Quasi-TEM Surface Impedance Approaches for the Analysis of MIC and MMIC Transmission Lines, Including Both Conductor and Substrate Losses

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Abstract—**The surface impedance approach is applied to the quasi-TEM spectral domain analysis of MIC and MMIC lines with lossy metallizations in both the weak skin effect and the strong skin effect regimes. The use of the spectral domain technique makes it possible the analysis of lines with multilayer iso/anisotropic substrates, including semiconductor and/or magnetic layers. PC computer codes have been developed following the proposed technique. Computation times within a few seconds are achieved.**

I. Introduction

The thickness of metallizations in MMIC circuits is about ten times smaller than the usual metallization thickness in hybrid MIC's. In this case the well known incremental inductance rule for conductor loss calculations may not be adequate. In the analysis of MMIC transmission lines, the use of a quasi-TEM approach is suggested because of the small dimensions of these structures [1]. The advantages of the quasi-TEM spectral domain techniques for the analysis of quasi-planar transmission lines are well known. The spectral domain approach makes it possible to analyze general multistrip transmission lines embedded in multilayered complex substrates within reasonably short computation times. Dispersion due to the dependence on frequency of the characteristic parameters of the substrate and/or metallizations, can be analysed by suitable quasi-TEM approaches. Following this method, some of the authors of the present paper have analysed the feasibility of the spectral domain quasi-TEM technique in planar transmision lines on lossy semiconductors and other complex multilayered substrates [2], [3]. The effects of finite thickness in lossless metallizations are analysed in [4] using an spectral domain quasi-TEM technique. In reference [5] a quasi-TEM spectral domain technique for the analysis of simple microstrip lines with conductor losses is developed. The method in [5] is able to deal with conductor strip thickness of the order of the skin depth by using the surface impedance approach [6]–[9].

In this paper a two-plate quasi-TEM surface impedance (quasi-TEM-SI) approach is presented for the analysis of lines with lossy conductors. Following this approach computer codes can be developed for the analysis of both conductor and substrate losses in MIC and MMIC lines on multilayer anisotropic substrates, including gyroelectric and gyromagnetic materials. These codes have been implemented on a personal computer (PC) and computation times of a few seconds have been achieved.

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II. Quasi-TEM-SI Approach

A. Basic Assumptions

The structure under analysis is shown in Fig. 1(a). It is a multistrip transmission line embedded in a multilayer medium, made of anisotropic dielectric and/or magnetic materials, including lossy as well as gyrotropic layers magnetized along the direction of propagation (i.e. the direction of propagation always coincides with a main axis of the permeability and permittivity tensors). We also regard the possibility of having strips made of two different conductors.

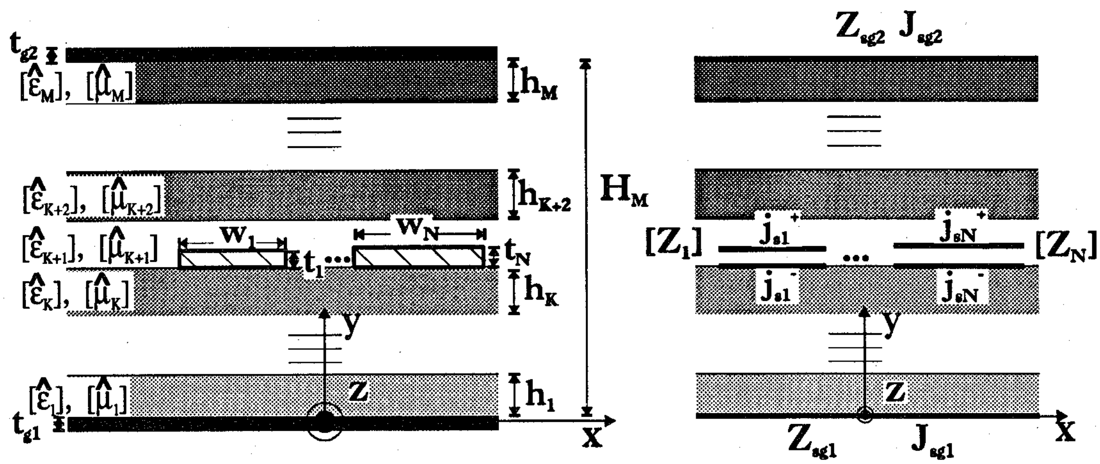
The quasi-TEM approach will remain valid outside the conductors as long as the transverse components (x and y components in Fig. 1(a)) of the electric and magnetic fields are much greater than their corresponding longitudinal components (z component). Nevertheless, since the longitudinal component of the magnetic field must be negligible with respect to the transverse components, the transverse components of all the ohmic currents inside the metallizations must be also negligible with respect to the longitudinal component. So, the validity of the quasi-TEM approach outside the conductors implies that the transverse components of the electric field inside the conductors must be much smaller than the longitudinal component [5].

Finally, we assume the validity of the *good conductor* approach for the line conductors ($\sigma_c \gg \omega\epsilon_c$), i.e. the diffusion equation is the equation that the fields inside the conductors must satisfy. All these aproximations can be summarized as follows:

1) *Outside the conductors*: The longitudinal field components are negligible with respect to the transverse components. The equation that the fields must satisfy is the wave equation for the TEM field.
2) *Inside the conductors*: The longitudinal magnetic field component is negligible with respect to the trasverse components and the transverse electric field components are negligible with respect to the longitudinal components.

B. Field Potentials and Basic Equations

We will suppose a quasi-TEM wave propagating along the line in Fig. 1(a). The space-time dependence of the fields is of the kind $\boldsymbol{E} = \boldsymbol{E}_0(x, y)\exp(j\omega t - j\gamma z)$. According to the



(a) (b)

Fig. 1. (a) Multistrip transmission line embedded in a multilayered medium, made of general anisotropic dielectric and/or magnetic layers, including lossy, as well as gyrotropic materiels. Each strip of the line can be made of a multilayer conducting media. (b) Approximately equivalent structure.

assumptions made above, the unique relevant components of the magnetic field, in the entire cross section of the line, are the transverse components. Therefore, they can be obtained from a longitudinal potential vector A_z in the usual way

$$\boldsymbol{H}_{t,n} = -[\mu]_n^{-1}\boldsymbol{a}_z \times \nabla_t A_z \tag{1}$$

where $[\mu]_n$ is the transverse permeability tensor

$$[\mu]_n = \begin{pmatrix} \mu_n^{xx} & \mu_n^{xy} \\ \mu_n^{yx} & \mu_n^{yy} \end{pmatrix}. \tag{2}$$

The subscript t stands for *transverse component* and the subscript n stands for each one of the different homogeneous layers of the line. The electric field can be obtained in terms of the scalar electric potential ϕ and the vector magnetic potential $A_z\boldsymbol{a}_z$ as usual

$$\boldsymbol{E}_n = -\nabla_t\phi + j\gamma\phi\boldsymbol{a}_z - j\omega A_z\boldsymbol{a}_z \tag{3}$$

where γ is the propagation constant along the line. Nevertheless, according to the basic assumptions made above, only the electric transverse components are relevant outside the conductors, and only the electric longitudinal components are relevant inside the conductors. Therefore, (3) can be simplified to

$$\boldsymbol{E}_{t,n} = -\nabla_t\phi \quad \text{outside the conductors} \tag{4}$$

and

$$E_{z,i} = j\gamma V_i - j\omega A_z \quad \text{inside the conductors} \tag{5}$$

where V_i are the constant values of the electric potential over the ith conductor of the line cross section. In (5) it has been taken into account that the transverse electric field components are negligible inside the conductors, and as a consequence of this, that the electric potential is constant.

The differential equations for the electric and magnetic potentials outside the conductors can be deduced from the divergence equations for the transverse electric and magnetic fields

$$\nabla_t \left\{ \begin{matrix} [\epsilon]_n \nabla_t\phi \\ [\epsilon^{eq.}]_n \nabla_t A_z \end{matrix} \right\} = 0 \quad \text{outside the conductors} \tag{6}$$

where $[\epsilon]_n$ is the transverse permitivity tensor for each layer and where a transverse permitivity equivalent tensor has been defined [2]

$$[\epsilon^{eq.}]_n = \frac{[\mu]_n^t}{\det([\mu]_n)} \tag{7}$$

where the superscript t stands for the transpose matrix.

The differential equations for the relevant magnetic potential components inside the conductors is deduced from (5) and the equation for the longitudinal potential vector

$$\{\nabla_t^2 - j\omega\mu_i\sigma_i\}A_z = -j\mu_i\sigma_i\gamma V_i \quad \text{inside the conductors.} \tag{8}$$

C. Two-Plate Surface Impedance Model

Equation (6) for the vector potential can be solved for the structure of Fig. 1(a) with the appropriate boundary conditions on the surfaces of the metallizations. Anyway we replace the actual problem by an other aproximately equivalent problem, where each strip is substituted by two surface current plates at the upper and the bottom strip interfaces (see Fig. 1(b)), which are related to the longitudinal electric fields at the interfaces by means of a surface impedance matrix

$$\begin{pmatrix} E_z^+ \\ E_z^- \end{pmatrix} = \begin{pmatrix} Z_{1,1} & Z_{1,2} \\ Z_{2,1} & Z_{2,2} \end{pmatrix} \cdot \begin{pmatrix} J_s^+ \\ J_s^- \end{pmatrix} \tag{9}$$

where $E_z^+(E_z^-)$ is the electric field component at the upper (bottom) strip interface, and the surface current $J_s^+(J_s^-)$ is equal to the tangential magnetic field $-H_x^+(H_x^-)$ at the upper (bottom) strip interface. The calculus of the surface impedance matrix will be discussed in Appendices A and B. By solving (6) as explained in [2], one obtains a Green's function in the spectral domain which relate $\tilde{A}_z$ and $\tilde{J}_s$, where $\tilde{}$ stands for spectral domain

$$\tilde{A}_z(k_x, y) = \tilde{G}(k_x; y, y') \cdot \tilde{J}_s(k_x, y') \tag{10}$$

k_x being the spectral variable.

Equations (5) and (9) make it possible to write the following expresion for the upper and bottom strip interfaces

$$\left[\begin{bmatrix}\tilde{G}_{1,1} & \tilde{G}_{1,2}\\ \tilde{G}_{2,1} & \tilde{G}_{2,2}\end{bmatrix} - \frac{j}{\omega}\begin{bmatrix}Z_{1,1} & Z_{1,2}\\ Z_{2,1} & Z_{2,2}\end{bmatrix}\right] \cdot \begin{bmatrix}\tilde{J}_s^+\\ \tilde{J}_s^-\end{bmatrix} = \mathcal{F}\left(\frac{j\gamma}{\omega}V_i\right) \cdot \begin{bmatrix}1\\ 1\end{bmatrix} \tag{11}$$

where $\mathcal{F}$ stands for the Fourier transform.

The method used for obtaining the Green's function [2] has been initializated with a resistive complex boundary condition at the ground planes, in order to take into account the effects of the finite conductivity of the ground plates. When anisotropic media are present, this equation in the frame of the spectral domain can be written as

$$\tilde{A}_z = -\frac{jZ_{m1(2)}}{\omega}\epsilon_{\text{int}}^{1(M)}\left.\frac{\partial \tilde{A}_z}{\partial n}\right|_{y=0,H_M} \tag{12}$$

where

$$Z_{m1(2)} = \frac{1+j}{\delta_{1(2)}\sigma_{1(2)}}\coth\left(\frac{1+j}{\delta_{1(2)}}t_{g1(2)}\right) \tag{13}$$

and

$$\epsilon_{\text{int}}^{1(M)} = \frac{\eta_{x,x}^{1(M)}}{\mu_0 - \dfrac{K_x}{\omega}Z_{m1(2)}\eta_{x,y}^{1(M)}} \tag{14}$$

and where n stands for the outward surface conductor direction, $\delta_{1(2)}$ is the skin depth, $\sigma_{1,(2)}$ is the conductivity, $t_{g,1(2)}$ is the thickness of the ground plate, $[\eta] = [\mu_r]^{-1}, k_x$ is the spectral variable, 1(2) stand for the bottom (upper) ground plate and $1(M)$ stand for the bottom (upper) layer of Fig. 1.

Equation (11) can be solved by using Galerkin method. Once the surface current densities are obtained, the total current of each strip can be calculated by integrating the surface current densities along the strip $I_j = \int_{-w/2}^{w/2}(J_{s,j}^+ + J_{s,j}^-)\,dx$. After some iterations in which the canonical excitations $\gamma V_i = \delta_{i,j}; j = 1, 2, \cdots N$ (N is the total number of strips) are used, the total currents I_i over the conductors can be linearly related with the excitations γV_i by means of a complex matrix $\hat{L}_{i,j}$ as follows:

$$\hat{L}_{i,j} = \left.\frac{\gamma V_i}{\omega I_j}\right|_{V_j=0\ \forall\ j\neq i} \tag{15}$$

where the $\hat{L}_{i,j}$ matrix has dimensions of induction per unit length. For conductors with finite conductivity, $\hat{L}_{i,j}$ is a complex matrix, whose imaginary part can be identified with the mutual per unit length resistance of the line, and its real part with the per unit length inductance matrix of the line

$$\hat{L}_{i,j} = L_{i,j} - \frac{j}{\omega}R_{i,j}. \tag{16}$$

The electrostatic problem for the determination of the complex $\hat{C}_{i,j}$ matrix from (6) with constant potentials at the conductors is solved by using the Galerkin method in the spectral domain. This method is developed in [4] for structures with thick strips. Two conducting plates are considered at each strip, because this approximation is good enough for the usual MIC and MMIC configurations [4]. The spectral Green's functions proposed in [2] are used to allow the presence of lossy and/or gyrotropic layers.

Once the inductance and the capacitance matrices have been calculated, the usual quasi-TEM expressions for the line complex propagation constant and impedance are obtained.

D. Improvement of the Two-Plate Model

Although the two-plate model suffices in many cases, certain lines having thick strips can not be properly approximated by using this model. Therefore it should be desirable to have a more accurate model for structures having high values of the ratio between the strips thickness and the strips width. This model has been successfully developed in [13] for lossless conductors. However, a direct generalization of the N-layer model proposed in [13] is not possible for lossy thick strips. Nevertheless, it is still possible to take some advantage of the model proposed in [13] for lossless strips. In fact, it is expected that the failure of the two-plate model for thick strips will be mainly related to the computation of the external inductance rather than to the computation of the internal inductance and resistance. Starting from this idea, a procedure to estimate the complex inductance $\hat{L}_{i,j}$ of thick-strip lossy lines is going to be proposed. First of all, the lossless thick-strip line complex inductance $\hat{L}_{i,j}^{0,N}$ is computed by using the N-layer model proposed in [13]. Then, the lossless thick-strip line complex inductance is computed again by using the two-plate model. The obtained value $\hat{L}_{i,j}^{0,2}$ is then subtracted from the two layers lossy thick-strip line complex inductance, $\hat{L}_{i,j}^{2}$, obtained by means of the method described in this section. Finally, the obtained incremental complex inductance matrix is added to the N-layer lossless complex inductance. The final expression for the Nth degree of approximation is

$$\hat{L}_{i,j}^{N} = \hat{L}_{i,j}^{0,N} + (\hat{L}_{i,j}^{2} - \hat{L}_{i,j}^{0,2}). \tag{17}$$

The complex capacitance matrix $\hat{C}_{i,j}$ is obtained by also using the N-layer model of [13], and the line parameters are computed by using the conventional expressions. This method provides improved results, with a small increase in the computation time.

III. Numerical Procedure and Results

The Galerkin method in the spectral domain was used for solving the integral equations for the charge density and the longitudinal current density on the plates. For the longitudinal current density a nonuniform linear piecewise aproximation, with triangular basis functions, [5] was used. The same basis functions were used for the electrostatic problem. Notice that the weighted Chebyshev polynomials [4] can not be used in the magnetostatic problem [10]. Triangular basis functions were also used in the electrostatic problem for completeness. Typical CPU times in a 486/66 PC are about 5–10 sec. for a

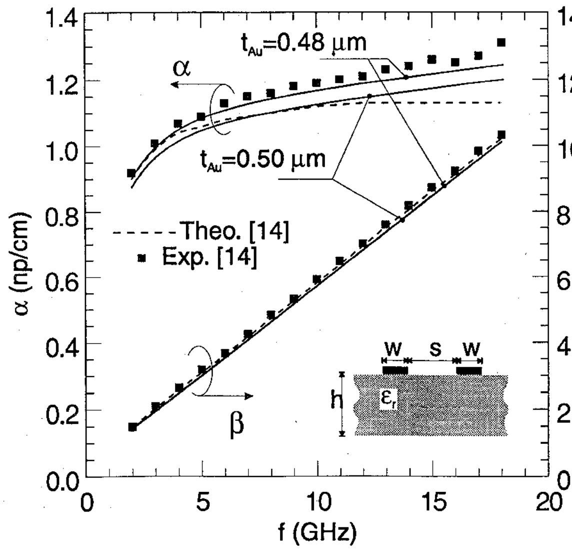


Fig. 2. Phase and attenuation constants for two coupled strips on a GaAs substrate. w = 4 μm, s = 8 μm, h = 670 μm, t_{Ti} = 0.05 μm, t_{Au} = (0.48–0.45) μm, ϵ_r = 13. (——) Our two-plates model results. (– – –) Theoretical results in [14]. (■) Experimental results in [14].

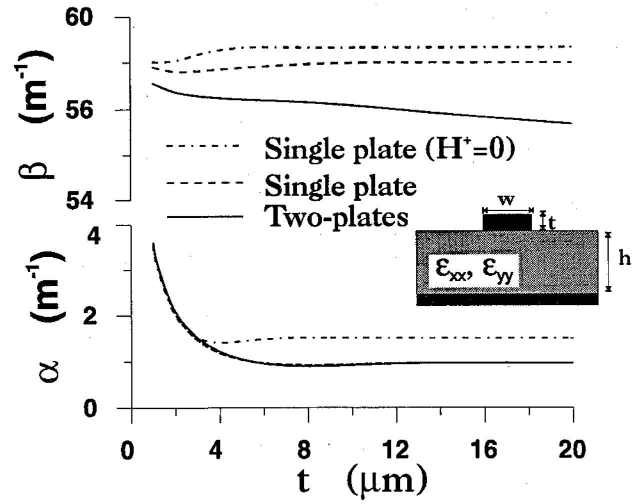


Fig. 3. Phase and attenuation constants for a microstrip line on sapphire substrate. w = 40 μm, h = 200 μm, ϵ_r^{xx} = 9.4, ϵ_r^{yy} = 11.6. (——) Two–plates results. (– – –) Single–late results using and (20) and (21). (– · –). Single plate results [8].

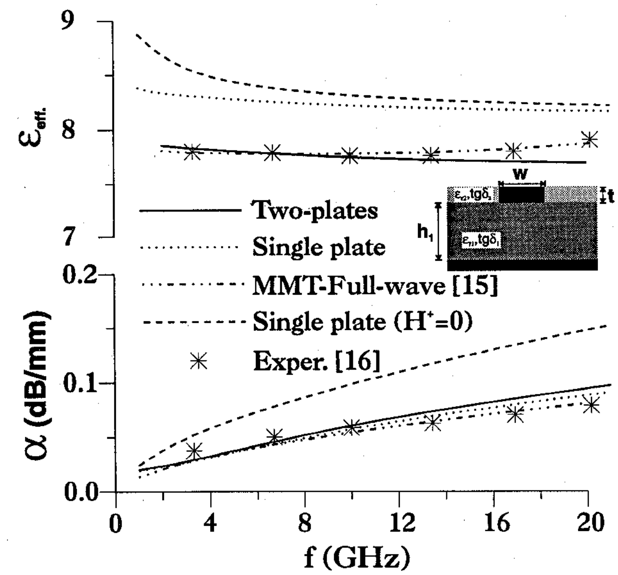


Fig. 4. Attenuation and effective dielectric constants for a microstrip line in a layered lossy substrate. $w = 30$ μm, $h = 200$ μm, $t = 6$ μm, $\epsilon_{r,1} = 12.9$, tg $\delta_1 = 3 \times 10^{-4}$, $\epsilon_{r,2} = 3.4$, tg $\delta_2 = 0.05$, $\sigma = 1.77 \times 10^7$ $(\Omega\text{m})^{-1}$. (——) Two-plates results. (– – –) Single-plate results. (– · –) Theoretical results in [15]. (*) Experimental results in [16].

single microstrip (Figs. 3, 4, and 6) and 20–30 sec. for coupled microstrip lines (Fig. 2) or for a coplanar waveguide (Fig. 5).

In Fig. 2 our results obtained by using the two-plate quasi-TEM-SI approach are compared with those provided in [14] for two coplanar strips on a GaAs substrate. The strips are made of gold with a thin titanium layer of 0.05 μm to promote adhesion with the substrate. Measured values of the metallization thickness are between 0.53 and 0.55 μm [14]. Thus the gold layer thickness varies between 0.48 and 0.50 μm. In the figure the computed values of both the attenuation and the phase constants are plotted. Both experimental and theoretical values given in [14] are also plotted. Good agreement with the experimental results is observed. Numerical computations show that the influence of the titanium layer thickness, with the poorest conductivity, is negligible in this frequency range. Nevertheless the influence of the gold layer conductivity is critical at the analyzed frequency and dimensions.

A comparison between the results obtained by using different single- and two-plate quasi-TEM-SI approaches is shown in Fig. 3 for a microstrip line on anisotropic sapphire substrate. The main features of the single-plate quasi-TEM-SI approach are reported in [5]. Two different expressions for the surface impedance of the single-plate quasi-TEM-SI approach have been used. The first one is that reported in [8]. The second one is an improved expression, which takes into account the presence of a non-zero tangential magnetic field on the upper strip interface (see Appendix B). A good agreement is observed between the two-plate results for the attenuation constant and the single-plate results when (20) and (22) are used. However, the results for the phase constant are different from the two-plates results for any single-plate SI model. These results show that a good single-plate SI model can give accurate results for the attenuation constant, provided that a good estimation is made for the H^+/H^- ratio. Nevertheless, good approximations for the phase constant can be only achieved by using a two-plates model.

Fig. 4 shows a comparison between our results for a microstrip line and those provided in [15]. The technique used in [15] is full-wave mode matching. In Fig. 4 the experimental data obtained in [16] are also shown. A good agreement is observed for the attenuation constant for both the two- and the single-plate models (using (20) and (21)). Once again good results for the effective dielectric constant are only provided by the two-plates model. The disagreement between our results and those of [15] for the highest values of the h/λ_0 ratio is believed to be due to the failure of the quasi-TEM approach owing to full-wave effects. This disagreement should be present even in the absence of conductor losses.

The two main limitations of the approach developed here are the presence of full-wave effects in the lossless problem,

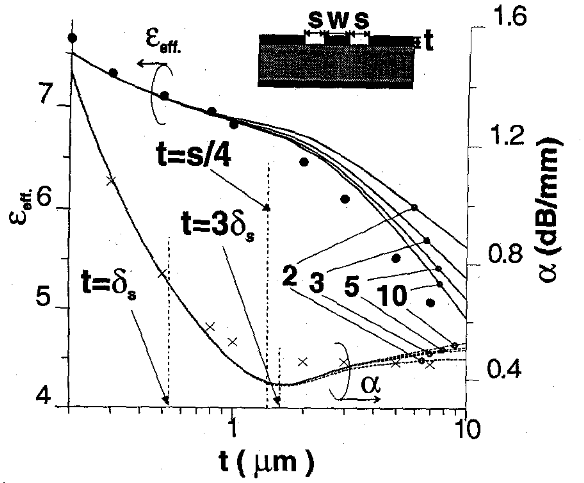


Fig. 5. Attenuation and effective dielectric constants for a coplanar waveguide on a AsGa substrate. $h = 200\ \mu$m, $s = 5\ \mu$m, $w = 40\ \mu$m, $\epsilon_r = 12.9$. (——) Phase constant results using a two-plates model and an $N = 3$, 5 and 10 improvement (17). (× •) Theoretical results in [15].

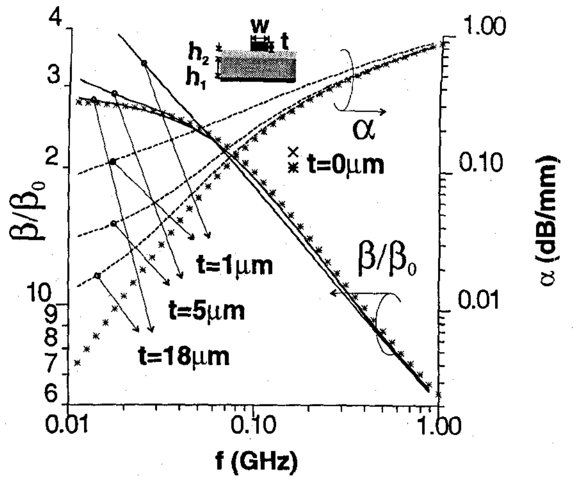


Fig. 6. Slow wave factor (——) and attenuation constant (– – –) for a microstrip line of gold ($\sigma = 4.094 \times 10^7\ (\Omega\text{m})^{-1}$) on a ferrimagnetic semiconductor. $h_1 = 100\ \mu$m, $h_2 = 1\ \mu$m, $w = 200\ \mu$m, $\epsilon_{r,1} = \epsilon_{r,2} = 15.5$. Layer 1 of saturated ferrimagnetic semiconductor with $\sigma_1 = 0.005\ (\Omega\text{mm})^{-1}$, $H_{0,1} = 2500$ Oe, $\Delta H_{0,1} = 75$ Oe.(*) Results in [2] with a lossless $t = 0$ strip.

and the presence of relatively high values for the metallization thickness. This latter limitation is expected to be more important in coplanar lines with high lateral coupling between the strips. In Fig. 5 our results for a coplanar waveguide are compared with those obtained in [15] by using a full-wave mode matching approach. Good agreement is observed for the attenuation constant for all the computed strip thicknesses. The results for the effective dielectric constant show a good agreement until values of the t/s ratio of about $t/s = 0.25$. For higher values of this ratio, our two-plates model results disagree. By using the improvement of the model given in (17), a much better convergence is achieved, although the effective dielectric constant is systematically overestimated.

One of the main advantages of the spectral domain technique is the ability to deal with multilayered and/or anisotropic substrates. In Fig. 6 the dispersion characteristics of a MIS line with a magnetic semiconductor layer are shown. At low frequencies the dispersion characteristics of the structure with ideal conductors can be obtained from a quasi-TEM approach [2], [3]. When conductor losses are also present, the proposed approach can be used to take into account both conductor and substrate magnetic and ohmic losses in a unified way.

IV. Conclusion

The surface impedance approach has been applied to the spectral domain quasi-TEM analysis of quasi-planar MIC and MMIC lines on dielectric, semiconductor and/or magnetic multilayered substrates having transverse anisotropy (i.e. the direction of propagation coincides with a main axis of the permittivity and permeability tensors). The proposed procedure takes advantage of the simplicity of both the quasi-TEM approach and the spectral domain analysis. The proposed method of analysis remains valid in both the weak and the strong skin effect frequency ranges.

Two surface impedance models have been used, a single-plate model [5] and a two-plates model. The single-plate model has shown to be useful for attenuation constant computations, provided that a good estimation of the ratio between the tangential magnetic field above and below the strips is made. The two-plates model also provides good results for the phase constant when thick strips are present. Therefore the two-plates model is useful in the analysis of MIC and MMIC lines with relatively thick strips. Moreover, no additional hypothesis on the magnetic field values at the neighbourhood of the strips must be made when aplying the two-plates model. This fact makes the two-plates model very useful in the analysis of multiconductor lines, in which a previous estimation of the magnetic field behavior near the strips is difficult or impossible. The use of asymptotic tails for the Green's function and other analytical preprocessing [13], makes it possible to develop computer codes capable of solving a great variety of configurations within short computing times. Typical computing times are within a few seconds in a 486 PC.

Appendix A

The surface impedance matrix appearing in (9) can be obtained solving the diffusion equation for the longitudinal component of the electric field inside the conductor. We assume that the conductor width is larger than conductor thickness, at least large enough to consider $\partial E_z^2/\partial x^2$ negligible with respect to $\partial E_z^2/\partial y^2$. Under that assumption, the expression for the impedance matrix can be found in [11], [9] and [12]

$$Z_{1,1} = Z_{2,2} = Z_m \operatorname{ctgh}\left(\frac{1+j}{\delta}t\right) \tag{18}$$

$$Z_{1,2} = Z_{2,1} = Z_m \frac{1}{\sinh(\frac{1+j}{\delta}t)}. \tag{19}$$

For strips made of several conducting layers, similar expressions can be obtained by following the same procedure. This makes possible to take into account the adhesive conducting layer (Ti) used in the technology.

APPENDIX B

The surface impedance used to improve (11) in [5] is

$$Z = Z_m\left(\frac{1+R_m}{1-R_m}\,\mathrm{ctgh}\left(\frac{1+j}{\delta}t\right) - \frac{R_m}{1-R_m}\,\mathrm{ctgh}\left(\frac{1+j}{2\delta}t\right)\right) \tag{20}$$

where $R_m = H_x^+/H_x^-$ is the ratio between the tangential magnetic fields on the upper and lower strip interfaces.

For an open microstrip line this ratio will be approximated by

$$\frac{H_x^+}{H_x^-} = -\left[1+\frac{1}{\pi}\,\mathrm{tg}^{-1}\left(\frac{2hw}{4h^2-(w/2)^2}\right)\right] \cdot \left[1-\frac{1}{\pi}\,\mathrm{tg}^{-1}\left(\frac{(2h+t)w}{(2h+t)^2-(w/2)^2}\right)\right]^{-1} \tag{21}$$

where h is the height of the substrate and w, t are the strip width and thickness respectively. Equation (21) is obtained by solving the magnetostatic problem of a uniform surface current layer of width w located at a distance h of an infinite ground plane. The H_x^+/H_x^- ratio in (21) is the ratio between the magnetostatic field at a distance t above the surface current layer, and the magnetostatic field just below the surface current layer. Notice that the surface impedance definition used in [8] is obtained from (20) and (21) by making $H_x^+ = 0$, i.e. by neglecting the magnetic field at the upper strip interface.

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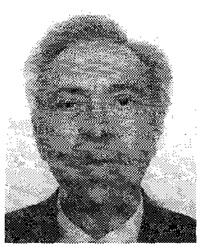

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