Conductor Loss Computation in Multiconductor MIC's by Transverse Resonance Technique and Modified Perturbational Method

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Abstract—**Rigorous computation of conductor loss in MMIC's transmission lines requires high computer expenditures, while conventional approaches become invalid for thin line conductors. Using a modified perturbational method, originally proposed by Horton *et al.*, in conjunction with the generalized transverse resonance technique, very accurate results are obtained with relatively modest computer effort.**

I. Introduction

EVALUATION of loss is of paramount importance for the accurate modeling of MMIC's. Dielectric loss can usually be evaluated by just assuming a complex dielectric permittivity. The accurate computation of conductor loss, on the contrary, requires a considerable effort, particularly when metal thickness t is of the order or less than the skin depth δ. In such cases, the conventional perturbational approach leads to a significant underestimation of the conductor loss. This is the case, for instance, of GaAs FET structures, which have received a considerable interest for the possible realization of distributed amplification [2]. Though feasible, the rigorous analysis of the propagation characteristics of MMIC's structures including conductor loss involves a formidable computational effort [3], [4].

In this letter, the generalized transverse resonance technique [5] is applied to compute the propagation characteristics of multiconductor quasiplanar transmission lines. To accurately compute the conductor loss without excessive computer effort, the conventional perturbational approach is modified using either a modified surface impedance (model 1) or a transmission line model for the metal layer (model 2). The latter model is coincident with that already proposed by Horton *et al.* [1]. It is found that model 1 correctly predicts the loss increase for $t/\delta < 1$, although quantitatively too small. Model 2, on the contrary, is found to give results in very close agreement with rigorous theories. With both models, however, the computational expenditure is reduced to only that for the lossless case.

Manuscript received January 22, 1992.

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IEEE Log Number 9200892.

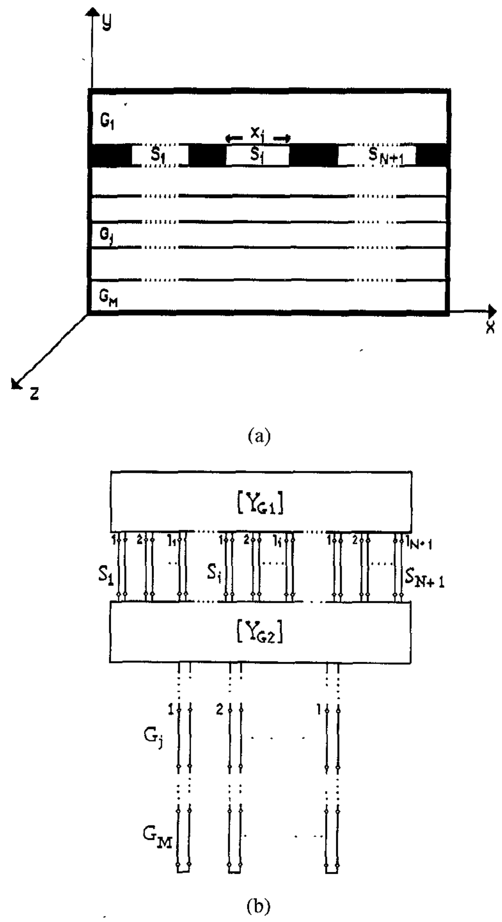

Fig. 1. (a) Schematic of a multiconductor multilayered quasiplanar transmission line; (b) equivalent transverse circuit of the structure.

II. Method Of Analysis

Fig. 1(a) shows the schematic of a multiconductor quasiplanar transmission line realized on a multilayered substrate. N metal strips, plus, possibly, two grounded fins, are deposited on top of the substrate, which consists of M lossy layers. The transmission line is enclosed in a waveguide housing. All metal strips are supposed to have a nonzero thickness t. The presence of holding grooves in the waveguide housing is neglected for simplicity, though it could be taken into account without too much a complication [6].

The method adopted is the generalized transverse resonance technique described in [5], [6]. Such a method is equivalent to a mode-matching technique applied in the transverse y-direction. In each homogeneous region of Fig. 1(a) the EM fields are expanded in terms of $TE_{(y)}$ and $TM_{(y)}$ modes. With the adoption of the microwave network formalism in conjunction with the admittance matrix representation, each region can be modeled as a generalized multiport network, the admittance matrix being computed without any matrix inversion [7]. In practice, applying the so-called transverse segmentation [8] to Fig. 1(a), only two generalized Y-matrices have to be computed, all the other regions being represented by a mere set of transmission lines representing the y-propagating (or evanescent) modes (Fig. 1(b)). In this manner, a substantial reduction in computing time is achieved.

To account for power lost in the conductors, the conventional perturbation technique can be applied provided the conductor thickness t is much larger than the skin depth δ. In this approach, one assumes that the tangential electric field $\boldsymbol{E}_\tau$ at the surface of an imperfect conductor is related to the tangential magnetic field $\boldsymbol{H}_\tau$ by

$$\boldsymbol{E}_\tau = Z_s \boldsymbol{H}_\tau \times \boldsymbol{n}, \quad (1)$$

$\boldsymbol{n}$ being the normal directed unit vector, Z_s equals the intrinsic impedance of the metal

$$Z_s = Z_c = (1+j)\sqrt{\frac{\omega\mu}{2\sigma}}. \quad (2)$$

The power lost in the conductor is then computed by the flow of the Poynting's vector entering the metal.

For thin conductors $t \cong \delta$, these assumptions lead to an underestimation of the conductor loss. Expression (2) is not valid any more, as the EM field penetrates deeply into the conductor reaching the opposite surface with finite amplitude. We can account for this phenomenon, in a first approximation, by still assuming the validity of (1) with unperturbed magnetic field at the metal surface, but replacing the impedance (2) seen at one side of the metal by the input impedance of a terminated lossy transmission line. In formulas

$$Z_s = Z_c \frac{Z_l + Z_c \tanh[k_c t]}{Z_c + Z_l \tanh[k_c t]} \quad (3)$$

with

$$k_c = (1+j)\sqrt{\pi \mu f \sigma} \quad (4)$$

being the complex propagation constant of a plane wave within the metal and $Z_l = \sqrt{\mu/\varepsilon}$ being the intrinsic impedance of the dielectric at the opposite side of the metal surface.

The previous expression constitutes a very simple modification of the conventional perturbation technique, requiring negligible additional computation with respect to the lossless case. Loss values computed through (3), though much better approximated than the conventional ones, are still underestimated compared to the rigorous computations by Heinrich [4].

The electric field at one metal surface, in fact, does not depend on the magnetic field at the same surface only, but, because of the magnetic field penetration in the metal, it depends also on the magnetic field at the opposite surface [9]. Accordingly, (1) is replaced by

$$\boldsymbol{E}_{\tau 1} = Z_c \frac{\boldsymbol{H}_{\tau 1} \cosh[k_c t] - \boldsymbol{H}_{\tau 2}}{\sinh[k_c t]} \times \boldsymbol{n} \quad (5)$$

and similarly for $E_{\tau 2}$. The previous expression, equivalent to that originally proposed by Horton et al. [1], can then be used to compute the power flow into the metal, thus the conductor loss.

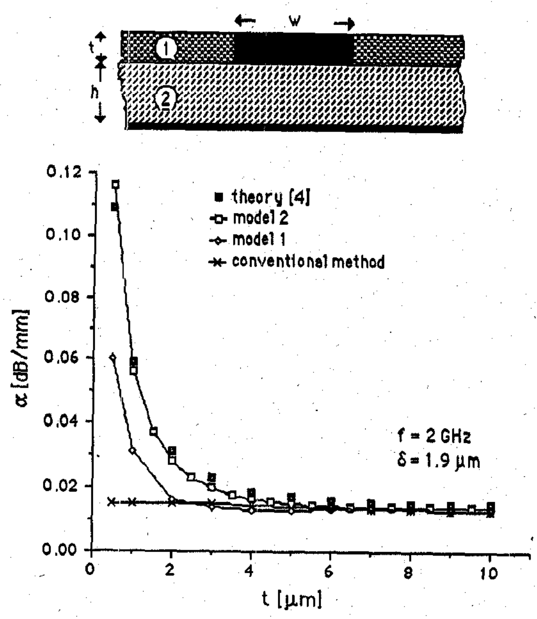

Fig. 2. Attenuation α of a MMIC microstrip line as a function of metallization thickness t. Line geometry: w = 30 μm, h = 200 μm. Material parameters of subregion are (1) dielectric layer: ϵ_r = 3.4, tanδ = 0.05; (2) GaAs: ϵ_r = 12.9, tanδ = 3 × 10^{-4}. Metallization conductivity σ = 3.333 × 10^7 $[\Omega \text{m}]^{-1}$.

III. Results

Besides the conventional perturbation technique, expressed basically by (1) and (2), two additional approximations for the conductor loss computation have been derived in the previous section. The simplest one (model 1), still based on the concept of surface impedance, simply consists of replacing the surface impedance (1) with (3). The other approach (model 2), still a perturbational one, evaluates the surface E-field in terms of the unperturbed H-fields on both sides of the metal. This approach is based on the use of (5). We have checked these approaches against the full-wave technique by Heinrich [4].

Fig. 2 shows the computed conductor loss attenuation in a microstrip line as a function of the conductor thickness. All theories lead to about the same attenuation for thicknesses of the order or larger than 3δ. The conventional method fails to predict the loss increase for smaller thicknesses. Model 1 does predict a loss increase for $t/\delta < 1$, though quantitatively too small. Model 2, on the contrary, is seen to provide results in very close agreement with [4]. Similar results are obtained for a boxed coplanar waveguide (Fig. 3). The conventional method is incorrect for thin conductors, while both models 1 and 2 lead to a sharp increase of conductor loss attenuation for $t/\delta < 1$. A small disagreement between model 2 and [4] is observed for extremely thin conductors ($t \cong 0.2$ μm). Finally,

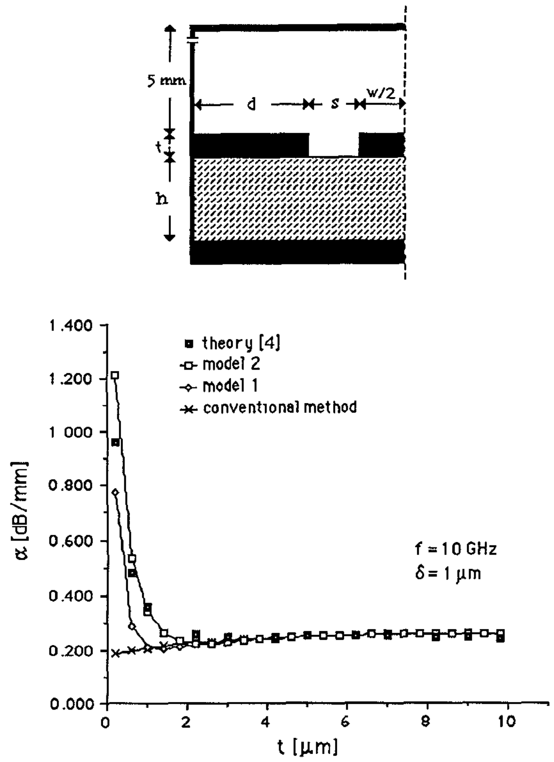

Fig. 3. Attenuation α of a CPW as a function of metallization thickness t. Line geometry: w = 40 μm, d = 50 μm, s = 5 μm, h = 600 μm. Substrate data: GaAs: ϵ_r = 12.9, tanδ =3 $\times$ 10^{-4}. Metallization conductivity σ = 3 $\times$ 10^7 $[\Omega m]^{-1}$.

it is worth mentioning that an excellent agreement has been found with the experimental results by Haydl *et al.* [10] on a CPW.

IV. Conclusion

The generalized transverse resonance technique has been applied to compute the propagation characteristics of multiconductor quasiplanar lines for application to MMIC's. Conductor loss is evaluated using a modified perturbational approach, originally proposed by [1], which yields very accurate results with negligible computational effort compared to rigorous full wave methods.

Acknowledgment

The authors wish to thanks the third reviewer for drawing their attention to [1].

References

[1] R. Horton, B. Easter, and A. Gopinath, "Variation of microstrip losses with thickness of strip," *Electron. Lett.*, vol. 7, no. 17, pp. 490–491, Aug. 1971.
[2] N. Sebati, P. Gamand, C. Varin, F. Pasqualini, and J. C. Meunier, "Continous active *T*-gate travelling-wave transistor," *Electron. Lett.*, vol. 25, pp. 403–404, Mar. 1989.
[3] C.-D. Chen, C.-K. C. Tzuang, and S. T. Peng, "Skin-effect current distribution of a unilateral finline with finite conductivity." in *IEEE Int. Microwave Symp. Dig.*, 1991, pp. 773–776.
[4] W. Heinrich, "Full-wave analysis of conductor losses," *IEEE Trans. Microwave Theory Tech.*, vol. 38, pp. 1468–1472, Oct. 1990.
[5] R. Sorrentino, "Transverse resonance technique," in *Numerical Techniques for Microwave and Millimeter Wave Passive Structures*, T. Itoh, Ed. New York: John Wiley, 1989, ch. 11, pp. 637–696.
[6] F. Alessandri, U. Goebel, F. Melai, and R. Sorrentino, "Theoretical and experimental characterization of nonsymmetrically shielded coplanar waveguides for millimeter-wave circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 37, pp. 2020–2027, Dec. 1989.
[7] F. Alessandri, G. Bartolucci, and R. Sorrentino, "Admittance-matrix formulation of waveguide discontinuity problems. Applications to the design of branch-guide couplers," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 394–403, Feb. 1988.
[8] F. Alessandri, M. Mongiardo, and R. Sorrentino, "Transverse segmentation: A novel technique for the efficient CAD of 2*N* -port branch-guide couplers," *IEEE Microwave Guided Wave Lett.*, vol. 1, pp. 204–207, Aug. 1991.
[9] T. Kitazawa, "Extended spectral domain approach," *Dig. MWE*, Tokyo, Japan, Sept. 1991, pp. 221–226.
[10] W. H. Haydl, T. Kitazawa, J. Braunstein, R. Bosch and M. Schlechtweg, "Millimeter-wave coplanar transmission lines on gallium arsenide, indium phosphide, and quartz with finite metallization thickness," in *IEEE Int. Microwave Symp. Dig.*, pp. 691–694, 1991.