Improved Calibration and Measurement of the Scattering Parameters of Microwave Integrated Circuits

RENATO R. PANTOJA, MICHAEL J. HOWES, SENIOR MEMBER, IEEE, JOHN R. RICHARDSON, MEMBER, IEEE, AND ROGER D. POLLARD, MEMBER IEEE

Abstract —A novel procedure for the calibration of microwave integrated circuit test fixtures, based on a generalization of the "through-reflect-line" algorithm, is presented. Its advantages compared with previous methods, namely bandwidth of validity and standards availability, are discussed. The approach is verified through the characterization of a particular microstrip verification standard using both the "generalized TRL" and precision 7 mm calibration techniques. Comparison of the results obtained from these schemes indicates that both the effective directivity and the source/load match are better than 30 dB.

I. Introduction

The ACCURATE characterization, in terms of scattering parameters, of microwave integrated circuits (MIC's) both in the component fabrication environment (device evaluation, wafer probing) and in the field of modeling and IC design is of considerable importance. A number of calibration schemes, mainly for active device characterization, have been proposed for MIC media.

The conventional open, short, load technique has proved unsatisfactory in MIC media due to the difficulties associated with both the modeling of open- and short-circuit parasitic effects and the construction of a fixed or sliding load. Another technique [1] uses offset opens of unknown reflection coefficient, but still makes use of a perfect short. Moreover its bandwidth of operation is limited to approximately two octaves [2] for a single set of standards. More recently the "through-short-delay" (TSD) [3] algorithm has been implemented, but this technique does not eliminate the problem of relying on either a short or open standard and suitable means of modeling these. Another factor which limits the accuracy of the above techniques arises from using a "perfect" short as a calibration standard when errors are introduced when obtaining the reference planes. The approach described here uses a generalization of the "through-reflect-line" (TRL) [4] algorithm in which the only standards required are two lengths of transmission line and two equal reflects of unknown reflection coefficient. Only one type of standard, therefore, is needed (a length of transmission line) and it is possible to define the reference planes arbitrarily as a function of the ratio of the lengths of the delay lines.

The detailed solution for the error terms which define the automatic network analyzer system is presented in Section II. It is then shown how the generalized TRL algorithm can be used to obtain both the conventional TRL and LRL ("line-reflect-line") techniques. In Section III the generalized TRL is presented together with some considerations on accuracy and this is followed, in Section IV, by experimental results from a particular microstrip standard.

Manuscript received November 21, 1988; revised July 6, 1989. The first author received financial support from the Brazilian Research Council (CNPq–Brazil).

The authors are with the Department of Electrical and Electronic Engineering, University of Leeds, Leeds LS2 9JT, U.K.

IEEE Log Number 8930656.

II. Error Terms Calculation

A. General Solution

This section is concerned with the procedure for obtaining the error terms describing a network analyzer system. The procedure presented here is a development from the original TRL publication [4]. The contribution here is the reformulation in terms of S parameters and the removal of the requirement to specify a line length. It is assumed that the errors associated with measurement-port mismatch changes, typical of switching scattering-parameter test sets, are negligible.

In Fig. 1 are shown the error boxes A and B, which represent the set of systematic errors present in an S-parameter measurement. Shown, also, are the standards to be measured, namely, two lengths of identical transmission lines and two identical reflects of unknown characteristics.

Use will be made of the wave cascading matrix $[R]$, defined by

$$\begin{pmatrix} b_1 \\ a_1 \end{pmatrix} = [R] \begin{pmatrix} a_2 \\ b_2 \end{pmatrix} \tag{1}$$

in which the $[R]$ matrix may be obtained from the $[S]$ matrix using

$$[R] = \frac{1}{S_{21}} \begin{pmatrix} -\Delta & S_{11} \\ -S_{22} & 1 \end{pmatrix} \tag{2}$$

where

$$\Delta = S_{11}S_{22} - S_{12}S_{21}. \tag{3}$$

0018-9480/89/1100-1675$01.00

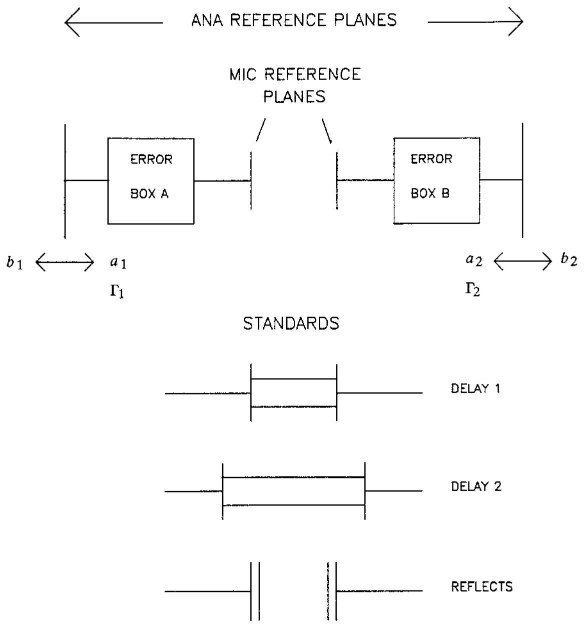


Fig. 1. Network analyzer model and standards for calibration.

If the cascading matrices of the error boxes A and B are denoted $[R_A]$ and $[R_B]$, and those for line 1 and line 2 are $[R_{L1}]$ and $[R_{L2}]$, connection of lines 1 and 2 between error boxes A and B yields

$$[R_{D1}] = [R_A]\cdot[R_{L1}]\cdot[R_B] \tag{4}$$

$$[R_{D2}] = [R_A]\cdot[R_{L2}]\cdot[R_B]. \tag{5}$$

Following the procedure outlined in [4], two quadratic equations are obtained after eliminating first $[R_B]$ and then $[R_A]$ in (4) and (5). The roots from these equations are

$$\frac{R_{11A}}{R_{21A}} = \frac{\Delta_A}{S_{22A}} = S_{11A} - \frac{S_{12A}S_{21A}}{S_{22A}} \tag{6}$$

$$\frac{R_{12A}}{R_{22A}} = S_{11A} \tag{7}$$

and

$$\frac{R_{21B}}{R_{22B}} = -S_{22B} \tag{8}$$

$$\frac{R_{11B}}{R_{12B}} = -\frac{\Delta_B}{S_{11B}} = -S_{22B} + \frac{S_{12B}S_{21B}}{S_{11B}}. \tag{9}$$

The correct root choice in (6) and (7) and in (8) and (9) is made from the observation that one of the roots has a magnitude much smaller than the other.

The only other result that may be obtained from the system of equations generated from the matrix equations (4) and (5) is [4], [5] the transmission parameter W of a delay line with the same propagation constant (γ) as that of lines l_1 and l_2 and length given by their difference $\Delta l = l_2 - l_1$, i.e.,

$$W = e^{\gamma(l_2 - l_1)}.$$

So far, expressions for the five terms $e^{2\gamma(l_2-l_1)}$, S_{11A}, S_{22B}, $S_{12A}S_{21A}/S_{22A}$, and $S_{12B}S_{21B}/S_{11B}$ have been introduced. To solve properly for the error terms in their normalized form, corresponding expressions for S_{22A}, S_{11B}, $S_{12A}S_{21A}$, and $S_{12B}S_{21B}$ are still needed. This can be done by using the same reflect Γ_R for both ports (MIC reference planes in Fig. 1), which gives

$$\Gamma_1 = S_{11A} + \frac{S_{12A}S_{21A}}{\dfrac{1}{\Gamma_R} - S_{22A}} \tag{10}$$

$$\Gamma_2 = S_{22B} + \frac{S_{12B}S_{21B}}{\dfrac{1}{\Gamma_R} - S_{11B}}. \tag{11}$$

Eliminating Γ_R^{-1} from (10) and (11) yields

$$S_{22A}\left(1 + \frac{X}{A}\right) = S_{11B}\left(1 + \frac{Y}{B}\right) \tag{12}$$

where

$$X \triangleq \frac{S_{12A}S_{21A}}{S_{22A}} \tag{13}$$

$$Y \triangleq \frac{S_{12B}S_{21B}}{S_{11B}} \tag{14}$$

$$A \triangleq \Gamma_1 - S_{11A} \tag{15}$$

$$B \triangleq \Gamma_2 - S_{22B}. \tag{16}$$

It should be noted, at this stage, that (12) relates S_{22A} and S_{11B}, the remaining unknowns, and that X, Y, A, and B are already known.

Now, with the shorter line inserted between error boxes A and B:

$$S_{11_{D1}} - S_{11A} = \frac{S_{12A}S_{21A}S_{11B}e^{-2\gamma l_1}}{1 - S_{22A}S_{11B}e^{-2\gamma l_1}} \triangleq C \tag{17}$$

so that

$$S_{22A}S_{11B}e^{-2\gamma l_1} = \left(1 + \frac{X}{C}\right)^{-1}. \tag{18}$$

Combining (12) and (18) provides the roots $S_{22A}e^{-\gamma l_1}$ and $S_{11B}e^{-\gamma l_1}$, namely

$$S_{11B}e^{-\gamma l_1} = \pm\left[\left(1 + \frac{X}{A}\right)\left(1 + \frac{Y}{B}\right)^{-1}\left(1 + \frac{X}{C}\right)^{-1}\right]^{1/2} \tag{19}$$

$$S_{22A}e^{-\gamma l_1} = S_{11B}e^{-\gamma l_1}\left(1 + \frac{Y}{B}\right)\left(1 + \frac{X}{A}\right)^{-1}. \tag{20}$$

Then, from (13) and (14) we have

$$S_{12A}S_{21A}e^{-\gamma l_1} = X\cdot S_{22A}e^{-\gamma l_1} \tag{21}$$

$$S_{12B}S_{21B}e^{-\gamma l_1} = Y\cdot S_{11B}e^{-\gamma l_1}. \tag{22}$$

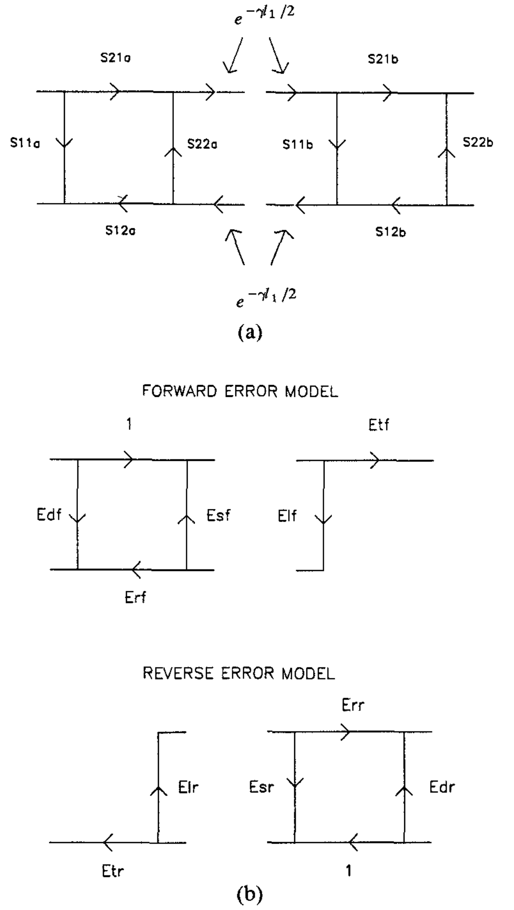


Fig. 2. Flow graph of error terms for generalized TRL calibration.

The correct choice of root in (19) is obtained as follows. From (10)

$$\left(\Gamma_R e^{\gamma l_1}\right)^{-1} = S_{22A}e^{-\gamma l_1} + \frac{S_{12A}S_{21A}e^{-\gamma l_1}}{\Gamma_1 - S_{11A}} \triangleq ke^{-j\alpha}. \quad (23)$$

Now, Γ_R and $e^{\gamma l_1}$ may be written

$$\Gamma_R = \Gamma e^{j\beta} \quad (24)$$

$$e^{\gamma l_1} = \rho e^{j\Theta} \quad (25)$$

so that

$$\Gamma \cdot \rho = k \quad (26)$$

$$\beta + \Theta = \alpha. \quad (27)$$

If the correct choice of sign in (19) is made, (27) will be satisfied. If the wrong choice is made, (27) will provide $\beta + \Theta$ with an offset of 180°. Equation (26), however, is always satisfied and no information may be extracted from it. Hence, if rough estimates of β and Θ are made, the correct root may be obtained. It is interesting to note that the error in this estimation may be as large as $\pm 90°$.

In Fig. 2(a) are shown the two error boxes A and B in which half the length of the shorter line l_1 has been included in each side. Expressions for the corresponding normalized error terms (Fig. 2(b)) are

$$E_{df} = S_{11A} \qquad E_{dr} = S_{22B} \quad (28)$$

$$E_{sf} = S_{22A}e^{-\gamma l_1} \qquad E_{sr} = S_{11B}e^{-\gamma l_1} \quad (29)$$

$$E_{rf} = S_{12A}S_{21A}e^{-\gamma l_1} \qquad E_{rr} = S_{12B}S_{21B}e^{-\gamma l_1} \quad (30)$$

$$E_{lf} = S_{11B}e^{-\gamma l_1} \qquad E_{lr} = S_{22A}e^{-\gamma l_1} \quad (31)$$

$$E_{tf} = S_{21A}S_{21B}e^{-\gamma l_1} \qquad E_{tr} = S_{12A}S_{12B}e^{-\gamma l_1}. \quad (32)$$

Comparison of (19)–(22) and (28)–(32) shows that the procedure so far developed provides the necessary error terms for calibration except for E_{tf} and E_{tr}, which are unknown at this stage. The important fact is that this calibration procedure will have its reference planes located at the middle of the shorter line.

The remaining error terms, E_{tf} and E_{tr}, are obtained from the analysis of the connection of the shorter line. The corresponding transmission parameters are

$$S_{21_{D1}} = \frac{S_{21A}S_{21B}e^{-\gamma l_1}}{1 - S_{22A}S_{11B}e^{-2\gamma l_1}} \quad (33)$$

$$S_{12_{D1}} = \frac{S_{12A}S_{12B}e^{-\gamma l_1}}{1 - S_{22A}S_{11B}e^{-2\gamma l_1}}. \quad (34)$$

Hence

$$E_{tf} = S_{21A}S_{21B}e^{-\gamma l_1} = S_{21_{D1}}\left(1 - S_{22A}e^{-\gamma l_1} \cdot S_{11B}e^{-\gamma l_1}\right) \quad (35)$$

$$E_{tr} = S_{12A}S_{12B}e^{-\gamma l_1} = S_{12_{D1}}\left(1 - S_{22A}e^{-\gamma l_1} \cdot S_{11B}e^{-\gamma l_1}\right). \quad (36)$$

B. TRL Procedure

In the TRL procedure the shorter line is "zero-length," meaning that $e^{-\gamma l_1}$ is equal to unity. This way the solution obtained in (7), (8), (19)–(22), (35), and (36) completely describes the two error boxes and the calibration is complete. It is important to note that the only information from the reflects is a rough estimate of its phase response with frequency.

As can be seen from (4) and (5), at the frequencies at which the delay line is half a wavelength the system of equations is ill conditioned, leading to significant errors. For the same reason, the best accuracy is obtained at the frequency at which the delay is a quarter wavelength long.

C. LRL Procedure

In the TRL procedure the necessary through connection and the use of the same reflect at both ports imply that the technique may be only used with sexless connectors of the same type. This problem is overcome with the LRL technique, as described in [7], and accurate measurements of devices with different type of connectors can be made.

From the previous development ((7), (8), (19)–(22), (35), and (36)), inspection of (28)–(32) shows that the value of $e^{\gamma l_1}$ is needed to obtain the correct value of the error terms. According to the original publication [6], [7] it may be solved in three ways: firstly, from a knowledge of both l_1 and l_2; secondly, from a measurement of a known reflect; and thirdly, from the precise knowledge of the phase response of a short. It is clear that all three possibilities will make use of high-quality reflects or well-defined length delays, which are not easily obtained. The manner of obtaining the solution here will become clear in the next section, but for the conventional LRL procedure, it is easily obtained from a knowledge of l_1 and l_2 as follows.

Recalling the transmission parameter W and using the fact that l_2/l_1 is known, it is easy to obtain $e^{\gamma l_1}$ from

$$e^{\gamma l_1} = e^{\ln W/(l_2/l_1 - 1)} = W^{1/(l_2/l_1 - 1)}. \tag{37}$$

The final solution for the error terms is then

$$S_{11B} = \pm e^{\gamma l_1}\left[\left(1+\frac{X}{A}\right)\left(1+\frac{Y}{B}\right)^{-1}\left(1+\frac{X}{C}\right)^{-1}\right]^{1/2} \tag{38}$$

$$S_{22A} = S_{11B}\left(1+\frac{Y}{B}\right)\left(1+\frac{X}{A}\right) \tag{39}$$

$$S_{12A}\cdot S_{21A} = X\cdot S_{22A} \tag{40}$$

$$S_{12B}\cdot S_{21B} = Y\cdot S_{11B} \tag{41}$$

$$S_{21A}S_{21B} = e^{\gamma l_1}\cdot S_{21_{D1}}\left(1 - S_{22A}S_{11B}/e^{2\gamma l_1}\right) \tag{42}$$

$$S_{12A}S_{12B} = e^{\gamma l_1}\cdot S_{12_{D1}}\left(1 - S_{22A}S_{11B}/e^{2\gamma l_1}\right). \tag{43}$$

The correct choice of root in (38) is accomplished through an estimate of the phase of the reflect, in the same way as before.

III. The Generalized TRL

A. Discussion

The generalized TRL algorithm differs from the original TRL in that there is no need for a "zero-length through." Moreover, it also differs from LRL in that no accurate information is needed about the delay lines. The only extra information required for the generalized TRL is an estimate of the phase shift through Δl in order to solve the ambiguity which arises when the square root of a complex number is calculated. It is interesting to note that the errors in both the estimates of the phase of the reflect and the phase shift through Δl need only be less than $\pm 90°$.

The advantage of this generalized TRL becomes apparent when the fixture used relies on repeatable microstrip-to-microstrip connections. In such cases, using the conventional TRL would mean employing very short lengths of transmission line and, more important, that the discontinuities present in the through connection would be different from those of the delay connection. Using the new procedure these discontinuities are lumped together with the rest of the imperfect test set [6]. If, on the other hand, repeatable coaxial-to-microstrip transitions are used, the generalized TRL provides a means of reducing the variation of the length of the fixture when measuring different standards. This avoids problems related to cable bending repeatability, since Δl is normally very small (typically 2.5 mm for a 2 to 18 GHz calibration on alumina).

Finally it should be noted that for the particular case of MIC calibration the reference planes can be chosen arbitrarily once they are a function of the chosen (or specified) ratio of the delay line lengths.

B. Accuracy Considerations

The degree of repeatability of connections involved in a complete sequence of calibration and measurement will affect the overall accuracy of any calibration procedure. In MIC media this problem is more severe than in coax or waveguide and care must be taken to ensure that this issue does not dominate the overall accuracy of the calibrated system.

With regard to the TRL technique, there are two aspects to be clarified. Firstly, as the phase shift through Δl approaches 0° or 180°, the error associated with the solution of the system of equations becomes larger. Once the maximum allowed error is limited, both the minimum and maximum phase shifts through Δl are defined together with the bandwidth over which the calibration is to be performed [6]. Typically, a phase margin of around 16° is sufficient to provide good calibration performance over a decade bandwidth.

The second aspect concerns the reference impedance of the calibrated system, defined by the impedance of the delay lines, irrespective of this value. It is, therefore, only important that the reference impedance be known accurately if absolute measurements are to be performed, as opposed to relative measurements using a particular set of transmission line standards. It is worth noting that whereas their impedance value is not used in the solution algorithm, they nevertheless still represent the calibrated system impedance reference Z_0. In the way the generalized TRL procedure has been solved, the only information needed comprises rough estimates of both the phase of the reflect and the value of the insertion phase of the shorter delay (γl_1); indeed, the reference plane is at the middle of the shorter line.

IV. Experimental Work

A comprehensive TRL/LRL algorithm has been implemented on a computer using measurement data collected from an HP8510 network analyzer system, and extensive tests, using short lengths of precision air lines, have been performed to ascertain the validity of the approach [8]. A microstrip fixture which relies on repeatable coaxial-to-microstrip transitions using tapered sections of air line has also been used with the system. Performance figures for each transition indicate the return loss and repeatability of connections to be better than 23 dB and 30 dB, respectively, over 2 to 18 GHz, as shown in Figs. 3 and 4.

To quantify the accuracy of the system a particular microstrip circuit has been fabricated and measured using both the generalized TRL technique and a conventional calibration at the test set connector planes. The corresponding microstrip calibration standards are shown in Fig. 5. The results using both calibration techniques are shown in Figs. 6 and 7. Trace *a* corresponds to measurements using the generalized TRL, trace *b* uses a 7 mm calibration at the test set connector planes, and trace *c* was

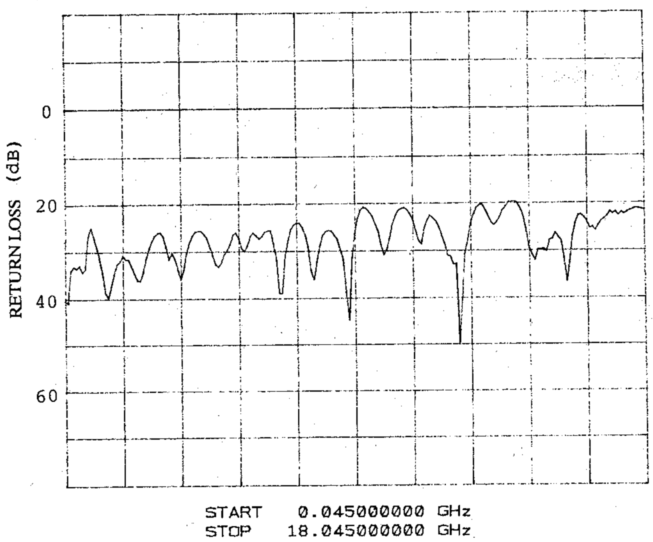


Fig. 3. Return loss of coaxial-to-microstrip transition (back to back).

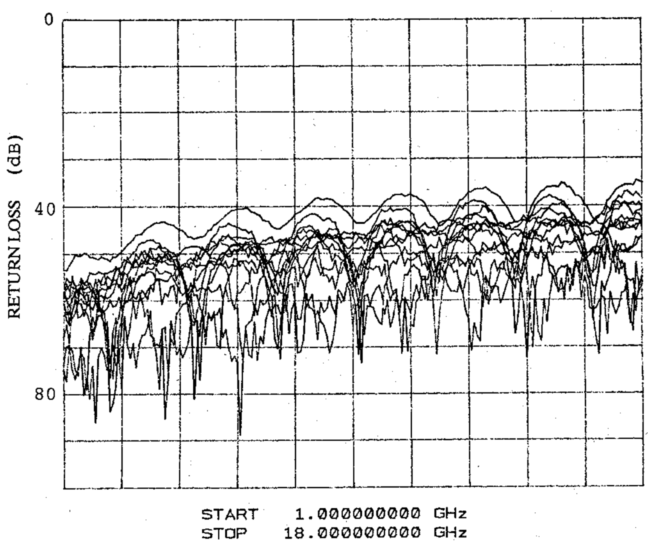


Fig. 4. Fixture connection repeatability.

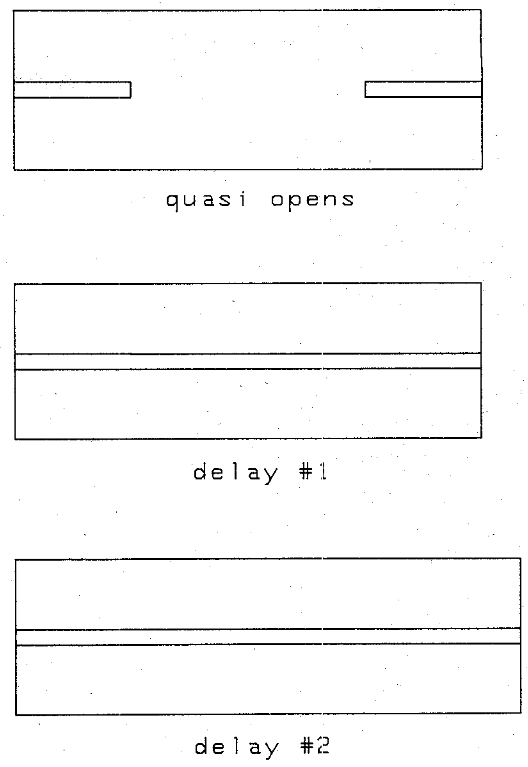


Fig. 5. Microstrip calibration standards.

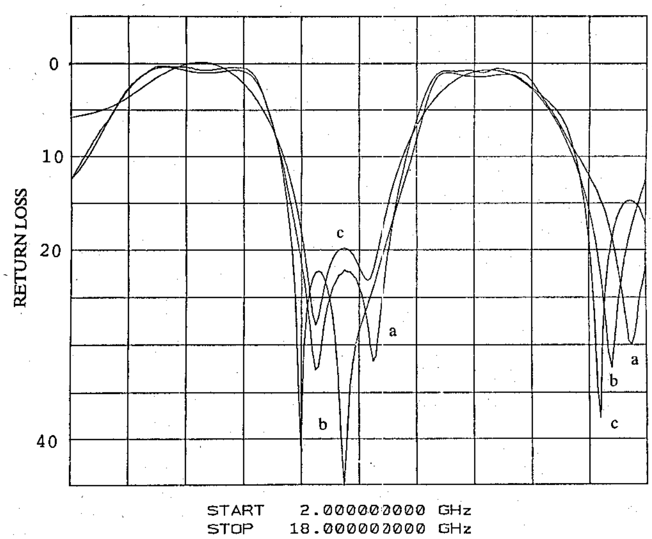


Fig. 6. Return loss of the microstrip verification standard: *a* measurement using the generalized TRL; *b* measurement using 7 mm connectors; *c* measurement *b* after use of time-domain gating.

obtained by removing the effects of the coaxial-to-microstrip transitions present on measurement *b* using time-domain techniques (gating). It is interesting to note that the circuit acts like a bandpass filter, providing both low and high return losses as well as low and high insertion losses. It is, consequently, extremely sensitive to the reference impedance in which its S parameters are evaluated.

Comparison of the results obtained with the two calibration procedures shows very good agreement. Return losses in excess of 30 dB indicate figures for effective directivity and source/load match of the same order.

V. Conclusions

A novel approach for accurate calibration and measurement of MIC's has been presented. The method is based on an extension of the TRL procedure and makes use of only one type of standard, a length of transmission line. The removal of a requirement for a precise knowledge of the electrical length of transmission lines makes the technique especially attractive for use in conjunction with on-wafer probes for MMIC evaluation. The measured performance of a verification circuit shows encouraging results when compared to those based on accurate calibration using 7 mm precision standards.

Acknowledgment

Useful technical discussions with M. Healy (University of Leeds) about the microstrip verification device are acknowledged.

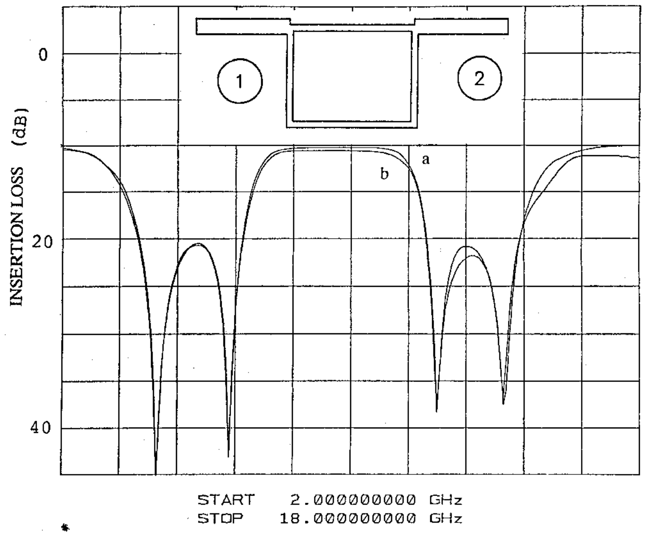


Fig. 7. Insertion loss of the microstrip verification standard.

References

[1] E. F. Da Silva, M. K. Mcphun, "Calibration of an automatic network analyser system using transmission lines of unknown characteristic impedances, loss and dispersion," *Radio Electron. Eng.*, vol. 48, no. 5, pp. 227–234, May 1978.

[2] P. R. Shepherd and R. D. Pollard, "Direct calibration and measurement of microstrip structures on gallium arsenide," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 1421–1426, Nov. 1986.

[3] D. Brubaker and J. Eisenberg, "Measure *S*-parameters with the TSD technique," *Microwaves & RF*, pp. 97–155, Nov. 1985.

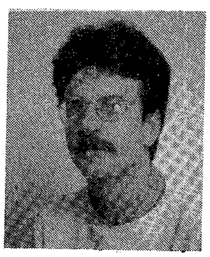
[4] G. Engen and C. Hoer, "Thru-reflect-line: An improved technique for calibrating the dual six-port automatic network analyser," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 987–993, Dec. 1979.

[5] J. Mondal and T. Chen, "Propagation constant determination in microwave fixture de-embedding procedure," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 706–713, Apr. 1988.

[6] C. Hoer, "Some questions and answers concerning air lines as impedance standards," in *ARFTG Dig.*, June 1987.

[7] C. Hoer and G. Engen, "Calibrating a dual-six-port or four-port for measuring two-ports with any connectors," in *1986 IEEE MTT-S Int. Microwave Symp. Dig.*, July 1986, pp. 665–668.

[8] R. Pantoja, M. Howes, J. Richardson, and R. Pollard, "The generalised TRL calibration technique," in *1988 Automated RF and Microwave Meas. Soc. (U.K.) Dig.*, Sept. 1988, pp. 2–10.



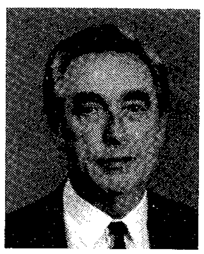
Renato R. Pantoja was born in Rio de Janeiro, Brazil, on March 14, 1961. He received the B.Sc. and M.Sc. degrees from the Military Engineering Institute in 1983 and 1986, respectively.

After graduating he worked as a Research Assistant at the Military Engineering Institute on the development of broad-band microwave voltage-controlled oscillators and antennas. He joined the Microwave Solid State Group at the University of Leeds, Leeds, U.K., in 1986, where he is currently studying for the Ph.D. degree. His main research interests are the nonlinear characterization of MESFET's incorporated in large-signal microwave circuits.

Michael J. Howes (SM'83), who is 47 and married with one child, was born in Lowestoft, Suffolk, England. A student at the University of Leeds during the 1960's, he gained a first in electronic engineering in 1965 and a Ph.D. two years later, having previously worked for the Ministry of Agriculture, Fisheries and Food developing new electronic systems for use in oceanographic studies and fish shoal detection.

In 1967 he was appointed Lecturer in Electronic Engineering at Leeds and promoted to a senior lectureship in 1976. He has had a great deal of industrial experience both in the United States and the U.K., where from 1980 to 1982 he was the Technical Director of MM Microwave, Kirkbymoorside, Yorkshire, England. He became Head of Department of Electrical & Electronic Engineering at Leeds University in 1984 and Professor of Electronic Engineering in 1988. In 1969 he set up a Microwave Solid-State Research Group, anticipating the importance of high-speed and high-frequency integrated circuits in future radar and communication systems. The group now has 40 active research workers, most of whom are working on collaborative projects with industrial establishments throughout the world. His personal research interests are concerned with the design of microwave devices and subsystems, with a particular interest in the computer-aided design of monolithic integrated circuits. He has also been actively involved in Professional Further Education and is closely involved in the Leeds Microwave Summer School, which attracts one hundred industrialists every year for intensive updating in modern electronic system design.

Dr. Howes has published over one hundred scientific papers and 16 books, and is currently the European Editor of *Microwave Systems News*. He has lectured widely all over the world and in 1980 was a Visiting Professor at Cornell University, Ithaca, NY. He is a fellow of the IEE and a member of the Institute of Physics.



John R. Richardson (M'58) was born in Durham, England, in 1930 and spent his undergraduate and postgraduate student years at Leeds University from 1949 to 1955, when he emigrated to Canada.

After one year as an instructor at the University of Toronto, he joined Sinclair Radio Laboratories (Toronto) as a research engineer and spent the next five years working on filters, duplexers, multiplexers, and antennas, the latter including responsibility for the development and design of the antenna systems on board the Canadian Alouette satellite. In 1961 he returned to Leeds University as a lecturer, and then Senior Lecturer in 1969, where his research interests have largely been concerned with antennas and microwave subsystems.

Roger D. Pollard (M'77) was born in London, England, in 1946. He received his technical education, graduating with the degrees of B.Sc. and Ph.D., in electrical and electronic engineering at the University of Leeds, Leeds, U.K.

He is currently a Senior Lecturer in the Department of Electrical and Electronic Engineering in the University of Leeds, where he has been engaged in research on topics connected with microwave solid-state devices and circuits since 1972. His experience has included work on microwave transistor circuits and two-terminal negative-resistance devices, and the use of automated techniques for the measurement of the parameters of active devices and circuits. He has acted as consultant to various organizations in the U.K. and the U.S., including periods spent with the Hewlett-Packard Company in Santa Rosa, CA.

Dr. Pollard is a member of the Institution of Electrical Engineers (U.K.) and the Institute of Physics.