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Practical Channel Modeling for High-speed Design

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Abstract

In order to ensure first time success in the GB/s regime, accurate modeling of PCB interconnect is a precursor to successful high-speed serial link designs. This is especially true for long backplane channels, or to meet industry standard chip to module specifications. Although many EDA tools include the latest and greatest models for conductor surface roughness and wideband dielectric properties, obtaining the right parameters to feed the models is always a challenge. So how do we get these parameters? Often the only sources are from data sheets alone. In most cases the numbers do not translate directly into parameters needed for the EDA tools. By using copper foil roughness and dielectric material properties, obtained solely from manufacturers' data sheets, a practical method of modeling high-speed channels is presented.

Author(s) Biography

Lambert (Bert) Simonovich graduated in 1976 from Mohawk College of Applied Arts and Technology, Hamilton, Ontario Canada, as an Electronic Engineering Technologist. Over a 32-year career, working at Bell Northern Research/Nortel, in Ottawa, Canada, he helped pioneer several advanced technology solutions into products. He has held a variety of engineering, research and development positions, eventually specializing in high-speed signal integrity and backplane design. After leaving Nortel in 2009, he founded Lamsim Enterprises Inc., where he continues to provide innovative signal integrity and backplane solutions as a consultant. He has also authored and coauthored several publications which are posted on his web site at www.lamsimenterprises.com. His current research interests include: high-speed signal/power integrity, modeling and characterization of high-speed serial link architectures.

Introduction

At high frequencies, the conductor and dielectric losses lead to dispersion of the transmitted signal. The total loss of the transmission path is the sum of dielectric and conductor losses. Predicting total loss using smooth copper and published loss tangent values is no longer adequate in the 10-plus GB/s regime.

Failure to account for conductor roughness can be problematic especially for 25 GB/s and above. As shown in Figure 1, with just 3.2 dB delta in insertion loss, at 12.5 GHz Nyquist frequency, results in an average reduction of 50 % in eye height when conductor roughness is taken into account. On top of that, failure to correct dielectric constant (D_k) from manufacturers' data sheets due to conductor roughness can lead to inaccuracy in phase delay [1].

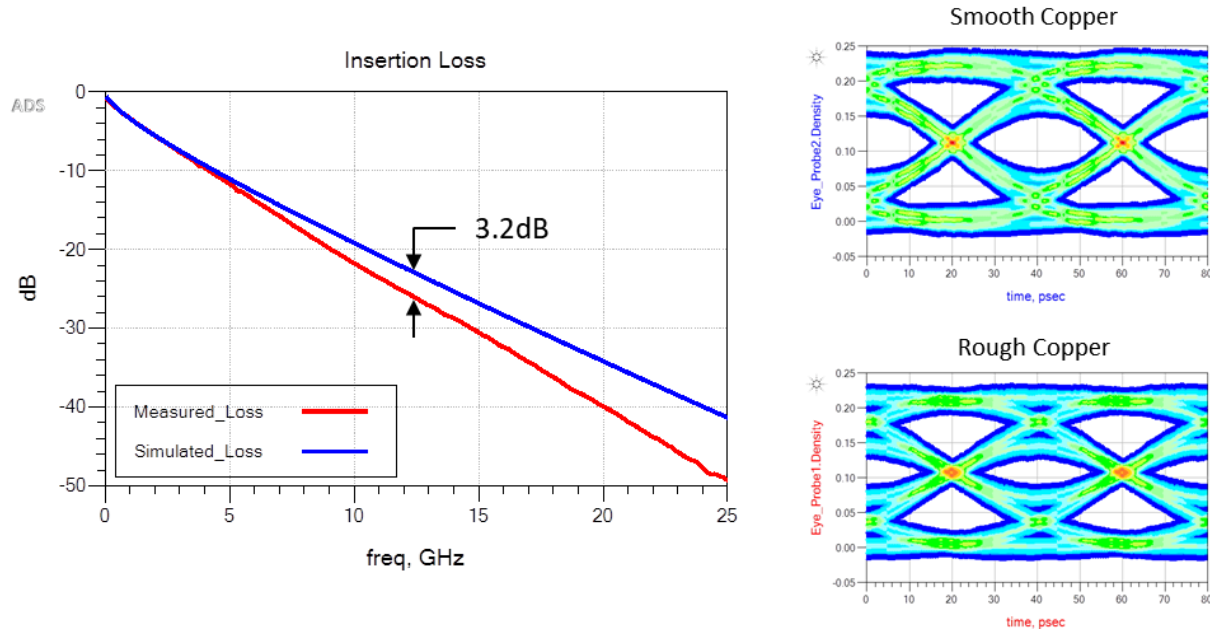


Figure 1 Simulated results of a differential transmission line with and without conductor roughness taken into account. With just 3.2 dB delta in insertion loss, as shown on the left, results in an average reduction of 50% in eye height.

In order to ensure first time success at these speeds, using the right parameters for dielectric and conductor roughness to feed into modern EDA tools is a prerequisite. This is especially true for long backplane channels.

Many electronic design automation (EDA) tools include the latest and greatest models for conductor surface roughness and wide-band dielectric properties. But obtaining the right parameters to feed the models is always a challenge. So how do we get these parameters?

One way is to follow the design feedback method which involves designing, building and measuring a test coupon. After modeling and tuning various parameters to best fit measured data, D_k , D_f and roughness parameters can be extracted. They are then used in channel modeling software to design the final product.

The benefits of this method is that it is pretty practical and accurate; as long as you use the exact same material, glass style and copper foil in your final board stack-up. On the other hand, a significant amount of expertise and equipment is required to design, build and measure the test coupon, which takes significant amount of time and money.

But, Eric Bogatin often likes to say, “*Sometimes an OK answer NOW! is better than a good answer late.*” As a high-speed signal integrity practitioner and backplane specialist, an answer sooner, rather than later is crucial because of the impact to time and cost to clients.

Often the only sources are from manufacturers’ data sheets. In most cases, the numbers do not translate directly into parameters needed for the EDA tools. For example, many conductor roughness models require input parameters not directly published in data sheets.

In this paper you will learn:

- How to take the copper foil roughness parameters from data sheets to correct the manufacturer’s published dielectric properties
- Use the Cannonball-Huray model to model conductor loss
- How to apply these parameters in the latest version of Polar Si9000e Field Solver
- How to pull it all together using Keysight ADS software

Overview

Alternating current (AC) causes conductor loss to increase in proportion to the square root of frequency. This is due to the redistribution of current towards the outer edges caused by skin-effect. The resulting skin-depth (δ) is the effective thickness where the current flows around the perimeter and is a function of frequency.

Skin-depth at a particular frequency is determined by:

Equation 1

$$\delta = \sqrt{\frac{1}{\pi f \mu_0 \sigma}}$$

Where:

δ = skin-depth in meters;

f = sine-wave frequency in Hz;

μ_0 = permeability of free space = 1.256E-6 Wb/A-m;

σ = conductivity in S/m. For annealed copper $\sigma = 5.80E7$ S/m.

Several modeling methods were developed over the years to determine a roughness correction factor (K_{SR}). When multiplicatively applied to the smooth conductor attenuation (α_{smooth}), the attenuation due to roughness (α_{rough}) can be determined by:

Equation 2

$$\alpha_{rough} = K_{SR} \alpha_{smooth}$$

Hammerstad-Jensen Model

The most popular method, for years, has been the Hammerstad and Jensen (H&J) model, based on work done in 1949 by S. P. Morgan. The H&J roughness correction factor (K_{HJ}), at a particular frequency, is solely based on a mathematical fit to S. P. Morgan's power loss data and is determined by [3]:

Equation 3

$$K_{HJ} = 1 + \frac{2}{\pi} \arctan \left(1.4 \left(\frac{\Delta}{\delta} \right)^2 \right)$$

Where:

K_{HJ} = H&J roughness correction factor;

Δ = RMS tooth height in meters;

δ = skin depth in meters.

The model has correlated well for microstrip geometries up to about 15 GHz, for surface roughness of less than 2 μ m RMS.

Huray Model

In recent years, the Huray model [4] has gained popularity due to the continually increasing data rate's need for better modeling accuracy. The model is based on a non-uniform distribution of spherical shapes resembling "snowballs" and stacked together forming a pyramidal geometry.

By applying electromagnetic wave analysis, the superposition of the sphere losses can be used to determine the total loss of the structure. Since the losses are proportional to the surface area of the roughness profile, an accurate estimation of a roughness correction factor (K_{SRH}) can be analytically solved by:

Equation 4

$$K_{SRH}(f) = \frac{A_{matte}}{A_{flat}} + \frac{3}{2} \sum_{i=1}^j \left(\frac{N_i \times 4\pi a_i^2}{A_{flat}} \right) \div \left(1 + \frac{\delta(f)}{a_i} + \frac{\delta^2(f)}{2a_i^2} \right)$$

Where:

$K_{SRH}(f)$ = roughness correction factor, as a function of frequency, due to surface roughness based on the Huray model;

$\frac{A_{matte}}{A_{flat}}$ = relative area of the matte base compared to a flat surface;

a_i = radius of the copper sphere (snowball) of the i^{th} size, in meters;

$\frac{N_i}{A_{flat}}$ = number of copper spheres of the i^{th} size per unit flat area in sq. meters;

$\delta(f)$ = skin-depth, as a function of frequency, in meters.

Although it has proven to be a pretty accurate model, it requires tuning of parameters for best fit to measured data, which is not a practical solution if all you have is roughness parameters from data sheets.

Cannonball-Huray Model

This leads to my Cannonball-Huray model. Using the Cannonball stack principle, the spheres radius (a_i) and tile area (A_{flat}) parameters for the original Huray model are easily estimated solely from roughness parameters published in manufacturers' data sheets [2].

As illustrated in Figure 2 there are three rows of spheres stacked on a square tile base. Nine spheres are on the first row, four spheres in the middle row, and one sphere on top. The height of the Cannonball stack is equal to the 10-point mean roughness (R_z) as published in foil manufacturer's datasheets.

If we can peer into the stack, and visualize a pyramid lattice structure connecting to all the centers of the spheres, then the total height of the Cannonball stack is equal to the height of two pyramids plus two radii.

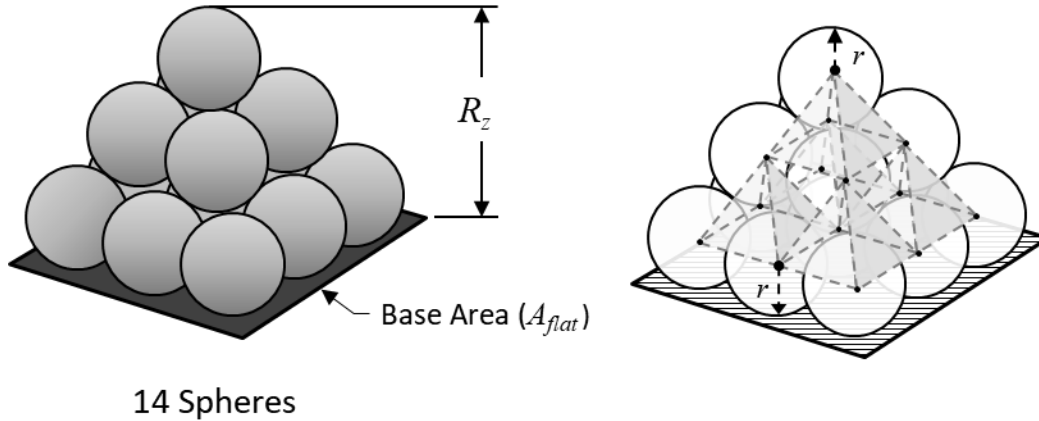


Figure 2 Cannonball-Huray physical model

Through simple geometry and a little bit of algebra, the radius of a single sphere (r) is approximated as [2]:

Equation 5

$$r \approx 0.06R_z$$

And base area (A_{flat}) as:

Equation 6

$$A_{flat} = 36(r)^2$$

Because the model assumes the ratio of $A_{matte}/A_{flat} = 1$, and there are only 14 spheres, the original the Cannonball-Huray model is simplified to:

Equation 7

$$K_{CH}(f) = 1 + 2.33\pi \left(1 + \frac{\delta(f)}{r} + \frac{\delta^2(f)}{2r^2} \right)^{-1}$$

Where:

$K_{CH}(f)$ = Cannonball-Huray roughness correction factor, as a function of frequency

$\delta(f)$ = skin-depth, as a function of frequency in meters

r = the radius of spheres in meters.

Effective Dk Due to Roughness:

Everyone involved in the design and manufacture of printed circuit boards (PCBs) knows one of the most important properties of the dielectric material is the dielectric constant (D_k). Many assume the value reported in manufacturers' data sheets is the intrinsic property of the material. But in actual fact, it is the effective dielectric constant (D_{keff}) generated by a specific test method. When you compare simulation against measurements, you will often see a discrepancy in D_{keff} , due to increased phase delay caused by surface roughness.

D_{keff} is highly dependent on the test apparatus and conditions of how it is measured. There are several methods used in the industry. One method, commonly used by many laminate suppliers, is a clamped stripline resonator test method, described by IPC-TM-650 Test Methods Manual [13]. Section 2.5.5.5 Rev C defines test methods to rapidly test dielectric material for permittivity and loss tangent over an X-band frequency range of 8-12.4 GHz in a production environment.

The measurements are made under stripline conditions using a carefully designed resonant element pattern card, made out of the same dielectric material to be tested. The card is sandwiched between two sheets of unclad dielectric material under test. The whole structure is then clamped between two large plates, lined with copper foils that are grounded. They act as reference planes for the stripline.

By measuring a resonant frequency of the cavity, the effective permittivity and loss tangent are determined. The value of this method is to assure consistency of product, when used in fabricated boards. It does not guarantee the values directly correspond to design applications.

This is a key point to keep in mind, and here is why.

Since the resonant element pattern card and material under test are not physically bonded together, as it would be the case in real life, there are small air gaps between the various layers that affect measured results. These air gaps are caused in part by:

- Etching away the copper on material under test, leaving the bare substrate complete with the micro void imprint of the copper roughness.
- The air gap between resonant element pattern card and material under test due to the copper thickness of the etch pattern.

The small air gaps result in a lower D_{keff} than what is measured in real applications using foil with different roughness bonded to the same core laminate, and is the primary reason for phase delay discrepancy between simulation and measurements.

If D_k and R_z roughness parameters from the manufacturers' data sheets are known, then the effective D_k due to roughness (D_{keff_rough}) of the fabricated core laminate can now be easily estimated by [1]:

Equation 8

$$D_{keff_rough} \approx \frac{H_{smooth}}{(H_{smooth} - 2R_z)} \times D_k$$

Where:

H_{smooth} is the thickness of dielectric from data sheet

R_z is 10-point mean roughness from data sheet

D_k is dielectric constant from data sheet

CMP-28 Case Study

There are many EDA tools that implement the Huray model. Polar Instruments Si9000e field solver [5] has included the Cannonball-Huray roughness model directly within the tool, and is the tool of choice for this study because it is a popular tool used by many board shops when designing stackups.

To test the accuracy of the model, measured data from a CMP-28 Channel Modeling Platform, courtesy of [12] was used for model validation. The 6 inch de-embedded S-parameter data was computed from 2 inch and 8 inch single-ended stripline traces.

The PCB was fabricated with Isola [9] FR408HR 3313 dielectric and 1 oz. MLS Grade 3, controlled elongation reverse treated foil (RTF), from Oak-mitsui [8]. The data sheet and PCB design parameters are summarized in Table 2.

Dielectric constant, D_k dissipation factor, D_f , and copper roughness, R_z are the values as reported in the respective manufactures' data sheets.

An oxide or oxide alternative (OA) treatment is usually applied to the copper surfaces prior to final PCB lamination. The etch treatment creates a surface full of micro-voids which follows the underlying rough profile and allows the resin to fill the voids, providing a good anchor.

Typically 50 μ in (1.27 μ m) of copper is removed by OA treatment [11] thereby reducing the roughness to 4.445 μ m.

Table 2 CMP-28 Test Board and Data Sheet Parameters

Parameter	Value
D_k Core/Prepreg @ 10GHz	3.65/3.59
D_f Core/Prepreg @ 10GHz	0.0094/0.0095
R_z Drum side	3.048 μm
R_z Before Micro-etch-Matte side	5.715 μm
R_z After 50 μin (1.27 μm) Micro-etch treatment - Matte side	4.445 μm
Trace Thickness, t	1.25 mils (31.73 μm)
Trace Etch Factor	60 deg taper
Trace Width, w	11 mils (279.20 μm)
Core thickness, H1	12 mils (304.60 μm)
Prepreg thickness, H2	10.6 mils (269.00 μm)
De-embedded trace length	6.00 in (15.24 cm)

From Table 2 and by applying Equation 8, D_{keff} of core and prepreg due to roughness were determined as:

$$D_{keff_core} = \frac{H_{smooth}}{(H_{smooth} - 2R_z)} \times D_{k_core} = \frac{304.6\mu\text{m}}{(304.6\mu\text{m} - 2 \times 3.048\mu\text{m})} \times 3.65 = 3.725 @ 10\text{GHz}$$

$$D_{keff_prepreg} = \frac{H_{smooth}}{(H_{smooth} - 2R_z)} \times D_{k_prepreg} = \frac{269\mu\text{m}}{(269\mu\text{m} - 2 \times 4.445\mu\text{m})} \times 3.59 = 3.713 @ 10\text{GHz}$$

Polar instruments Si9000 [5] was used for transmission line modeling and simulation. The causal model option was used to model dielectric properties over frequency. Effective D_k due to roughness for core and prepreg, calculated above, were substituted instead of data sheet values. Respective R_z of drum side and matte side after micro-etch were used directly in the Cannonball-Huray model.

Figure 3 compares the simulated results vs measurement of a 6 inch, de-embedded stripline trace. The red plots are simulated and green are measured. Insertion loss (IL) is shown on the left and phase is shown on the right. As can be seen, there is excellent correlation.

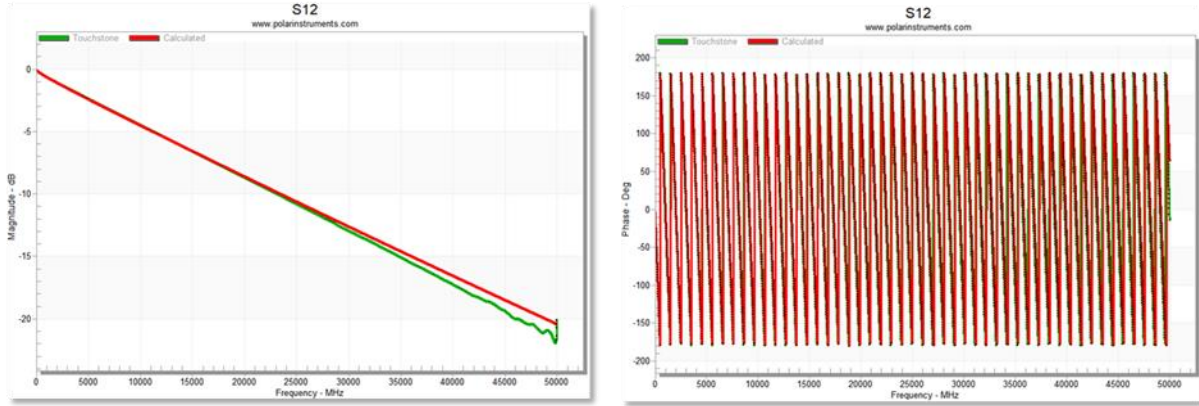


Figure 3 Simulated (red) vs measured (green) insertion loss (left) and phase (right).

D_{keff} can be derived from phase delay, which is also known as time delay (TD), and is often used as a metric for simulation correlation accuracy for phase. TD , as a function of frequency, in seconds, is calculated from the unwrapped measured transmission phase angle, and is given by [14]*:

Equation 9

$$TD(f) = -1 \left[\frac{\text{unwrap}(\text{phase}(S21))}{360 \times \text{freq}} \right]$$

And:

D_{keff} , as a function of frequency, is then given by:

Equation 10

$$D_{keff}(f) = \left(TD(f) \frac{c}{\text{Length}} \right)^2$$

Where:

c = speed of light (m/s);

Length = length of conductor (m).

Results for D_{keff} due to roughness are shown in Figure 4. On the left graph, D_{keff} measured (red) was 3.761 compared to simulated D_{keff} (blue) of 3.626, at 10 GHz, when data sheet values for core and prepreg were used. This gave an error of -3.6%.

* Keysight ADS equation syntax [6]

But when the respective D_{keff_rough} was used for core and prepreg there was better correlation, $D_{keff}=3.727$ vs 3.761, for an error of only -0.9%, as shown on the right graph.

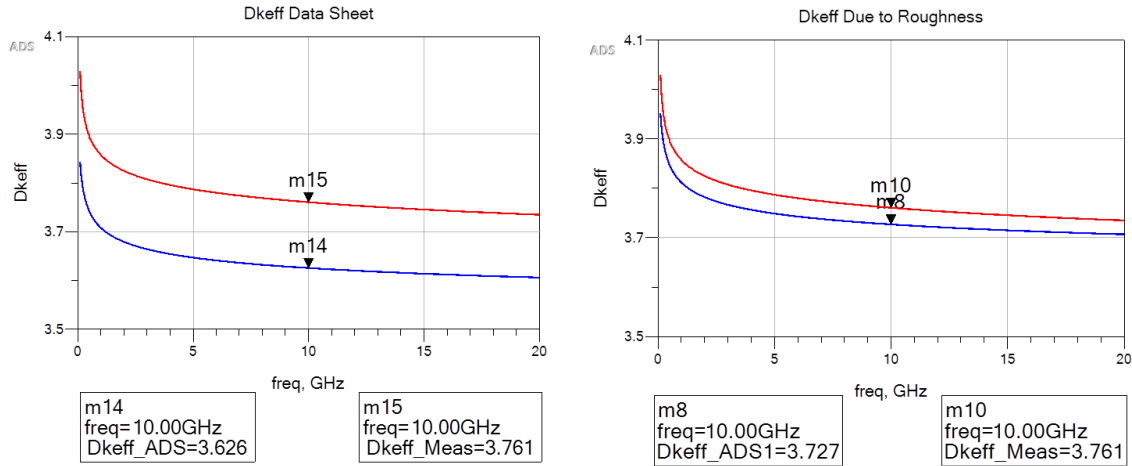


Figure 4 Measured vs simulated D_{keff} using FR408HR data sheet values for core and prepreg (left) and using D_{keff_rough} (right).

Although D_{keff} is closer, to measurements, it is non-causal because the roughness correction factor has only been applied to the real part of the internal impedance of the metal. In our DesignCon 2018 paper [15], we showed that when the roughness correction factor was also applied to the imaginary part of the internal impedance of the metal, it corrects the inductance due to roughness and improves the simulated transmission line characteristics.

Mentor HyperLynx [16] was used to include a causal conductor models with Cannonball-Huray correction factors. Keysight ADS [6] was used for simulation analysis and comparison with the non-causal conductor model. The results are shown in Figure 5.

The non-causal model is consistent with Figure 4, but when the causal version of conductor roughness model was applied it can be seen that simulated D_{keff} matches measured almost exactly. This is remarkable, considering there was no additional tuning or curve of fitting parameters from manufacturers' data sheet values.

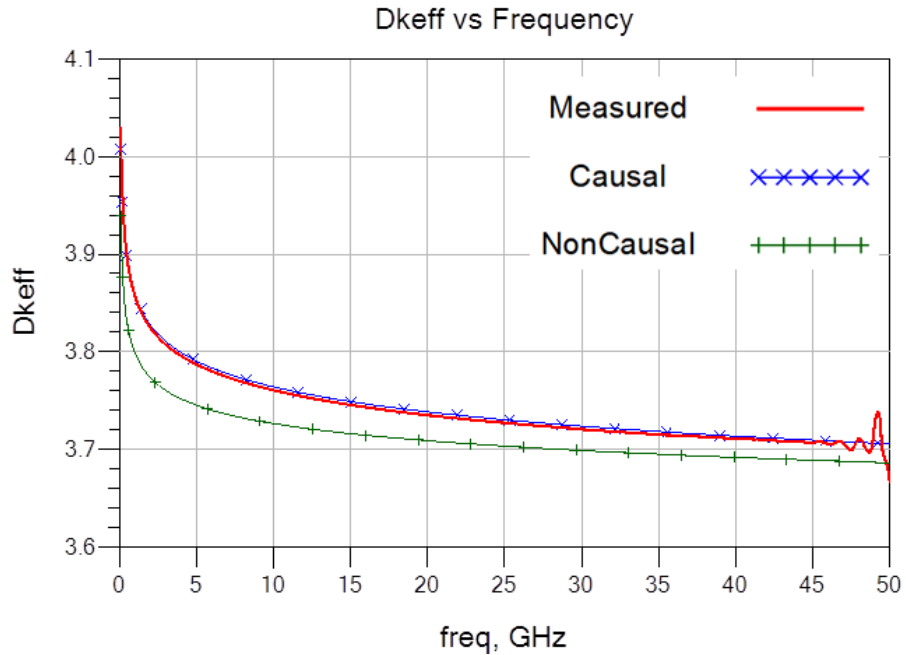


Figure 5 D_{keff} corrected due to roughness and complex roughness correction factor applied.

Figure 6 shows simulated vs measured results for time domain transmission (TDT) impulse response on the left graph. The causal model shows almost an exact fit to measurements. Even though the non-causal impulse response shows slightly faster rise/fall time and a slight difference in delay, the results are still acceptable for analysis.

The time domain reflectometry (TDR) impedance is shown on the right graph. As can be seen, there is excellent correlation between causal models and measured data for both graphs. The causal model has higher characteristic impedance and is a better fit to measured results, but the non-causal model is still acceptable.

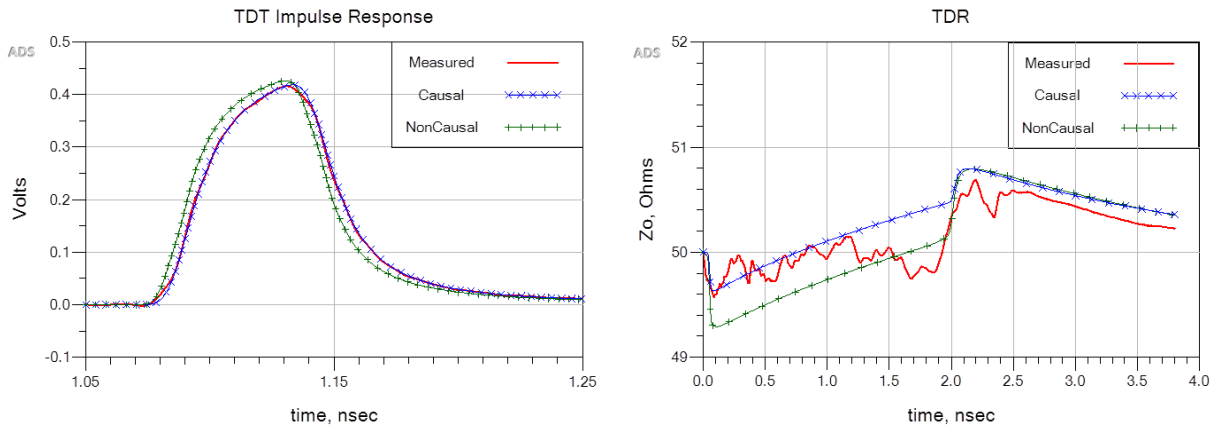


Figure 6 Causal / non-causal vs measured time domain transmission (TDT) impulse response (left) and time domain reflected (TDR) response (right).

Practical Channel Modeling of a High-speed Backplane Case Study

A traditional high-speed serial link backplane channel model has three separate parts. They are two plug-in circuit cards and a backplane. Neglecting vias, the high-speed channel can be quickly modeled as three separate transmission line segments with connectors in-between.

The best way to demonstrate this is through a practical case study example. An Amphenol-FCI Examax demo platform, I helped design in 2013 to showcase the Examax connector performance at 28GB/s NRZ, was used.

A picture of the platform is shown in Figure 7. Among other test structures designed into the backplane, there were four channels with different overall lengths. For simplicity only one channel topology shown is used for comparison in this case study.

The PCBs were fabricated with Nelco N4000-13epsi material [7] clad with MW-G-VSP foil from Oak-mitsui [8]. The respective transmission line design and data sheet parameters are summarized in Table 1.

Each transmission line segment was modeled separately using Polar Si9000e field solver [5]. The respective s-parameters were then saved as touchstone format and later brought into Keysight ADS [6] to model and simulate the entire channel.

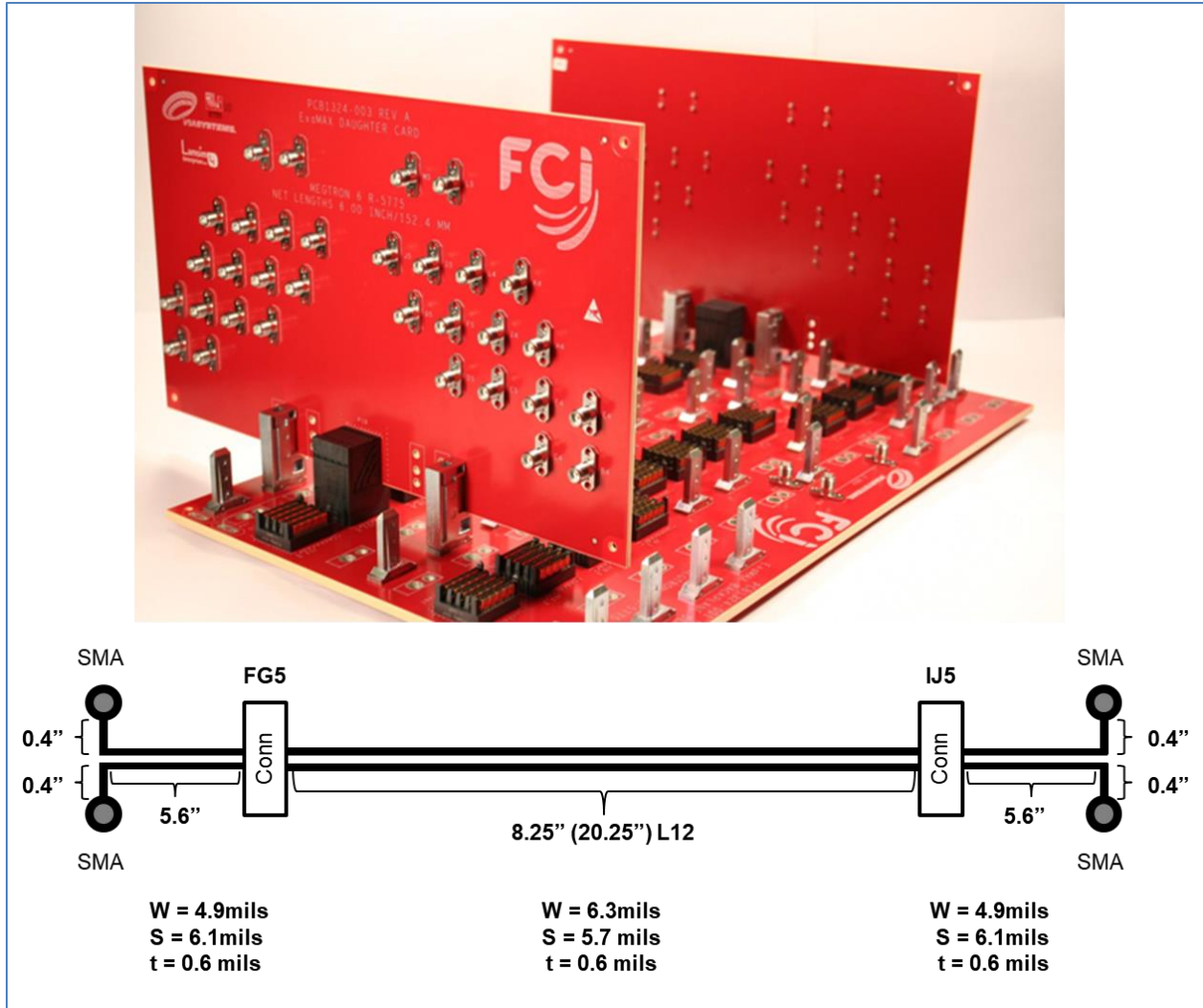


Figure 7 Amphenol-FCI Examax demo platform and channel topology summary used for case study.

Table 1 Design parameters and data sheet summary

Parameter	N4000-13EPSI Backplane	N4000-13EPSI Daughter Card
D_k Core/Prepreg @ 10GHz	3.08/3.06	3.04/3.06
D_f Core/Prepreg @ 10GHz	0.0083/0.0084	0.0085/0.0084
R_z Matte side	2.5 μ m	2.5 μ m
R_z Drum side with OA	1.5 μ m	1.5 μ m

Parameter	N4000-13EPSI Backplane	N4000-13EPSI Daughter Card
Trace Thickness, t	0.6 mils (15.2 μm)	0.6 mils (15.2 μm)
Trace Width, w_1	6.3 mils (160.0 μm)	4.9 mils (124.5 μm) (Diff) 5.4 mils (137.2 μm) (SE)
Trace Width, w_2	5.7 mils (144.8 μm)	4.3 mils (109.2 μm) (Diff) 4.8 mils (121.9 μm) (SE)
Trace Separation, s	5.7 mils (144.8 μm)	6.1 mils (154.9 μm)
Core thickness, H1	6 mils (152.4 μm)	4 mils (101.6 μm)
Prepreg thickness, H2	5.8 mils (147.3 μm)	5.8 mils (147.3 μm)

Method

The first step is to determine the effective D_k due to roughness for the cores and prepregs used on the daughter cards and backplane. By applying Equation 8 to the respective values in Table 1, the D_{keff} results are summarized in Figure 8.

Daughter Card	Backplane
$D_{keff_prepreg} = \frac{H_{smooth}}{(H_{smooth} - 2R_{z_drum})} \times D_{k_prepreg}$ $= \frac{6.2\text{mils} \times 25.4}{(6.2\text{mils} \times 25.4 - 2 \times 1.5\mu\text{m})} \times 3.06$ $= 3.12$	$D_{keff_prepreg} = \frac{H_{smooth}}{(H_{smooth} - 2R_{z_drum})} \times D_{k_prepreg}$ $= \frac{6.2\text{mils} \times 25.4}{(6.2\text{mils} \times 25.4 - 2 \times 1.5\mu\text{m})} \times 3.06$ $= 3.12$
$D_{keff_core} = \frac{H_{smooth}}{(H_{smooth} - 2R_{z_matte})} \times D_{k_core}$ $= \frac{4.0\text{mils} \times 25.4}{(4.0\text{mils} \times 25.4 - 2 \times 2.5\mu\text{m})} \times 3.04$ $= 3.20$	$D_{keff_core} = \frac{H_{smooth}}{(H_{smooth} - 2R_{z_matte})} \times D_{k_core}$ $= \frac{6.0\text{mils} \times 25.4}{(6.0\text{mils} \times 25.4 - 2 \times 2.5\mu\text{m})} \times 3.08$ $= 3.18$

Figure 8 Effective D_k due to roughness summary of cores and prepregs used on the daughter cards and backplane.

The next step is to determine the radius of spheres and base tile area for Huray model, as summarized in Figure 9. Because electro-deposited (ED) foil has a matte side and drum side,

with different roughness parameters, the sphere radius for each side is calculated separately. Since most EDA tools using the Huray model only allow one input for radius, the average effective radius (r_{eff}) is used.

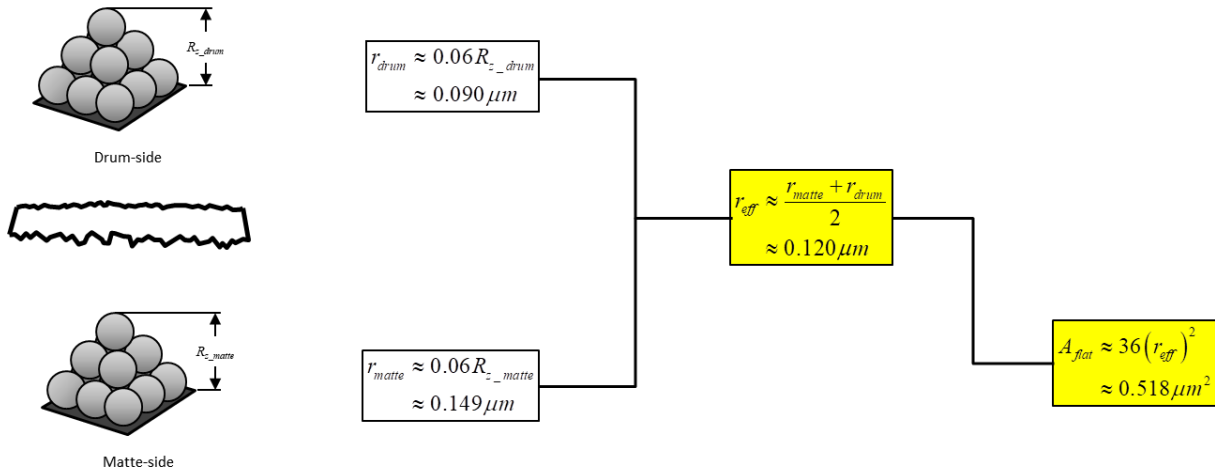


Figure 9 Summary to determine effective sphere radius (r_{eff}) and flat tile area (A_{flat}) parameters for input into Polar Si900s Huray model.

For each length of transmission line of the topology shown in Figure 7, separate touchstone files are created. Using Polar Si9000 [5], the following steps for each file are described below:

1. The first step for modeling each transmission line segment is to select the “Lossless Calculation” tab at the bottom of the Si9000e main input window. When the pop-up window appears, choose the appropriate transmission line geometry from the pallet along the left-hand side, and then enter the specific design parameters in the boxes. Change the various parameters until the desired impedance is reached
2. The next step is to select the “Frequency Dependent Calculation” tab at the bottom of the input panel and enter line length, conductivity and frequencies in the appropriate boxes in the main window.
3. Under the “Extended Substrate Data” section, choose “Causally Extrapolate Er / TanD” radio button and click “Edit” to enter the D_{keff} parameters in a pop-up window as shown. Click “Calculate” to view causal D_{keff} over frequency. Click “Close” to return to main window.
4. Under the “Surface Roughness Compensation” section from the main window, select Huray radio button and click “Edit” to enter the appropriate roughness parameters in the pop-up window. Enter r_{eff} and A_{flat} in the boxes shown. Enter 1.00 for “Ratio of Areas” and 14 for the “Number of Balls in Area” boxes. Click “Apply” to return to main window.
5. In the main window, hit calculate. Once the simulation has run, then export the respective touchstone file under the “File” menu.

After repeating this for each transmission line segment of the topology, there should be 3 touchstone files generated. One file is for the single-ended traces on the daughter card. Another one is for the differential pair on the daughter card and the last one is for the differential pair on the backplane.

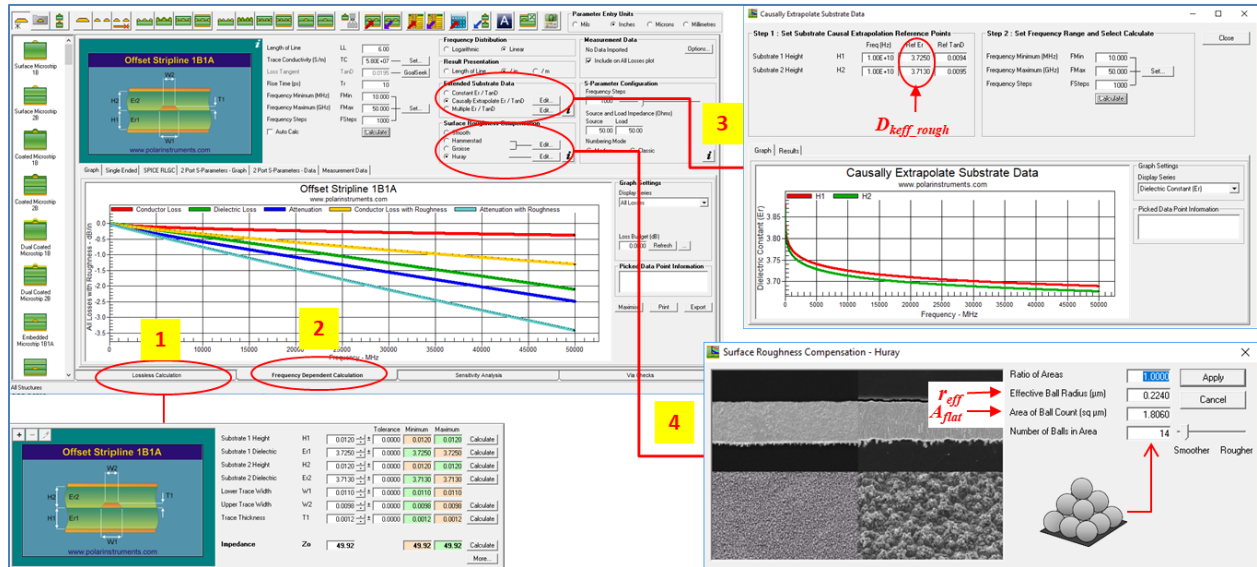


Figure 10 Generic Polar Si900e user interface window and pop-up panels to enter required parameters.

In this case study, Keysight ADS [6] was used to model and simulate the entire backplane channel, as shown in Figure 11 . All the s-parameter files were concatenated together as shown, including the Examax connector s-parameter files. If you want the “OK answer NOW!” you can ignore the vias, since you can always model and add them later to get that “good answer late.” if need be.

The results of the simulation are plotted in Figure 12. As you can see there is excellent correlation for both differential SDD21 insertion loss on the left and differential TDR on the right.

Figure 13 show plots of transmit eyes on top and receive eyes on bottom at 28GB/s for the 20.25” channel. The simulated channels are on the left and the measured channels are on the right. As can be seen, even though the measured transmit eyes show slightly more noise and jitter, the received eyes are virtually the same, suggesting the “OK answer NOW!” is probably good enough to make an engineering decision.

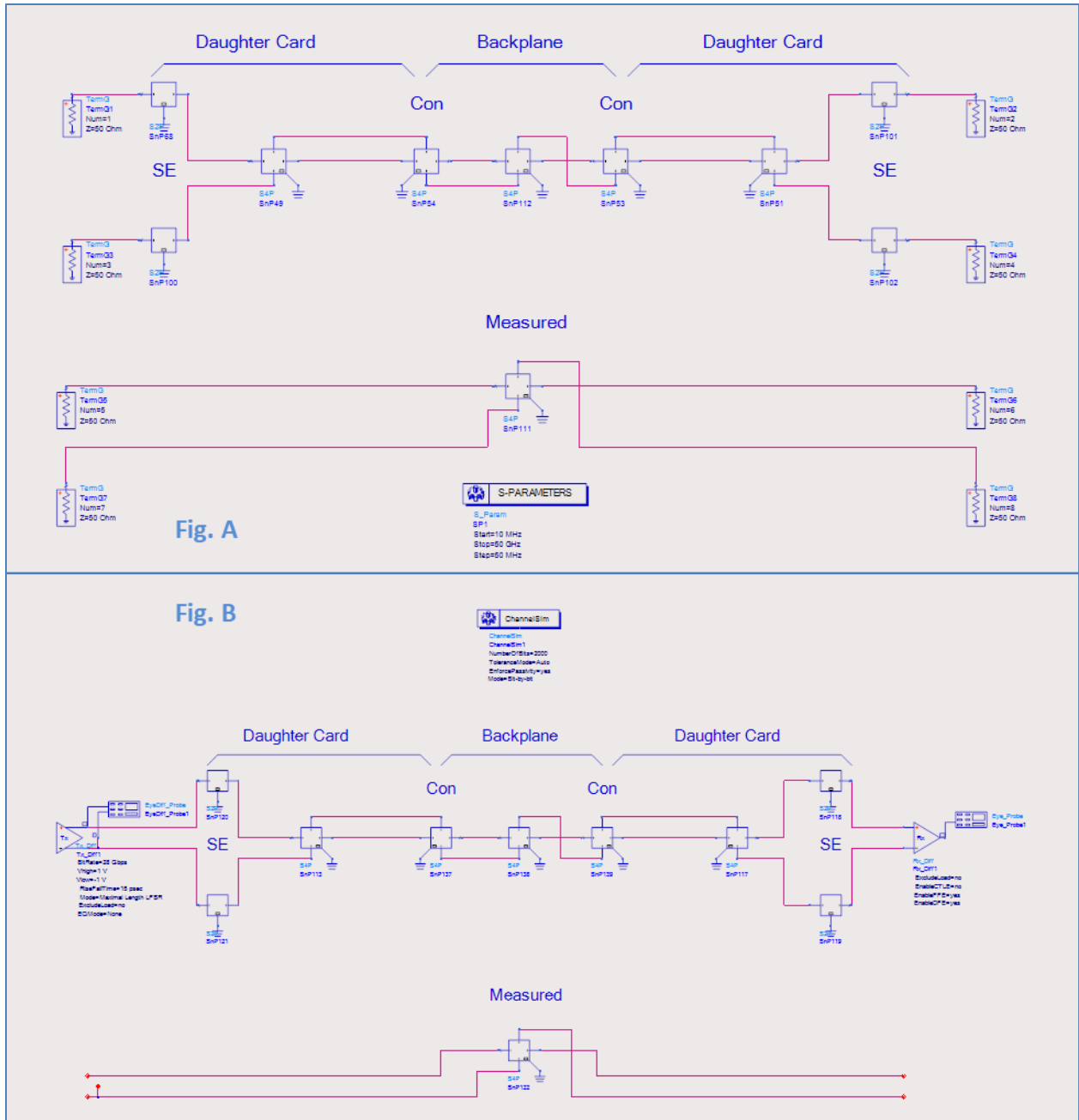


Figure 11 Keysight ADS generic topology models for frequency domain (Fig. A) and transient channel simulation analysis (Fig. B).

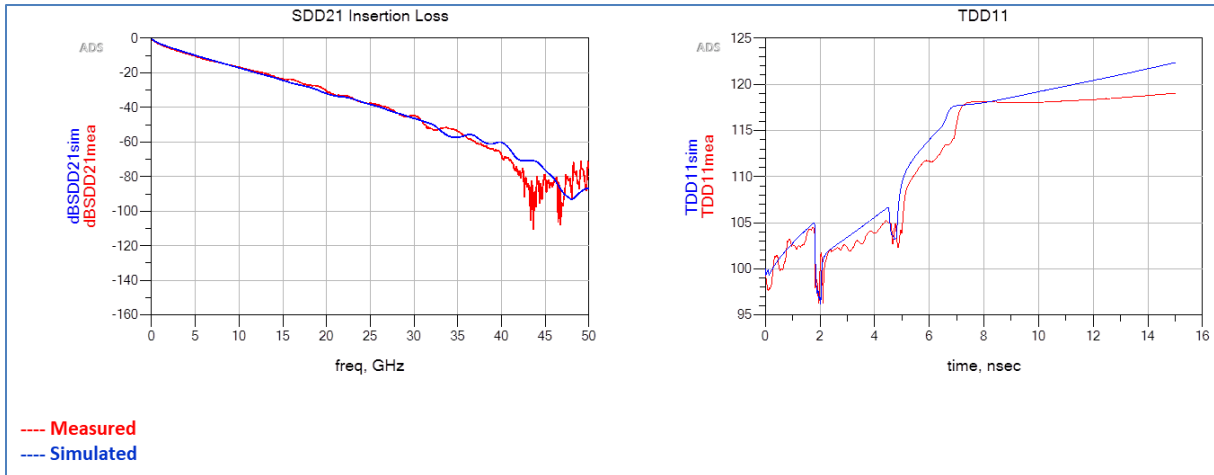


Figure 12 Simulated differential insertion loss (SDD21) shown on left and differential TDR (TDD11) shown on right

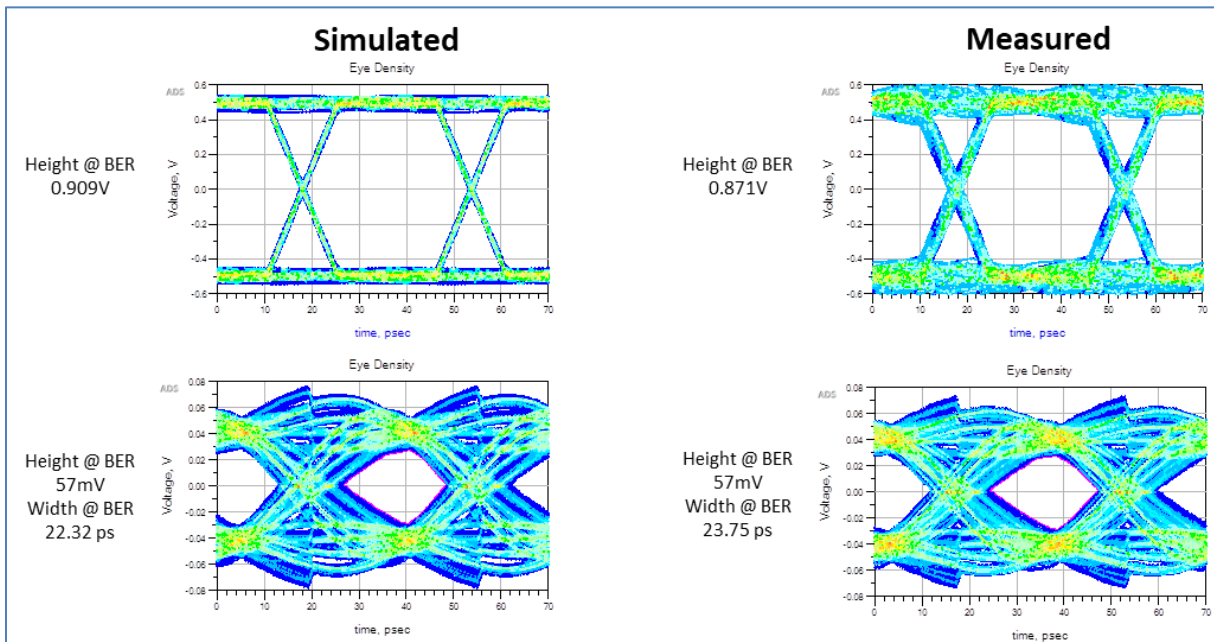


Figure 13 Channel simulation at 28 GB/s. Total length = 20.25”.

Summary

By using Cannonball-Huray model, with copper foil roughness and dielectric material properties obtained solely from manufacturers’ data sheets, practical channel modeling for high-speed design is now achievable using commercial field-solving software employing Huray model.

The non-causal model for conductor loss does not adversely affect simulation results when compared to measurements and shouldn't disqualify EDA tools that have not included a causal metal loss model.

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