An Adaptive Spectral Response Modeling Procedure for Multiport Microwave Circuits

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Abstract —An adaptive scheme is proposed to generate the spectral response of waveguide junctions in minimum computation time. The procedure uses the newly developed transfinite element method to determine the fields in junctions at a few adaptively selected frequencies and then employs these solutions to generate the spectral response throughout the frequency range of interest. In typical problems, the method converges in five or six iterations to the full spectral response evaluated at 100 points. We show by solving example problems that the new procedure is orders of magnitude faster than the alternatives.

I. Introduction

MICROWAVE CIRCUITS in use today often employ planar geometries that may be represented mathematically as a multiport microwave junction. A multiport microwave junction is, in general, defined as a structure that consists of an arbitrarily shaped cavity, with or without dielectrics, and has ports coupling in and out of the cavity. A number of studies have been made of microwave junction problems [1]–[4]. However, the classical analyses have been largely confined to networks of simple shape or of geometry that lends itself to analytical or semianalytical methods of solution.

The highly complex geometries used in microwave circuits today makes it necessary to use numerical methods for analysis. Multiport microwave junctions are solved numerically in the literature by using one of two approaches: the eigensolution method [5], [6] and the deterministic method [7]–[9]. In the eigensolution method, either the finite element method or the finite difference method is used to compute the eigenvalues and the eigenvectors of the normal modes of the junction, and then circuit theory is used to determine the circuit parameters [10]. In the deterministic method, the field solution is computed at a single specified frequency and scattering parameters are computed at that frequency only; the entire process must be repeated to determine the solution at other frequencies.

While the eigensolution method is mathematically elegant, it has the disadvantage of requiring the solution of large matrix eigenvalue equations. Since the solution of matrix eigenvalue problems is expensive, recent work has focused on the deterministic approach, which requires the solution of deterministic matrix equations only [7], [8]. Recently, Webb [8] has used the finite element method for the analysis of H-plane rectangular waveguide problems. In his procedure, a set of boundary value problems is solved in order to get the field solution at a single frequency point. Two similar procedures have been developed by Koshiba. In [7], the boundary element is combined with modal analysis to solve waveguide discontinuity problems, and the finite element method is combined with modal analysis to solve for fields in an H-plane waveguide circulator in [9]. Unfortunately, however, both of these procedures result in nonsymmetric matrix equations that are expensive to solve.

In this paper, we introduce a new, highly efficient procedure for modeling multiport waveguide junctions. The basis of the procedure is the transfinite element method [11], [12], in which modal basis functions are combined with finite element basis functions to provide solutions for open boundary problems. This procedure results in symmetric sparse matrix equations that can be solved very efficiently by using the preconditioned biconjugate gradient algorithm. Further, we develop a spectral response estimation procedure by which solutions at a few adaptively selected frequencies are used to generate the full solution in the frequency range of interest.

Numerical results are given for a T junction, a screen filter containing metal inserts, and a dielectric filter to show the validity of the present procedure. A comparison of the computation times required by the adaptive procedure and by the direct deterministic procedure for the T junction problem is also presented.

II. Formulation

A. The Transfinite Element Method

The structure to be analyzed consists of a cavity coupled with some rectangular waveguides. The shape of the cavity and the dimensions of the waveguides are arbitrary, but the overall structure must be uniform so that the problem can be approximated by two-dimensional analysis. To simplify the formulation, we assume that the junction is in the H plane: problems involving E-plane junctions can be

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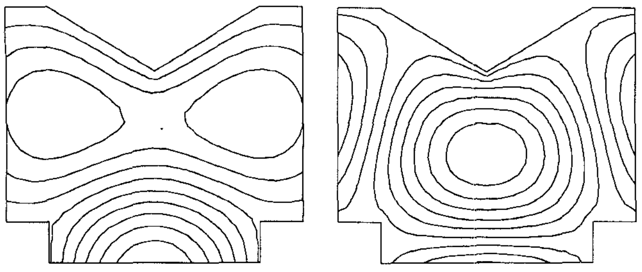

(a) (b)

Fig. 1. Transfinite element solution of a microwave T junction at 225 MHz. (a) Real part of the constant E field. (b) Imaginary part.

treated in much the same way. Microwave planar circuit problems differ in that the electric field is taken to be a constant perpendicular to the plane of the circuit provided that effective dimensions are used to account for the fringing fields.

Consider exciting port 1 by the dominant TE_{10} mode. The field over the cavity region Ω_i and the port regions extending to infinity must satisfy the Helmholtz equation:

$$\nabla^2 E + k^2 \epsilon_r E = 0$$

where

$$k^2 = \omega^2 \epsilon_0 \mu_0. \quad (1)$$

Here ω is the angular frequency of the excitation, ϵ_0 and μ_0 are the permittivity and the permeability of free space, respectively, and ϵ_r is the relative dielectric constant of the material.

In the transfinite element method [11], [12], the problem region is divided into two parts. An interior region Ω_i of finite extent and an exterior region Ω_e that is homogeneous and unbounded. Within Ω_i finite element basis functions are used to approximate the field; in Ω_e analytical solutions of the Helmholtz equation provide a basis set for the field. Employing both sets of basis functions in a variational procedure and requiring continuity along the boundary between Ω_i and Ω_e gives a symmetric matrix equation that is solved for the field.

The application of the transfinite element method to multiport microwave circuits is presented in [12]. In this procedure, (1) is converted into the matrix equation

$$([\hat{S}] - k^2[\hat{T}] + [\hat{\gamma}])\tilde{\Psi} = -\tilde{f} + k^2\tilde{g} + \tilde{\delta} \quad (2)$$

where $[\hat{S}]$ and $[\hat{T}]$ are complex symmetric matrices, $[\hat{\gamma}]$ is a diagonal matrix, $\tilde{\Psi}$ is the solution vector, k is the wavenumber, $\tilde{f}$ and $\tilde{g}$ are known vectors, and $\tilde{\delta}$ has only one nonzero entry.

With respect to spectral modeling, we note that the matrices $[\hat{S}]$ and $[\hat{T}]$ and the vectors $\tilde{f}$ and $\tilde{g}$ are frequency independent and can be computed once and stored for any problem geometry. Only the wavenumber k, the vector $\tilde{\delta}$, and the matrix $[\hat{\gamma}]$ depend on frequency. However, since $[\hat{\gamma}]$ is a diagonal matrix with only $L \times M$ nonzero entries, where L is the number of ports in the circuit and M is the number of basis functions used to approximate the fields in each port, it requires very little work to compute $[\hat{\gamma}]$.

Fig. 1 provides the transfinite element solution of a microwave T junction at 225 MHz.

B. Spectrum Modeling

In designing microwave components, the frequency response over a given frequency range is often required. With the deterministic approach in which (1) is solved at a given frequency, (2) must be solved N times to get N points on the spectrum. For problems where the response changes very quickly, the number of points N used to generate the spectrum must be large in order to get satisfactory answers. The adaptive scheme proposed here for modeling the spectral response is best explained by referring to Fig. 2. Instead of solving (2) N times, we solve the matrix equation at a few adaptively selected optimal frequencies and then use these solutions as basis functions to generate the entire spectral response.

A preliminary illustration of the procedure is as follows: In Fig. 2(a), the three-port junction of Fig. 1 is solved by using the transfinite element method at the two limiting frequencies indicated by the squares on the horizontal axis. These two solutions are then used as basis functions to generate a crude spectral response curve throughout the region of interest. This is plotted as solid lines in Fig. 2(a). Next, we compute the error in the solution throughout the frequency range by substituting the crude solution values into the governing equations for the system and by evaluating the residual. As explained below, this may be done very efficiently. We then solve the system once again using the transfinite element algorithm *at the frequency that gave the maximum residual on the last pass*. The new square in Fig. 2(b) shows the location of this solution as well as the new spectral response curve computed by using the three transfinite element solutions as basis functions. This process is repeated for six iterations in Fig. 2 until the error in the entire spectral response curve is within acceptable limits. As is evident from Fig. 2(e), the procedure converges to the solution given in [4], but as shown below, in much less computer time.

C. The Adaptive Algorithm

In the eigensolution method, once the set of eigenfunctions $\tilde{\Psi}_i$ is known, the solution of (2) at frequency ω can be approximated by [5]

$$\tilde{\Psi}(\omega) = \sum_{i=1}^{n} a_i(\omega)\tilde{\Psi}_i \quad (3)$$

where n is the number of basis functions and a_i are unknown coefficients. Equation (3) is valid since the eigenfunctions of the Helmholtz equation are linearly independent and orthogonal. The problem with this approach, as noted earlier, is that it is very expensive to compute these eigenfunctions. Note, however, that orthogonality of basis functions is not required in (3); all that is required is that basis functions be linearly independent. The basis functions $\tilde{\Psi}_i$ in our case are taken to be the solutions of (2) at n specified frequencies ω_i. The purpose of the adaptive selec-

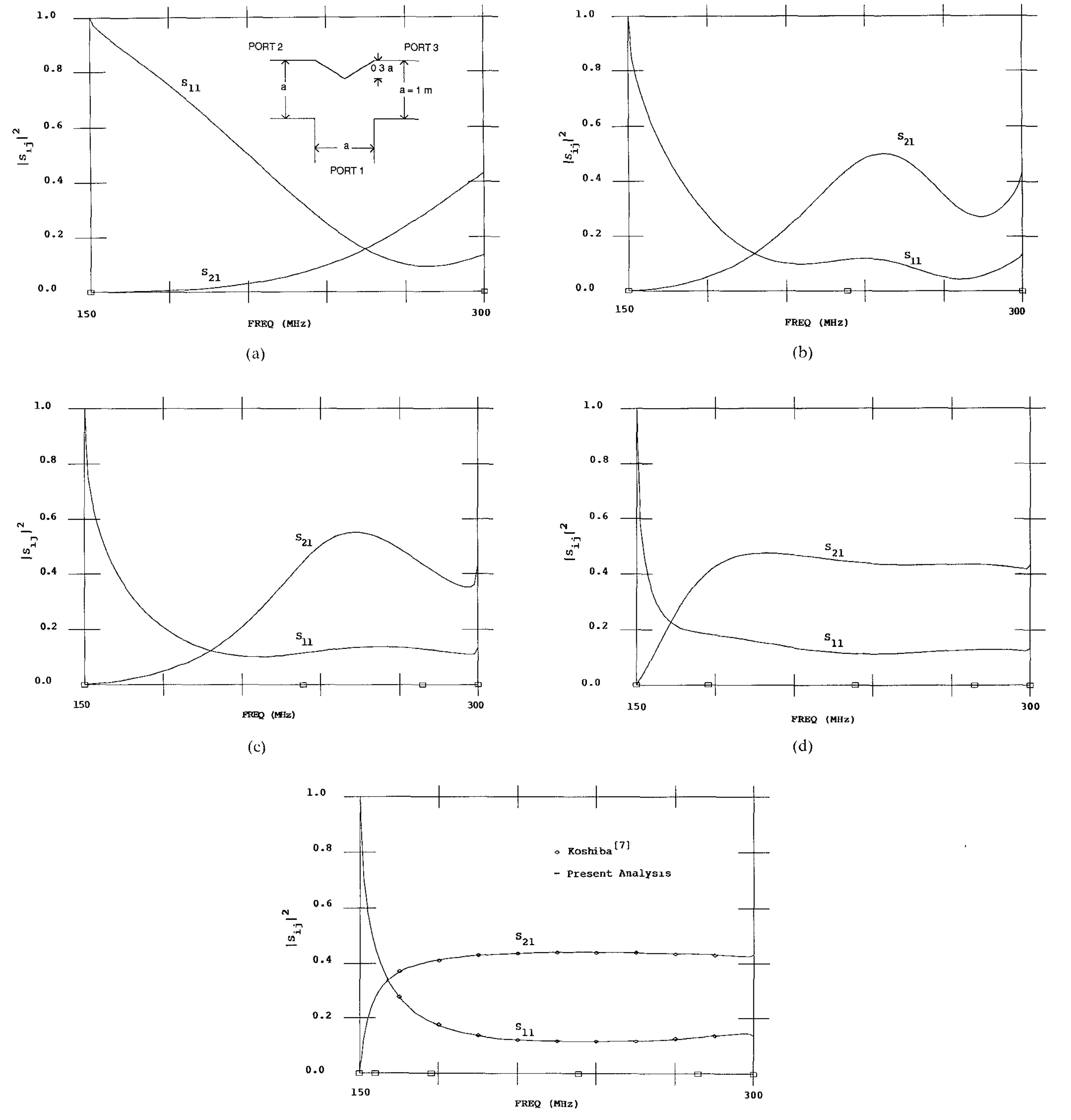

Fig. 2 Power reflection and transmission coefficients of the T junction in Fig. 1 for (a) $n=2$, (b) $n=3$, (c) $n=4$, (d) $n=5$, and (e) $n=6$.

tion scheme proposed here is to determine those frequencies ω_i that will generate the most linearly independent solutions $\tilde{\Psi}_i$ as basis functions. The rigorous proof of the validity of the algorithm is shown in the Appendix.

Substituting (3) into (2) gives

$$([\hat{S}]-k^2[\hat{T}]+[\hat{\gamma}])\cdot\sum_{i=1}^{n} a_i(\omega)\tilde{\Psi}_i = -\tilde{f}+k^2\tilde{g}+\tilde{\delta}. \quad (4)$$

Since $[\hat{S}]$, $[\hat{T}]$, $\tilde{f}$, and $\tilde{g}$ are frequency independent, we can rewrite (4) as

$$\sum_{i=1}^{n} a_i(\omega)\left(\tilde{\phi}_i - k^2\tilde{\Phi}_i + [\hat{\gamma}]\cdot\tilde{\Psi}_i\right) = -\tilde{f}+k^2\tilde{g}+\tilde{\delta} \quad (5)$$

where

$$\tilde{\phi}_i = [\hat{S}]\cdot\tilde{\Psi}_i$$
$$\tilde{\Phi}_i = [\hat{T}]\cdot\tilde{\Psi}_i. \quad (6)$$

Now applying the method of least-squares to (5) leads to the following matrix equation for the coefficients a_i:

$$([A]-k^2[B]+k^4[C]+[D])\tilde{a} = -\tilde{F}+k^2\tilde{G}-k^4\tilde{H}-\tilde{d} \tag{7}$$

where

$$A_{ij} = \underset{\sim}{\phi}_i^* \cdot \tilde{\phi}_j$$

$$B_{ij} = \underset{\sim}{\phi}_i^* \cdot \tilde{\Phi}_j + \underset{\sim}{\Phi}_i^* \cdot \tilde{\phi}_j$$

$$C_{ij} = \underset{\sim}{\Phi}_i^* \cdot \tilde{\Phi}_j$$

$$D_{ij} = \underset{\sim}{\phi}_i^*[\hat{\gamma}]\tilde{\Psi}_j + \underset{\sim}{\Psi}_i^*[\hat{\gamma}]^*\tilde{\phi}_j - k^2\left(\underset{\sim}{\phi}_i^*[\hat{\gamma}]\tilde{\Psi}_j + \underset{\sim}{\Psi}_i^*[\hat{\gamma}]^*\tilde{\phi}_j\right)$$
$$+ \underset{\sim}{\Psi}_i^*[\hat{\gamma}]^*[\hat{\gamma}]\tilde{\Psi}_j$$

$$F_i = \underset{\sim}{\phi}_i^* \cdot \tilde{f}$$

$$G_i = \underset{\sim}{\Phi}_i^* \cdot \tilde{f} + \underset{\sim}{\phi}_i^* \cdot \tilde{g}$$

$$H_i = \underset{\sim}{\Phi}_i^* \cdot \tilde{g}$$

$$d_i = \underset{\sim}{\Psi}_i^*[\hat{\gamma}]^*\tilde{f} - k^2\underset{\sim}{\Psi}_i^*[\hat{\gamma}]^*\tilde{g} - \left(\underset{\sim}{\phi}_i^* - k^2\underset{\sim}{\Phi}_i^* + \underset{\sim}{\Psi}_i^*[\hat{\gamma}]^*\right)\tilde{\delta}. \tag{8}$$

Notice that in (8), only the matrix $[D]$ and the vector $\tilde{d}$ depend on the frequency ω. The computation time for evaluating $[D]$ and $\tilde{d}$ for each frequency is negligible because $[\hat{\gamma}]$ is a diagonal matrix. Matrix equation (7) is therefore trivial to set up and it is inexpensive to solve since it is only of order n.

The residual associated with the frequency ω after solving for the coefficients $a_i(\omega)$ is evaluated as follows:

$$\tilde{r} = -\tilde{f}+k^2\tilde{g}+\tilde{\delta} - \sum_{i=1}^{n} a_i(\omega)\left(\tilde{\phi}_i - k^2\tilde{\Phi}_i + [\hat{\gamma}]\tilde{\Psi}_i\right). \tag{9}$$

The norm of the residual is given by

$$|r(\omega)| = \{\underset{\sim}{r}^* \cdot \tilde{r}\}^{0.5}. \tag{10}$$

The entire adaptive solution process is presented in the flowchart in Fig. 3. First, input the desired frequency limits ω_s, ω_e and N, the number of equal divisions in the frequency range to be used. Next solve (2) at the lowest frequency to generate the basis function $\tilde{\Psi}_1$, and the auxiliary functions $\tilde{\phi}_1$, $\tilde{\Phi}_1$. From the basis functions and the auxiliary functions, compute the components of matrices $[A]$, $[B]$, $[C]$ and of vectors $\tilde{F}$, $\tilde{G}$, $\tilde{H}$. Then generate N approximate solutions at the other frequencies and compute the residual at each frequency to indicate the corresponding error. At the maximum residual, solve (2) again and add the new solution to the basis set for generating the spectrum. Repeat the process until the maximum residual is smaller than the prescribed error tolerance η.

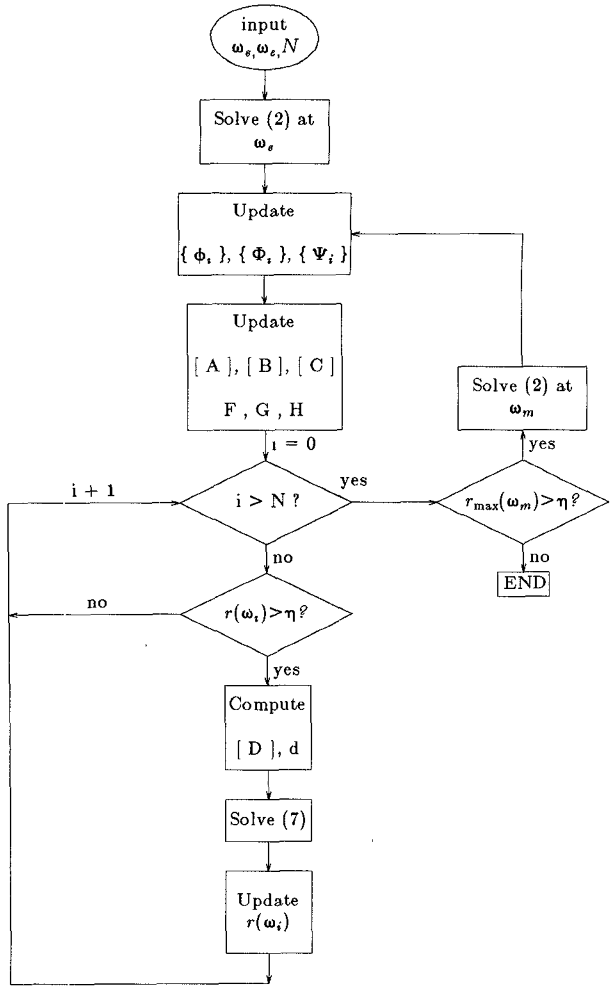

Fig. 3. Flow chart of the adaptive spectral modeling procedure.

III. Numerical Results

Figs. 3–5 provide examples of spectral response obtained by using the new adaptive spectrum modeling procedure. The number of modes used in each port to generate the results is 3 for all of the problems shown. Solutions of (2) are obtained in the computer program by using the preconditioned biconjugate gradient method and have relative residual L_2 norms smaller than 1.0e-5. The error tolerance η employed is 1.0e-4. In general, the algorithm will pick up the highest frequency as the second frequency point to solve for the basis function. So the figures shown here are plotted from $n = 2$.

Fig. 2 shows the convergence of the procedure for modeling the T junction in Fig. 1 throughout the frequency range of dominant-mode propagation. The number of frequency points N used was 50, and the frequencies that are solved for and used as the basis functions in the procedure are indicated by squares on the axis. As shown in Fig. 2, the procedure converges when $n = 6$. Comparing the final adaptively produced spectral response with the boundary element deterministic solution of [7] shows excellent agreement.

The analysis presented here is not limited to dominant-mode propagation. Fig. 3 shows the spectral response

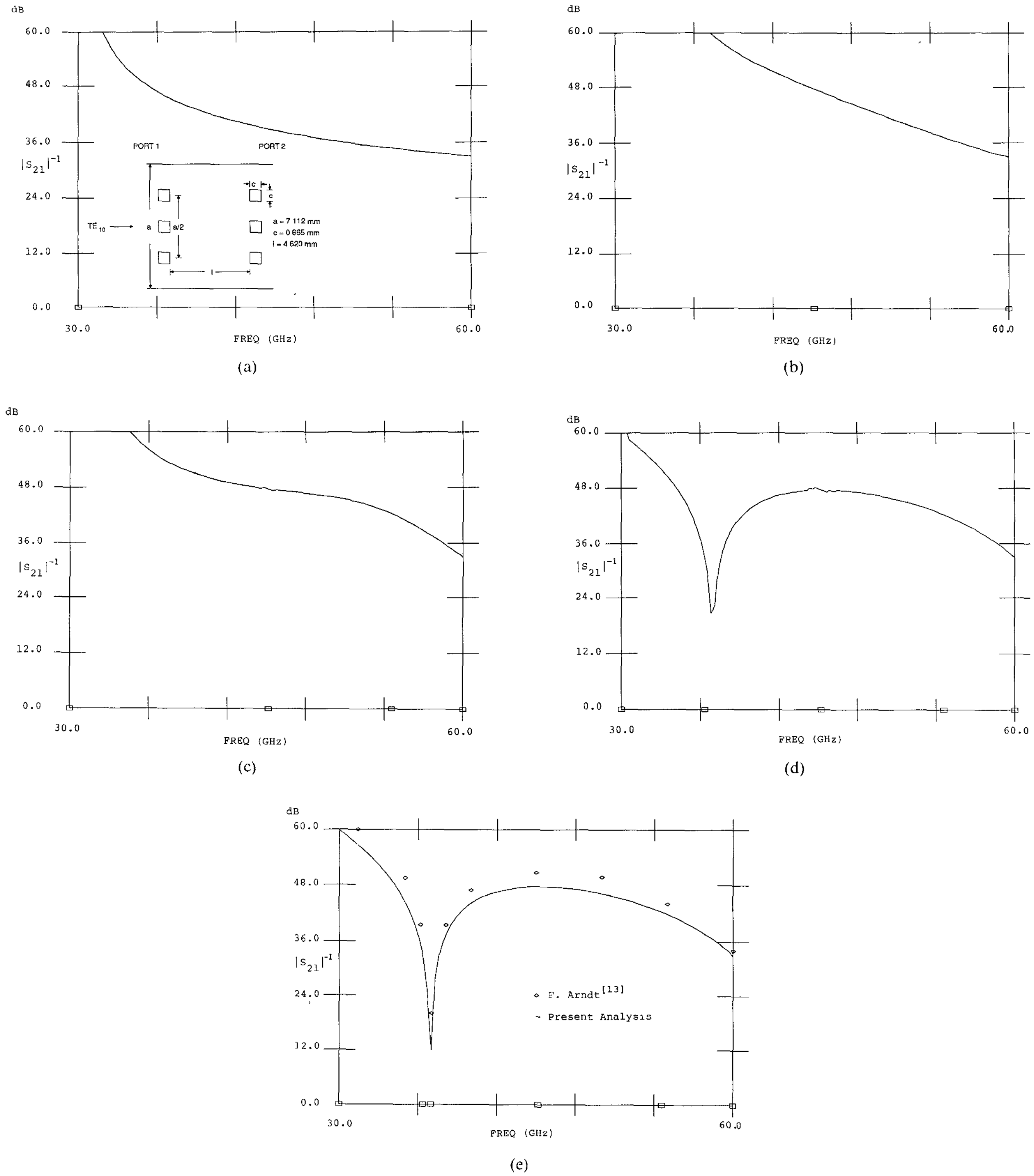

Fig. 4 Power transmission coefficient of a metal insert filter for (a) $n=2$, (b) $n=3$, (c) $n=4$, (d) $n=5$, and (e) $n=6$.

computed by the adaptive spectral modeling procedure for the metal insert filter shown in Fig. 4(a) from 30 GHz to 60 GHz. The number of frequency points used was 100 for the plots in Fig. 4(a)–(e), and the procedure terminates in six iterations. A comparison of the final spectral response with that obtained by the field expansion calculation in [13] is given in Fig. 4(e). Fig. 5 presents the adaptive solution of a waveguide dielectric filter modeled by Koshiba and Suzuki [7]. Here we used $N=100$ spectral points; the total number of solved frequency points was only 5.

Fig. 6 shows a comparison of the computer time required by the new procedure and that of the deterministic approach. The time for the deterministic approach is based

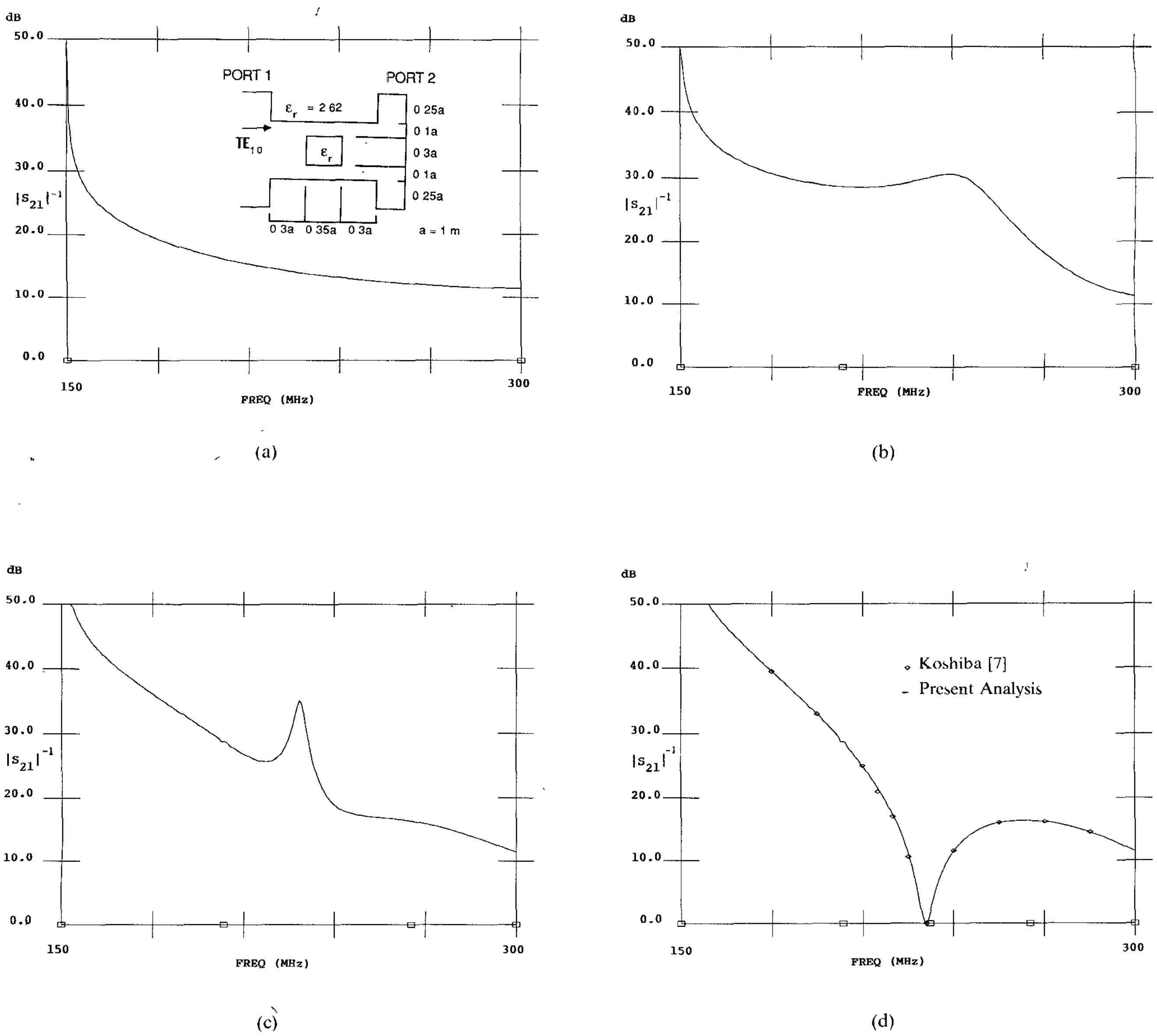

Fig 5. Power transmission coefficient of a waveguide dielectric filter for (a) $n = 2$, (b) $n = 3$, (c) $n = 4$, and (d) $n = 5$.

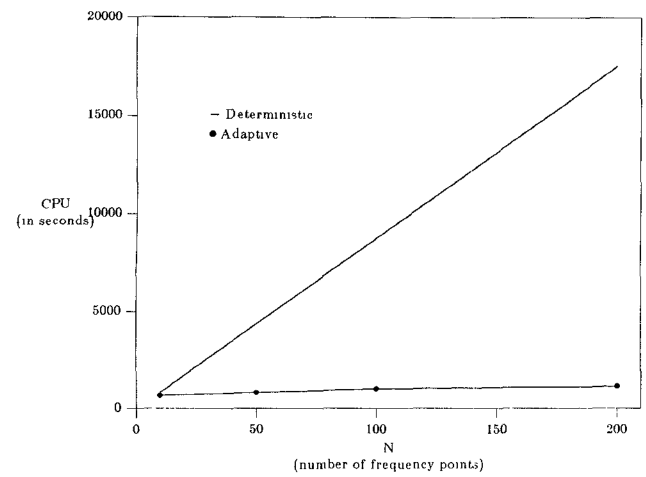

Fig. 6. Comparison of the computer time required to solve a 500×500 system using the adaptive modeling procedure with the deterministic approach.

on the total time needed to solve (2). The times reported are for the T junction problem using a DEC VAX 11/780 computer; the matrix size was 500 by 500.

IV. Conclusions

A very efficient procedure for determining the spectral response of microwave circuits has been developed. The procedure may be applied to waveguide junctions involving either E-plane or H-plane discontinuities or to microwave planar circuits. The spectral response evaluation procedure employs the transfinite element method to solve for the field at a few adaptively selected frequencies and then constructs the solution at any frequency by using the computed solutions as basis functions. In typical problems, only five or six transfinite element solutions are required to converge to the full spectral response evaluated at 100 points throughout the frequency range of interest.

In the past, there were two basic alternatives to computing the spectral response of microwave circuits: one could employ the eigensolution approach, which required expensive eigenvalue problems to be solved but gave solutions at in-between frequencies very economically, or one could employ the existing deterministic approach, which provided solutions at a specified frequency relatively efficiently but had to be reapplied at every frequency of interest. The new spectral response modeling procedure combines the advantages of both approaches and gives the full spectral response in orders of magnitude less computing time than the alternatives.

Appendix
The Linear Independence of Adaptively Selected Basis Functions

Consider the linear system

$$A(\omega)X(\omega)=Y(\omega) \quad \text{(A1)}$$

where $A(\omega)$ is a linear operator, $Y(\omega)$ is the forcing function, and $X(\omega)$ is the response, all at frequency ω. We want to be able to evaluate the response at N frequency points ω_n,

$$A(\omega_n)X(\omega_n)=Y(\omega_n), \qquad n=1\cdots N \quad \text{(A2)}$$

but have only computed the response at M frequencies ω_m,

$$A(\omega_m)\phi_m=Y(\omega_m) \quad \text{(A3)}$$

where $M\leqslant N$, $\{\omega_m\}\subseteq\{\omega_n\}$, and $\phi_m=X(\omega_m)$. We wish to use the solution ϕ_m to evaluate the response at the frequencies ω_n,

$$X(\omega_n)=\sum_{m=1}^{M}a_m\phi_m \quad \text{(A4)}$$

with a unique set of coefficients a_m.

Corresponding to equation (A4) there will be residual

$$r^M(\omega_n)=A(\omega_n)X_a(\omega_n)-Y(\omega_n) = \sum_{m=1}^{M}a_mA(\omega_n)\phi_m-Y(\omega_n). \quad \text{(A5)}$$

In the method of weighted residuals, we multiply (A5) by $\phi_k^HA^H(\omega_n)$, where ^{H} denotes the Hermitian and require that the residual $r^M(\omega_n)$ be orthogonal to the $\{A(\omega_n)\phi_m\}$ so that the left-hand side vanishes. This yields a matrix equation for the coefficients a_m:

$$B(\omega_n)\tilde{a}=\tilde{y} \quad \text{(A6)}$$

where

$$B_{km}=\phi_k^HA^H(\omega_n)A(\omega_n)\phi_m$$
$$y_k=\phi_k^HA^H(\omega_n)Y(\omega_n). \quad \text{(A7)}$$

The validity of equation (A4) is established by the following theorem.

Theorem: Provided that the frequencies ω_m are selected at points $\omega_m=\omega_n$ such that $|r^M(\omega_n)|\neq0$, the functions ϕ_m, $m=1,\cdots,M+1$, will be linearly independent.

Proof: The proof is by induction:

1) The first function ϕ_1 is linearly independent except in the trivial case.

2) Assume that after M steps, M independent basis functions $\{\phi_m\}$ exist so that (A4) is valid. Then because of the independence of the basis functions, the matrix B is positive definite and nonsingular and the coefficients a_m are uniquely determined.

3) We now show that under the conditions of the theorem, the basis function ϕ_{M+1} is independent of the set of functions $\{\phi_m\}$. There are two possibilities: either the norm of the residual $|r^M(\omega_n)|$ in (A5) is zero or it is nonzero. If it is zero, then the response ϕ_n is linearly dependent on the $\{\phi_m\}$ and cannot be used as the next basis function.

Now suppose that $|r^M(\omega_n)|\neq0$ so that we determine ϕ_{M+1} by solving (A1) at the frequency ω_n:

$$A(\omega_n)\phi_{M+1}=Y(\omega_n). \quad \text{(A8)}$$

By definition, the functions ϕ_m $(m=1,\cdots,M+1)$ are linearly independent provided that

$$\sum_{m=1}^{M+1}c_m\phi_m=0 \quad \text{implies that} \quad c_m=0. \quad \text{(A9)}$$

Multiplying (A9) by $A(\omega_n)$ and using (A8) and (A5) yields

$$\sum_{m=1}^{M+1}c_mA(\omega_n)\phi_m = \sum_{m=1}^{M}c_mA(\omega_n)\phi_m+c_{M+1}Y(\omega_n) = \sum_{m=1}^{M}c_mA(\omega_n)\phi_m + c_{M+1}\sum_{m=1}^{M}a_mA(\omega_n)\phi_m-c_{M+1}r^M(\omega_n)=0. \quad \text{(A10)}$$

However, $r^M(\omega_n)$ is orthogonal to $\{A(\omega_n)\phi_m\}$ and is nonzero; thus we conclude that $c_m=0$ $(m=1,\cdots,M+1)$ and the proof is complete.

One logical way to select the frequencies ω_m is to always take the one which maximizes $|r(\omega_n)|$. This will ensure that the conditions of the theorem are met. It will also help to select the most independent basis functions and to resolve the zero–nonzero tolerance η in the computer.

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