Deterministic Approach to Full-Wave Analysis of Discontinuities in MIC's Using the Method of Lines

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Abstract — A deterministic approach to the full-wave analysis of discontinuities in MIC's using the method of lines is described. Arbitrarily shaped one-port and in-line two-port discontinuities in any quasi-planar configurations can be analyzed by this "one step" method. Using the hybrid homogeneous boundary conditions, this approach simplifies the calculation, with ψ^e and ψ^h being discretized only in and near the discontinuity regions. Illustrative examples of S-parameter calculation are given. Computed results are compared with measured data and with the published results of other authors.

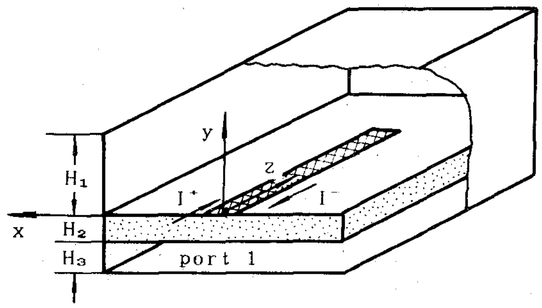

Fig. 1. Discontinuity in MIC shielded structure.

I. Introduction

ACCURATE computation of the frequency-dependent properties of the discontinuities in MIC's (Fig. 1) is important for extending the applicability of existing CAD tools to higher microwave frequencies and millimeter-wave frequencies.

At higher operating frequencies full-wave analysis is more accurate than quasi-static analysis. Due to the complexity of the problem, a rigorous numerical 3-D full-wave approach has not yet been elaborated. In recent years, some full-wave analyses of the discontinuities in MIC's/MMIC's have been published [1]–[18]. A unified user-oriented full-wave computation method for the analysis of arbitrarily shaped discontinuities in any quasi-planar configuration is an attractive subject.

The method of lines [19]–[25] is more flexible than other methods. It avoids the choice of basis functions and the relative convergence problem. It provides a simple means for dealing with complex structures and was applied to discontinuity problems by Worm *et al.* [21], [23], [24]. However, for nonperiodic discontinuity problems, if only the Dirichlet and the Neumann homogeneous boundary conditions are applied, ψ^e and ψ^h must be discretized in both the discontinuity region and the connected open/short ended transmission line sections. The maximum of the electrical length of the transmission line sections is π, so the number of discretization lines is very large.

Recently, the authors [26] presented an eigenvalue approach using the method of lines to discontinuity problems in quasi-planar configurations by introducing the hybrid homogeneous boundary conditions. The approach requires that ψ^e and ψ^h be discretized only in and near the discontinuity regions, so it markedly reduces the computation effort.

In this paper, we present a deterministic full-wave approach to the problem for the case of shielded structures. The deterministic approach avoids solving the eigenvalue equation, which requires many computation steps to find the root. It is a "one-step" solution and is more suitable for practical applications than the eigenvalue approach.

Manuscript received June 13, 1988; revised October 5, 1988. This work was supported by the National Science Foundation of China.
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IEEE Log Number 8825393.

II. Analysis

The method of lines combines the advantages of the analytic and the discrete methods. The principles underlying this method have been described by Worm *et al.* [19]–[24]. The electromagnetic field in each layer of the dielectric is described by two scalar potential functions ψ^e and ψ^h:

$$\vec{E} = \nabla \times \nabla \times (\psi^e \hat{z})/(j\omega\epsilon) - \nabla \times (\psi^h \hat{z}) \tag{1}$$

$$\vec{H} = \nabla \times (\psi^e \hat{z}) + \nabla \times \nabla \times (\psi^h \hat{z})/(j\omega\mu). \tag{2}$$

The functions ψ^e and ψ^h are discretized in the x and the z direction shown in Fig. 1 with nonequidistant mesh width h_{xi} $(i=1,2,\cdots,N_{hx})$, e_{xi} $(i=1,2,\cdots,N_{ex})$, h_{zi} $(i=1,2,\cdots,N_{hz})$, e_{zi} $(i=1,2,\cdots,N_{ez})$; then they will be denoted by the matrices $[\psi^e]$ and $[\psi^h]$, respectively. The transformation of matrix notation into vector notation can be expressed as follows:

$$[A][X][B] \rightarrow [B]^T \otimes [A]\overline{X} \tag{3}$$

where $\overline{X} = [x_{11}, x_{21}, \cdots, x_{m1}, x_{12}, x_{22}, \cdots, x_{m2}, x_{13}, x_{23}, \cdots, x_{mn}]^T$; $\otimes$ denotes the Kronecker product [24]; and

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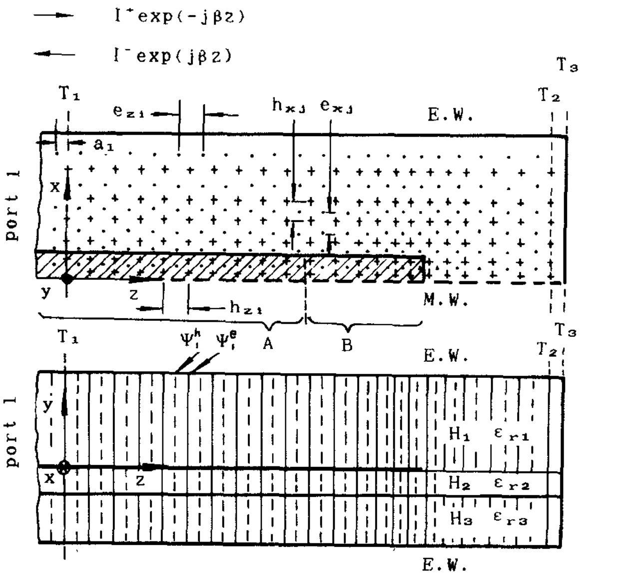

Fig. 2. Discretization of ψ^e and ψ^h for one-port discontinuity + ψ^e lines, ● ψ^h lines.

$[A]$, $[B]$, and $[X]$ are $M \times M$, $N \times N$, and $M \times N$ matrices, respectively.

Fig. 2 shows the discretization of ψ^e and ψ^h lines in one-port discontinuity (open end) structure. It is assumed that only one propagation mode exists in the transmission line connected to port 1 and that the plane T_1 is so far from the discontinuity region that the higher order modes excited by the discontinuity will vanish at T_1.

We take the incident wave as the impressed wave, denoted by $\psi_I^{e,h}$, and the reflected wave as the excited wave, denoted by $\psi_E^{e,h}$. According to transmission line theory, near the plane T_1 $\psi_{I,E}^e$ can be expressed as

$$\psi_{I,E}^e = C_{I,E}(x,y)e^{\mp j\beta_1 z} \tag{4}$$

where β_1 is the propagation constant of the transmission line at port 1.

At $z=0$, ψ_I^e and ψ_E^e satisfy the hybrid homogeneous boundary conditions:

$$\left.\frac{\partial \psi_{I,E}^e}{\partial z}\right|_{z=0} = \mp j\beta C_{I,E} = \mp j\beta e^{\pm j\beta_1 a_1}\psi_{I,E_1}^e \equiv A^{\pm}\psi_{I,E_1}^e \tag{5}$$

where A^+ and A^- are complex constants. The derivative of $[\psi_I^{e,h}]$ and $[\psi_E^{e,h}]$ with respect to z can be transformed to matrix operation forms:

$$\frac{\partial[\psi_{I,E}^e]}{\partial z} \to \frac{1}{hz}[\Phi_{I,E}^e][D_z^{I,E}]^T[r_{hz}^{I,E}] \tag{6}$$

with

$$[D_z^{I,E}] = [r_{hz}^{I,E}]\begin{bmatrix} 1 & & 0 \\ -1 & \ddots & \\ 0 & -1 & 1 \end{bmatrix}[r_{ez}^{I,E}]$$

for Fig. 2 with E.W. at T_3 (7)

or

$$[D_z^{I,E}] = [r_{hz}^{I,E}]\begin{bmatrix} 1 & & 0 \\ -1 & \ddots & 1 \\ 0 & & -1 \end{bmatrix}[r_{ez}^{I,E}]$$

for Fig. 2 with M.W. at T_2 (8)

$$[r_{hz}^{I,E}] = \begin{bmatrix} \sqrt{h_{z1}A^{+,-}}\cdot r_{hz1} & & & 0 \\ & r_{hz2} & & \\ & & \ddots & \\ 0 & & & r_{hzNhz} \end{bmatrix} \tag{9}$$

$$[r_{ez}^{I,E}] = \begin{bmatrix} r_{ez1} & 0 & \ddots \\ 0 & & r_{ezNez} \end{bmatrix} \tag{10}$$

$$[\Phi_{I,E}^e] = [r_{ex}]^{-1}[\psi_{I,E}^e][r_{ez}^{I,E}]^{-1} \tag{11}$$

$$\frac{\partial[\psi_{I,E}^h]}{\partial z} \to -\frac{1}{h_z}[\Phi_{I,E}^h][D_z^{I,E}][r_{ez}^{I,E}] \tag{12}$$

$$[\Phi_{I,E}^h] = [r_{hx}]^{-1}[\psi_{I,E}^h][r_{hz}^{I,E}]^{-1} \tag{13}$$

$$\frac{\partial^2[\psi_{I,E}^e]}{\partial z^2} \to -\frac{1}{h_z^2}[\Phi_{I,E}^e][D_z^{I,E}]^T[D_z^{I,E}][r_{ez}^{I,E}] = \frac{1}{h_z^2}[\Phi_{I,E}^e][D_{zz}^{eI,E}][r_{ez}^{I,E}] \tag{14}$$

$$\frac{\partial^2[\psi_{I,E}^h]}{\partial z^2} \to -\frac{1}{h_z^2}[\Phi_{I,E}^h][D_z^{I,E}][D_z^{I,E}]^T[r_{hz}^{IE}] = \frac{1}{h_z^2}[\Phi_{I,E}^h][D_{zz}^{hI,E}][r_{hz}^{I,E}]. \tag{15}$$

The second-order operators $[D_{zz}^{eI,E}]$ and $[D_{zz}^{hI,E}]$ are complex symmetric tridiagonal matrices. By means of the Kronecker product it has been shown that the discretization in the x and z directions is independent and that operator equations for one direction have an obvious analogy in the other direction. Therefore, with the exception of (9), equations (6)–(15) are the same as (5)–(14) in [22] but are formulated for the z direction.

The transformations of the derivatives of $[\psi_{I,E}^e]$ and $[\psi_{I,E}^h]$ with respect to the x direction are the same as those in [22]. By the transformations $[V_{I,E}^e] = [T_{ex}]^T[\Phi_{I,E}^e][T_{ez}^{I,E}]$ and $[V_{I,E}^h] = [T_{hx}]^T[\Phi_{I,E}^h][T_{hz}^{I,E}]$, Helmholtz equations for $[\psi_{I,E}^e]$ and $[\psi_{I,E}^h]$ can be transferred to uncoupled differential equations:

$$\frac{d^2V_{I,Eik}^e}{dy^2} + \left(\frac{[\lambda_{xx}^e]_{ii}}{h_x^2} + \frac{[\lambda_{zz}^{eI,E}]_{kk}}{h_z^2} + \epsilon_r k_0^2\right)V_{I,Eik}^e = 0$$

$$(i=1,2,\cdots,N_{ex};\ k=1,2,\cdots,N_{ez}) \tag{16}$$

$$\frac{d^2V_{I,Eik}^h}{dy^2} + \left(\frac{[\lambda_{xx}^h]_{ii}}{h_x^2} + \frac{[\lambda_{zz}^{hI,E}]_{kk}}{h_z^2} + \epsilon_r k_0^2\right)V_{I,Eik}^h = 0$$

$$(i=1,2,\cdots,N_{hx};\ k=1,2,\cdots,N_{hz}) \tag{17}$$

where $[T_{ez}^{I,E}]$ denotes the normalized eigenvector matrices of $[D_{zz}^{eI,E}]$, $[\lambda_{zz}^{eI,E}]$ denotes their eigenvalue diagonal matrices, and $[T_{hz}^{I,E}]$, $[\lambda_{zz}^{hI,E}]$, $[T_{ex}]$, $[\lambda_{xx}^e]$, $[T_{hx}]$, and $[\lambda_{xx}^h]$ are defined in a similar way. We employ the subroutines in EISPACK [30] to compute $[T]$ and $[\lambda]$. Using the method proposed by Worm *et al.* [21], [22] we can obtain

$$\begin{bmatrix} [\tilde{Z}_{11}^{I,E}] & [\tilde{Z}_{12}^{I,E}] \\ [\tilde{Z}_{21}^{I,E}] & [\tilde{Z}_{22}^{I,E}] \end{bmatrix} \begin{bmatrix} \tilde{\bar{J}}_z^{I,E} \\ \tilde{\bar{J}}_x^{I,E} \end{bmatrix} = \begin{bmatrix} \tilde{\bar{E}}_z^{I,E} \\ \tilde{\bar{E}}_x^{I,E} \end{bmatrix} \quad \text{for strip structures} \tag{18}$$

or

$$\begin{bmatrix} [\tilde{Y}_{11}^{I,E}] & [\tilde{Y}_{12}^{I,E}] \\ [\tilde{Y}_{21}^{I,E}] & [\tilde{Y}_{22}^{I,E}] \end{bmatrix} \begin{bmatrix} \tilde{\bar{E}}_z^{I,E} \\ \tilde{\bar{E}}_x^{I,E} \end{bmatrix} = \begin{bmatrix} \tilde{\bar{J}}_z^{I,E} \\ \tilde{\bar{J}}_x^{I,E} \end{bmatrix} \quad \text{for slot structures} \tag{19}$$

and then

$$[Z^{I,E}]_{\text{red}} \begin{bmatrix} \bar{J}_z^{I,E} \\ \bar{J}_x^{I,E} \end{bmatrix}_{\text{strip}} = \begin{bmatrix} \bar{E}_z^{I,E} \\ \bar{E}_x^{I,E} \end{bmatrix}_{\text{strip}} \tag{20}$$

or

$$[Y^{I,E}]_{\text{red}} \begin{bmatrix} \bar{E}_z^{I,E} \\ \bar{E}_x^{I,E} \end{bmatrix}_{\text{slot}} = \begin{bmatrix} \bar{J}_z^{I,E} \\ \bar{J}_x^{I,E} \end{bmatrix}_{\text{slot}}. \tag{21}$$

The details of deriving (18)–(21) can be found in [19]–[24]. In our approach $[\delta_z^{I,E}] = [T_{hz}^{I,E}][D_z^{I,E}][T_{ez}^{I,E}]$ are complex matrices which retain quasi-diagonal properties of the existing representations [21], [22]. $[\tilde{Y}_{ij}]$ or $[\tilde{Z}_{ij}]$ can be carried out using only quasi-diagonal matrices. Obviously, the large matrix storage requirement can be alleviated and the computation time can be reduced. Sometimes we have to suitably rearrange the columns of $[T_{hz}^{I,E}]$ or $[T_{ez}^{I,E}]$ and the corresponding $[\lambda_{e,hz}^{I,E}]$ in order to get a standard form of the quasi-diagonal matrix $[\delta_z^{I,E}]$.

Now we discuss the strip structures shown in Fig. 2. We split up the strip region into two parts: transmission line region A and discontinuity region B. The sum of the number of ψ^e and ψ^h lines in region A should be larger than that in region B. By rearranging the rows and the columns of the matrix $[Z^{I,E}]$, we obtain

$$\begin{bmatrix} [Z_{11}^{I,E}] & [Z_{12}^{I,E}] \\ [Z_{21}^{I,E}] & [Z_{22}^{I,E}] \end{bmatrix} \begin{bmatrix} \bar{J}^{AI,E} \\ \bar{J}^{BI,E} \end{bmatrix} = \begin{bmatrix} \bar{E}^{AI,E} \\ \bar{E}^{BI,E} \end{bmatrix} \tag{22}$$

where

$$\bar{J}^{AI,E} = \begin{bmatrix} \bar{J}_z^{AI,E} \\ \bar{J}_x^{AI,E} \end{bmatrix} \qquad \bar{J}^{BI,E} = \begin{bmatrix} \bar{J}_z^{BI,E} \\ \bar{J}_x^{BI,E} \end{bmatrix} \tag{23}$$

$$\bar{E}^{AI,E} = \begin{bmatrix} \bar{E}_z^{AI,E} \\ \bar{E}_x^{AI,E} \end{bmatrix} \qquad \bar{E}^{BI,E} = \begin{bmatrix} \bar{E}_z^{BI,E} \\ \bar{E}_x^{BI,E} \end{bmatrix}. \tag{24}$$

The superposition of the impressed field and the excited field must satisfy the tangential condition for the conductor surface:

$$\bar{E}_z^{A,B} = \bar{E}_z^{A,BI} + \bar{E}_z^{A,BE} = \bar{0} \tag{25}$$

$$\bar{E}_x^{A,B} = \bar{E}_x^{A,BI} + \bar{E}_x^{A,BE} = \bar{0} \tag{26}$$

so we have

$$\begin{bmatrix} [Z_{11}^{E}] & [Z_{12}^{E}] \\ [Z_{21}^{E}] & [Z_{22}^{E}] \end{bmatrix} \begin{bmatrix} \bar{J}^{AE} \\ \bar{J}^{BE} \end{bmatrix} = - \begin{bmatrix} \bar{E}^{AI} \\ \bar{E}^{BI} \end{bmatrix}. \tag{27}$$

Since region A is a transmission line section connected to port 1, $\bar{J}_z^{AI}$ and $\bar{J}_x^{AI}$ can be assigned according to the results of the 2-D analysis [20], [22] of the transmission line at port 1. In the z direction the factor $e^{-j\beta z}$ should be taken into account. $\bar{J}_z^{BI}$ and $\bar{J}_x^{BI}$ are determined in such a way that the electromagnetic field described by $\psi_I^{e,h}$ in region A remains in the "incident" state, so the equations $\bar{E}_z^{AI} = \bar{0}$ and $\bar{E}_x^{AI} = \bar{0}$ must be satisfied. In this way, the equations are obtained as follows:

$$[Z_{11}^{I}]\bar{J}^{AI} + [Z_{12}^{I}]\bar{J}^{BI} = \bar{0} \tag{28}$$

$$[Z_{11}^{E}]\bar{J}^{AE} + [Z_{12}^{E}]\bar{J}^{BE} = \bar{0} \tag{29}$$

$$[Z_{21}^{I}]\bar{J}^{AI} + [Z_{22}^{I}]\bar{J}^{BI} = \bar{E}^{BI} \tag{30}$$

$$[Z_{21}^{E}]\bar{J}^{AE} + [Z_{22}^{E}]\bar{J}^{BE} = -\bar{E}^{BI} \tag{31}$$

and then

$$\bar{J}^{BI} = -[Z_{12}^{I}]^{-1}[Z_{11}^{I}]\bar{J}^{AI} \tag{32}$$

$$\bar{J}^{BE} = \left([Z_{21}^{E}][Z_{11}^{E}]^{-1}[Z_{12}^{E}] - [Z_{22}^{E}]\right)^{-1} \cdot \left([Z_{21}^{I}] - [Z_{22}^{I}][Z_{12}^{I}]^{-1}[Z_{11}^{I}]\right)\bar{J}^{AI} \tag{33}$$

$$\bar{J}^{AE} = -[Z_{11}^{E}]^{-1}[Z_{12}^{E}]\bar{J}^{BE} \tag{34}$$

where $[\;]^{-1}$ denotes the Moore–Penrose generalized inverse matrix. The S parameter of the one-port can be calculated by

$$S_{11} = I_1^- / I_1^+ \tag{35}$$

$$I_1^+ = \sum_{i=1}^{N_1} J_i^{AI} e_{xk_1(i)} \tag{36}$$

$$I_1^- = -\sum_{i=1}^{N_1} J_i^{AE} e_{xk_1(i)} \tag{37}$$

where N_1 is the total number of ψ^e lines on the conductor strip at plane T_1, and $k_1(i)$ is the x-direction subscript of the ith ψ^e line on the conductor strip at T_1.

Fig. 3 shows the discretization of ψ^e and ψ^h lines for an in-line two-port discontinuity structure. To calculate S_{11} and S_{21}, we consider that there is no incident wave at port 2, so the boundary conditions of port 2 for both impressed and excited fields are the same, i.e., the reflect types. Similar to the one-port treatment, the following results are

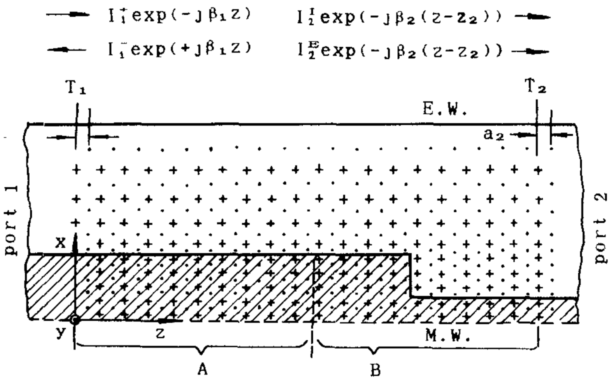

Fig. 3. Discretization of ψ^e and ψ^h for in-line two-port discontinuity. + ψ^e lines, ● ψ^h lines.

obtained:

$$\left.\frac{\partial \psi^h_{I,E}}{\partial z}\right|_{z=0} = \mp j\beta_1 e^{\pm j\beta_1 a_1}\psi^h_{I,E_1} = G^{\pm}\psi^h_{I,E_1} \tag{38}$$

$$\left.\frac{\partial \psi^e_{I,E}}{\partial z}\right|_{z=z_2+a_2} = -j\beta_2 e^{-j\beta_2 a_2}\psi^e_{I,ENez} = -B^{-}\psi^e_{I,ENez} \tag{39}$$

$$\left[r^{I,E}_{ez}\right] = \begin{bmatrix} \sqrt{e_{z1}G^{+,-}}\cdot r_{ez1} & & & 0 \\ & r_{ez2} & & \\ & & \ddots & \\ 0 & & & r_{ezNez} \end{bmatrix} \tag{40}$$

$$\left[r^I_{hz}\right] = \left[r^E_{hz}\right] = \begin{bmatrix} r_{hz1} & & & 0 \\ & \ddots & & \\ & & r_{hz(Nhz-1)} & \\ 0 & & & \sqrt{h_{zNhz}\cdot B^{-}}\cdot r_{hzNhz} \end{bmatrix} \tag{41}$$

$$\left[D^{I,E}_z\right] = \left[r^{I,E}_{hz}\right]\begin{bmatrix} -1 & 1 & & 0 \\ & \ddots & \ddots & \\ & & & 1 \\ 0 & & & -1 \end{bmatrix}\left[r^{I,E}_{ez}\right]. \tag{42}$$

S_{11} can be calculated in a way similar to that of the one-port discontinuity case. For S_{21}, we have

$$S_{21} = \frac{I_2^-}{I_1^+}\sqrt{\frac{Z_{L2}}{Z_{L1}}} \tag{43}$$

where Z_{L1} and Z_{L2} are the characteristic impedances of transmission lines at port 1 and port 2, respectively, based on the current–power definition:

$$I_2^- = \sum_{i=1}^{N_2}\left(J^{BI}_{M_2-N_2+i} + J^{BE}_{M_2-N_2+i}\right)\cdot e_{xk_2(i)} \tag{44}$$

where M_2 denotes the number of ψ^e lines in region B, N_2 equals the number of ψ^e lines on the conductor strip at T_2, and k_2 (i) is the x-direction subscript of the ψ^e line, which is the ith on the conductor strip at T_2.

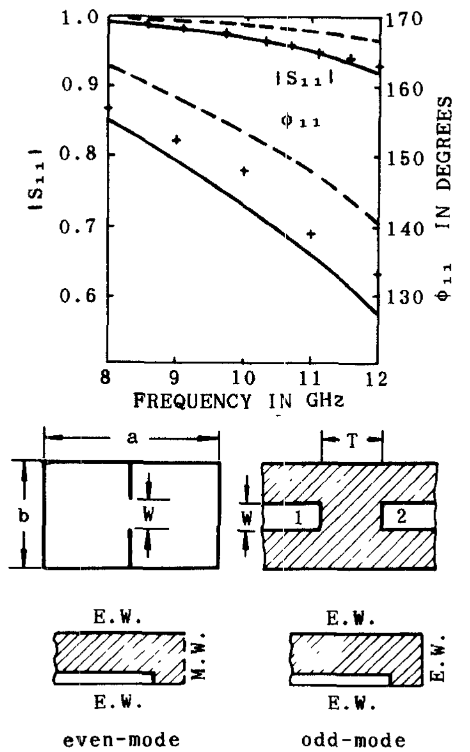

Fig. 4. Analysis of a finline inductor strip. $T = 5.08$ mm, $W/b = 0.25$, WR(90) waveguide. + + + measured by Knorr *et al.* [4]; --- computed by Knorr et al. [4], ——— computed by this method.

The equations for the slotline structures are dual to those of strip structures. We use U instead of I. The characteristic impedance Z_L is based on the voltage–power definition.

III. Results

Based on the method described above, we have developed a computer program for calculating the S parameters of arbitrarily shaped one-port and in-line two-port discontinuities in quasi-planar configurations with three dielectric layers and one conductor strip layer. The computations were performed on the ELXSI 6400 system. Another version of the program with smaller array dimensions can be run on the IBM PC/AT.

Since we assumed that only one propagation mode exists at T_1 shown in Fig. 2 and at T_1, T_2 shown in Fig. 3, the locations of the above-mentioned planes T_i should be far enough from the discontinuity. By applying principles for the nonequidistant discretization [22], we can achieve this condition without using an extremely large number of discretization lines.

A finline inductive strip is analyzed by the even- and the odd-mode method (Fig. 4). Our results are closer to the measured results by Knorr *et al.* [4].

An example of a more complex structure is shown in Figs. 5 and 6. It contains a branch line coupler. Four branch lines are located close to each other. As Jansen [7] pointed out, parasitic coupling and spurious parameters are so strong that the usefulness of a decomposition to four T junctions and four transmission line sections is

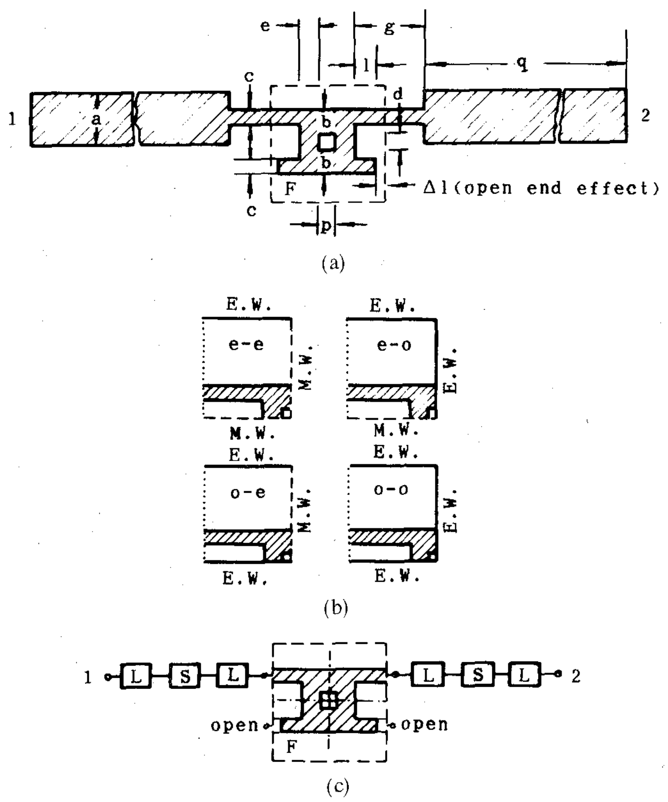

Fig. 5. A two-port network containing a branch line coupler. $H = 0.75$ mm, $\epsilon_r = 2.86$, $a = 2.24$ mm, $b = 0.81$ mm, $c = 0.47$ mm, $d = 0.84$ mm, $e = 0.53$ mm, $g = 3.05$ mm, $l = 1.00$ mm, $p = 0.87$ mm, $q = 26.00$ mm. L: microstrip transmission line; S: microstrip step discontinuity; F: four-port network.

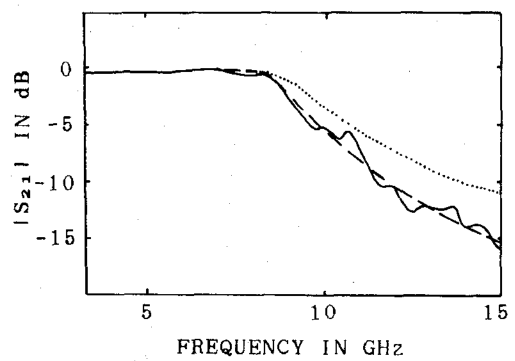

Fig. 6. $|S_{21}|$ of the two-port network shown in Fig. 5(a). ——— measured, ---- computed by this method, ······· computed by decomposition of the branch line coupler.

questionable. In our computation, the branch line coupler is treated as a whole. The one-port S parameters of four modes, i.e., $e-e, e-o, o-e, o-o$ (see Fig. 5(b)) are computed first and the four-port S parameters are obtained by the existing method [29]. After obtaining four-port S parameters, we calculated the two-port S parameters of the circuit shown in Fig. 5(a). Fig. 6 shows the computed results using the present method compared to measured results and the results obtained by adopting the decomposition of the branch line coupler. A general-purpose microstrip CAD program, developed by the authors [27] in 1985 using the microstrip waveguide model mode-matching method [28], is applied to consider the microstrip discontinuity effects (Fig. 5(c)).

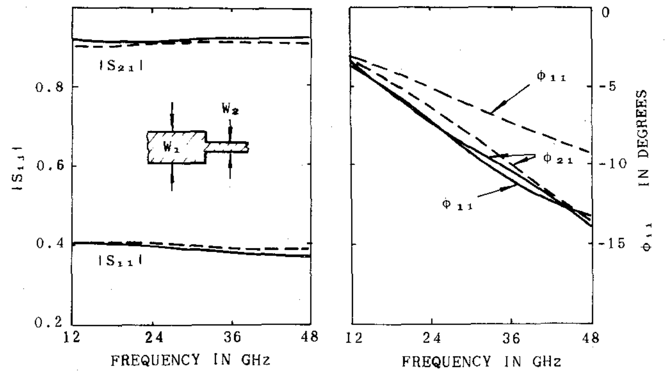

Fig. 7. Computation results of the S parameters of a microstrip step discontinuity. $W_2/H = 1.0$, $W_1/H = 4.0$, $\epsilon_r = 10.0$, $H = 0.25$ mm. ——— this method; ---- Jansen [11].

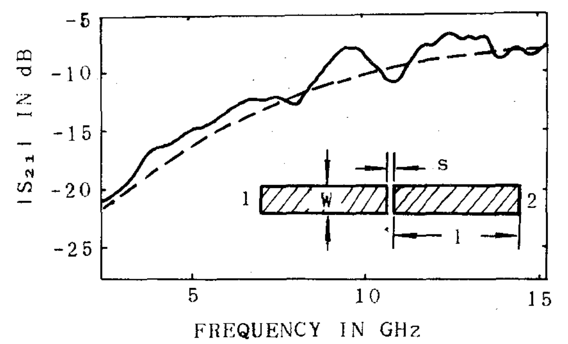

Fig. 8. Computed and measured $|S_{21}|$ of a microstrip gap discontinuity. $S = 0.15$ mm, $W = 2.23$, mm, $H = 0.75$ mm, $\epsilon_r = 2.86$. ——— measured; ---- computed.

The analysis of a microstrip step discontinuity up to millimeter frequencies is shown in Fig. 7 compared to results by Jansen.

Fig. 8 shows $|S_{21}|$ of a microstrip gap discontinuity. One of the reasons for the difference between the computed and the measured results may be the neglect of the finite thickness of the strip.

IV. Conclusions

A deterministic approach to the full-wave analysis of discontinuities in MIC's using the method of lines has been proposed. This method is flexible for arbitrarily shaped discontinuities in any quasi-planar configurations and valid up to millimeter-wave frequencies. It provides useful results with sufficient accuracy for practical applications. The derivation procedure can be easily extended to the structures consisting of more strip layers and more dielectric layers.

Acknowledgment

The authors would like to thank Dr. Gong Ke for his help in revising this paper.

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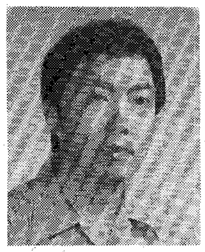

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