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# Which one is better? Comparing Options to Describe Frequency Dependent Losses


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## Abstract

In any channel operating at 2 Gbps and above, conductor and dielectric losses can dominate channel performance. These effects must be included in any accurate system simulation. The problem isn't that simulators don't do this; there are several choices in interconnect loss mathematical expressions and it's difficult to decide how to transform fab information into simulator input.

There are different combinations of parameterized mathematical expressions for dielectric and conductor loss which are in popular use in the industry. Each works to some extent. This paper takes each mathematical expression, explains its origin, evaluates its predicted insertion loss magnitude and phase then explores how the expression scales.

This is useful when translating test coupon results into accurate simulation predictions.


## Author(s) Biography

Dr. Eric Bogatin received his BS in physics from MIT and MS and PhD in physics from the University of Arizona in Tucson. He has held senior engineering and management positions at Bell Labs, Raychem, Sun Microsystems, Ansoft and Interconnect Devices. Eric has written 6 books on signal integrity and interconnect design and over 300 papers. His latest book, Signal and Power Integrity- Simplified, was published in 2009 by Prentice Hall. He is currently a signal integrity evangelist with Bogatin Enterprises, a wholly owned subsidiary of Teledyne LeCroy. He is also an Adjunct Associate Professor in the ECEE department of University of Colorado, Boulder. Many of his papers and columns are posted on the www.beTheSignal.com web site.

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Dr. Paul G. Huray is Professor of Electrical Engineering at the University of South Carolina and has worked at the Oak Ridge National Laboratory, Intel, and the White House. Huray introduced the first graduate program on signal integrity and is the author of ***Maxwell's Equations*** and ***The Foundations of Signal Integrity***.

Dr. Yuriy Shlepnev is President and Founder of Simberian Inc., where he develops Simbeor electromagnetic signal integrity software. He received M.S. degree in radio engineering from Novosibirsk State Technical University in 1983, and the Ph.D. degree in computational electromagnetics from Siberian State University of Telecommunications and Informatics in 1990. He was principal developer of electromagnetic simulator for Eagleware Corporation and leading developer of electromagnetic software for simulation of signal and power distribution networks at Mentor Graphics. The results of his research are published in multiple papers and conference proceedings.

## Introduction

In any channel operating at 2 Gbps and above, conductor and dielectric losses can dominate channel performance. Conductor loss is dominated by the resistive losses from the current redistribution related to skin depth effects and the impact of surface texture on one or all copper surfaces increasing the loss due to absorption of the propagating electromagnetic field.

The dielectric loss is dominated by the material properties described by the dissipation factor and dielectric constant of the laminates, and their relative distribution in the stack up.

Both mechanisms contribute to frequency dependent loss and to dispersion in the speed of the signal. The dispersion can be easily described by an effective dielectric constant.

These mechanisms must be included in any accurate system simulation. The problem isn't that simulators don't do this; there are several choices for interconnect loss mathematical expressions. While it is often possible to get accurate information about the cross section geometry information, it is a challenge to get material properties information in a format that immediately translates into mathematical parameters and results in accurate simulation.

A number of studies [1], [2], [3] have reported success in fitting parameterized mathematical expressions for loss to specific measured test lines. There is no guarantee that measured data in a high volume manufacturing environment will be R&D laboratory quality. With noise added, while a good match may be obtained, there may not be a unique solution.

Rather than take specific measurements and fit parameters of a model, in this study each of the popular mathematical expressions are evaluated to compare the sensitivity of their parameters to the predicted frequency dependence of loss and dispersion.

A few examples are offered for how to fit parameters to measured data.

## Mathematical Expressions for interconnect loss

Any real interconnect will have a causal performance. The most valuable mathematical expressions that are the basis of simulating insertion loss in transmission lines should be causal.

To first order the conductor and dielectric losses are independent. This may not always be a good assumption. There may be some connection between the tooth structure of the copper surface texture and the dielectric material, changing the effective Dk or Df of the laminate. [4], [5] In this study, we assume the two mechanisms are independent.

### Conductor Texture Power Loss Mechanisms:

There are four popular mathematical expressions [6] used to describe conductor power loss:

1. A smooth copper-skin depth based power loss [7].

2. The Hammerstad empirical fit for surface texture which includes dependence on an rms deviation term.

3. A modified Hammerstad empirical fit for surface texture which includes an additional surface area factor and an rms deviation term.

4. The Huray snowball model [8] for surface texture which models the surface as a collection of copper spherical balls electrodeposited on a Matte or Flat base copper surface.  This model is independent of rms deviation.

In each mathematical expression, the macroscopic parameters which define the base conductor cross section and material properties are the same:

Line width, $w$
Conductor thickness, $t$
Bulk conductivity, $\sigma$
They differ in their description of copper surface texture.

The Hammerstad empirical fit [9] is based on a copper surface texture proposed by Samuel Morgan that has a 2 dimensional transverse triangular distortion, shown in Figure 1.  The Morgan model was based on a two-dimensional numerical solution of Maxwell's equations.

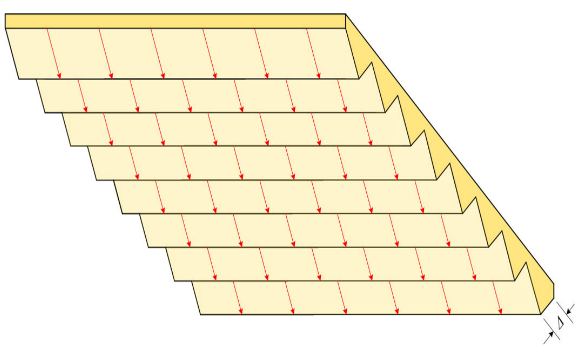

**Figure 1. Morgan's concept of transverse equilateral triangular conductor grooves on a flat base copper surface; Δ is the RMS deviation from flatness.  The red arrows show the direction of surface current flow if the signal electromagnetic field propagates from the upper left (Port 1 input) toward the lower right (Port 2 output).**

Morgan intuitively ***guessed*** that the power loss due to the various surface textures was correlated with the ratio of the RMS deviation to the skin depth at various frequencies so he plotted his rough power loss results (compared to his smooth power loss results) for transverse grooves as a function of the ratio, *Δ/δ*.

Hammerstad did not know how to incorporate the Morgan parallel groove loss results (up to 30% of the transverse groove losses) so he ignored them. He then estimated that a mathematical function was a "good" fit to the Morgan data

$$\frac{P_{Rough}}{P_{Flat}} = 1 + \frac{2}{\pi}\arctan\left[1.4\left(\frac{\Delta}{\delta}\right)^2\right],$$

Where

Δ is the rms deviation from a flat surface
δ is the skin depth of copper as a function of frequency

There was **no** theoretical basis for this mathematical function.

The modified Hammerstad empirical fit [5] adds a scale factor which is basically related to the added surface area from the roughness over a flat surface. This is illustrated in Figure 2.

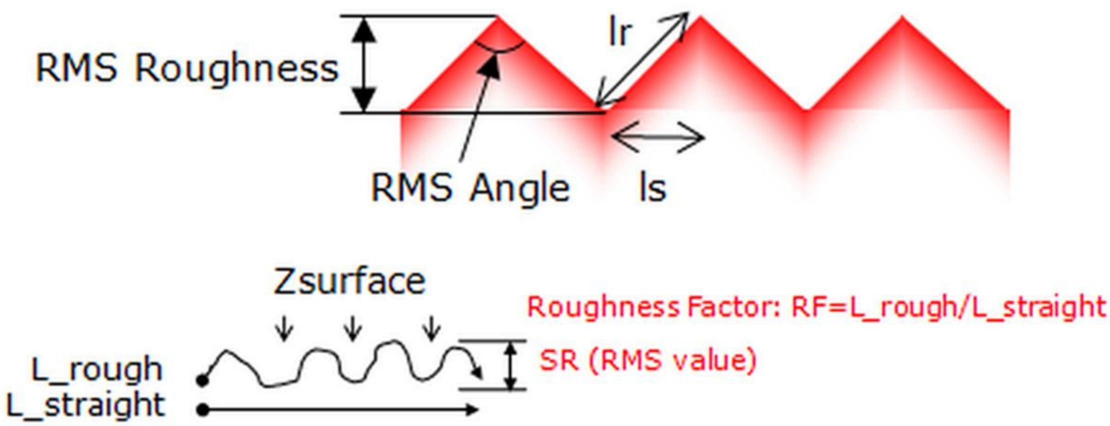


**Figure 2. Illustration of the roughness factor scaling term to account for any angle of tooth and the resulting increased surface area.**

This higher surface area is integrated into the modified Hammerstad approximation as a scaling factor

$$\frac{P_{Rough}}{P_{Flat}} = 1 + \frac{2}{\pi}\arctan\left[1.4\left(\frac{\Delta}{\delta}\right)^2\right](SF-1)$$

When *SF = 2*, this expression reduces to the Hammerstad empirical mathematical equation.

The Huray snowball Model uses a first principles analysis with no fudge factors or scaling factors to describe the copper surface texture in terms of a collection of small copper spheres electrodeposited on a Matte or Flat copper surface. In the case of a Matte surface consisting of a hexagonal lattice of relatively smooth oscillations the area of the surface, $A_{Matte}$, is larger than the area of a Flat hexagonal surface, $A_{Flat}$. In the case of electrodeposition on a Flat surface, the collection of small copper spheres is randomly electrodeposited on a unit area. In both cases, the number of spheres per unit flat area, $N_i/A_{Flat}$, along with the radius of each sphere, $a_i$ (including their area, $4\pi a_i^2$) determine the additional power lost due to the textured copper. This effect, for a hexagonal Matte surface is illustrated in Figure 3.

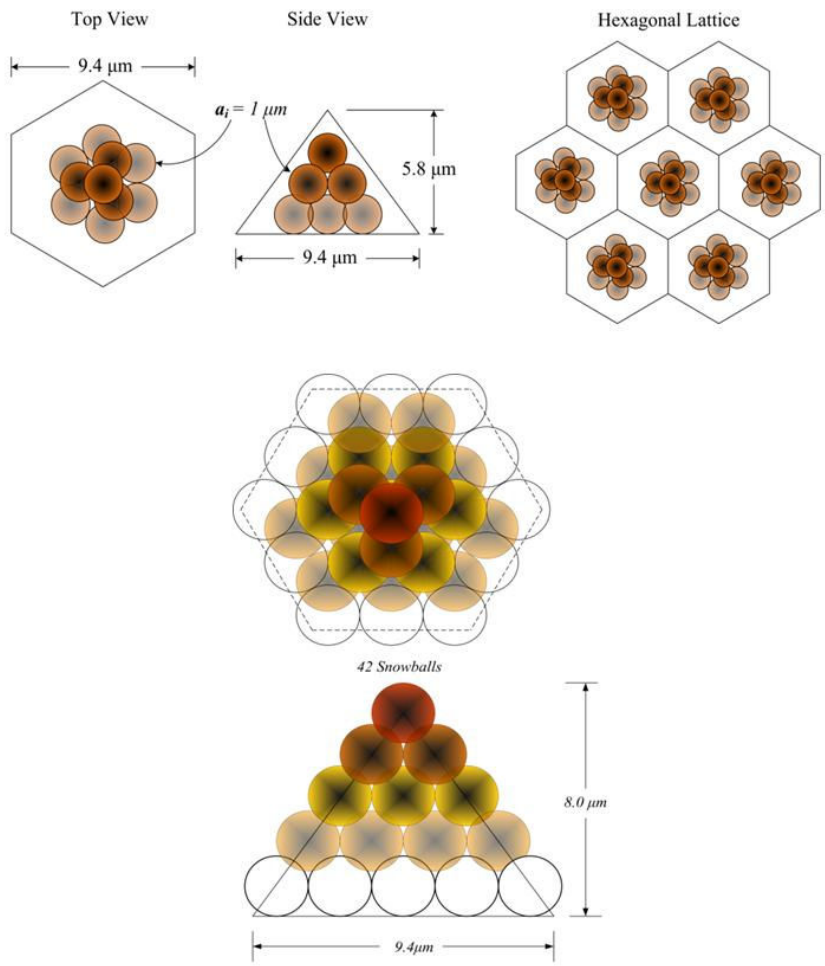


**Figure 3. Illustration of the features of the Huray model for a hexagonal Matte surface of copper upon which copper spheres have been electrodeposited. For the illustration, the spheres have been taken to be of uniform radius, but in general there is a distribution of spherical (snowball) sizes.**

This model is very similar to the actual close up SEM views of Matte copper surfaces after electrodeposition treatment for adhesion promotion, as show in Figure 4.

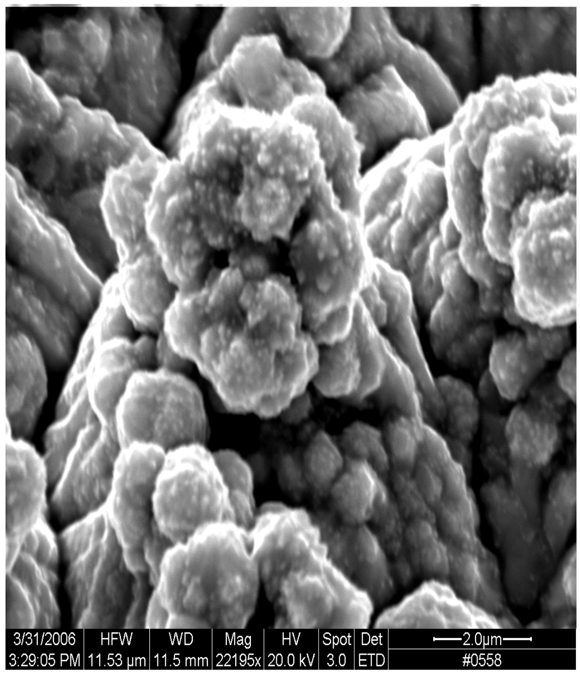


**Figure 4. SEM photograph of a High Profile surface copper Matte foil following additional "anchor nodule" electrodeposition (snowballs) on top of a heat treated base foil (with large micrograins arranged in an approximately hexagonal geographic pattern).**

Form the geometrical distribution of spheres, and the absorption and scattering properties of a single sphere, the impact on this textured surface can be analytically solved. The power loss is given by:

$$\frac{P_{rough}}{P_{Flat}} \approx \frac{A_{Matte}}{A_{Flat}} + \frac{3}{2}\sum_{i=1}^{j} \left(\frac{N_i 4\pi a_i^2}{A_{Flat}}\right) \Bigg/ \left[1 + \frac{\delta}{a_i} + \frac{\delta^2}{2a_i^2}\right]$$

Where the first term relates the relatively larger fractional area of the Matte base surface before electrodeposit of the copper nodules.

$A_{Matte}/A_{Flat}$ is the relative area of the Matte base compared to a flat surface
$a_i$ is the radius of the copper sphere (snowball) of the $i^{th}$ size.
$N_i / A_{Flat}$ is the number of copper spheres of the $i^{th}$ size per unit Flat area.
$\delta$ is the electromagnetic skin depth for copper at a particular frequency.

In this model, a rougher surface is obtained with more snowballs per unit area, whether electrodeposited on a Matte or Flat copper base but does **not** depend upon the deviation from flatness, Δ. The radius of the balls and their associated total area $N_i (4\pi a_i^2)$ per unit flat area, $A_{Flat}$, affects the scaling of the loss with frequency.

## Dielectric Power Loss Mechanisms:

The model used in this study to describe dielectric loss is the wideband Debye model [10]. The simplest and most commonly used approach to implement a wideband, or an infinite pole Debye model, is using the Svensson-Djordjevic approximation [11].

In this model, the real part of the dielectric constant is assumed to take a log dependence on frequency. The imaginary part of the dielectric constant is related to the real part with the Kramers-Kronig relationship. These parameters take the form of:

$$\mathrm{Dk}(f) = \mathrm{Dk}_2 + \frac{\Delta \mathrm{Dk}}{\log(f_2) - \log(f_1)}\left(\log(f_2) - \log(f)\right)$$

And

$$\mathrm{Df}(f) = \frac{\varepsilon''(f)}{\mathrm{Dk}(f)} = \frac{0.682}{\mathrm{Dk}(f)}\,\frac{\Delta \mathrm{Dk}}{\log(f_2) - \log(f_1)} = \frac{-0.682}{\mathrm{Dk}(f)}\,\mathrm{slope}$$

There are five parameters that define the Svensson-Djordjevic approximation to the wideband Debye model:

f1 is the low frequency range for the model
f2 is the high frequency range for the model
f is the frequency at which Dk and Df are defined
Dk at a frequency
Df at a frequency

From these five terms, the functional dependence of Dk(f) and Df(f) can be calculated. This is inherently a causal model. Other causal models, neglected in this paper, are the multipole Debye over-damped model and the multipole Lorentz relaxation models.

## The Fundamental Problem and Solution

### A First Principles Design Flow

In a perfect world, design information such as cross section information, material properties and manufacturing processes, would be input to a simulation tool and from first principles, with no feedback from the fab vendor, and an accurate prediction of the performance of the interconnects would be available for integration into a system level simulation.

Before there was so much concern about loss at frequencies higher than *1 GHz*, the industry had come close to this goal. The characteristic impedance, time delay and cross talk features are straightforward to accurately predict. When the details about the specific glass yarn, resin composition of each laminate layer and the process conditions like etch back are taken into consideration, interconnect performance can be controlled and predicted to better than 5% for frequencies below *1 GHz*.

In the design for a target impedance or cross talk level, an accurate simulation environment allows an engineer to realistically explore design space and make performance, practical manufacturing capabilities and cost tradeoffs.

But, when signal data rates are in excess of 2 Gbps (with inclusion of up to the $5^{th}$ harmonic for signal bandwidths) and interconnects can exceed 40 inches in length, an accurate representation of loss at frequencies above *5 GHz* is essential. The challenge is in being able to simulate the performance of a specific interconnect structure based on the first principles input information from a fab vendor about their processes and material choices.

This is the industry goal for lossy interconnects: to establish a methodology, set of tools and a set of specific measureable input properties which will output an accurate prediction of a specific interconnect's performance which will closely match a real measurement.

With such a tool, design space can be explored and the most cost effective balance in materials choices manufacturing processes and design for acceptable loss could be found.

Finalizing this process requires more investigation in determining the best way of characterizing surface texture and material properties, parameterizing the models and accurately measuring the features of the manufacturing process and specific intrinsic of materials features which affect loss.

In the mean time, another approach is being adopted to fill the gap and provide some degree of predictability or final lot acceptance.

## A Practical Approach: Feedback Based Design Flow

As an intermediate goal, one approach is to take the information from a fabricated test coupon and extract from it the parameters and their values which would be used as input to a simulation tool which would then accurately predict the performance of all interconnects fabricated in the same way.

Two additional features of this process would increase the value of this approach. First would be to allow scaling to different cross section geometries rather than just those specific lines that look like the features of the coupon's test line, assuming the same materials and processes for all the layers. This allows the possibility of exploring design space to optimize the cross section design.

As an added bonus of this process, it would be great if there was a strong correlation between the parameter values of the mathematical expressions which results in a good fit and specific manufacturing processes or material features. This way, the specific root cause of the loss could be identified and direct decisions made about adjusting the design, the manufacturing process or the materials selection to optimize the cost and performance based on the actual root cause. This design process could provide some feedback to the fab house on where to look to bring a board into closer compliance to a total loss spec.

A feedback based design flow involves the following steps:

1. Measure the S-parameters of the test lines on the coupon
2. Convert them into a useful form
3. Select the mathematical expression
4. Fit the parameters of the mathematical expression to the measured data
5. Evaluate the quality of the mathematical fit
6. Interpret the parameters in terms of design, materials or process
7. Use the parameters in a circuit simulation of the board level interconnects
8. Adjust the design to balance cost, manufacturability and performance

A key step in this process is converting the raw, measured, S-parameters of various test lines into a format from which the material properties can be more easily extracted. This means the artifacts from non 50-ohm lines and non-transparent launches are removed.

A variety of techniques are available to accomplish this task. For example, the launches can be de-embedded, and then the ports re-normalized. Two lines of different length can be measured and the generalized modal S-parameters (GMS) extracted. Or, the measurements from multiple length lines can be combined to directly extract the complex propagation constant for the interconnect medium using the multi-line approach [12].

These techniques are not the topic of this paper. Instead, in this paper, we explore some of the other elements of this process; in particular, the analysis of the S-parameters with self normalized ports obtained from both measurement and simulation, and the properties of the various loss descriptions.

## Exploring Design Space

In this study, the simulated or measured 2-port S-parameters for candidate uniform stripline transmission lines were created. In most cases, the simulations in this study were performed with Simberian's Simbeor [13].

To be consistent, the following geometry features were used in all simulations:

- Line width = 7 mils
- Conductor thickness = 0.7 mils ( ½ oz copper)
- Rectangular cross section
- Bulk copper conductivity
- Dielectric thickness above and below the signal conductor = 8 mils
- The typical interconnect length simulated = 1 inch
- Nominal dielectric constant @ 1 GHz = 4
- In all examples, the surface texture was applied to all copper surfaces

This results in close to a 50 Ohm line. The maximum return loss from 1 MHz to 40 GHz was less than -30 dB in all cases. This means reflections have no impact on the insertion loss results.

A simple analysis process was used to quickly identify the features of the mathematical expression or measured data.

The time delay of the interconnect, in nsec, was extracted from the unwrapped phase of the insertion loss using:

$$TD(f)[nsec] = -1 \text{ x } \left(unwap\left(phase(S21)\right) / 360\right) / frequency[GHz] \text{ x } 1e9$$

Where
TD(f) is the time delay of the interconnect in nsec
S21 is the complex insertion loss
Frequency is the frequency of each value, in GHz

It should be noted that when unwrapping the phase of the insertion loss to get the phase delay, a short enough frequency interval must be used so that the transitions from the lower to the upper half of the unit circle can be accurately counted. This is especially important when using a log freq increment.

From the time delay and the interconnect length, the effective dielectric constant can be calculated from:

$$Dk_{eff}(f) = \left(\frac{11.8}{Len[in]} TD(f)[nsec]\right)^2$$

Where
$Dk_{eff}(f)$ is the effective dielectric constant of the interconnect
11.8 is the speed of light in air in inches/nsec
Len is the interconnect length in inches
TD(f) is the time delay in nsec

The interconnect losses are described by the insertion loss. Assuming that the conductor and dielectric losses can be separately described by a resistance per length and a conductance per length the insertion loss can be written as

$$S21(f)[dB] = 4.24 \text{ x } \left(\frac{R_{Len}(f)}{Z_0} + G_{Len}(f) Z_0\right)$$

Where
S21 is the insertion loss in dB

$R_{Len}$ is the resistance per length
Z0 is the characteristic impedance
$G_{Len}$ is the conductance per length

While we may be able to separate the conductor and dielectric losses when synthesizing the S-parameters, there is no practical way of separating the conductor and dielectric contributions to loss or dispersion from already existing S-parameter data. The egg has been scrambled.

This factor creates the most significant challenge when interpreting the S-parameters of a uniform transmission line. The impact from both conductor and dielectric mathematical expressions are inseparably intertwined in the loss and dispersion. The only hope of gaining some insight into their relative contributions lies in their different frequency dependences.

Each conductor and dielectric mathematical expressions will contribute a different frequency dependence in both loss and dispersion depending on the parameter values selected. This is the focus of the analysis here. The analysis of the frequency dependence of each loss mathematical expression can only be done as a numerical study, where each term can be isolated and the impact on the frequency dependence of loss and dispersion evaluated.

The frequency dependence of the insertion loss, in dB, arises from the separate frequency dependence of the resistance per length and conductance per length. To first order, the resistance should vary roughly with the square root of frequency, dominated by skin depth effects and the propagating medium permittivity should vary linearly with frequency due to the motion of the dipoles.

It's possible to quickly identify the frequency dependence of the insertion loss and which term dominates by taking the unusual step of plotting the insertion loss, in dB, and frequency on a log-log scale and comparing the slope of the curve to a slope of ½ and slope of 1. Of course to plot the insertion loss on a log scale, the absolute value of the insertion loss must be used.

An example of the insertion loss vs. frequency on a log-log scale for the case of a measured 10 inch long stripline in an FR4 type material is shown in Figure 5.

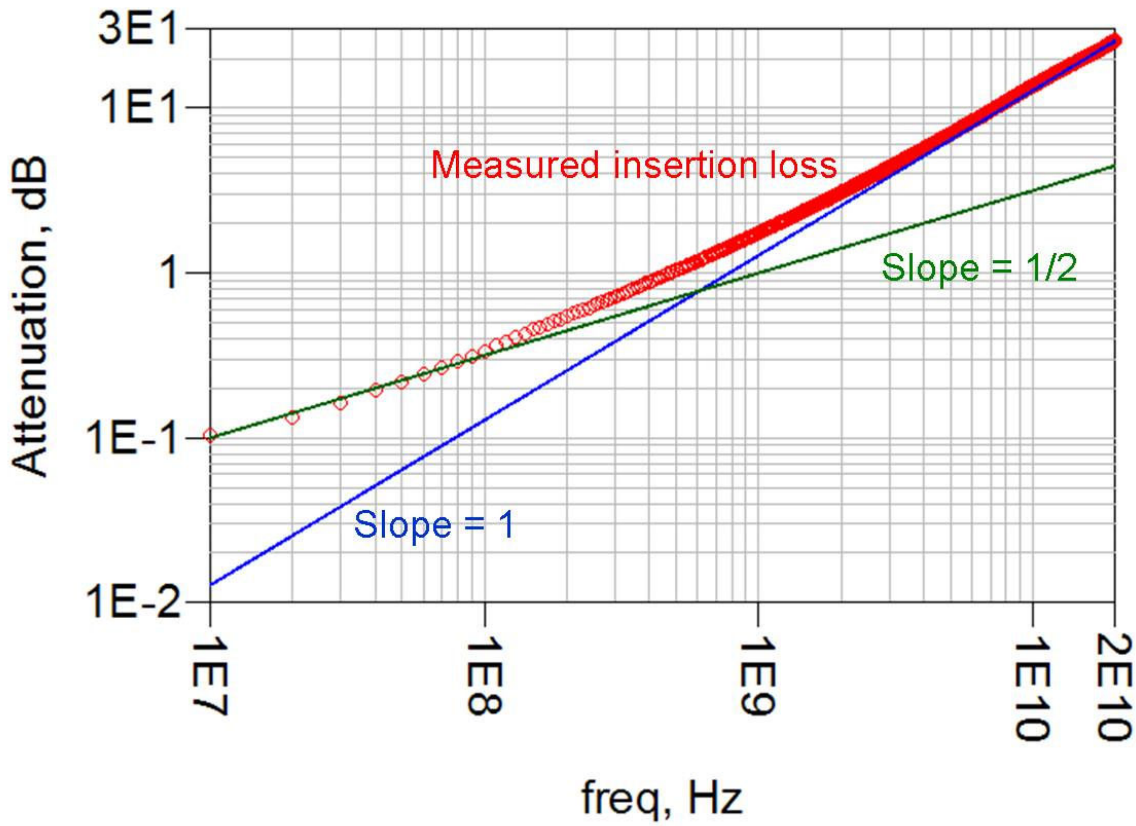


**Figure 5. Measured insertion loss on a log-log scale with reference lines of slope 1 and 1/2.**

In this example, the value of a log-log plot is apparent. This clearly shows the skin depth related effects dominating at low frequency and the dielectric effects dominating at higher frequency.

In the following case studies, each of the mathematical expressions for conductor loss and dielectric loss are separately investigated and how the frequency dependence of the loss and dispersion vary as some of the parameter values are adjusted.

## Case 1: lossless dielectric with smooth copper

To establish a baseline, the simplest case of a lossless dielectric, with Dk = 4, flat with frequency, is used as the laminate with smooth copper which includes the frequency dependence of resistance from skin depth. The simulation using Simbeor includes the losses from the return plane.

The insertion loss should vary with the square root of frequency, above about 10 MHz, where the skin depth drops below the geometric thickness. There will be dispersion due to the current redistribution.

To verify this dispersion in the effective dielectric constant is due to the inductance varying with frequency, a second simulation was created using a perfect conductor so that the skin depth is significantly less than 1 micron at 1 MHz. This effectively makes the current distribution in the simulated frequency range constant and just on the surface of the conductors so there should be no dispersion.

Figure 6 shows the results for case 1, giving an indication of the sort of insertion loss and dispersion just from smooth copper and a perfect conductor for this 7 mil wide uniform transmission line.

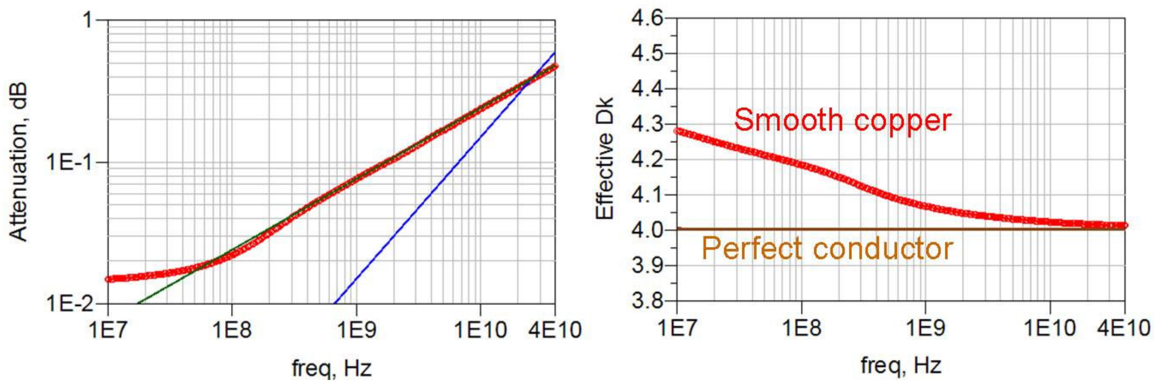


**Figure 6. Insertion loss and dispersion for smooth copper. The square root of frequency is an excellent fit to the insertion loss, while the linear dependency with frequency is a poor fit even at the highest frequency.**

The dispersion from the current re-distribution dominates the effective dielectric constant below 1 GHz. Above 5 GHz is it negligible.

## Case 2: Hammerstad mathematical expression for surface roughness

The Hammerstad mathematical expression is an empirical approximation to the 2 dimensional triangular surface features used by Samuel Morgan [14] and with a linear distance up and down over the peaks and valleys that is twice the straight line distance. This effectively means the triangles have a 60 degree angle from the surface.

In this example, a lossless dielectric was used but the surface roughness was fit with the Hammerstad mathematical expression using three values of the rms surface roughness, 1 micron, 3 microns and 5 microns. The insertion loss and effective dielectric constant are shown in Figure 7.

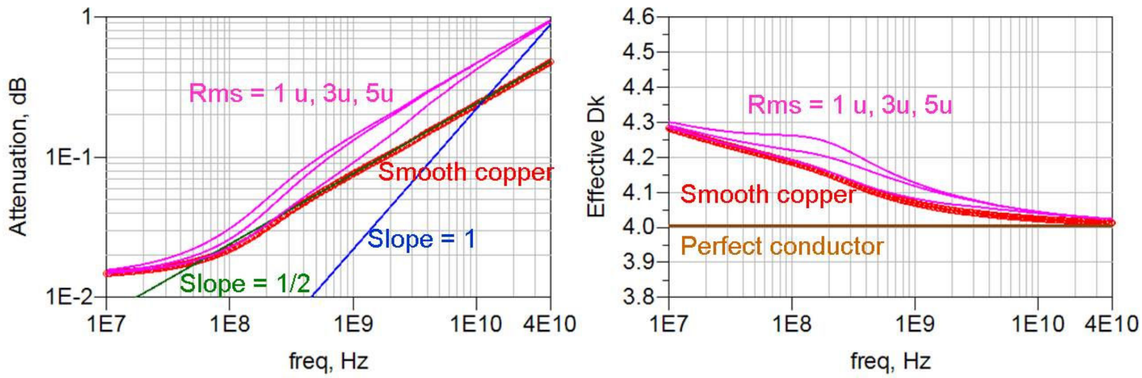


**Figure 7. Simulated insertion and effective dielectric constant for the case of lossless dielectric but surface roughness loss described by the Hammerstad mathematical equation.**

As the rms surface roughness parameter, Δ, increases from 1 micron to 3 microns to 5 microns, the insertion loss increases steeper than the square root of frequency, but not really approaching a linear with frequency dependence. The Hammerstad mathematical expression predicts a loss which saturates when the skin depth is very small compared to the rms roughness parameter, roughly above 10 GHz in this case.

The dispersion is larger at lower frequency with surface roughness.

## Case 3: Modified Hammerstad mathematical expression

In this expression, the Hammerstad mathematical expression is modified to allow any angle for the triangular teeth structures. Rather than the angle, another parameter is the roughness factor, the ratio of the total roughened surface distance to the linear distance along the surface. Effectively, there is twice as much surface area from the teeth structure than just the flat surface.

In this example, the rms roughness was held constant at 1 micron and the scaling factor (*SF*) changed to 2, 4 and 6. With a scaling factor of 2, this is the Hammerstad mathematical expression. A scaling factor of 4 and 6, would correspond to an empirical fit to a higher effective surface area created by sharper teeth and would cause additional power loss increases to 4x and 6x.

As the teeth structure get sharper, and take up more surface area, they will provide a larger surface to absorb tangential H field into the surface and the power losses will increase. Figure 8 shows the impact of increasing scaling factor for constant 1 micron rms peak height.

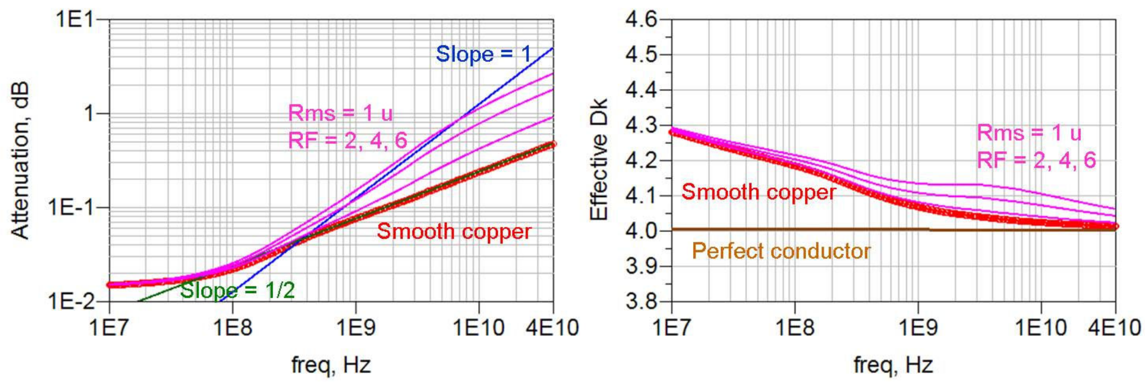


**Figure 8. Simulated insertion and effective dielectric constant for the case of lossless dielectric but surface roughness losses given by a modified Hammerstad mathematical expression in which only the scaling factor is increasing for a fixed 1 u rms peak height.**

In this example, increasing scaling factor means increasing losses from the conductor, and more importantly, in some frequency regions, the losses increase close to a linear frequency rate. The dispersion also increases in the high frequency region.

## Case 4: Huray snowball model

There are three parameter that define the Huray first principles snowball model:

$a_i$, the radius of the $i^{th}$ sphere,
$A_{Matte}/A_{Flat}$, the relative area of a Matte base compared to a Flat base area
$N_i\,(4\pi a_i^2)/A_{Flat}$, the total area of the additional spheres compared to a Flat base area.

The total area of the additional spheres per unit Flat area means more surface area and absorption of tangential H fields into the electrodeposited conductor anchor nodules than that provided by $A_{Matte}/A_{Flat}$. However, each additional loss term is dependent upon frequency according to $\frac{3}{2}\sum_{i=1}^{j}\left(\frac{N_i 4\pi a_i^2}{A_{Flat}}\right)\Big/\left[1+\frac{\delta}{a_i}+\frac{\delta^2}{2a_i^2}\right]$

The factor of 3/2 and the frequency dependence in the denominator occur from the dipole approximation for copper snowballs (i.e. to first order they can be modeled as copper spheres).

In exploring the impact of these parameters on the loss and dispersion, two ranges are considered. In the first example, a fixed ball diameter, 1 micron, and number of balls, 30, was used for various base areas of (10 microns)$^2$, (7 microns)$^2$ , (5 microns)$^2$ and (3 microns)$^2$. This effectively increases the surface area density for absorption. The smaller the base area the higher the expected loss. Figure 9 shows the simulated insertion loss and dispersion for these four values of base areas.

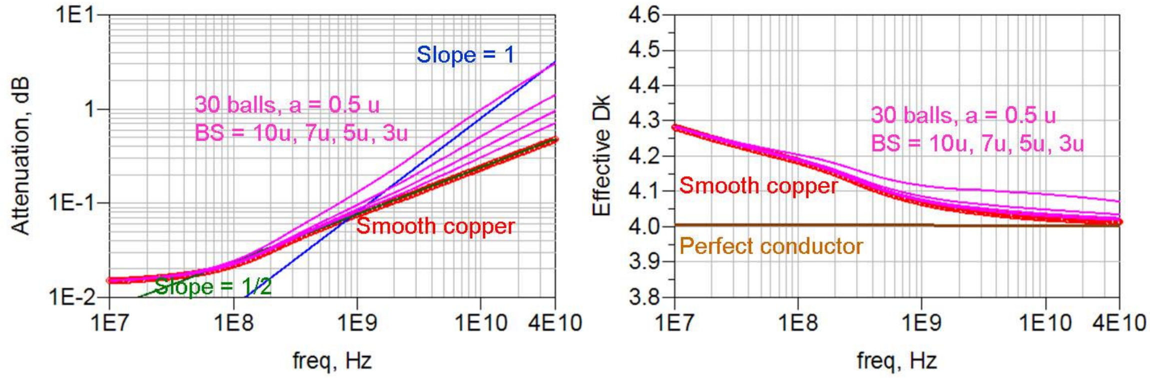


**Figure 9. Simulated results using the Huray snowball model for the case of 30 balls, each 1 micron in diameter, with different flat base areas.** I would like to see the code that deduced this chart to verify its validity.

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It is interesting to note that as the density of ball area per unit flat area increases, the slope of the insertion loss increases and approaches a linear frequency dependence. The dispersion is only slightly affected by the density until the density gets very high.

As a second example, the tile base area of (5 microns)$^2$ was selected, with 30 balls of diameter 2 microns, 1 micron and 0.5 micron. Again the assumption of uniform ball size was made.

As the ball diameter decreases two effects happen. The frequency at which the excess loss turns on increases in frequency as it scales with the ball radius to skin depth value. Secondly, as the ball diameter decreases, the effective surface area available for absorption and loss decreases and the total excess loss decreases. This is related to the square of the ball diameter.

These two effects are seen in the simulated examples shown in Figure 10.

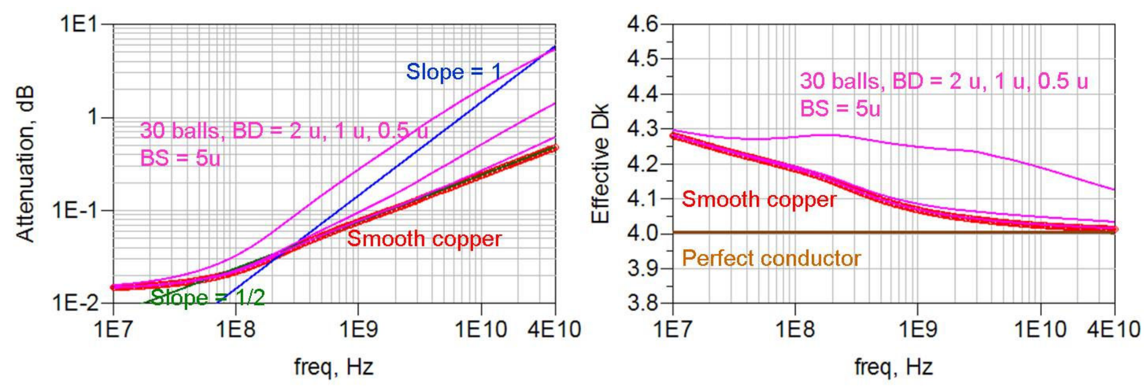


**Figure 10. Simulated insertion loss and dispersion with the Huray snowball model, changing just the ball diameter from 2 microns to 1 micron to 0.5 micron but assuming a constant base area of (0.5 microns)$^2$ and uniform size balls.**

In this example, it's clear how sensitive the results are to the ball diameter. This one parameter affects when the loss turns on, how much loss and how much dispersion results. Large nodule ball area per unit base area generates a lot of loss and has a big impact on dispersion.

These mathematical expressions of loss all point out that it's the increased roughened surface area which most strongly affects the amount of loss from surface texture. This suggests that to engineer the lowest loss surface texture while still providing some adhesion, a selectively patterned surface should be used. A nominally flat surface should be the starting place, with a patterned surface treatment spaced every mil or more with large features (small surface area compared to the conductor volume). This would cut the surface area by as much as 10x to 20x, reducing the surface power loss by a comparable amount.

While the modified Hammerstad mathematical expression and the Huray snowball model use different parameters, they both have a parameter which describes a surface deviation and a surface area factor. Either approach might be a candidate for comparing to the measured loss contribution from surface texture.

For example, using the values of ball diameter of 0.5 microns, number of balls = 40 and base area of (9 microns)$^2$, the Huray snowball model can be matched with the modified Hammerstad expression using the parameters of rms feature of 0.4 microns and surface factor 1.23. This match is shown in Figure 11.

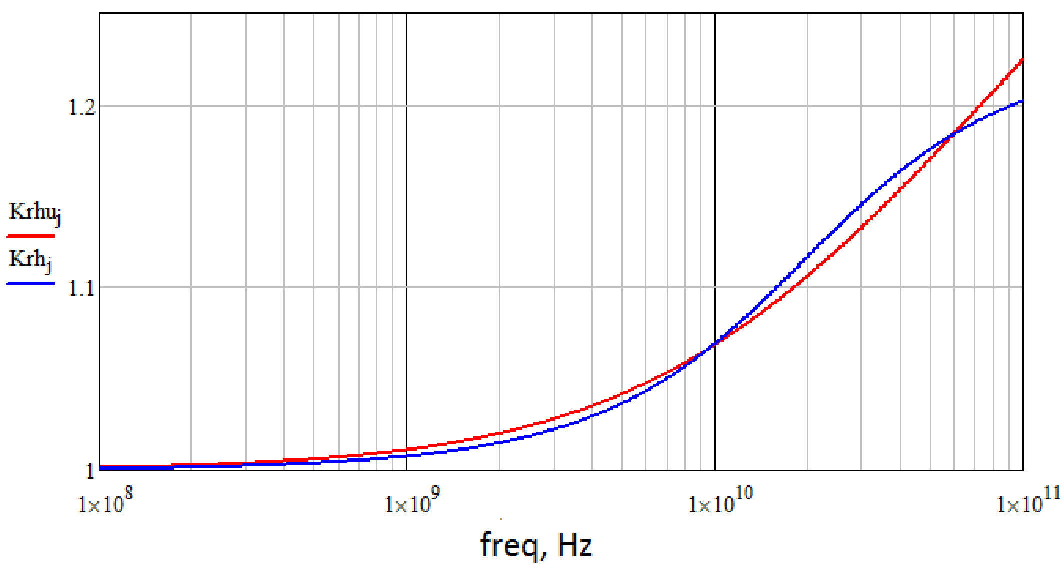


**Figure 11. Huray snowball model (red line) and modified Hammerstadt expression (blue line) with roughness correction coefficient matched up to 60 GHz.**

## Case 5: Dielectric loss model: relationship between Dk slope and Df

In the wide band Debye model with the Svensson-Djordjevic approximation, the dielectric constant varies linearly with the log of frequency. The slope of the dielectric constant with the log of the frequency, is a direct measure of the dissipation factor. In fact, the dissipation factor is given by:

$$Df(f) = \frac{\varepsilon''(f)}{Dk(f)} = \frac{0.682}{Dk(f)} \frac{\Delta Dk}{\log(f_2) - \log(f_1)} = \frac{-0.682}{Dk(f)} \text{slope}$$

In the following example, the five parameters for the wideband Debye model were used:

Dk = 4
Df = 0.02, 0.01, 0.005, 0.002
f = 1 GHz
f1, the low frequency limit = 100 kHz
f2, the high frequency limit = 100 GHz.

Figure 12 shows the simulated insertion loss and dispersion for the case of a perfect conductor but the dielectric properties above.

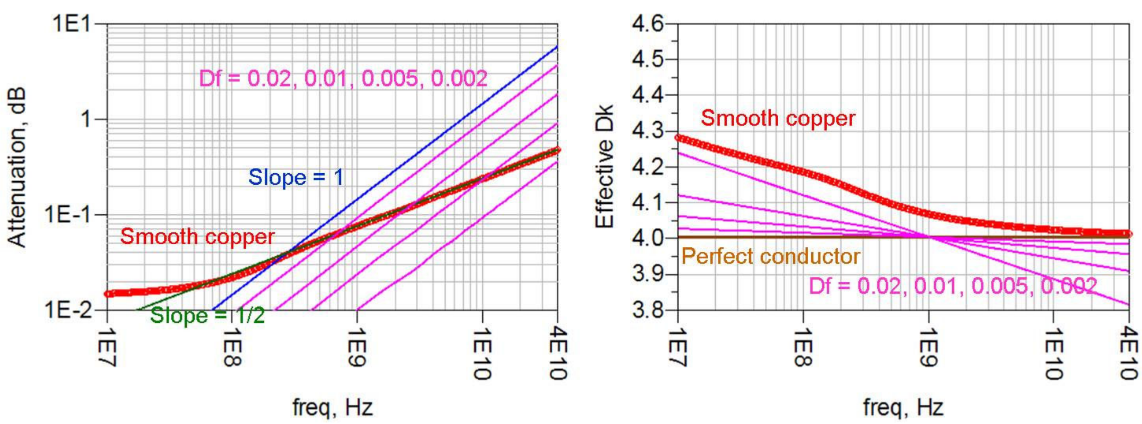


**Figure 12. Simulated loss and dispersions for perfect conductor and wide band Debye model where only the Df is varied from 0.02, 0.01, 0.005, 0.002.**

There are two important features of the wideband Debye model in this example. The insertion loss from just dielectric loss, shows a linear dependence on frequency. The slope on the log-log scale matches the reference slope of 1.

Secondly, as expected, the slope of the effective dielectric constant over frequency is related to the dissipation factor. The lower the dissipation factor, the flatter the dielectric constant over frequency and the less the dispersion. Though the vertical scale on the graph is linear, for small variations, the log of a number and the number have the same relative difference.

## A Suggested Methodology

This analysis points out that while it is not possible to directly separate the conductor and dielectric loss in the measured response of a transmission line sample, it may be possible to find a set of parameters which match the measured performance of measured samples.

The input to a typical model that most simulators understand is:

<u>the cross section parameters:</u>
h, dielectric thickness
w, line width
t, conductor thickness

<u>The conductor loss parameters:</u>
Copper Conductivity
Modified Hammerstad model: rms value, scale factor
Huray snowball model: relative Matte to Flat base area, ball diameter, number of balls per unit Flat area

<u>Dielectric loss parameters:</u>
Dk
Df
At f
f1 low freq limit
f2 high freq limit

Some insight into the behavior of a sample can be gained by comparing the measured response with the simulated response which includes just some of the mathematical expression features. The starting place can be the cross section parameters based on the known sample properties, the smooth copper losses and the lossless dielectric properties.

The copper conductivity parameter can be matched to the low frequency loss and the Dk parameter matched to the 1 GHz measured dielectric constant. Using values of copper conductivity and Dk = 4.33, the comparison between measured and simulated responses are shown in Figure 13.

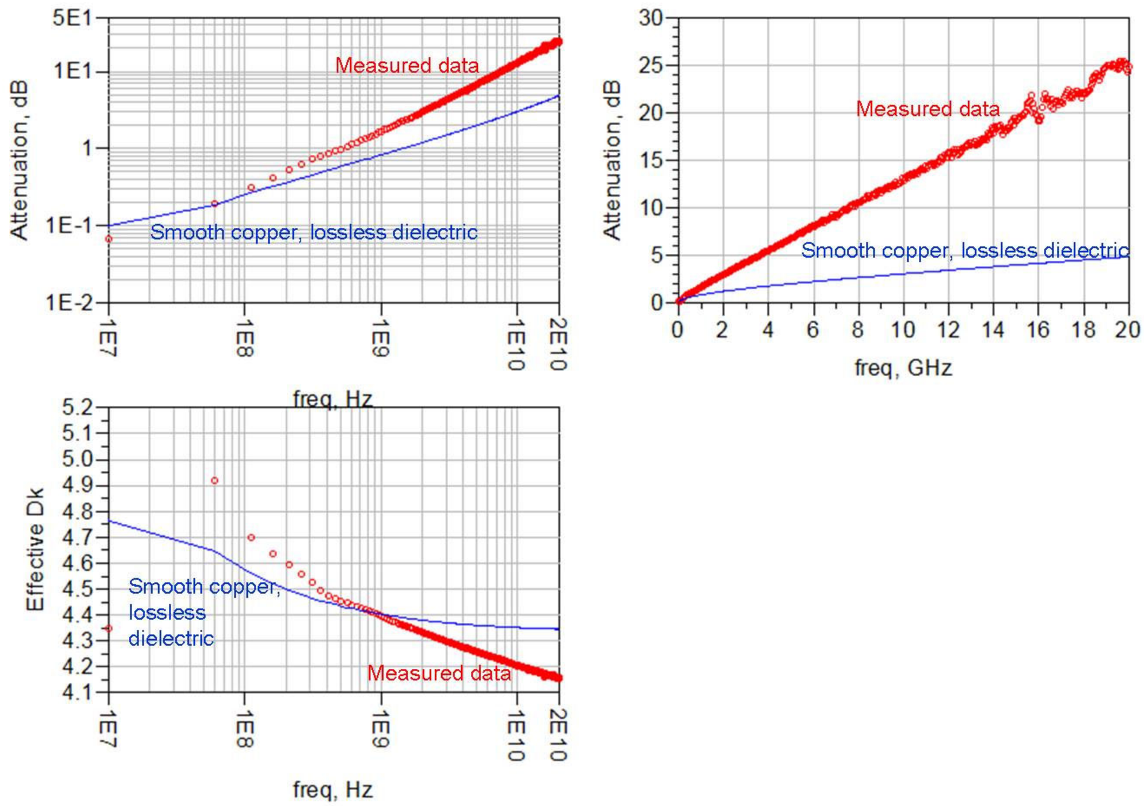


**Figure 13. Example of the measured insertion and dielectric constant for an FR4 type sample compared to the simulation using smooth copper and lossless dielectric. Note, for typical samples measured in a high volume manufacturing environment, there will be noise, especially at the low frequency and high frequency ranges.**

It is surprising how much of the dispersion is due to the smooth copper properties. In this example, most of the high frequency loss seems to be contributed by the dielectric. This suggests that the avenue to lower loss, in the 4-10 GHz range, is by focusing on material selection for lower dissipation factor.

The next step is to optimize the dielectric loss to match the insertion loss. Figure 14 shows the simulation using a dissipation factor of 0.022.

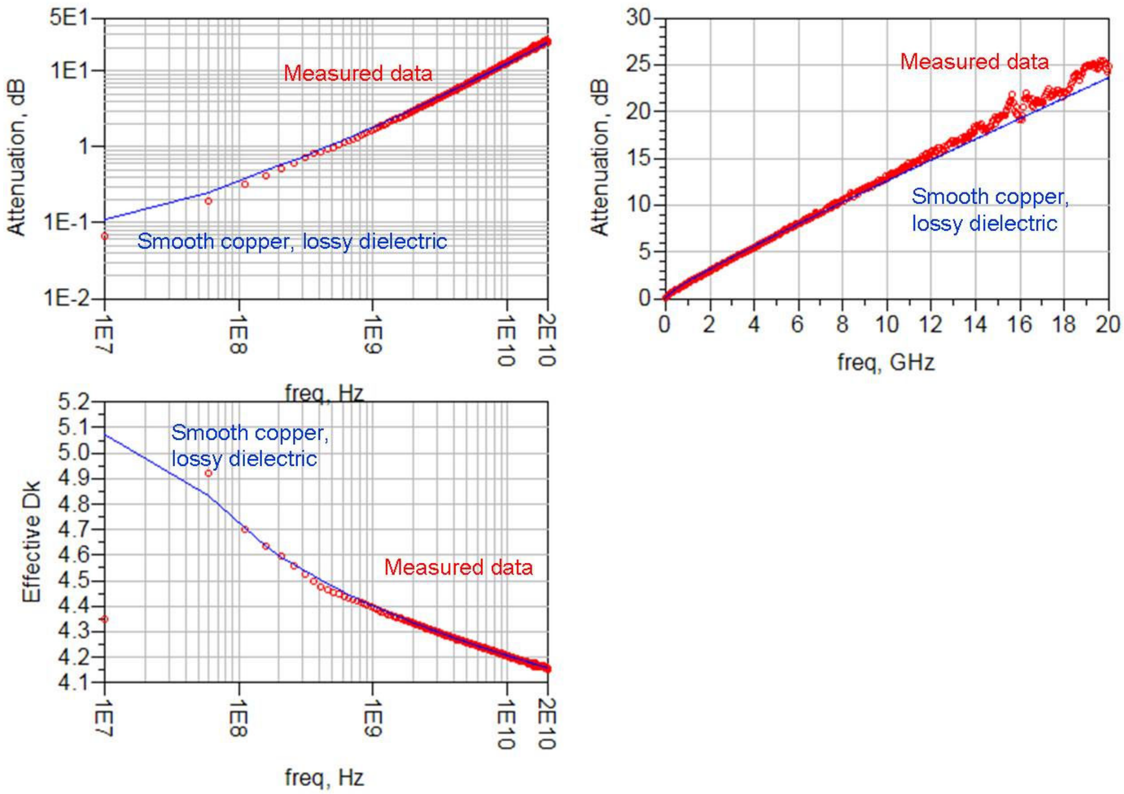


**Figure 14. Measured insertion loss and dielectric constant with smooth copper and lossy dielectric. The slight deviation at the lowest frequency is due to the difficulty of accurately measuring total attenuation less than 0.1 dB related to calibration and reproducible, low contact resistance connections.**

The dispersion in the dielectric constant is seen to be very well described with just smooth copper and dielectric loss. As seen in the linear attenuation plot, the smooth copper loss and dielectric loss are able to match the measured performance up to about 10 GHz. Above this value, the loss is larger than predicted. This is probably due to the impact of surface texture. In this example, the Huray snowball model is used to match the measured performance.

With a little optimization, a set of parameters were found to match the complete performance. Figure 15 shows the final match of the measured and simulated response for this sample.

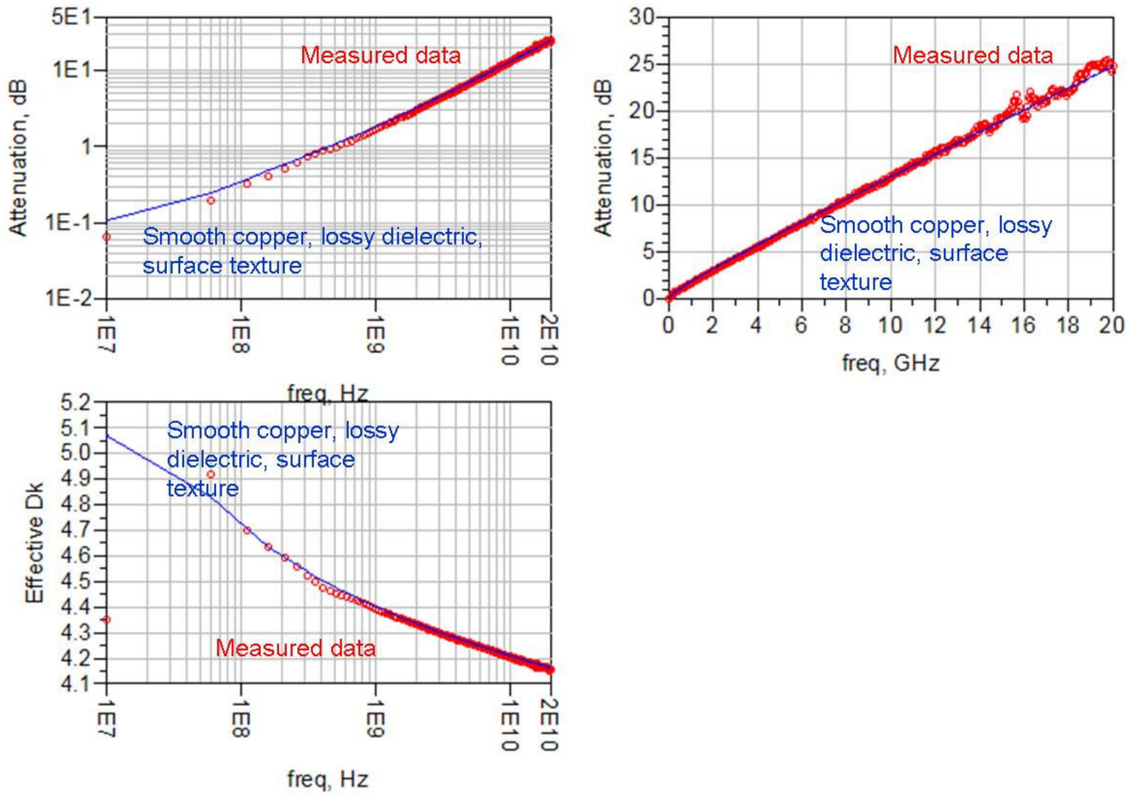


**Figure 15. Final match of the measured response and simulated response using wideband Debye model, smooth copper and Huray snowball model.**

This same process can be applied to a variety of samples. An example of another measured transmission line sample with a low profile ($A_{Matte}/A_{Flat}=1$) is shown in Figure 16.

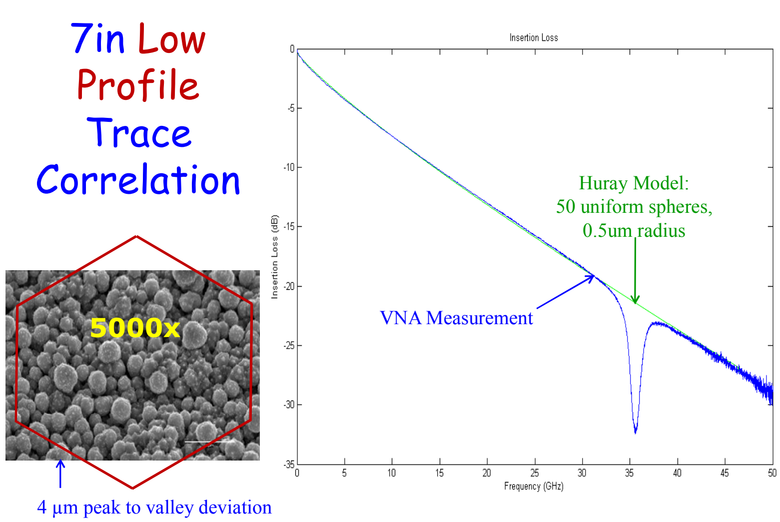


**Figure 16. Insertion loss measurements on a 7 inch Low Profile trace as a function of frequency (shown in blue) compared to the Huray snowball model (in green) with the further approximations $A_{Matte}/A_{Flat}\approx 1$ and $N_i/A_{Flat}=50$ uniform spheres per (100 micron)$^2$ area shown in the red hexagonal structure at lower left.**

As a final example, 2-inch and 4-inch long Megtron6 stripline transmission lines, courtesy of Molex Corp. were measured. A wideband Debye model with the modified Hammerstad expression were fit to the insertion loss and group delay. Figure 17 shows the excellent comparison between measured and simulated results.

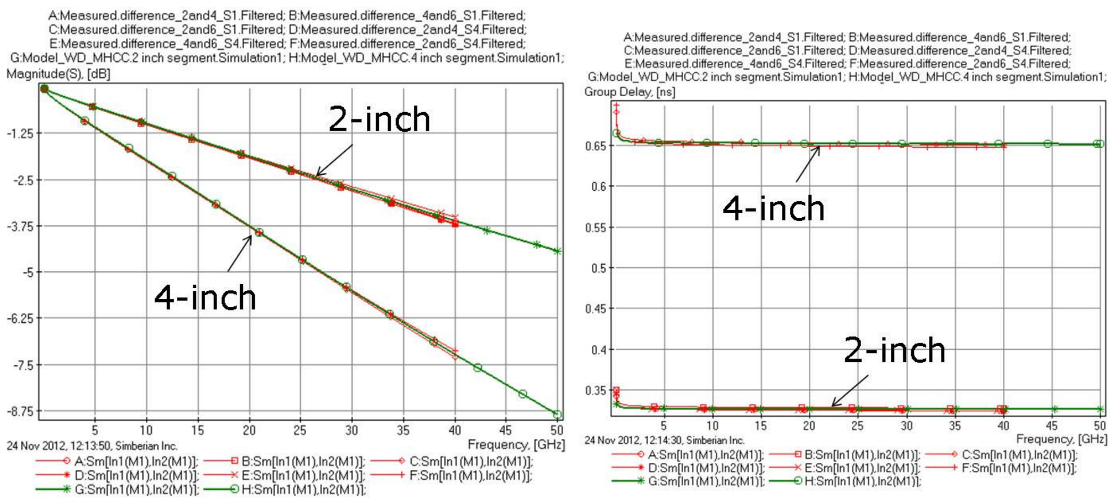


**Figure 17. Measured insertion loss and group delay for Megtron6 samples compared with the simulated values using a wideband Debye model and a modified Hammerstad approximation with values of Dk = 3.7, Df = 0.002 and SR = 0.3 u and RF = 5.**

## Conclusion

While it is still not practical to take information obtained directly from a fab vendor and turn this into a first principles model which accurately describes the complete performance of an interconnect, it is practical, in a variety of material systems to take the measured response from a test coupon and fit parameters associated with conductor and dielectric loss.

The dielectric loss can be modeled with a wideband Debye model and the surface texture of copper can be described with a variety of mathematical expressions. By fitting the measured insertion loss and effective dielectric constant, a complete description of a transmission line can be extracted from manufacturing test coupons in some samples using just a few parameters.

These parameters can be used as input to a variety of popular simulators to create scalable transmission line models for any interconnect structure. These parameters also may be useful in providing some insight into where to attack for reduced loss.

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