Direct Calibration and Measurement of Microstrip Structures on Gallium Arsenide

PETER R. SHEPHERD, MEMBER, IEEE, AND ROGER D. POLLARD, MEMBER, IEEE

Abstract —**Measurements on coupled microstrip line structures on gallium arsenide (GaAs) are described which use a direct launch and calibration standards in the same microstrip medium. Novel techniques are adopted to ensure repeatability and overcome problems of fragility of the substrates. Measurements of the S-parameters show good agreement with theoretical analysis. The theoretical derivation of the S-parameters is summarized in an appendix.**

I. Introduction

WITH INCREASING INTEREST in microwave integrated circuits (MIC's) on GaAs, as well as other new ceramic and plastic substrates, there is a need for accurate measurement techniques in order to characterize both the materials and circuits. The most popular and useful transmission-line medium on these circuits is microstrip, so it is necessary for the circuit designer to have good models for this transmission medium and its associated structures. Such models are often derived empirically from measurements performed on test structures, and it follows that measurement techniques are required which are capable of accurate characterization in the microstrip medium. Microwave characterization of circuits and devices is most often performed using a vector network analyzer, but to obtain accurate results this equipment must be calibrated using "standards," i.e., devices with known response. To avoid the additional uncertainty due to the use of adapters, it is necessary that such calibration standards be fabricated in the same transmission-line medium as the circuit to be measured. For characterization of MIC's on GaAs, an "on-circuit" probing system has been demonstrated [1] which utilizes calibration standards in coplanar waveguide (CPW). CPW has the advantages that the test signals can be launched easily from the source onto the circuit and good calibration standards can readily be fabricated. However, most MIC's necessarily employ microstrip as their transmission-line medium and it is preferable to have a measurement system in this medium. The development of such a system is the basic objective of this work.

Of specific interest is the coupling between microstrip lines, and a problem in GaAs in particular is that of spurious coupling between different parts of a circuit which may cause unexpected degradation of the circuit performance. It is necessary that these effects be well understood in order to reduce the spurious coupling or to be able to design around its effects. It was these types of structures, therefore, that were examined in the experimental part of this work.

Manuscript received March 29, 1986; revised June 18, 1986. This work was supported in part by the Procurement Executive (DCVD), U.K. Ministry of Defence.

P. R. Shepherd was with the Department of Electrical and Electronic Engineering, University of Leeds, England. He is now with the School of Electrical Engineering, University of Bath, Bath BA2 7AY, England.

R. D. Pollard is with the Department of Electrical and Electronic Engineering, University of Leeds, Leeds LS2 9JT, England.

IEEE Log Number 8610553.

II. Measurement Techniques

A measurement fixture has been developed (Fig. 1) for direct calibration and measurement of microstrip structures. Fig. 1(a) shows a cutaway 3-D drawing of the fixture showing a microstrip substrate fitted inside. Fig. 1(b) shows section and plan views of the fixture for added clarity. The fixture has provision for up to four measurement ports, one on each side, and the transition from coaxial line to microstrip is achieved with SMA tab-type launchers. The fixture has an adjustable ground plane, which serves two purposes. Firstly, the fixture is capable of accommodating substrates of arbitrary thickness; secondly, a good short circuit can be created in the absence of a substrate by bringing the ground plane into contact with the launching tabs. It is this short circuit which is used to define accurately the reference plane for the measurement. The problem with an adjustable ground plane is that of achieving a good electrical contact between that ground plane and the side walls of the fixture; this can cause problems of spurious resonances, which were evident in earlier versions of the fixture. The solution is to have slits at the corners of the fixture and a beveled pinch collar to grip the four walls together by adjusting the tightening screw once the circuit substrate is in position.

The standards used for network analyzer calibration are short and open circuits, matched loads, and transmission lines of known length and characteristic impedance. Good-quality standards are readily available in coaxial and waveguide media, but the realization of good-quality standards in other media can present problems. In microstrip, for example, short circuits to the ground plane can be formed by using via hole technology or wraparound connections at the edge of the substrate. These structures may have parasitic inductances and loss which may have to be taken into account in the calibration procedure. Resistive loads using cermet or chip devices are also of dubious quality. Open circuits are easy to fabricate, but again there are parasitic effects associated with the capacitance at the end of the line and also radiation losses. Good-quality transmission lines may also be realized, and

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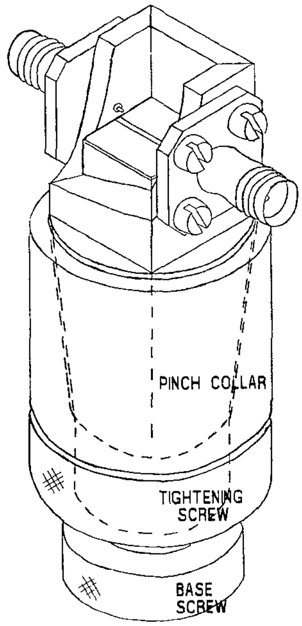
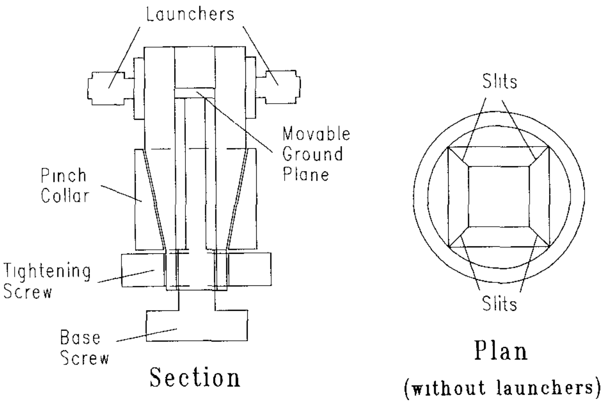


(a)



(b)

Fig. 1. Measurement fixture. (a) Cutaway view. (b) Section and plan views.

calibration routines have been devised which can take account of the losses, dispersion, and characteristic impedance of the lines [2].

With the fabrication technology available, it was decided to base the reflection calibration procedure on microstrip open circuits at the ends of lengths of transmission lines (offset opens). In addition, the short circuit realized by bringing the movable ground plane into contact with the launching tabs was used to define accurately the reference plane of measurement. In addition, a microstrip through line was used in the transmission calibration. The end effects of the microstrip open circuits are not assumed but are taken into account by using a semi-redundant technique involving an extra calibration standard. The reflection calibration procedure is summarized in Fig. 2. The end effect of the open circuit is treated as an unknown reflection coefficient Γ, and this value, as well as the error terms, is derived by measuring the four reflection coefficients ρ_s, ρ_1, ρ_2, and ρ_3 and solving the four resulting

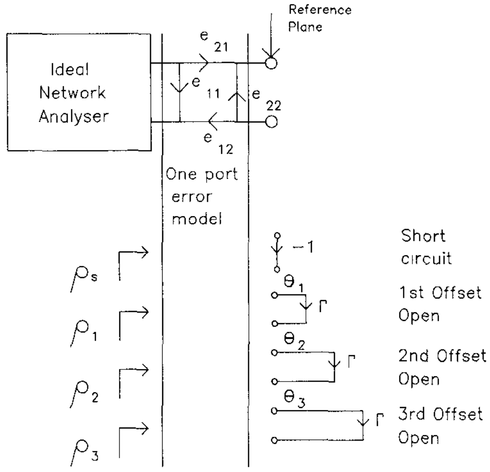


Fig. 2. Reflection calibration technique

equations

$$\rho_s = e_{11} - \frac{e_{21}e_{12}}{1+e_{22}} \tag{1}$$

$$\rho_1 = e_{11} + \frac{e_{21}e_{12}\Gamma e^{-j2\theta_1}}{1-e_{22}\Gamma e^{-j2\theta_1}} \tag{2}$$

$$\rho_2 = e_{11} + \frac{e_{21}e_{12}\Gamma_e^{-j2\theta_2}}{1-e_{22}\Gamma e^{-j2\theta_2}} \tag{3}$$

$$\rho_3 = e_{11} + \frac{e_{21}e_{12}\Gamma e^{-j2\theta_3}}{1-e_{22}\Gamma e^{-j2\theta_3}}. \tag{4}$$

One drawback with using a set of offset open circuits is that the calibration is inherently inaccurate at certain frequencies. For the set of calibration standards used for these measurements, the accuracy deteriorates below 2 GHz and above 10 GHz. These frequencies could be covered by using an alternative set of standards. This measurement technique has been used successfully on alumina substrates [3] but the important step here has been its application to GaAs. The overwhelming problem is the fragility of the GaAs substrates, since considerable contact pressure is required to obtain a good and, more importantly, repeatable contact between the tab and the circuit. This pressure will invariably fracture normal unmounted GaAs substrates. The solution adopted here is to cement the substrates down onto metal shims using a conductive adhesive to make the structure sufficiently robust. There remains the problem of the actual launching point, and here it is found that the metal tabs tend to crush the substrate, which disintegrates with the repeated pressure, resulting in a poor, unrepeatable contact as well as limited life for the calibration standards. This problem has been overcome by cementing a thin sliver of conductive rubber onto the underside of the tab to spread the pressure of the contact. A thin layer of this rubber is also placed between the substrate and the ground plane of the fixture to counteract variations in flatness of these and to ensure good electrical contact over the whole area of the

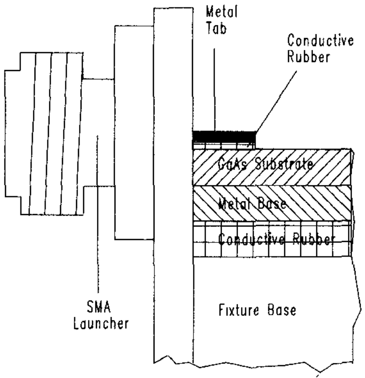


Fig. 3. Direct launching onto GaAs.

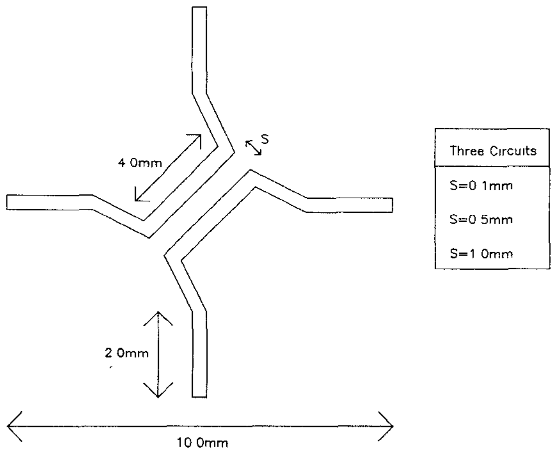


Fig. 4. Dimensions of coupled line structures.

ground plane. The cross section of the overall launching structure is shown in Fig. 3.

Using the measured results from a large number of contacts between launcher and line, the repeatability of the contact has been estimated as better than ± 0.2 dB and $\pm 1°$ at 2 GHz, ± 0.3 dB and $\pm 2°$ at 10 GHz, and ± 0.3 dB and $\pm 3°$ up to 18 GHz.

III. Results

A series of test structures was fabricated on 400-μm-thick semi-insulating GaAs substrates. The relevant dimensions of the microstrip lines are shown in Fig. 4. The coupling and isolation of the structures were measured over the frequency range 2 to 10 GHz. The results of the magnitude of the coupling measurements are shown in Fig. 5. The error limits on the measured results were derived solely from repeatability data and so represent random errors but not systematic errors. The coupling phase for the 0.1-mm-spaced lines are shown in Fig. 6. The corresponding results for the isolation are shown in Figs. 7 and 8; the results are shown for the 1.0-mm-spaced structure. These results are compared with the theoretical S-parameters derived from an analysis of the microstrip structure. The S-parameters are derived from the voltages and currents at the four ports of the structure, which are in turn obtained from a superposition of the even and odd modes of propagation. The derivation of these S-parameters is summarized in the

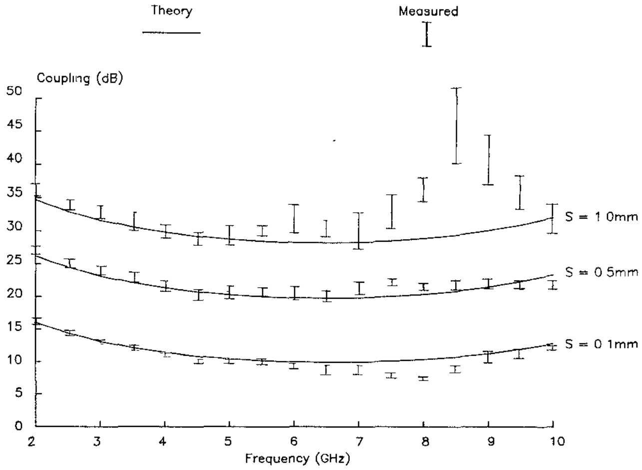


Fig. 5. Theoretical and measured coupling of coupled lines on GaAs.

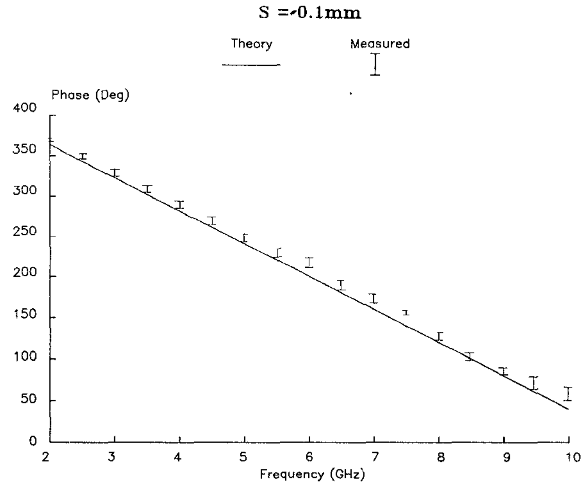


Fig. 6. Theoretical and measured coupling phase of coupled lines on GaAs.

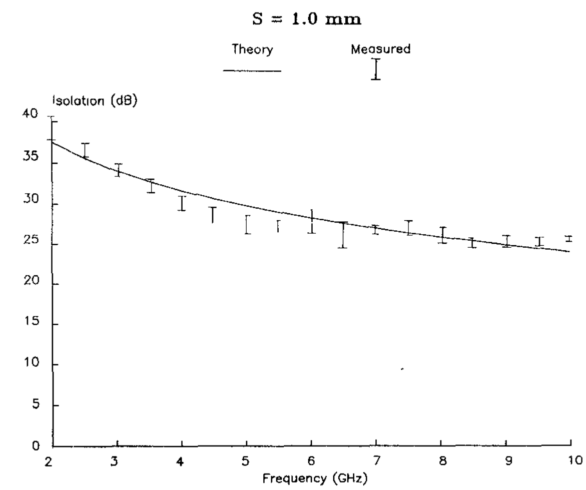


Fig. 7. Theoretical and measured isolation of coupled lines on GaAs

Appendix. The characteristic impedances and effective dielectric constants of these two modes are calculated using the routine of Bryant and Weiss [4], while the losses are calculated using [5].

There is good agreement between theory and measurements, particularly in the lower part of the frequency

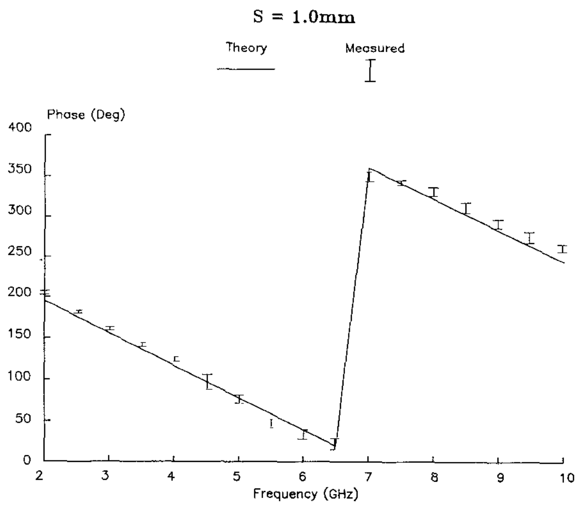


Fig. 8. Theoretical and measured isolation phase of coupled lines on GaAs.

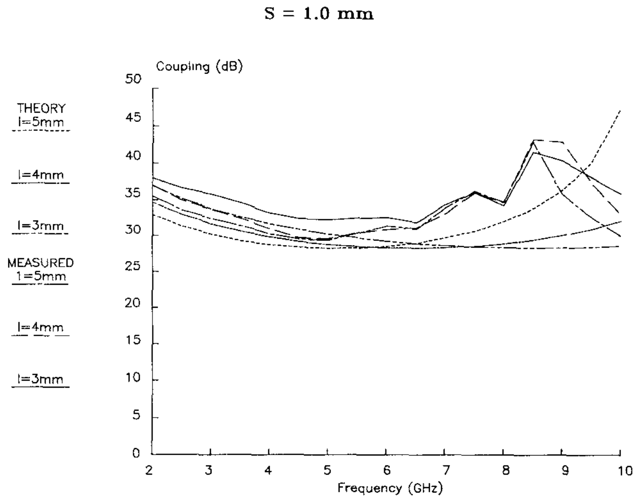


Fig. 9. Theoretical and measured coupling of coupled lines on GaAs.

range. Deviation at higher frequency may be due to the fact that the analysis assumes a quasi-TEM mode of propagation and no account is taken of dispersion effects. There is an apparent resonance at 8.5 GHz for the 1.0-mm-spaced structure which is not predicted by the theory. To investigate this further, more coupled structures were fabricated with the same 1.0-mm spacing, but with coupled lengths of 3 mm and 5 mm. The magnitude of the coupling results for these 1.0-mm-spaced structures are shown in Fig. 9, and it can be seen that these also exhibit the discrepancy around 8.5 GHz. This demonstrates that the effect is not a function of the length of coupled line section, but does appear to be related to the spacing between the lines, as this behavior has not been seen in the 0.1-mm- or 0.5-mm-spaced structures. The effect is thought to be a fixture resonance, the resonant mode only being excited for the particular case of the 1.0-mm-spaced structure. Alternatively, it may be caused by the launchers, since similar effects at 8.5 GHz have been noted by other workers [6] employing fixtures of different dimensions from the one described here.

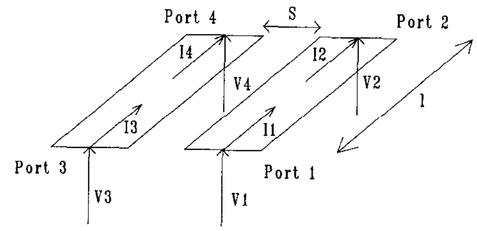


Fig. 10. Layout of a pair of coupled microstrip lines.

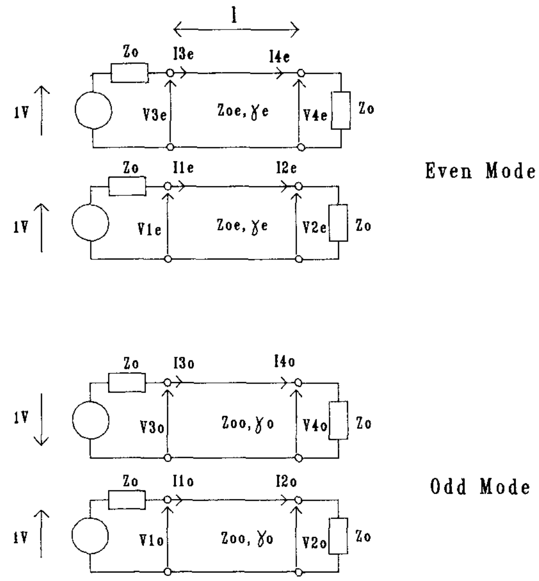


Fig. 11. Odd- and even-mode circuit representations.

IV. Conclusions

A measurement technique has been presented for the direct calibration and measurement of microstrip structures and has been applied to GaAs circuits. Results have been presented for coupled line structures and the authors believe that these represent the first published results of a calibration and measurement performed *directly in microstrip on GaAs*. The technique provides a basis for improved characterization, not only of GaAs structures and active-device circuits but also in the broader field of a wide range of microwave ceramic and plastic substrates which are becoming available and which could achieve widespread use if their properties were well characterized and reliable techniques were available for measurements of circuits employing these materials.

Appendix

Derivation of the S-Parameters of a Pair of Coupled Microstrip Lines

A pair of coupled microstrip lines, as shown in Fig. 10, is a four-port configuration which is purely symmetrical and so can be characterized by a set of four S-parameters, S_{11}, S_{21}, S_{31}, and S_{41}. The remaining elements of the 4×4 S-matrix will be equivalent to one of these parameters.

The analysis of this type of structure is performed by considering the odd- and even-mode excitations of the structure. Each mode has an associated characteristic impedance and propagation constant which are dependent on the physical dimensions of the structure and the dielectric constant of the substrate. These two modes are shown diagrammatically in Fig. 11, which also shows the terminal voltages and currents at each port for the two modes [7]. The total terminal voltages and currents are derived from superposition of the two cases and taking into account the symmetry of the structure, so that

$$V_1 = V_{1e} + V_{1o} \tag{A1}$$
$$V_2 = V_{2e} + V_{2o} \tag{A2}$$
$$V_3 = V_{1e} - V_{1o} \tag{A3}$$
$$V_4 = V_{2e} - V_{2o} \tag{A4}$$
$$I_1 = I_{1e} + I_{1o} \tag{A5}$$
$$I_2 = I_{2e} + I_{2o} \tag{A6}$$
$$I_3 = I_{1e} - I_{1o} \tag{A7}$$
$$I_4 = I_{2e} - I_{2o}. \tag{A8}$$

We can now write down the chain matrix for the even- and odd-mode cases.

$$\begin{bmatrix} V_{1e} \\ I_{1e} \end{bmatrix} = \begin{bmatrix} \cosh\gamma_e l & Z_{oe}\sinh\gamma_e l \\ Y_{oe}\sinh\gamma_e l & \cosh\gamma_e l \end{bmatrix}\begin{bmatrix} V_{2e} \\ I_{2e} \end{bmatrix} \tag{A9}$$

$$\begin{bmatrix} V_{1o} \\ I_{1o} \end{bmatrix} = \begin{bmatrix} \cosh\gamma_o l & Z_{oo}\sinh\gamma_o l \\ Y_{oo}\sinh\gamma o l & \cosh\gamma_o l \end{bmatrix}\begin{bmatrix} V_{2o} \\ I_{2o} \end{bmatrix} \tag{A10}$$

where

$$Y_{oe} = 1/Z_{oe} \tag{A11}$$

and

$$Y_{oo} = 1/Z_{oo} \tag{A12}$$

and we have at the terminals

$$V_{2e} = Z_o I_{2e} \tag{A13}$$
$$V_{2o} = Z_o I_{2o} \tag{A14}$$
$$V_{1e} + Z_o I_{1e} = 1 \tag{A15}$$
$$V_{1o} + Z_o I_{1o} = 1. \tag{A16}$$

Expanding the matrices and substituting in the terminal conditions, we get

$$V_{1e} = \frac{Z_{oe}Z_o\cosh\gamma_e l + Z_{oe}^2\sinh\gamma_e l}{(Z_o^2 + Z_{oe}^2)\sinh\gamma_e l + 2Z_oZ_{oe}\cosh\gamma_e l} \tag{A17}$$

$$I_{1e} = \frac{Z_{oe}\cosh\gamma_e l + Z_o\sinh\gamma_e l}{(Z_o^2 + Z_{oe}^2)\sinh\gamma_e l + 2Z_oZ_{oe}\cosh\gamma_e l} \tag{A18}$$

$$V_{2e} = \frac{Z_oZ_{oe}}{(Z_o^2 + Z_{oe}^2)\sinh\gamma_e l + 2Z_oZ_{oe}\cosh\gamma_e l} \tag{A19}$$

$$I_{2e} = \frac{Z_{oe}}{(Z_o^2 + Z_{oe}^2)\sinh\gamma_e l + 2Z_oZ_{oe}\cosh\gamma_e l} \tag{A20}$$

$$V_{1o} = \frac{Z_{oo}Z_o\cosh\gamma_o l + Z_{oo}^2\sinh\gamma_o l}{(Z_o^2 + Z_{oo}^2)\sinh\gamma_o l + 2Z_oZ_{oo}\cosh\gamma_o l} \tag{A21}$$

$$I_{1o} = \frac{Z_{oo}\cosh\gamma_o l + Z_o\sinh\gamma_o l}{(Z_o^2 + Z_{oo}^2)\sinh\gamma_o l + 2Z_oZ_{oo}\cosh\gamma_o l} \tag{A22}$$

$$V_{2o} = \frac{Z_oZ_{oo}}{(Z_o^2 + Z_{oo}^2)\sinh\gamma_o l + 2Z_oZ_{oo}\cosh\gamma_o l} \tag{A23}$$

$$I_{2o} = \frac{Z_{oo}}{(Z_o^2 + Z_{oo}^2)\sinh\gamma_o l + 2Z_oZ_{oo}\cosh\gamma_o l}. \tag{A24}$$

For the purpose of deriving the S-parameters from the terminal voltages and currents, we use the power waves given by

$$a_i = \frac{V_i + Z_o I_i}{2\sqrt{Z_o}} \tag{A25}$$

$$b_i = \frac{V_i - Z_o I_i}{2\sqrt{Z_o}} \tag{A26}$$

where V_i is the ith terminal voltage and I_i is the current flowing *into* the ith terminal. The S-parameters are then given by

$$S_{11} = \frac{b_1}{a_1} \tag{A27}$$

$$S_{21} = \frac{b_2}{a_1} \tag{A28}$$

$$S_{31} = \frac{b_3}{a_1} \tag{A29}$$

$$S_{41} = \frac{b_4}{a_1} \tag{A30}$$

when $a_2, a_3, a_4 = 0$ in all cases.

If we adopt the shorthand notation

$$S_o = \sinh\gamma_o l \tag{A31}$$
$$C_o = \cosh\gamma_o l \tag{A32}$$
$$S_e = \sinh\gamma_e l \tag{A33}$$
$$C_e = \cosh\gamma_e l \tag{A34}$$

and also

$$A = S_e(Z_o^2 + Z_{oe}^2) + 2Z_oZ_{oe}C_e \tag{A35}$$
$$B = S_o(Z_o^2 + Z_{oo}^2) + 2Z_oZ_{oo}C_o \tag{A36}$$

then using all the previous equations gives

$$S_{11} = \frac{S_e(Z_{oe}^2 - Z_o^2)B + S_o(Z_{oo}^2 - Z_o^2)A}{2AB} \tag{A37}$$

$$S_{21} = \frac{2Z_oZ_{oe}B + 2Z_oZ_{oo}A}{2AB} \tag{A38}$$

$$S_{31} = \frac{S_e(Z_{oe}^2 - Z_o^2)B + S_o(Z_o^2 - Z_{oo}^2)A}{2AB} \tag{A39}$$

$$S_{41} = \frac{2Z_oZ_{oe}B - 2Z_oZ_{oo}A}{2AB}. \tag{A40}$$

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Peter R. Shepherd (M'83) was born in Exeter, England, in 1957. He received the BEng degree in electronic engineering from the University of Sheffield in 1978.

He then worked for Plessey Semiconductors, Plymouth, England, manufacturing silicon VLSI circuits. In 1982 he completed the Ph.D. degree at Portsmouth Polytechnic, based on research into microwave oscillator design. He then joined the Department of Electrical and Electronic Engineering, University of Leeds, where he was employed as a research fellow investigating the modeling and measurement of passive components for GaAs monolithic microwave integrated circuits, funded by the U.K. Ministry of Defence. In September 1986, he joined the School of Electrical Engineering, University of Bath, as a lecturer.

Dr. Shepherd is an associate member of the Institution of Electrical Engineers (U.K.).

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.

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. Experience has included work on microwave transistor circuits, 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.