Calibration Methods for Time Domain Network Analysis

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***Abstract*—Accurate calibration techniques for the characterization of general one- and two-port networks using Time Domain Reflection/Transmission (TDR/T) measurements are presented. Simple one-port open-short-match corrections formulated in reference [1] are generalized for three arbitrary known loads and extended to the two-port case. Known general frequency-domain techniques are shown to be directly applicable to the time-domain measurements including the use of redundancy to reduce the number of required calibration standards. A time-domain thru-match-short method similar to the well known TRL method [2] is presented. Examples of the measured results for typical one- and two-port devices are included and compared with Vector Network Analyzer measurements to validate the Time Domain Network Analysis algorithms.**

I. Introduction

The DOMAIN NETWORK ANALYSIS (TDNA) has been proposed over the last few decades [3]–[9] but has been slow to gain wide acceptance as evidenced by the lack of a commercially available instrument using the time domain for network characterization.

In the meantime, the frequency-domain vector network analyzer (VNA) has become a sophisticated measurement system with metrology applications. Error correction techniques such as the TRL method [2], the LRL method [10], and the generalized theory for these and associated techniques described in [11]–[12], have greatly refined VNA accuracy. Extensive application of these techniques along with microwave wafer probes and associated calibration substrates has greatly increased VNA capabilities, particularly in the area of semiconductor device characterization.

Previous efforts in TDNA predated the more sophisticated TDR systems that have recently become available. These systems use equivalent time sampling and have bandwidths that rival VNA systems. Recent work with nonlinear transmission line based samplers [13] has obtained 150 GHz TDR bandwidth. Improved multiple port sampling heads and multiple sampling head systems that are now available should lead to true multi-port calibrations and measurements. In addition, instrument and controller processing capabilities make possible rapid signal acquisition, translation, and display in frequency- as well as time-domain form.

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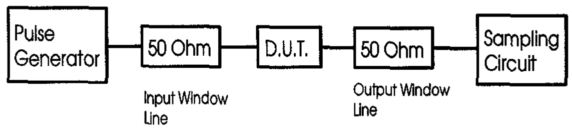

Fig. 1. Previously published calibration techniques used 50 ohm "window" lines to isolate the D.U.T.

TDNA has unique characteristic features that provide a useful alternative to VNA particularly at higher frequencies. TDNA systems are less complex than the normal VNA system, with only a portion of the system using microwave components. A combined pulse source and sampling head can be located remotely from the TDNA system with all interconnects being either low frequency or slow logic. It is conceivable that this portion could be integrated with a microwave probe, making extremely high frequency measurements possible since the cable and connector losses, as well as the losses associated with the VNA test set, are eliminated. Close proximity can be maintained between the sampling circuitry and the device under test (d.u.t.).

Most of the calibration techniques used for TDNA rely on long, matched transmission lines to isolate the d.u.t. [3], [4], [8], as shown in Fig. 1. The time-domain response is truncated to select only the portion of the response due to the d.u.t. interacting solely with these 'window' lines. Outside of this window the response includes the effects of the terminations associated with the pulse generator source, and sampling circuit load leading to erroneous, inaccurate results.

An error correction procedure for one-port TDNA measurements was described in [1]. Here the source-interconnect-sampling system was modeled as shown in Fig. 2. The associated (frequency-domain) flow diagram allows for an unknown excitation, source reflection coefficient and unknown interconnection two-port S-parameters. Equations (3)–(5) of [1] allow complete determination of an unknown ρ_L when the measurement from the unknown load is combined with measurements from ideal short, match, and known non-ideal open terminations.

In this paper, a reduced error model for the TDNA system is obtained by using network transformations. From this model the procedure given in [1] is generalized for any set of three known terminations. This error model is then extended to the important case of two-port measurements and it is shown that general VNA calibration techniques apply directly in TDNA, the solution for the specific case of Thru-Match-Short (TMS) calibration is presented as an example. TMS is a natural set for

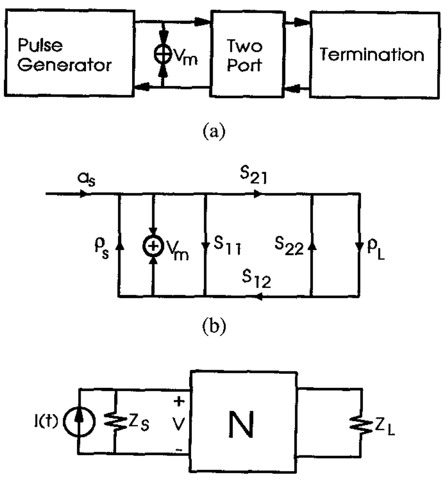

Fig. 2. Model of the one-port measurement system from [1]: (a) Block diagram. (b) Signal flow graph. (c) Schematic.

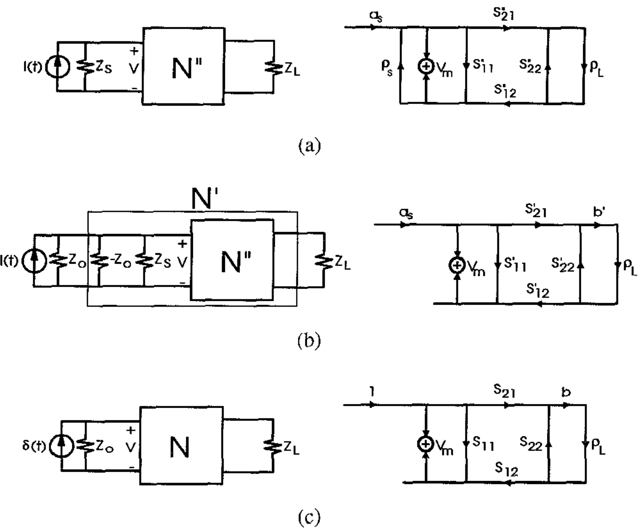

Fig. 3. Variable elimination by network transformation: (a) The original network. (b) Fictitious positive and negative reference impedances are added in shunt creating a new unknown network N'. (c) Source waveshape absorbed into final network N.

the TDNA application as the match may be implemented using a reference transmission line that is as long as the desired time response measurement. This TMS method is equivalent to the VNA TRL method of [2] in its use of the line impedance as reference; however, complete knowledge of the reflect standard is required.

These calibration methods are practical and example measurements are presented. Frequency-domain results obtained from the measured time-domain data are compared with VNA measurements for validation.

II. Reduction of Variables by Network Transformation

The system of Fig. 2 is repeated in Fig. 3(a) with the original two-port re-labeled as N''. The source is a Norton current source with unknown waveshape and unknown shunt impedance. By adding both positive and negative reference shunt impedances (Z_o), the circuit is not affected and can be rearranged as shown in Fig. 3(b) and a new unknown Network N' identified. In the flow diagram the previously unknown ρ_S has been absorbed into N'.

The source wave shape is also a candidate for variable elimination. The source is causal and is ideally a step but for convenience an impulse (unity in the frequency-domain) will be assumed. The transformation must not change the measured voltage, $V = V'$, the impedance looking back toward the source, $S_{22} = S'_{22}$, or the wave incident on the load, $b = b'$. Using these relations the transformation is readily found from the flow diagrams, but it is sufficient to know that such a transformation exists. The flow diagram and schematic of the transformed source system are shown in Fig. 3(c).

All of the system non-idealities are grouped into a single two-port network N. The number of complex unknowns in the original system of [1] shown in Fig. 2 has been reduced from six to four. In VNA calibration terminology the network N is an 'error box,' and the voltage measurement corresponds to an ideal (corrected) system limited only by the repeatability and precision of the knowledge of the error box terms.

III. One-Port Calibration

The simplified system can now be easily solved for the general solution to the one-port calibration. Using an FFT method suitable for step-like waveforms, (e.g., [14]) TDR measurements are converted to frequency-domain voltages, V_x. The voltage curve measurements (V_u, V_1, V_2, V_3) from the unknown and the known terminations (ρ_u and ρ_1, ρ_2, ρ_3) allow determination of S_{22} which is used to solve for ρ_u.

Using the flow diagram of Fig. 3(c) the observed frequency-domain voltage V_x for a termination ρ_x (where x is u, 1, 2, or 3) is

$$V_x = 1 + S_{11} + \frac{S_{12}S_{21}\rho_x}{1 - S_{22}\rho_x}. \tag{1}$$

The difference between two such measurements,

$$V_{xy} \equiv V_x - V_y = S_{12}S_{21}\frac{\rho_x - \rho_y}{(1 - S_{22}\rho_x)(1 - S_{22}\rho_y)}, \tag{2}$$

eliminates the S_{11} terms. Dividing two independent differences eliminates the $S_{12}S_{21}$ product, or

$$M_1 \equiv \frac{V_{13}}{V_{23}} = \frac{(\rho_1 - \rho_3)(1 - S_{22}\rho_2)(1 - S_{22}\rho_3)}{(\rho_2 - \rho_3)(1 - S_{22}\rho_1)(1 - S_{22}\rho_3)}. \tag{3}$$

Letting

$$\alpha \equiv \frac{\rho_1 - \rho_3}{\rho_2 - \rho_3}, \tag{4}$$

then

$$S_{22} = \frac{M_1 - \alpha}{\rho_1 M_1 - \rho_2 \alpha}, \tag{5}$$

and

$$\rho_u = \frac{M_u(\rho_2 - \rho_3) + \rho_3(1 - S_{22}\rho_2)}{M_u(\rho_2 - \rho_3)S_{22} + (1 - S_{22}\rho_2)}, \tag{6}$$

with

$$M_u = \frac{V_{u3}}{V_{23}}. \tag{7}$$

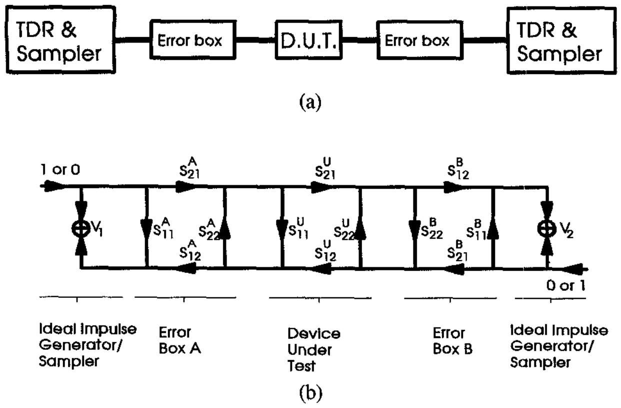

Fig. 4. The two-port Time Domain Network Analysis System: (a) Block diagram. (b) Signal flow diagram.

For the open-short-match case that is considered in [1], $\rho_1 = \rho_o, \rho_2 = -1$, and $\rho_3 = 0$, and (6) reaches to the same results as in [1].

$$\rho_u = -\frac{M_u}{1 - S_{22}(M_u - 1)} \tag{8}$$

with

$$S_{22} = \frac{\frac{M_o}{\rho_o} + 1}{M_0 - 1}. \tag{9}$$

IV. Two-Port Calibration

The two-port system is represented by the block diagram and flow diagram of Fig. 4. The elimination of the source impedance and waveshape variables allows the determination of the S-parameters of the cascaded error boxes and d.u.t. from the FFT of the voltage measurements for each port excited in turn. This information allows the use of any of the well known VNA calibration techniques. A generalization of these techniques and a variety of possible methods have been reported by Eul and Schiek [11]. In the derivation that follows, V_{pe}^N denotes the FFT of the voltage measured at port p when port e is excited and a device N is between the error boxes.

Performing the one-port calibration described in the previous section on both ports provides S_{22}^A and S_{22}^B as well as ρ_A and ρ_B which are the reflection coefficients at ports 1 and 2 respectively in the environment of the measurement system including the error boxes. That is,

$$\begin{aligned} \rho_A &= S_{11}^U + \frac{S_{12}^U S_{21}^U S_{22}^B}{1 - S_{22}^U S_{22}^B} \\ \rho_B &= S_{22}^U + \frac{S_{12}^U S_{21}^U S_{22}^A}{1 - S_{11}^U S_{22}^A} \end{aligned} \tag{10}$$

are obtainable using (6) and (7) [or (8) and (9)].

Two additional equations are needed to determine all four S -parameters, these are obtained from the forward and reverse transmission terms along with the through calibration. The through terms are

$$\begin{aligned} V_{21}^T &= \frac{S_{21}^A S_{12}^B}{1 - S_{22}^A S_{22}^B} \\ V_{12}^T &= \frac{S_{21}^B S_{12}^A}{1 - S_{22}^A S_{22}^B}, \end{aligned} \tag{11}$$

while the measured unknown transmission voltages are

$$\begin{aligned} V_{21}^U &= S_{12}^B \frac{S_{21}^U}{1 - S_{22}^U S_{22}^B} \frac{S_{21}^A}{1 - S_{22}^A \rho_A} \\ V_{12}^U &= S_{12}^A \frac{S_{12}^U}{1 - S_{11}^U S_{22}^A} \frac{S_{21}^B}{1 - S_{22}^B \rho_B}. \end{aligned} \tag{12}$$

From (11) the $S_{21}^A S_{12}^B$ and $S_{21}^B S_{12}^A$ products can be obtained, leaving only the S_{xx}^U terms undetermined in (12).

The solution is obtained by first identifying the known middle terms of (12),

$$\begin{aligned} T_1 &\equiv \frac{V_{21}^U(1 - S_{22}^A \rho_A)}{S_{12}^B S_{21}^A} = \frac{S_{21}^U}{1 - S_{22}^U S_{22}^B} \\ T_2 &\equiv \frac{V_{12}^U(1 - S_{22}^B \rho_B)}{S_{12}^A S_{21}^B} = \frac{S_{12}^U}{1 - S_{11}^U S_{22}^A} \end{aligned} \tag{13}$$

and forming the product

$$T_1 T_2 = \frac{S_{12}^U S_{21}^U}{(1 - S_{22}^U S_{22}^B)(1 - S_{11}^U S_{22}^A)} \tag{14}$$

which allows rewriting (10) as

$$\begin{aligned} \rho_A &= S_{11}^U + T_1 T_2 (1 - S_{11}^U S_{22}^A) S_{22}^B \\ \rho_B &= S_{22}^U + T_1 T_2 (1 - S_{22}^U S_{22}^B) S_{22}^A. \end{aligned} \tag{15}$$

The reflection terms are then extracted as

$$\begin{aligned} S_{11}^U &= \frac{\rho_A - S_{22}^B T_1 T_2}{1 - S_{22}^A S_{22}^B T_1 T_2} \\ S_{22}^U &= \frac{\rho_B - S_{22}^A T_1 T_2}{1 - S_{22}^A S_{22}^B T_1 T_2} \end{aligned} \tag{16}$$

which are used with (13) to obtain the transmission terms

$$\begin{aligned} S_{21}^U &= T_1(1 - S_{22}^U S_{22}^B) \\ S_{12}^U &= T_2(1 - S_{11}^U S_{22}^A). \end{aligned} \tag{17}$$

V. Thru-Match-Short Calibration

Measurement in the time domain has the advantage of natural time windowing since only a finite duration of response is acquired. A match termination is readily obtained with a reference impedance transmission line of sufficient length. This suggests that a likely choice for TDNA calibration could be the line-obtained match as part of a thru-match-short (TMS) calibration using the standards for TRL. Here, however, the reflection coefficient of the short, ρ_S, must be fully known since the line-obtained match behaves like a one-port termination. Unlike the one-port calibration described above, the requirement of a known open has been eliminated.

The open termination is eliminated by using the thru calibration to obtain S_{22}^A and S_{22}^B which are used to obtain ρ_A and ρ_B. The measurements from the thru standard are

$$V_{11}^T = 1 + S_{11}^A + \frac{S_{12}^A S_{21}^A S_{22}^B}{1 - S_{22}^A S_{22}^B}$$
$$V_{22}^T = 1 + S_{11}^B + \frac{S_{12}^B S_{21}^B S_{22}^A}{1 - S_{22}^A S_{22}^B} \tag{18}$$
$$V_{21}^T = \frac{S_{21}^A S_{12}^B}{1 - S_{22}^A S_{22}^B}$$
$$V_{12}^T = \frac{S_{21}^B S_{12}^A}{1 - S_{22}^A S_{22}^B}. \tag{19}$$

The match standard gives

$$V_{11}^M = 1 + S_{11}^A$$
$$V_{22}^M = 1 + S_{11}^B, \tag{20}$$

and the short standard terms are

$$V_{11}^S = 1 + S_{11}^A + \frac{S_{12}^A S_{21}^A \rho_S}{1 - S_{22}^A \rho_S}$$
$$V_{22}^S = 1 + S_{11}^B + \frac{S_{12}^B S_{21}^B \rho_S}{1 - S_{22}^B \rho_S}. \tag{21}$$

The S_{11} terms are eliminated using differences,

$$V_{11}^{TM} \equiv V_{11}^T - V_{11}^M = \frac{S_{12}^A S_{21}^A S_{22}^B}{1 - S_{22}^A S_{22}^B}$$
$$V_{22}^{TM} \equiv V_{22}^T - V_{22}^M = \frac{S_{12}^B S_{21}^B S_{22}^A}{1 - S_{22}^A S_{22}^B} \tag{22}$$
$$V_{11}^{SM} \equiv V_{11}^S - V_{11}^M = \frac{S_{12}^A S_{21}^A \rho_S}{1 - S_{22}^A \rho_S}$$
$$V_{22}^{SM} \equiv V_{22}^S - V_{22}^M = \frac{S_{12}^B S_{21}^B \rho_S}{1 - S_{22}^B \rho_S}. \tag{23}$$

The $S_{22}^A S_{22}^B$ product is obtained by

$$\frac{V_{11}^{TM} V_{22}^{TM}}{V_{21}^T V_{12}^T} = \frac{\left(\frac{S_{12}^A S_{21}^A S_{22}^B}{1 - S_{22}^A S_{22}^B}\right)\left(\frac{S_{12}^B S_{21}^B S_{22}^A}{1 - S_{22}^A S_{22}^B}\right)}{\left(\frac{S_{21}^A S_{12}^B}{1 - S_{22}^A S_{22}^B}\right)\left(\frac{S_{21}^B S_{12}^A}{1 - S_{22}^A S_{22}^B}\right)} = S_{22}^A S_{22}^B. \tag{24}$$

Two useful measurement terms are defined as,

$$M_a \equiv \frac{V_{11}^{SM}}{V_{11}^{TM} - V_{11}^{SM}} = \frac{\rho_S(1 - S_{22}^A S_{22}^B)}{S_{22}^B - \rho_S}$$
$$M_b \equiv \frac{V_{22}^{SM}}{V_{22}^{TM} - V_{22}^{SM}} = \frac{\rho_S(1 - S_{22}^A S_{22}^B)}{S_{22}^A - \rho_S}. \tag{25}$$

The solution for S_{22}^A and S_{22}^B then is readily obtained as

$$S_{22}^A = \rho_S\left(1 + \frac{1 - S_{22}^A S_{22}^B}{M_b}\right)$$
$$S_{22}^B = \rho_S\left(1 + \frac{1 + S_{22}^A S_{22}^B}{M_a}\right). \tag{26}$$

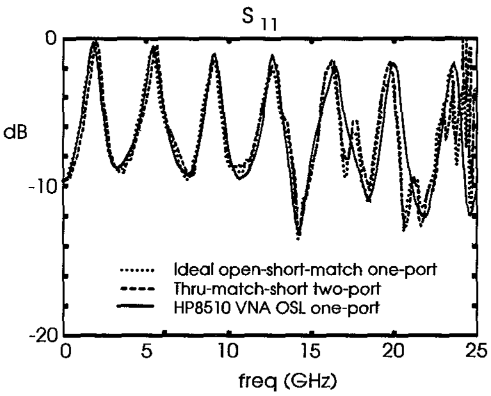

Fig. 5. One-port measurements on a power divider with both ports terminated with short circuits. The one-port TDNA calibration described in [1] and generalized in this discussion and the one-port TDNA result from a two-port T-M-S calibration are shown along with a VNA measurement obtained using standard open-short-load one-port calibration.

Using (6) and (7) with $\rho_2 = \rho_S, \rho_3 = 0$ and a change to the corresponding two-port notation gives

$$\rho_A = \frac{M_u^A \rho_S}{M_u^A \rho_S S_{22}^A + 1 - S_{22}^A \rho_S}$$
$$\rho_B = \frac{M_u^B \rho_S}{M_u^B \rho_S S_{22}^B + 1 - S_{22}^B \rho_S} \tag{27}$$

with

$$M_u^A = \frac{V_{11}^{UM}}{V_{11}^{SM}}$$
$$M_u^B = \frac{V_{22}^{UM}}{V_{22}^{SM}} \tag{28}$$

The solution procedure for the S_{xx}^U terms follows the method described above for the general two-port case.

VI. Experimental Procedure and Results

In order to validate the TDNA procedure described above, one- and two-port measurements were made on a TDR/T system and compared with the results obtained from a frequency-domain VNA system. A measurement system has been developed using a Tektronix 11801 TDR system equipped with an SD-24 20 GHz dual sampling head. For coaxial measurements APC-3.5 calibration standards (characterized reference terminations) from the Hewlett-Packard 8501B vector network analyzer were used. Wafer probe measurements used standards available on a Cascade Microtech calibration tile. VNA measurements using short-open-match-thru calibrations were also taken for comparion with the TDNA results.

Fig. 5 shows the TDNA results from a one-port measurement of a power divider (with both output ports terminated in short circuits) used as a test structure rich in reflections. Curves for calibration assuming ideal open-short-match calibration standards, and one-port results based on the two-port TMS calibration as well as an equivalent VNA measuement are shown.

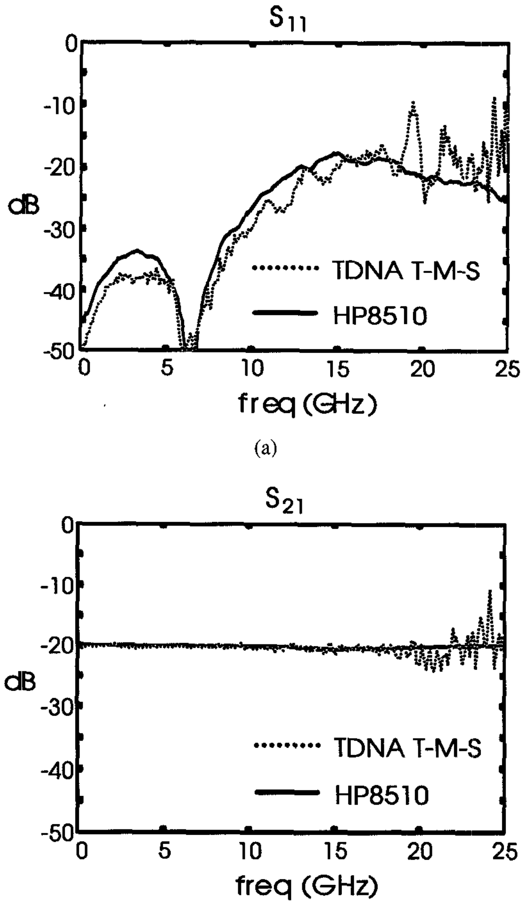

Fig. 6. TDNA and VNA measurements of an SMA 20 dB attenuator. Thru-Match-Short calibration was used with measurements from a Tektronix 11801 TDR system with an SD-24 dual TDR sampling head. Standard Open-Short-Load-Thru calibration was used with the Hewlett-Packard 8510B Vector Network Analyzer.

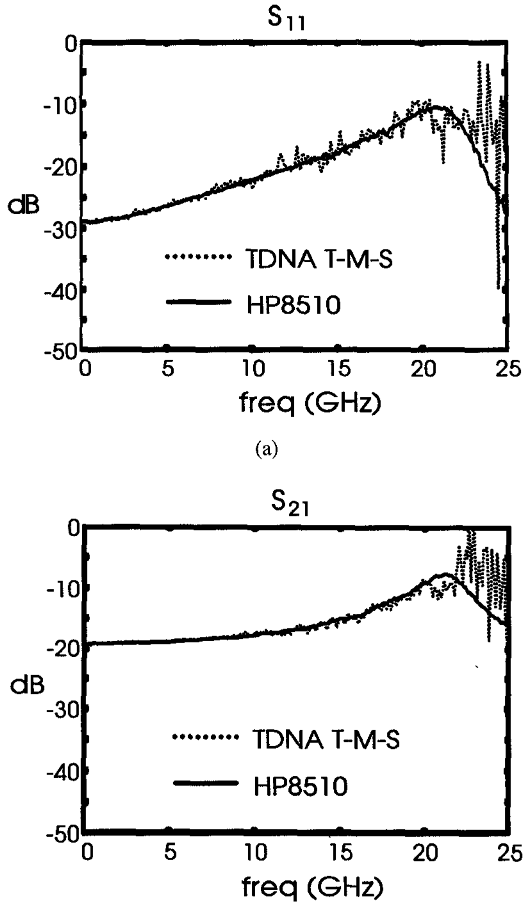

Fig. 7. TDNA and VNA wafer probe measurements of a 20 dB attenuator on a Cascade Microtech calibration tile.

Fig. 6 shows the two-port TDNA and VNA results from an SMA 20 dB attenuator, while Fig. 7 shows the results from measurement of the 20 dB pad on the calibration tile. For clarity only the results for S_{11} and S_{21} are shown, S_{22} and S_{12} are similar. Accurate performance is obtained up to and above the 15 GHz TDR bandwidth and return loss dynamic range exceeds 40 dB (at least at low frequencies) for both cases. Sampling head smoothing and waveform averaging were used to reduce the effects of noise in the measurement system. Trigger drift can result in phase error which becomes significant about 20 GHz, a correction technique for this drift has been described in [15] but has not been used for these results.

VII. Conclusion

Accurate calibration techniques for Time Domain Network Analysis using a reduction of unknowns by network transformation have been demonstrated. The calibration methods are complete and account for both source and load mismatches and pass-thru errors as well as the shape of the excitation waveform. When combined with trigger drift corrections [15], very precise measurements are possible.

The entire spectrum is obtained with a single TDNA measurement and increased frequency resolution requires increasing the number of points recorded and transformed. The overhead associated with this long record length makes TDNA impractical for narrow-bandwidth measurements but well-suited to wide-bandwidth structures such as interconnections and packaging for high speed digital circuits. De-embedded time-domain results obtained from inverse FFT of the results described here can be used for characterization of these passive structures [16].

Recent developments in broadband TDR systems such as the nonlinear transmission line system described in [13] suggest that network analysis in the time domain may become a practical alternative to frequency-domain methods, particularly at higher frequencies where the ability to locate the sampling head and d.u.t. in close proximity may allow greater measurement dynamic range.

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