Full-Wave Analysis of Aperture-Coupled Microstrip Lines

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***Abstract*—Two methods are presented for the analysis of aperture-coupled microstrip lines. Assuming a quasi-TEM traveling wave incident on the feeding line, an expression for the wave on the coupled line is derived. First, the moment method is used and the current on the coupled line is represented by a traveling wave propagating away from the slot. In the second method, the reciprocity theorem is applied to the coupled line. An equivalent circuit is derived and the S parameters are computed. Theoretical results are verified with measurements.**

I. Introduction

This paper presents a full-wave analysis and an equivalent circuit for the problem of two parallel microstrip lines on parallel substrates coupled by a small rectangular slot in the common ground plane [1]. In the most general case, this circuit can be represented as a four-port network as shown in Fig. 1. In operation, an input signal applied to port 1 is partially coupled to the top microstrip line, and this coupled power divides equally between ports 3 and 4, with a 180° phase shift. The remainder of the input goes out the through line of port 2. By terminating one or two of the microstrip lines with stubs, the four-port network can be reduced to a three-port or a two-port coupler. By proper design, these circuits can be made to couple all of the input power to the output port (or ports) on the top substrate. If more than one coupling aperture is used, directionality between the output ports can be obtained.

The two-, three-, and four-port forms of this circuit find application in a variety of systems that require power division or coupling circuits in planar form, including planar antenna feed networks and noncontacting RF interconnects between multilayer integrated circuits. The 180° phase shift between the outputs of ports 3 and 4 can be useful in balanced mixer circuits, and in some types of antenna feed networks (e.g., a sequentially phased feed for circular polarization). In fact, ports 3 and 4 have a perfect equal-amplitude, opposite-polarity property over a very wide bandwidth owing to the nature of the slot feed.

To date, there has been little analysis on this problem, although related problems have been treated. These include the use of aperture coupling between a rectangular waveguide and a microstrip line or a stripline [2], [3] and a resonant cavity aperture coupled to a microstrip line [4]. In [5], a full-wave analysis of an aperture-coupled microstrip antenna was given; the present work is an extension of [5].

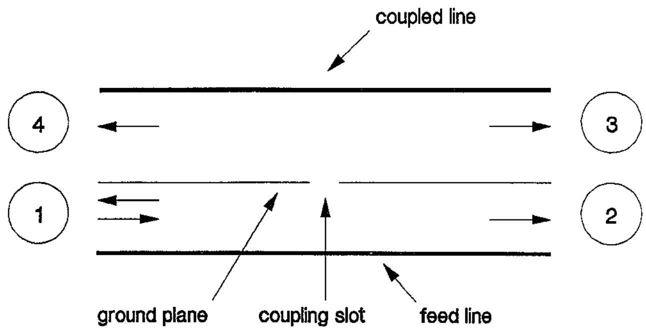

Fig. 1. Aperture-coupled microstrip lines.

In this paper, the four-port geometry in Fig. 2 is analyzed using two methods. In both methods, the reciprocity technique developed in [5] is used to find the reflected and transmitted waves on the feed line in terms of the coupling slot field; the slot field is represented in terms of a single piecewise-sinusoid (PWS) mode with an unknown voltage amplitude. In what we refer to as the moment method solution, the unknown current on the coupled line is expanded in terms of a traveling wave mode, and the electric field integral equation is invoked to determine the amplitude of this current and the unknown slot mode amplitude. In the method we refer to as the reciprocity solution, we apply reciprocity to the coupled line in the same way as was done for the feed line, to obtain an equation for the unknown modal fields of the coupled line in terms of the slot field. In both cases, continuity of the tangential magnetic field in the aperture is enforced. Numerical results from these two methods are virtually identical, but the reciprocity method is computationally more efficient. Both models account for power loss by radiation from the slot, which should be negligible in a practical design.

As shown in Fig. 2, the geometry of this problem consists of two back-to-back parallel microstrip lines. The feed substrate has a dielectric constant of ϵ_f and a thickness d_f; the width of the feed line is w_f. The top substrate has a dielectric constant of ϵ_c and a thickness d_c; the width of the coupled line is w_c. The coupling slot

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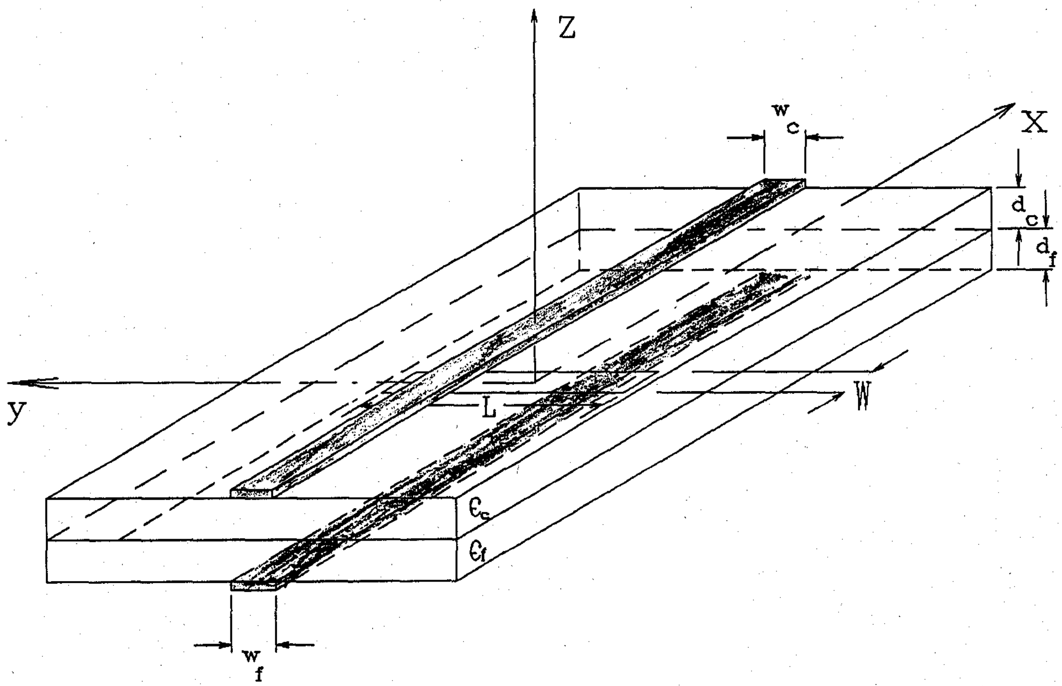

Fig. 2. The geometry of two aperture-coupled microstrip lines.

is oriented perpendicular to the microstrip lines, and has a length L and a width W. Section II describes the theory of the moment method solution and the reciprocity method solution. Section III discusses the two methods and presents an equivalent circuit. An example comparing calculated results with measured data is given.

II. Analysis

A. Moment Method Solution

This analysis follows the method and assumptions of [5], and the basic relationships relevant to the case studied here are brought without derivation (the letter f will be used for quantities related to the feed line and the letter c will denote quantities related to the coupled line). The microstrip lines are assumed to be infinitely long and propagating quasi-TEM waves with a normalized transverse modal magnetic field given by [5]

$$\boldsymbol{H} = h_y \hat{y} + h_z \hat{z}. \tag{1}$$

The modal reflection (R) and transmission (T) coefficients on the feeding line are, from [5],

$$S_{11} = R = -\frac{V_0}{2} \Delta v_f \tag{2}$$

$$S_{21} = T = 1 - R. \tag{3}$$

The electric field in the slot is assumed to be of a piecewise-sinusoidal form with amplitude V_0:

$$\boldsymbol{E}^{\text{slot}} = E^a \hat{x} = V_0 e_x^a \hat{x} \tag{4}$$

where

$$e_x^a = \begin{cases} \dfrac{\sin k_e(L/2 - |x|)}{W \sin k_e L/2} & |x| < L/2, |y| < W/2 \\ 0 & \text{elsewhere} \end{cases} \tag{5}$$

In (5) $k_e = k_0\sqrt{\epsilon_{\text{eff}}}$ is the effective wavenumber of the PWS mode, with ϵ_{eff} chosen as the average of the two substrate dielectric constants. As with [5, eq. (11)], Δv_f is the reaction between the slot field and the feeding line, representing the voltage discontinuity across the slot, and is given by

$$\Delta v_f = \iint_{s_a} e_x^a(x, y) h_y^f(x, y)\, dx\, dy \tag{6}$$

where s_a is the aperture (slot) area. In the slot at $z = 0$, the tangential magnetic field is continuous:

$$H_y^c + H_y^{ac} = H_y^f + H_y^{af} \tag{7}$$

where

- H_y^c is the tangential magnetic field caused by the coupled line at $x = 0^-$, $z = 0^+$ in the absence of the slot;
- H_y^{ac} is the tangential magnetic field from the slot for $z = 0^+$;
- H_y^f is the tangential magnetic field arising from the feeding line at $x = 0^-$, $z = 0^-$ in the absence of the slot;
- H_y^{af} is the tangential magnetic field caused by the slot for $z = 0^-$.

The magnetic fields in (7) are obtained using the appropriate Green functions [5] for a grounded dielectric slab and the equivalent magnetic currents created by the fields in the slot. Thus,

$$H_y^c = \int_{-\infty}^{\infty}\int_{-w_c/2}^{w_c/2} G_{yx}^{HJ}(x,y;x_0,y_0)J_x^c\, ds_0 \tag{8}$$

$$H_y^{ac} = \iint_{s_a} G_{yy}^{HMac}(x,y;x_0,y_0)M_y^{ac}\, ds_0 \tag{9}$$

$$H_y^f = (1-R)h_y^f \tag{10}$$

$$H_y^{af} = \iint_{s_a} G_{yy}^{HMaf}(x,y;x_0,y_0)M_y^{af}\, ds_0 \tag{11}$$

where

$$M_y^{af} = -V_0 e_x^a(x,y) \tag{12}$$

is the magnetic current for $z = 0^+$,

$$M_y^{ac} = -M_y^{af} \qquad \text{for } z = 0^- \tag{13}$$

and J_x^c is the current induced on the coupled line. Multiplying (7) by e_x^a and integrating over s_a, we get

$$V_0(Y_{af}+Y_{ac}) = (1-R)\Delta v_f + \iint_{s_a}\int_{-\infty}^{\infty}\int_{-w_c/2}^{w_c/2} G_{yx}^{HJ}(x,y;x_0,y_0)J_x^c e_x^a\, ds_0\, ds \tag{14}$$

where

$$Y_{af} = \frac{1}{V_0}\iint_{s_a} e_x^a \iint_{s_a} G_{yy}^{HMf}(x,y;x_0,y_0)M_y^{af}\, ds_0\, ds \tag{15a}$$

$$Y_{ac} = -\frac{1}{V_0}\iint_{s_a} e_x^a \iint_{s_a} G_{yy}^{HMc}(x,y;x_0,y_0)M_y^{ac}\, ds_0\, ds. \tag{15b}$$

Y_{af} is the slot admittance looking at the top substrate ($z > 0$), and Y_{ac} is the slot admittance looking at the bottom substrate ($z < 0$).

At this point, we need another equation to solve for J_x^c in (14), which can be obtained from the boundary condition imposed on the electric field. Enforcing $\boldsymbol{E}_{\tan} = 0$ on the coupled line, we obtain an integral equation relating the equivalent magnetic current source M_y^c and the electric current density J_x^c induced on the coupled line:

$$\iint_{s_c} G_{xx}^{EJc}(x,y;y_0)J_x^c\, ds_0 + \iint_{s_a} G_{xy}^{EMc}(x,y;x_0,y_0)M_y^{ac}\, ds_0 = 0. \tag{16}$$

This equation is valid for all $x, |y| < w_c$. Assuming that the dominant mode is of a quasi-TEM form, the current on the coupled line can be expanded as an entire-domain traveling-wave mode:

$$J_x^c(x,y) = I_0 f^{\text{trav}}(x) f^{\text{unif}}(y) \tag{17}$$

$$f_{\text{trav}}(x) = \begin{cases} e^{-j\beta x} & x > 0 \\ e^{+j\beta x} & x < 0 \end{cases} \tag{18a}$$

$$f_{\text{unif}}(y) = \begin{cases} \dfrac{1}{w_c} & |y| < w_c/2 \\ 0 & |y| > w_c/2. \end{cases} \tag{18b}$$

This mode is based on the line modal field and is also dictated by the continuity of H_y at $x = 0$ and the discontinuity of E_z at $x = 0$; thus it accounts for traveling waves propagating on the coupled line away from the slot. This result leads to the property of perfect antisymmetry (equal magnitude, opposite polarity) of the outcoming coupled signals over a wide bandwidth. In the region of the slot discontinuity, however, we might expect the need for some decaying current modes to represent reactive energy near the slot. As for any discontinuity, one would expect the excitation of additional currents besides the dominant mode current. The requirement for such non-TEM current modes was tested by using subsectional PWS modes, which resulted in a change of less than 10% in the equivalent admittance. This result implies that the transverse nature of the slot discontinuity does not lead to a significant excitation of non-TEM currents on the lines. If we assume only one expansion mode of the form given by (17), then (16) can be solved in a least-mean-square sense by testing with a PWS mode. The testing function used is

$$f_p^t(x) = \begin{cases} \dfrac{\sin k_e(h-|x|)}{\sin k_e h} & |x| < h \\ 0 & \text{elsewhere} \end{cases} \tag{19a}$$

$$f_u^t(y) = \begin{cases} \dfrac{1}{w_c} & |y| < w_c/2 \\ 0 & |y| > w_c/2 \end{cases} \tag{19b}$$

where $2h$ is the PWS mode length. Then (16) reduces to

$$I_0 \iint_{s_c} G_{yx}^{EJc} f_{tr}^c f_u^c f_p^t f_u^t\, ds_0 - V_0 \iint_{s_a} G_{yy}^{EMc} f_u^s f_p^s f_p^t f_u^t\, ds_0 = 0. \tag{20}$$

If we define

$$N^s = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \tilde{G}_{yy}^{EMc}(k_x,k_y) \cdot \tilde{F}_p^s(k_y)\tilde{F}_u^s(k_x)\tilde{F}_u^{*t}(k_y)\tilde{F}_p^{*t}(k_x)\, dk_x\, dk_y \tag{21}$$

$$Z = -\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \tilde{G}_{yx}^{EJc}(k_x,k_y)\tilde{F}_{tr}^c(k_x) \cdot \tilde{F}_u^c(k_y)\tilde{F}_u^{*t}(k_x)\tilde{F}_p^{*t}(k_y)\, dk_x\, dk_y \tag{22}$$

where

$f_p^s(y)$ slot field (magnetic current) y variation (PWS) (eq. (5));

$f_u^s(x)$ slot field (magnetic current) x variation (uniform) (eq. (5));

$f_p^t(x)$ testing mode x variation (PWS) (eq. (19.a));

$f_u^t(y)$ testing mode y variation (uniform) (eq. (19.b));
$f_{tr}^c(x)$ coupled line current x variation (traveling wave) (eq. (18.a));
$f_u^c(y)$ coupled line current y variation (uniform) (eq. (18b));

and $\tilde{F}_p^s$ is the Fourier transform of f_p^s, etc., then (20) reduces to

$$ZI_0 + V_0 N^s = 0. \tag{23}$$

Finally, the current coefficient is

$$I_0 = -V_0 \frac{N^s}{Z}. \tag{24}$$

We now define

$$N^c = \iint_{s_a} \int_{-\infty}^{\infty} \int_{-w_c/2}^{w_c/2} G_{yx}^{HJ}(x, y; x_0, y_0) \cdot f_{\text{trav}}(x) f_{\text{unif}}(y) e_x^a \, ds_0 \, ds. \tag{25}$$

Since by reciprocity [5],

$$G_{yx}^{HJ}(x, y; x_0, y_0) = -G_{xy}^{EM}(x, y; x_0, y_0),$$

(25) becomes

$$N^c = -\iint_{s_a} \int_{\infty}^{\infty} \int_{-w_c/2}^{w_c/2} G_{xy}^{EM}(x, y; x_0, y_0) \cdot f_{\text{trav}}(x) f_{\text{unif}}(y) e_x^a \, ds_0 \, ds. \tag{26}$$

Using (14) and (25) in (20) gives

$$V_0(Y_{af} + Y_{ac}) = (1 - R)\Delta v_f - I_0 N^c. \tag{27}$$

If we define an equivalent admittance, Y_c, as seen by the feed line,

$$Y_c = \frac{N^s N^c}{Z} \tag{28}$$

then (27) becomes

$$V_0(Y_{af} + Y_{ac}) = (1 - R)\Delta v_f - V_0 Y_c. \tag{29}$$

We can define a total equivalent admittance as

$$Y_{\text{tot}} = Y_{af} + Y_{ac} + Y_c. \tag{30}$$

Using (2) in (29) we obtain an expression for V_0 and the other parameters characterizing the circuit:

$$V_0 = \frac{-2\Delta v_f}{2Y_{\text{tot}} + \Delta v_f^2} \tag{31}$$

$$R = -\frac{V_0}{2}\Delta v_f = \frac{\Delta v_f^2}{2Y_{\text{tot}} + \Delta v_f^2}. \tag{32}$$

The power output at port 3 (or 4) is

$$P_3 = |I_3|^2 Z_{0c} = |I_0|^2 Z_{0c} \tag{33}$$

and since the input power at port 1 is assumed to be 1 W [5], the coupling factor, C_p (the power ratio between port 3 and port 1), is

$$C_p = \frac{P_3}{P_1} = |I_0|^2 Z_{0c}. \tag{34}$$

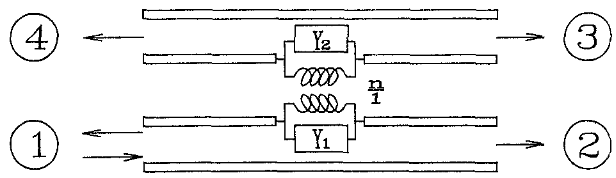

Fig. 3. The equivalent circuit of two aperture-coupled microstrip lines.

B. Solution Using Reciprocity for the Coupled Line

The reciprocity theorem was used in [5] to obtain a relationship between R, T, and V_0 for the feeding line of a microstrip fed slot. In a similar manner we can obtain a relationship between the coupled line magnetic modal field amplitude and V_0.

In order to apply reciprocity we choose two sets of fields, both of which satisfy Maxwell's equations and the boundary conditions in the region of interest. The boundary conditions applied in the plane $x = 0$ suggest the following:

Set 1—The total field of the coupled line:

$$\boldsymbol{E}_1 = \begin{cases} A\boldsymbol{E}^- & x < 0 \\ A\boldsymbol{E}^+ & x > 0 \end{cases} \qquad \boldsymbol{H}_1 = \begin{cases} A\boldsymbol{H}^- & x < 0 \\ A\boldsymbol{H}^+ & x > 0 \end{cases} \tag{35}$$

where

$$\boldsymbol{E}^{\pm} = \pm \boldsymbol{e} e^{\mp j\beta x} \qquad \boldsymbol{H}^{\pm} = \boldsymbol{h} e^{\mp j\beta x}. \tag{36}$$

Set 2—The coupled line modal fields:

$$\boldsymbol{E}_2^+ = \boldsymbol{e} e^{-j\beta x} \qquad \boldsymbol{H}_2^+ = \boldsymbol{h} e^{-j\beta x} \tag{37}$$

where $\boldsymbol{e}$ and $\boldsymbol{h}$ are the normalized modal fields of the top line subject to the normalization

$$\int_{y=-\infty}^{\infty} \int_{z=0}^{\infty} \boldsymbol{h} \times \boldsymbol{e} \, dy \, dz = 1. \tag{38}$$

Applying reciprocity for the coupled line and following the same procedure that was used in [5] for the feeding line, we obtain an expression for the amplitude of the field on the coupled line:

$$A = -\frac{V_0}{2}\Delta v_c \tag{39}$$

where Δv_c is given by an expression similar to that for Δv_f in (6) for the feeding line:

$$\Delta v_c = \iint_{s_a} e_x^a(x, y) h_y^c(x, y) \, dx \, dy. \tag{40}$$

The field for the coupled line is expressed in terms of the coupled line modal fields (using the same normalization as for the feeding line). Thus, the power carried by the coupled line is A^2. However, A is not the ratio between the amplitudes of the total fields on the coupled and field

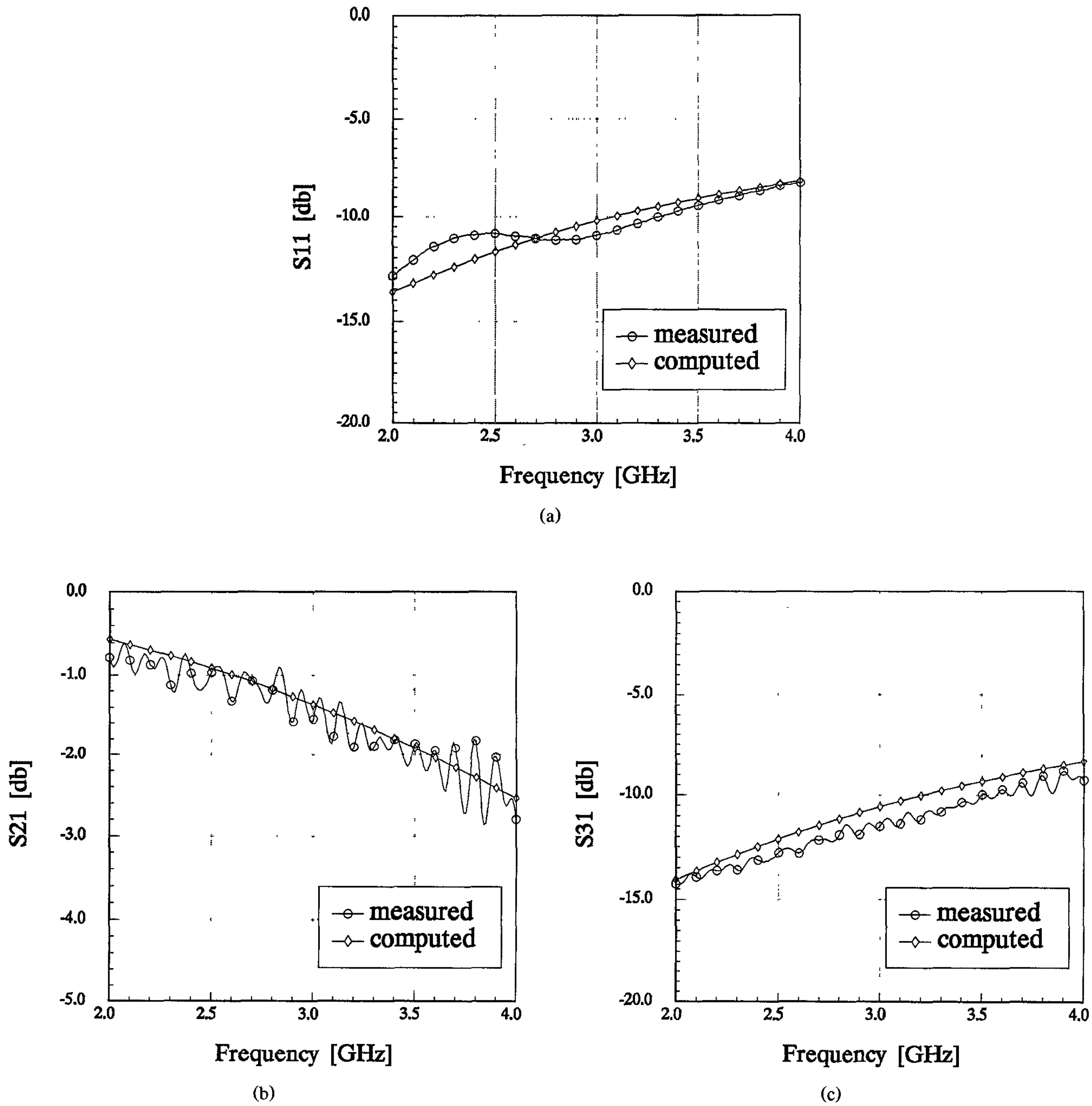

Fig. 4. Measured and calculated S parameters [dB] for two aperture-coupled microstrip lines. $w_c = w_f = 0.254$ cm, $\epsilon_c = \epsilon_f = 2.22$, $d_c = d_f = 0.0762$ cm, $L = 1.5$ cm, $W = 0.11$ cm: (a) S_{11}, (b) S_{21}, (c) S_{31}.

lines (unless the lines are identical), since the modal fields $\boldsymbol{e}$ and $\boldsymbol{h}$ are different for different lines.

The tangential magnetic field is continuous in the slot:

$$H_y^c + H_y^{ac} = H_y^f + H_y^{af} \tag{41}$$

with

$$H_y^c + H_y^{ac} = \iint_{s_a} G_{yy}^{EMc}(x, y; x_0, y_0) M_y^{ac} \, ds_0 + A h_y^c(x, y) \tag{42a}$$

$$H_y^f + H_y^{af} = \iint_{s_a} G_{yy}^{EMf}(x, y; x_0, y_0) \cdot M_y^{af} \, ds_0 + (1 - R) h_y^f(x, y) \tag{42b}$$

where M_y^{ac} and M_y^{af} are defined in (12) and (13) respectively. Multiplying (41) by e_x^a and integrating over s_a gives

$$V_0(Y_{af} + Y_{ac}) = (1 - R)\Delta v_f + A \Delta v_c \tag{43}$$

where Y_{af} and Y_{ac} are given in (15). Using (39) and (43) to solve for V_0 gives

$$V_0 = \frac{-2\Delta v_f}{2Y_e + \Delta v_f^2 + \Delta v_c^2} \qquad (Y_e = Y_{af} + Y_{ac}). \tag{44}$$

Then, using this result in (2), (3), and (39), we obtain the following expressions:

$$R = -\frac{V_0}{2}\Delta v_f = \frac{\Delta v_f^2}{2Y_e + \Delta v_f^2 + \Delta v_c^2} \tag{45}$$

$$T = 1 - R = \frac{2Y_e + \Delta v_c^2}{2Y_e + \Delta v_c^2 + \Delta v_f^2} \tag{46}$$

$$A = -\frac{V_0}{2}\Delta v_c = \frac{\Delta v_f \Delta v_c}{2Y_e + \Delta v_c^2 + \Delta v_f^2}. \tag{47}$$

Finally the coupling coefficient S_{31} is

$$S_{31} = A\sqrt{\frac{Z_{0f}}{Z_{0c}}} = \sqrt{C_p} \tag{48}$$

where Z_{0f} and Z_{0c} are the characteristic impedances of the feed and coupled lines, respectively.

III. Discussion

Following the formulation in [5], the problem consists in solving for four unknowns:

1) R the reflection coefficient on the feeding line;
2) T the transmission coefficient on the feeding line;
3) V_0 the amplitude of the electric field in the slot;
4) C_p or A the excitation of the coupled line.

In addition, the moment method solves for the current on the coupled line, while the reciprocity method finds the magnetic field on the coupled line. In the first method, the boundary condition $E_{\tan} = 0$ is enforced on the coupled line; the current is represented by one mode (18) and its amplitude is found in terms of V_0 by solving the moment method equation (16). In the second method, the reciprocity theorem is applied to the coupled line relating the amplitude of the magnetic field on the coupled line to V_0 (39). The two methods given very similar results.

It appears that the use of only a TEM mode in the moment method could lead to some error if in reality some other (non-TEM) modes are excited. For this particular geometry, however, this does not happen. Intuitively, the slot field, being parallel to the transmission line axis, creates a reflected wave which is the traveling mode used. This is the case even for lines of different widths provided the ratio w_c / w_f is smaller than the ratio L / W. If the slot were oriented obliquely with respect to the couple line, or the lines were not parallel, additional modes would probably be needed in order to obtain good results, and in this case the reciprocity method would probably be less accurate. However, the reciprocity method is computationally more efficient: the CPU time required is about 60% less than that needed by the moment method. On a Cyber 850 computer, the solution based on the moment method required 117 s, while the solution based on the reciprocity method required only 47 s.

The equivalent circuit (see derivation in the Appendix) is shown in Fig. 3. As shown in [5], the slot discontinuity appears as a series impedance to the microstrip line. That means that the feed line as well as the coupled line has a series element accounting for the slot. The coupling effect between the two lines suggests the need for an ideal transformer. For example, the case in which the two lines are identical would correspond to a turns ratio equal to 1; the magnetic field created by the slot as a result of the incident wave on the feed line would be identical on the two lines.

A four-port coupler was built and tested to verify the theory. Fig. 4 shows a comparison between the measured S parameters and those computed using the moment method. The two substrates were fabricated with two separate ground planes; this might explain the ripples appearing in the measured data if the two ground planes were not in perfect contact. The surface mount connectors used could also account for some deviation between the theoretical and experimental results, especially for S_{11}. The values obtained using the reciprocity method agreed to within 0.1 dB.

IV. Conclusion

Two methods have been presented for the analysis of two parallel aperture-coupled microstrip lines. Both the moment method and the reciprocity method make use of the exact Green functions and produce results that are in very close agreement. A four-port coupler has been built and tested to verify the theory. An equivalent circuit, which should aid in the design of practical coupled line circuits, has been developed.

Appendix

The induction of current on the coupled line suggests the use of an ideal transformer to couple the two lines. From the case of a microstrip line fed slot we infer the presence of two normalized admittances, $\bar{Y}_1$ and $\bar{Y}_2$ (see Fig. 3). These admittances (each normalized to the respective line characteristic admittance) represent the slot self-admittance looking toward the coupled and feed lines respectively. Assuming all ports are matched, the normalized admittance seen by the ideal transformer on the feed line side is

$$\bar{Y}_{12} = n^2(\bar{Y}_2 + 0.5) \tag{A1}$$

where the turns ratio of the transformer is

$$n \triangleq \frac{n_c}{n_f}. \tag{A2}$$

The normalized series admittance on the feed line is

$$\bar{Y}_s = n^2(\bar{Y}_2 + 0.5) + \bar{Y}_1. \tag{A3}$$

The reflection coefficient is

$$R = \frac{\bar{Z}_{in} - 1}{\bar{Z}_{in} + 1} = \frac{\bar{Z}_s}{\bar{Z}_s + 2} = \frac{1}{1 + 2\bar{Y}_s}. \tag{A4}$$

Using (A3) and comparing it with (41), we arrive at the following results for the parameters of the equivalent circuit of Fig. 3:

$$n = \frac{\Delta v_c}{\Delta v_f} \tag{A5}$$

$$Y_2 = \frac{1}{Z_{0c}} \frac{Y_{ac}}{\Delta v_c^2} \tag{A6}$$

$$Y_1 = \frac{1}{Z_{0f}} \frac{Y_{af}}{\Delta v_f^2} \tag{A7}$$

where Δv_f, Δv_c, Y_{af}, and Y_{ac} are defined in (6), (40), (15a), and (15b) respectively.

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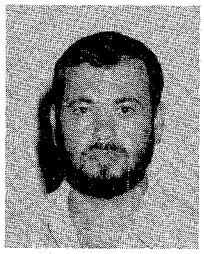

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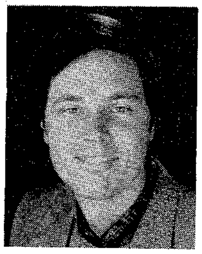
In 1978 he joined the Israeli Navy, where he served as project supervisor and Head of the Radar Section of the Navy Technical School. Between 1983 and 1989 he was employed by Rafael, Haifa, Israel, as an Antenna Engineer; there he was engaged in research and development work on microwave antennas. His research interests include microstrip arrays, reflector antennas and feeds, shaped beam antennas, and antenna modeling. Since 1989 he has been with the University of Massachusetts at Amherst, where he is working toward the Ph.D. degree.



David M. Pozar (S'74–M'80–SM'88–F'90) was born in Pittsburgh, PA, in 1952. He received the B.S. degree in 1975 and the M.S. degree in 1976, both in electrical engineering, from the University of Akron. He received the Ph.D. degree from Ohio State University in 1980. During the course of his undergraduate studies, he spent one year as an Engineering Assistant at the National Security Agency, Fort Meade, MD. He was a Graduate Research Assistant at the ElectroScience Lab of Ohio State University while pursuing the Ph.D. degree, and became a Research Assistant upon completion of the Ph.D. degree.

He joined the faculty at the University of Massachusetts in 1980 and in 1989 became Professor of Electrical and Computer Engineering. In 1988 he spent a six-month sabbatical leave at the Ecole Polytechnique Federale de Lausanne, in Lausanne, Switzerland, participating in several short courses and workshops on printed antennas.

Dr. Pozar is an Associate Member of the International Union of Radio Science (URSI), Commission B. He has served as an Associate Editor of the IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION (1983–1986) and as an Associate Editor of the *IEEE Antennas and Propagation Society Newsletter* (1982–1984). In 1981 he received the Outstanding Professor for 1981 award from Eta Kappa Nu, the student honor society. In 1984 he received an NSF Presidential Young Investigator Award, as well as the Keys to the Future Award from the IEEE Antennas and Propagation Society. In 1985 he received the Engineering Alumni Association's Outstanding Junior Faculty Award. In 1986 he received the R.W.P. King Best Paper Award from the IEEE Antennas and Propagation Society. In 1987 he received the URSI Issac Koga Gold Medal for his work on printed antennas and phased arrays. He again received the R.W.P. King Best Paper Award in 1988. In 1989 he received the Outstanding Teaching Award from the United Technologies Corporation. He presently serves as an Associate Editor for the IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION. Besides publishing papers on printed antennas and phased arrays, Dr. Pozar is the author of *Antenna Design Using Personal Computers* (Artech House, 1985) and *Microwave Engineering* (Addison-Wesley, 1990).