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<td>著者</td>
<td>Koshiba, M.; Hayata, K.; Suzuki, M.</td>
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<tr>
<td>引用</td>
<td>IEEE Transactions on Microwave Theory and Techniques, 33(10), 900-905</td>
</tr>
<tr>
<td>発行日</td>
<td>1985-10</td>
</tr>
<tr>
<td>URL</td>
<td><a href="http://hdl.handle.net/2115/6037">http://hdl.handle.net/2115/6037</a></td>
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<tr>
<td>東京大学コレクションの学術研究論文 : HUSCAP</td>
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Finite-Element Formulation in Terms of the Electric-Field Vector for Electromagnetic Waveguide Problems

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Abstract — A vector finite-element method for the analysis of anisotropic waveguides with off-diagonal elements in the permeability tensor is formulated in terms of all three components of the electric field. In this approach, spurious, nonphysical solutions do not appear anywhere above the "air-line." The application of this finite-element method to waveguides with an abrupt discontinuity in the permittivity is discussed. In particular, we discuss how to use the boundary conditions of the electric field at the interface between two media with different permittivities. To show the validity and usefulness of this formulation, examples are computed for dielectric-loaded waveguides and ferrite-loaded waveguides.

I. INTRODUCTION

THE VECTOR finite-element method is widely used either in an axial-component (E, - H.) formulation [1]-[4] or in a three-component (either the electric field E or the magnetic field H) formulation [5], [6], which enables one to compute accurately the mode spectrum of an electromagnetic waveguide with arbitrary cross section. The most serious difficulty in using the vector finite-element analysis is the appearance of spurious, nonphysical solutions [1]-[6]. Hano [7] has presented a three-component finite-element formulation with rectangular elements. In his formulation, spurious solutions, except for zero eigenvalues, do not appear, but a diagonal permittivity tensor and a diagonal permeability tensor are assumed. Recently, an improved finite-element method with triangular elements has been formulated for the analysis of anisotropic dielectric waveguides in terms of all three components of H [8]-[11]. In dielectric waveguides, the permeability is always assumed to be that of free space. Therefore, each component of H is continuous in the whole region and it is more advantageous to solve for H rather than for E [12]. In this improved H-field formulation, no spurious solutions appear anywhere above the "air-line" corresponding to $\beta/k_0 = 1$ in a $\beta/k_0$ versus $k_0$ diagram [11], where $k_0$ is the wavenumber of free space and $\beta$ is the phase constant in the z-direction. The appearance of spurious solutions is limited to the region $\beta/k_0 < 1$ and these solutions are equivalent to the TM modes of "hollow" waveguides. To show the validity and usefulness of this formulation, examples are computed for dielectric-loaded waveguides and ferrite-loaded waveguides.

II. FUNCTIONAL FORMULATION

We consider an anisotropic waveguide with a tensor permeability and a scalar permittivity. With a time dependence of the form $\exp(j\omega t)$ being implied, Maxwell's equations are

\begin{align}
\nabla \times E &= -j\omega \mu_0 [\mu_r] H \\
\nabla \times H &= j\omega \varepsilon_0 [\varepsilon_r] E
\end{align}

where $\omega$ is the angular frequency, $\mu_0$ and $\varepsilon_0$ are the permeability and permittivity of free space, respectively, $[\mu_r]$ is the relative permeability tensor, $[\cdot]$ denotes a matrix, and $[\varepsilon_r]$ is the relative permittivity which is assumed to be constant in each material.

From (1) into (2), the following wave equation is derived:

\begin{equation}
\nabla \times \left(([\mu_r]^{-1}) \nabla \times E\right) - k_0^2 \varepsilon_r E = 0
\end{equation}

Manuscript received February 1, 1985; revised May 22, 1985.

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The functional \([12,13]\) for (3) is known to be
\[
\mathcal{F} = \int_\Omega ((\nabla \times E)^* \cdot ([\mu_r]^{-1} \nabla \times E)) \, d\Omega
- k_0^2 \int_\Omega \epsilon_r E^* \cdot E \, d\Omega
\]
\[\text{(5)}\]
where \(\Omega\) represents the cross section of the waveguide and the asterisk denotes complex conjugation. In the finite-element analysis using (5), spurious solutions appear scattered all over the propagation diagram \([5]-[12], [14], [15]\). These spurious solutions belong to two distinct categories \([11]\). The first one (S1) can be characterized as follows:
\[
\nabla \times E = 0 \quad \nabla \cdot E \neq 0 \quad \text{for } k_0^2 = 0.
\]
(6)
The second group (S2) can be characterized as follows:
\[
\nabla \times E \neq 0 \quad \nabla \cdot E \neq 0 \quad \text{for } k_0^2 > 0.
\]
(7)
In order to eliminate the spurious solutions S1 and S2, we propose the following functional according to the \(H\)-field formulation \([8]-[11]\):
\[
\delta \mathcal{F} = \int_\Omega \delta \mathcal{E}^*
\cdot \left[ \nabla \times \left( [\mu_r]^{-1} \nabla \times E \right) - \nabla (\nabla \cdot E) - k_0^2 \nabla \right] \, d\Omega
\]
\[\text{(8)}\]
\[\text{where } \delta \mathcal{F} = 0 \text{ yields}\]
\[
\nabla \times \left( [\mu_r]^{-1} \nabla \times E \right) - \nabla (\nabla \cdot E) - k_0^2 \nabla = 0,
\]
(10a)
as the Euler equation and
\[
\nabla \cdot (\nabla \cdot E) = 0 \quad \text{on perfect electric conductor}
\]
(10b)
\[
n \times ( [\mu_r]^{-1} \nabla \times E ) = 0 \quad \text{on perfect magnetic conductor}
\]
(10c)
as natural boundary conditions, since \(\delta \mathcal{E}^*\) in (9) is arbitrary. The spurious solutions S1 and S2 are not included in (8), but (8) may have other solutions than (3). This new group (S3), characterized by
\[
\nabla \times E = 0 \quad \nabla \cdot E \neq 0 \quad \text{for } k_0^2 > 0
\]
(11)
where \(k_0^2 = \omega^2 \epsilon_0 \mu_0\).

III. FINITE-ELEMENT DISCRETIZATION AND BOUNDARY CONDITIONS

Dividing the cross section \(\Omega\) of the waveguide into a number of second-order triangular elements as shown in Fig. 1 and using the finite-element method on (8), we can write the functional for the whole region \(\Omega\) in the form
\[
\delta \mathcal{F} = \sum_{e} \delta \mathcal{F}_e
\]
\[\text{(13)}\]
\[\mathcal{F}_e = [E]_e^T [A]_e [E]_e
\]
(14)
\[\text{where } [A]_e = [S]_e + [U]_e - k_0^2 [T]_e
\]
(15)
where \([E]_e\) is the electric-field vector corresponding to the nodal points within each element, the matrices \([S]_e, [T]_e,\) and \([U]_e\) for each element are related to the first, second, and third terms on the right-hand side of (8), respectively, \(T, \{ \cdot \},\) and \(\{ \cdot \}^T\) denote a transpose, a column vector, and a row vector, respectively, and the summation \(\Sigma_e\) extends over all different elements. Variation of (13) with respect to the nodal variables leads to the eigenvalue problem.

In (14), the nodal electric-field vector \([E]_e\) should be forced to satisfy the boundary conditions at the interface between two media with different permeabilities. We consider the interface \(\Gamma'\) with an abrupt discontinuity in the permittivity as shown in Fig. 1, where \(\epsilon_1\) and \(\epsilon_2\) are permeabilities of the regions 1 and 2, respectively, the unit vector \(n\) normal to \(\Gamma'\) makes an angle \(\theta\) with respect to the \(x\)-axis in the \(xy\)-plane, and the elements related to \(\Gamma'\) are grouped into two classes: the elements \((e_1)\) in region 1 and the elements \((e_2)\) in region 2.
If the functional for $e_1$ is used in its original form (14), we should modify the functional for $e_2$ in order to satisfy the boundary conditions of the electric field $E$ on $\Gamma'$. For $e_2$, the functional (14) can be rewritten as

$$
\{E\}_2 = \begin{bmatrix}
\{E_x\}_2 \\
\{E_y\}_2 \\
\{E_z\}_2
\end{bmatrix}
\quad (16a)
$$

Using (17), (16) can be transformed as follows:

$$
\bar{\{E\}}_2 = (\bar{E})_2 \{A\}_2 \{E\}_2
$$

where

$$
\{E\}_2 = \begin{bmatrix}
\{E_x\}_2 \\
\{E_y\}_2 \\
\{E_z\}_2
\end{bmatrix}
\quad (19a)
$$

where the components of the $\{E_i\}_2$ vector are the values of the electric field $E_i$ ($i = x, y, z$) at the nodal points within the element $e_2$ except $\Gamma'$, the components of the $\{E_i\}_2$ vector are the values of $E_i$ at the nodal points on $\Gamma'$ included in the element $e_2$, and the $[A_{xx}]_2$, $[A_{xy}]_2$, $[A_{y}x]_2$, and $[A_{zz}]_2$ are the submatrices of the matrix (15) for $e_2$.

The tangential components of $E$ and the normal component of $\epsilon E$ should be continuous at the interface $\Gamma'$. These boundary conditions can be written as

$$
\{E_x\}_2 = q_{xx} [A_{xx}]_2 + q_{xy} [A_{xy}]_2 + q_{xz} [A_{xz}]_2
$$

$$
\{E_y\}_2 = q_{xy} [A_{xy}]_2 + q_{yy} [A_{yy}]_2 + q_{yz} [A_{yz}]_2
$$

$$
\{E_z\}_2 = q_{xz} [A_{xz}]_2 + q_{yz} [A_{yz}]_2 + q_{zz} [A_{zz}]_2
$$

where

$$
q_{xx} = \sin^2 \theta + (\epsilon_1 / \epsilon_2) \cos^2 \theta
$$

$$
q_{xy} = [(\epsilon_1 / \epsilon_2) - 1] \sin \theta \cos \theta
$$

$$
q_{yy} = \cos^2 \theta + (\epsilon_1 / \epsilon_2) \sin^2 \theta.
$$
By using the original functional (14) for \( e_1 \) and the modified functional (19) for \( e_2 \), the boundary conditions of the electric field \( E \) at the interface with an abrupt discontinuity in the permittivity are satisfied.

### IV. NUMERICAL RESULTS

#### A. Dielectric-Loaded Waveguides

First, let us consider a rectangular waveguide half-filled with a dielectric of permittivity \( \epsilon_1 \) (relative permittivity \( \epsilon_{r1} = \epsilon_1 / \epsilon_0 \)).

We subdivide one half of the cross section into second-order triangular elements as shown in the insert in Fig. 2, where \( \epsilon_{r1} = 1.5 \), the plane of symmetry is assumed to be a perfect magnetic conductor, 36 elements \( (N_p) \) are used, and the number of the nodal points \( (N_x) \) is 91. Computed results (solid lines) for the LSM \( mn \) and LSE \( mn \) modes agree well with the exact results [16]. Spurious solutions \( S_1 \) and \( S_2 \), which are included in (5), do not appear. Spurious solutions \( S_3 \) (dashed lines) corresponding to the solutions of (12) appear only in the region \( \beta/k_0 < 1 \). The solutions \( S_3 \) with cutoff frequencies \( k_0a = \sqrt{2} \pi \) and \( \sqrt{5} \pi \) are equivalent to the TM\(_{11} \) and TM\(_{12} \) modes of a “hollow” waveguide of square cross section, respectively.

One can control the solutions \( S_3 \) by changing the functional (8) as follows [10], [15]:

\[
\tilde{F}_p = \int_\Omega (\nabla \times E)^* \cdot \left( [\mu_r]^{-1} \nabla \times E \right) d\Omega \\
- k_0^2 \int_\Omega \epsilon E^* \cdot E d\Omega \\
+ p^2 \int_\Omega \epsilon^{-1} (\nabla \cdot E)^* (\nabla \cdot E) d\Omega \tag{21}
\]

where \( p \) is a positive number. If \( p \) is set equal to 1, \( \tilde{F}_p \) becomes \( F \). For (21), (12) is reduced to

\[
\left( p^2 \nabla^2 + k_0^2 \right) \psi = 0 \quad \text{in region } \Omega \\
\psi = 0 \quad \text{on perfect electric conductor} \\
\frac{\partial \psi}{\partial n} = 0 \quad \text{on perfect magnetic conductor.} \tag{22}
\]

The appearance of the solutions of (22) is limited to the region \( \beta/k_0 < 1/p \) and the cutoff frequencies of these solutions vary in proportion to the value of \( p \).

Fig. 3 shows the \( p \)-dependence for the solutions \( S_3 \) in the same waveguide as shown in Fig. 2. Solid and dashed lines in Fig. 3 correspond to the TM\(_{11} \) and TM\(_{12} \) modes in Fig. 2, respectively. When \( p = 2 \), the solutions \( S_3 \) appear in the region \( \beta/k_0 < 0.5 \) and the cutoff frequencies of the solutions corresponding to the TM\(_{11} \) and TM\(_{12} \) modes in Fig. 2 become \( k_0a = 2\sqrt{2} \pi \) and \( 2\sqrt{5} \pi \), respectively. When \( p = 0.5 \), the solutions \( S_3 \) appear in the region \( \beta/k_0 < 2 \) and the cutoff frequencies of the solutions corresponding to the TM\(_{11} \) and TM\(_{12} \) modes in Fig. 2 become \( k_0a = 0.5\sqrt{2} \pi \) and \( 0.5\sqrt{5} \pi \), respectively. The \( p \)-dependence is very small for the physical solutions. For larger values of \( p \), however, the degree of accuracy for the physical solutions becomes poorer. For smaller values of \( p \), on the other hand, more spurious solutions appear, because the cutoff frequencies of these spurious solutions become lower. Hereafter, we use \( p = 1 \), namely the functional (8).

Fig. 4 shows the dispersion characteristics for the fundamental mode of half-filled dielectric waveguides, where the plane of symmetry is assumed to be a perfect magnetic conductor. For both \( \epsilon_{r1} = 1.5 \) and 10.0, our results agree well with the results of the H-field finite-element formulation [11].

In Figs. 2 and 4, the normal direction of the interface with an abrupt change in the permittivity coincides with the direction of a coordinate axis.

Next, let us consider a rectangular waveguide with a diamond-shaped dielectric insert [17], as shown in Fig. 5. In this waveguide, there are abrupt changes in the permittivity at the interface whose normal direction does not coincide with the direction of a coordinate axis. Fig. 5 shows the dispersion characteristics for the fundamental
mode, where two planes of symmetry are assumed to be perfect magnetic conductors and one quarter of the cross section is divided into second-order triangular elements. In Fig. 5, the results of the \( \mathbf{H} \)-field formulation with \( N_E = 50 \) and \( N_p = 121 \) and the results of the modal approximation techniques [17] are also presented. For \( \epsilon_r = 1.5 \), the results of the \( \mathbf{E} \)-field formulation with \( N_E = 50 \) and \( N_p = 121 \) agree well with those of the \( \mathbf{H} \)-field formulation. For a larger value of relative permittivity, \( \epsilon_r = 10.0 \), the results of the \( \mathbf{E} \)-field formulation with \( N_E = 50 \) and \( N_p = 121 \) deviate from those of the \( \mathbf{H} \)-field formulation at higher frequencies. However, the \( \mathbf{E} \)-field finite-element solutions can be improved by increasing the number of the elements. The results of the \( \mathbf{E} \)-field formulation with \( N_E = 128 \) and \( N_p = 289 \) are closer to those of the \( \mathbf{H} \)-field formulation.

The computed results in Figs. 2, 4, and 5 prove the validity of (19) and (20).

B. Ferrite-Loaded Waveguides

We consider a ferrite-loaded waveguide as shown in Fig. 6. The ferrite material is characterized by the relative permeability tensor

\[
[
\mu_r
\] = \begin{bmatrix}
3.0 & 0 & j0.8 \\
0 & 1.0 & 0 \\
-j0.8 & 0 & 3.0
\end{bmatrix}
\]  

and a relative permittivity of 2.0 [5]. Here, \( \mu_r \) is independent of frequency, although this assumption is not valid for ferrites in general [5]. Table I shows the dispersion characteristics for the fundamental mode, where \( a = 2b, N_E = 64 \), and \( N_p = 153 \). For both \( a_1 = a \) (completely filled) and \( a_1 = a/4 \), agreement between our results and the exact results [5], [18] is good. In the case of \( a_1 = a/4 \), the value of \( k_0 b \) for \( \beta b = -1 \) is different from that for \( \beta b = 1 \). This fact implies that when \( k_0 \) is given, the modes propagating in this structure (partially filled) in the opposite directions have different phase constants, namely these modes are nonreciprocal [18]. The modes propagating in the completely filled waveguide in the opposite directions, on the other hand, have the same phase constants, and therefore, these modes are reciprocal [18].

V. CONCLUSION

The finite-element method was formulated for the analysis of anisotropic waveguides with a nondiagonal permeability tensor in terms of all three components of the electric field \( \mathbf{E} \). In this approach, spurious solutions do not appear anywhere above the “air-line.” The application of this \( \mathbf{E} \)-field formulation to waveguides with an abrupt discontinuity in the permittivity was discussed in detail.

REFERENCES

References


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