A New Technique for In-Fixture Calibration Using Standards of Constant Length

Changhua Wan, Bart Nauwelaers, Dominique Schreurs, Walter De Raedt, and Marc Van Rossum

***Abstract*—This paper presents a new technique for in-fixture calibration using standards of constant length. The technique uses a through line, reflective load, symmetric two-port at a reference position, and the same two-port at a different position, all produced on substrates of the same electrical properties and physical length. When compared with the through-reflect line (TRL) technique, this one eliminates the need for a length change during calibration and device measurements while retaining comparable accuracy. Moreover, in contrast with the line-network network (LNN) technique, it provides easy resolution of all error coefficients without ambiguities and does not require physical movement of a reference two-port, but reproduction of a reference two-port on microwave integrated circuit (MIC) substrates, which is easy to realize. All these features make the new technique useful for in-fixture measurements requiring a constant distance between input and output connections. The validity of the proposed technique is illustrated by experimental results.**

***Index Terms*—Calibration, error compensation, measurement errors, microwave circuits, microwave measurements, millimeter-wave measurements.**

I. INTRODUCTION

Since the introduction of vector automatic-network analyzers (VANA's) for microwave measurements, a number of calibration techniques have been proposed. Typical ones are short-open-load through (SOLT) [1], through-short delay (TSD) [2], through-reflect line (TRL) [3], and line-reflect line (LRL) [4]. Due to the use of simple and realizable standards, TRL/LRL exhibits higher accuracy. A multiline version of TRL delivers even higher accuracy [5]. Recently, several general approaches to network-analyzer calibration have also been reported [6]–[8].

Some in-fixture applications may require a constant distance between input and output connections during system calibration and device measurements [9].[1] For these applications, the TRL technique becomes difficult and other suitable techniques have to be used. Techniques using a matched load or an attenuator [7] in place of the longer line in the TRL scheme are applicable, but the substituting standards are difficult to produce, particularly at high frequencies and in a planar format. Quite recently, a technique called the line-network network (LNN), which uses a line and symmetric network moved to three positions on the line, has been suggested [10]. This technique avoids connections/disconnections of calibration standards to ensure connection repeatability. However, it requires moving the network with precisely equal distances between two adjacent positions and without changing the connection conditions between the network and line. This is not easy. Furthermore, the LNN scheme needs some prior information to resolve a sign ambiguity in its algorithm.

This paper aims at developing a calibration technique for applications with a constant distance between input and output connections, which is comparable with the TRL in accuracy and simplicity. It is built upon, and is superior to, our earlier techniques using only reflective loads [11] or symmetric two-ports along with a through line [12]. Sections II and III of this paper present the theory and experimental results, respectively. Section IV presents the conclusion.

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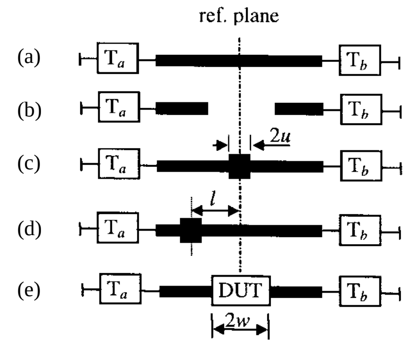

Fig. 1. Calibration measurements using: (a) thru, (b) reflect, (c) symmetric two-port, and (d) the same two-port shifted. (e) Device measurement.

II. THEORY

The following formulation is based on the well-established eight-term error model and wave cascading matrix (WCM) description [3]. The model uses two error two-ports to represent removable systematic errors, and the WCM description provides convenience in deriving error coefficients.

Fig. 1 schematically shows the measurements for obtaining the deembedded two-port scattering parameters of a device-under-test (DUT) with a microstrip as a sample transmission medium. Fig. 1(a) represents the through measurement in which a thru microstrip line connects the input transition T_a and the output transition T_b. Fig. 1(b) depicts the reflection measurement at each port using a pair of identical loads (open end). Fig. 1(c) shows the measurement of the system with a symmetric two-port discontinuity i of length $2u$ at the reference plane. Fig. 1(d) shows the measurement with the discontinuity offset by a distance l from the reference plane. Fig. 1(e) is the final measurement in which a device d of length $2w$ is mounted. All these measurements use connecting transmission lines of the same electrical property and substrates of the same material and length. Maintaining the same input and output conditions during calibration and device measurements is essential. Otherwise, any calibration techniques will not work.

The WCM of a two-port is defined as [3]

$$\boldsymbol{R} = \frac{1}{s_{21}} \begin{pmatrix} -\Delta & s_{11} \\ -s_{22} & 1 \end{pmatrix} \tag{1}$$

with

$$\Delta = s_{11}s_{22} - s_{12}s_{21}. \tag{2}$$

In (1) and (2), s_{11}, s_{12}, s_{21}, and s_{22} are the scattering parameters of the two-port. Therefore, the WCM of a nonreflecting transmission line of length x becomes

$$\boldsymbol{R}(x) = \begin{pmatrix} e^{-\gamma x} & 0 \\ 0 & e^{\gamma x} \end{pmatrix} \tag{3}$$

where γ is the complex propagation constant of the line. With the aforementioned error model and WCM description, the cascaded system with the error-two-ports corresponds to a matrix product.

For the full two-port measurements in Fig. 1(a)–(e), we have the following corresponding equations:

$$\boldsymbol{R}_t = \boldsymbol{R}_a \boldsymbol{R}_b \tag{4}$$

$$\boldsymbol{R}_I = \boldsymbol{R}_a \boldsymbol{R}(-u) \boldsymbol{R}_i \boldsymbol{R}(-u) \boldsymbol{R}_b \tag{5}$$

$$\boldsymbol{R}_f = \boldsymbol{R}_a \boldsymbol{R}(-l-u) \boldsymbol{R}_i \boldsymbol{R}(l-u) \boldsymbol{R}_b \tag{6}$$

$$\boldsymbol{R}_D = \boldsymbol{R}_a \boldsymbol{R}(-w) \boldsymbol{R}_d \boldsymbol{R}(-w) \boldsymbol{R}_b. \tag{7}$$

On the left-hand side of (4)–(7), the subscripts t, I, f, and D denote the measured matrices of the system using thru, symmetric two-port, symmetric two-port shifted, and device, respectively, while on the right side, a and b stand for the input error-two-port and the output error-two-port, respectively.

For the reflection measurement in Fig. 1(b), we have the following equation by eliminating an unknown reflection coefficient of the load

$$\frac{s_{22}^a(\Gamma_a - \alpha)}{\Gamma_a - s_{11}^a} = \frac{s_{11}^b(\Gamma_b - \beta)}{\Gamma_b - s_{22}^b} \tag{8}$$

with

$$\alpha = \frac{\Delta_a}{s_{22}^a} \quad \beta = \frac{\Delta_b}{s_{11}^b}. \tag{9}$$

In (8), Γ_a and Γ_b are, respectively, the reflection coefficients measured at the input of error-two-port a and at the output of error-two-port b, while in (8) and (9) (also from now on), both subscripts and superscripts in alphabetical letters mean the same thing.

Expanding (4) produces

$$\alpha = \frac{s_{11}^t s_{22}^b - \Delta_t}{s_{22}^b - s_{22}^t} \tag{10}$$

$$\beta = \frac{s_{22}^t s_{11}^a - \Delta_t}{s_{11}^a - s_{11}^t} \tag{11}$$

$$s_{22}^a s_{11}^b \left(\alpha - s_{11}^t\right) = s_{11}^a - s_{11}^t. \tag{12}$$

Since the reference two-port is symmetric, (5) gives

$$s_{22}^a \left[\Delta_I - s_{11}^I s_{22}^b - \alpha(s_{22}^I - s_{22}^b)\right] = s_{11}^b \left[\Delta_I - s_{11}^I \beta - s_{11}^a (s_{22}^I - \beta)\right]. \tag{13}$$

After taking the ratio of (8)–(13) and making some manipulations using (10) and (11), we obtain

$$s_{22}^b = \frac{A_1 s_{11}^a + B_1}{C_1 s_{11}^a + D_1} \quad \beta = \frac{A_1 \alpha + B_1}{C_1 \alpha + D_1} \tag{14}$$

where

$$A_1 = B\Delta_t + C - E s_{22}^t \tag{15}$$

$$B_1 = A s_{22}^t - C s_{11}^t + F\Delta_t \tag{16}$$

$$C_1 = B s_{11}^t - D s_{22}^t + F \tag{17}$$

$$D_1 = A + D\Delta_t - E s_{11}^t \tag{18}$$

$$A = \Gamma_a \left(\Gamma_b s_{11}^I - \Delta_I\right) \tag{19}$$

$$B = \Gamma_b - s_{22}^I \tag{20}$$

$$C = \Gamma_b \left(\Gamma_a s_{22}^I - \Delta_I\right) \tag{21}$$

$$D = \Gamma_a - s_{11}^I \tag{22}$$

$$E = \Gamma_a \Gamma_b - \Delta_I \tag{23}$$

$$F = \Gamma_a s_{22}^I - \Gamma_b s_{11}^I. \tag{24}$$

Considering the same two-port at two positions, one can use (5) and (6) to obtain

$$s_{22}^b = \frac{A_2 s_{11}^a + B_2}{C_2 s_{11}^a + D_2} \quad \beta = \frac{A_2 \alpha + B_2}{C_2 \alpha + D_2} \tag{25}$$

where

$$A_2 = s_{21}^I s_{22}^f - s_{21}^f s_{22}^I \tag{26}$$

$$B_2 = s_{21}^f \Delta_I - s_{21}^I \Delta_f \tag{27}$$

$$C_2 = s_{21}^I - s_{21}^f \tag{28}$$

$$D_2 = s_{21}^f s_{11}^I - s_{21}^I s_{11}^f. \tag{29}$$

Eliminating s_{22}^b or β between (14) and (25) immediately leads to the following quadratic equation:

$$(A_1C_2 - A_2C_1)z^2 + (A_1D_2 + B_1C_2 - A_2D_1 - B_2C_1)z + (B_1D_2 - B_2D_1) = 0 \tag{30}$$

where the unknown z represents both s_{11}^a and α or, in other words, one root of (30) is s_{11}^a and the other is α. Apparently, the criterion for root assignment in [3] can also be used here. Upon determining s_{11}^a and α, one can calculate s_{22}^b and β using (10) and (11) or using (14) or (25). Owing to the shift of the symmetric two-port in Fig. 1(c) and (d), (5) and (6) yield an expression relating the complex propagation constant to s_{11}^a, β, and other measured information. The expression is

$$e^{2\gamma l} = \frac{s_{21}^I \left[\Delta_f - s_{11}^f \beta - s_{11}^a (s_{22}^f - \beta)\right]}{s_{21}^f \left[\Delta_I - s_{11}^I \beta - s_{11}^a (s_{22}^I - \beta)\right]}. \tag{31}$$

Multiplying (8) by (12) and manipulating the resulting equation, one has

$$s_{22}^a = \pm \sqrt{\frac{(s_{11}^a - s_{11}^t)(\Gamma_a - s_{11}^a)(\Gamma_b - \beta)}{(\alpha - s_{11}^t)(\Gamma_a - \alpha)(\Gamma_b - s_{22}^b)}}. \tag{32}$$

Like the TRL scheme, the sign ambiguity here can be resolved using the following estimate based on a nominal value of the reflection coefficient of the load Γ_l and other measured information:

$$s_{22}^a \approx \frac{\Gamma_a - s_{11}^a}{\Gamma_l(\Gamma_a - \alpha)}. \tag{33}$$

Then, s_{11}^b can be determined from (8) or (12), and Δ_a and Δ_b from (9). The product $s_{12}^a s_{12}^b$ is also a necessary error coefficient, which is derived from (4) and given by

$$s_{12}^a s_{12}^b = \frac{(s_{11}^a - s_{11}^t)(\beta - s_{22}^b)}{s_{21}^t}. \tag{34}$$

Finally, the WCM of the device is obtained from (7) and calculated by

$$\boldsymbol{R}_d = \boldsymbol{R}^{-1}(-w)\boldsymbol{R}_a^{-1}\boldsymbol{R}_D\boldsymbol{R}_b^{-1}\boldsymbol{R}^{-1}(-w) = \frac{1}{s_{12}^a s_{12}^b}\boldsymbol{R}(w)\begin{pmatrix} 1 & -s_{11}^a \\ s_{22}^a & -\Delta_a \end{pmatrix}\boldsymbol{R}_D\begin{pmatrix} 1 & -s_{11}^b \\ s_{22}^b & -\Delta_b \end{pmatrix}\boldsymbol{R}(w). \tag{35}$$

III. Experimental Verification

To verify the above theory, calibration standards for the current technique have been designed at a center frequency of 27.5 GHz and produced on an indium–phosphide wafer. For comparison, it would be desirable to have some results from other accurate techniques, such as thru-reflect-match (TRM) [7] and the LNN [10], which also maintain a constant distance between input and output connections during calibration. However, TRM fails to provide the complex propagation constant for rotating the reference plane and requires an idealized nonreflecting load, while the LNN involves ambiguities which are not easy to resolve. Thus, standards for the TRL calibration have also been produced on the same wafer. The TRL-based results are used only for verifying the proposed technique not intended for a parallel-performance comparison between these techniques

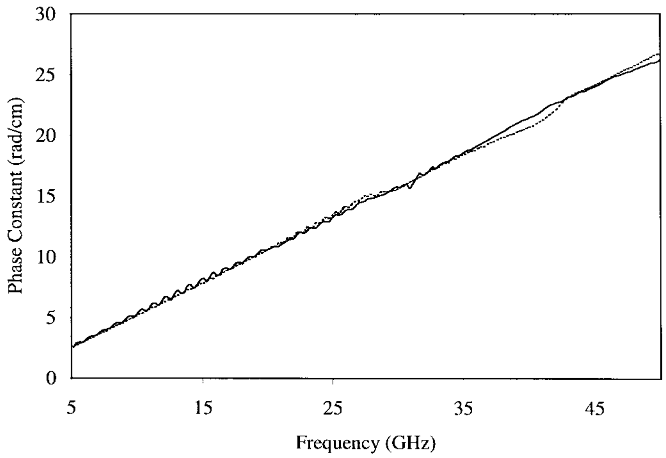

Fig. 2. A comparison between phase constants resulting from the new calibration (solid line) and the TRL (dashed line).

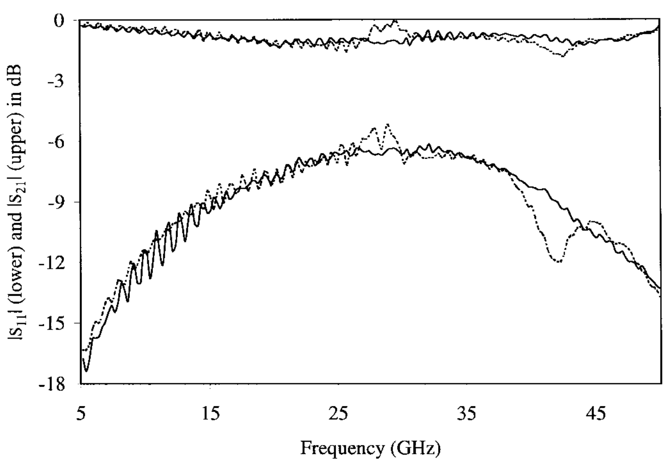

Fig. 3. A comparison between the s-parameters of a CPW discontinuity after the new calibration (solid line) and the TRL technique (dashed line).

because the two techniques are to be used in different situations. On-wafer measurements have been performed in the frequency range of 5–50 GHz using a probe station from Cascade Microtech.

For TRL, the propagation constant can be extracted directly from two line measurements without calculating any error coefficients [3], [5] or calculated upon obtaining some error coefficients [3]. In contrast, for the current technique, the propagation constant can be determined only after obtaining some error coefficients.

All the calibration standards for the new technique have a length of 3324 μm, and the symmetric two-port used is simply a discontinuity of 1000-μm length on the reference transmission line.

Fig. 2 compares the phase constants calculated from measured information using the new calibration technique and TRL technique. The two curves are in excellent agreement.

Both calibrations have also been used to generate error-corrected s-parameters of a coplanar waveguide (CPW) discontinuity. The results are shown in Fig. 3 and, again, an excellent agreement between the two calibrations is observed.

It should be noted that the magnitude of the reflection parameters of the symmetric two-port should not be too small because the two-port is equivalent to a one-port reflective load in the calibration [11]. In other words, a reflective load can be used to replace the two-port. The modification of the current calibration algorithm corresponding to this replacement is straightforward [11].

IV. Conclusions

A new network-analyzer calibration technique for in-fixture measurements requiring a constant distance between input and output connections to calibration standards and devices has been proposed and experimentally verified in this paper. In common with the TRL technique, the new technique uses a nominal short or open to resolve a sign ambiguity and the well-established criterion for root assignment during calibration. The significant distinction between the proposed technique and the TRL technique is that the former utilizes an unknown symmetric two-port at two positions on its reference transmission line in place of a longer line, as employed by the latter. It is this distinction that makes this paper's technique particularly useful for applications not allowing a length change during calibration as well as device measurements. The accuracy of this paper's technique has been found to be comparable with that of the TRL technique through actual measurements. This is an expected result because both techniques use simple and realizable standards. Like the design criterion for line lengths in the TRL technique, the relative shift of the reference two-port on two standards should be designed to be an effective quarter-wavelength-long at the center frequency for best accuracy. The proposed technique in its original form covers an $8:1$ frequency bandwidth. To cover a wider bandwidth, one needs to use more standards with the same reference two-port at corresponding positions. This is analogous to the case with the TRL scheme. Also, to ensure accuracy, the symmetric two-port should not have too low reflection, or another reflective load can be used instead.

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