Method to Reduce Coupon Lengths for Transmission Line S-parameter Measurements through Elimination of Guided-wave Multiple Reflections

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Abstract— A new approach is proposed to reduce the length requirements for the measurement of transmission line S-parameters in printed circuit boards (PCBs). In addition, the approach improves the accuracy of measurements. We demonstrate that guided-wave multiple reflections of the transmission lines in the signal propagation direction limits use of short transmission lines samples. The elimination of the guided-wave multiple reflections is introduced to overcome the coupon length limit for accurate measurements of transmission lines S-parameters. We introduce implementations of the proposed approach in the time and frequency domains. For time-domain measurements, we do corrections to remove the reflection noise signals from received signals: the first is the waveform manipulation for removing the internal reflections, and the second is the novel manipulation of S-parameters to compensate for reflected and transmitted losses. For native frequency-domain measurements, that manipulation of the S-parameters is used as a de-embedding correction method to remove the reflected and transmitted signals' contributions from the S-parameters. Simulation results are presented to demonstrate the efficiency and accuracy of the new approach. Results show that the new approach can reduce the minimum length of high-speed transmission lines from 8 inches in traditional industry standards to only 0.5 inch for both 4-port TDR/VNA measurements and 2-port SET2DIL.

1. INTRODUCTION

The signal distortion and attenuation in transmission lines have been a key issue for printed circuit board (PCB) design and application. The issue escalates as the PCB operates at faster speed and higher frequency. Therefore, the PCB industry needs accurate and efficient methodologies to qualify and monitor the performance of PCB manufactured by the supply chains. This is extremely important in high volume manufacturing.There are several methodologies recommended by IPC specification [1], for example, the single-ended TDR/TDT to differential insertion loss (SET2DIL) method [2].

In existing methods, a trace length of 8 inches is typically chosen to achieve accurate measurements of S-parameters, particularly of insertion loss. For manufacturing, shorter coupons are desired, but two major effects require longer coupons: 1) limits of receiver sensitivity require insertion loss to be high enough to exceed the noise floor of the instrument, and 2) for non-50 ohm DUTs (or 100 ohms differential), return loss effects can cause errors in insertion loss measurements. A longer trace is chosen so that the effects of multiple reflections on the insertion loss measurements are negligible, compared to the insertion loss measured, and signal amplitude is sufficient for the instrument used.

In this work, we propose a new approach to reduce the length requirements for the measurement of transmission line S-parameters in printed circuit boards (PCBs). In addition, the approach improves the accuracy of measurements. In this work, we demonstrate that guided-wave multiple reflections of the transmission lines in the signal propagation direction limits use of short transmission lines samples. The reflections can be also interpreted as standing waves, waveguide modes, waveguide resonances, or layered medium multiple reflections. The elimination of the guided-wave multiple reflections is introduced to overcome the coupon length limit for accurate measurements of transmission lines S-parameters. We introduce implementations of the proposed approach in the time and frequency domains. For time-domain measurements, we do corrections to remove the reflection noise signals from received signals: the first is the waveform manipulation for removing the internal reflections, and the second is the novel manipulation of S-parameters to compensate for reflected and transmitted losses. For native frequency-domain measurements, that manipulation of the S-parameters is used as a de-embedding correction method to remove the reflected and transmitted signals' contributions from the S-parameters. Simulation results are presented to demonstrate the efficiency and accuracy of the new approach. Results show that the new approach can reduce the minimum length of high-speed transmission lines from 8 inches in traditional industry standards to only 0.5 inch for both 4-port TDR/VNA measurements and 2-port SET2DIL.

This can significantly reduce coupon area, allowing more compact PCB designs and enabling the ability to include coupons in the actual design, rather than as "outriggers" (dedicated portions of a PCB panel for manufacturing testing).

2. METHODOLOGY

2.1. Multiple Reflections of Transmission Lines with Finite Length

Figure 1 shows the diagram for a transmission line with length "d" and the signal propagation in z direction. For non-magnetic materials $\mu_1 = \mu_2 = \mu_3 = \mu_0$, $\varepsilon_1 = \varepsilon_{r1}\varepsilon_0$, $\varepsilon_2 = \varepsilon_{r2}\varepsilon_0$, $\varepsilon_3 = \varepsilon_{r3}\varepsilon_0$, with μ_0 and ε_0 as the free space permeability and permittivity respectively. The wave numbers are: $k_1 = k_0\sqrt{\varepsilon}_{r1}$, $k_2 = k_0\sqrt{\varepsilon}_{r2}$, and $k_3 = k_0\sqrt{\varepsilon}_{r3}$, where $k_0 = 2\pi f\sqrt{\mu_0\varepsilon_0}$ and f is a frequency of interest. In general transmission lines problems we can assume the wave propagation directions are same as the normal direction of the interfaces, which are the z direction. It is notable that different approaches can be used to describe the phenomena of the multiple reflections of transmission lines. In this section, we calculate the effects of the multiple reflections in term of varying dielectric materials in transmission lines. In the appendix, we show another way to evaluate the effects through using impedance discontinuities. Note that the analysis in terms of impedance will give same prediction if the loss/attenuation and phase is taken into account.

The total reflection coefficient from Region 1 to Region 2 for TM wave is [3]

$$R^{total} = \frac{R_{12} + R_{23}\exp(-2jk_2d)}{1 + R_{12}R_{23}\exp(-2jk_2d)} \tag{1}$$

And the total transmission coefficient from Region 1 to Region 3 for TM wave is

$$T^{total} = \frac{T_{12}T_{23}\exp(-jk_2d)\exp(jk_3d)}{1 + R_{12}R_{23}\exp(-2jk_2d)} \tag{2}$$

where R_{12} is the reflection coefficient from Region 1 to Region 2, R_{23} the reflection coefficient from Region 2 to Region 3, T_{12} the transmission coefficient from Region 1 to Region 2, and T_{23} the transmission coefficient from Region 2 to Region 3. Using Taylor series expansion, "$1/(1+x) = 1 - x + x^2 - x^3 + \ldots$", and truncating it up to first order, (1) and (2) can be casted to

$$R^{total} = R_{12} + R_{23}T_{12}T_{23}\exp(-2jk_2d) - R_{23}^2T_{12}T_{23}\exp(-4jk_2d) + \ldots \tag{3}$$

and

$$T^{total} = T_{12}T_{23}\exp(-jk_2d)\exp(jk_3d) - T_{12}T_{23}R_{12}R_{23}\exp(-3jk_2d)\exp(jk_3d) + \ldots \tag{4}$$

For the characterization of transmission lines, we represent the transmission lines as uniform and infinitely long in the propagation direction. Thus, we only need R_{12} as the reflection coefficient, and $T_{12}T_{23}\exp(-jk_2d)\exp(jk_3d)$ as the transmission coefficient. All other terms are noise signals, which exist due to the multiple reflections between the two interfaces. These noise signals will impair the accuracy of the measurements for s-parameters of a realistic (of finite length) transmission line. We further illuminate the problem using a ladder diagram showed in Figure 2.

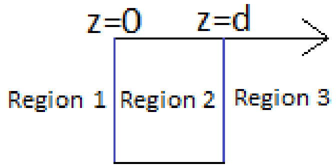


Figure 1: Diagram of a uniform transmission line with a finite length of d. The signal propagates in z direction. Region 1 has ε_1, μ_1, k_1, Region 2 has ε_2, μ_2, k_2, and Region 3 has ε_3, μ_3, k_3.

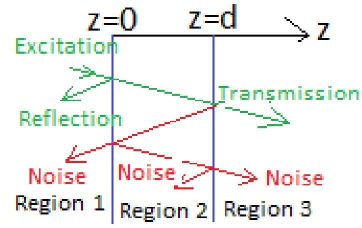


Figure 2: Ladder diagram for the measurement system of a uniform transmission line of finite length, including excitation signal, targeted reflection signal, targeted transmission signal, and noise signals due to the multiple reflections.

2.2. The Effects of Finite Length on the Characterization of Reflection Signals

Next, we will discuss the length effects on the measurements. The single interface reflection and transmission coefficients can be written as

$$R_{12} = \frac{\varepsilon_{r2}k_1 - \varepsilon_{r1}k_2}{\varepsilon_{r2}k_1 + \varepsilon_{r1}k_2} = \frac{\sqrt{\varepsilon}_{r2} - \sqrt{\varepsilon}_{r1}}{\sqrt{\varepsilon}_{r2} + \sqrt{\varepsilon}_{r1}} \tag{5a}$$

$$R_{23} = \frac{\varepsilon_{r3}k_2 - \varepsilon_{r2}k_3}{\varepsilon_{r3}k_2 + \varepsilon_{r2}k_3} = \frac{\sqrt{\varepsilon}_{r3} - \sqrt{\varepsilon}_{r2}}{\sqrt{\varepsilon}_{r3} + \sqrt{\varepsilon}_{r2}} \tag{5b}$$

$$T_{12} = 1 + R_{12} = \frac{2\sqrt{\varepsilon}_{r2}}{\sqrt{\varepsilon}_{r2} + \sqrt{\varepsilon}_{r1}} \tag{5c}$$

$$T_{23} = 1 + R_{23} = \frac{2\sqrt{\varepsilon}_{r3}}{\sqrt{\varepsilon}_{r3} + \sqrt{\varepsilon}_{r2}} \tag{5d}$$

Note, Equations 5(a)–5(d) indicate that the reflection and transmission coefficients are determined by the ratios of $\varepsilon_{r3}/\varepsilon_{r2}$ and $\varepsilon_{r1}/\varepsilon_{r2}$. Thus, in the modeling of multiple reflections, the Region 1 and Region 3 can be simplified as air as reference permittivity, which has $\varepsilon_{r1} = 1$ and $\varepsilon_{r3} = 1$. Then

$$R_{12} = \frac{\sqrt{\varepsilon}_{r2} - 1}{\sqrt{\varepsilon}_{r2} + 1} \tag{6a}$$

$$R_{23} = \frac{1 - \sqrt{\varepsilon}_{r2}}{1 + \sqrt{\varepsilon}_{r2}} = -R_{12} \tag{6b}$$

$$T_{12} = 1 + R_{12} = \frac{2\sqrt{\varepsilon}_{r2}}{\sqrt{\varepsilon}_{r2} + 1} \tag{6c}$$

$$T_{23} = 1 - R_{12} = \frac{2}{1 + \sqrt{\varepsilon}_{r2}} \tag{6d}$$

$$R^{total} = \frac{R_{12} - R_{12}\exp\left(-2jk_0 d\right)}{1 - R_{12}^2 \exp\left(-2jk_2 d\right)} \tag{7}$$

Let's define the reflection signal-to-noise SNR_R, as the ratio of the first noise reflection signal to the targeted reflection signal,

$$SNR_R = \frac{R_{12}}{R_{23}T_{12}T_{23}\exp\left(-2jk_2 d\right)} = -\exp\left(2jk_0\sqrt{\varepsilon}_{r2} d\right) \tag{8}$$

Examining the relative permittivity of region 2

$$\varepsilon_{r2} = {\varepsilon'}_{r2}\left(1 - j\tan\delta_2\right) \tag{9}$$

$$\sqrt{\varepsilon}_{r2} = \sqrt{\varepsilon'_{r2}\left(1 - j\tan\delta_2\right)} = \sqrt{\varepsilon'_{r2}}\sqrt{\left(1 - j\tan\delta_2\right)} \tag{10}$$

Since the loss tangent $\tan\delta_2 \ll 1$ (about $0.005 \sim 0.03$) for typical PCB dielectric materials, rewrite (10) using Taylor series expansion, "$\sqrt{1-x} = 1 - \frac{1}{2}x - \frac{1}{4}x^2 - \ldots$", and truncating it up to first order, we have

$$\sqrt{\varepsilon}_{r2} \approx \sqrt{\varepsilon'}_{r2}\left(1 - \frac{1}{2}j\tan\delta_2\right) \tag{11}$$

Substituting (10) into (8) yields

$$SNR_R = T_{12}T_{23}\exp\left(2jk_0\sqrt{\varepsilon'_{r2}}d\right)\exp\left(k_0\sqrt{\varepsilon'_{r2}}d\tan\delta_2\right) \tag{12}$$

The free space wave number $k_0 = \frac{2\pi}{\lambda_0}$, λ_0 is the free space wavelength. Then

$$SNR_R = T_{12}T_{23}\exp\left(4\pi j\sqrt{\varepsilon'_{r2}}\frac{d}{\lambda_0}\right)\exp\left(2\pi\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}\right) \tag{13}$$

The ratio of trace length to wavelength $\frac{d}{\lambda_0}$ depends on the frequency. $\frac{d}{\lambda_0}$ is small at low frequency while is large at high frequency. If we write SNR_R in dB scale we arrive at

$$20\log_{10}|SNR_R| = 20\log_{10}\exp\left(2\pi\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}\right) - 20\log_{10}|T_{12}T_{23}| \tag{14}$$

Thus,

$$20\log_{10}|SNR_R| \approx 289\cdot\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0} - 20\log_{10}\left|\frac{4\sqrt{\varepsilon_{r2}}}{\left(\sqrt{\varepsilon_{r2}}+1\right)^2}\right| \tag{15}$$

SNR_R. The term $289\cdot\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}$ is the critical parameter, the loss of the transmission line. It increases with $\frac{d}{\lambda_0}$, (the electrical length of the trace). The term $-20\log_{10}|\frac{4\sqrt{\varepsilon_{r2}}}{(\sqrt{\varepsilon_{r2}}+1)^2}|$ is an inherent error term, a constant, and is due to the multiple reflections at the interfaces. In Table 1, we show the SNR_R for a typical PCB transmission lines of 8 inches and 1 inch at 4 GHz.

Table 1: SNR_R for differential transmission lines at 4 GHz: (a) length of transmission line = 8 inches; (b) length of transmission line = 1 inch.

Dielectric Material Types	ε'_{r2}	$\tan\delta_2$	d	λ_0 at 4 GHz	$289\cdot\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}$	$-20\log_{10}\left\|\frac{4\sqrt{\varepsilon_{r2}}}{(\sqrt{\varepsilon_{r2}}+1)^2}\right\|$	$20\log_{10}\|SNR_R\|$
Regular FR4	4	0.02	8 inches	2.953 inches	31.32 dB	1.02 dB	32.34 dB
Mid-loss	3.7	0.01	8 inches	2.953 inches	15.06 dB	0.91 dB	15.97 dB
Low-loss	3.5	0.005	8 inches	2.953 inches	7.32 dB	0.84 dB	8.16 dB

(a) 8 inches

Dielectric Material Types	ε'_{r2}	$\tan\delta_2$	d	λ_0 at 4 GHz	$289\cdot\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}$	$-20\log_{10}\left\|\frac{4\sqrt{\varepsilon_{r2}}}{(\sqrt{\varepsilon_{r2}}+1)^2}\right\|$	$20\log_{10}\|SNR_R\|$
Regular FR4	4	0.02	1 inches	2.953 inches	3.91 dB	1.02 dB	4.94 dB
Mid-loss	3.7	0.01	1 inches	2.953 inches	1.88 dB	0.91 dB	2.80 dB
Low-loss	3.5	0.005	1 inches	2.953 inches	0.92 dB	0.84 dB	1.75 dB

(b) 1 inch

2.3. The Effects of Finite Length on the Characterization of Transmission Signals

Similarly, let's define the transmission signal-to-noise SNR_T, as the ratio of the first noise transmission signal to the targeted transmission signal,

$$SNR_T = \frac{T_{12}T_{23}\exp(-jk_2d)\exp(jk_3d)}{-T_{12}T_{23}R_{12}R_{23}\exp(-3jk_2d)\exp(jk_3d)} = \frac{1}{R_{12}^2}\exp\left(2jk_0\sqrt{\varepsilon_{r2}}d\right) \tag{16}$$

thus

$$SNR_T = \frac{1}{R_{12}^2}\exp\left(2jk_0\sqrt{\varepsilon'_{r2}}d\right)\exp\left(k_0\sqrt{\varepsilon'_{r2}}d\tan\delta_2\right) \tag{17}$$

Then

$$SNR_T = \frac{1}{R_{12}^2}\exp\left(4\pi j\sqrt{\varepsilon'_{r2}}\frac{d}{\lambda_0}\right)\exp\left(2\pi\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}\right) \tag{18}$$

Write SNR_T in dB scale

$$20\log_{10}|SNR_T| = 20\log_{10}\exp\left(2\pi\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0}\right) - 40\log_{10}(|R_{12}|) \tag{19}$$

then

$$20\log_{10}|SNR_T| \approx 289\cdot\sqrt{\varepsilon'_{r2}}\tan\delta_2\frac{d}{\lambda_0} - 40\log_{10}\left(\left|\frac{\sqrt{\varepsilon_{r2}}-1}{\sqrt{\varepsilon_{r2}}+1}\right|\right) \tag{20}$$

SNR_T has terms with identical properties as SNR_R: the critical term $289 \cdot \sqrt{\varepsilon'_{r2}} \tan\delta_2 \frac{d}{\lambda_0}$ due to the loss of transmission line and the constant, inherent error term $-40\log_{10}(|\frac{\sqrt{\varepsilon}_{r2}-1}{\sqrt{\varepsilon}_{r2}+1}|)$ due to the multiple reflections at the interfaces. In Table 2, we list the SNR_T for a typical PCB transmission lines of 8 inches and 1 inches at 4 GHz.

Table 2: SNR_T for differential transmission lines at 4 GHz: (a) length of transmission line = 8 inches; (b) length of transmission line = 1 inch.

Dielectric Material Types		ε'_{r2}	$\tan\delta_2$	d	λ_0 at 4 GHz	$289 \cdot \sqrt{\varepsilon'_{r2}} \tan\delta_2 \frac{d}{\lambda_0}$	$-40\log_{10}\left(\left\lvert\frac{\sqrt{\varepsilon}_{r2}-1}{\sqrt{\varepsilon}_{r2}+1}\right\rvert\right)$	$20\log_{10}\lvert SNR_R\rvert$
Regular FR4	FR4	4	0.02	8 inches	2.953 inches	31.32 dB	19.08 dB	50.40 dB
Mid-loss	IS-415	3.7	0.01	8 inches	2.953 inches	15.06 dB	20.02 dB	35.08 dB
Low-loss	IT-150DA	3.5	0.005	8 inches	2.953 inches	7.32 dB	20.72 dB	28.05 dB

(a) 8 inches

Dielectric Material Types		ε'_{r2}	$\tan\delta_2$	d	λ_0 at 4 GHz	$289 \cdot \sqrt{\varepsilon'_{r2}} \tan\delta_2 \frac{d}{\lambda_0}$	$-20\log_{10}\left\lvert\frac{4\sqrt{\varepsilon}_{r2}}{\left(\sqrt{\varepsilon}_{r2}+1\right)^2}\right\rvert$	$20\log_{10}\lvert SNR_R\rvert$
Regular FR4	FR4	4	0.02	1 inches	2.953 inches	3.91 dB	19.08 dB	23.00 dB
Mid-loss	IS-415	3.7	0.01	1 inches	2.953 inches	1.88 dB	20.02 dB	21.90 dB
Low-loss	IT-150DA	3.5	0.005	1 inches	2.953 inches	0.92 dB	20.72 dB	21.64 dB

(b) 1 inch

2.4. Compensation of Two Transmission Losses for Insertion Loss Measurements

Note that, in the measurement of single ended insertion loss (S_{21}) or differential insertion loss (Sdd21), the loss results S_{21} or Sdd21 need to comprehend the effects of T_{12} and T_{23} when short coupons ($< 8''$) are used. The waveform manipulation used in the current SET2DIL method only removes internal reflections, but does not compensate the two transmission losses, which is critical for shorter length coupons. As shown in Figure 1, because of the impedance mismatch, the transmission signal loses energy when propagating through the two interfaces: (1) the first interface from reference media (Region 1) to transmission line (Region 2), (2) the second interface from transmission line (Region 2) to reference media (Region 3).

Assuming the transmission line has the differential impedance Z_d, the two reference media have the same differential impedance Z_0, the total transmission coefficients $= T_1 * T_2 = (2 * Z_d/(Z_d + Z_0)) * (2 * Z_0/(Z_d + Z_0))$. The final SDD21 needs to be re-calculated as:

$$\text{Final SDD21} = [\text{SET2DIL SDD21 obtained with waveform manipulation}]/(T_1 * T_2). \tag{21}$$

In dB scale, it is:

$$\text{Final SDD21\_dB} = [\text{SET2DIL SDD21\_dB}] - 20 * \log 10(T_1 * T_2) \tag{22}$$

Normalized to unit inch length, then it is

$$\text{Final SDD21\_dB/inch} = [\text{SET2DIL SDD21\_dB/inch}] - 20 * \log 10(T_1 * T_2)/(\text{\# of inches}) \tag{23}$$

Note, the term of $20 * \log 10(T_1 * T_2)/(\text{\# of inches})$ is relatively small (compared to SDD21) for longer coupons (hence the choice of $8''$ for the original coupons), but it is significant for short coupons. For example, if $Z_d = 80$ ohms and $Z_0 = 100$ ohms, then $-20 * \log 10(T_1 * T_2) = 0.108$ dB. Then using different lengths of coupons, we have different amount of loss per inch, as shown in the following Table 3. We see it is only 0.013 dB/inch for 8 inch length coupons, however, it becomes 0.216 dB/inch for 0.5 inch coupons. It should be pointed out that this term is constant for all frequencies.

With the waveform manipulation for removing the internal reflections and the correction term to compensate the two transmission losses, weobtain the simulation results of 2-port SET2DILtime domain measurement and 4-port traditional scope time domain measurement, with 0.5 inch coupons.

They are in good agreement with all the simulation results of VNA frequency domain measurement results with three different lengths, 8 inches, 4 inches, and 0.5 inches. These results will be presented in Section 3.

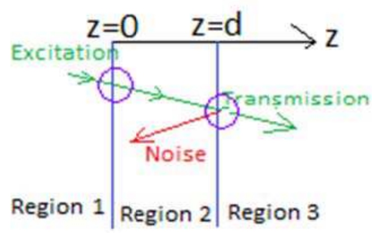


Figure 3: Diagram to show the two transmission losses, which are due to the two discontinuities (shown as two pink circles) in the interfaces between transmission line (Region 2) and reference media (Region 1 and Region 3).

Thus, for time domain measurements, either 2-port SET2DIL, or 4-port traditional scope, we need to do two corrections: to manipulate waveform for removing the internal reflections, and to compensate the two transmission losses. With these two corrections, we can use coupons much shorter than 8 inches to accurately measure the SDD21. In the Appendix, we also use simple circuit formula to further explain the error calculation.

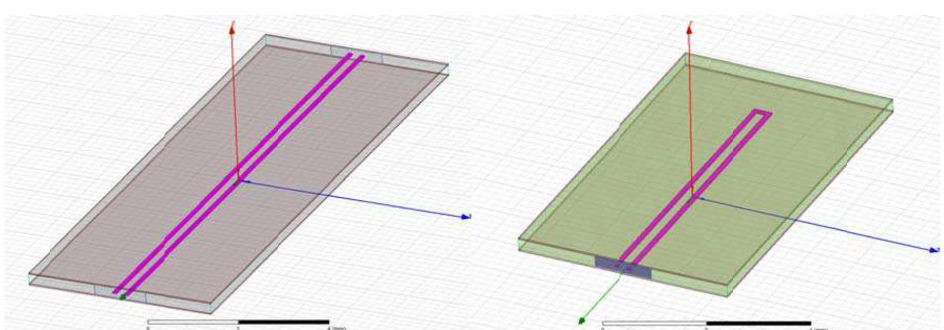
Table 3: $-20 * \log 10(T_1 * T_2)/(\#\text{ of inches})$, for different coupon lengths, assuming $Z_d = 80$ ohms and $Z_0 = 100$ ohms.

	8 inches	4 inches	2 inches	1 inches	0.5 inches
$-20 * \log 10(T_1 * T_2)/(\#\text{ of inches})$	0.013 dB	0.027 dB	0.054 dB	0.108 dB	0.216 dB

3. SIMULATIONS

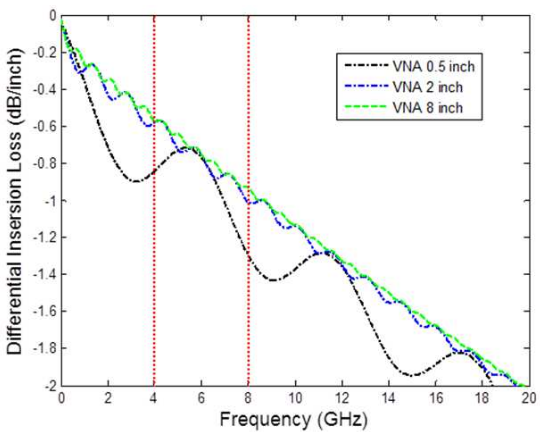
Two different simulations, frequency domain simulation and time domain simulation, are performed to verify the proposed method for frequency domain measurement and time domain measurement, respectively.

In the frequency domain simulation, we use 3D full wave simulator HFSS to directly solve the S-parameters for two coupons, as showed in Figure 4. Figure 4(a) is a 4-port coupon, which is used in in characterization of differential transmission line. In the simulation, we used three different lengths; 0.5 inch, 2 inches, and 8 inches. Figure 4(b) is a 2-port SET2DIL coupon, which can be used for charactering differential loss. The simulation results are showed in Figure 5(a). The label "VNA" (Vector Network Analyzer) refers to results obtained directly from frequency domain simulation. Note, appropriate de-embedding have been done to the 0.5 inch and 2 inches S-parameters in Figure 5(a).

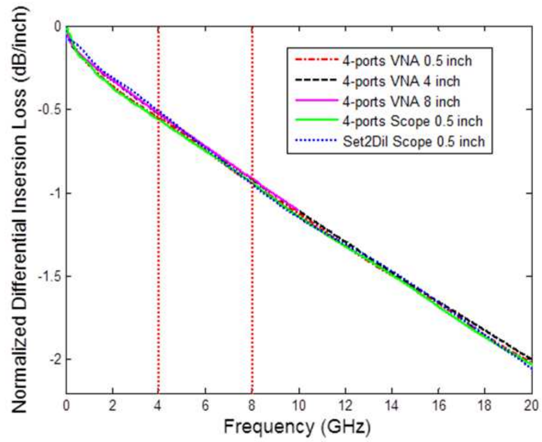


(a) (b)

Figure 4: 3D full wave simulation using HFSS: (a) a traditional 4-port coupon for characterizing two adjacent transmission lines; (b) SET2DIL coupon with 2 ports for characterizing a differential transmission line pair.



(a)



(b)

Figure 5: (a) Original Differential Insertion Loss obtained using frequency domain HFSS simulations. VNA" refers to the frequency results; (b) Frequency results obtained by using de-embedding(black, blue, and green curves), and time domain results obtained by removing the reflection signals. "Scope" (Oscilloscope) refers to the time domain results, while "VNA" refers to the frequency results.

In the time domain simulation, we obtain the step responses transformed from the frequency simulation S-parameter results. After removing the reflection signals, we transform the step response back to differential insertion loss in S-parameters. The results are shown as the curves in Figure 5(b). "Scope" (Oscilloscope) refers to the time domain results, while "VNA" refers to the frequency results. We see that time domain and frequency domain simulations are in good agreement. The results support the new approach, a trace as short as 0.5 inch can accurately give differential insertion loss.

4. CONCLUSION

A new approach is proposed to reduce the length requirements for the measurement of transmission line S-parameters in printed circuit boards (PCBs). The method also improves the accuracy of measurements. The proposed approach is implemented in the time and frequency domains. Simulations are performed to verify the **new** approach. Results show that the new approach can reduce the minimum length of high-speed transmission lines from 8 inches in traditional industry standards to only 0.5 inch for both 4-port TDR/VNA measurements and 2-port SET2DIL.

5. APPENDIX

In this section, we use simple circuit formula to estimate the errors due to the impedance discontinuities of first order reflection. In the Figure 6, the reflection coefficient due to the first impedance discontinuity is

$$\Gamma_1 = \frac{\mathrm{Z}_{DUT} - Z_{in}}{\mathrm{Z}_{DUT} + Z_{in}}$$

and transmission coefficient due to the first impedance discontinuity is

$$\mathrm{T}_1 = \frac{2\mathrm{Z}_{DUT}}{\mathrm{Z}_{DUT} + Z_{in}}$$

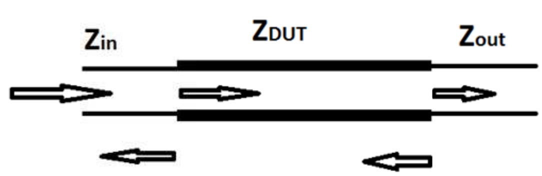


Figure 6: Diagram to show the impedance discontinuities.

And the reflection coefficient due to the second impedance discontinuity is

$$\Gamma_2 = \frac{Z_{out} - Z_{DUT}}{Z_{out} + Z_{DUT}}$$

the transmission coefficient due to the second impedance discontinuity is

$$\begin{aligned} T_2 &= 1 + \Gamma_2 \\ T_2 &= \frac{2Z_{out}}{Z_{DUT} + Z_{out}} \end{aligned}$$

The total error in the insertion loss measurement is

$$Error_T \cong T_1 T_2$$

The total error in the return loss measurement is

$$Error_R \cong -T_1 \Gamma_2 T_1$$

The relative error in the return loss is

$$Error'_R \cong \frac{Error_R}{\Gamma_1}$$

In dB scale,

$$\begin{aligned} 20\log_{10}|Error_T| &\cong 20\log_{10}|T_1 T_2| \\ 20\log_{10}\left|Error'_R\right| &\cong 20\log_{10}\left|\frac{T_1\Gamma_2 T_1}{\Gamma_1}\right| \end{aligned}$$

Numerical results:

1. Assume the interconnect is for single end transmission line, $Z_{DUT} = 40\,\text{ohm}$, $Z_{in} = 50\,\text{ohm}$, and $Z_{out} = 50\,\text{ohm}$, then

$$\begin{aligned} \Gamma_1 &= \frac{80 - 100}{80 + 100} = -0.1111 \\ T_1 &= 1 - 0.0526 = 0.8889 \\ \Gamma_2 &= \frac{50 - 45}{50 + 45} = 0.1111 \\ T_2 &= 1 + 0.0526 = 1.1111 \\ 20\log_{10}|Error_T| &= -0.11\,\text{dB} \\ 20\log_{10}\left|Error'_R\right| &= -2.05\,\text{dB} \end{aligned}$$

1. Assume the interconnect is for differential transmission line, $Z_{DUT} = 80\,\text{ohm}$, $Z_{in} = 100\,\text{ohm}$, and $Z_{out} = 100\,\text{ohm}$, then

$$\begin{aligned} \Gamma_1 &= \frac{80 - 100}{80 + 100} = -0.1111 \\ T_1 &= 1 - 0.0526 = 0.8889 \\ \Gamma_2 &= \frac{50 - 45}{50 + 45} = 0.1111 \\ T_2 &= 1 + 0.0526 = 1.1111 \\ 20\log_{10}|Error_T| &= -0.11\,\text{dB} \\ 20\log_{10}\left|Error'_R\right| &= -2.05\,\text{dB} \end{aligned}$$

Note, for PCB transmission lines, the errors $Error_T$ and $Error'_R$ due to the impedance discontinuities are approximately constant (1) over different frequencies in frequency domain, and (2) for different lengths of DUT transmission lines.

REFERENCES

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