Precisely Calibrated Coaxial-to-Microstrip Transitions Yield Improved Performance in GaAs FET Characterization

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Abstract —**A new approach for calibrating coaxial-to-microstrip transitions up to 26.5 GHz with high precision is presented. An ideal through, noncritical open, noncritical short, and a surface absorber are used as microstrip standards for the calibration. The calibration measurement and a novel approach in extracting the scattering parameters of the transitions are described. Error-corrected results on broad-band measurements of the scattering coefficients of packaged FET's in a hybrid circuit configuration are given.**

I. Introduction

THE PRECISE broad-band characterization of GaAs FET's from measurement data is necessary for the development of physically based self-consistent device models. This in turn will be helpful in understanding the physical properties of the FET and in studying the influence of small variations in geometry and material parameters on the electrical characteristics. In addition, these device models may serve as a basic tool for computer-aided design (CAD) techniques of nonlinear microwave circuits such as amplifiers, oscillators, and mixers. As was confirmed by Vaitkus [1], the levels of uncertainty of extracted model parameters depend strongly on the bandwidth of the S parameters considered and on the S-parameter measurement errors as well.

In general, the attainable accuracy is restricted to the performance of the coaxial measurement system used, mostly a vector automatic network analyzer (ANA), and in particular to the calibration method applied to characterize the coaxial-to-microstrip transitions of a microstrip test fixture. In a one-step calibration procedure, the complete measurement setup comprising the ANA and the connected coaxial-to-microstrip transitions is used for calibration. This approach has not been followed in this paper, because any calibration verification test would give only information on the total measurement system uncertainty. On the other hand, a two-step calibration of the measurement system, which means network analyzer calibration prior to system calibration, is more suited to separate the residual calibration errors due to the ANA used and the coaxial-to-microstrip error-parameter extraction procedure employed.

Manuscript received August 27, 1988; revised July 31, 1989.

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IEEE Log Number 8931553.

Many authors have been engaged in the field of ANA calibration techniques with microstrip reference planes involved. One of the main shortcomings of the proposed procedures is the repeated disconnection and reassembly of the coaxial-to-microstrip transitions within a set of calibration runs, which may result in uncontrollable reproducibility errors. This has recently been outlined by Shepherd *et al.* [2], [3]. Bauer *et al.* proposed a redundant set of measurements in order to minimize the repeatability problem [4]. It should be noted that similar problems exist with microwave wafer probes for on-wafer measurement [5], [6]. Shurmer placed the reference standards directly in the microstrip medium, thus avoiding any disturbance of the coaxial-to-microstrip transitions during calibration [7]. However, as reported by Dunleavy *et al.* [8], the repeatability of the required microstrip-to-microstrip interconnections is considered to be unsatisfactory.

Furthermore the attainable measurement accuracy depends largely on the calibration procedure applied, the reference standards referred to, and the quality of conditioning of the defined set of equations for the error-parameter extraction. For instance, Shurmer [7] and Da Silva *et al.* [9] proposed open- and short-circuit microstrip calibration standards for the extraction of the reflection coefficients of the transitions. As is well known, such highly reflective standards cannot be described precisely. Consequently, even small uncertainties in the description of the standards lead to extremely large errors, as has been demonstrated e.g. by Moser [10], using four short-circuited microstrip lines with different incremental lengths. Similar error-parameter extraction problems exist with the calibration proposals given in [2], [3], [7], and [11]. These experiences agree well with the conclusion of Glasser [12] that the error grows in regions of the Smith chart that are remote from the location of the standard loads.

Novel calibration procedures have been developed recently based on transmission line standards, eliminating the requirement for precise knowledge of the terminations used [13]–[15]. In addition to noncritical highly reflective standards, such as open and short, two different lengths of

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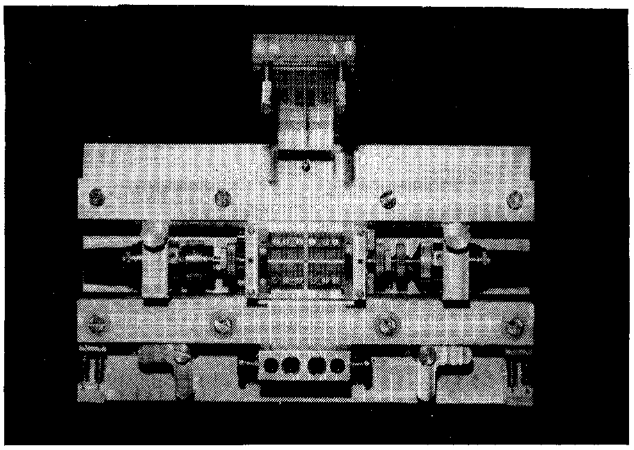


Fig. 1. The developed microstrip test fixture.

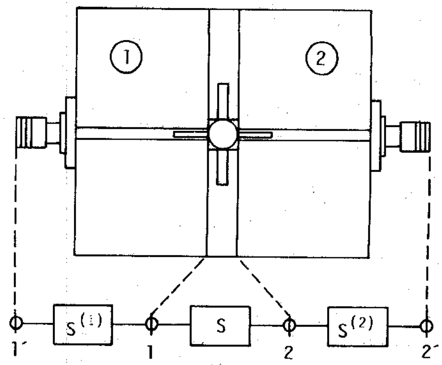


Fig. 2. Definition of the reference planes and scattering parameters of the microstrip test fixture.

STEP	MEASUREMENT CONFIGURATION	MICROSTRIP STANDARD
1	$S^{(1)}$ $S^{(2)}$ $\Gamma_1^{(1)}$ t_{21} $\Gamma_1^{(2)}$	IDEAL THROUGH
2	$S^{(1)}$ l l $S^{(2)}$ $\Gamma_2^{(1)}$ Γ_{SL} Γ_{SL} $\Gamma_2^{(2)}$	SLIDING LOAD
3	$S^{(1)}$ $S^{(2)}$ $\Gamma_3^{(1)}$ Γ_O Γ_O $\Gamma_3^{(2)}$	NON-IDEAL OPEN
4	$S^{(1)}$ $S^{(2)}$ $\Gamma_4^{(1)}$ Γ_S Γ_S $\Gamma_4^{(2)}$	NON-IDEAL SHORT
5	$S^{(1)}$ $S^{(2)}$ $\Gamma_5^{(1)}$ Γ_{FL} Γ_{FL} $\Gamma_5^{(2)}$	FIXED LOAD

Fig. 3. Series of calibration steps and used standards.

transmission lines are generally involved. The line differences are usually chosen to be approximately a quarter wavelength at the center of the frequency range to be covered. The equations derived for the transition parameters will become ill conditioned for length differences of the order of a multiple of half-wavelength, as is demonstrated by the results given in [16]. An example of the successful approach of the TRL method (thru-reflect-line) in microstrip media at low frequencies of 2–8 GHz is reported in [17], which examines the response of a microcircuit amplifier.

The motivation for developing the calibration method proposed in this paper comes from the following. Any random error which may originate in any variation of the coaxial-to-microstrip transitions should be avoided. The electromagnetic field of the active device interacts with the environment of the installation; i.e., the measured scattering coefficients are dependent on the substrate material and the grounding techniques used. With respect to hybrid technology, it is necessary to test the active device in a situation which is similar to the final microstrip application. The principal ideas of the novel approach were published earlier by the authors [18] neglecting the radiation loss of the open standards. This restriction has been overcome in this paper, which proposes an iteration procedure for the determination of the error-parameters of the coaxial-to-microstrip transitions.

II. Method of Calibration

A. Measurement Procedure

The calibrated microstrip test fixture developed for highly accurate characterization and modeling of planar microwave transistors is shown in Fig. 1. The two halves of the fixture are compact mechanical and electrical units yielding high transition rigidity. A simplified representation of the microstrip text fixture is shown in Fig. 2. The coaxial input and output reference planes 1′ and 2′ coincide with the ANA calibration reference planes. The microstrip reference planes are denoted by ports 1 and 2. The S-parameter notation of the cascaded two-ports has been chosen according to [19]. Each of the microstrip lines has a length of about 25 mm, thus avoiding any interaction of the device and adapter discontinuities via higher order microstrip modes. The measurement of the device is performed after calibration with regard to the microstrip reference planes which coincide with the gate and drain terminals of the device mounted on a fixture insert. Transistor chips are eutectically bonded on the central carrier between the two halves of the fixture and are wire-bonded to the microstrip lines.

The calibration procedure of the microstrip test fixture is performed as follows (Fig. 3). In a first step the two halves of the fixture are assembled so as to support a whole microstrip substrate with a continuous 50 Ω homogeneous microstrip line. In this case the microstrip reference planes 1 and 2 are identical; i.e., the configuration represents an ideal "through" standard. The calibrated network analyzer measures the forward and reverse reflection, as well as the transmission coefficient. The ideal through configuration of step 1 is then used to measure the outer eigenreflection coefficients of the transitions by employing the sliding load technique (step 2). At low frequencies this measurement is replaced by a fixed load measurement (step 5). In a third step the microstrip substrate is carefully cut through along the defined microstrip reference planes. The nonideal open reference standards, which are assumed to be identical, are then measured. Finally, the open-ended microstrip lines are terminated one after another with the same short-circuiting device, and the input reflection coefficients are measured (step 4).

B. Error-Parameter Extraction

The reflection coefficient $S_{11}^{(\mu)}$ with $\mu = 1,2$ is first derived using either the sliding load technique (step 2) or the fixed load measurement at low frequencies (step 5). The measured input reflection coefficient $\Gamma_2^{(\mu)}$ of transition μ terminated with the sliding load reflection coefficient $\Gamma_{SL}(l) = \Gamma_{SL}(0)\exp(-2\gamma l)$ can be written as

$$\Gamma_2^{(\mu)}(l) = S_{11}^{(\mu)} + \frac{\left(S_{21}^{(\mu)}\right)^2 S_{22}^{(\mu)*}|\Gamma_{SL}(0)|^2}{1-|\Gamma_{SL}(0)S_{22}^{(\mu)}|^2} + \frac{|\left(S_{21}^{(\mu)}\right)^2\Gamma_{SL}(0)|}{1-|\Gamma_{SL}(0)S_{22}^{(\mu)}|^2}\exp\left(j\phi(l)\right) \quad (1)$$

with the phase angle $\phi(l)$ dependent on the absorber displacement l and transmission line loss neglected. With $|S_{22}^{(\mu)}|, |\Gamma_{SL}(0)| \ll 1$, the center of the reflection circle is approximately $S_{11}^{(\mu)}$. Under worst-case conditions with $|S_{11}^{(\mu)}| = |S_{22}^{(\mu)}| = |\Gamma_{SL}(0)| = 0.2$ and $|S_{21}^{(\mu)}| = 1$ this assumption yields a magnitude error of 0.008, or 4 percent, and a phase uncertainty of 2.4°. With regard to random small variations in the magnitude of the measured input reflection coefficients due to displacement errors of the sliding load as well as economy of calibration time, five different positions of the sliding load have been proved adequate. The evaluation of the center of the input reflection circle is performed using a least-squares fitting procedure, as described e.g. in [20] and [21]. With respect to the defined small number of load positions, the evaluation of the center is always verified by a repeat measurement.

At low frequencies the sliding load is replaced by a fixed load, terminating the open-ended microstrip lines with a 50 Ω chip resistor. In this case the attenuation of the surface absorber would be insufficient so that parasitic reflections at the opposite transition would contribute to the input reflection coefficient $\Gamma_2^{(\mu)}$:

$$\Gamma_2^{(\mu)}(l) = S_{11}^{(\mu)} + \left(S_{21}^{(\mu)}\right)^2\left\{\zeta^2 S_{22}^{(\nu)}\exp\left(2\gamma l_{SL}\right) + \xi\exp\left(-2\gamma l\right)\right\} \quad (2)$$

with sliding load transmission coefficient ζ and reflection coefficient ξ. The surface absorber was constructed from ferrite material MF 124. The absorber length l_{SL} was 20 mm. With worst-case values for the magnitudes of $S_{11}^{(\mu)}$ and $S_{22}^{(\mu)}$ equal to 0.1 at 10 GHz and a permitted error of 5 percent in the evaluation of $S_{11}^{(\mu)}$, a minimum attenuation of 12.6 dB is required. As could be shown by measurement this is fulfilled by the sliding load configuration at frequencies of 7.5 GHz and above. Fig. 4 shows an anticipated result on the error-corrected measured frequency-dependent magnitude of the fixed load reflection coefficient Γ_{FL}. Defined as an ideal matched load at low frequencies during calibration, the error-corrected data simulate zero reflection up to about 7.5 GHz. At higher frequencies the fixed load is unknown in the sense of calibration; i.e., the smooth curve describes the physical features of the load realistically. By fitting the curve of the error-corrected data, an error estimation for the lower frequency range can be made. A maximum magnitude error of 0.016 and a maximum phase error of 8° are attained at 7.5 GHz.

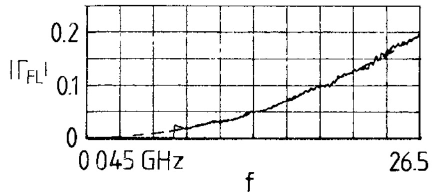


Fig. 4. Error-corrected measured magnitude of reflection coefficient of the 50 Ω fixed load and curve fitting by least squares.

The determination of the transmission coefficients of the transitions starts with the measured input reflection coefficients $\Gamma_3^{(\mu)}$ and $\Gamma_4^{(\mu)}$ of shorted and open-ended microstrip line (steps 3 and 4), which may be written as

$$\Gamma_3^{(\mu)} = S_{11}^{(\mu)} + \frac{\left(S_{21}^{(\mu)}\right)^2\Gamma_o}{1-S_{22}^{(\mu)}\Gamma_o} \qquad (\mu = 1,2) \quad (3)$$

$$\Gamma_4^{(\mu)} = S_{11}^{(\mu)} + \frac{S_{21}^{(\mu)2}\Gamma_s}{1-S_{22}^{(\mu)}\Gamma_s} \qquad (\mu = 1,2). \quad (4)$$

It should be noted that a knowledge of the square of $S_{21}^{(\mu)}$ is sufficient for the de-embedding procedure of one-port devices, whereas two-port measurement corrections require the product of the two transmission coefficients of the coaxial-to-microstrip transitions. Eliminating $S_{22}^{(\mu)}$, the square of $S_{21}^{(\mu)}$ is

$$\left(S_{21}^{(\mu)}\right)^2 = \left(\frac{1}{\Gamma_s} - \frac{1}{\Gamma_o}\right)\frac{\left(\Gamma_4^{(\mu)} - S_{11}^{(\mu)}\right)\left(S_{11}^{(\mu)} - \Gamma_3^{(\mu)}\right)}{\left(\Gamma_4^{(\mu)} - \Gamma_3^{(\mu)}\right)} \qquad (\mu = 1,2). \quad (5)$$

With the physical open and short reflection coefficients Γ_o and Γ_s described as

$$\Gamma_o = (1-\delta_o)\exp\left(j\phi_o\right) \quad (6)$$

$$\Gamma_s = -(1-\delta_s)\exp\left(j\phi_s\right) \quad (7)$$

and with $\delta_o, \delta_s, |\phi_o|, |\phi_s| \ll 1$, the mixed termination term

$(\Gamma_s^{-1}-\Gamma_o^{-1})$ can be approximated as

$$\Gamma_s^{-1}-\Gamma_o^{-1}\approx -2\left(1+\frac{\delta_o+\delta_s}{2}-j\frac{\phi_o+\phi_s}{2}\right) \tag{8}$$

where δ_o represents radiation loss of the air-terminated microstrip lines, and δ_s skin loss in the short circuit. As will be evident from error-corrected open and short measurements, δ_s turns out to be zero in a first-order estimation and $\delta_o \leqslant 0.06$ for the substrates used over the full frequency range. As is well known, the fringing field at the microstrip open end increases the electrical length of the line. The short-circuiting device comprises a small spring-loaded gib key which ensures a good contact to the microstrip. This special construction yields a decrease in the electrical length of the line. Moreover, as will be shown later, the absolute values of the excess phases are approximately equal. Thus, (8) is simplified as

$$\Gamma_s^{-1}-\Gamma_o^{-1}\approx -2\left(1+\frac{\delta_o}{2}\right). \tag{9}$$

In an earlier paper [18], the mixed term has been chosen equal to -2, meaning that extraction errors for $S_{21}^{(\mu)}$ and $S_{22}^{(\mu)}$ result mainly from neglecting the radiation loss of the open-ended reference standards. In this paper improved extraction values are obtained by a subsequent iteration procedure taking into account the measurement result of t_{21}. Regarding the ideal through configuration, the measured overall transmission coefficient t_{21} refers to the true transition scattering coefficients:

$$t_{21}=\frac{S_{21}^{(1)}S_{21}^{(2)}}{1-S_{22}^{(1)}S_{22}^{(2)}}. \tag{10}$$

The calculated transmission based on the extracted first-order error parameters for the transitions as described in [17] may be written as

$$t_{21,\mathrm{cal}}=\left(\frac{S_{21}^{(1)}S_{21}^{(2)}}{1-S_{22}^{(1)}S_{22}^{(2)}}\right)_{\mathrm{cal}}. \tag{11}$$

Assuming

$$S_{22}^{(1)}S_{22}^{(2)}\approx\left(S_{22}^{(1)}S_{22}^{(2)}\right)_{\mathrm{cal}} \tag{12}$$

the following relationship for an improved value for the product $S_{21}^{(1)}S_{21}^{(2)}$ can be found:

$$S_{21}^{(1)}S_{21}^{(2)}\approx\left(S_{21}^{(1)}S_{21}^{(2)}\right)_{\mathrm{cal}}\frac{t_{21}}{t_{21,\,\mathrm{cal}}}. \tag{13}$$

From (5) the ratio of the square value of the transmission coefficients $S_{21}^{(1)}$ and $S_{21}^{(2)}$ yields

$$\left(\frac{S_{21}^{(1)}}{S_{21}^{(2)}}\right)^2=\frac{\Gamma_3^{(1)}-S_{11}^{(1)}}{\Gamma_3^{(2)}-S_{11}^{(2)}}\,\frac{\Gamma_4^{(1)}-S_{11}^{(1)}}{\Gamma_4^{(2)}-S_{11}^{(2)}}\,\frac{\Gamma_4^{(2)}-\Gamma_3^{(2)}}{\Gamma_4^{(1)}-\Gamma_3^{(1)}}=C. \tag{14}$$

As can be seen, the mixed termination term has been dropped. From (14), the following relation can be derived:

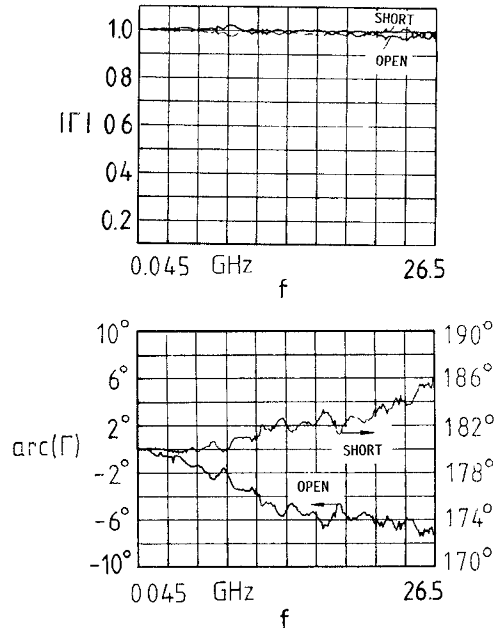


Fig. 5. Error-corrected measurement of the magnitude and phase of the short and open standard.

$$\left.\begin{aligned}\left(S_{21}^{(1)}\right)^2&=C^{1/2}S_{21}^{(1)}S_{21}^{(2)}\\ \left(S_{21}^{(2)}\right)^2&=C^{-1/2}S_{21}^{(1)}S_{21}^{(2)}\end{aligned}\right\}. \tag{15}$$

With (13), this results in

$$\left.\begin{aligned}\left(S_{21}^{(1)}\right)^2&=C^{1/2}S_{21}^{(1)}S_{21}^{(2)}t_{21}/t_{21,\mathrm{cal}}\\ \left(S_{21}^{(2)}\right)^2&=C^{-1/2}S_{21}^{(1)}S_{21}^{(2)}t_{21}/t_{21,\mathrm{cal}}\end{aligned}\right\}. \tag{16}$$

The sign of $S_{21}^{(1)}S_{21}^{(2)}$ for two-port error correction can be directly derived from the measured transmission coefficient t_{21} of the ideal through. With $t_{21}\approx S_{21}^{(1)}S_{21}^{(2)}$ the following relation can be found for the sign determination:

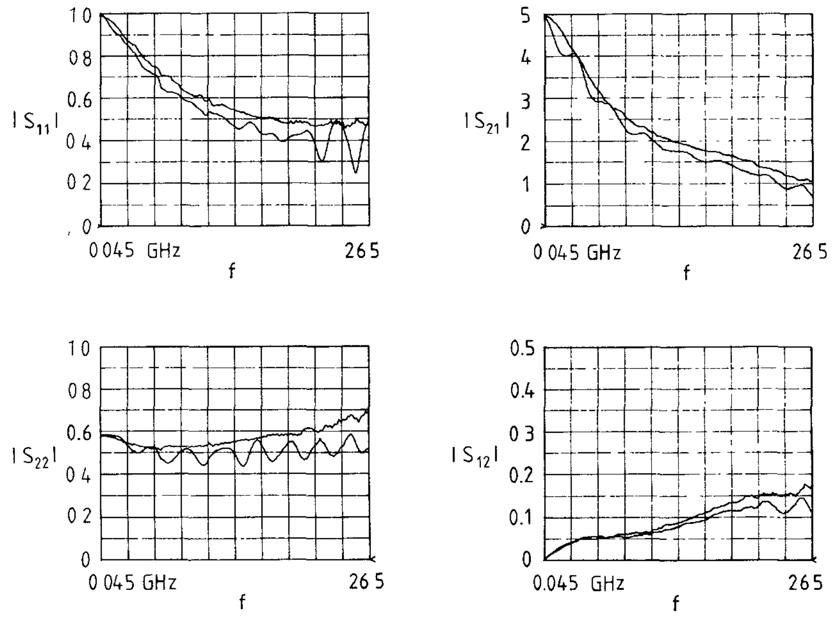
$$|\theta+m\pi-\mathrm{arc}(t_{21})|\leqslant\epsilon \tag{17}$$

with the decision interval ϵ chosen to be 35° according to experience and θ as the phase of $S_{21}^{(1)}S_{21}^{(2)}$. The quantity m is chosen equal to 0 (positive sign for the transmission product) or 1 (negative sign) to satisfy the relation. The error $\delta(S_{21}^{(\mu)2})$ resulting from small uncertainties in the determination of the reflection coefficients $S_{11}^{(\mu)}$ and $(\Gamma_s^{-1}-\Gamma_o^{-1})$ can be estimated as

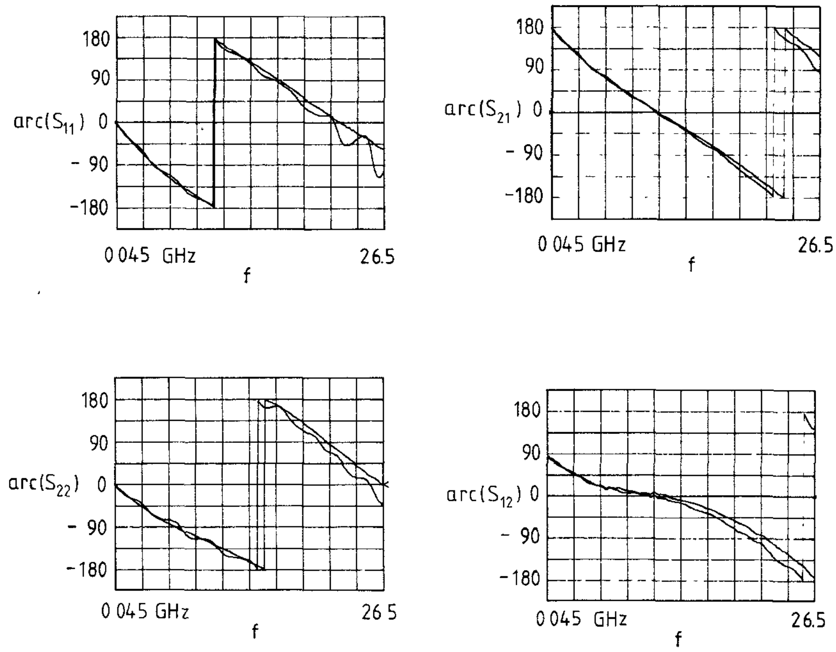
$$\delta\left(S_{21}^{(\mu)2}\right)\approx-\frac{2S_{11}^{(\mu)}}{\Gamma_4^{(\mu)}-\Gamma_3^{(\mu)}}\delta S_{11}^{(\mu)}. \tag{18}$$

As will be proved, (18) yields only a second-order error effect on the total calibration uncertainty of the fixture used.

The reflection parameter $S_{22}^{(\mu)}$ is derived from the reflection measurements of the through-line configuration and



(a)



(b)

Fig. 6. (a) Measured and error-corrected amplitudes of S parameters of packaged GaAs FET NE 71083 on CuClad substrate material. (b) Measured and error-corrected phases of S parameters of packaged GaAs FET NE 71083 on CuClad substrate material.

all error parameters thus far evaluated. The input reflection coefficients of the cascaded transitions can be expressed as

$$\Gamma_1^{(\mu)} = S_{11}^{(\mu)} + \frac{S_{21}^{(\mu)}S_{22}^{(\nu)}}{1 - S_{22}^{(\mu)}S_{22}^{(\nu)}} \tag{19}$$

with $(\mu, \nu) = (1, 2)$ and $\mu \neq \nu$. After some mathematical manipulation the following equation for $S_{22}^{(\mu)}$ can be derived:

$$S_{22}^{(\mu)} = \frac{S_{21}^{(\mu)2}S_{22}^{(\nu)}}{2\left(S_{11}^{(\mu)} - \Gamma_1^{(\mu)}\right)} \cdot \left\{ 1 \mathop{-}\limits_{(+)} \left(1 + 4\frac{\left(\Gamma_1^{(\mu)} - S_{11}^{(\mu)}\right)\left(\Gamma_1^{(\nu)} - S_{11}^{(\nu)}\right)}{S_{21}^{(\nu)2}} \right)^{\frac{1}{2}} \right\}. \tag{20}$$

The upper negative sign is valid, as can be proved e.g. for $\Gamma_1^{(\nu)} = S_{11}^{(\nu)}$. With $|S_{22}^{(\mu)}S_{22}^{(\nu)}| \ll 1$ equation (20) can be approximated as

$$S_{22}^{(\mu)} \approx \left(\Gamma_1^{(\nu)} - S_{11}^{(\nu)}\right)/S_{21}^{(\nu)2} \tag{21}$$

with $(\mu, \nu) = (1,2)$ and $\mu \neq \nu$. This approximate formula involves an error multiplication factor of $(1 - S_{22}^{(\mu)} S_{22}^{(\nu)})$. With a typical maximum value for the magnitude of the reflection coefficients of about 0.2 for higher frequencies, the approach results in a maximum magnitude error of 0.008, or 4 percent, and a maximum phase error of 2.3°. To save computer time, (21) has been used for calibration in this paper.

With the uncertainties of all extracted parameters known, a total error estimation for the test fixture can be attempted. A critical test is given by the input reflection measurement Γ of a highly reflective load Γ_L. Starting with the bilinear transformation

$$\Gamma = S_{11} + \frac{S_{21}^2 \Gamma_L}{1 - S_{22}\Gamma_L} \tag{22}$$

with the braced transition indices neglected for clarity, the uncertainties of Γ_L dependent on the extraction errors of the scattering coefficients can be derived from the expression

$$\delta\Gamma_L = S_{21}^{-2}\left(-\delta S_{11} - \Gamma^2 \delta S_{22} + \frac{1-\Gamma}{S_{21}^2}\delta S_{21}^2\right). \tag{23}$$

With

$$\delta S_{22} = -S_{21}^{-2}\delta S_{11} \tag{24}$$

and the last term neglected because of the second-order effect, $\delta\Gamma_L$ becomes

$$\delta\Gamma_L \approx -S_{21}^{-2}\left(1 - S_{21}^{-2}\Gamma^2\right)\delta S_{11}. \tag{25}$$

An upper bound of the error magnitude is given by

$$|\delta\Gamma_L| \leqslant |S_{21}^{-2}||\delta S_{11}|\left(1 + |S_{21}^{-2}||\Gamma|\right). \tag{26}$$

With (1) and (2), $|\delta\Gamma_L|$ can be evaluated assuming typical values of $|S_{11}| = |S_{22}| = 0.15$, $|S_{21}^2| = 0.75$, $|a|^2 = 0.016$, $|\Gamma_{SL}| = 0.2$, and $|\Gamma| = 1$:

$$|\delta\Gamma_L| \leqslant \left(\frac{|\Gamma_{SL}|}{1 - |\Gamma_{SL} S_{22}|^2} + |a|^2\right)\left(1 + \frac{|\Gamma|^2}{|S_{21}^2|}\right)|S_{22}| = 0.0196. \tag{27}$$

As a result, regarding highly reflective terminations such as short and open circuits, an uncertainty of about 0.02 should be expected.

III. Experimental Results

Measurement has been carried out on a low-dielectric substrate material such as CuClad 250LX with a dielectric constant $\epsilon_r = 2.5$ and a substrate thickness $t = 0.0129$ in and on the high-dielectric substrate material DiClad 810

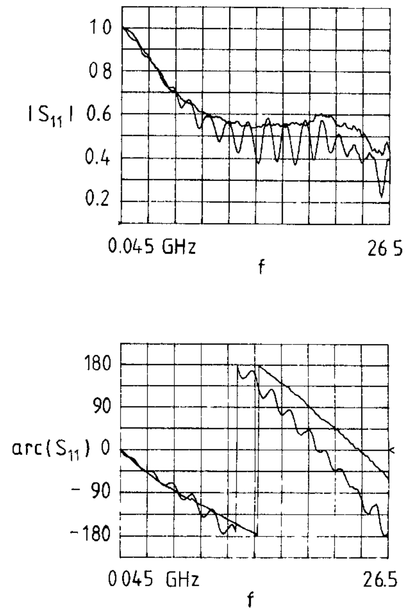


Fig. 7. Measured and error-corrected reflection coefficient S_{11} of packaged GaAs FET NE 71083 on DiClad substrate material.

with a dielectric constant of 10.2 and a thickness of 0.025 in. The latter was specially chosen as a preliminary substitute for the conventional rigid Al_2O_3 ceramics.

Fig. 5 shows typical results of the measured and error-corrected frequency-dependent reflection coefficients of the open and short circuit on CuClad material used for calibration. It should be emphasized that the individual open and short standards are initially unknown regarding their reflection coefficients. Only the mixed term $(\Gamma_s^{-1} - \Gamma_o^{-1})$ has been used for error parameter extraction. Therefore the smoothness of the amplitude and phase traces of the plotted curves gives information on the quality of calibration. The ripples of the error-corrected magnitude of the reflection coefficients can be found to be approximately 0.025, which agrees well with the derived fixture uncertainty under consideration of any residual errors due to the calibrated analyzer system. The plotted error-corrected phase curves of the open and short show a small ripple range of only 2° within the discussed frequency range.

Fig. 6(a) shows the measured and error-corrected amplitude of the scattering parameters of a packaged 0.3 μm MESFET of the NE 71083 type using CuClad substrate material. The calculated de-embedded smooth traces indicate the high-quality performance of the procedure. Fig. 6(b) shows the comparison of the measured and de-embedded phases of the S parameters. The electrical delay of the NA had been adjusted for the measured phases to compensate the phase shift of the transition for clearer representation. Fig. 7 shows the scattering coefficient S_{11} of the same transistor using high-dielectric DiClad. The different results of the de-embedded data seem to confirm a severe influence of the microstrip substrate on the S parameters of the device under test.

IV. Conclusion

In this paper a calibration technique is proposed that yields a highly accurate characterization of coaxial-to-microstrip transitions in a broad-band frequency range from 45 MHz up to 26.5 GHz. The method has been tested with the HP 8510A network analyzer. The accuracy attained is based on the fundamental idea that low-value transition parameters should be derived only from low-value measured reflection terminations. Physical open-circuit and short-circuit standards could be applied to the extraction of high-value transmission coefficients of the transitions. The high performance of the calibration approach discussed was demonstrated by comparing the measured and error-corrected scattering parameters of a packaged GaAs FET in various microstrip media.

Acknowledgment

The authors wish to thank M. Novotny for his help in providing the experimental data, and they are very much obliged to the Fernmeldetechnisches Zentralamt (FTZ)/Darmstadt, which supported the instrumentation.

References

[1] R. L. Vaitkus, "Uncertainty in the values of GaAs MESFET equivalent circuit elements extracted from measured two-port scattering parameters," presented at 1983 IEEE Cornell Conference on High Speed Semiconductor Devices and Circuits, Cornell Univ., Aug. 1983.
[2] P. R. Shepherd and P. Daly, "Modeling and measurement of microstrip transmission-line structure," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1501–1506, Dec. 1985.
[3] 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, Dec. 1986.
[4] R. F. Bauer and P. Penfield, "De-embedding and unterminating," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 282–288, Mar. 1974.
[5] E. W. Strid, "26 GHz wafer probing for MMIC development and manufacture," *Microwave J.*, pp. 71–82, Aug. 1986.
[6] D. Harvey, "A lumped coplanar to microstrip transition model for de-embedding *S*-parameters measured on GaAs wafers," in *29th ARFTG Conf. Dig.*, June 1987, pp. 204–217.
[7] H. V. Shurmer, "Calibration procedure for computer-corrected *S* parameter characterisation of devices mounted in microstrip," *Electron. Lett.*, vol. 9, no. 14, pp. 323–324, July 12, 1973.
[8] L. Dunleavy and P. B. Katehi, :"Eliminate surprises when de-embedding microstrip launches," *Microwaves & RF*, pp. 117–122, Aug. 1987.
[9] E. F. da Silva and M. K. McPhun, "Calibration techniques for one port measurement," *Microwave J.*, pp. 97–100, June 1978.
[10] M. Moser, "Entwurf und Aufbau einer Messfassung in Mikrostreifenleitungstechnik zur messtechnischen Untersuchung der *S*-Parameter von Mikrowellen-FETs," M. Eng. thesis, High Frequency Eng., University of Kassel, 1986 (unpublished).
[11] J. Staudinger, "MMIC tests improved with standards on chip," *Microwaves & RF*, pp. 107–114, Feb. 1987.
[12] L. A. Glasser, "An analysis of microwave de-embedding errors," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 379–380, May 1978.
[13] N. R. Franzen and R. A. Speciale, "A new procedure for system calibration and error removal in automated *S*-parameter measurements," in *Proc. 5th European Microwave Conf.*, Sept. 1975.
[14] G. F. Engen and C. A. Hoer, "Thru-reflect-line: An improved technique for calibrating the dual six-port reflectometer automatic network analyzer," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 987–993, Dec. 1979.
[15] C. A. Hoer, "Some questions and answers concerning air lines as impedance standards," in *29th ARFTG Conf. Dig.*, June 1987, pp. 161–173.
[16] B. Bianco, M. Parodi, S. Ridella, and F. Selvaggi, "Launcher and microstrip characterization," *IEEE Trans. Instrum. Meas.*, vol. IM-25, pp. 320–323, Dec. 1976.
[17] M. Pervere and J. Barr, "Advances in vector error correction," in *29th ARFTG Conf. Dig.*, June 1987, pp. 113–129.
[18] G. Kompa, M. Schlechtweg, and F. van Raay, "Novel redundant calibration procedure for computer-corrected *S*-parameter characterization of microstrip components," presented at Microwave Congress at the 8th Int. Congress LASER 87, Munich, June 1987.
[19] W. Kruppa and K. F. Sodomsky, "An explicit solution for the scattering parameters of a linear two-port measured with an imperfect test set," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 122–123, Jan. 1971.
[20] I. Kasa, "A circle fitting procedure and its error analysis," *IEEE Trans. Instrum. Meas.*, vol. IM-25, pp. 8–14, Mar. 1976.
[21] M. Berman and P. I. Somlo, "Efficient procedures for fitting circles and ellipses with application to sliding termination measurements," *IEEE Trans. Instrum. Meas.*, vol. IM-35, pp. 31–35, Mar. 1986.

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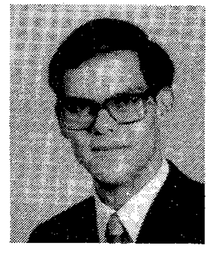

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