of performance degradation; that is, the channel penalty of introducing a connector of certain reflection was approximately the same for both 112G-PAM4 and 56G-NRZ.

6.2 ICN vs. COM

A constant connector-thru response is studied with increasing far end crosstalk levels to assess the impact at both 112G-PAM4 and 56G-NRZ data rates. Comparing the signaling methods, PAM4 demonstrates a very high sensitivity to noise and begins to attenuate channel performance above 5mV of ICN_{Conn}. In contrast, 56G-NRZ channel performance is slow to degrade.

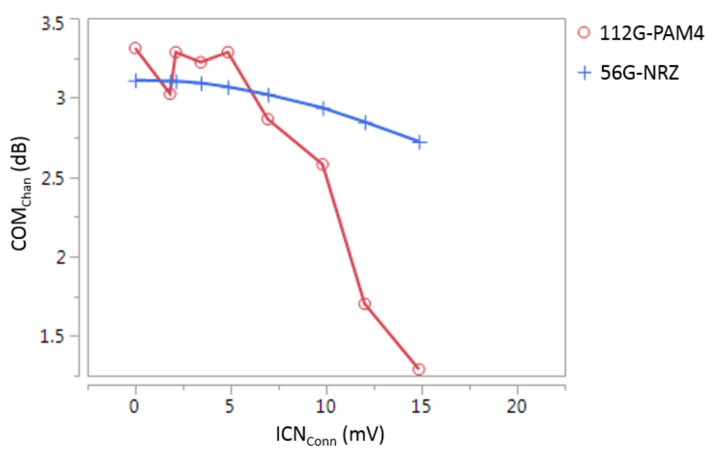

Figure 16: Channel performance against ICN_{Conn}, 112G-PAM4 and 56G-NRZ

6.3 ICN vs. ERL vs. COM

Full link simulations for all the 225 combinations, 25 throughs and 9 crosstalks, were completed. These 225 simulations are charted in Figure 17 for 112G-PAM4 and Figure 18 for 56G-NRZ with the full link COM_{Chan} results.

In both Figure 17 and Figure 18, contour lines reveal positions of equal COM_{Chan} performance yet have significant differences in the balance of crosstalk and reflection. This analysis shows that a new tool for component selection with the dynamic ability to exchange noise sources and meet the same performance goal is a viable option. It is possible, for example, to select smaller and more dense components with high crosstalk for small form factor applications at the exchange of reduced reflection and still meet performance needs.

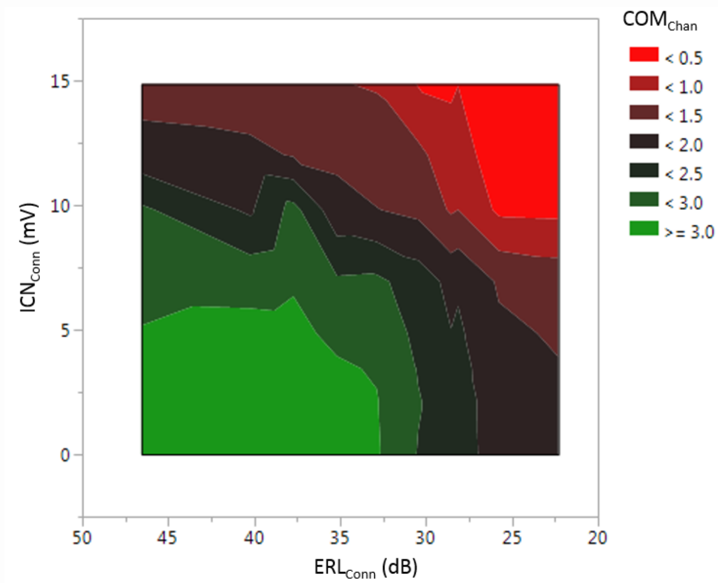

Figure 17: 112G-PAM4 Channel COM response across connector ERL and ICN levels

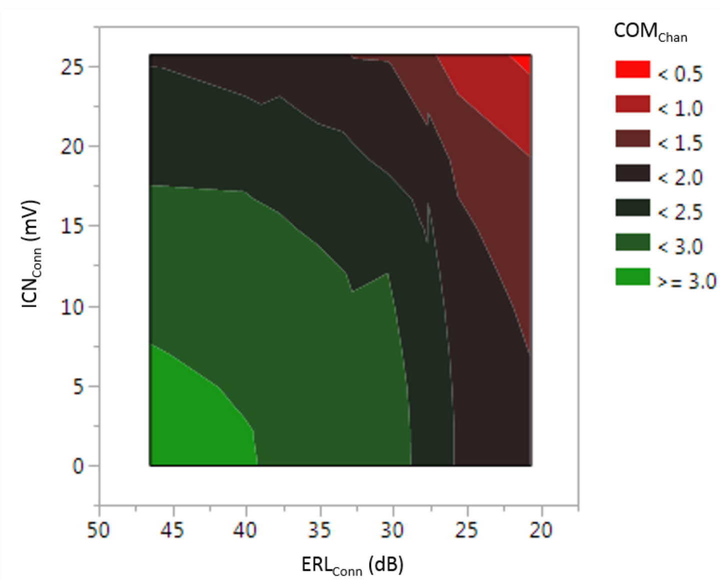

Figure 18: 56G-NRZ Channel COM response across connector ERL and ICN levels

7. Proposed Connector Evaluation/Selection Metric

At the time of writing, the complete channel loss budget and silicon budgets for 112G-PAM4 and 56G-NRZ are not known; it is, therefore, difficult to determine how much budget is available for connectors. To aid system design, the authors assess a connector's viability for 112G-PAM4 and

56G-NRZ through comparison to COM_{chan} for a perfect connector. As the COM_{chan} difference between the device under consideration and a perfect connector increases, capability of the connector to operate at high speed decreases.

The difference in COM_{Chan} between the perfect connector and the evaluated connector it's COM_{Chan} impact. Due to lack of actual budgets, the authors have chosen the following three (3) connector COM_{Chan} impact levels: 0.5dB for easy design implementation, 0.75dB for moderate, and 1.0dB for difficult implementation. An increased difficulty might mean reduced channel levels. The contours for each of the three (3) levels, easy, moderate and hard, at 112G-PAM4 are shown in Figure 19. These are the same contours in Figure 17, redrawn to show the proposed metric. From these three (3) contours limits, two-dimensional limits that trace the sensitivity of a connector's reflection and crosstalk in the full link is derived.

Simplified curves with monotonic slopes were derived from the 112G-PAM4 and 56G-NRZ contours (Figure 17 and Figure 18) and are shown in Figure 20. At the easy implementation limit, 112G-PAM4 ERL_{Conn} may be up to 32 dB for ICN levels of 5mV or below. Higher ICN levels are permissible for ERL_{Conn} levels better than 30dB. At 56G-NRZ, ERL_{Conn} is reduced to 30dB and ICN raised to 7mV.

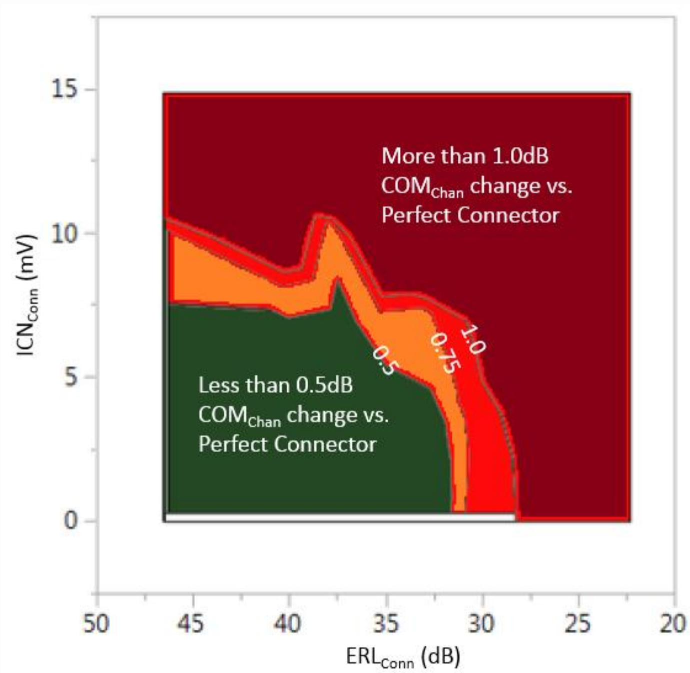

Figure 19: 112G-PAM4 COM_{Chan} contours: 0.5, 0.75, and 1.0 dB less than perfect connector

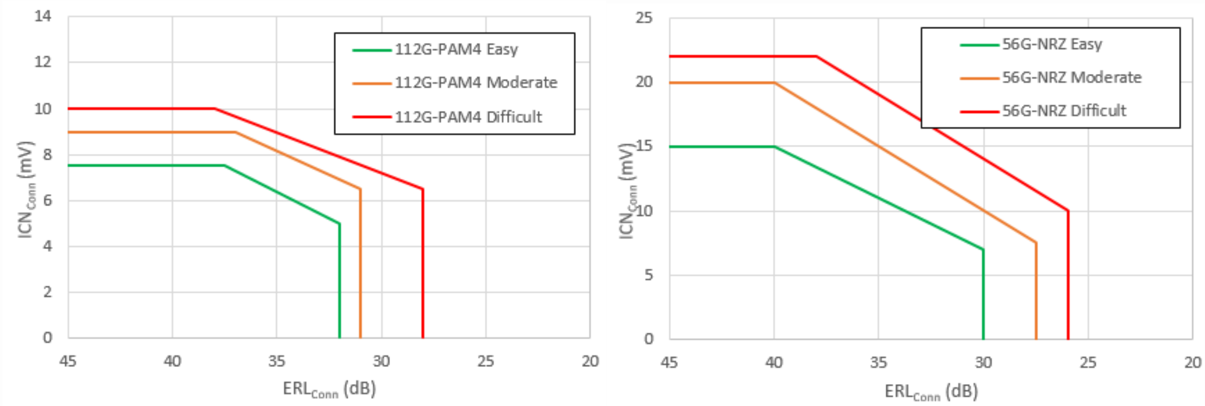

Figure 20: Proposed 112G-PAM4 and 56G-NRZ Connector Limits

There are a couple of caveats to the proposed limits: Any change in loss budget can modulate the appearance of a passing solution: in a shorter channel, a poor interconnect can operate with a passing COM. Uncertainties in future specifications may change the outcome; additional equalization may lessen the connector noise impacts and differences in channel budget, and aggressor magnitude limits will scale the crosstalk sensitivity. The inclusion of near-end crosstalk is a function of application pinout and system location, and it is not analyzed here. Of lesser importance than the methodology used in the derivation of the metrics is the presented values of the metric.

7.2 NRZ vs PAM4

Comparing between signaling levels, the NRZ design permits a 7mV ICN_{Conn} increase between the easy and difficult limits, while the PAM4 increment is only 2mV. This difference continues to provide examples of the reduced noise tolerance of PAM4. The change in reflection allowance between easy and difficult limits is minor (4dB for 56G-NRZ and 5dB for 112G-PAM4) and consistent with earlier results, demonstrating minimal slope differences in COM_{Chan} with ERL_{Conn} change.

In practice, form factors for 56-NRZ may be developed with larger channel loss budgets than PAM4. In such a case, sensitivity to crosstalk and reflection – and therefore connector selection limit recommendations – may relax.

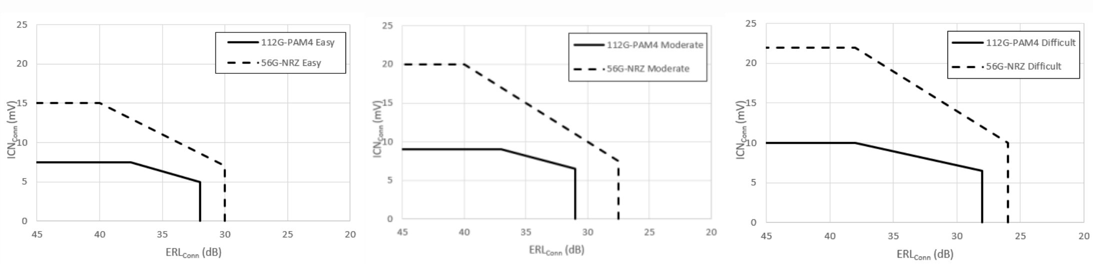

Figure 21: Date Rate Comparison of Proposed Limits: (left) Easy (middle) Moderate (right) Difficult

7.3 Realistic Connector Examples:

Figure 22 shows the NOVARAY connector system from Samtec. This scalable mezzanine and cable series uses a propriety pin to ground configuration to enable low crosstalk and tight impedance control, resulting in an industry leading aggregate data rate average of 1.33 Tbps/sq inch.

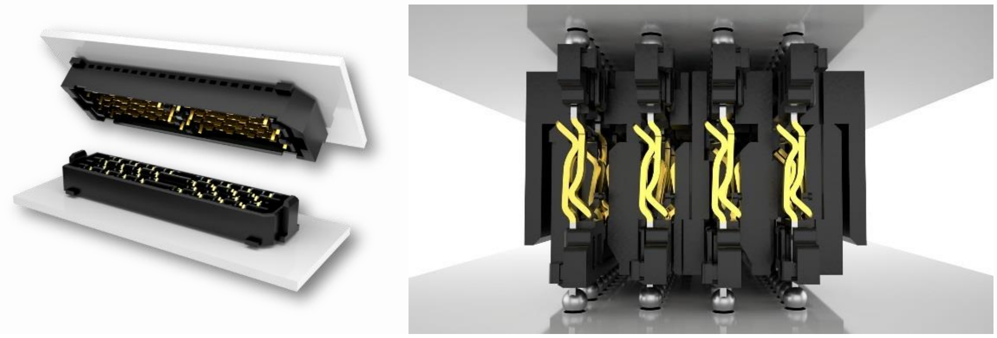

Figure 22: NVAx Connector from Samtec

Shown in Figure 23 is the comparison of performance of NovaRay, a Samtec high-speed mezzanine product that is expected to be capable of 112G-PAM4 to the recommended limits. Unlike the example channels evaluated, the NovaRay channel also includes two (2) realistic PCB breakout regions, one on each end. The connector break-out region includes 40 mil deep plated through holes and eight (8) surrounding crosstalk aggressors. This 7mm mezzanine product has an ERL_{conn} of 34dB and a FEXT ICN_{Conn} of 4mV.

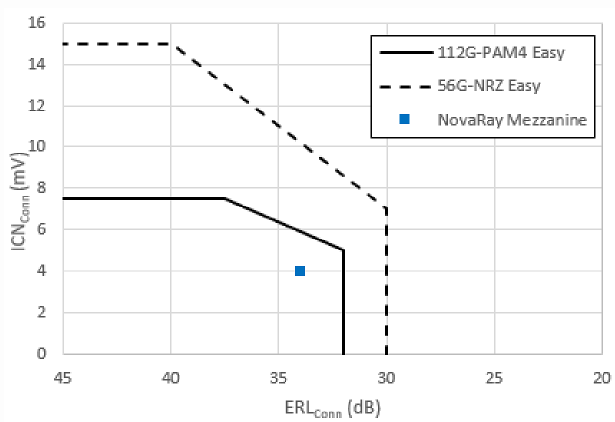

Figure 23: Samtec NovaRAY passing easy 112G-PAM4 and 56G-NRZ limits

Figure 24 shows the APX6 connector, a High Speed Micro Array interconnect from Samtec. This open pin field product is low profile and uniquely designed for low crosstalk noise.

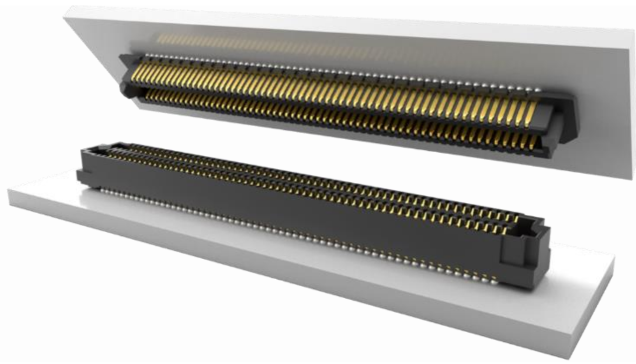

Figure 24 Samtec High Speed Micro Array

An analysis of APX6 performance and the results compared to the 56G-NRZ recommended moderate limit is shown in Figure 25. The analysis includes a PCB breakout region on each end, and has an ICN_{conn} of 1.2 mV when including eight (8) crosstalk aggressors, and an ERL_{conn} of 29.1 dB.

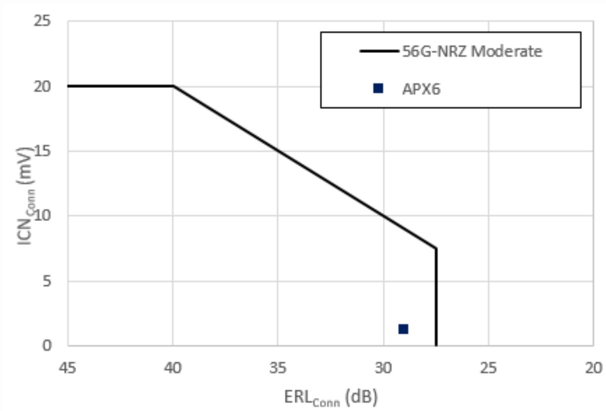

Figure 25 APX6 Passing 56G-NRZ Moderate Limits

8. Conclusions

The metrics presented in this paper can be used as a component design guide for specifying ERL_{conn} and ICN_{conn} thresholds that would constrain channel COM impacts to a modest level. These quantitative results can also assist in design decisions where subtle frequency domain differences would be subject to judgement as potential channel impacts are known without the burden of full link model development and simulation.

System designers can utilize the metrics during the component selections process.The metrics shown in this paper will enable ranking of components by their potential channel impairments. While one component may offer the best reflection loss and another the best crosstalk, the net trade-offs at the channel level will be realized when evaluated by the proposed limits in this paper.

This paper provides an improved and practical understanding of the potential noise sensitivity changes that occur during the transition from NRZ to PAM4 signaling. A significant increase in FEXT crosstalk sensitivity for PAM4 signaling is observed through the reference COM simulations, while the reflection sensitivity change from NRZ to PAM4 is less pronounced.

9. Future Work

This paper presented a basic framework for channel (COM) correlated component metrics. Areas of further study include the study of shorter channels, including NEXT in the evaluation, consideration of components with varying levels of IL, and assessment of the impact of trace impedance variations. These metrics should be re-visited, and the limits set accordingly, when realistic silicon budgets become available. Other data rates can also be considered.

10. References

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