On the Quasi-TEM and Full-Wave Approaches Applied to Coplanar Multistrip on Lossy Dielectric Layered Media

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Abstract—**The characteristic parameters of coplanar multistrip lines embedded in multilayered lossless/lossy substrates are commonly computed by using either quasi-TEM or full-wave models. Several methods are provided in the literature to deal with this type of structure. In this paper a comparative study of quasi-TEM and rigorous solutions is carried out in order to establish criteria for the validity of the quasi-TEM approach. Reliable quasi-TEM and full-wave numerical data have been properly generated by applying an enhanced spectral domain analysis. We conclude that the quasi-TEM model yields satisfactory results for many MIC and MMIC practical cases. However, significant errors arise when high conductivity substrates are involved in MMIC applications. A discussion about the computation of the dynamic modal characteristic impedance is also reported, showing how the insertion of the modal orthogonality can save computational effort in a lossy multiport system.**

I. Introduction

It is well known that microstrip-like multilayered transmission lines play a vital role in MIC, MMIC and high speed VLSI technologies [1], [2]. A great deal of technical literature has been devoted to this subject. Conductor and substrate losses have been either neglected or analysed by means of perturbational techniques in many of these works. Nevertheless, new demands of the microwave technology (i.e., the joint packing of passive and active devices and the reduction in the size of the circuits) and the modeling of lossy interconnects, require on the one hand the extensive use of lossy dielectric substrates such as semiconductors and on the other hand to take into account the thickness and the losses of the metallizations. Conductor losses have been treated in [3]–[4] and references therein for MIC or MMIC structures. Since [3] discussed the validity of the quasi-TEM model in case of conductor losses, we are now concerned with the analysis of planar transmission lines just including substrate losses. This type of line, see Fig. 1, consists of N layers with lossless/lossy dielectric substrates, N_c zero thickness ideal metal strips and one or two ground planes. Semiconductors, the most common lossy substrates, show a large conductivity giving rise to relevant substrate losses. This fact precludes the use of perturbational techniques in the analysis of these problems. Substrate losses must be therefore explicitly considered in the formulation of the problem. This has been already carried out in the quasi-TEM [5], [6] and the full-wave approaches [7], [8] including the lossy nature of the substrate by means of the imaginary part of the complex dielectric permittivity.

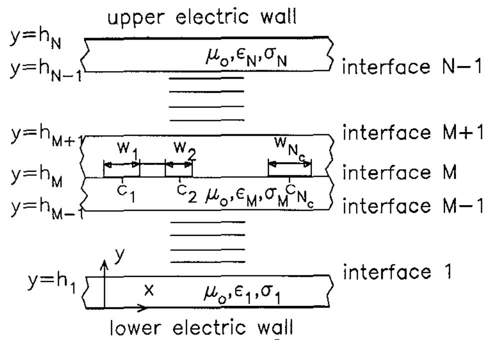

Fig. 1. Transverse section of a multilayered, multiconductor coplanar transmission line.

To our knowledge, the limits of validity of the simplified quasi-TEM model have not been so clearly established for lossy structures as much as for low losses MIC ones. A previous knowledge of this validity range is extremely useful in order to avoid the use of the much more involved full-wave analysis when the simpler quasi-TEM model is sufficient. In the low-loss case, a wavelength significantly larger than the cross dimensions is required to assure the validity of the quasi-TEM model, in such a way that quasi-TEM model is equivalent to quasi-static approach. However, when arbitrary lossy substrates are involved, the quasi-TEM concept should be revised. Thus, the operation frequency and cross-sectional dimensions are not the sole parameters to be considered. For instance, the quasi-TEM model fails even at very low frequencies if high conductivity substrates are present (even with cross sectional dimensions as small as those involved in MMIC technology). Moreover, the range of validity is different for different parameters (phase constant, atten-

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uation factor, characteristic impedance). Thus, the main purpose of the present paper is to accomplish a comparative analysis (quasi-TEM versus full-wave) regarding the different characteristic parameters of the lossy multistrip structure, namely phase constants, attenuation factors and complex modal characteristic impedances. The definition of the modal characteristic impedances is also discussed for the multiconductor lossy case. It is shown how it is possible to obtain the modal characteristic impedances without computing the cross powers defined in [9] when the full-wave model is used.

Prior to carrying out the comparative analysis, we have developed computer codes based on spectral domain formulations of the quasi-TEM and full-wave problems. Although the details of the implementation of these programs are out of the scope of this paper, it should be pointed out that the accuracy and reliability of the numerical results has been conveniently checked. The numerical convergence has been also significantly enhanced by using various analytical procedures [5], [10]. These numerical schemes are specially convenient when very thin layers are involved (i.e., metal-insulator-semiconductor (MIS) structures).

II. On the Validity of the Quasi-TEM Model

In this section, we briefly show how to determine the validity range of the quasi-TEM model quantitatively. The structure to be considered is shown in Fig. 1. The lossy nature of the (i)-th layer is taken into account by means of a complex dielectric permittivity, i.e., $\epsilon_i = \epsilon_0 \epsilon_{ri}(1 - j \tan \delta_i)$, with $\tan \delta_i = \sigma_i/(\omega \epsilon_o \epsilon_{ri})$ (a complex formulation of the problem is then required). This consideration means that lossy substrates are assumed to be dispersive and makes quasi-TEM model accounts directly for certain dispersion arising from the lossy nature of the substrates.

The quasi-TEM model is based on the assumption that the electromagnetic fields propagating along the transmission line are essentially transverse. This condition refers to the spatial average of the field over the cross section rather than to the values of the fields at each point, that is

$$\langle |\boldsymbol{E}_t| \rangle \gg \langle |E_z| \rangle \quad \text{and} \quad \langle |\boldsymbol{H}_t| \rangle \gg \langle |H_z| \rangle, \tag{1}$$

where $\langle \bullet \rangle$ stands for the spatial average of the argument.

From a dimensional analysis applied to the Maxwell's equations, the condition (1) can be reformulated, following [5], as

$$d \ll \frac{1}{\omega \sqrt{\mu_0 \langle |\epsilon(\boldsymbol{r}, \omega)| \rangle}}, \tag{2}$$

ω being the angular frequency, $\langle |\epsilon(\boldsymbol{r}, \omega)| \rangle$ the average value of the module of the complex permittivity (ω dependent) of the multilayered medium and d the largest transverse distance between the conductors. In a two-conductor line ($N_c = 1$) d would be the vertical distance from the conductor to the ground plane and in a multiconductor line ($N_c > 1$), d would be the distance between the most distant conductors. Expression (2) establishes that the quasi-TEM assumption can be considered correct provided the distance between the conductors in the transmission line is much smaller than the *average wavelength* of the propagating fields. This condition is more stringent than requiring transversal dimensions to be much smaller than the free space wavelength. Note that the denominator in the second member of the inequality can be considered as the propagation constant of the fundamental mode in a structure such as that in Fig. 1 when the internal metallized interface is removed and the layered medium is replaced by an equivalent medium with permittivity $\langle |\epsilon(\boldsymbol{r}, \omega)| \rangle$. As an example, for the practical MIS configuration, expression (2) can be rewritten (after simple algebraic manipulations), requiring that d should be a magnitude order minor than the *average wavelength* as d(in mm) $< d_{\max}$, being

$$d_{\max} = \frac{15}{f\sqrt{\epsilon_{rs}}} \left[1 + 3.24 \cdot 10^8 \left(\frac{\sigma}{f\epsilon_{rs}} \right)^2 \right]^{-1/4}, \tag{3}$$

ϵ_{rs}, σ the values of the relative permittivity and conductivity (in $(\Omega\text{mm.})^{-1}$) of the semiconductor layer, and f the operation frequency in GHz. We have checked that the quasi-TEM model yields accurate data for the phase constant (relative errors smaller than 1% with respect to rigorous full-wave analysis) when condition (3) is fulfilled. Nevertheless, the model drastically falls down in the computation of the attenuation constant when high conductivities (approximately $\sigma > 1(\Omega\text{mm.})^{-1}$) are involved, even though condition (3) is already fulfilled. These facts can be explained from physical arguments. The quasi-TEM model only implies that the *average transverse* fields are much stronger than the *average longitudinal* ones, namely the electromagnetic field distribution is essentially transversal. Since the field distribution determines the phase constant and modal impedances, these parameters should be accurately computed in the quasi-TEM frame whenever (2) is fulfilled. This is also true for any parameter depending on the field distribution (for example, the current eigenvectors). On the contrary, the quasi-TEM model does not account for the part of the Joule effect losses in the semiconductor layer due to the axial current. This effect arises from the existence of relatively important longitudinal currents produced by a weak longitudinal electric field—the average electromagnetic field still satisfying (1)—when very lossy substrates are present. In this case, the quasi-TEM value of the attenuation constant—which basically depends on the total Joule losses—is not realistic. A full-wave analysis is then unavoidable to compute the attenuation factor if σ is large enough even though frequency is low and cross dimensions are small.

III. Quasi-TEM and Full-Wave Spectral Domain Analysis

Since the particular method used to generate the numerical data is not the subject of the present work, we restrict ourselves to sketch a brief outline.

The quasi-TEM propagation parameters are computed from the complex capacitance, $[C]$, and vacuum capacitance, $[C_o]$, matrices per unit length (p.u.l.) of the transmission line system. In this work, we have applied the Galerkin method in the spectral domain to compute the elements of those matrices. Theoretical and numerical details on this method are reported in [5]. In that paper, a very general multilayered multiconductor system was treated.

The spectral domain analysis has also been employed to compute the full-wave results. Thus, the spectral Green's dyad of a layered configuration can be obtained by following either the *transverse propagation matrix* technique reported in [11] or the EBM method proposed in [12]. The Galerkin method is also used to solve the resultant eigenvalue problem, that is

$$[A(\omega, \beta - j\alpha)] \cdot c = 0, \tag{4}$$

with $[A]$ being the Galerkin matrix and c the vector whose elements are the different coefficients of the current density expansion.

The full-wave numerical scheme has been recently improved to be efficiently applied to MMIC structures including highly lossy substrates and thin layers [10]. Modal propagation constants and field distributions are found by solving (4). This equation has been solved by adapting the search zeroes method proposed in [13]. As it will be discussed in the following section, modal powers need to be computed. These modal powers are obtained by means of the scheme proposed in [10], [14] which makes it possible to handle systematically arbitrary multilayered and multiconductor configurations. It should be noted here that the total *complex power* must be used rather than the real part of it [9].

Particular attention was paid to all the numerical aspects of the implementation of the methods in order to ensure accuracy and numerical efficiency. The use of unsuitable or too few basis functions as well as inaccurate computation of the spectral integrals can result in serious numerical errors, specially when high conductivity and/or very thin layers are involved.

IV. Modal Characteristic Impedances and Voltage Current Eigenvectors

In the quasi-TEM frame, the complex modal propagation constants, $\gamma_n = \beta_n - j\alpha_n$ (arising from the complex nature of the capacitance matrix), and the modal voltage, V_n, and current, I_n, eigenvectors of a $N_c + 1$ conductor system are defined by the following wave equations:

$$\left.\begin{aligned}\{\omega^2[L] \cdot [C] - \gamma_n^2\} V_n = 0\\ \{\omega^2[C] \cdot [L] - \gamma_n^2\} I_n = 0\end{aligned}\right\} (n = 1, \cdots, N_c). \tag{5}$$

with $[C]$, $[L]$ being the $N_c \times N_c$ capacitance and inductance matrices p.u.l. of the multiconductor system and ω the angular frequency. The current and voltage complex eigenvectors satisfy two important rules:

$$V_n \cdot I_m = 0 \qquad \text{for } m \neq n, \gamma_n \neq \gamma_m. \tag{6}$$

$$V_n \cdot I_n^* = P_n. \tag{7}$$

The so-called biorthogonality condition (6) directly arises from the eigenvector theory when applied to (5) [15]. Physically speaking, (6) stands for the reciprocity properties of the multiconductor system [16]. The second condition (7)—namely the power condition—stands for the usual complex power-voltage-current relationship in transmission line theory.

Since eigenvalues and eigenvectors are uniquely defined from (5), modal characteristic impedances are defined from voltage and current eigenvectors in the following way:

$$Z_n^j = \frac{V_n^j}{I_n^j} \tag{8}$$

where n stands for the mode and j for the conductor, i.e., V_n^j, I_n^j are the jth entries of the V_n and I_n column vectors in (5). Consequently, propagation constants, current and voltage eigenvectors and impedance parameters—required in a circuital description of the system—are directly defined in this frame. All the above quantities are readily computed from $[C]$ and $[L]$ matrices.

When a full-wave model is used, a circuital description of the lossy multiconductor system is also possible. In the following, our discussion is restricted to the quasi-TEM type fundamental modes rather than higher order modes which are assumed to be evanescent. Complex modal propagation constants are computed by solving the eigenvalue equation (4), as stated above, and therefore, these quantities are unambiguously defined. However, the imprecise definition of the remainder circuital parameters (current and voltage eigenvectors and dynamic impedances) has been subject of discussion in the literature, even in the simple two-conductor line case [16]–[19]. In this way, some comments should be pointed out here concerning the significance of V_n^j and I_n^j in the full-wave frame. Actually, certain ambiguity arises from the hybrid nature of propagating fields. As a result of this, the voltage, V, and the current, I, definitions are not unique. Nevertheless, in the two-conductor case, we know that V is related to a certain integral of the transverse electric field E_t and I to a certain integral of the transverse magnetic field H_t. According to this, we can write the following relations from the electromagnetic field standpoint:

$$\iint_{S_t} (E_t \times H_t^*) \cdot dS \leftrightarrow VI^* \tag{9}$$

$$\iint_{S_t} (E_t \times H_t) \cdot dS \leftrightarrow VI. \tag{10}$$

The V and I quantities have been obtained in the literature by three different ways for a single line: a) V and I are separately defined as integrals of the electric and magnetic fields along proper paths, b) I is defined as the total z-directed current and V is computed from I and z-power, $V = 2P/I$, and c) V is defined as a proper path integral of the electric field and I is computed from V and z-power, $I = 2P/V$. All of the three options lead to the same numerical results at the low frequency limit. They also agree with quasi-TEM values assuming this approximation to be valid. However, significant discrepancies between the above three definitions are found when frequency increases sufficiently. These differences stem from different choices of the *primary* or independent and the *derived* quantities [18]. From a practical point of view, the primary quantity (either current or voltage) is chosen to be the most accesible one. In microstrip-like structures, currents are primary and voltages are derived, as it is commonly accepted nowadays. So, we can now compute the voltage from the definition of complex power as follows:

$$P = \frac{1}{2} \int\int_{S_t} (\boldsymbol{E}_t \times \boldsymbol{H}_t^*) \cdot d\boldsymbol{S} = \frac{1}{2} VI^*, \tag{11}$$

S_t being the cross section of the structure.

The extension of the impedance concept to a multistrip system—in the full-wave frame—requires a little more discussion. Apart from the propagation constants, the only quantities uniquely obtained from a full-wave analysis of a multistrip system are the complex modal powers, P_n and the complex current eigenvectors $\boldsymbol{I}_n$. The definition of the complex modal power is given by:

$$P_n = \frac{1}{2} \int\int_{S_t} (\boldsymbol{E}_{t,n} \times \boldsymbol{H}_{t,n}^*) \cdot d\boldsymbol{S} = \frac{1}{2} \boldsymbol{V}_n \cdot \boldsymbol{I}_n^* \tag{12}$$

where we have identified the scalar product of the modal eigenvoltage, $\boldsymbol{V}_n$, and modal eigencurrent, $\boldsymbol{I}_n$, with the modal power in order to obtain a suitable coupled transmission line model (CTL) of the system. It should be noted that expression (12) does not completely define the modal eigenvoltage which is the opposite to what happens with (11) in the single microstrip case. Thus, we need more equations to determine all the eigenvoltages. The reciprocity theorem [20] can be used to complete the number of required equations, that is,

$$\boldsymbol{V}_n \cdot \boldsymbol{I}_m = \int\int_{S_t} (\boldsymbol{E}_{t,n} \times \boldsymbol{H}_{t,\mathrm{m}}) \cdot d\boldsymbol{S} = 0 \qquad \text{for } m \neq n. \tag{13}$$

If expressions (12) and (13) are combiend, we can write the following system of equations to obtain the eigenvoltages:

$$\left.\begin{aligned} P_k &= \tfrac{1}{2} \boldsymbol{V}_k \cdot \boldsymbol{I}_k^* \\ 0 &= \boldsymbol{V}_l \cdot \boldsymbol{I}_m (l \neq m) \end{aligned}\right\} (k, l, m = 1, \cdots, N_{\mathrm{modes}}). \tag{14}$$

Once $\boldsymbol{V}_n$ vectors are obtained from (14), the modal impedances are computed from (8).

The eigenvoltages could also be determined by computing the modal power matrix [9], defined by

$$[\boldsymbol{P}] = \tfrac{1}{2} [\boldsymbol{V}]^T \cdot [\boldsymbol{I}]^*, \tag{15}$$

$[\boldsymbol{V}]$ and $[\boldsymbol{I}]$ being the eigenvoltage and eigencurrent matrices (that is, matrices whose columns are the voltage, $\boldsymbol{V}_n$, and the current, $\boldsymbol{I}_n$, eigenvectors of the problem). Note that in the lossless case $[\boldsymbol{P}]$ is a real diagonal matrix whose entries are the power associated with each propagating mode thus resulting that equations (15) and (14) are exactly the same. However, if lossy materials are involved, the $[\boldsymbol{P}]$ matrix is no longer diagonal because the cross powers $P_{l,m}$ associated with the electric field of the lth mode and the magnetic field of the mth mode ($l \neq m$) are not equal to zero [9]. Hence in a multiconductor lossy system, the procedure exposed in this work causes considerable reduction of analytical and numerical effort ($P^{k,l}$, $k \neq l$ does not need to be computed) reducing consequently CPU time. The higher the number of coupled conductors is, the lesser CPU time required is.

Notice that, in the lossless case, the formulation in the present work reduces to that reported in [16]. In the lossy case, expression (4) in [16] should be replaced by (14), where the total complex modal power must be considered instead of its real part. The definition of the dynamic modal impedances discussed in this work has been incorporated in [10], where the analytical and computational aspects related to their evaluations are stressed.

When (14) is used, the equivalence between the full-wave impedances and the quasi-TEM ones is assured at the low frequency limit for both, lossless and arbitrarily *lossy* structures. Moreover, from a conceptual point of view, the connection between quasi-TEM and dynamic definitions is quite obvious.

Finally, it should be remembered here that other authors [19], [21] have used a slightly different definition of the impedance based on the concept of the partial power associated to each line for each mode. This latter definition is indistinguishable from (15) assuming the quasi-TEM approximation to be valid, but discrepancies can be otherwise detected, as it has been reported in [16].

V. Results

The numerical data included in this work have been obtained by using two computer codes (quasi-TEM and full-wave), which are able to deal with the generalized multilayered microstrip-like transmission lines considered in this work. Both programs have been systematically checked with previous results. In Figs. 2(a)–(c), we present our quasi-TEM and full-wave data [10] together with some experimental data (reported in [22]). Normalized wavelength (λ/λ_0), attenuation constant (α) and complex characteristic impedance (Z_0) of a simple MIS structure are shown for three values of semiconductor conductivity (σ). As it was predicted, a good agreement between full-wave and quasi-TEM results is found for λ/λ_0 at frequencies satisfying (3) (Fig. 2(a)). The quasi-TEM model provides accurate results for this quantity up to 18 GHz, 220 MHz and 30 MHz for σ values of 0.005, 1 and 10 $(\Omega \cdot \mathrm{mm})^{-1}$ respectively. However, we can see from Fig. 2(b) that, except for the lowest conductivity case, the attenuation constant should not be computed by using the simple quasi-TEM model. The great quantitative and

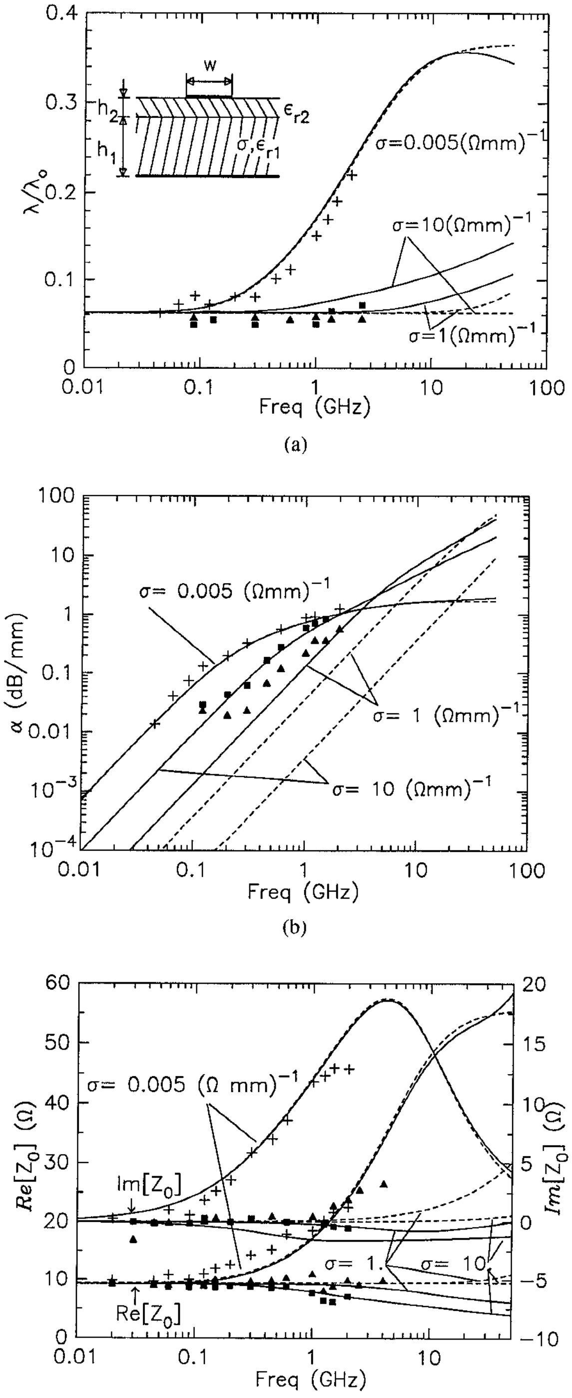

Fig. 2. (a) Normalized wavelength. (b) Attenuation constant. (c) Complex characteristic impedance for a simple MIS configuration (SiO_2 on Si). $h_1 = 250$ μm., $h_2 = 1$ μm., $w = 160$ μm., $\epsilon_{r1} = 12$, $\epsilon_{r2} = 4$. (+, △, ●): Experimental values of [22], (———): Full-wave data (-------): Quasi-TEM data.

qualitative discrepancies existing for these high conductivity values make the quasi-TEM model unsuitable to determine α *at any operation frequency*. The real and imaginary parts of Z_0, plotted in Fig. 2(c), show a behavior similar to that observed for λ/λ_0 in Fig. 2(a). At this point, we can therefore conclude that phase constant and characteristic impedance can be accurately computed by

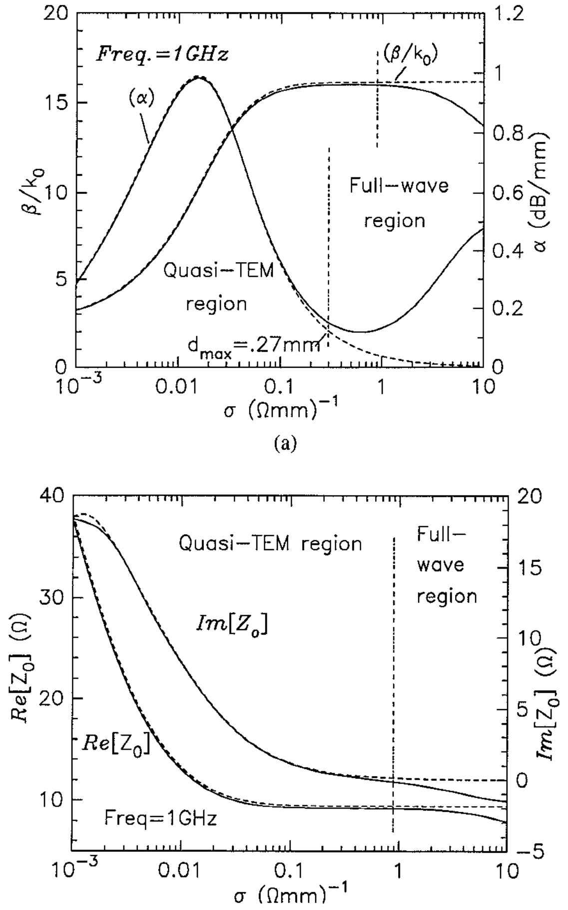

Fig. 3. (a) Slow-wave factor and attenuation constant. (b) Complex characteristic impedance versus conductivity for the configuration of Fig. 2. (———): Full-wave data, (-------): Quasi-TEM data.

using the quasi-TEM approach as long as condition (2) is fulfilled. However, the attenuation factor can not be predicted by means of a quasi-TEM analysis is σ is too high. This fact is ratified by the curves of Fig. 3(a)–(b). They show the behavior of the slow-wave factor (β/k_0), (α) and Z_0 versus σ at a fixed frequency value (1 GHz) for the same structure. We can note, in particular, that quasi-TEM and full-wave data show a relative difference (concerning to the quasi-TEM value) of 1% and 50% for β/k_0 and α respectively if $\sigma = 0.6(\Omega \cdot \text{mm})^{-1}$. These discrepancies become 10% for β/k_0 and 6700% for α when $\sigma = 6(\Omega \cdot \text{mm})^{-1}$. Thus, quasi-TEM values for β and Z_0 are accurate enough in the slow-wave and dielectric mode regions [23] but not in the skin-effect region. On the other hand, quasi-TEM values of α are not valid even in the slow-wave mode region if σ is high enough. The theoretical explanation of all these facts was pointed out in Section III.

Next, we compare the results of the two models when applied to the analysis of a multistrip transmission line. In particular, we choose a three strip structure printed on a three layer composite medium (semi-insulator, semiconductor and insulator). Firstly, we hope to verify the validity range of the quasi-TEM model in this type of

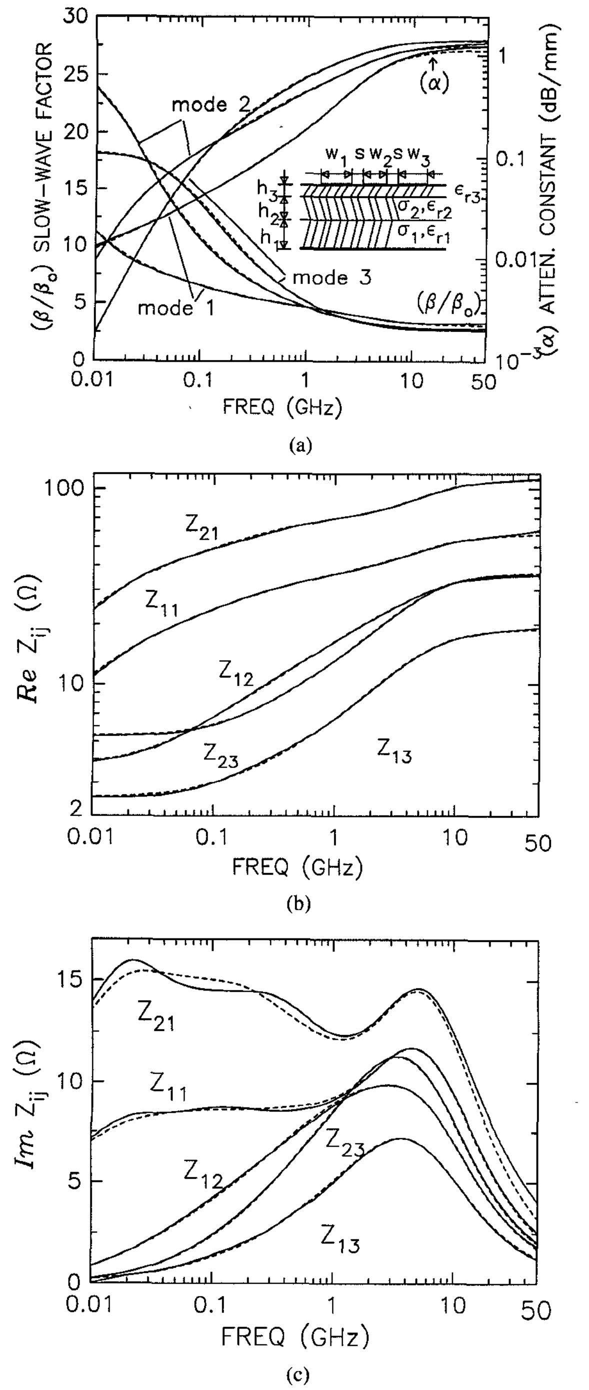

Fig. 4. Modal parameter. (a) Slow-wave factor and attenuation constant. (b) Real part of impedances. (c) Imaginary part of impedances, of a three-conductor MIS configuration. $h_1 = h_2 = 100$ μm. $h_3 = 1$ μm., $\epsilon_{r1} = \epsilon_{r2} = \epsilon_{r3} = 12.9$, $\sigma_1 = 10^{-5}(\Omega\text{mm})^{-1}$ $\sigma_2 = 5 \cdot 10^{-3}(\Omega\text{mm})^{-1}$, $w_1 = w_3 = 200$ μm $w_2 = 100$ μm., $s_1 = s_2 = 50$ μm. (———): Full-wave data, (-------): Quasi-TEM data.

structure and, secondly, we are also concerned about checking that the procedure followed to compute the full-wave modal impedances is compatible with the quasi-TEM one. Thus, the conductivities are choosen to make longitudinal Joule effect losses negligible for the structure analyzed in Fig. 4(a)–(c). The quasi-TEM and full-wave data plotted in Fig. 4(a) show very good agreement for the three modal slow-wave factors and attenuation constants for frequencies below 15 GHz, that is, when condition (3) is fulfilled. Regarding the characteristic modal impedances, we can see from Fig. 4(b) very satisfactory agreement for the real and imaginary parts.

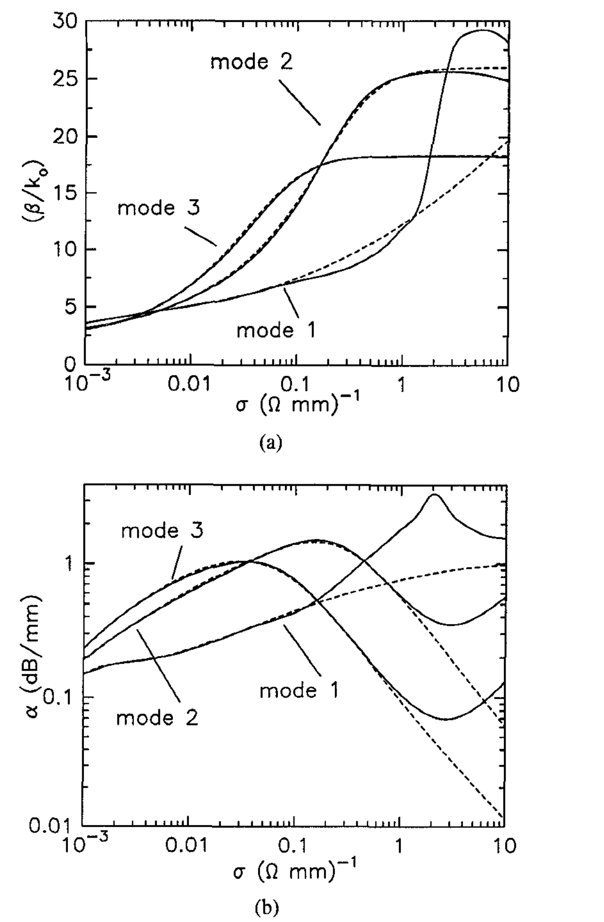

Fig. 5. (a) Modal slow-wave factors. (b) Modal attenuation constants versus conductivity for the configuration of Fig. 4 with freq = 1 GHz. (———): Full-wave data, (-------): Quasi-TEM data.

Figs. 5(a)–(b) show the quasi-TEM and full-wave predictions for β / k_0 and α as a function of the semiconductor conductivity for the three-strip structure. We find that by applying (2) to this configuration that $\sigma = 0.08(\Omega\text{mm})^{-1}$ causes $d_{\max}$ to be 400 μm (the largest distance between the center of the strips). Therefore, this conductivity value shows the upper limit in which the quasi-TEM approximation can be used properly. This fact is clearly reflected in the curves of Fig. 5(a)–(b) where we can see that this upper limit has its main effect on mode #1. The quasi-TEM results for the other two modes are valid at a slightly higher conductivity value. This can be explained from the field distribution associated with these modes. Modes #2 and #3 have a higher concentration of electric field in the lossless layer—the sign of the components of eigencurrents are (+, 0, −) for mode #2 and (+, −, +) for mode #3—than mode #1—(+, +, +)—. Anyway, the quasi-TEM model should be used only if the all three modes are correctly described in this frame. So, the condition in (2) can not be considered excessively conservative for multiconductor lines.

V. Conclusions

In this paper we have carried out a comparative analysis in order to establish the conditions under which a quasi-TEM model can be used in the study of the fundamental propagating modes in planar microstrip lines currently used in (M)MIC and high speed digital technologies. The operation frequency, cross sectional dimensions and conductivities are the main parameters to be taken into account. Phase constants and characteristic impedances are adequately predicted under quasi-TEM assumption whereas condition (3) is fulfilled. Attenuation constants can be also computed by using the quasi-TEM model in many practical cases involving lossy semiconductor substrates. However, high conductivity substrates precludes the use of the quasi-TEM approach for attenuation factor computations even if (3) is fulfilled. A full-wave model must be used in this case.

The dynamic modal characteristic impedances of a lossless/lossy multiconductor structure are obtained from primary complex mode parameters: current eigenvectors and modal powers. Voltage eigenvectors are defined by imposing two conditions: the bi-orthogonality of voltage and current eigenvectors and the usual complex power/voltage/current relationship for each propagating mode. These conditions ensure compatibility with the transmission line theory, preserve fundamental reciprocity requirements of the multiport system and match quasi-TEM results in its validity range. In case lossy substrates are involved, this general procedure is especially advantageous because avoids the computation of cross-powers (reported in [9]) involving different modes.

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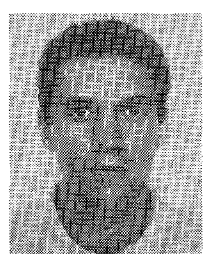

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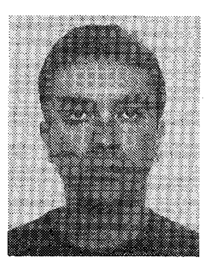

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