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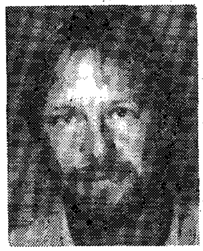

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Rectangular Bars Coupled Through a Finite-Thickness Slot

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Abstract —**A rigorous new solution, based on fringing capacitances and conformal mapping, is presented for the coupling problem in the parallel-coupled TEM transmission-line structure formed by two rectangular bars coupled through a rectangular slot, cut longitudinally in a finite-thickness ground plane. The conformal mapping solution is summarized in Appendix I and is based on the procedure published by Getsinger [1]. Graphs of the capacitances needed to calculate the coupling for a given physical geometry, without solving the complicated equations associated with the theoretical solution, are also presented. A measurement procedure and experimental results are presented which verify the accuracy of the theoretical coupling data.**

I. Introduction

THE TEM STRUCTURE with cross section shown in Fig. 1 is used in interdigital linear-phase filters [2] and directional couplers with weak coupling, for example. For these applications, it is necessary to have an accurate procedure for designing the slot width that will realize a desired coupling, for a geometry that is otherwise specified. The current status of the literature on this problem is reviewed below.

In 1958, Shimizu and Jones [3] published an approximate formula, valid for weak coupling and based on Bethe's small aperture theory, for the case where both the coupling plate and center conductors have zero thickness, i.e., $a = 0$ and $t = 0$. Although their formula is only valid for zero-thickness slots, Shimizu and Jones argued that, for finite-thickness coupling plates, the slot acts as a waveguide below cutoff, and suggested the use of a simple correction term to take the effect of finite thickness into account.

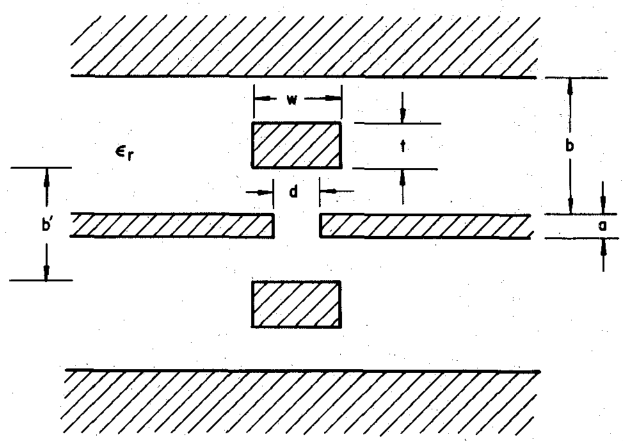

Fig. 1. Slot-coupled rectangular-bar geometry.

Rhodes [2] invented the linear-phase filter and, in 1970, he used Shimizu and Jones' formula to compute the cross-coupling slot widths of rectangular-bar interdigital linear-phase filters, and obtained good agreement between the theoretical and measured performance of 14- and 18-element filters.

However, in 1976, Levy [4] stated that the early linear-phase filters did not have good agreement between theory and experiment as far as the group-delay characteristics were concerned. The given reason was the failure to obtain the correct values for the cross couplings, due to the inadequate electromagnetic theories for the prediction of

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the coupling between the conductors. Levy "developed a new formula for the coupling capacitance between two bars coupled through a slot of finite thickness, and soon obtained generalized linear-phase filters having accurately predictable characteristics." The formula was not published for proprietary reasons [4].

In 1978, Jokela [5] used equations published by Matthaei, Young, and Jones [6, p. 177] to design a stripline-generalized interdigital filter. This expression was published in 1964, six years after the paper by Shimizu and Jones. The new equation also is valid only for the case with zero-thickness center conductors and coupling plates, but can be expected to give good results in Jokela's application, which is for stripline.

In 1966, Yamamoto, Azakami, and Itakura [7] published a rigorous conformal mapping solution for the zero-thickness geometry, with $a=0$ and $t=0$.

The 1970 procedure suggested by Rhodes [2] for designing the cross-sectional dimensions of interdigital linear-phase filters by using the 1958 formula of Shimizu and Jones was refined by Cloete and published by Malherbe [8] in 1979. In 1982, the use of this procedure was recommended by Zabalawi [9] for the design of a printed circuit linear-phase filter.

In summary, designs based on published data for zero-thickness striplines coupled through a zero-thickness ground plane [3], [6] can be expected to give good results for printed circuit structures [5], [9]. However, as pointed out by Levy [4], the coupling problem involving a finite-thickness slot had not been solved satisfactorily in the literature by 1976. The situation still prevails, and the purpose of this paper is to present data which can be used with confidence for the design of rectangular bars which are coupled through a thick ground plane.

II. Coupling in Terms of Fringing Capacitances

In considering the various analytic and numerical techniques which could be used to solve the coupling problem for the geometry of Fig. 1, the question arose whether the conformal mapping technique used by Getsinger [1] for the problem of two-side coupled rectangular bars could be adapted to solve the problem of two-slot coupled rectangular bars.

It will be shown that this is indeed the case, and, furthermore, that only one new fringing capacitance problem has to be solved in order to complement Getsinger's even-mode and isolated-bar fringing capacitance data, and thus obtain an accurate solution for the slot-coupled rectangular-bar geometry.

A general expression for the coupling between the two rectangular cross-sectional bars through a slot in the finite-thickness ground plane can be derived in terms of the static capacitances per unit length of the bars to ground, for the odd and even modes of TEM propagation. The capacitances and boundary conditions for the two modes are shown in Fig. 2, where the notation for the parallel plate and fringing capacitances has been kept consistent with Getsinger's notation wherever possible.

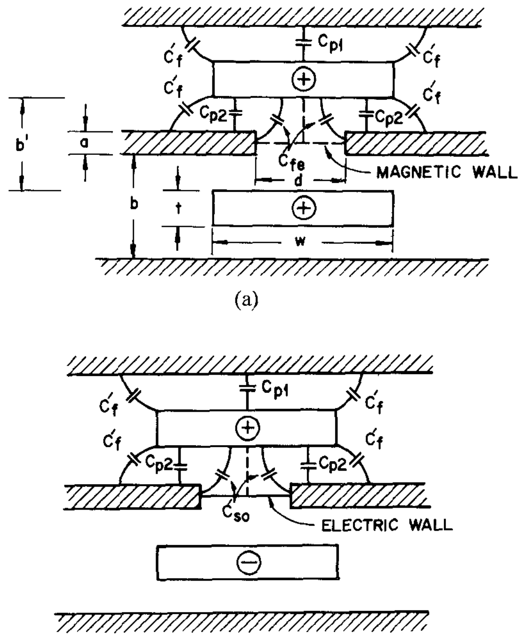

Fig. 2. (a) Even- and (b) odd-mode excitation of two slot-coupled rectangular bars.

For the even-mode excitation, the total even-mode capacitance from one bar to the ground is

$$\frac{C_{oe}}{\epsilon}=4\frac{C_f'}{\epsilon}+2\frac{C_{fe}'}{\epsilon}+\frac{C_{p1}}{\epsilon}+2\frac{C_{p2}}{\epsilon} \tag{1}$$

where the fringing capacitances C_f'/ϵ from the corners of the bar, and the fringing capacitances C_{fe}'/ϵ from the reentrant corners of the slot, are identical to the isolated-bar and even-mode fringing capacitances of Getsinger's geometry, subject to the constraint that interaction between the fringing fields is negligible.

The expressions for the parallel-plate capacitances follow by inspection as

$$\frac{C_{p1}}{\epsilon}=\frac{w}{(b-t)/2} \tag{2}$$

and

$$2\frac{C_{p2}}{\epsilon}=\frac{(w-d)}{(b-t)/2} \tag{3}$$

and substitution into (1) yields

$$\frac{C_{oe}}{\epsilon}=4\frac{C_f'}{\epsilon}+2\frac{C_{fe}'}{\epsilon}+\frac{2(2w-d)}{(b-t)}. \tag{4}$$

A similar expression results for the total odd-mode capacitance from one bar to ground

$$\frac{C_{oo}}{\epsilon}=4\frac{C_f'}{\epsilon}+2\frac{C_{so}'}{\epsilon}+\frac{2(2w-d)}{(b-t)} \tag{5}$$

where C_{so}'/ϵ is the odd-mode fringing capacitance from the reentrant corners of the slot.

The voltage coupling coefficient between the two bars is given by [6, p. 779]

$$k = \frac{Z_{oe} - Z_{oo}}{Z_{oe} + Z_{oo}} \tag{6}$$

and since, for symmetrical coupling geometries

$$Z_{oe} = \frac{1}{vC_{oe}} \tag{7}$$

$$Z_{oo} = \frac{1}{vC_{oo}} \tag{8}$$

where v is the velocity of propagation, (6) can be manipulated into the form

$$k = \frac{\Delta C/\epsilon}{C_{oe}/\epsilon + \Delta C/\epsilon} \tag{9}$$

where the coupling capacitance $\Delta C/\epsilon$ is given by

$$\Delta C/\epsilon = \{C_{oo}/\epsilon - C_{oe}/\epsilon\}/2. \tag{10}$$

From (4) and (5), it follows that the coupling capacitance can be written entirely in terms of the odd-mode and even-mode fringing capacitances as

$$\Delta C/\epsilon = C'_{so}/\epsilon - C'_{fe}/\epsilon. \tag{11}$$

It is now evident that if a solution for the odd-mode fringing capacitance C'_{so}/ϵ can be found to be used in conjunction with Getsinger's data for C'_f/ϵ and C'_{fe}/ϵ, the coupling problem can be completely solved in terms of (4), (9), and (11).

In Appendix I, a conformal mapping solution for C'_{so}/ϵ is presented, based on the assumption that the slot is narrow enough, relative to the bar width, to prevent significant interaction between the fringing fields at the corner of the slot and the corner of the bar. A simple modification of Cohn's criterion [1, sec. IV] results in the following restriction on slot width:

$$d < w - 0.7\,(b - t) \tag{12}$$

in order to keep the total capacitance error due to interaction of the fields less than 1.24 percent. The practical implications of the restriction on slot width are discussed in Section V.

As shown in Appendix I, a deceptively simple-looking expression for $\Delta C/\epsilon$ results from (11), (19) for C'_{so}/ϵ, and Getsinger's equation for C'_{fe}/ϵ [1, (36)]. However, evaluation of $\Delta C/\epsilon$ is, in fact, computationally complex because of the Jacobian elliptic-, zeta-, and theta-functions which are implicit in (20).

The need to compute $\Delta C/\epsilon$ for a particular physical geometry is circumvented by Fig. 3, where a family of graphs of the coupling capacitance $\Delta C/\epsilon$ is plotted as a function of normalized slot width d/b' for various choices of normalized slot thickness a/b'. To make the graphs self contained, corresponding graphs of the even-mode fringing capacitance C'_{fe}/ϵ are also presented, to allow C'_{so}/ϵ to be recovered from (11). C'_{fe}/ϵ was computed using Getsinger's equation (36).

With the graphs in Fig. 3 and the graph of isolated bar-fringing capacitance C'_f/ϵ, published by Getsinger [1, fig. 5], the voltage coupling coefficient for two rectangular bars coupled through a finite-thickness slot can be simply calculated using (4), (9), and (11). The effect of sidewalls at finite distances can be taken into account using the procedure described by Getsinger [1].

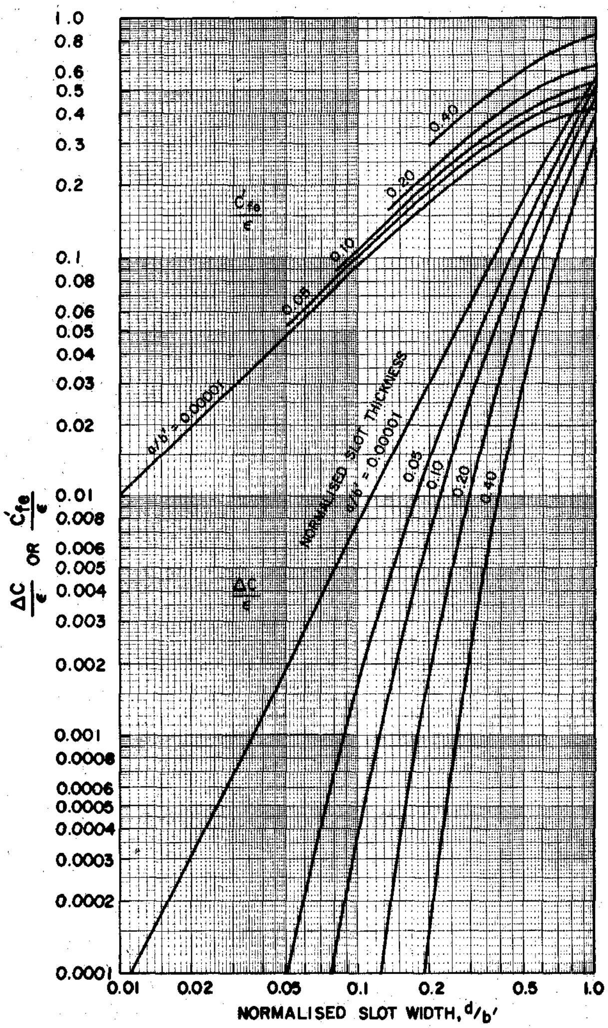

Fig. 3. Graphs of the fringing capacitance $\Delta C/\epsilon$ and C'_{fe}/ϵ.

III. Experimental Microwave Circuit

The accuracy of the theoretical solution was verified by comparison with measured coupling data.

The experimental circuit was designed to allow coupling measurements to be made for a variety of slot widths and thicknesses in order to compare the measured results with theoretical values. The circuit is shown in Fig. 4. The "interdigital coupler" configuration was selected instead of the conventional quadrature coupler, where all four ports have to be terminated in identical resistive loads, for the following reasons:

1) Only two connectors are required with the remaining two ports terminated in well-defined short circuits instead of less reliable 50-Ω loads.

2) As will be shown, the output power of the "interdigital coupler" is 6 dB more than the coupled output ob-

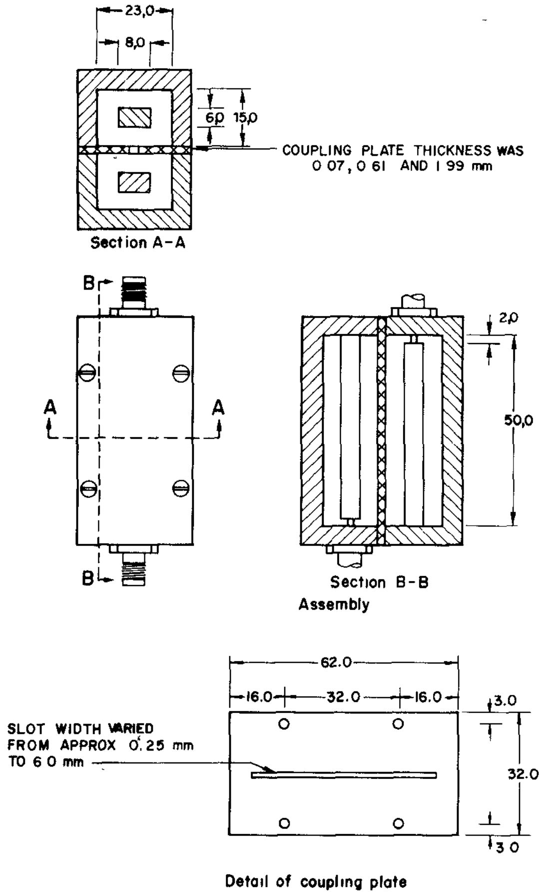

Fig. 4. Experimental slot-coupled line pair.

tained with a quadrature coupler for small coupling coefficients, thus allowing narrower slots to be characterized for the same measurement noise level.

3) The configuration is the basic structure in the interdigital linear-phase filter.

The circuit was designed for a center frequency of 1500 MHz. The choice of frequency was determined by considering size, discontinuity effects, and measurement convenience. The ground plane spacing was chosen to be $b = 15$ mm, being a compromise between the criterion that $b < 0.1 \lambda_0 = 20$ mm to minimize discontinuity effects [6, p. 798], and the choice of SMA connectors with square flanges of dimensions 12.7 mm. The center conductor thickness was chosen to give $t/b = 0.4$, and the width w was designed to give an uncoupled characteristic impedance of 50 Ω for each bar. As usual, the circuit was designed to operate between a 50-Ω load and generator impedances.

The magnitude of the transmission response at the center frequency ω_0 of the experimental circuit

$$t \triangleq |S_{21}|_{\omega = \omega_0} \tag{13}$$

can be measured directly. However, the purpose of the circuit is to measure the voltage coupling coefficient [6, p. 779]

$$k = \frac{Z_{oe} - Z_{oo}}{Z_{oe} + Z_{oo}} \tag{14}$$

where Z_{oe} and Z_{oo} are the even- and odd-mode impedances of the circuit.

In Appendix II, it is shown that k can be computed from the measured value of t by

$$k = \frac{1 - \sqrt{1 - t^2}}{t - \left(1 - \sqrt{1 - t^2}\right)}. \tag{15}$$

For weakly coupled lines, $t \ll 1$ and (15) can be simplified to the form

$$k \cong \frac{t}{2 - t} \cong \frac{t}{2} \qquad \text{for } t \ll 1. \tag{16}$$

Thus it follows from (16) that, for weak coupling, the transmission coefficient t is approximately twice the voltage coupling coefficient k. Therefore, by measuring t, instead of k directly, an increase of 6 dB in measurement range is achieved.

IV. Measurements

Three sets of brass coupling plates with rectangular slots were manufactured for the experimental slot-coupled line pair using a numerically controlled Charmilles DCNC F40 Wire Spark Eroder.

At the start of the measurement program, all the plates were cleaned using Brasso, followed by an ethyl alcohol bath in an ultrasonic cleaner for 10 min and hand polishing. High pressure air was blown through the slots to ensure that they were clean.

The nominal thickness of the brass plates in each of the three sets was 0.07 mm for Set No. 1, 0.61 mm for Set No. 2, and 1.99 mm for Set No. 3, as measured using a micrometer. Inspection of the slots using a microscope revealed that, for practical purposes, the edges were perfectly parallel and perpendicular to the plate surfaces. The peak-to-peak surface roughness of the two parallel slot edges was measured using a Zeiss Light Section Microscope and found to be approximately 20 μm, or 6 skin depths for brass at 1500 MHz. It was shown that the resulting increase in surface resistivity has a negligible effect on the accuracy of the coupling measurement [10, pp. 213–218]. The widths of the slots were measured using a Bausch and Lomb Binocular Microscope.

The transmission measurements were carried out at five frequencies, namely 1.0, 1.1, 1.2, 1.3, 1.4, and 1.5 GHz, using the HP8568A spectrum analyzer as a receiver and two calibrated-step attenuators as attenuation standards.

The measurement setup is shown in Fig. 5. Because the slot-coupled line pair is severely mismatched, fixed attenuators were inserted at the input and output ports to prevent measurement inaccuracies. The measurement procedure consisted of two steps. Having set the signal source to the desired frequency, the first step was to set the step attenuator to $A_0 = 90$ dB with the through line in the measurement circuit and to measure the reference signal power level P_0 (dBm) at the spectrum analyzer. With appropriate settings, the spectrum analyzer noise level could be reduced to -117 dBm if necessary, but most measurements were carried out with the noise level at -94 dBm.

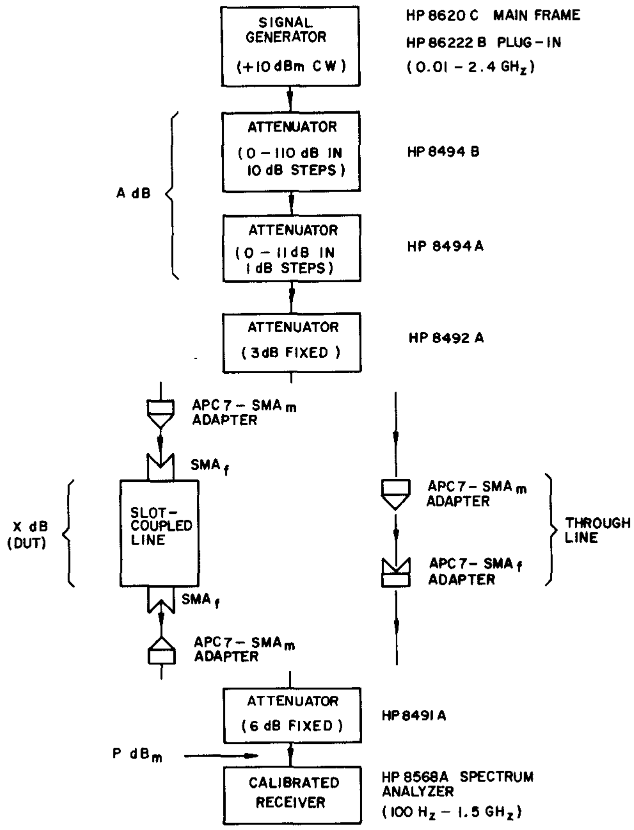

Fig. 5. The system for measuring coupling.

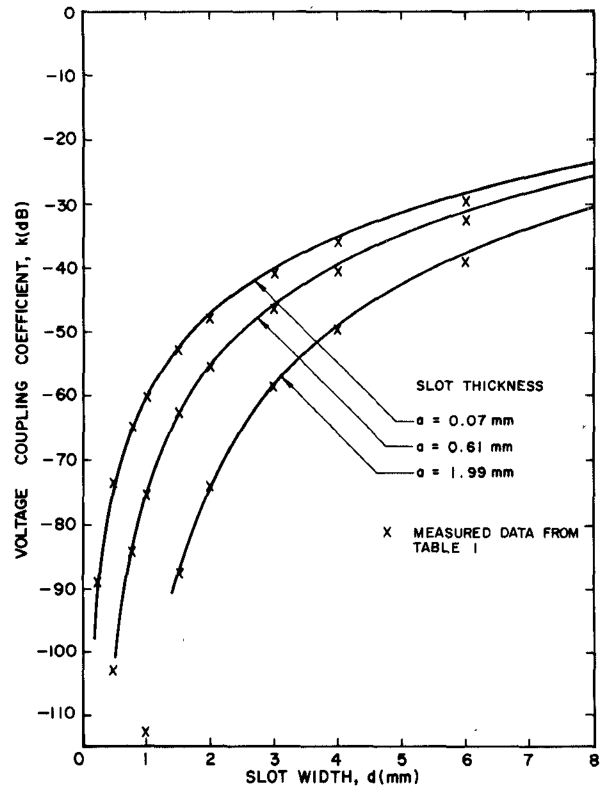

Fig. 6. Comparison of measured data with theoretical coupling, computed using fringing capacitances.

TABLE I
MEASURED COUPLING k (dB) FOR THREE SETS OF COUPLING PLATES

Slot width d (mm)	Slot thickness a (mm)		
	Set No 1 a = 0.07	Set No 2 a = 0.61	Set No 3 a = 1.99
0.24	-88.6	-	-
0.48	-73.4	-102.9	-
0.76	-65.0	-84.2	-
1.00	-60.3	-75.2	-112.1
1.50	-52.9	-62.9	-87.3
2.00	-47.6	-55.5	-74.0
3.00	-40.7	-46.1	-58.6
4.00	-35.7	-40.1	-49.5
6.00	-29.5	-32.2	-38.9

The second step was to remove the through line, insert the slot-coupled line pair, and then reduce the attenuation of the calibrated step attenuators to A_1 (dB) until the signal level P_1 (dBm) had been restored to within a few dBm of the reference signal level. The attenuation X (dB) due to the slot-coupled line pair was then calculated using the equation

$$X=\{P_0-P_1\}+\{A_0-A_1\}\text{ dB}. \qquad (17)$$

Throughout the procedure, care was taken to ensure that the spectrum analyzer remained calibrated.

The measured results are summarized in Table I. The peak value of the transmission coefficient measured at the five frequencies was used to compute k from (15). The large influence of slot thickness on the coupling for a given slot width is evident from the data, stressing the importance of taking slot thickness into account in any theoretical approach to the problem.

V. Comparison of Theoretical and Experimental Results

Fig. 6 presents theoretical graphs of the coupling between two rectangular-bar transmission lines for the geometry of the experimental circuit shown in Fig. 4, with the slot thicknesses chosen equal to the thicknesses of the three sets of experimental coupling plates. The theoretical coupling was computed as a function of slot width, using the solution discussed in Section II. The measured data from Table I are also shown in Fig. 6.

In the theoretical solution, it was assumed that interaction between the fringing fields at the slot corners and the bar corners was negligible. This gave rise to restriction (12) on slot width. For the experimental circuit, the restriction is satisfied if $d<1.7$ mm. It is clear that for slot widths less than 1.7 mm, the agreement between theoretical and measured data is excellent. Furthermore, discrepancies of greater than 1 dB only become evident for slot widths in excess of twice this value, while the indication is that the theoretical coupling will be within 2 dB of the actual coupling, even when the slot width is equal to the width of the rectangular bar. The accuracy of the solution, even for this extreme case, can be ascribed to the circuit being a member of the class with cross-sectional geometries in which interaction between the fringing fields is weak, even when the discontinuities that give to the fields are close to each other [11, p. 10].

VI. Conclusion

The solution presented here was inspired by Getsinger's widely used paper [1] and uses conformal mapping to obtain equations for the relevant capacitances associated with the slot-coupled rectangular-bar geometry. The derivation of the equations is summarized in Appendix I.

Because of the complexity of the capacitance equations, graphs of the coupling capacitance and even-mode fringing capacitance are presented to simplify the design of rectangular bars coupled through a longitudinal slot in a thick ground plane.

The measurement program has verified the accuracy of the theoretical solution, subject to the constraint that interaction between the fringing fields is minimal. It has been shown that, even for slots of width comparable to the rectangular-bar width, the theoretical procedure results in coupling errors that may be acceptable in a practical design.

Appendix I
Conformal Mapping Solution

After using symmetry arguments, the exact boundary conditions which apply when the slot-coupled rectangular bars are excited in the odd mode are shown in Fig. 7(a). A solution for the total odd-mode capacitance-to-ground could be obtained by the use of numerical techniques, such as the finite-difference or finite-element methods, but only with great difficulty by the use of analytical techniques, such as conformal mapping. In the case of conformal mapping, the use of the Schwarz–Christoffel transformation requires the nine corners of the polygon to be mapped onto the t-plane, and the resulting integral would be practically intractable.

In contrast, by using the approach described in Section II the problem can be considerably simplified, since only the reentrant corner odd-mode fringing capacitance C'_{so} need be found, and not the total odd-mode capacitance-to-ground C_{oo}.

The odd-mode reentrant corner fringing capacitance C'_{so} is defined, in the simplified geometry of Fig. 7(b), as the difference between the total capacitance which exists between the electric walls, FG and EDCA, and the parallel plate capacitance. This problem is tractable by conformal mapping.

In fact, the method described by Getsinger [1] for the even-mode fringing capacitance C'_{fe} in the side-coupled case can be modified to obtain C'_{so} for the slot-coupled case. The steps in the mapping procedure are shown in Fig. 8, where notation consistent with Getsinger's Fig. 7 and Fig. 9 have been used. Comparison of the boundary conditions shows that in the case of C'_{fe}, the boundary ED is a magnetic wall, whereas in the case of C'_{so} it is an electric wall. Accordingly, the magnetic wall EF must be mapped onto the imaginary axis of the w_1 plane, instead of DEF, as in Getsinger's case.

The appropriate mapping from the z-plane onto the

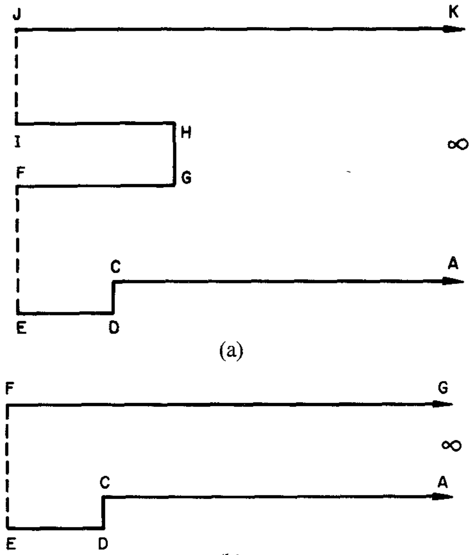

Fig. 7. Boundary conditions at the reentrant corner for the odd-mode excitation. (a) Exact boundary conditions. (b) Simplified boundary conditions.

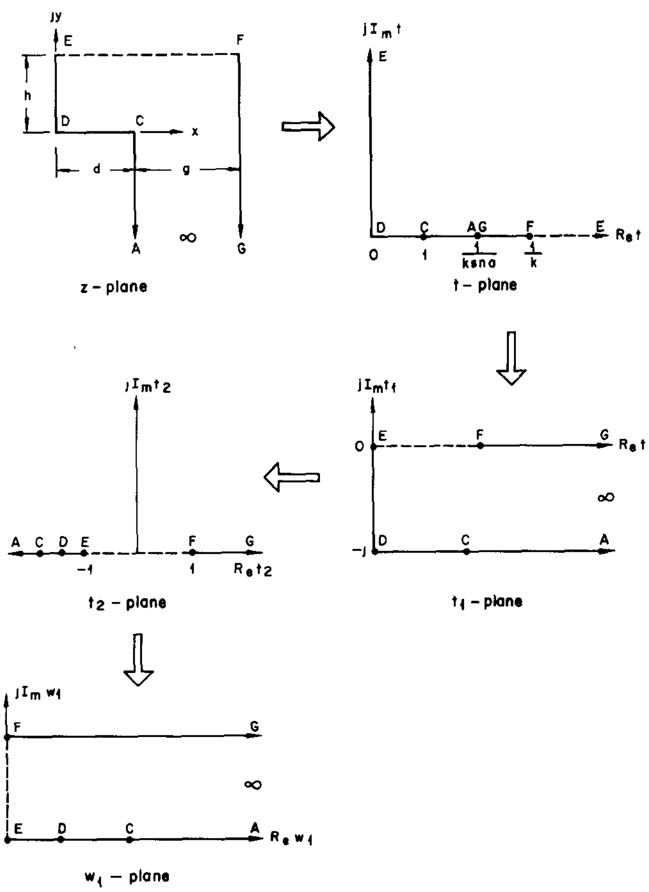

Fig. 8. Steps in the conformal mapping solution for C'_{so}/ϵ.

w_1-plane is accomplished by using the t_1-plane to t_2-plane mapping

$$t_2 = -\operatorname{sn}^2 a - \operatorname{cn}^2 a \cosh \pi t_1 \tag{18}$$

instead of Getsinger's equation (30), with all the other

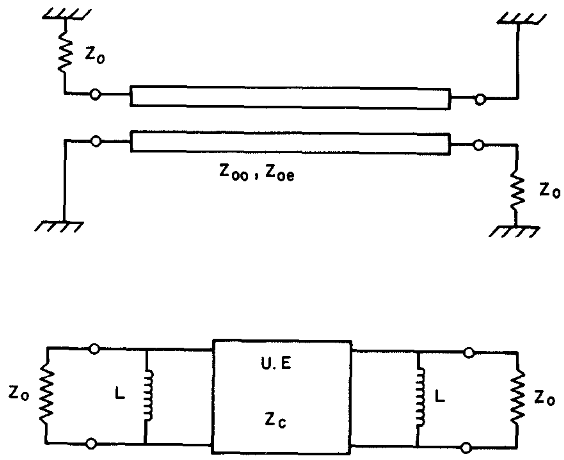

Fig. 9. Equivalent circuit of a slot-coupled line pair. (a) Interdigital coupler. (b) Equivalent circuit using commensurate length-distributed circuit symbols.

mappings unchanged [10, pp. 141–155]. The resulting expression for the odd-mode fringing capacitance is

$$\frac{C'_{so}}{\epsilon}=\frac{1}{\pi}\ln 2+\frac{1}{\pi}\ln\left\{\frac{\operatorname{sn}a\operatorname{cn}a}{j\operatorname{dn}a}\right\}-(K'-\beta)\left\{\frac{1}{g}-\frac{2j}{\pi}Z(a)\right\}$$
$$+\frac{K'}{4K}+\frac{1}{2\pi}\ln\frac{2kk'K}{\pi}-\frac{1}{\pi}\ln\Theta(jK'-j2\beta). \quad (19)$$

The equation for the coupling capacitance is obtained by substituting (19) and Getsinger's equation for C'_{fe}/ϵ [1, (36)] into (11) to obtain

$$\frac{\Delta C}{\epsilon}=\frac{1}{\pi}\ln(\operatorname{sn}^2 a). \quad (20)$$

The variables and functions in (19) and (20) are as defined by Getsinger [1].

Appendix II
Relationship Between k and t

The equivalent circuit of the coupled line pair is given in Fig. 9. The even- and odd-mode impedances Z_{oo} and Z_{oe} are related to the circuit element values by the following equations [12]:

$$L=Z_{oe} \quad (21)$$

$$Z_c=\frac{2Z_{oe}Z_{oo}}{Z_{oe}-Z_{00}}. \quad (22)$$

The transmission response of the interdigital coupler can be shown to be

$$S_{21}(S)=\frac{2S\sqrt{1-S^2}}{(Z_o/Z_c+Z_c/Z_o)S^2+2(1+Z_c/L)S+(2+Z_c/L)Z_o/L} \quad (23)$$

where

$$S=j\tan\left(\frac{\pi\omega}{2\omega_o}\right). \quad (24)$$

Therefore, the magnitude of the transmission response at the center frequency is

$$|S_{21}|_{\omega=\omega_o}\triangleq t=\frac{2}{Z_o/Z_c+Z_c/Z_o} \quad (25)$$

which is of the form

$$t=\frac{2x}{1+x^2} \quad (26)$$

where

$$x=Z_o/Z_c. \quad (27)$$

Solving the quadratic equation in x yields

$$x=\left(1\pm\sqrt{1-t^2}\right)/t. \quad (28)$$

The purpose of the experimental slot-coupled line pair is to enable the voltage coupling coefficient k to be measured. Since the transmission response t can be measured directly, it is necessary to find an expression for k in terms of t.

The voltage coupling coefficient is given by (14)

$$k=\frac{Z_{oe}-Z_{oo}}{Z_{oe}+Z_{oo}} \quad (14)$$

and the characteristic impedance of the coupled line pair is given by

$$Z'_o=\sqrt{Z_{oe}Z_{oo}} \quad (29)$$

where the prime is used to distinguish between the characteristic impedance of the coupled line pair and the impedance of the terminating loads Z_o.

From (14), it follows that

$$Z_{oe}=Z_{oo}\left(\frac{1+k}{1-k}\right) \quad (30)$$

and from (22) and (30) that

$$Z_c=Z_{oo}\left(\frac{1+k}{k}\right) \quad (31)$$

and, finally, from (27) and (31) that

$$x=\frac{Z_o}{Z_{oo}}\left(\frac{k}{1+k}\right). \quad (32)$$

Although, in general, the odd-mode characteristic impedance Z_{oo} is not known exactly, it may be assumed, to a high degree of accuracy, that

$$Z_{oo}\cong Z'_o|_{d=0} \quad (33)$$

where $Z'_o|_{d=0}$ is the characteristic impedance of the lines when the coupling slot width is zero, i.e., the uncoupled characteristic impedance [10, pp. 209–212].

Since

$$Z'_o|_{d=0}=Z_o=50\ \Omega \quad (34)$$

by design, it follows from (32)–(34) that

$$x \cong \frac{k}{1+k} \tag{35}$$

for all practical slot widths and thicknesses.

From physical considerations, it is known that $0 \leqslant k < 1$ and, hence from (35), the range of x is $0 \leqslant x < 0.5$. In order to satisfy this constraint, it is necessary to select the solution with the negative sign before the square root in (28). Therefore

$$x = \left(1 - \sqrt{1 - t^2}\right)/t \tag{36}$$

and from (35) it follows that

$$k = \frac{1 - \sqrt{1 - t^2}}{t - \left(1 - \sqrt{1 - t^2}\right)} \tag{37}$$

which is the desired relationship between t and k.

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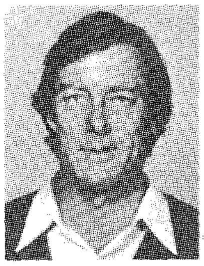

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Experimental Study of the *W*-Band Dielectric-Guide *Y*-Branch Interferometer

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***Abstract*—Applicability of a *W*-band dielectric-guide *Y*-branch (DGY) interferometer as a modulator and switching element for the millimeter-wave planar dielectric-guide integrated circuits is established. The switching element includes a phase shifter formed by a metal wall proximate to a dielectric guide.**

The influence of the branches asymmetry on the on/off switching ratio was measured and found to comply with theory. On/off switching ratios of about 25 dB were obtained over a 98–104-GHz frequency range.

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I. Introduction

ALTHOUGH the dielectric-guide Y-branch (DGY) interferometer is extensively applied in the optical integrated circuits for power splitting, light intensity modulation, and switching [1], to our knowledge no experimental data on this subject has been published with respect to the millimeter-wave range. This study seeks to establish the applicability of the Y-branch element in the W-band integrated circuits. Low radiation losses under even excitation, near-3-dB power division, and low VSWR values were observed in the Y-junction.

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