Analysis of Test Coupon Structures for the Extraction of High Frequency PCB Material Properties

White Paper

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Abstract

The driving forces behind low cost RF/Microwave mponents and the ever increasing data rates of the digital world have resulted in a proliferation of printed circuit board (PCB) materials and manufacturing techniques to provide cost effective solutions for modern day applications. This has resulted in the need for practical ways of identifying the frequency dependent loss properties of the after fabrication or "as-fabricated" PCB designs for accurate pre- and post- layout simulation. Variations in key parameters such as dielectric constant, loss tangent, dielectric height, etched trace width, surface roughness, glass weave, moisture content, etc. can easily reduce the effectiveness of simulations to predict the final design performance. Few companies have the time, money, or equipment to fully dissect a fabricated PCB and determine all of these key dimensions and material loss properties for a design. There is a strong need for simpler techniques, such as measurable test structures, that can enable extraction of material properties for improved accuracy of simulations. Recent papers using two transmission line lengths do demonstrate measurements of dielectric constant and complex loss tangent [1,2,3,4], but this method makes it difficult to predict the as-fabricated PCB trace width and dielectric height. The intent of this paper is to explore the addition of Beatty series resonant impedance structures [5] to improve the accuracy of extracting the as-fabricated PCB material properties for the purpose of constructing 3D-EM simulations.

Keywords—material properties, PCB test structures, high speed digital, 3D-EM

Introduction

The proliferation of epoxy based multi-layer laminate PCBs for high speed digital circuits running at microwave frequencies puts new demands on the world of EM simulations. Recent publications have clearly shown that assuming dielectric loss to be constant with frequency may have worked well for narrow band microwave applications, but for wide band digital simulations it produces non-causal results. Surface roughness which could be ignored with high quality thin film circuits is sparking new debates [6,7] to understand the physics and accurately predict the variations seen between standard copper foil and low profile versions. One issue that has not been widely addressed is the as-fabricated transmission line etched trace widths and the dielectric height which are critical for the set-up of accurate multi-Gigahertz EM simulations.

PCB fabrication can be very repeatable for a specific PCB design and fabrication process, but accurately predicting the nominal value for a given fabricator's process is best done with the measurement of simple test structures. This is especially true when designing to the limits of PCB "lines and spaces" with narrow transmission lines that are sensitive to fabrication tolerances and require the fabricator to adjust the original CAD data in order to meet impedance targets. This is becoming a common practice in the PCB industry, and yet few EM simulations are done with the after fabrication as-fabricated trace widths and dielectric height since it is far easier to obtain the pre-fabricator CAD data and laminate manufacturer's generalized data sheets. It is proposed that by using the 2-Line test structure method for material properties with the addition of a simple impedance varying Beatty resonant test structure, one can determine the PCB transmission line dimensions as well as benefit from additional insights into the electrical performance of the as-fabricated PCB material.

The Beatty structure [5] utilizes a simple change in trace width to create a measurable delta change in impedance. The measured impedance change for a given increase in physical transmission line trace width helps to define unique dielectric height and trace width dimensions for improved EM simulation accuracy of the as-fabricated structures. This paper provides a description of the methodology and then demonstrates the process utilizing measurements from fabricated test coupons.

Description of The Methodology

The two key factors in determining the performance of a transmission line are the impedance and the loss. The IPC recommended approximation for the impedance of an internal stripline trace with grounds on either side help to provide insight into the controlling features that are needed for an accurate EM simulation [8].

$$Zo = \frac{60}{\sqrt{\varepsilon_r}} \ln\left(\frac{2b+t}{0.8w+t}\right) \tag{1}$$

Where:

Z_o, characteristic impedance (Ohm)
b, the dielectric height between reference planes (mil)
t, copper thickness of the PCB trace (mil)
w, trace width (mil)
ε_r, dielectric constant

The approximation for loss when a sine wave travels down a stripline with 1oz copper for the conductor is: [8]

$$\alpha_{dB} = \alpha_{cond} + \alpha_{diel} \tag{2}$$

$$\alpha_{cond} = \frac{36}{wZ_o} \sqrt{f} \tag{3}$$

$$\alpha_{diel} = \frac{\pi}{c_o} f \tan\delta \sqrt{\varepsilon_r} \tag{4}$$

Where:

Z_0, characteristic impedance (Ohm)
f, is the sine wave frequency (GHz)
w, trace width (mil)
c_o, is the speed of light in vacuum
$\tan\delta$, loss tangent
ε_r, dielectric constant

The equations show how impedance relates to the physical dimensions of the dielectric height between the two reference planes and the trace width for a given dielectric constant.

Two Transmission Line Segments

To calculate the loss tangent and the dielectric constant one can use the 2-Line method that enables the removal of connector fixturing losses and impedance mismatches so that the total attenuation for a given length L of transmission line can be measured. At high frequencies impedance reflections from the fixturing can be significant and must be removed in order to see the true loss of the transmission line [9].
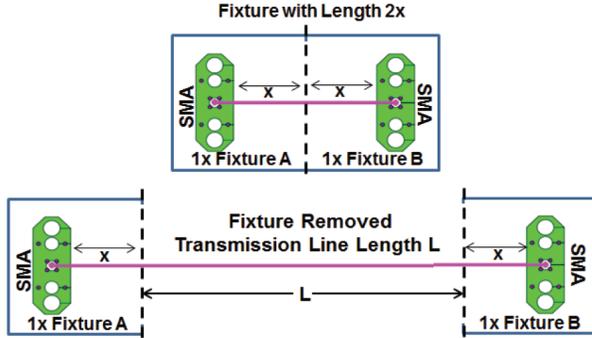

Figure 1: 2-Line Test Coupon Structures for measuring the S-Parameters of the Length L of transmission line with the connector fixturing removed.

Since the transmission line is typically low loss, so it is helpful to select L such that it is always longer (higher loss) then the connecting fixture and thus less sensitive to the fixture removal quality. There are numerous methods for removal of the connector fixture effects including NIST multi-line cal [3], optimization to split the 2x fixture into two equal S-Parameters, and mathematical closed form expressions like the Automatic Fixture Removal (AFR) used by Keysight Technologies PLTS software [10]. The AFR is a fast technique for obtaining the S-Parameters for Fixture A and Fixture B which can then be de-embedded from the longer line to obtain the S-Parameters for the length L transmission line. The S-Parameter data can then be used to find the dielectric constant as a function of frequency by using the definition of how fast a signal travels in a dielectric material for a given length L.

$$v(f) = \frac{c_o}{\sqrt{\varepsilon_r'(f)}} \tag{5}$$

$$\varepsilon_r'(f) = \left(\frac{c_o \times TimeDelay(f)}{L} \right)^2 \tag{6}$$

$$\varepsilon_r = \varepsilon_r' - j\varepsilon_r'' = \varepsilon_r'(1 - j\tan\delta) \tag{7}$$

Where the dissipation factor $\tan\delta = \left(\dfrac{\varepsilon_r''}{\varepsilon_r'} \right)$

By enforcing causality, the real and imaginary permittivity must satisfy the Hilbert transform relationship between the two [1,2] which enables one to create a model that only requires measurement at two frequencies f1 and f2 where f1 is the main frequency of interest and f2 can be set arbitrarily high at 1 Terahertz. This wide bandwidth causal model or the Djordjevic model then determines the behavior of the dielectric constant versus frequency [11].

$$\varepsilon_r{}'(f) = \varepsilon_2' + \frac{\Delta(\varepsilon_2' - \varepsilon_1')}{\log(f_2) - \log(f_1)} \times (\log(f_2) - \log(f)) \qquad (8)$$

And since "to know the real part is to know the imaginary part" then the Kramers-Kronig relationship [1] gives:

$$\tan\delta = \frac{1}{\varepsilon_r{}'(f)} \times \frac{\Delta(\varepsilon_2' - \varepsilon_1')}{\log(f_2) - \log(f_1)} \times (0.682) \qquad (9)$$

The measured S-Parameters for the Length L of transmission line are then used to determine $\varepsilon_r{}'(f)$ and tan δ.
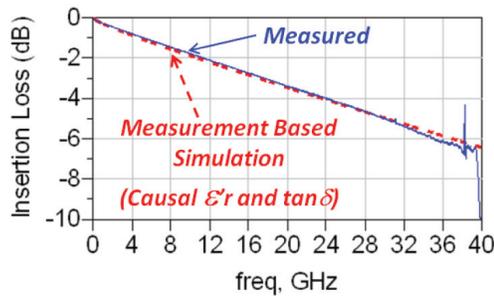
Figure 2: Plotting the loss in dB shows linear loss vs. frequency. At higher frequencies it may be difficult to separate out trace loss from the accuracy of the fixture removal. (11.25 cm of 17.8 mil stripline with R4350/R4450B 36 mil dielectric height. Dielectric constant at 10 GHz=3.78, and Tand=0.006)
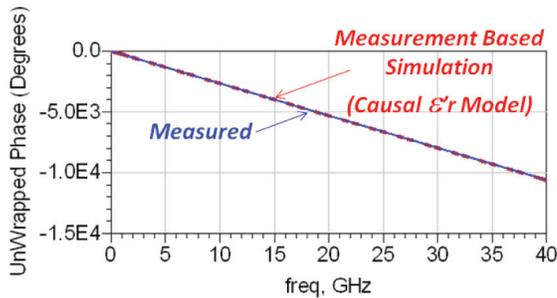

Figure 3: The measured vs. simulated for unwrapped phase shows how the correct selection of dielectric constant insures accurate prediction of the phase or electrical delay. (11.25 cm of 17.8 mil stripline with R4350/R4450B 36 mil dielectric height. Dielectric constant at 10 GHz=3.78, and Tand=0.006)

This 2-Line method works well for determining the dielectric properties and loss models, but this does not address the need for accurate as-fabricated dimensions for creating accurate 3D-EM models.

Beatty Resonant Standard

To find the remaining as-fabricated variables including dielectric height, etched trace width, and copper thickness. The proposed Series Resonant Beatty structure [5] has the advantage of allowing additional data which includes the delta change in impedance for a given delta change in physical trace width along with a resonant ripple in the insertion loss which is affected by dielectric constant and transmission line losses.
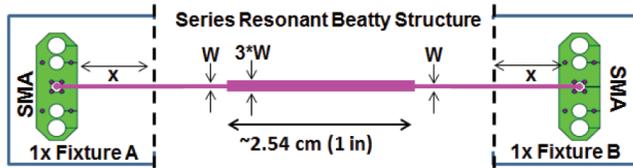

Figure 4: The series resonant Beatty structure provides a delta impedance change for additional information in determining the as-fabricated dielectric height and trace width.

The Beatty standard is constructed by increasing the trace width of the transmission line for a specific distance that results in reflection resonances and broad band ripple on the insertion loss.

Simulation of the methodology

To test out the methodology and explore the sensitivities of the simulated Beatty standard to the dielectric height and the trace width it is useful to use a closed form model such as equations 1, 3, and 4 or a 2D-Planar EM model. It will also be shown that this is a quicker way of predicting the settings needed for the full 3D-EM simulation and avoids trying to tune the material properties with the longer simulation times of the 3D-EM simulator.
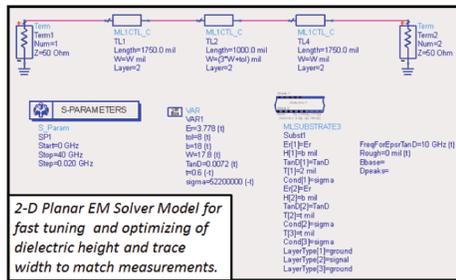

Figure 5: Frequency Domain Simulation for tuning the transmission line parameters. The use of 2D-Planar EM models have improved accuracy over closed form models for wide bandwidth simulations and yet still provide fast optimization and tuning of the variables to match with measured data.

The initial simulation starts with the vendor material data sheets and PCB CAD data to see how far off this method is in predicting the as-fabricated measured performance. The prediction is okay at a few GHz, but for applications going beyond 10 GHz it begins to lose accuracy.
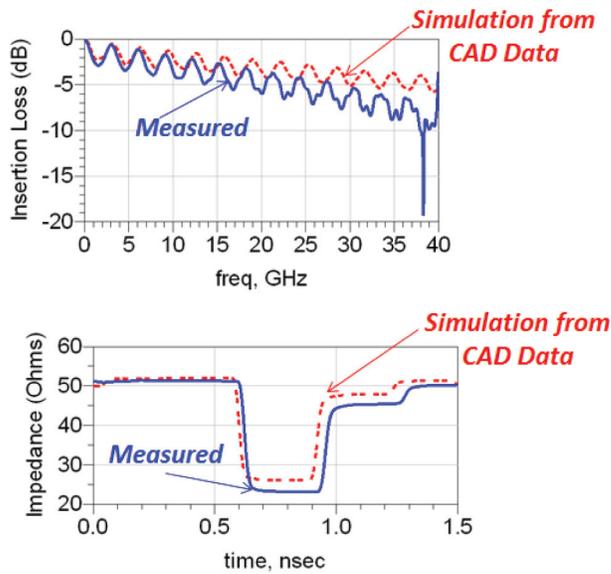
Figure 6: The simulation using CAD design data and vendor supplied data sheets works fine up to a few GHz, but beyond 10 GHz it is helpful to make use of test structures for measurement based modeling.

Updating with the measured loss properties and then tuning to the measured results of the Beatty test structure for both insertion loss and time domain TDR impedance results in much better correlation between simulation and measurements.
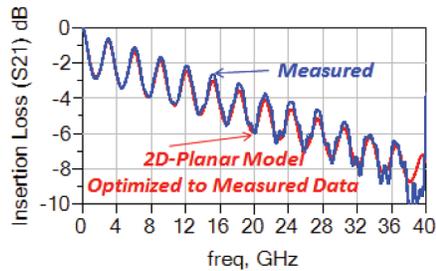
Figure 7: The 2D-Planar model is optimized to match the measurements.

Optimizing the impedance to match the measured Beatty standard helps to understand how the fabrication tolerances impact the PCB performance.
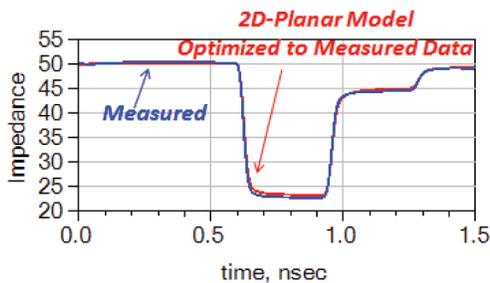
Figure 8: The 2D-Planar model is also tuned in the time domain to match with the measured Beatty test structure impedances.

Measurement vs. simulation

The same settings for the transmission line properties that were determined from the tuning of the 2D-Planar EM model can now be entered into the 3D-EM simulation. These settings immediately provide a very good approximation to the measured data.
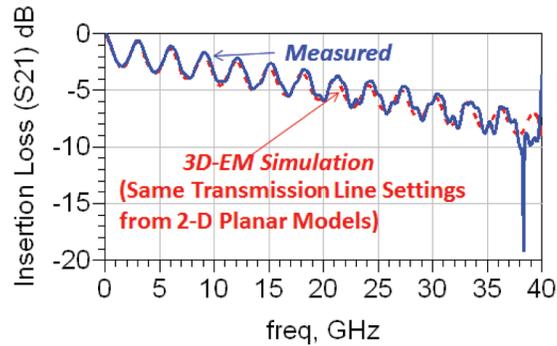
Figure 9: Successfully simulating a wide bandwidth resonant Beatty standard with a 3D-EM simulator using measurement based as-fabricated properties.

The 2D-planar optimized settings also provide a good match with measured impedance data in the time domain for the 3D-EM simulator. It is useful to note that if the meshing density is too low in an EM simulator it can produce inaccurate results like lower then expected impedance.
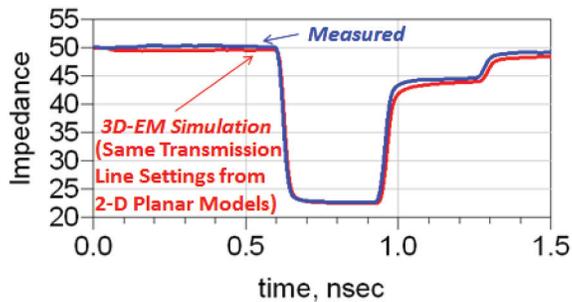
Figure 10: The 3D-EM simulator using measurement based as fabricated properties also predicts the correct impedances of the Beatty test structure.

Conclusion

The well-established 2-Transmission Line or multi-line method of determining dielectric and conductor losses goes a long way towards improving the causal time domain simulations that are needed in today's high speed multi-gigabit systems. However, full 3D-EM simulations also require information on the physical dimensions of the as-fabricated PCB. It has been shown that with the addition of a series resonant Beatty test standard one can improve the accuracy of using a 3D-EM simulator to predict measured performance.

The stepped impedances and resulting frequency domain broadband resonances provide additional data points for optimizing to the fabricated dimensions of the transmission line structure. There is also the added benefit of having a single structure which can verify the PCB fabrication process and run to run repeatability. Future plans are to explore a double resonant structure stepping from 1x trace width, up to 3x width, back down to 2x trace width, and then back to the reference 1x trace width.

Acknowledgment

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