MMIC DEVICES FOR ACTIVE PHASED ARRAY ANTENNAS

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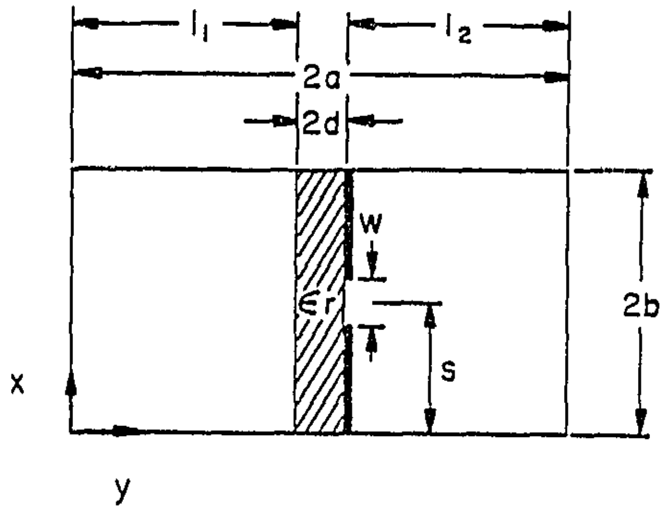

Figure 1. Unilateral fin-line structure.

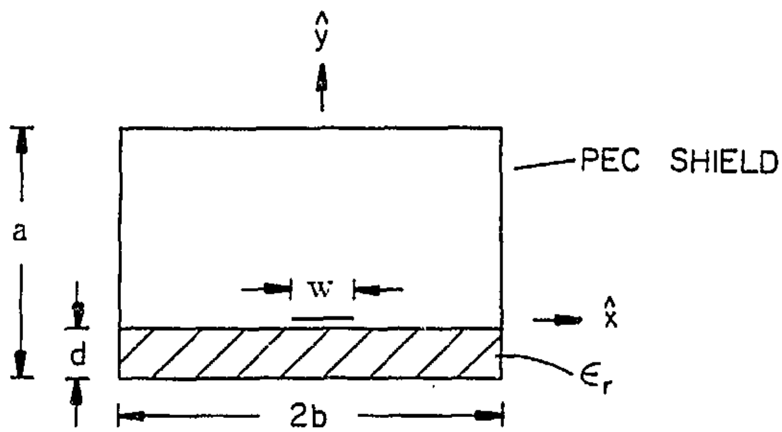

Figure 2. Shielded microstrip structure.

III. MODAL SOLUTIONS FOR FIN-LINE AND MICROSTRIP

Modal solutions for many printed circuit waveguides, such as the fin-line and microstrip, are most simply formulated in the spectral domain. This involves the use of a Fourier series representation of the field quantities with respect to the transverse direction (x-direction), parallel to the metallization. A coordinate transformation can then be used to enable a transverse resonance analysis [7]. The result is a spectral relationship, in the plane of the metallization, between the electric field and current. For the fin line, this relationship may be written as

$$\bar{\bar{Y}}(k_x, \beta)\, \tilde{E}\, (k_x) = \tilde{J}\, (k_x) \tag{1}$$

where

$$\bar{\bar{Y}} = \begin{vmatrix} Y_{xx} & Y_{xz} \\ Y_{zx} & Y_{zz} \end{vmatrix}$$

$$\tilde{E} = \begin{vmatrix} \tilde{E}_x \\ \tilde{E}_z \end{vmatrix}$$

$$\tilde{J} = \begin{vmatrix} \tilde{J}_x \\ \tilde{J}_z \end{vmatrix},$$

and for the microstrip,

$$\bar{\bar{Z}}(k_x, \beta)\, \tilde{J}\, (k_x) = \tilde{E}\, (k_x) \tag{2}$$

where

$$\bar{\bar{Z}} = \begin{vmatrix} Z_{xx} & Z_{xz} \\ Z_{zx} & Z_{zz} \end{vmatrix}$$

$$\tilde{J} = \begin{vmatrix} \tilde{J}_x \\ \tilde{J}_z \end{vmatrix}.$$

$$\tilde{E} = \begin{vmatrix} \tilde{E}_x \\ \tilde{E}_z \end{vmatrix}.$$

The spectral variable k_x is discrete, and $k_x = 2n\pi/2b$. The dyadics $\bar{Y}$ and $\bar{Z}$ are forms of spectral Green's functions and are of similar form. The derivation of these Green's functions is discussed in [6], [7].

The application of Galerkin's method to eq. (1) or (2) results in a homogeneous equation of the form $Lf = 0$, where f represents the coefficients of the x- and z- unknown-quantity basis functions. The basis functions used to represent the slot fields in the case of the fin-line and the strip currents in the case of the microstrip are

$$\left.\begin{aligned} \zeta_p(x) &= \frac{\cos\left[\dfrac{(p-1)\pi}{w}\left(x-s-\dfrac{w}{2}\right)\right]}{1-\left[\dfrac{2(x-s)}{w}\right]^2}, \quad p = 1, 2, 3, \ldots \\ \eta_q(x) &= \frac{\sin\left[\dfrac{q\pi}{w}\left(x-s-\dfrac{w}{2}\right)\right]}{1-\left[\dfrac{2(x-s)}{w}\right]^2}, \quad q = 1, 2, 3, \ldots \end{aligned}\right\} \quad s-\frac{w}{2} < x < s+\frac{w}{2} \tag{3}$$

$$\left.\begin{aligned} \zeta_p(x) &= 0 \\ \eta_z(x) &= 0 \end{aligned}\right\} otherwise,$$

together with $\zeta_o(x) = P_x(s, w)$. The representation becomes

$$J_z(x) \; or \; E_x(x) = \sum_{p=1}^{P} a_p \, \zeta_p(x) \tag{4a}$$

$$J_x(x) \; or \; E_z(x) = \sum_{q=1}^{Q} b_q \, \eta_q(x). \tag{4b}$$

For a centrally located slot or strip, only even ζ and odd η terms (with respect to the center of the slot) are required, i.e., $p = 0, 2, 4, \ldots$ and $q = 2, 4, 6, \ldots$.

The matrix equation is formed using the inner product

$$<\tilde{X}_i(k_x), \tilde{Y}_j(k_x)> = \sum_{n=-N/2}^{N/2} \tilde{X}_i^*(k_x)\tilde{Y}_j(k_x) \tag{5}$$

where the number of spectral terms has been truncated to N. The eigenvalue solutions for k_z are found by solving $det(L) = 0$. This can be achieved satisfactorily by iterative procedures that involve a detection of the change in the sign of the determinant, followed by Newton's method. The relative basis function coefficients can now be found following substitution for the eigenvalues. This results in the electric field in the plane of the fin for the fin-line, or the strip current for the microstrip, which when Fourier transformed and operated on by $\bar{Z}$ give the electric field in the plane of the strip. The electric and magnetic fields can be found in the spectral domain as a function of y by returning to the transverse resonance analysis [7]. Therefore, for each mode, this analysis gives $\tilde{E}_x(k_x, y)$, $\tilde{E}_y(k_x, y)$, $\tilde{H}_x(k_x, y)$ and $\tilde{H}_y(k_x, y)$.

The fields for the nth mode may be expressed as

$$\bar{E}(x, y, z) = \bar{e}_n(x, y)\, e^{-jk_{zn}z} + \bar{e}_{zn}(x, y)\, e^{-jk_{zn}z} \tag{6a}$$

$$\bar{H}(x, y, z) = \bar{h}_n(x, y)\, e^{-jk_{zn}z} + \bar{h}_{zn}(x, y)\, e^{-jk_{zn}z} \tag{6b}$$

where $\bar{e}_n$ and $\bar{h}_n$ are transverse vector mode functions, and $\bar{e}_{zn}$ and $\bar{h}_{zn}$ are longitudinal vector mode functions. These mode functions, when normalized, satisfy the following orthogonality relation [10]

$$\int_s\int \bar{e}_n \times \bar{h}_m \cdot \hat{z}\; ds = \delta_{nm}$$

where δ_{nm} is the Kronecker delta which is defined by

$$\delta_{nm} = \begin{cases} 1 & n = m \\ 0 & n \neq m \end{cases} . \tag{7}$$

Using Parseval's theorem, the orthogonality condition in eq. (6) can be written as

$$\frac{1}{2b}\,\mathrm{Re}\left\{ \sum_{n=-\infty}^{\infty} \int_0^{2a} \left[\tilde{E}_x(k_x, y)\, \tilde{H}_y^*(k_x, y) - \tilde{E}_y(k_x, y)\tilde{H}_x^*(k_x, y) \right] dy \right\} = \delta_{nm}. \tag{8}$$

The fields are determined at (m, n) points over the waveguide cross-section. Therefore, m spectral terms are used in the summation and a discrete integral in y is performed over n points. The space domain fields can be determined by applying the inverse Fourier transform using an FFT.

IV. DISCUSSION OF MODAL SOLUTIONS AND ORTHOGONALITY OF MODES

Upon investigation of the convergence of the fin-line phase constant solutions, two basis functions from eq. (3) ($p = 1, 3, q = 2, 4$) give sufficient accuracy for most purposes. A rule of thumb for the number of spectral terms required in the inner product of eq. (4) is $N \geq 16b/w$. This indicates that more spectral terms are required for a larger waveguide width to slot-width ratio. This number of basis functions gives good results without excessive computation time.

The solutions for the propagation constant are very stable when a relatively small number of basis functions and spectral terms are used, i.e., a small number of spectral terms are required for convergence in the propagation constant. However, at least an order of magnitude more spectral terms are required for convergence of the basis function coefficients. Also, an increase in the number of basis functions necessitates an increase in the number of spectral terms. Figures 3 and 4 show plots of four field components for the dominant mode over the fin-line waveguide cross-section. The fields are essentially confined to the region around the slot.

It is quite difficult to find a satisfactory representation for the propagating and evanescent mode functions. In particular, the modes do not satisfy the orthogonality relationship of eq. (7) well. This means that only several of these modes are likely to be useful in the analysis of discontinuity problems and the like. Example sets of computed values for eq. (7) are shown in Table 1 for the fin-line and Table 2 for the microstrip.

In summary, the solution of a rather large class of discontinuity problems in printed circuitry may depend upon the accurate determination of the propagation constants and modal fields of these guides. It is shown in this paper, that, while the spectral Galerkin technique is very good for finding the propagation constants of these modes, it may not be as accurate for finding the modal fields, when

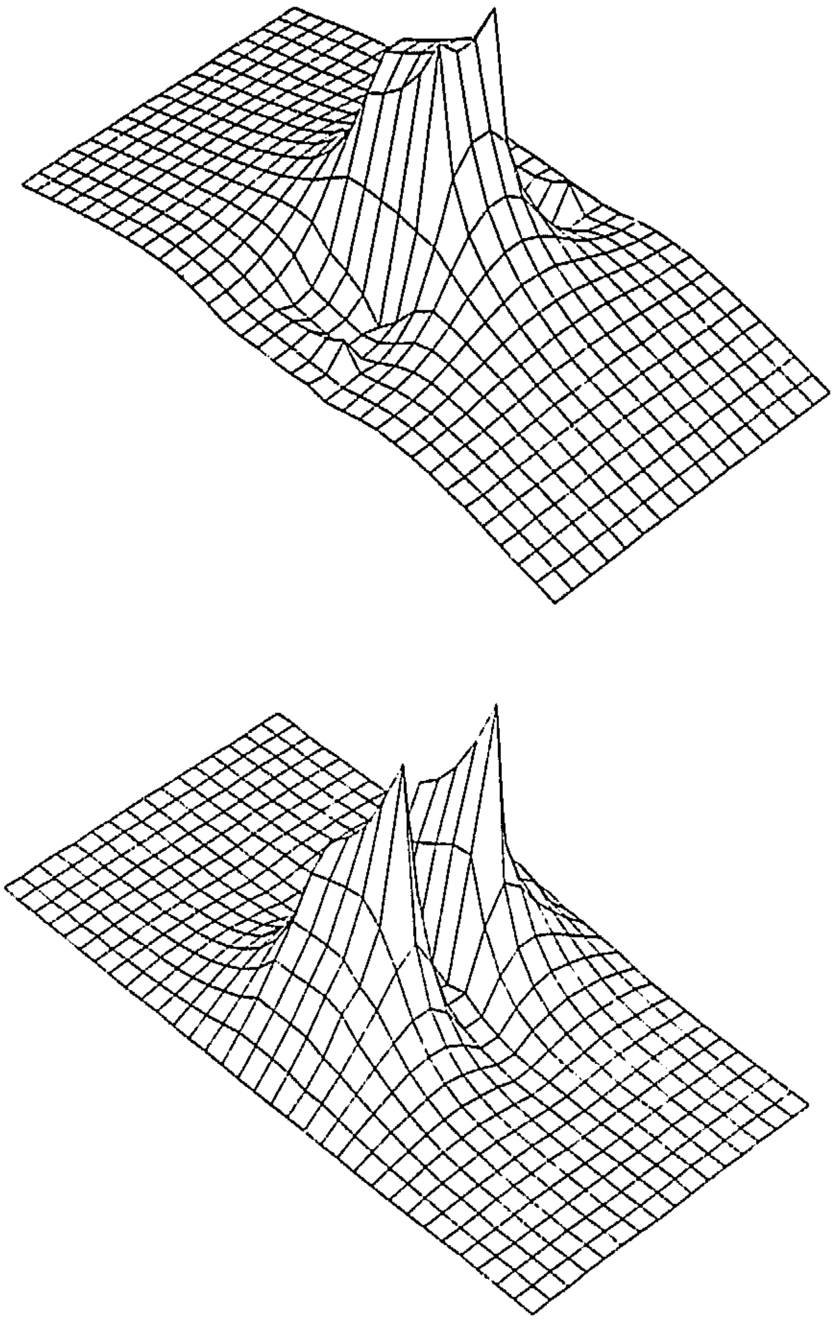

Figure 3. Fin-line mode functions $|e_x(x,y)|$ (top) and $|e_y(x,y)|$ (bottom).

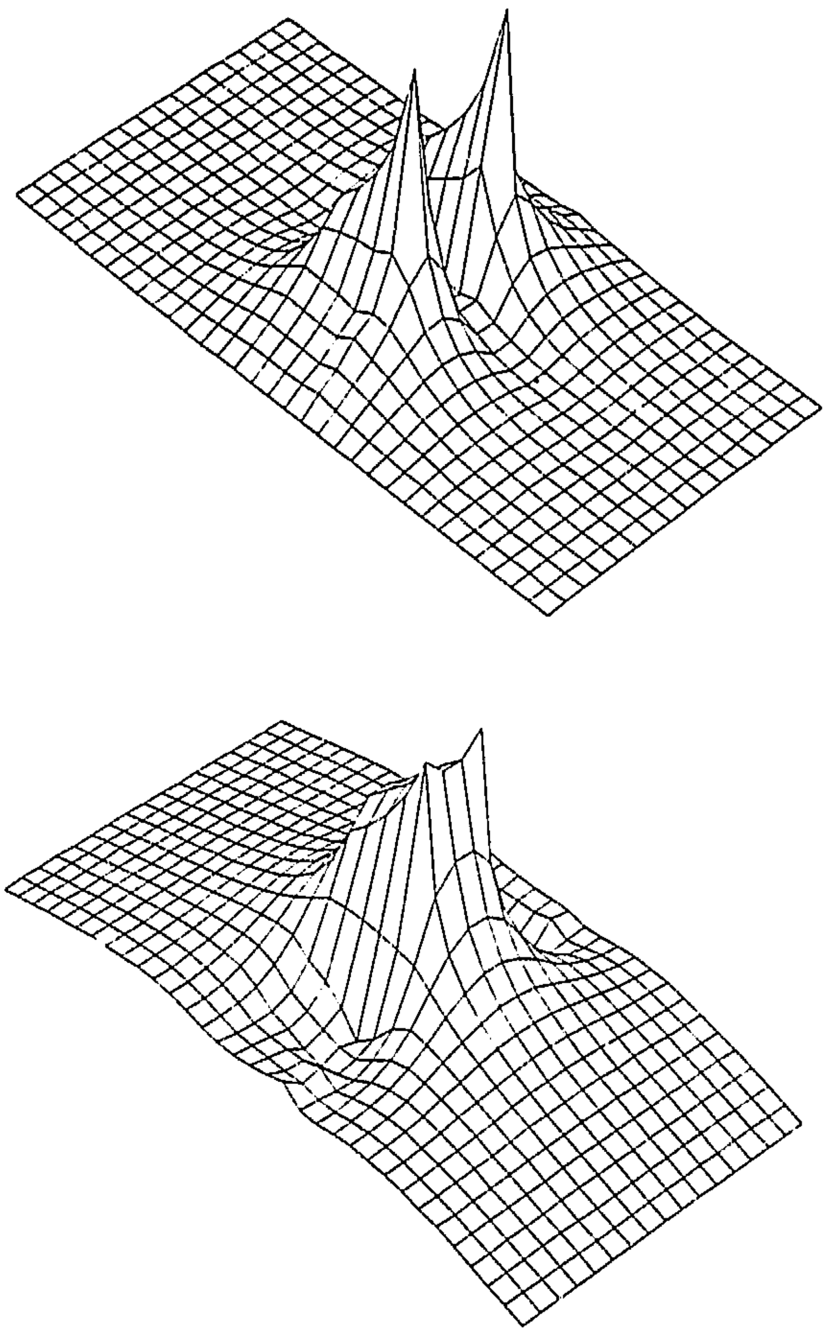

Figure 4. Fin-line mode functions $|h_x(x,y)|$ (top) and $|h_y(x,y)|$ (bottom).

Table 1

Normalized fin-line inner products, w=1 mm, WR-28 shield, f=30 GHz, 2d=.254 mm, m=17, n=32, 2 basis functions (p=0,1 q=2,4), ϵ_r =2.2.

i	j	$\left\| \int\int \bar{e}_i \times \bar{h}_j \cdot \hat{z} \; dx \; dy \right\|$
1	1	1.00
2	2	1.00
3	3	1.00
4	4	1.00
1	2	0.03
1	3	0.18
1	4	0.34

Table 2

Normalized microstrip inner products, w=.318mm, d=.127mm, 2b=.762mm, a=.445mm, f=20 GHz, 2 basis functions (p=1,3 q=2,4), ϵ_r =9.6.

i	j	$\left\| \int\int \bar{e}_i \times \bar{h}_j \cdot \hat{z} \; dx \; dy \right\|$
1	1	1.00
2	2	1.00
3	3	1.00
4	4	1.00
1	2	0.05
1	3	0.01
1	4	0.18

the criterion of mode orthogonality is used.

We are currently studying possible modifications to the spectral Galerkin procedure which should improve the orthogonality of the modes. This would lead to a procedure for solving any problem involving an abrupt printed circuit discontinuity, which is the ultimate goal of this work.

V. EXPERIMENTAL STUDY OF MMIC TRANSITION

Investigations towards the characterization of MMIC devices are now in progress. The scope of the experiments was set to include the complete evaluation of strip-to-substrate and substrate-to-waveguide transitions. Efforts are being directed towards the generation of a data base and developing a test procedure that will provide the user with reproducible test results. Moreover, experimental verification of computed results is of prime importance to this whole program.

Single abrupt discontinuities on a microstrip are presently being evaluated during this initial stage of the experimental program. These discontinuities may be considered as being analogous to lumped reactive elements. Once properly modeled, such elements can be included as part of impedance matching networks.

Various circuit boards have been laid down and tested. Quarter-wave sections with characteristic impedances of 40 and 25 Ω were inserted in series with a 50 Ω line. High permittivity duroid ($\epsilon_R = 10.5$) was used to obtain adequate proportions between the various circuit elements. Swept frequency measurements ranging from 1 to 18 GHz were made. The lower frequency response, i.e., 1 to 5 GHz, is of particular interest since the behavior in this range is more easily predicted by the theoretical models. Direct scaling of these results will then lead us towards predicting the behavior of these circuits at high frequencies.

The actual patterns laid down are shown in Fig. 5. These boards are then placed in the test fixture designed, as shown in Figs. 6a and 6b. It is important to point out that the present experimental setup was designed so as to minimize any discontinuities in the transition from the coaxial connector to the microstrip. The added discontinuities produced by the pin contact and the dielectric pressure bar are

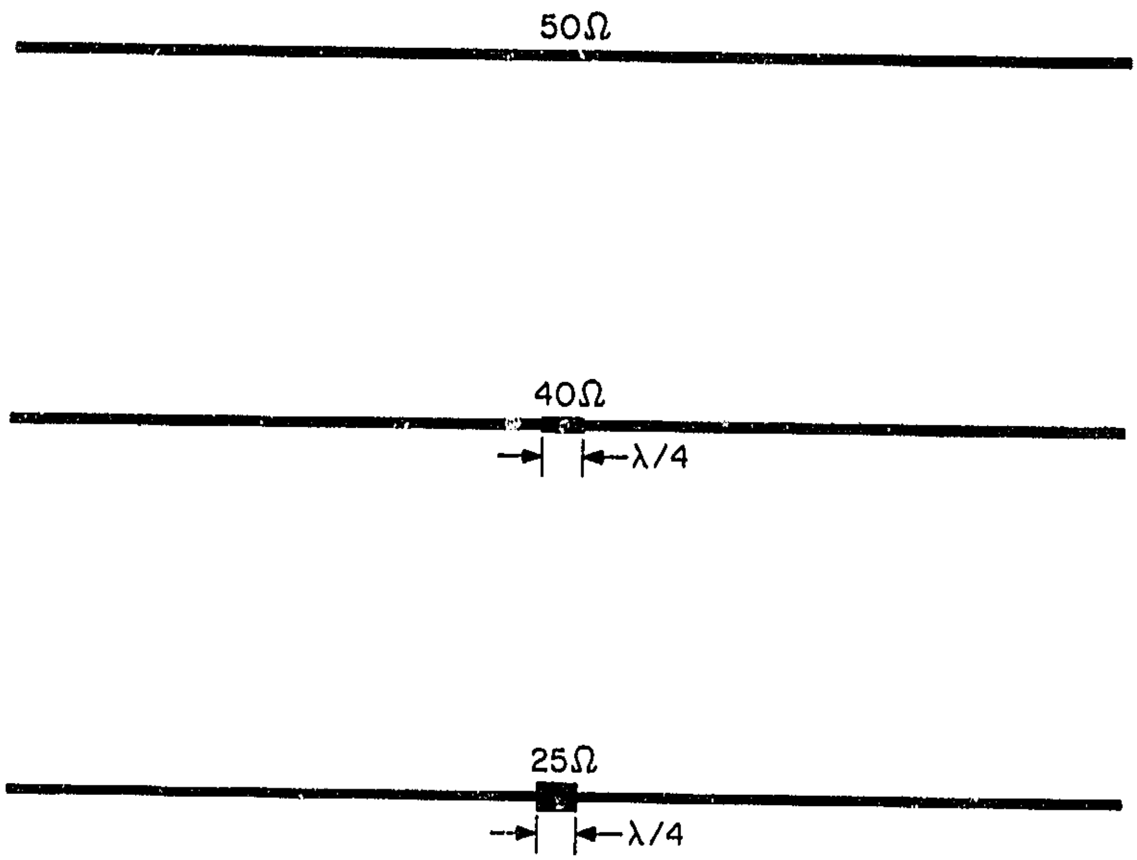

Figure 5. Types of discontinuities being evaluated. 50 Ω line is used as reference.

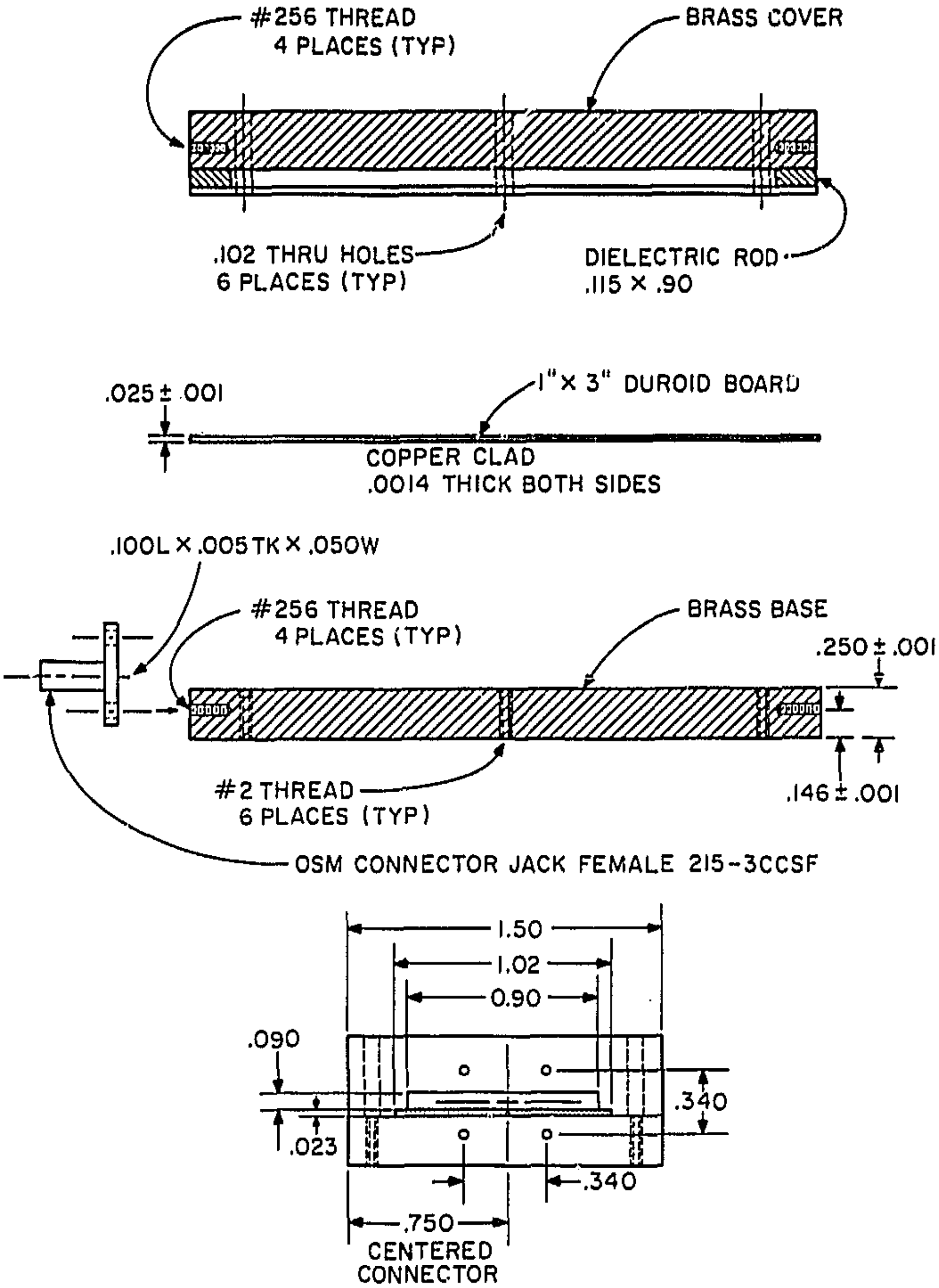

Figure 6a. Mechanical drawing of test fixture.

I. INTRODUCTION

During the course of this contract, the study of printed circuit discontinuities has been carried out using both analytical and experimental approaches. An understanding of these discontinuities is necessary in order to design, for example, transitions between rectangular waveguides and printed circuits. In this section, new developments with respect to the analytical approaches to this problem are discussed. In the section that follows, a summary of the progress in the experimental approach is presented.

The accurate solution for the modes in various millimeter-wave waveguides is essential in the analysis of many integrated circuit components, such as filters and impedance transformers [1]-[5]. Problems associated with the numerical computation of these modes in two frequently used waveguide forms, namely, the finline and microstrip (see Figures 1 and 2), are presented. The spectral domain method of formulation, with a moment method solution, is considered [6]-[8]. This approach can be readily extended to analyze an arbitrary configuration of dielectric and metallized regions in a shielded enclosure. Galerkin's method is used, where the testing and basis functions are the same. It is shown that the mode functions, or eigenfunctions, are more sensitive to errors than the phase constants, or eigenvalues. The approximate mode functions do not satisfy the orthogonality relationship well, resulting in difficulties when these modal solutions are used to form an approximate Green's function or are used in a mode matching analysis [2]-[5], [9]. The significance of using approximate mode functions and procedures to improve the accuracy of modal solutions is described.

II. PERSONNEL

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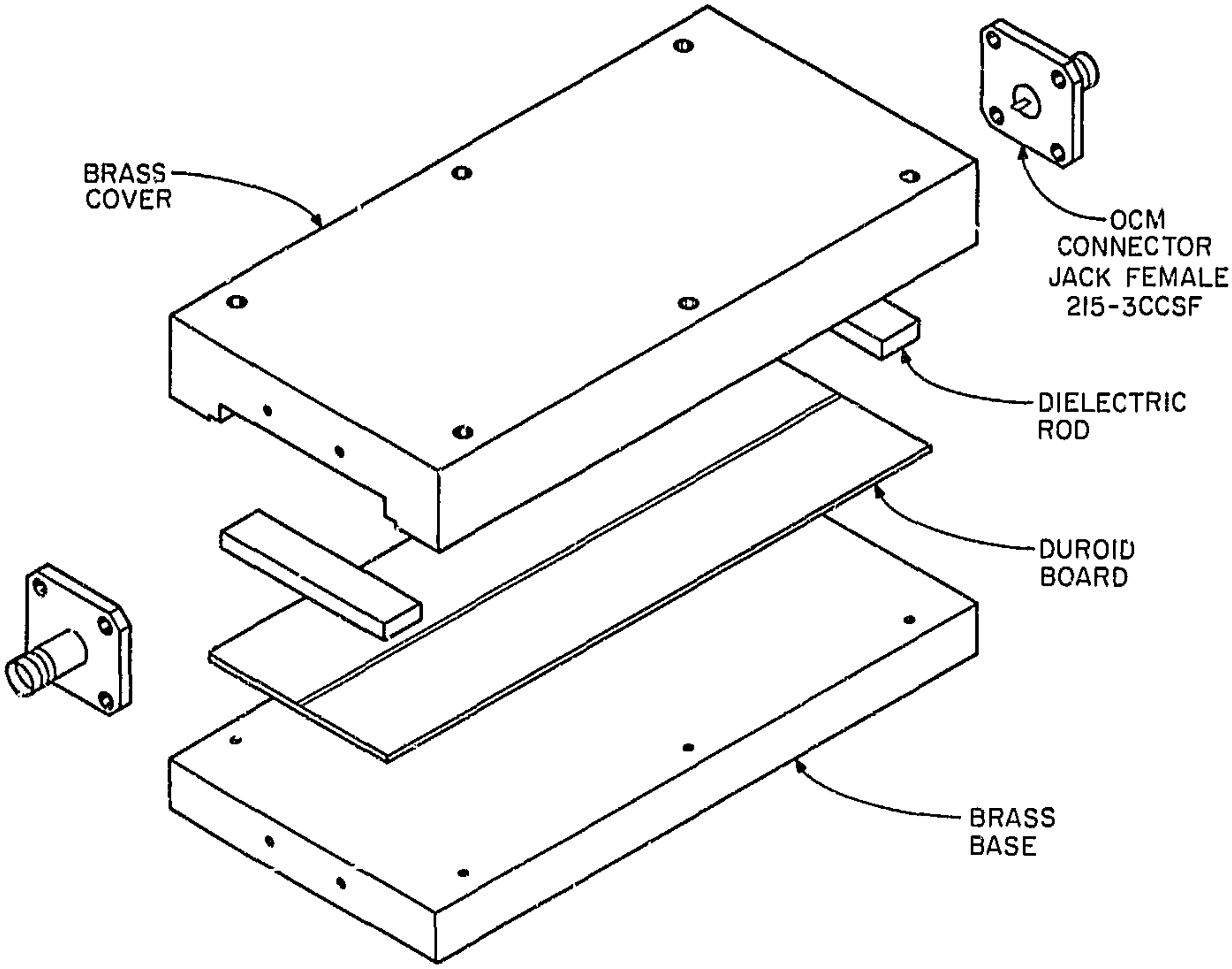

Figure 6b. Assembly drawing of test fixture. Note pressure connection between connector pin and circuit board.

also under investigation. An efficient transition into the microstrip is vital to the success of this project. Some variations of this transition are being studied, including the replacement of the pressure bar by a nylon screw.

Additional valuable information is being generated by the Time Domain Reflectometry measurements. The TDR response permits us to visually determine the magnitude of the various microstrip discontinuities, particularly those at the connector level.

Upon completion of these initial experiments, more elaborate transitions will be evaluated. Preliminary designs of fin-line circuits are being readied. This represents the next step in the determination of the MMIC transitions. During the final stages of the program the Van-Heuven structure will be investigated in detail.

VI. CROSS-TALK IN MULTICONDUCTOR TRANSMISSION LINES

We have investigated the problem of cross-talk and pulse distortion in multiconductor transmission lines with n conductors. To calculate the cross-talk in such a system when it is excited with a pulse whose rise time is short, and when the line terminations are non-linear circuits, e.g., logic gates, is a very difficult problem indeed. First of all, it is difficult to compute the capacitance and inductance matrices of multiconductor lines when the number of conductors is large, say five or more, because the determination of the charge distribution in a large system represents a formidable task. We have investigated this problem using the iterative techniques described in the last section. The second step in the analysis is the computation of the eigenvalues and eigenvectors of the propagation matrix for the modal vectors and propagation constants in the multiconductor system. The third and perhaps the most challenging step is to model the non-linear gates and to incorporate their effects into the reflection of digital signals at the terminations. It should be noted that the conventional Fourier transform techniques for solving transient problems are not applicable here because of the non-linear nature of the terminations. Hence the time-domain differential equations must be solved using some type of a time-stepping procedure. We have developed one such procedure and have obtained some preliminary results for the pulse distortion and cross-talk.

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