Analysis of Lossy Planar Transmission Lines by Using a Vector Finite Element Method

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***Abstract*—The vector finite element method with hybrid edge/nodal triangular elements is extended for the analysis of lossy planar transmission lines. In order to handle lossy conductor transmission lines, the present approach includes the effect of finite conductivity of a lossy area, and the dissipations in metallic conductors and dielectrics are calculated directly by considering a complex permittivity for the lossy region of interest. Propagation constant formulation is used in the FEM, which avoids spurious solutions absolutely and can handle sharp metal edges in inhomogeneous electromagnetic waveguides. Numerical examples are computed for microstrip lines, finlines, and triplate strip lines. The results obtained agree well with the earlier theoretical and experimental results, and thus show the validity of the method. Also, the current distributions on the lossy microstrip lines with finite strip thickness and isotropic substrates are presented.**

I. Introduction

To SATISFY the need to wide range of applications, the microwave integrated circuits (MIC's) and the microwave monolithic integrated circuits (MMIC's) are becoming very complex in miniaturized form, giving rise to many complicated waveguiding structures, and thus also complicating the theoretical analyses. Therefore, there is still a great demand for more accurate and flexible computer modeling techniques for analysis and design of wide range of practically used waveguiding structures. In this context, the finite element method (FEM) is found to be most versatile method in the simulations of many electromagnetic field problems [1]–[4]. With the introduction of the hybrid edge/nodal elements [1]–[4], the FEM is now well established and sufficiently accurate method for the analysis of different planar transmission lines [5]–[10], and many problems that clouded the previous FEM analyses are no longer existing.

Obviously, accurate computation of transmission-line parameters is an important part of MMIC design, and a significant step in this calculation is the estimation of losses due to conductor and/or dielectric dissipations. While the propagation characteristics of lossless planar transmission lines have been studied extensively by many different methods, comparatively a little information is available on lossy planar transmission lines. Of the analyses on lossy transmission lines, in most cases, the method is incorporated with the perturbation approaches to calculate transmission loss characteristics [11], [12]. However, a great interest has been found recently in the direct calculation of transmission loss by assuming the permittivity of the lossy area to be a complex quantity [13], [14].

Olyslager *et al.* reported the propagation characteristics of a lossy shielded microstrip line [13], where they used a boundary integral equation technique, and also, Tzuang *et al.* analyzed a lossy finline by using a mode matching technique [14]. The FEM has also been applied directly to lossy microstrip lines [7]–[9]. In these reports, losses are included in an exact way without making use of a perturbation approach. In [7] and [8], Helal *et al.* used a high-order hybrid element with facial variables in their analysis, and calculated only the conductor loss in a microstrip line by using two formulations with the FEM [8]. Lee used a lowest-order hybrid element, and calculated substrate loss in a microstrip line [9]. From the numerical results reported in [8], we see that the attenuation constant of a microstrip line varies greatly with the formulations in the FEM. So, accurate formulation to model lossy microstrip line is still a problem to be solved.

The FEM approach [4] has also been successfully applied to lossless microstrip line [5], lossless finline [6], and more recently to lossy microstrip lines [10]. In [10], a combined approach of the FEM and the perturbation technique has been shown to be very efficient to handle arbitrary metallization cross section in lossy microstrip lines. However, the approach has not been applied yet directly to lossy planar transmission lines.

In this paper, we adopted the FEM approach with high-order hybrid edge/nodal triangular elements without any facial variables presented in [4], and extended it to include losses which are very important for many practical applications in MMIC's, in an exact way. As a matter of fact, the inclusion of losses becomes much important, as frequency limits are raised into the millimeter-wave bands, where often such losses are the limiting parameters for successful applications. The novelty of this paper is that both dielectric losses and conductivity losses are included in an exact way. In our analysis, we consider a complex permittivity for the lossy material, and obtain a matrix eigenvalue problem, which gives solutions directly for the complex propagation constants that include the phase and attenuation constants at a given frequency. In the formulation [4], since the edge and nodal elements are used to describe the transverse and axial fields, respectively, the approach ensures tangential continuity of electric fields

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at the sharp perfectly conducting edges without using any special singular functions, and also avoids spurious solutions completely. Numerical results are shown for microstrip lines, finlines, and triplate strip lines in order to verify the approach. Also, the validity of the numerical results has been checked with previous theoretical and experimental data.

For a meaningful realization of performance of a microstrip line, clarification of current distributions produced by the electromagnetic fields is very important. The current distributions on microstrip lines for lossless strips are discussed by Kobayashi and Takaishi in [15] and in the references therein. Also, How *et al.* reported that for a lossy microstrip line with a thin metallization [16]. However, the current distributions on lossy microstrip line with finite metallization thickness has not been discussed yet to our knowledge. In our analysis, we calculated the current from the electric fields in the lossy microstrip line, and the longitudinal current distributions have been presented for various conductor thickness on isotropic substrates.

II. Finite Element Method

A. Basic Equations

Let us consider a shielded electromagnetic waveguide having uniform cross section along the direction of propagation, the z axis, and the shielding walls are assumed to be perfect electric conductors. The electromagnetic wave that propagates through the waveguide has a phase factor of the form $\exp(j\omega t - j\gamma z)$, where ω is the angular frequency, t is the time, and $\gamma = \beta - j\alpha$ is the (complex) propagation constant with β being the phase constant and α being the attenuation constant. The relative permittivity of the region of interest is $\tilde{\epsilon_r}$, which is assumed to be a complex number for lossy media.

We start with the Maxwell's equations for the lossy case as follows

$$\nabla \times \mathbf{E} = -j\,\omega\,\mu_0\,\mathbf{H} \tag{1}$$

$$\nabla \times \mathbf{H} = j\,\omega\,\epsilon_0\,\tilde{\epsilon_r}\,\mathbf{E} \tag{2}$$

where $\mathbf{E}$ and $\mathbf{H}$ are electric and magnetic field vectors, respectively, ϵ_0 and μ_0 are permittivity and permeability of free space, respectively.

The vectorial wave equation in terms of the electric field is derived as

$$\nabla \times (\nabla \times \mathbf{E}) - k_0^2\,\tilde{\epsilon_r}\,\mathbf{E} = 0 \tag{3}$$

where $k_0 = \omega\sqrt{\epsilon_0\mu_0}$ is the free space wave number. The functional for the wave equation is given by

$$F = \iint_\Omega \left[(\nabla \times \mathbf{E})^* \cdot (\nabla \times \mathbf{E}) - k_0^2\,\tilde{\epsilon_r}\,\mathbf{E}^* \cdot \mathbf{E}\right] dxdy \tag{4}$$

where Ω is the cross section of the waveguide in the transverse plane, and the asterisk denotes a wave with the z-dependence of the form $\exp(j\gamma z)$.

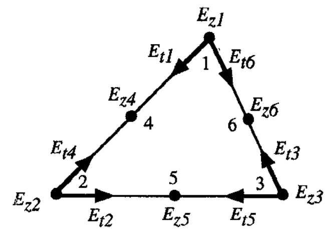

Fig. 1. High-order hybrid edge/nodal triangular element.

B. Hybrid Edge/Nodal Elements

Fig. 1 shows the high-order hybrid edge/nodal triangular element [4], which is composed of a linear edge element with six tangential unknowns at the three vertices of the triangle, E_{t1} to E_{t6}, and a quadratic nodal (conventional Lagrange) element with six axial unknowns, E_{z1} to E_{z6}. We use electric field as the working variable because it will provide some additional advantage over the magnetic field for the case of electrically shielded transmission lines. Since the tangential electric field over a perfect electric conductor vanishes, working with the electric field will considerably reduce the system size and computation time as well.

The transverse field components E_x and E_y are described by the edge element shape function vectors, and the axial components E_z are described by the nodal element shape function vectors. Thus the electric field $\mathbf{E}$ in each element is expressed as

$$\mathbf{E} = \begin{bmatrix} E_x \\ E_y \\ E_z \end{bmatrix} = \begin{bmatrix} \{U\}^T\{E_t\}_e \\ \{V\}^T\{E_t\}_e \\ j\{N\}^T\{E_z\}_e \end{bmatrix}. \tag{5}$$

Here $\{U\}$ and $\{V\}$ are the shape function vectors for the linear edge elements, $\{N\}$ is the ordinary shape function vector for the quadratic nodal elements, and T denotes a transpose. The interpolating functions for the shape functions are detailed in [4]. The description of transverse field components in each element, apparently let them change their direction infinitely rapidly at the edge. Hence, it ensures tangential continuity of electric fields at the sharp perfectly conducting edges of planar transmission lines.

Here it might be interesting to note that different hybrid type elements have been introduced in the FEM world [1]–[4]. A very similar high-order element as that we used in this paper has been reported by Lee *et al.* [2], where the second order approximation is attained by using eight edge variables, which in our present case is six, in each element. Actually, Lee *et al.* used two extra facial local variables with two of the three sides of an element in order to provide a quadratic approximation of the normal component of the field. Although, Lee's element may provide better accuracy for many cases because of the use of those facial variables, the FEM approach [4] with the element as depicted in Fig. 1 is also found to be efficient to handle planar transmission lines [5], [6], [10]. These inspired us to extend the method for the analysis of lossy microwave problems.

C. Eigenvalue Problem

We divide the waveguide cross section into a number of high-order hybrid edge/nodal triangular elements. With the introduction of the complex permittivity for the lossy materials and γ in the formulation, and applying the finite element method discussed in [3], the following matrix equations are derived

$$[K_{tt}]\{E_t\} - \gamma\,[K_{tz}]\{E_z\} - \gamma^2\,[M_{tt}]\{E_t\} = \{0\} \quad (6)$$

$$-\gamma\,[K_{zt}]\{E_t\} + [K_{zz}]\{E_z\} = \{0\} \quad (7)$$

with

$$[K_{tt}] = \sum_e \iint_e \Big[\tilde{\epsilon_r}\,k_0^2\{U\}\{U\}^T + \tilde{\epsilon_r}\,k_0^2\{V\}\{V\}^T + \Big(\{U_y\} - \{V_x\}\Big)\Big(\{V_x\}^T - \{U_y\}^T\Big)\Big]dxdy \quad (8a)$$

$$[K_{tz}] = \sum_e \iint_e \Big[\{U\}\{N_x\}^T + \{V\}\{N_y\}^T\Big]\,dxdy = [K_{zt}]^T \quad (8b)$$

$$[K_{zz}] = \sum_e \iint_e \Big[\tilde{\epsilon_r}\,k_0^2\{N\}\{N\}^T - \{N_x\}\{N_x\}^T - \{N_y\}\{N_y\}^T\Big]dxdy \quad (8c)$$

$$[M_{tt}] = \sum_e \iint_e \Big[\{U\}\{U\}^T + \{V\}\{V\}^T\Big]\,dxdy \quad (8d)$$

where $\{U_y\} \equiv \partial\{U\}/\partial y$, $\{V_x\} \equiv \partial\{V\}/\partial x$, $\{N_x\} \equiv \partial\{N\}/\partial x$, and $\{N_y\} \equiv \partial\{N\}/\partial y$, and their explicit forms are detailed in [4].

Eliminating the axial component from (6) and (7), the following final eigenvalue equation is obtained

$$[K_{tt}]\,\{E_t\} - \gamma^2\Big([M_{tt}] + [K_{tz}][K_{zz}]^{-1}[K_{zt}]\Big)\{E_t\} = \{0\}. \quad (9)$$

We see that the eigenvalue problem involves only the edge variables in the transverse plane, $\{E_t\}$, and all the matrices comprising the eigenvalue problem are complex. Therefore, the eigenvalue equation can be solved to give a solution for the complex propagation constant as the eigenvalue by using a complex eigenvalue solver. For the lossless case, we have seen that the matrices in the eigenvalue equation are real but the eigenvalues and eigenvectors can still possibly be complex conjugate (as in the case of complex modes in lossless guides) [5], [6]. Since no fields exist inside the lossless conductor, it is reasonable to discretize the problem geometry neglecting the conductor area in the lossless case, but, for the lossy case, the conductor area should also be discretized as the fields exist inside the conductor in this case.

To choose a solver for complex eigensystems, which usually appear when lossy systems are studied, is of great importance. The complex eigenvalue solvers in [17] and [18] are useful for large, sparse complex matrices, and are based on subspace iteration algorithms. As it is described there that they provide better accuracy in relatively short time with reasonable memory requirement, hence they are very efficient. However, in case of dense complex matrices, they can not be used. In our case, since the complex matrices are not sparse, we have no other choice but to use a dense complex eigenvalue solver even though it is a time and memory consuming process. It is needless to say that the FEM approach of this paper would be very useful if appropriate sparse matrix solvers are unavailable.

D. Dissipations in Conductor and Dielectrics

For the lossy case, current flowing along the metallic strip may have all the three vector components as [19]

$$\mathbf{J} = \mathbf{x}_0 J_x + \mathbf{y}_0 J_y + \mathbf{z}_0 J_z \quad (10)$$

where $\mathbf{x}_0$, $\mathbf{y}_0$, and $\mathbf{z}_0$ are the unit vectors. The current vector $\mathbf{J}$ at any point on the strip is related with the eletric field vector $\mathbf{E}$ as

$$\mathbf{J} = \sigma\mathbf{E} \quad (11)$$

where σ is the conductivity of the metallic conductors.

Therefore, as it is evident from (10) and (11), all the current components may be determined from the FEM field solutions easily. The transverse electric field components may be obtained directly from the eigenvectors of (9), and thus the transverse current components may be calculated. The longitudinal electric field components may be calculated from the transverse electric fields as [10]

$$\{E_z\} = \gamma\,[K_{zz}]^{-1}\,[K_{zt}]\,\{E_t\} \quad (12)$$

and thus the longitudinal current at a point may be calculated from the longitudinal field at that point, easily.

In this analysis, the relative permittivity of the lossy conductor is assumed to be a complex number and contains the conductivity σ of the conductor as

$$\tilde{\epsilon_r} = 1 - j\,\frac{\sigma}{\omega\epsilon_0} \quad (13)$$

and for lossy dielectrics, the complex permittivity is expressed as

$$\tilde{\epsilon_r} = \epsilon_r\,(1 - j\tan\delta) \quad (14)$$

where ϵ_r is the relative permittivity and $\tan\delta$ is the loss tangent. In fact, the complex nature of the relative permittivity as shown in (13) and (14) are responsible to make the propagation constants complex in nature, and lead us to calculate the phase and attenuation constants directly by using the FEM procedure.

Of course, confusion arises when σ is considered to be a scalar quantity. For some materials, σ may depend on both the magnitude and direction of $\mathbf{E}$, and hence is a nonlinear dyadic quantity [20]. However, the present approach may give sufficient accurate results for many practical cases, when it is assumed that the transmission conductors are homogeneous materials of conductivity σ [21].

III. Computational Results

The complex eigenvalue problem (9) has been solved with γ^2 treated as the eigenvalue using a (dense) QR algorithm, as implemented in the double precision version. For all the following examples, we calculated the dominant mode characteristics. Even though, an ambiguity arises in judging the

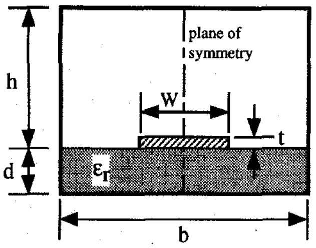

Fig. 2. Cross section of a shielded microstrip line.

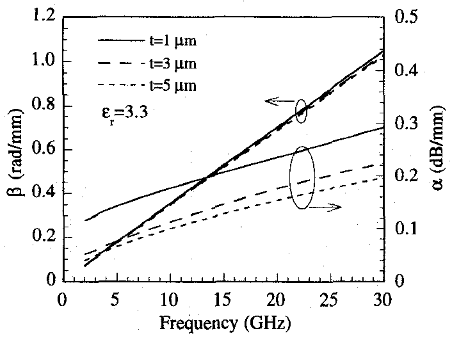

Fig. 3. Phase and attenuation constants versus frequency for a lossy microstrip line on isotropic substrate.

dominant mode of the lossy structure as because the propagation constant is complex number for both the propagating and evanescent modes, we determined the dominant mode by comparing the phase constants for different modes [9]. However, this could be a severe problem when other phenomena such as leakage effects or surface wave losses are to be accounted for in the millimeter-wave treatment of planar transmission lines, and are not included in this paper.

First, we consider a shielded microstrip line shown in Fig. 2. The microstrip line has a lossless isotropic dielectric substrate with the thickness d=10 μm, $\epsilon_r = 3.3$, and a gold strip with the width W= 22 μm and the conductivity $\sigma = 3.9 \times 10^7$ S/m. To analyze this structure, we take the advantage of symmetry and discretize half of the cross section into elements shown in Fig. 1.

Fig. 3 shows the phase and attenuation constants of lossy microstrip lines with three different thickness of the strip: t= 1, 3, or 5 μm. We can see that the thickness of the strip induces only a small variation in the phase constant, especially at high frequency, but, the attenuation constants vary greatly with the thickness over the whole frequency range under consideration. Therefore, in the miniaturized MIC's and MMIC's, the thickness of the strip should be considered very carefully.

Fig. 4 shows the longitudinal current over the symmetric half space on the lower surface of the strip at 10 GHz. We see that the amplitude of current increases near the strip edge, especially for thin strip. In the calculation, we have seen that the transverse components of current are very, very small compared to the axial components, which is the same as that reported by Kobayashi *et al.* for a lossless microstrip line [15].

A lossy unilateral finline shown in Fig. 5 is also analyzed, and the results are presented in Fig. 6. For the finline, we

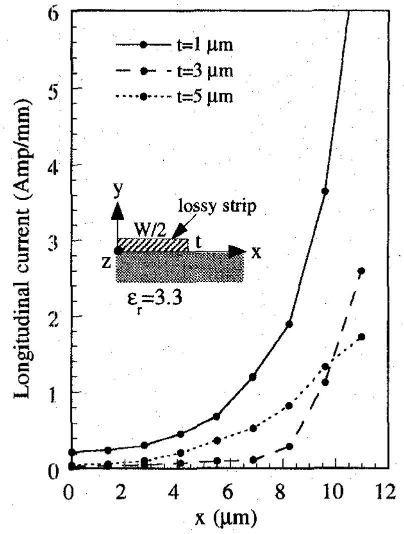

Fig. 4. Longitudinal current on the lower surface of the strip at 10 GHz for the microstrip line of Fig. 2.

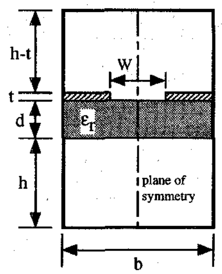

Fig. 5. Cross section of a lossy unilateral finline.

use b= 3.556 mm, d= 0.127 mm, h= 3.4925 mm, t= 5 μm, and ϵ_r= 2.22. Here the metallic fins are assumed to be lossy materials with conductivity $\sigma = 3.333 \times 10^7$ S/m and the dielectric material is assumed to be lossless. For this structure also, we discretize half of the cross section. In Fig. 6, the solid lines show the results of FEM and the closed circles show the results of Tzuang *et al.* [14], where they used a mode matching approach for the analysis of lossy finlines. The crosses represent the phase constants of the lossless finline. From the FEM calculation, we see that the phase constants of lossless and lossy finlines do not vary too much, and the agreement between our results and those of Tzuang *et al.* [14] is better at higher frequencies.

Finally, we apply the FEM to the calculation of transmission losses of a triplate strip line shown in Fig. 7. Since this is a symmetrical-strip transmission line having two-fold symmetry, we discretize only one-fourth of the cross section by applying appropriate boundary conditions on the symmetry planes and calculate the transmission loss. Here we use $2h+t$= 910 μm, t= 10 μm, W= 200 μm, ϵ_r= 7.55, tanδ= 0.005, and $\sigma = 2.5 \times 10^7$ S/m. The transmission loss includes both the conductor loss and the dielectric loss. The upper and lower shielding plates are assumed to be lossless. Fig. 8 shows the frequency dependence of transmission loss, where the broken and dashed-dotted lines show, respectively, the results of experiment and simulation reported by Taguchi *et al.* [22]. Here the higher computed loss

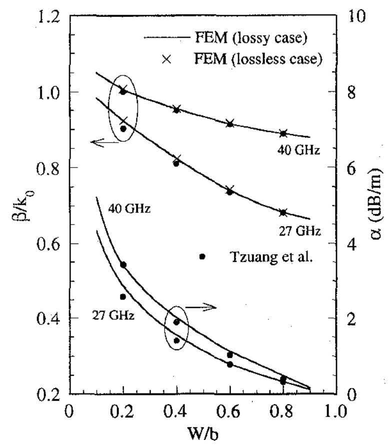

Fig. 6. Phase and attenuation constants versus slot-width.

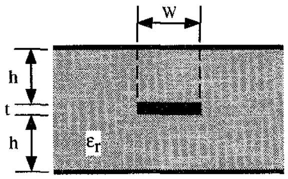

Fig. 7. Cross section of a triplate strip line.

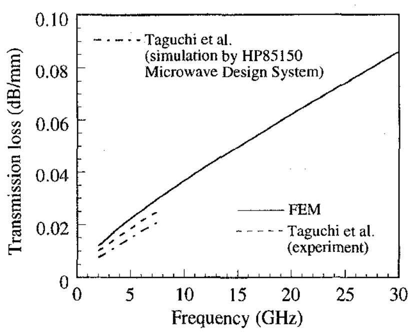

Fig. 8. Transmission loss versus frequency.

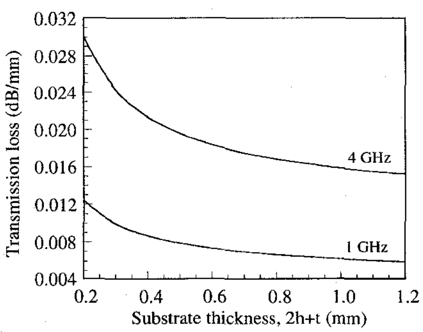

Fig. 9. Transmission loss versus substrate thickness.

could be due to the upper and lower plates being assumed to be lossless. Also, we see that with the increase in frequency, the transmission loss increases. Fig. 9 shows how the transmission loss varies with the substrate thickness, where $t = 20$ μm. We see that transmission loss decreases with the increase in substrate thickness.

In the calculation, we used around 200 hybrid edge/nodal triangular elements. The size of eigenvalue matrix equation follows about 600 and required memory is about 15 MB. The computations were performed on a NEWS-5000 UA workstation with main memory of 64 MB and virtual memory of 300 MB, where the time required is about 20 minutes to obtain a complex propagation constant at a given frequency.

It is worth mentioning that calculating loss directly by using the FEM has an added advantage, as because the approach is applicable to planar transmission lines with metallization thickness of the order of or smaller than the skin depth, the approach could be successfully applied to different micron-sized waveguides in recent MIC and MMIC applications.

IV. Conclusion

The finite element method with high-order hybrid edge/nodal triangular elements was employed to calculate the propagation characteristics of lossy planar transmission lines, such as microstrip lines, finlines, and triplate strip lines, where no spurious (nonphysical) solutions appeared anywhere. The analysis includes the finite metallization thickness and losses of the metal conductor, and also losses in dielectric material directly without using any perturbation approach. The accuracy of numerical results was checked by comparing with the earlier theoretical and experimental results. It has been observed that the approach is very efficient especially for high frequency applications, and it does not work very well when the operating frequency is too low. Current distributions on microstrip lines with isotropic substrates are also presented by using the FEM field solutions. Even though we have considered a few examples in this paper, many other composite structures with multilayered isotropic/anisotropic substrates may also be handled with this approach.

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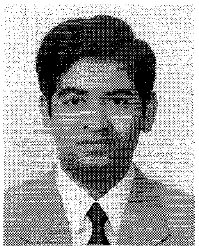

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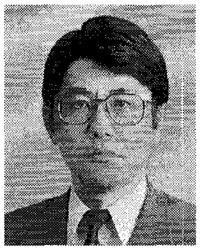

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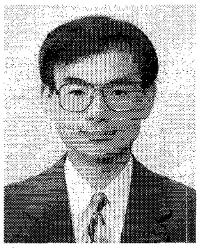

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