Application of the FD-TD Method to the Analysis of Circuits Described by the Two-Dimensional Vector Wave Equation

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***Abstract*—In the paper a class of microwave circuits described by a two-dimensional vector wave equation is defined. It is proposed to refer to them as vector two-dimensional or 2-DV circuits to distinguish them from circuits described by a two-dimensional scalar wave equation (typically referred to as 2-D circuits). It is shown that the 2-DV class contains some types of: planar circuits filled with anizotropic medium, two-dimensional waveguide discontinuities and circular waveguide discontinuities. Calculation of dispersion characteristics of inhomogeneously filled hollow waveguides is an eigenvalue problem belonging to the 2-DV class. Application of the finite-difference time-domain (FD-TD) method to the analysis of 2-DV circuits is described. The efficiency of this method is shown by means of several examples of various kinds of circuits**

I. Introduction

RECENTLY, full-wave three-dimensional time-domain methods found big interest [1]–[3]. However, a variety of practically used waveguide discontinuities can be described by a two-dimensional wave equation. For these circuits obviously, two-dimensional calculations take much less time than three-dimensional. Assuming that the solution in one of the three directions is exact, the two-dimensional analysis using particular (and always limited) computer resources produces more accurate results. Usually in 2-D methods modeling of matched ports, as well as curved boundaries is more straightforward.

A class of circuits described by the scalar wave equation has been thoroughly investigated [4]–[11]. In this paper we further distinguish a class of circuits described by the two-dimensional vector (2-DV) wave equation. This class contains some types of: planar circuits filled with anizotropic medium, two-dimensional waveguide discontinuities and circular waveguide discontinuities. Calculation of dispersion characteristics of the inhomogeneously filled hollow waveguides is an eigenvalue problem belonging to the 2-DV class. Many of these circuits have been analyzed before (see e.g. [12], [13]) but the fact that they belong to the same category has not been stressed. The formal distinguishing of the class of 2-DV circuits should help in extension of applications of the methods proved useful in the analysis of some circuits belonging to the same class.

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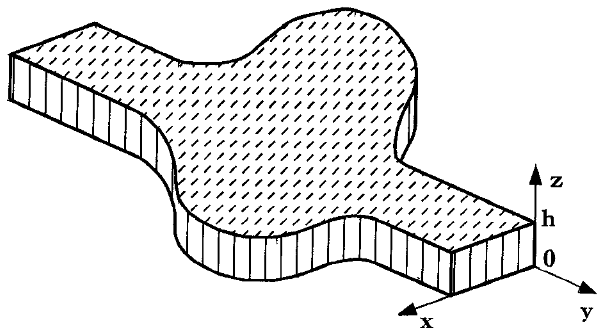

Fig. 1. A planar circuit.

In the paper we formulate basic equations describing 2-DV problems and define categories of circuits to which these equations can be applied. We also show how an algorithm written for solving scalar 2-D circuits can be efficiently used for solution of 2-DV circuits. Results of several exemplary calculations are also presented.

II. Two-Dimensional Vector Circuits

In a previous paper [5] a general theory of two-dimensional scalar circuits was presented. The two-dimensional FD-TD implementation and its application to different classes of circuits were also described in [5]–[8]. We will refer to notions and definitions used in [5].

Consider a planar circuit in the sense of [4], [5]. The circuit is bounded by two planes: $z = 0$ and $z = h$ (Fig. 1). The medium filling the circuit is homogeneous along the z-axis and is described by diagonal tensors:

$$\mu(x,y) = \begin{bmatrix} \mu_x & 0 & 0 \\ 0 & \mu_y & 0 \\ 0 & 0 & \mu_z \end{bmatrix};$$
$$\epsilon(x,y) = \begin{bmatrix} \epsilon_x & 0 & 0 \\ 0 & \epsilon_y & 0 \\ 0 & 0 & \epsilon_z \end{bmatrix}. \quad (1)$$

Evolving upon general properties of the Hertzian potentials it was proposed in [5] to develop the fields in the so defined planar circuit into a series of height modes E_n, H_n. For a nth mode, all field components are proportional either to $\cos(\beta_{zn} z)$ or $\sin(\beta_{zn})$, where $\beta_{zn} = n\pi/h$.

A scalar 2-D algorithm [5] can be applied to a planar circuit with a single E_n or H_n mode. However, if the medium filling the circuit is inhomogeneous in the (x, y) plane, then a single E_n or H_n mode (other then E_O) cannot propagate [14]. We must consider a combined EH_n or HE_n mode with all six field components. From now on we will refer to a planar circuit supporting a combined EH_n or HE_n mode as a vector two-dimensional circuit (2-DV).

We consider propagation of a single height mode EH_n (dual analysis leads to the solution for the HE_n mode). In this case the dependence of all six field components on the z-coordinate is known:

$$\underline{E}_x = E_x \sin(\beta_{zn} z), \quad \underline{E}_y = E_y \sin(\beta_{zn} z),$$
$$\underline{E}_z = E_z \cos(\beta_{zn} z), \tag{2}$$

and

$$\underline{H}_x = H_x \cos(\beta_{zn} z), \quad \underline{H}_y = H_y \cos(\beta_{zn} z),$$
$$\underline{H}_z = H_z \sin(\beta_{zn} z), \tag{3}$$

For coherence with the previous paper [5], let us define electric and magnetic currents and potentials:

$$\boldsymbol{J} = -\boldsymbol{i}_z \times \boldsymbol{H}_t|_{z=h} = -\boldsymbol{i}_z \times (\boldsymbol{i}_x H_x + \boldsymbol{i}_y H_y);$$
$$V = -hE_z \tag{4}$$
$$\boldsymbol{J}^h = \boldsymbol{i}_z \times \boldsymbol{E}_t|_{z=h/(2n)} = -\boldsymbol{i}_z \times (\boldsymbol{i}_x E_x + \boldsymbol{i}_y E_y);$$
$$V^h = -hH_z, \tag{5}$$

where $\boldsymbol{i}_x, \boldsymbol{i}_y, \boldsymbol{i}_z$ are unit vectors along axis x, y, z respectively,
z is the distinguished axis of the circuit,
$\boldsymbol{E}_t, \boldsymbol{H}_t$ are the field components tangential to the plane z = const,
$E_x, E_y, E_z, H_x, H_y, H_z$ are amplitudes as defined by (2) and (3).

Considering a vector two-dimensional assumptions the Maxwell's equations (6), (7):

$$\nabla \times \underline{\boldsymbol{H}} = \epsilon \frac{\partial \underline{\boldsymbol{E}}}{\partial t} \tag{6}$$
$$\nabla \times \underline{\boldsymbol{E}} = -\mu \frac{\partial \underline{\boldsymbol{H}}}{\partial t} \tag{7}$$

are equivalent to the following four equations using the defined notions (4) and (5):

$$\nabla_t \cdot \boldsymbol{J} = -C \frac{\partial V}{\partial t} \tag{8}$$
$$\nabla_t V - \beta'(\boldsymbol{i}_z \times \boldsymbol{J}^h) = -\boldsymbol{L} \frac{\partial \boldsymbol{J}}{\partial t} \tag{9}$$
$$\nabla_t \cdot \boldsymbol{J}^h = -C^h \frac{\partial V}{\partial t}; \tag{10}$$
$$\nabla_t V^h - \beta'(\boldsymbol{i}_z \times \boldsymbol{J}) = -\boldsymbol{L}^h \frac{\partial \boldsymbol{J}^h}{\partial t} \tag{11}$$

where:

$$C = \frac{\epsilon_z}{h}; \quad \boldsymbol{L} = h \begin{bmatrix} \mu_x & 0 \\ 0 & \mu_y \end{bmatrix}; \tag{12}$$
$$C^h = \frac{\mu_z}{h}; \quad \boldsymbol{L}^h = h \begin{bmatrix} \epsilon_x & 0 \\ 0 & \epsilon_y \end{bmatrix} \tag{13}$$
$$\beta' = n\pi. \tag{14}$$

Solving the set of equations (8)–(11) with proper boundary conditions is equivalent to solving any of the practical problems of microwave engineering belonging to the following classes:

(a) characterization of E-plane discontinuities in rectangular waveguides filled with medium inhomogeneous in the E-plane;

(b) characterization of planar circuits filled with anisotropic medium (described by diagonal tensors of ϵ and μ);

(c) characterization of inhomogeneously filled circular waveguide discontinuities, maintaining axial symmetry of boundary conditions and excited by not axially symmetrical modes;

(d) finding propagation constants and field distributions of the modes in hollow waveguides of arbitrary shapes and of inhomogeneously filled cross sections.

Although, application of (8)–(11) to the classes **a** and **b** is straightforward, the classes **c** and **d** need further comments.

III. Axially Symmetrical Circuits

As it was previously pointed out [6] the circuits maintaining axial symmetry of boundary conditions and excited by fields of axial symmetry can be treated as inhomogeneously filled scalar 2-D circuits. Let us now abandon the assumption about the axial symmetry of excitation. We assume that an axially symmetrical circuit is filled with medium characterized by $\mu'(\rho, y)$ and $\epsilon'(\rho, y)$—Fig. 2(a). The circuit is excited by one mode in which all the fields depend on the angle ϕ proportionally to $\sin(n\phi)$ or $\cos(n\phi)$, where n is the mode number. We have introduced here the cylindrical coordinates in the unconventional order (ρ, y, ϕ) to get results compatible with the planar circuit model. Defining:

$$\boldsymbol{J} = -\boldsymbol{i}_\phi \times \boldsymbol{H}_t|_{\phi=0}; \qquad V = -\rho E_\phi \tag{15}$$
$$\boldsymbol{J}^h = \boldsymbol{i}_\phi \times \boldsymbol{E}_t|_{\phi=\pi/(2n)}; \qquad V^h = -\rho H_\phi \tag{16}$$

where H_ϕ and E_ϕ are normal to the plane ϕ = const) we obtain equations identical with (8)–(11) but with the following meaning of the parameters and operators:
∇_t is a two-dimensional operator in (ρ, y) coordinates,
the role of $\boldsymbol{i}_z$ is played by $\boldsymbol{i}_\phi$,

$$\beta' = n; \quad C = \epsilon'/\rho; \quad C^h = \mu'/\rho \tag{17}$$
$$\boldsymbol{L} = \begin{bmatrix} \mu'\rho & 0 \\ 0 & \mu'\rho \end{bmatrix}; \quad \boldsymbol{L}^h = \begin{bmatrix} \epsilon'\rho & 0 \\ 0 & \epsilon'\rho \end{bmatrix} \tag{18}$$

It means that we can analyze a circuit of axial symmetry as a planar circuit (see Fig. 2) filled with inhomogeneous anizotropic medium described by (17) and (18). This analogy opens a vast area of applications for which the FD-TD method has not been used up to now. We can analyze various discontinuities of cylindrical waveguides. Our preliminary investigation shows also that the presented approach can lead to powerful tools of numerical analysis of waveguide antennas of axial symmetry [15], including radiating horns with dielectric inserts.

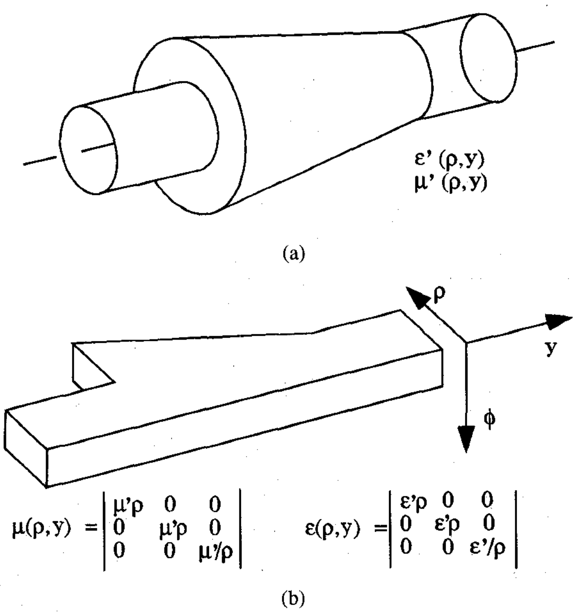

Fig. 2. A circuit of axial symmetry and its planar equivalence.

IV. Eigenvalue Problems for Hollow Waveguides

Let us imagine a hollow waveguide (Fig. 3(a)) with the electromagnetic wave propagating in it with a propagation constant β_z. Let us cut a part of the waveguide of length $l = \lambda/2$ and bound it with two electric walls at both ends. After such manipulation we obtain a resonator satisfying the vector two-dimensional assumption. Electromagnetic field in

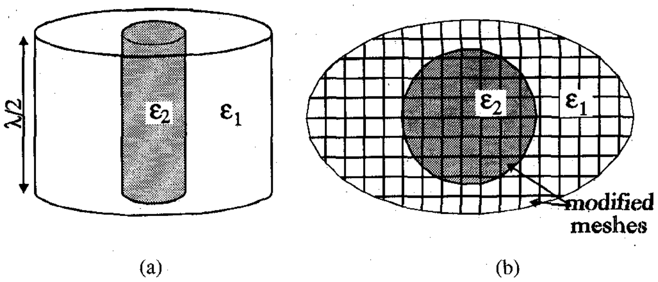

Fig. 3. A section of a hollow waveguide and its two-dimensional representation.

this resonator can be described by the equations of the EH_1 mode. The resonant frequencies of the resonator are equal to the eigenfrequencies of the waveguide. Computing the resonant frequencies for different values of β_z we can obtain the dispersion characteristics of all modes propagating in the waveguide. However, it should be noted that the method does not distinguish complex modes having the same β_z.

V. Numerical Solution

For solution of (8)–(11) we apply the finite-difference time-domain algorithm. Using the central finite-difference scheme equations (8)–(11) are discretized below (see Fig. 4): where $a = \Delta x = \Delta y$ is a step of the space discretization and after (12) and (13):

$$L_x = h\mu_x, \qquad L_y = h\mu_y,$$
$$L_x^h = h\epsilon_x, \qquad L_y^h = h\epsilon_y.$$

$$J_x(x,y,t+\Delta t/2) = J_x(x,y,t-\Delta t/2) - \frac{\Delta t}{L_x}\left(\frac{V(x+a/2,y,t)-V(x-a/2,y,t)}{a} - \beta' J_y^h(x,y,t)\right); \tag{19}$$

$$J_y(x+a/2,y+a/2,t+\Delta t/2) = J_y(x+a/2,y+a/2,t-\Delta t/2) - \frac{\Delta t}{L_y}\left(\frac{V(x+a/2,y+a,t)-V(x+a/2,y,t)}{a} - \beta' J_x^h(x+a/2,y+a/2,t)\right); \tag{20}$$

$$V^h(x,y+a/2,t+\Delta t/2) = V^h(x,y+a/2,t-\Delta t/2) - \frac{\Delta t}{C^h}\left(\frac{J_x^h(x+a/2,y+a/2,t)-J_x^h(x-a/2,y+a/2,t)}{a} + \frac{J_y^h(x,y+a,t)-J_y^h(x,y,t)}{a}\right); \tag{21}$$

$$V(x+a/2,y,t+\Delta t) = V(x+a/2,y,t) - \frac{\Delta t}{C}\left(\frac{J_x(x+a,y,t+\Delta t/2)-J_x(x,y,t+\Delta t/2)}{a} + \frac{J_y(x+a/2,y+a/2,t+\Delta t/2)-J_y(x+a/2,y-a/2,t+\Delta t/2)}{a}\right); \tag{22}$$

$$J_x^h(x+a/2,y+a/2,t+\Delta t) = J_x^h(x+a/2,y+a/2,t) - \frac{\Delta t}{L_x^h}\left(\frac{V^h(x+a,y+a/2,t+\Delta t/2)-V^h(x,y+a/2,t+\Delta t/2)}{a} - \beta' J_y(x+a/2,y+a/2,t+\Delta t/2)\right); \tag{23}$$

$$J_y^h(x,y,t+\Delta t) = J_y^h(x,y,t) - \frac{\Delta t}{L_y^h}\left(\frac{V^h(x,y+a/2,t+\Delta t/2)-V^h(x,y-a/2,t+\Delta t/2)}{a} - \beta' J_x(x,y,t+\Delta t/2)\right); \tag{24}$$

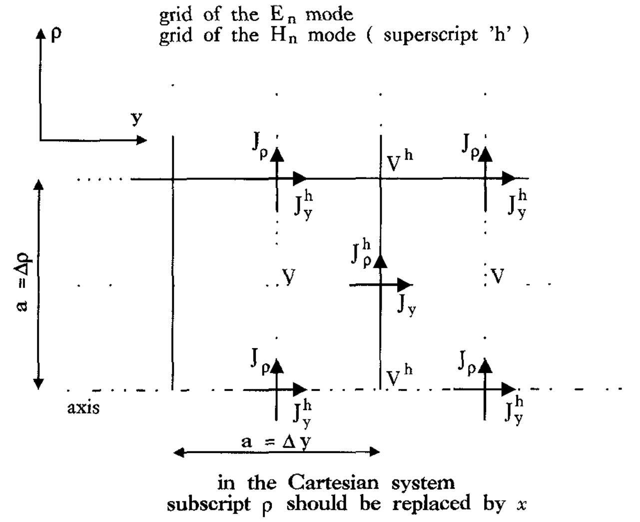

Fig. 4. A fragment of the grid used in the 2-DV FD-TD calculations.

Equations (19)–(24), shown on the previous page, are the kernel of the 2-DV FD-TD algorithm. The accuracy and efficiency of this algorithm are further improved by introducing so-called modified meshes to modeling arbitrarily shaped boundaries [6], [7] and media interfaces. The theory of modified meshes will be presented in a separate paper. In the existing software for the analysis of arbitrarily shaped scalar 2-D circuits by the FD-TD method a great programming effort has been put to calculate the coefficients C and L for the modified meshes. Thus, it was important to find a correspondence between the scalar 2-D and the 2-DV software for optimal application of existing numerical tools.

Correspondence of the Scalar 2-D and 2-DV Software

The form of equations (8)–(11) has the advantage of exhibiting direct correspondence to the scalar wave equations [5]:

$$\nabla_t \cdot \boldsymbol{J} = -C \cdot \frac{\partial V}{\partial t}; \tag{25}$$

$$\nabla_t V = -\boldsymbol{L}\frac{\partial \boldsymbol{J}}{\partial t}; \tag{26}$$

Equations (9) and (11) are modified with respect to (26) by a term describing the coupling between them. The pair of equations (8)–(9) is identical to the pair (10)–(11) transformed according to duality relations. This observation is important from the practical point of view. It follows from the form of (8)–(11) that the same software tools (after minor modifications) can be used to calculate the coefficients C, L, and C^h, L^h of 2-DV circuits. The modification consists of introducing two grids of meshes shifted by half of a mesh in x and y directions, as presented in Fig. 5. For calculation of C^h and L^h dual medium parameters should be taken (ϵ instead μ and vice versa) and dual boundary condition (open exchanged with short). The two grids are coupled according to (9) and (11).

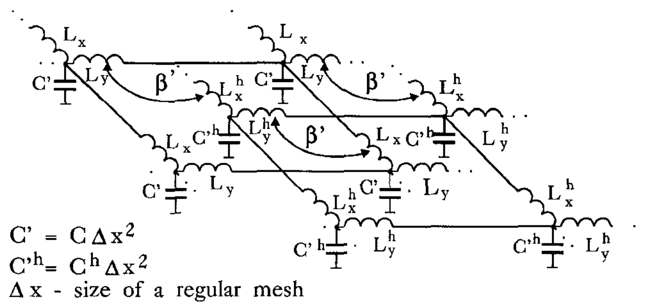

Fig. 5. A model of a 2-DV circuit composed of two coupled 2-D models.

From the physical viewpoint, in different classes of circuits the coupling factor β' depends on:
in the classes **a** and **b**—on the cutoff frequency of the mode propagating in the planar structure,
in the class **c**—on the mode number n characterizing the dependence of fields on the angle coordinate ϕ,
in the class **d**—on the propagation constant β_z along the waveguide.

Circuits Axially Symmetrical

With definitions (17) and (18) we use (19)–(24) to simulate wave propagation in circuits of the class **c**. We can rewrite (24) in the cylindrical coordinates shown in (27) at the bottom of the page.

Boundary conditions at the axis are: $J_\rho(\rho = 0) = 0, J^h_\rho(\rho = 0) = 0$. The other values at the axis can differ from 0, in particular $J_y(\rho = 0) \neq 0$. Since at the axis $L_y(\rho = 0) = 0$ and $V^h(\rho = 0) = 0$ (see (16), (18) and Fig. 4) we can not apply directly (27). Instead, at the axis we calculate $J^h_y(\rho = 0)$ from (28), which is shown at the bottom of the page.

VI. Examples

The FD-TD method in the form described in [5] and [7] and modified according to (8)–(11) was used by the authors to calculate several examples belonging to all the classes marked **a..d**. In the cases where comparative data were available very good agreement was reached. We will present here three examples.

Example 1

As an example of a problem of the class **a** we present a dielectric post situated in the E-plane of a rectangular waveguide (Fig. 6). The results of calculations by the FD-TD method in the form presented above are compared in Fig. 7 with the results obtained by the finite element method presented in [16].

$$J^h_y(\rho, y, t + \Delta t) = J^h_y(\rho, y, t) - \frac{\Delta t}{L^h_y}\left(\frac{V^h(\rho, y + a/2, t + \Delta t/2) - V^h(\rho, y - a/2, t + \Delta t/2)}{a} - \beta' J\rho(\rho, y, t + \Delta t/2)\right). \tag{27}$$

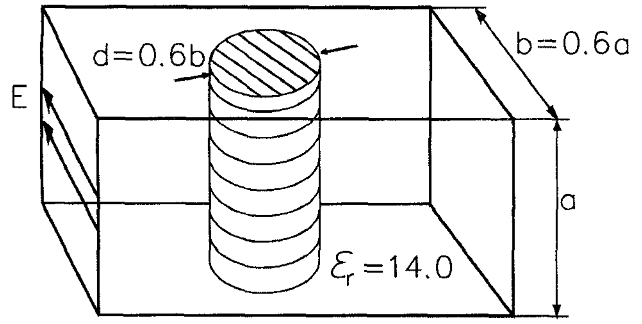

Fig. 6. A dielectric post in a rectangular waveguide analyzed in Example 1.

TABLE I
EIGENFREQUENCIES OF THE RESONATOR OF FIG. 8 FD-TD DENOTES CALCULATION BY THE PRESENT METHOD, MM RESULTS AFTER [17] OBTAINED BY THE MODE MATCHING METHOD (THE TABLE CONTAINS FREQUENCIES f' NORMALIZED IN SUCH A WAY THAT: $f' = b\sqrt{\epsilon}2\pi f/c$)

Mode	p=0 3		p=0.5		p=0.7	
	FD-TD	MM	FD-TD	MM	FD-TD	MM
HE 111	2.631	2.636	2.843	2.846	3.119	3.121
EH 110	3.373	3.381	4.235	4 241	4.343	4.347
EH 111	3.818	3 825	4 945	4.954	5.141	5.149
HE 121	4.504	4 515	5.231	5 242	5 583	5.591

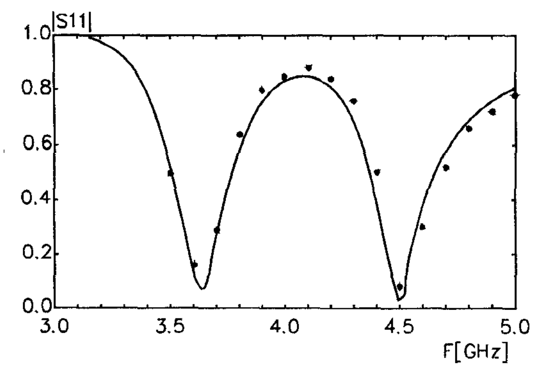

Fig. 7. Results of analysis of the circuit from Fig. 6 for $d = 0.4b$ (a) Continuous line—the present method, (b) Points—finite element method [16].

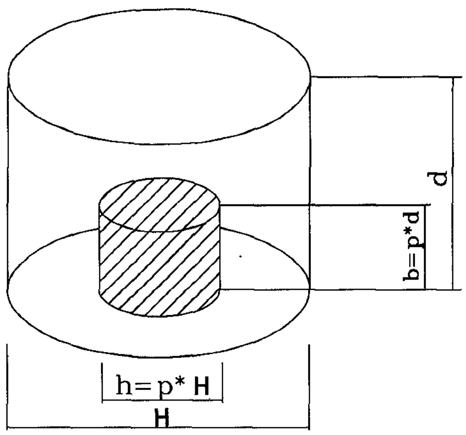

Fig. 8. A shape of the axially symmetrical resonator considered in Example 2.

As it was shown in [7], to perform the FD-TD calculations efficiently we have to simulate the pulse excitation of the circuit. Since input impedance of the rectangular waveguide is frequency dependent, a model of a port should ensure good matching in a broad frequency band. Assuming that at the port plane only the fundamental mode exists we simulate a matched load by a one-dimensional model of a lossy waveguide [8].

Example 2

Let us consider an inhomogeneously filled resonator of axial symmetry (Fig. 8). In the resonator we assume $b = h$ and $d = H$, and we modify the filling factor $p = b/d$. The results of calculations of normalized resonant frequencies for several modes, assuming different values of p are compared in Table I with the results obtained by Krupka [17] by the mode matching method. In the FD-TD calculations a grid of 40×40 meshes was used. This example is treated as an accuracy test before designing a resonator of a really complicated shape and filling which is planned to be used as a part of a cyclotron.

Example 3

The dispersion characteristics of the modes in various image waveguides have been calculated. This is a problem of the class **d** mentioned above.

The results of our calculations are compared with those published by Strube and Arndt [18] and very good agreement is obtained (Figs. 9 and 10). The advantage of our approach is its great flexibility in the calculation of complicated profiles. Its disadvantage (in the present state), as it was mentioned before, is inability to analyze complex modes which in some cases appear in inhomogeneously filled guiding structures.

$$J_y^h(0,y,t+\Delta t) = J_y^h(0,y,t) - \frac{\Delta t}{\epsilon'}\left(\frac{H_\phi(0,y+a/2,t+\Delta t/2) - H_\phi(0,y-a/2,t+\Delta t/2)}{a}\right); \tag{28}$$

where

$$\begin{aligned} H_\phi(0,y+a/2,t+\Delta t/2) &= H_\phi(0,y+a/2,t-\Delta t/2) \\ &\quad - \frac{\Delta t}{\mu'}\left(\frac{2J_\rho^h(a/2,y+a/2,t)}{a} + \frac{J_y^h(0,y+a,t) - J_y^h(0,y,t)}{a}\right); \end{aligned} \tag{29}$$

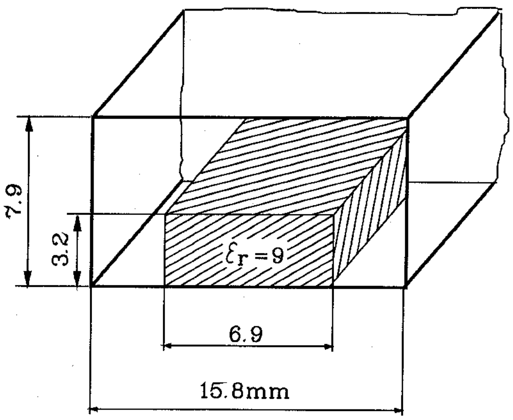

Fig. 9. A section of an inhomogeneously filled waveguide analyzed in Example 3.

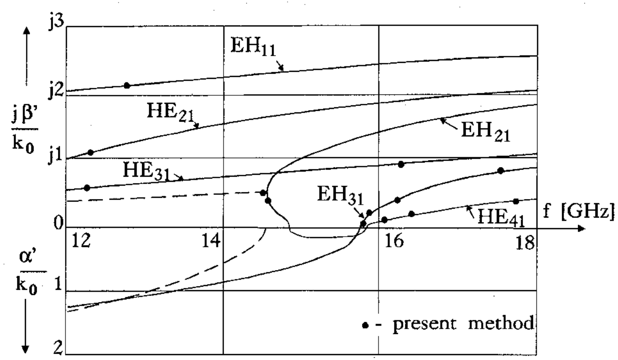

Fig. 10. Results of calculations in Example 3 with comparison to Strube–Arndt [18].

VII. Conclusions

We have presented a class of microwave circuits, which can be described by the two-dimensional vector wave equation. This class includes some planar circuits filled with inhomogeneous or anisotropic medium, as well as circular waveguide discontinuities which can be analyzed using a planar model. We can also assign to this class the problem of finding the modes in inhomogeneously filled waveguides. We have shown that all these circuits can be analyzed by the FD-TD method and in the analysis the method retains all its advantages previously shown in application to scalar 2-D circuits. When we already have a computer program for analyzing arbitrarily shaped scalar 2-D circuits, the paper shows how to modify it with relatively small programming effort to obtain a program analyzing vector 2-D circuits.

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