PROPAGATION CHARACTERISTICS OF LOSSY DISTRIBUTED GaAs FET STRUCTURES

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Abstract-The generalized transverse resonance technique is applied to compute the propagation characteristics of distributed FET structures. To account for conductor thickness of the order or less than the skin depth, a modified perturbational approach is applied, which yields very accurate results with negligible computational effort. Loss behaviors of the three dominant modes and the contributions of the various electrodes to the total loss have been investigated.

I. INTRODUCTION

Evaluation of loss is of paramount importance for the accurate modeling of MMIC's. While dielectric loss can be evaluated by just assuming a complex dielectric permittivity, conductor loss computation involves a considerable analytical and numerical effort. When metal thickness is of the order of the skin depth, in fact, the conventional perturbation approach becomes invalid. This is the case of GaAs FET structures, which are receiving a considerable interest for the possible realization of distributed amplification [1]. Though feasible, the rigorous analysis of the propagation characteristics of MMIC's structures including conductor loss involves a formidable computational effort.

In this paper the generalized transverse resonance technique [2] is applied to compute the propagation characteristics of multiconductor transmission line realized on multilayered lossy substrate. The conventional perturbation formula is modified to account for the finite thickness of the conductors. In this manner, the computational expenditure is reduced to that for the lossless case. Results are in very close agreement with experiments [3] and with those computed by rigorous but more time consuming methods [4]. The method is applied to compute the propagation characteristics of distributed GaAs FET structures. The three dominant quasi-TEM modes are found to possess different loss behaviors because of the uneven current distributions on the electrodes. The contributions to the overall loss of the various electrodes including the housing are pointed out.

II. METHOD OF ANALYSIS

Fig. 1a shows the schematic of a distributed FET structure. It consists of a multiconductor quasi-planar transmission line realized on a GaAs multilayered substrate. Three metal strips, representing the Source, Gate and Drain electrodes, plus, possibly, two grounded fins, are deposited on top of the substrate. The fins can be used to modify the lossy behavior by varying the source and drain capacitances. The structure is enclosed in a waveguide housing. All metal strips are supposed to have a finite non-zero thickness.

The analysis method adopted is the generalized transverse resonance technique described in [2], [5]. Such a method is equivalent to a mode-matching technique applied in the transverse y-direction. In each homogeneous region of Fig.1a the EM fields are expanded in terms of $TE^{(y)}$ and $TM^{(y)}$ modes:

$$\begin{aligned}
\mathbf{E}_t(x,y,z) &= \sum_{n=0}^{N} \left[V'_n(y)\, e'_{xn}(x,z) + V''_n(y)\, e''_{xn}(x,z)\right] \mathbf{x}_o + \left[I''_n(y)\, e''_{yn}(x,z)\right] \mathbf{y}_o \\
\mathbf{E}_z(x,y,z) &= \sum_{n=0}^{N} \left[V'_n(y)\, e'_{zn}(x,z) + V''_n(y)\, e''_{zn}(x,z)\right] \mathbf{z}_o \\
\mathbf{H}_t(x,y,z) &= \sum_{n=0}^{N} \left[I'_n(y)\, h'_{xn}(x,z) + I''_n(y)\, h''_{xn}(x,z)\right] \mathbf{x}_o + \left[V'_n(y)\cdot\, h'_{yn}(x,z)\right] \mathbf{y}_o \\
\mathbf{H}_z(x,y,z) &= \sum_{n=0}^{N} \left[I'_n(y)\, h'_{zn}(x,z) + I''_n(y)\, h''_{zn}(x,z)\right] \mathbf{z}_o
\end{aligned} \tag{1}$$

where the prime stands for TE modes and double prime for TM modes; the suffix 't' indicates the transverse to z components. The boundary conditions at the interfaces lead to a homogeneous system of equations in the equivalent voltage and current amplitudes. The condition for non trivial solutions constitutes the characteristic equation for the structure.

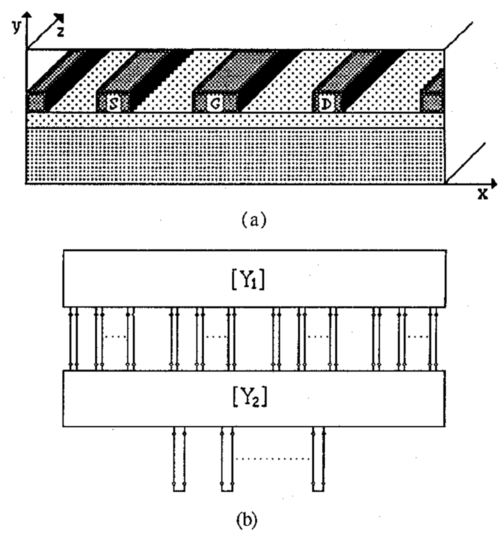


Fig. 1 a) Schematic of a distributed FET structure on a GaAs multilayered substrate; b) equivalent circuit for the structure transverse analysis.

It is worth mentioning that, 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 [6]. In practice, applying the so-called transverse segmentation [7] to Fig. 1a, 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. 1b). In this manner a substantial reduction in computing time is achieved. In terms of the network formalism, the characteristic equation is obtained as the resonance condition of the overall equivalent circuit of Fig. 1b.

Because of the presence of lossy layers, the solutions for the characteristic equations must be searched for in the complex plane. This has been done using the ZEPLS routine [8].

To account for power lost in the conductors, the conventional perturbation technique can be applied as long as the conductor thickness is much larger than the skin depth. One assumes that the tangential electric field $\mathbf{E}_t$ at the surface of an imperfect conductor is related to the tangential magnetic field $\mathbf{H}_t$ by

$$\mathbf{E}_\tau = Z_s\, \mathbf{H}_\tau \times \mathbf{n} \qquad (2)$$

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

$$Z_s = Z_c = (1+j)\sqrt{\frac{\omega\mu}{2\sigma}} \qquad (3)$$

The power lost in the conductor is then computed by the flow of the Poynting's vector. One obtains:

$$P_c = \frac{1}{2}\,\mathrm{Re}\left[Z_s \int_C |\,\mathbf{H}_\tau|^2\, dl\right] \qquad (4)$$

For thin conductors, specifically when the thickness t is comparable with the skin depth, the above assumptions lead to an underestimation of the conductor loss. Expression (3) is not valid any more, as the EM field penetrates deeply into the conductor reaching the opposite surface with finite amplitude. In a first approximation, the validity of (2) with unperturbed magnetic field at the metal surface can still be assumed, but with the impedance (3) replaced by [9]

$$Z_s = Z_c \frac{Z_l + Z_c \tanh[k_c t]}{Z_c + Z_l \tanh[k_c t]} \qquad (5)$$

with

$$k_c = (1+j)\sqrt{\pi\mu f\sigma} \qquad (6)$$

being the complex propagation constant of a plane wave within the metal. Although (5) correctly predicts the conductor loss increase with decreasing metal thickness, it still leads to loss values lower than those predicted by Heinrich [4]. This is due to the electric field at one metal surface depending on the magnetic field on both surfaces. This idea was suggested by Kitazawa in [10]. Accordingly, Eqn. (2) is replaced by

$$\mathbf{E}_{\tau 1} = Z_c \frac{\mathbf{H}_{\tau 1}\cosh[k_c t] - \mathbf{H}_{\tau 2}}{\sinh[k_c t]} \times \mathbf{n} \qquad (7)$$

The above expression can be used to compute the power flow into the metal, thus the conductor loss. As shown in the next section, excellent results are obtained.

III. RESULTS

The above approach has been first checked against the full wave analysis by Heinrich [4] (Fig. 2) and the experiments by Haydl et al. [3] (Fig. 3).

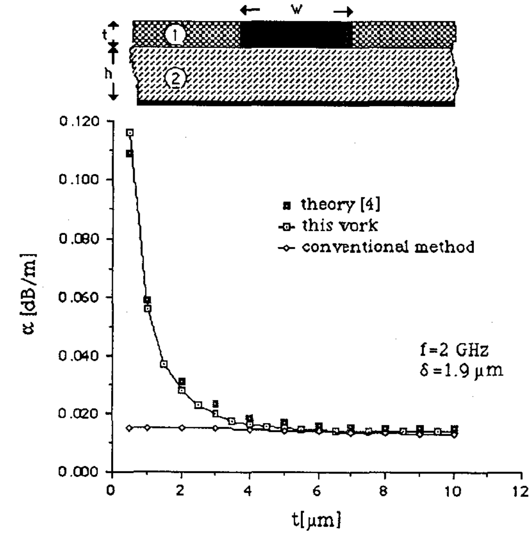


Fig. 2 Attenuation α of a MMIC microstrip line as a function of metallization thickness t. Line geometry: W=30 μm, h=200 μm. The dielectric characteristics of the subregions are: dielectric layer (1) (ε_r=3.4, tanδ=0.05); GaAs (2) (ε_r=12.9, tanδ=3x10^{-4}). Metallization conductivity σ=3.333x10^7[Ωm]$^{-1}$.

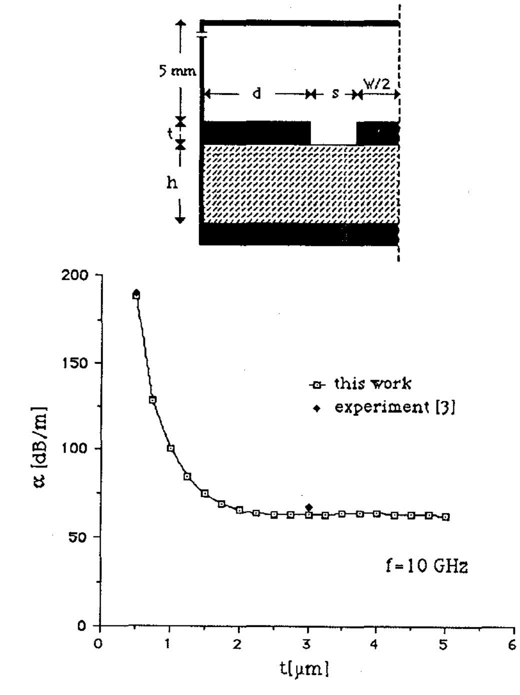
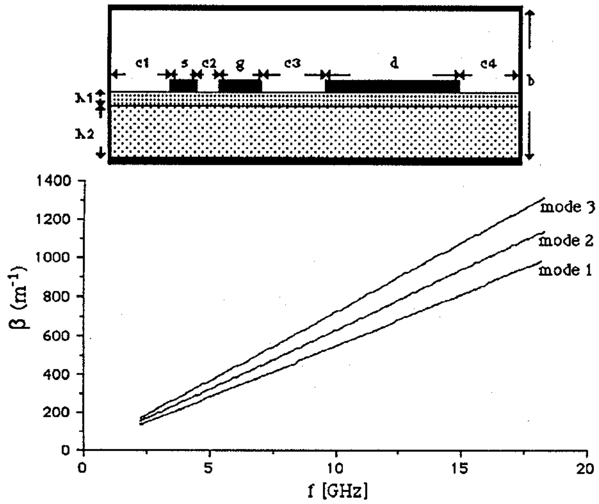


Fig. 3 Attenuation α of a CPW as a function of metallization thickness t. Line geometry: W=50 μm, d=150 μm, s=37.5 μm, h=500 μm. Substrate data: GaAs: ε_r=12.9, tanδ=3x10^{-4}. Metallization conductivity σ=4.1x10^7[Ω m]$^{-1}$.

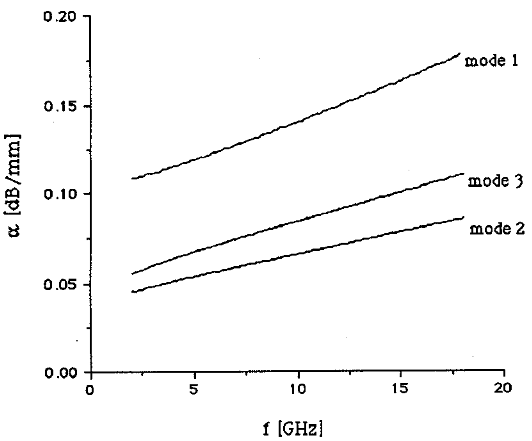
It is observed that while the conventional perturbational approach leads to erroneous results for thin strip conductors, our theory is in excellent agreement with [3] and [4] for all metal thicknesses.

The method has then been applied to compute the propagation characteristics of distributed FET structures. In this case, the accurate knowledge of the propagation constants of the three coupled modes is necessary to determine the conditions for attaining distributed amplification. It is also important to determine the single contributions of the various conductors to the total loss.

An example of the computed results is shown in Fig. 4a, b, the insert shows the schematic of the three conductor structure corresponding to the T-gate FET studied in [11].



(a)

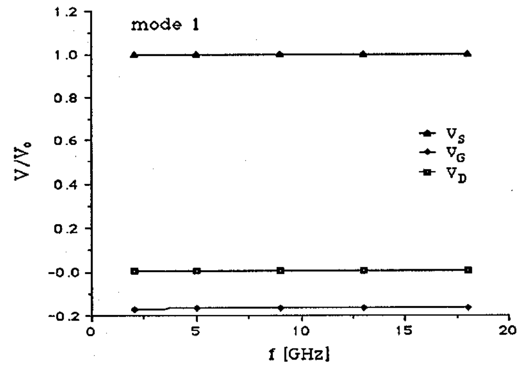


(b)

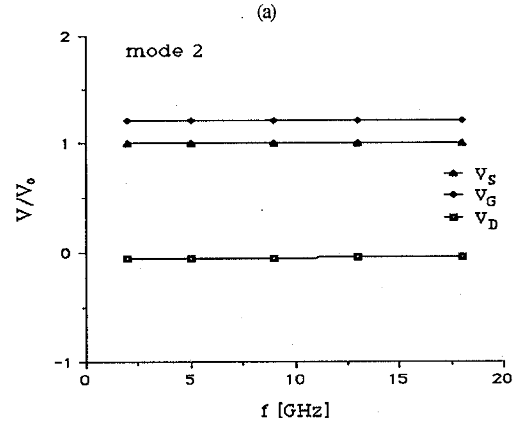
Fig. 4 Schematic of the T-gate FET structure from [11]: s=2μm, g=30μm, d=500μm, c1=c4=800μm, c2=10μm, c3=100μm, h1=0.1μm, h2=30μm, b=5.03mm, metallization thickness t=1μm; dielectric characteristics of the substrate layers: depletion region ε_r=12.9, channel ε_r=12.9, tanδ=3x10^{-4}; metallization conductivity σ=3.333x10^{7} [Ωm]$^{-1}$. a) Propagation characteristics, b) attenuation of the three modes in the 2-18 GHz band.

The complete lack of symmetry of the structure can be observed. Dielectric losses have been evaluated by simply assuming a complex permittivity for the semiconductor layer. Dielectric losses, however, have been found to be negligible compared to the conductor loss.

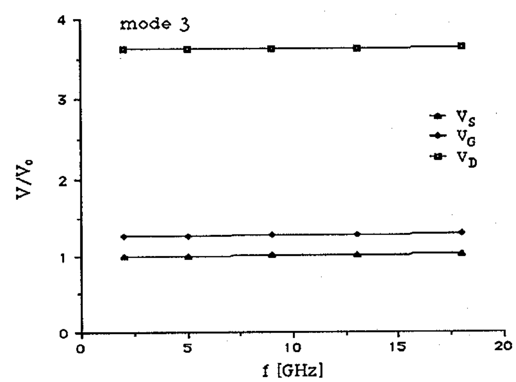
It is observed that each mode presents a quasi-TEM behavior in the all 2-18 GHz band and that mode 1 exhibits a much higher attenuation than modes 2 and 3. The voltage distributions associated to the three modes are shown in Fig. 5a, b, c.



(a)



(b)



(c)

Fig. 5 Voltage distribution associated to the three modes of Fig. 4a normalized to the 2 GHz source voltage V_0: a) mode 1, b) mode 2, c) mode 3.

The lossy mode (mode 1) corresponds to a field distribution with the drain electrode connected to ground.

Finally, the contributions of the single electrodes (D, G, S) and of the waveguide housing (H) to the total loss of the FET have been evaluated for each mode. Attenuation of mode 1 is considered as an example in Fig.6. These results prove that, for this mode, conductor loss mainly arises from the gate electrode and, partially, from the source. Additional computations (not shown here), on the contrary, have demonstrated that the waveguide housing have a significant role in determining the total losses of modes 2 and 3.

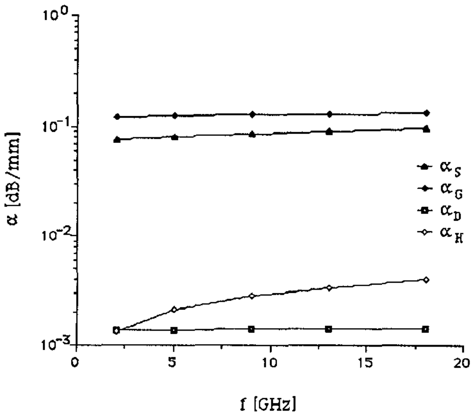


Fig. 6 Contribution of each electrode (source (α_S), gate (α_G), drain (α_D)) and of the waveguide housing (α_H) to the total loss for mode 1 in the FET structure.

IV. CONCLUSIONS

The generalized transverse resonance technique has been applied to compute the propagation characteristics of multiconductor quasi-planar lines for application to MMIC's and particularly to distributed FET structures. To account for conductor thickness of the order or less than the skin depth, a modified perturbational approach is applied, which yields very accurate results with negligible computational effort. Loss behaviors of the three dominant modes and the contributions of the various electrodes to the total loss have been investigated.

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