Multimode TRL—A New Concept in Microwave Measurements: Theory and Experimental Verification

Christophe Seguinot, Patrick Kennis, Jean-Franiços Legier, Fabrice Huret, Erick Paleczny, and Leonard Hayden, *Member, IEEE*

***Abstract*— An original multimode thru-reflection line (TRL) algorithm is used to derive the generalized scattering parameters of multimode two-port networks. Theoretical developments are detailed for calibration procedures based on two ports as well as multiple-port vector network analyzer (VNA) measurements. First-run experimental results demonstrate the validity of this technique. This method allows the experimental characterization of multiconductor transmission-line devices. It could also be used to characterize power coupling to undesired modes in monolithic microwave integrated circuit (MMIC) structures using conductor-backed coplanar waveguides.**

***Index Terms*—Calibration, multiconductor transmission lines, multimode waveguides, scattering parameters, scattering parameters measurement.**

I. Introduction

SCATTERING-PARAMETER measurements have become common and essential for microwave engineers. Vector network analyzers (VNA's) associated with efficient calibration techniques such as the thru-reflection line (TRL) technique [1] offer almost rigorous measurements up to millimeter-wave range. However, conventional calibration techniques are based on a major assumption. Waveguides must propagate only one mode at the measurement reference planes. Most waveguides are designed so that only one dominant mode is present. If some evanescent modes are excited at discontinuities, reference planes are chosen suitably far enough away so that their amplitude can be neglected.

Nevertheless, multiconductor lines used in logic bus [2] or in monolithic microwave integrated circuit (MMIC) technology [3] are designed to drive several signals at the same time. As these lines propagate multiple modes, conventional calibration techniques generally fail. Similar situations are encountered with conductor-backed coplanar waveguides which propagate three dominant modes. One expects that only the symmetric coplanar mode will be excited in such structures. For this reason, coplanar probes are used and transitions are optimized in order to minimize undesired modes. However, it is well known from engineers that these type of circuits sometimes exhibit strange behavior as compared to computer-aided design (CAD) tool predictions. This can arguably be attributed to microstrip-like mode excitation.

In such situations, we need new calibration techniques able to handle multimode waveguides [4]. For conductor-backed coplanar lines, corresponding characterizations would at least allow the optimization of technologies and devices in limiting the influence of undesirable modes. For multiconductor lines, these new techniques could be used to build component libraries for interconnections, bends, crossing lines, via holes, and other discontinuities.

With this in mind, we propose a generalization of the TRL calibration technique to multimode waveguides. In the first part of this paper, we will introduce the basic principle of this approach, which is based on a multiport scattering-parameter measurement. All measurements will be represented by a simple model. Generalized scattering and transfer parameters [5] which are used throughout this paper will be presented. The original multimode TRL algorithm will then be discussed. The theoretical developments will use notation similar to that used in the previous work of Eul and Schiek [1], which may be familiar to engineers involved in microwave calibration techniques. Finally, an experimental verification will be described in order to point out the validity of the multimode TRL approach.

II. Theory

A. Basic Principle of Multimode Calibration

Multimode calibration intends to derive the generalized scattering parameters of multimode waveguide devices. As an example, we will consider the multiple conductor structure device under test (DUT), shown in Fig. 1. In general, this circuit may be a logic bus, coupled microstrip lines, or any coplanar waveguide. It may include crossing, open-end, bend, and other discontinuities, as well as active devices. We define two reference planes for which we assume that only N modes propagate. In order to extract the generalized scattering parameters, we need $2N$ measurement channels, N on each of the two ports. With this in mind, the multimode DUT is connected to a VNA, as indicated in Fig. 2. N single-mode lines are used on each generalized port. For N equal to one, this simply describes the conventional single-mode calibration

Manuscript received October 4, 1996; revised December 1, 1997.

C. Seguinot, P. Kennis, J.-F. Legier, F. Huret, and E. Paleczny are with the Institut d'Electronique et de Microélectronique du Nord (IEMN), Citè scientifique, 59652 Villeneuve d'Ascq cedex, France (e-mail: Seguinot@iemn.univ-lille1.fr).

L. Hayden is with Cascade Microtech, Inc., Beaverton, OR 97005 USA.

Publisher Item Identifier S 0018-9480(98)03167-6.

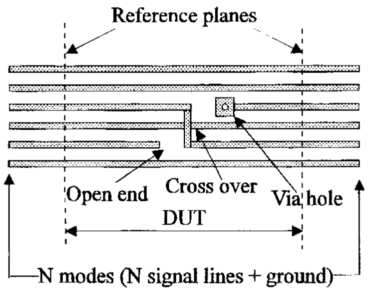


Fig. 1. A multiple-conductor waveguide DUT.

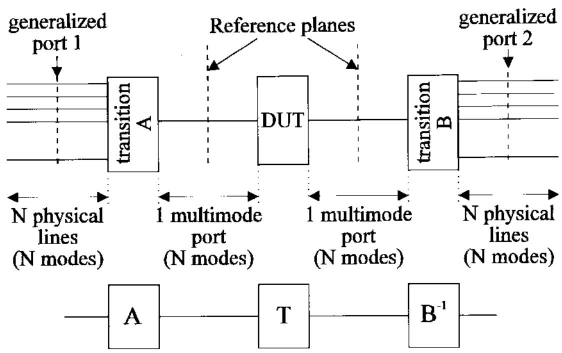


Fig. 2. The basic setup for multimode TRL measurements.

technique. Transitions A and B are quite unusual. They relate one multimode waveguide supporting N modes to N single-mode lines. A possible realization of transitions A and B will be presented in the experimental section. As indicated in Fig. 2, such measurements can be represented by simply cascading transfer matrices. Due to multimode propagation, we will require a generalized transfer and scattering matrices concept.

B. Generalized Transfer and Scattering Matrices

The DUT has two physical ports (1 and 2). Each port has N propagating modes. Such devices are commonly represented as $2N$-ports, each port being associated with one mode (see Fig. 3). Ports denoted $\mathbf{1i}$ are located on physical port 1. More exactly, port $\mathbf{1i}$ refers to the ith mode of generalized port 1. All incident (a_{ji}) and reflected (b_{ji}) modal-wave amplitudes are combined to define the generalized wave vectors A_j and B_j as follows:

$$[A_j] = \begin{bmatrix} a_{j,1} \\ a_{j,2} \\ \vdots \\ a_{j,N} \end{bmatrix} \qquad [B_j] = \begin{bmatrix} b_{j,1} \\ b_{j,2} \\ \vdots \\ b_{j,N} \end{bmatrix}. \tag{1}$$

Throughout this paper, uppercase letters refer to matrices, while corresponding lowercase letters refer to simple scalar matrix coefficients. Generalized waves are used to define the generalized transfer matrices [5]

$$\begin{bmatrix} B_1 \\ A_1 \end{bmatrix} = T \begin{bmatrix} A_2 \\ B_2 \end{bmatrix} = \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} \begin{bmatrix} A_2 \\ B_2 \end{bmatrix}. \tag{2}$$

In this expression, T_{ij} are $(N \times N)$ square-block submatrices which correspond to the DUT generalized transfer coefficients.

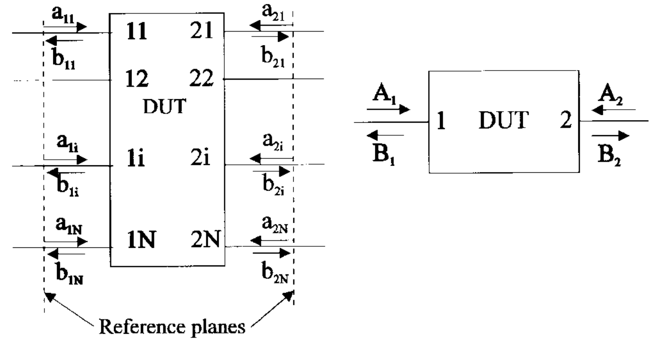


Fig. 3. Representations of a two-port multimode DUT.

In the same way, we define the generalized scattering matrix (S) and coefficients (S_{ij}) with

$$\begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} = S \begin{bmatrix} A_1 \\ A_2 \end{bmatrix}. \tag{3}$$

Using simple linear algebra, S and T matrices are related by

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} S_{12} - S_{11}S_{21}^{-1}S_{22} & S_{11}S_{21}^{-1} \\ -S_{21}^{-1}S_{22} & S_{21}^{-1} \end{bmatrix}$$
$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} T_{12}T_{22}^{-1} & T_{11} - T_{12}T_{22}^{-1}T_{21} \\ T_{22}^{-1} & -T_{22}^{-1}T_{21} \end{bmatrix}. \tag{4}$$

This notation is close to the one used in [1] despite the large number of modes taken into account. In the single-mode case ($N = 1$), all block matrices (generalized coefficients) become single scalar so that these equations simplify to the conventional ones.

As we have built a $2N$-port network, we require measurements of $2N$-port scattering parameters. At this step, we propose two different approaches. Under some reasonable approximations, a calibrated two-port VNA can be used to derive the required $2N$-port scattering parameters. This requires that all ports not connected to the VNA are loaded with perfectly matched terminations. Imperfect loads must be accounted for using an appropriate deembedding algorithm [6]. Moreover, calibration procedures are available to extract multiport S-parameters from multiport VNA measurements [7]. Using these techniques, each measured transfer matrix M is related to the DUT T matrix by

$$M = ATB^{-1} \tag{5}$$

where A and B^{-1} are the transfer matrices of transitions A and B, as described in Fig. 2. In this approach, the two-port VNA is first calibrated. The multimode TRL algorithm is then applied. As the measurement system is calibrated twice, this will be referred to as the "two-step calibration procedure."

Using a multiport VNA (with $2N$ ports) could greatly ease the multimode TRL implementation in getting a "one-step calibration procedure." This is investigated below for the completeness although multiport VNA's are not commonly available. This paper has been organized so that readers who are not interested in the "one-step calibration procedure" can skip Section II-C.

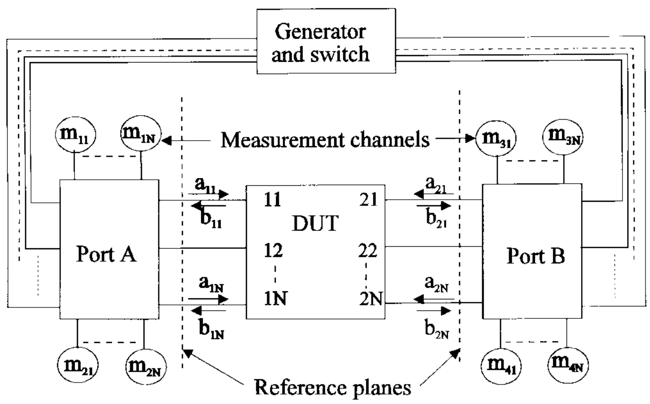


Fig. 4. A general model for multimode TRL measurements.

C. A General Model for Multimode TRL Measurements

The simple model proposed in Fig. 2 and (5) is based on the assumption that at some measurement planes "exact" multiport parameters are available. In this section, we will show the ability of such a simple model to describe all measurement situations involving a multiport VNA corresponding to Fig. 4. This model is a generalization of the single-mode double reflectometer model presented in [1]. The multiport VNA source and loads are successively switched to each port of the VNA. $4N$ measurements channels are used. Transitions A and B represent the linear mapping between measured complex-wave amplitudes (m_{ij}) and reference plane wave amplitudes ($A_i,\ B_i$) as well as waves incoming from the generator and loads. Quantity m_{1i} (respectively, m_{2i}) may be, for example, reflected (respectively, incident) complex-wave amplitudes measured on each single-mode input line (port 1 in Fig. 2). We define measurements column vectors by

$$\left[\mathrm{Meas}_i^{(k)}\right]^t = \left[m_{i,1}^{(k)} \quad m_{i,2}^{(k)} \quad m_{i,N}^{(k)}\right], \qquad i = 1, \cdots, 4 \quad (6)$$

where the superscript (k) refers to one possible configuration of the switch. As in [1], it can be shown that measured values are only dependent on reference plane waves

$$\begin{bmatrix}\mathrm{Meas}_1^{(k)} \\ \mathrm{Meas}_2^{(k)}\end{bmatrix} = A \begin{bmatrix} B_1^{(k)} \\ A_1^{(k)} \end{bmatrix} \quad \begin{bmatrix}\mathrm{Meas}_3^{(k)} \\ \mathrm{Meas}_4^{(k)}\end{bmatrix} = B \begin{bmatrix} A_2^{(k)} \\ B_2^{(k)} \end{bmatrix} \quad (7)$$

where A and B are unknown $(2N \times 2N)$ matrices accounting for all imperfections of the VNA such as imperfect load, couplers, etc. Transitions from the VNA's ports and the reference planes are also taken into account by (7). The only assumptions which have been made concern the linearity of detectors and the repeatability of connections [1]. According to this model, each measurement must satisfy the equation

$$\begin{bmatrix}\mathrm{Meas}_1^{(k)} \\ \mathrm{Meas}_2^{(k)}\end{bmatrix} = ATB^{-1} \begin{bmatrix}\mathrm{Meas}_3^{(k)} \\ \mathrm{Meas}_4^{(k)}\end{bmatrix}. \quad (8)$$

where T is the $(2N \times 2N)$ generalized transfer matrix of the DUT. Each configuration of the switch [referenced by the superscript (k)] gives one set of measurements. Combining $2N$ sets of measurements gives a $(2N \times 2N)$ matrix M, which will be called the measurements matrix

$$\mathrm{Meas}_i = \left[\mathrm{Meas}_i^{(1)} \quad \mathrm{Meas}_i^{(2)} \quad \cdots \quad \mathrm{Meas}_i^{(2^N)}\right], \qquad i = 1 \text{ to } N$$

$$M = \begin{bmatrix}\mathrm{Meas}_1 \\ \mathrm{Meas}_2\end{bmatrix} \begin{bmatrix}\mathrm{Meas}_3 \\ \mathrm{Meas}_4\end{bmatrix}^{-1}. \quad (9)$$

Using (7) and (8), matrix M is simply related to A and B by the same equation as (5). Thus, the same TRL algorithm will be developed for "one-step" as well as "two-step" calibration procedures.

D. Multimode TRL Algorithm

Starting from (5), each DUT of generalized transfer matrix N_x will lead to the following measurements matrix M_x:

$$M_x = AN_xB^{-1} \qquad N_x = A^{-1}M_xB. \quad (10)$$

We start calibration by measuring thru line (T) and line (L) standards. This choice is justified by the fact that only these components parameters are fully or partly known. This is especially true for multimode waveguides for which simple assumptions seem unrealistic. A line may be used instead of the thru line, as indicated by Eul and Schiek [1]. In our case, transfer parameters of thru (N_1) and line (N_2) standards are given by

$$N_1 = \begin{bmatrix} I & 0 \\ 0 & I \end{bmatrix}$$

$$N_2 = \mathrm{diag}\,[\exp(-\gamma_1 l), \cdots, \exp(-\gamma_N l), \exp(+\gamma_1 l), \cdots, \exp(+\gamma_N l),] \quad (11)$$

where I (respectively, 0) denotes an $(N \times N)$ identity (respectively, null) matrix. γ_i is the ith-mode complex propagation constant and l is the line length. As in [1], we define two similar matrices P and Q as follows:

$$P = N_2N_1^{-1} \quad \text{and} \quad Q = M_2M_1^{-1} \quad \text{with} \quad P = A^{-1}QA. \quad (12)$$

We use the fact that similar matrices have identical eigenvalues to define an eigenvalue matrix Λ as follows:

$$\mathrm{eig}(P) = \mathrm{eig}(Q) = \lambda_i \quad \Lambda = \mathrm{diag}(\lambda_i), \qquad i = 1, N. \quad (13)$$

Since P is equal to N_2, N_2 and Λ have identical eigenvalues. This is used to derive the unknown propagation constants of the line

$$\exp(-\gamma_i l) = \lambda_i. \quad (14)$$

Each mode is associated with two opposite propagation constants and a pair of inverse eigenvalues. Solving (14) may lead to numerical difficulties, as in the conventional TRL. The phase shift introduced by the line must be different from multiples of one half-wavelength. However, as this criteria can be fulfilled for one mode at one given frequency, it can fail for other modes with different propagation constants. We hope that multiple line standards can be used [8] in order to overcome such problems and to improve the bandwidth and accuracy of the calibration, but this will not be investigated here. It is important to note that all propagation constants which are calculated by (14) must be correctly ordered in Λ so that N first (respectively, last) eigenvalues are related to N incident (reflected) waves.

In fact, the algorithm which computes the $2N$ eigenvalues (13) cannot attribute a pair of eigenvalues to their specific existent modes (e.g., even and odd modes). In our algorithm, we have proceeded as follows. First, we identify N pairs of opposite eigenvalues. Each pair corresponds to one incident and the corresponding reflected mode. Then, the N eigenvalues attributed to incident modes are sorted in ascending order. Therefore, the "first" mode (mode one) appears as the highest velocity mode. If the frequency band is limited, the phase shift does not exceed 180° so that the algorithm does not need any prior estimate of the mode's propagation constants. For highest frequencies, an estimate is required and can be provided by interpolating lower frequencies values.

Up to this step, our notation is very close to that used by Eul and Schiek [1]. The derivation of propagation constants has been extended to multimode waveguide structures. As we are only concerned with thru-line-based calibration, we will now solve the problem in a different way from [1]. We use the following relation between P, Q, and Λ:

$$\Lambda = X^{-1}PX = Y^{-1}QY \tag{15}$$

where the columns of X (respectively, Y) are composed of the eigenvectors of P (respectively, Q). Since eigenvectors are known except for an arbitrary constant, we can write X and Y as

$$\begin{aligned} X &= X_0\beta, \qquad \beta = \mathrm{diag}(\beta_i) \\ Y &= Y_0\delta, \qquad \delta = \mathrm{diag}(\delta_i), \qquad i = 1, N. \end{aligned} \tag{16}$$

Since P and Q are fully known, X_0 and Y_0 can be computed from (15) while β and δ remain, for the moment, unknown. As P is diagonal, one possible solution for X_0 is a unit matrix ($X_0 = I$). Taking into account the relation between XY and A

$$P = XY^{-1}QYX^{-1} = A^{-1}QA. \tag{17}$$

We use (16) to partly determine A as follows:

$$\begin{aligned} A &= YX^{-1} = Y_0KX_0^{-1} = A_0K \\ K &= \mathrm{diag}(\beta_i/\delta_i) = \begin{bmatrix} K_1 & 0 \\ 0 & K_2 \end{bmatrix} \end{aligned} \tag{18}$$

where A_0 is fully known while K remains unknown. The measurement of the thru is also used to partly derive B as follows:

$$B = M_1^{-1}AN_1 = M_1^{-1}A_0K = B_0K. \tag{19}$$

At this step, the T and L standard measurements yield a partial determination of A and B while $2N$ unknowns in K remain to be solved. Any additional measurement of lines would lead to the same result. We need additional standard measurements to proceed with the calibration. We use a nontransmitting third standard (R). This one-port is characterized by its unknown generalized reflection coefficient denoted Γ ($N \times N$ matrix), which is defined in the same way as the generalized reflection coefficient S_{11} in (3). This generalized one-port (R), which is connected at both reference planes, is related to reference plane waves by

$$B_1 = \Gamma A_1 \qquad B_2 = \Gamma A_2. \tag{20}$$

Corresponding measurements on generalized port 1 and 2 (Fig. 2) yield two measured reflection coefficients denoted Γ_1 and Γ_2 defined and related to Γ by

$$\begin{aligned} \mathrm{Meas}_1 &= \Gamma_1 \mathrm{Meas}_2 \\ \mathrm{Meas}_4 &= \Gamma_2 \mathrm{Meas}_3 \end{aligned} \tag{21a}$$

$$\begin{aligned} \Gamma_1 &= (A_{11}\Gamma + A_{12})(A_{21}\Gamma + A_{22})^{-1} \\ \Gamma_2 &= (B_{22}\Gamma + B_{21})(B_{12}\Gamma + B_{11})^{-1}. \end{aligned} \tag{21b}$$

Equation (21a) is used only if a "one-step calibration procedure" is performed. In this case, the Meas_i are defined similarly as in (9), except that we only need N sets of measurements (instead of $2N$) to derive a reflection coefficient. Using (18) and (19), it can readily be shown that (21b) can be rewritten as

$$\Gamma = K_2^{-1}\hat{\Gamma}_2K_1 = K_1^{-1}\hat{\Gamma}_1K_2 \tag{22}$$

where the nonnormalized reflection coefficients $\hat{\Gamma}_1$ and $\hat{\Gamma}_2$ are known and defined by

$$\begin{aligned} \hat{\Gamma}_1 &= [A_{0,11} - \Gamma_1 A_{0,21}]^{-1}[\Gamma_1 A_{0,22} - A_{0,12}] \\ \hat{\Gamma}_2 &= [B_{0,22} - \Gamma_2 B_{0,12}]^{-1}[\Gamma_2 B_{0,11} - B_{0,21}]. \end{aligned} \tag{23}$$

In these expressions $A_{0,ij}$, $B_{0,ij}$ denote ($N \times N$) submatrices of A_0 and B_0, as in (2) and (3). Solving the right side of (22) yields

$$\begin{aligned} K_2K_1^{-1}\hat{\Gamma}_1K_2K_1^{-1} &= L\hat{\Gamma}_1L^{-1} = \hat{\Gamma}_2 \\ l_il_j\hat{\Gamma}_{1,ij} &= \hat{\Gamma}_{2,ij}. \end{aligned} \tag{24}$$

Where l_i denotes the ith term of diagonal ($N \times N$), matrix L and $\hat{\Gamma}_{k,ij}$ is one coefficient of the matrix $\hat{\Gamma}_k$. In (24), N^2 equations relate the N unknown diagonal terms of L. Nevertheless, solving (24), L is known except for one sign ambiguity ($L = sL_0$ with L_0 fully known and $s = \pm 1$). This is used to relate K_1 and K_2 as follows:

$$K_2 = sL_0K_1, \qquad \text{with } s = \pm 1. \tag{25}$$

In the single-mode case, removing this sign ambiguity (s) is sufficient to complete the calibration. This is no longer true in the multimode case for which we have to provide an additional piece of information to solve (22). Any additional measurement of any arbitrary generalized one- or two-port network will not give additional information. Thus, the only way to proceed is to use the reciprocity of the reflection coefficient [3], [5], [9] which now has to be assumed for the third standard

$$\Gamma = \Gamma^t. \tag{26}$$

Combining (22), (25), and (26), we obtain

$$K_1^2 L_0 \hat{\Gamma}_k^t L_0 K_1^{-2} = \hat{\Gamma}_k = K_1^2 \tilde{\Gamma}_k K_1^{-2}$$
$$k_{1,i}^2 \, k_{1,j}^{-2} \, \tilde{\Gamma}_{k,ij} = \hat{\Gamma}_{k,ij}. \qquad (27)$$

It then appears that K_1 is determined except for one arbitrary complex constant α and an $(N-1)$-order sign ambiguity

$$K_1^2 = \alpha^2 K_{10}^2. \qquad (28)$$

Inserting (28) and (25) in (22), we now get two estimates of the previously unknown reflection coefficient Γ valid in the multimode as well as in the single-mode cases

$$\Gamma' = (\alpha K_{10})^{-1} \hat{\Gamma}_1 (s\alpha L_0 K_{10}) = s K_{10}^{-1} \hat{\Gamma}_1 L_0 K_{10}$$
$$\Gamma'' = (s\alpha L_0 K_{10})^{-1} \hat{\Gamma}_2 (\alpha K_{10}) = s (L_0 K_{10})^{-1} \hat{\Gamma}_2 K_{10}. \qquad (29)$$

We finally find as in the single-mode case $(N = 1)$ [1] that an Nth-order sign ambiguity remains unsolved by the TRL algorithm. Comparing Γ' and Γ'' in (29) does not help in removing the sign ambiguity. Thus, the user must provide an estimate of the R standard to proceed. Assuming that any sign ambiguity is removed, K is known except for the one-dimensional ambiguity $(K = \alpha K_0)$. Inserting this in (10), we complete the calibration procedure and derive the transfer parameters of any DUT as follows:

$$N_x = \alpha^{-1} K_0^{-1} A_0^{-1} M_x A_0 K_0 \alpha = K_0^{-1} A_0^{-1} M_x A_0 K_0. \qquad (30)$$

This is a generalization of (23) in [1]. Thus, α appears (as in previous works) as the one-dimensional ambiguity which is not removed by the TRL algorithm. However, this is sufficient for the determination of the DUT parameters. It can be shown that scattering parameters obtained using (4) and (30) are reciprocity normalized [5], [9]. In fact, reciprocity was essential to complete the calibration. At this step, important observations must be made. Equations (24) and (27) leading to the determination of K_1 and K_2 are redundant (except in the single-mode case). Thus, these equations may be solved using a least-square approximation. Furthermore, if unrepeatable connections of standards to the VNA ports leads to measurement uncertainties, Γ' and Γ'' will exhibit different values. This is used to define the following figure of merit F:

$$||\Gamma' - \Gamma''|| = F \qquad (31)$$

where double vertical bars indicate any norm of a square matrix. This figure-of-merit must be as small as possible. This gives qualitative information about calibration precision which is evaluated without need of prior knowledge of the third standard scattering parameters. One may also observe that if the third standard is matched $(\Gamma = 0)$, (22) becomes trivial and K_i cannot be determined. This might seem obvious. Since the first two standards are perfectly matched, the third one must be reflecting and preferably highly reflecting. Furthermore, solving (27) requires that at least one off-diagonal term of each line of Γ must be nonnull. Thus, the R standard must be designed to exhibit coupling between modes ($\Gamma_{ij} \neq 0$ for $i \neq j$). For this reason, an "ideal" open or short cannot be used.

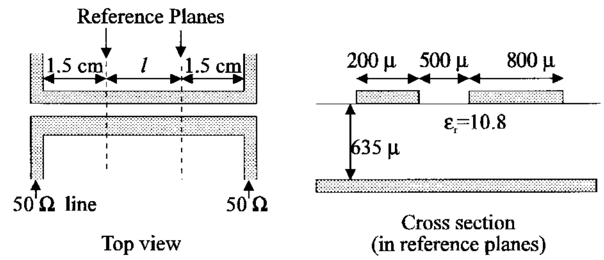


Fig. 5. Multimode TRL standards used for experiments.

Realizations other than a nontransmitting third standard (R) are possible. A transmitting network (N) would lead to a multimode TLN [1] calibration procedure. It can be shown that this would lead to conclusions similar to the generalized TRL procedure described above, and for this reason will not be presented here. Nevertheless, we must point out that TLN calibration offers the advantage of increased equation redundancy as compared to the TRL.

At this step, we have demonstrated the theoretical feasibility of TRL calibration under the multimode propagation condition. We will now present our first experimental verifications.

III. Experiments

Experimental verification of the multimode TRL method has been made for two asymmetrically coupled microstrip lines. Because two dominant modes are propagated, two single-mode connections are used on each port, as indicated in Fig. 5. A 1-cm-long line (L) standard which is approximately half-a-wavelength long at 5 GHz was used to allow TRL calibration up to 4 GHz. The third standard was realized by terminating and connecting together the two microstrip lines in the reference plane. Measurements were carried out at IEMN using a conventional two-port VNA. The VNA was first calibrated using the built-in "full two-port" menu. Connection of multimode standards to the VNA was made using microstrip to coaxial test fixtures. No corrections have been made to account for imperfect 50-Ω terminations used to measure the four-port scattering parameters. All results presented here correspond to first-run experiments. Our goal here is to prove the feasibility of multimode TRL and to illustrate potential experimental difficulties. All results have been derived without using any curve smoothing or averaging.

Using our TRL algorithm, we have experimentally determined the two propagation constants of the coupled microstrip lines. Mode "one" is chosen to have the lowest propagation constant. Slowing factors are presented in Fig. 6. Experimental values are compared to theoretical ones (dotted lines) obtained with the quasi-TEM model implemented in MDS.[1] The relative error between experimental and theoretical results is approximately 4%. Thus, the accuracy of these results is obviously not sufficient to derive precise calibration. This may be attributed to poor repeatability of microstrip-to-coaxial test fixtures as well as imperfect matching of 50-Ω loads. Better results can

[1] "HP 85150B microwave and RF design systems, components catalog," *Microwave Library Component,* vol. 3, pp. 32–33, ch. 12.

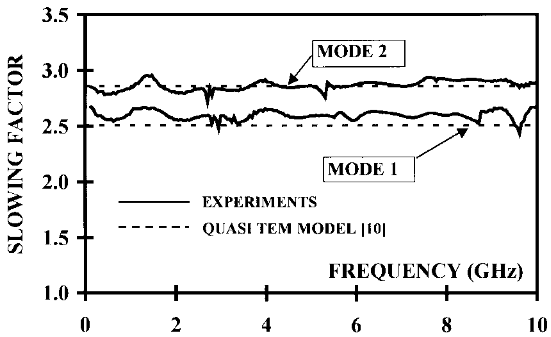
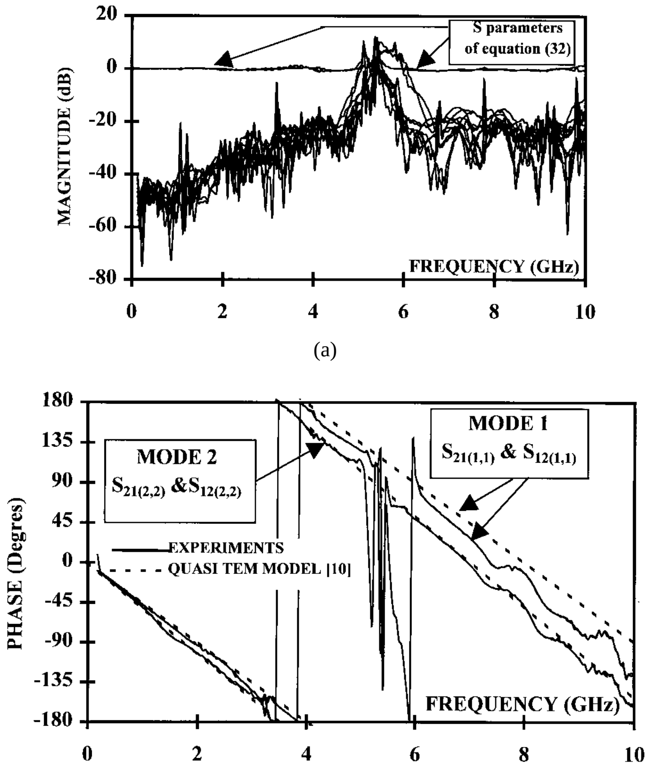


Fig. 6. Slowing factors of coupled microstrip lines. Comparison of experimental values derived from multimode TRL calibration to theoretical simulation.



(a)

(b)

Fig. 7. Generalized scattering parameters of a delay line. (a) Magnitude of the 16 S-parameters. (b) Phase of the four nonnull parameters appearing in (32).

be expected using microwaves probes and accounting for imperfect terminations [7].

Completing the calibration, we have also derived the scattering parameters of a 1.5-cm-long delay line (see Fig. 7). Using our notation, parameter $S_{21(1,1)}$ is, for example, the transmission coefficient of the lowest velocity mode (mode one) from port 1 to port 2. For an ideal transmission line, all parameters are null except the four parameters indicating the traveling of waves from one port to the other:

$$S_{ij(k,k)} = \exp(-\gamma_k l), \qquad \text{for } i \neq j \text{ and } k = 1, 2. \quad (32)$$

Our results indicate that the four above-mentioned parameters magnitudes are close to 0 dB, except around 5 GHz. The 12 other parameters have negligeable amplitude (< -20 dB) up to 5 GHz and above 6 GHz. Measurement errors occurring between 5–6 GHz are inherent to TRL calibration. In this frequency range, the length of the line standard used in the calibration (1 cm) is close to a half-wavelength for the two propagating modes. This phenomenon which is consistent with conventional single-mode TRL is also observed in the phase of the scattering parameters. Measured phase shift of the two propagating modes [four scattering parameters of (32)] are in good agreement with the quasi-TEM model,[1] except in the 5–6-GHz range.

We have pointed out previously that off-diagonal terms of the reflecting standard must be nonnull. Therefore, we performed a full-wave analysis to determine the S-parameters of our R standard. We found a coupling between modes (Γ_{12} and Γ_{21}) which was close to -20 dB. Such small reflection coefficients are sensitive to measurement uncertainties. At the same time, $\tilde{\Gamma}_{12}$ and $\tilde{\Gamma}_{21}$ (in 27) are in the same order of magnitude as Γ_{12} and Γ_{21}, and are used to solve (22). Such a situation might obviously be a source of calibration errors.

Although the precision of these measurements is not sufficient, they do demonstrate the validity of our TRL algorithm. At the same time, these first-run results indicate that further efforts must be made to improve the accuracy of multimode TRL calibration.

IV. Conclusion

An original multimode TRL calibration procedure has been proposed. This procedure is based on measurements of scattering parameters of a multiport circuit, including the multimode DUT. The corresponding algorithm is a generalization of the conventional single-mode TRL. The three multimode reciprocal standards used during calibration are a thru line, line, and reflect. The multimode TRL procedure leads to the determination of reciprocity normalized multimode scattering parameters. Using this original technique, we have experimentally determined propagation constants and generalized scattering parameters of asymmetric coupled microstrip lines. These first-run results exhibit insufficient accuracy for precise calibration; however, they demonstrate the validity of our algorithm.

For MMIC's, multimode TRL could help us in analyzing the amount of energy coupled to undesired modes. For example, conductor-backed coplanar structures which propagate several dominant modes could be optimized to avoid any energy conversion from the coplanar mode to the microstrip one. For multiconductor transmission lines and logic buses, multimode TRL would allow precise characterization of crosstalk and discontinuities such as via-hole and crossing strips. With this in mind, we now have to improve this new method and extend it to other practical structures. This can be accomplished by taking into account multiple standards or symmetry of structures. A first possible problem arising from an improper choice of the reflecting standard has been outlined. We must also design a practical calibration kit for microwave probe-based measurements. A sensitivity analysis may also be helpful to point out the origin of calibration errors and to determine the limits of this new technique. We hope that multimode TRL will open new paths in microwave measurements.

ACKNOWLEDGMENT

The authors would like to thank an anonymous reviewer for his detailed correction of the English grammar.

REFERENCES

[1] H. J. Eul and B. Schiek, "A generalized theory and new calibration procedures for network analyzer self calibration," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 724–731, Apr. 1993.

[2] R. H. Voelker, "Transposing conductors in signal buses to reduce nearest-neighbor crosstalk," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 1095–1099, May 1995.

[3] M. D. Wu, S. M. Deng, R. B. Wu, and P. Hsu, "Full-wave characterization of the mode conversion in a coplanar waveguide right angled bend," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 2532–2538, Nov. 1995.

[4] G. David, W. Schroeder, D. Jager, and I. Wolff, "2D Electro-optic probing combined with field theory based multimode wave amplitude extraction: A new approach to on wafer measurements," in *IEEE MTT-S Dig.*, Orlando, FL, June 1995, pp. 1049–1052.

[5] V. Eleftheriades, A. S. Omar, and L. P. Katehi, "Some important properties of waveguide junction generalized scattering matrices in the context of mode matching technique," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1896–1903, Oct. 1994.

[6] S. Sercu and L. Martens, "Characterizing N-port packages and interconnections with a 2-port network analyzer," presented at the 6th Topical Meeting Elect. Performances Electron. Packaging, San Jose, CA, Oct. 27–29, 1997.

[7] R. A. Speciale, "A generalization of the TSD network-analyzer calibration procedure, covering n-port scattering-parameter measurements, affected by leakage errors," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 1100–1115, Dec. 1977.

[8] R. B. Marks, "A multiline method of network analyzer calibration," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 1205–1214, July 1991.

[9] F. Olyslager, D. De Zutter, and A. T. De Hoop, "New reciprocal model for lossy waveguide structure based on the orthogonality of the eigenmodes," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 2261–2269, Dec. 1994.

Christophe Seguinot received the Ph.D. degree in electronics from the University of Lille, Villeneuve d'Ascq, France, in 1988.

In 1981, he joined the Institut d'Electronique et de Microélectronique du Nord (IEMN), Villeneuve d'Ascq, France, where he is currently an Assistant Professor. His work has included the characterization and modeling of microwave transmission lines and waveguides laid on semiconducting substrates. His current research is concerned with high-frequency discontinuities and interconnects simulation, and digital and mobile communications.

Patrick Kennis received the Ph.D. degree in electronics from the University of Lille, Villeneuve d'Ascq, France, in 1977.

While at the University of Lille, he worked on high-efficiency X-band GaAs IMPATT diodes. He has been with the Centre Hyperfrequences et Semiconducteurs (CHS) since 1972, and in 1978, he joined the Electromagnetic and Circuits Group of the CHS. He is currently a Professor at the University of Lille, and Group Leader of the Electromagnetic and Circuits Group, Institut d'Electronique et de Microlectronique du Nord (IEMN), Villeneuve d'Ascq, France. He is involved in the theoretical simulation and experimental characterization of propagating structures, interconnects and related electromagnetic discontinuities.

Prof. Kennis is member of the editorial board of the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES.

Jean-François Legier received the Ph.D. degree in electronics from the University of Lille, Villeneuve d'Ascq, France, in 1984.

In 1979, he joined the Institut d'Electronique et de Microélectronique du Nord (IEMN), Villeneuve d'Ascq, France, where he is currently an Assistant Professor. His work has included the numerical modeling of microwave-losses phenomena in planar transmission lines. His current research is focused on experimental metallic losses measurements in case of interconnects and metallic strips cross section of various shapes. He is also concerned with vector finite-element simulation for the evaluation of integrated-circuit electromagnetic-propagation characteristics.

Fabrice Huret was born in Hénin Beaumont, France, on April 2, 1965. He received the Ph.D. degree from the University of Lille, Villeneuve d'Ascq, France, in 1991.

He is currently at the Institut d'Electronique et de Microélectronique du Nord (IEMN), Villeneuve d'Ascq, France. His current research interests are numerical methods for solving electromagnetic-field problems, and uniplanar and 3-D modelization of MMIC's interconnects and packaging. He is also an Associate Member of the Laboratoire d'Etude des Matériaux et des Composants pour l'ELectronique (LEMCEL) de l'Université du Littoral Côte d'Opale, France, where his research interests are new substrate materials and their effect on integrated-circuit structures and printed antennas, and an Assistant Professor at the Université du Littoral Côte d'Opale.

Erick Paleczny received the Ph.D. degree in electronics from the University of Lille, Lille, France, in 1992.

In 1988, he joined the Institut d'Electronique et de Microélectronique du Nord (IEMN), Villeneuve d'Ascq, France, where he is currently an Assistant Professor. His research has focused on the electromagnetic modeling of planar integrated circuits using the finite-element method. His current research is concerned with high-frequency discontinuities and interconnects simulation.

Leonard Hayden (S'85–M'85) received the Ph.D. degree from Oregon State University, Corvallis, in 1993.

After spending a few years in the Frequency-Domain Instruments Division, Tektronix, Inc., he became a National Research Council Post-Doctoral Research Associate in the Microwave Metrology Group, National Institute of Standards and Technology, Boulder, CO. In November 1994, he joined Cascade Microtech, Inc., Beaverton, OR, as a Senior Applications Engineer, and is currently a Senior Engineer in the High Frequency Probing Business Unit. He has authored or co-authored 25 professional publications and holds one U.S. patent.