NN-7

Introducing the Through-Line Deembedding Procedure

Michael B. Steer†, Steven B. Goldberg†, Glenn Rinne‡, Paul D. Franzon†, Iwona Turlik‡, and Jeff S. Kasten†

†High Frequency Electronics Laboratory and the Picosecond Digital Systems Laboratory
Department of Electrical and Computer Engineering,
North Carolina State University, Raleigh, NC 27695-7911.

‡ Microelectronics Center of North Carolina,
Research Triangle Park, NC 27709-2889.

ABSTRACT

The through-line (TL) method is introduced to replace the through-reflect-line (TRL) deembedding procedure. TL utilizes measurements of two lengths of line following approximate open-short-line calibration. TL accounts for the frequency dependent characteristic impedance of the line and avoids the periodic glitches inherent to the TRL procedure.

I Introduction

Through-Reflect-Line- (TRL-) related methods of calibration are the most accepted techniques for deembedding fixtures in planar microwave measurement systems [2]. However TRL has two major shortcomings:

1. The measurement uncertainties that result when the phase delay along the line is an integer $n \times 360° \pm 20°$ resulting in glitches in the deembeded response of a DUT. We have determined that these periodic glitches are largely due to the the reflection standard being arbitrary. The phase uncertainty is largely removed if the reflection standard is known precisely or otherwise removed from TRL calibration.

2. The accuracy depends predominantly on the estimate of the characteristic impedance, Z_C, of the line. Estimation of Z_C is usually made using TDR measurements or through careful design of a transmission line to minimize reflection so that Z_C approximates the system measurement impedance Z_{ref}.

The second point is a matter of some confusion. Error in determining Z_C leads to a systematic measurement error. It has been argued several times that without a known lossy impedance reference inserted in the measured line or additional physical insight the characteristic impedance of an embedded transmission line can not be determined from measurements made at the external ports of the test fixtures. Thus, if the estimated characteristic impedance of the line is in error, all deembedded impedance measurements of a DUT will be in error by the same factor [1].

The above problems are addressed via two techniques introduced in this paper. The first technique to be discussed is the through-line deembedding procedure. This new measurement technique can replace the through-reflect-line (TRL) deembedding procedure in many situations and eliminates both of the problems referred to above. Otherwise TL is very similar to TRL. The second technique determines the characteristic impedance of the line from scattering parameter measurements and the measured free-space capacitance of the reference line. In particular the effective permittivity of the line is determined from high frequency asymtotic behavior.

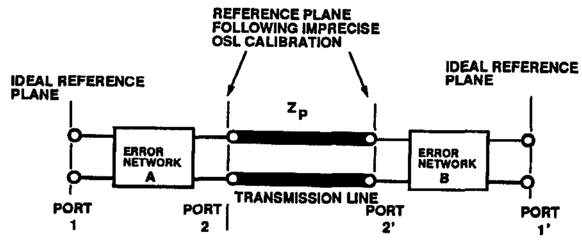


Figure 1: Error networks following approximate OSL calibration.

II Through Line Calibration

Through-line calibration is implemented in three steps:

1. Apply open-short-load (OSL) calibration to each of the two test ports. The OSL calibration need not be precise, the only requirement is that the same reference impedances be presented to the two ports. The test ports will not be precisely calibrated but now the test ports can be modeled as shown in Fig. 1 where the error networks A and B are identical. The error networks may include fixturing parasitics when the fixtures are used with a media, or pad arrangement other than that used during OSL calibration.

2. Perform through and line measurements. Because of symmetry, as explained in the following section, the through measurement yields the reflection coefficient of an ideal short placed at the fixture reference plane, see section II. This removes the need for using the arbitrary reference in the TRL calibration algorithm thus removing the major source of phase ambiguity. This will be mathematically demonstrated in the full paper. We now have sufficient information to implement conventional TRL calibration (with a fixed predetermined Z_C of the reference line) if we so choose.

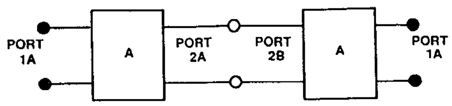
3. Determine the frequency-dependent characteristic impedance of the line used in step 2. We propose a new method for determining the characteristic impedance based on the high-frequency asymptotic behavior of a transmission line and free-space capacitance calculation.

The three steps together yield precise in-situ calibration. Furthermore the glitches associated with TRL deembedding are largely eliminated and this is achieved by implementing

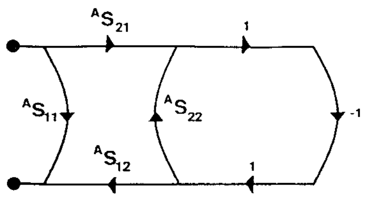
NN

CH3141-9/92/0000-1455$01.00 © 1992 IEEE

1992 IEEE MTT-S Digest



(a)



(b)

Figure 2:
TSL signal flow graphs: (a) for symmetric fixture, and (b) with ideal short circuit placed at fixture port 1.

steps 1 and 2 alone. The combined methods may be implemented using existing calibration substrates. All that is required is the free-space capacitance of the line standards. This is readily obtained once the conductor cross-section is determined. Alternatively, a characteristic impedance versus frequency characteristic could be supplied by the vendor of the calibration substrate.

III Symmetry Applied To The Through Measurement

As a result of approximate OSL calibration the calibrated fixtures of the test system are identical so that in a Through connection the embedded device under test is symmetrical. Symmetry permits synthesis of TRL reflection standards. For the symmetrical through connection, we can calculate the reflection coefficient with the microstrip ports terminated in ideal open or short circuits.

In the signal flow graphs of fig. 2b, we consider the through connection S parameters given by, $S_{11f}(=S_{22f})$ and $S_{21f}(=S_{12f})$, and the actual parameters of the A network being δ, α and γ for $S_{11a}, S_{21a} = S_{12a}$ and S_{22a} respectively. In addition, the input reflection coefficient with an ideal short circuit placed at port 1b of A is

$$\rho_{SC} = \delta - \frac{\alpha^2}{1+\gamma} \tag{1}$$

Using Mason's rule to relate the fixture parameters to the individual network parameters we have results

$$S_{11f} = \delta + \frac{\alpha^2\gamma}{1-\gamma^2} \tag{2}$$

$$S_{21f} = \frac{\alpha^2}{1-\gamma^2} \tag{3}$$

Combining (1), (2) and (3), the short circuit reflection coefficient can be expressed as a function of measured fixture s parameters.

$$\rho_{SC} = S_{11f} - S_{21f} \tag{4}$$

Thus it is possible to "insert" ideal short circuits in a non-insertable medium such as a dielectric waveguide. A similar development holds for an ideal open circuit.

IV Determination of Z_C

Recently several workers [3] – [7] have used non-microwave measurements or additional knowledge to determine Z_C. Kasten and Goldberg et al. [3], [6] calculate the free space capacitance, C_0, of the transmission line and derive Z_C assuming the line resistance, R, to be negligible and the inductance, L, to be frequency independent. Marks and Williams [4] calculate the capacitance, C, of the line and, by assuming that the conductance, G, of the line is negligible, determine Z_0. In a later paper [5] they present two techniques for determining C from measurements. The first of these uses extrapolated low frequency S-parameter measurements and DC resistance measurement to determine the quasi-static line capacitance. (The second technique neglects the difference between Z_C and the measurement reference impedance, Z_{ref}. This is examined in the appendix.) In this paper we present a technique that uses C_0 and S-parameter measurements of two lengths of the line to determine the complex effective permittivity of the line in the skin effect regime. The C_0 calculation is reasonably straight-forward as only the conductor cross-section is required.

The aim of the work presented in this section is to determine the complex characteristic impedance of the transmission line standard used in TRL-type deembedding techniques. The approach taken here is to first neglect R and internal conductor inductance, L_{int}, of the conductors so that an approximate effective relative permittivity, $\hat{\epsilon}_{r,\mathrm{eff}}$, can be determined from the measured propagation constant, γ, of the line. The high frequency asymptote of $\hat{\epsilon}_{r,\mathrm{eff}}$ is then taken as the actual effective permittivity, $\epsilon_{r,\mathrm{eff}}$, at all frequencies. In terms of the per unit length parameters

$$\begin{aligned}\gamma &= \sqrt{\left(R+j\omega(L+L_{\mathrm{int}})\right)(G+j\omega C)} \\ &= j\omega\sqrt{\mu_0\mu_{r,\mathrm{eff}}\,\epsilon_0\epsilon_{r,\mathrm{eff}}} = \frac{j\omega}{c}\sqrt{\mu_{r,\mathrm{eff}}\,\epsilon_{r,\mathrm{eff}}}\end{aligned} \tag{5}$$

and

$$Z_C = \sqrt{\frac{R+j\omega(L+L_{\mathrm{int}})}{G+j\omega C}} = Z_0\sqrt{\frac{\mu_{r,\mathrm{eff}}}{\epsilon_{r,\mathrm{eff}}}} \tag{6}$$

where $\omega = 2\pi f$, $\mu_{r,\mathrm{eff}} = [R+j\omega(L+L_{\mathrm{int}})]/j\omega L$, f is frequency and $\epsilon_{r,\mathrm{eff}}$ are the effective relative permeability and permittivity respectively, and Z_0 is the free space impedance of the line with ideal conductors

$$Z_0 = \frac{1}{C_0 c} \tag{7}$$

L_{int} is asymptotically zero at high frequencies as then the skin effect is fully established. Consequently, if the line is embedded in a nonmagnetic media,

$$\mu_{r,\mathrm{eff}} = \frac{R+j\omega(L+L_{\mathrm{int}})}{j\omega L} \tag{8}$$

and

$$\epsilon_{r,\mathrm{eff}} = \frac{G+j\omega C}{j\omega C_0} \tag{9}$$

L, C, and $G/\omega C$ are relatively independent of frequency but R increases with a $\sqrt{f}$ dependence because of the skin effect. As well $R << \omega L$ at high frequencies since R has a $\sqrt{f}$ dependence so that $\lim_{f\to\infty}\mu_{r,\text{eff}}(f) = 1$. Thus

$$\epsilon_{r,\text{eff}} = \lim_{f\to\infty} \hat{\epsilon}_{r,\text{eff}}(f) \tag{10}$$

where the approximate dielectric constant

$$\hat{\epsilon}_{r,\text{eff}}(f) = -\gamma^2(f)c^2/\omega^2 \tag{11}$$

is obtained by substituting $\mu_{r,\text{eff}} = 1$ in (5) and rearranging. In γ is the measured propagation constant and is a by product of the conventional TRL calibration procedure [8]. At lower frequencies L_{int} and R become important so that $\mu_{r,\text{eff}} \neq 1$. In particular, rearranging (5)

$$\mu_{r,\text{eff}}(f) = -\frac{\gamma^2(f)c^2}{\epsilon_{r,\text{eff}}\,\omega^2} \tag{12}$$

Combining (6), (7) and (12) yields, the expression for the characteristic impedance as a function of frequency is

$$Z_C(f) = -j\frac{\gamma(f)}{\epsilon_{r,\text{eff}}\,\omega C_0} \tag{13}$$

IV Measurements

The above method was used in conjunction with a conventional TRL set of measurements, to determine Z_C of a 4 cm. long embedded microstrip line on a thin-film substrate for use in an advanced multichip module. C_0 was found to be 54.5 pF/m using boundary element analysis and the skin depth, δ is half the microstrip thickness at 733 MHz. The dimensions of the line are shown in Fig. 3. From measurements of a through and the 4 in. section of line γ was determined [8] and the approximate dielectric constant $\hat{\epsilon}_{r,\text{eff}}(f)$ evaluated, see Fig. 4. Subsequently Z_C was evaluated, see Fig. 5. It was determined that $\epsilon_{r,\text{eff}} = 2.62$ and the effective dielectric loss tangent was 0.145. (The high loss tangent is attributed to the infusion of metallic particles into the polyimide dielectric.)

The RLGC parameters of the line were calculated using

$$\gamma Z_C = R + \jmath\omega L \tag{14}$$

and

$$\gamma / Z_C = G + \jmath\omega C \tag{15}$$

derived from (5) and (6) obtaining As expected with the relative permittivity fixed, $G/\omega C$ and C were frequency independent with $G = 19$ mS/m at 1 GHz and $C = 101.4$ pF/m. The L and R per unit length parameters are presented in Fig. 6. L falls off with frequency as the skin effect is established and L_{int} shrinks. The R per unit length is extracted with considerable less certainty at all but the lowest frequencies as the dielectric conductance is the dominant loss mechanism so that the phase ambiguity common in TRL deembedding is amplified. Never-the-less R has the expected $\sqrt{f}$ behavior at low frequencies as expected from simple skin effect considerations.

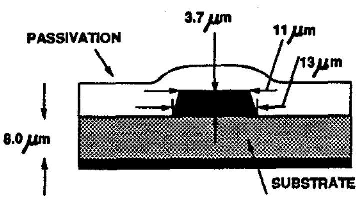


Figure 3: Cross-section of microstrip transmission line.

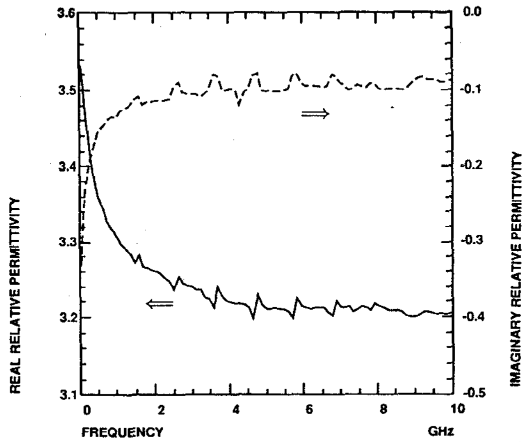


Figure 4: Approximate relative dielectric constant $\hat{\epsilon}_{r,\text{eff}}$ versus frequency.

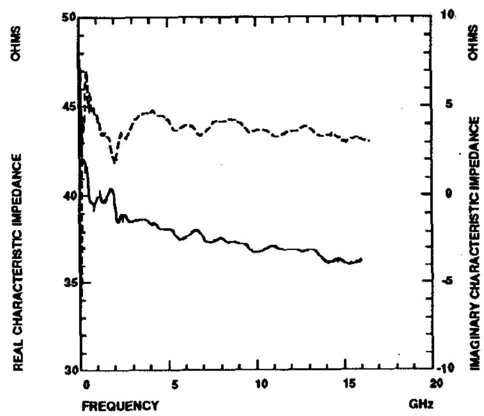


Figure 5: Characteristic impedance versus frequency.

IV Discussion and Conclusion

The above method of determining the characteristic impedance of a line uses the measured propagation constant and high frequency estimation of the effective permittivity. The inherent assumption is that the capacitance of the line is independent of frequency. However, at very high microwave frequencies this capacitance will change as the field distribution changes and more field lines are concentrated in the substrate. Never-the-less, this technique presented here yields the best estimate of the characteristic impedance of the line. This estimate, combined with approximate OSL calibration of the test fixtures and measurement of a through and line yields a calibration procedure which is a viable replacement for TRL.

Acknowledgement

This work was supported in part by the National Science Foundation under Grant MIPS-9017054.

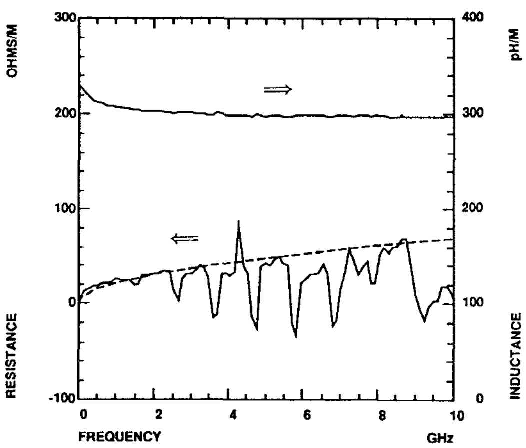


Figure 6: Per unit length resistance and conductance parameters of the transmission line.

References

[1] G.F. Engen and C.A. Hoer, "Thru-reflect-line: an improved technique for calibrating the dual six-port automatic network analyzer," *IEEE Trans. Microwave Theory Tech.*, MTT-27, December 1979, pp. 987-993.

[2] M.B. Steer, S.B. Goldberg and P.D. Franzon, "Comment on an accurate measurement technique for line properties, junction effects and dielectric and magnetic parameters," *IEEE Trans. Microwave Theory Tech.*, Vol. 39, Feb. 1992, pp. 410-411.

[3] J.S. Kasten, M.B. Steer, and R. Pomerleau, "Enhanced through-reflect-line characterization of two-port measuring systems using free-space capacitance calculation," *IEEE Trans. Microwave Theory and Tech.*, Vol. 38, No. 2, Feb. 1990, pp. 215-217.

[4] R.B. Marks and D.F. Williams, "Characteristic impedance determination using propagation constant measurements," *IEEE Microwave and Guided Wave Letters* vol. 1, June 1991, pp. 141-143.

[5] D.F. Williams and R.B. Marks, "Transmission line capacitance measurement," *IEEE Microwave and Guided Wave Letters*, Vol. MTT-1, Sept. 1991, pp. 243-345.

[6] S.B. Goldberg, M.B. Steer, P.D. Franzon and J.S. Kasten, "Experimental electrical characterization of high speed interconnects," *41st Electronic Components and Technology Conference Digest*, May 1991, pp. 85-88.

[7] S.B. Goldberg, M.B. Steer, P.D. Franzon and J.S. Kasten, "Experimental electrical characterization of interconnects and discontinuities in high speed digital systems," *IEEE Trans. on Components Hybrids and Manufacturing Technology*, Dec. 1991, pp. 761-765.

[8] J.P. Mondal and T-H. Chen, "Propagation constant determination in microwave fixture de-embedding procedure," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-36, April 1988, pp. 706-714.

Appendix

The second technique presented in [5] for determining the capacitance of a line neglects the difference between the characteristic impedance of the line, Z_C, and the reference characteristic impedance, Z_{ref}, of the measurement system. Following the development in [5], for a small lumped load resistor, $R_{\text{load,dc}}$, at low frequencies we have

$$Z_{\text{ref}}\left(\frac{1+\Gamma_{\text{load}}}{1-\Gamma_{\text{load}}}\right) \equiv Z_{\text{load}} \approx R_{\text{load,dc}} \quad (16)$$

where Γ_{load} is the measured reflection coefficient of the load referred to Z_{ref}. Substituting (16) in (15)

$$C\left[1-\jmath(G/\omega C)\right] \approx \frac{\gamma}{\jmath\omega\alpha R_{\text{load,dc}}}\left(\frac{1+\Gamma_{\text{load}}}{1-\Gamma_{\text{load}}}\right) \quad (17)$$

where $\alpha = Z_C/Z_{\text{ref}}$ and in general is complex. The important point to note is that Z_C of the line must be known or assumed before C can be calculated. Thus the technique can not be used to determine Z_C. In contrast, the corresponding equation in [5], their (6), does not include α. However the lines examined had Z_C's of approximately 50 Ω so that $\alpha \approx 1$, since $Z_{\text{ref}} = 50\Omega$, and the error is not evident.