

Accurate Characterization and Modeling of Transmission Lines for GaAs MMIC's

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Abstract — The benefit of the spectral-domain hybrid mode approach in the design of multidielectric-media transmission lines is described. Using GaAs ring resonator techniques covering 2 to 24 GHz, accuracies in effective dielectric constant and loss of 1 percent and 15 percent respectively, are presented. By combining theoretical and experimental techniques, a generalized MMIC microstrip design data base is outlined.

I. INTRODUCTION

MODERN SYSTEM designs are demanding performance, size, yield, and cost specifications, which point to the use of GaAs MMIC technology. To meet these complex demands the IC manufacturer must now use improved computer-aided design (CAD) tools based on more accurate MMIC component models. This paper describes the use of new CAD tools together with high-accuracy microwave measurements to realize improved design data for GaAs MMIC's. In particular, this paper details a combined theoretical and experimental approach to the generation of an accurate design data base for transmission lines on GaAs MMIC's. The theoretical approach is based on an improved transmission line theory which is part of the spectral-domain hybrid mode computer program MCLINE [3], [4]. The program is capable of accurately predicting the effective dielectric constant for single and multiple dielectric layers in the presence of lid shielding. Furthermore, it is a generalized multiconductor strip analysis program and can be applied to other MMIC components such as couplers, interdigital capacitors, and spiral inductors. A detailed description is given in [4]. An extension of this work is now reported and theoretical emphasis has been placed on the treatment of loss mechanisms.

Accurate prediction of transmission line loss in MMIC's is complicated by the multilayered dielectric construction and complex metallization schemes used in typical GaAs IC technology. The dielectric layers used in addition to the semi-insulating GaAs substrate may exhibit comparatively high dielectric loss factors. Consequently, these thin multi-

layer dielectrics can have a strong influence on the field distribution of narrow lines and coupled line structures [1]. Also, conductors are often composed of different metals while the ground metallization may again be different from the top face metals. Frequently the ground metallization may consist of a very thin layer deposited on the base of a substrate having a relatively high surface roughness. In this complex situation and particularly for frequencies below or in *X*-band, the conventional microwave perturbation approach for the determination of conductor loss [2] fails since skin depth is not small when compared to the metallization thickness. Therefore, a new, more sophisticated perturbation method has been developed and used. The program has been designed to include loss mechanisms in all dielectric layers and to include conductor and surface roughness loss contributions. The ability to handle such multidielectric sandwich structures makes it one of the most suitable CAD tools for GaAs IC lines.

II. THEORETICAL APPROACH

The new treatment of loss presented below is new, as far as conductor loss is concerned, in that a modified microwave perturbation approach is used. It makes use of the ideal electromagnetic field, evaluated by means of the MCLINE computer program [3] for the lossless case. From the ideal electric field distribution and the dielectric loss factors of the layers involved, the absorbed power per unit length is determined in the conventional way (see, for example, [2]). The validity and high quality of this common approximate type of dielectric loss analysis have been demonstrated by Mirshekar-Syakhal in a comparative study [5]. The situation is completely different if conductor loss is considered. When conductor loss is required, direct application of the conventional perturbation approach is not possible since the magnetic field singularities at the microstrip edges [6], [7] have to be taken into account using the assumption that skin depth α is much smaller than metallization thickness t . In the modified perturbation approach, the electromagnetic field computation is firstly performed assuming zero strip thickness, ideal conductors, and ideal ground and cover metallizations; i.e., as in the classic treatment. This is done purely for reasons of numerical efficiency. The tangential magnetic field resulting from the ideal, lossless case with zero strip thickness is then processed in the following way:

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1) The singularity complication is resolved applying the assumption that the modification from zero to finite strip thickness essentially changes only the singular field near the conductor edges. This applies to a good approximation if the strip width w is larger than about 5 times the strip thickness t . Accordingly, the transition from $t = 0$ to non-zero values of t implies a modification of only the asymptotic behavior of the spectral-domain magnetic field quantities in the planes below and on top of the conductor metallization. For both fields this is achieved simply by the introduction of suitable spectral damping factors in conjunction with a spectral threshold beyond which these become effective and make the modified field square-integrable [6]. The criteria defining this modification have been found to yield stable results which are insensitive to the chosen threshold values.

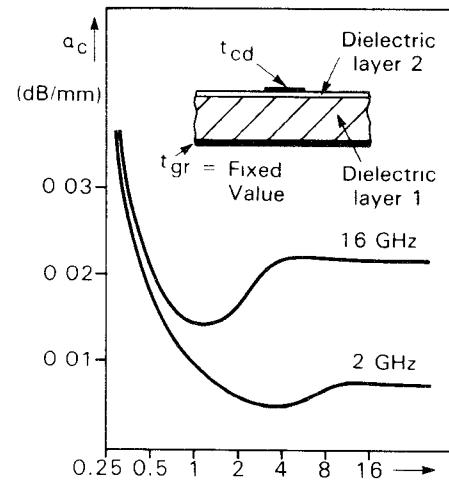
2) The complication associated with the thickness-to-skin-depth ratio t/a is resolved using the modified tangential magnetic field as a boundary condition in a lossy wave problem for each of the metallizations considered, ground and cover. In the case of the strip conductors, this means that the modified ideal magnetic field at the conductor top face is used as an impressed boundary condition for a fictitious strip metallization of thickness t with its top face located at $z = t$. An analogous boundary condition is formed for the conductor bottom face at $z = 0$ and for the ground and cover metallizations. Bearing in mind that in these regions on the substrate surface which are not covered by metal strips, the magnetic field is either vertical (near the strips) or much smaller than the components tangential to the strips, then an approximate analytical solution can be derived. This derivation is similar to the case of a plane metallic sheet of thickness t and the resultant strip current density therefore has the form

$$\mathbf{J}(x, z') = \mathbf{J}(x, 0) \cdot \sinh(t' - z') \sinh(t') + \mathbf{J}(x, t') \cdot \sinh(z') / \sinh(t') \quad (1)$$

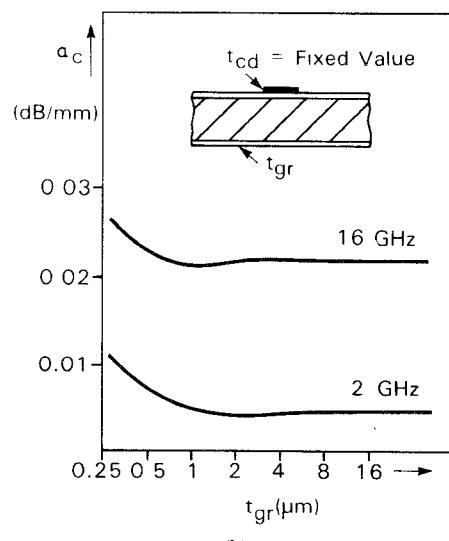
with $z' = (1 + j)z/a$, and $t' = (1 + j)t/a$, respectively.

In this relationship x is the coordinate transverse to the MMIC transmission line considered, and j is the imaginary unit, while a is the skin depth, already mentioned. The analysis for the ground and cover metallizations leads to similar analytical expressions. The x dependency of the current density contributions in (1) is derived from the numerically computed magnetic field by applying the integral relationship between strip current and magnetic field in the spectral domain. Though this is only an approximation, the results obtained have been found to agree well with measurements. A further improvement of this description has now been implemented into the next generation of MMIC transmission line computer analysis programs [8]. While (1) describes a homogeneous, single-material metallization, it can also be extended to composite, layered metallic structures. In any case, the power absorbed per unit length by the nonideal metallization is

$$P'_c = 1/2 \cdot \iint \rho(z) \cdot |\mathbf{J}(x, z)|^2 dx dz \quad (2)$$



(a)



(b)

Fig. 1. Theoretical predictions of conductor loss of MMIC sandwich microstrip versus (a) strip thickness t_{cd} and (b) ground metallization thickness t_{gr} .

where the integration is over the total metallization cross section of finite thickness t and $\rho(z)$ denotes the piecewise-constant resistivity. Surface roughness is taken into account using a well-known correction formula. The description given includes interaction of the tangential magnetic fields on both sides of a conductor if the thickness t is not large compared to the skin depth a . This interaction results in the interference effect already predicted by Horton [9] and a $1/t$ loss behavior for small thickness, which cannot be described by the classic perturbation approach. The loss analysis procedure briefly outlined above, however, does allow the consistent description of effects associated with finite strip thickness even though the analysis starts originally from zero-thickness magnetic field data.

This can be seen clearly in Fig. 1 for a sandwich microstrip structure (polyimide on top of a GaAs substrate) of about 50Ω characteristic impedance. In this

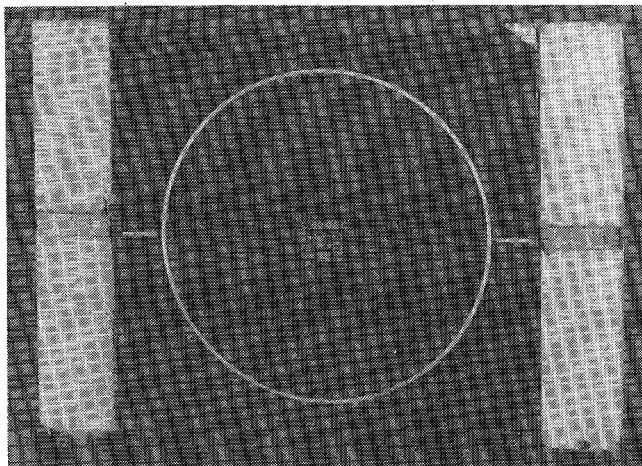


Fig. 2. GaAs MMIC ring resonator on measurement carrier.

case, the usual perturbation approach is correct only if the respective thicknesses are larger than about $4 \mu\text{m}$ (16 GHz) and $12 \mu\text{m}$ (2 GHz). It has to be expected that the improved theory described here predicts a $1/t$ law for very small thicknesses independent of frequency and down to dc. The behavior is less pronounced in Fig. 1(b) since the loss contribution of the ground plane is relatively small for the structure considered.

III. EXPERIMENTAL VERIFICATION

Having developed an advanced transmission line description capable of dealing with the geometries and structures found on GaAs MMIC's, it is then clearly necessary to verify the theory's accuracy for practical MMIC circuits. Of necessity this verification demanded a measurement procedure capable of much greater accuracy than the more usual substitution approach. For this reason, a ring resonator experimental technique was chosen as a proven method of obtaining accurate data on transmission line loss and effective dielectric constant [10]. A set of ring resonators were therefore produced based upon established design techniques. The resonators were realised with characteristic impedances in the range 20Ω to 100Ω and Fig. 2 shows a photograph of one of the GaAs IC ring resonator chips.

RF coupling to the ring was designed by introducing a "trident" gap capacitor at either side of a diameter. It was found that the gap capacitors could not be designed using established theories [11], [12] or software [13]. This was partly due to the fact that dielectrics were small relative to metal thicknesses and also due to the presence of multiple dielectric layers. Therefore a number of accurate measurements were made on various gap capacitors and modeled for the multiple-dielectric cases considered. The value of capacitor chosen had a typical value in the region of 0.028 pF . This provided adequate coupling to the resonators over the frequency range considered.

These resonators (10.03 mm outer diameter) were fabricated on $200 \mu\text{m}$ thick, 50.8 mm diameter semi-insulating GaAs wafers. A variety of multilayer dielectrics were used ranging from GaAs only to complex structures of GaAs,

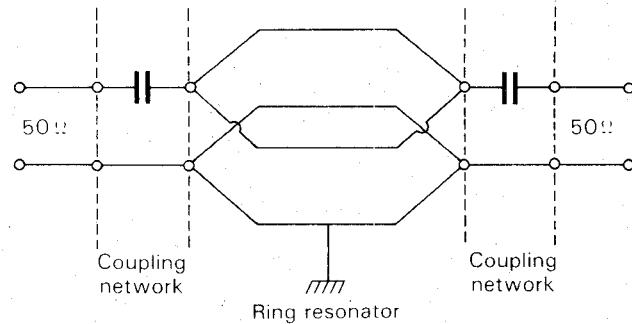


Fig. 3. Equivalent circuit model of ring resonator.

Si_3N_4 , polyimide, and Si_3N_4 . The processing of the ring resonators was carried out in our GaAs IC front-end production facility using standard, fixed fabrication procedures.

This large range of test ring structures were characterized using the latest network analyser techniques (HP8510T) from 2 to 24 GHz. The measured S parameters around each resonance frequency were optimized to the equivalent circuit model of Fig. 3 using commercially available CAD software. This method took account of the loading effects of the coupling capacitors, effective dielectric constant, loss and ring characteristic impedances with frequency. Estimations of the experimental errors involved showed effective dielectric constant measurement uncertainties of better than 1 percent. This high accuracy is due to the excellent frequency resolution of the synthesized output from the HP8510T since the determination of ϵ_{eff} relies only on resonant frequency measurements.

Using the above approach it was possible to identify the optimum low-loss IC technology and to determine the values of crucial material constants. To illustrate this, Fig. 4 shows the close agreement between the computed (MCLINES) and measured loss behavior of various ring resonators mounted on a GaAs substrate only. This particular MMIC transmission line structure (T1) had a thick composite metal top track designed for low-loss, high current handling applications. The close correlation between the measured and predicted results shown in Fig. 4 (15 percent average 2–20 GHz) was achieved using measured metal and dielectric thicknesses, metal resistivities, and surface roughness together with a GaAs loss tangent of 0.0003 as the MCLINES input data.

Fig. 5 presents a similar correlation for the case of various rings mounted on a GaAs/ Si_3N_4 /polyimide/ Si_3N_4 dielectric sandwich structure (T3). Losses in these structures are slightly higher due to reduced track thicknesses and the underlying polyimide layer. In this case the loss tangent for the thin polyimide layer was determined as 0.055 ± 0.005 , a value which agrees with subsequent large area MMIC polyimide capacitor measurements ($\tan \delta = 0.047$). Similar results have been obtained for 26Ω and 94Ω resonator structures as well as for other GaAs/polyimide sandwich structures (T2). In addition, the measured and predicted effective dielectric constants have been within 1 to 2 percent for both GaAs and multilayer dielectric

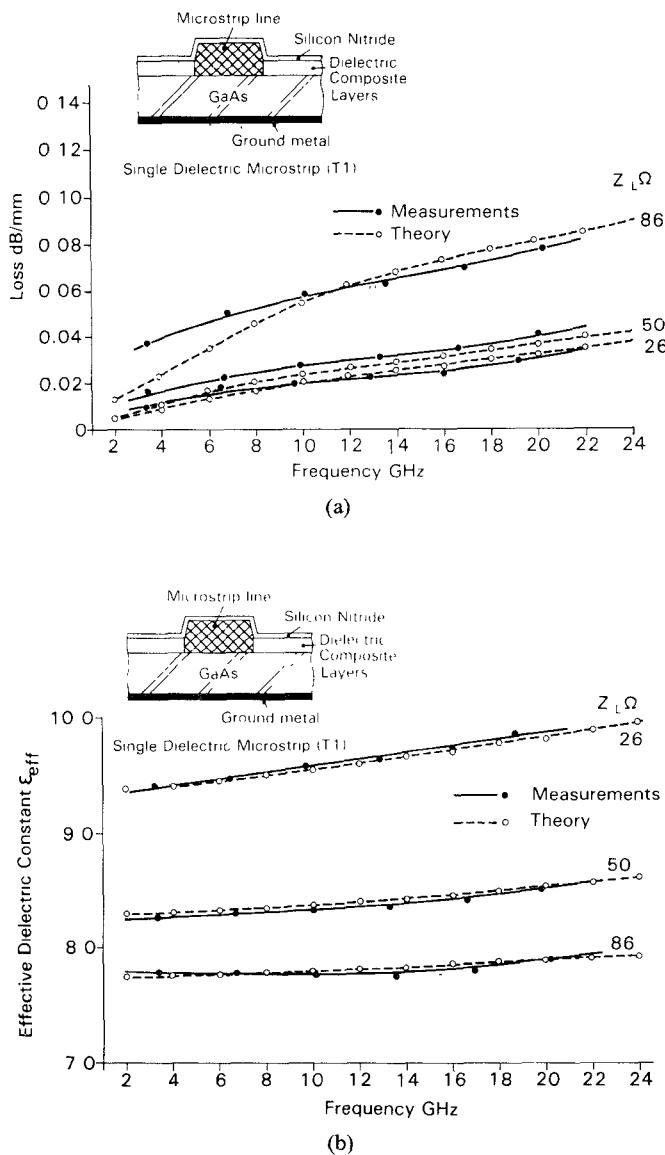


Fig. 4. (a) Predicted and measured loss versus frequency for line on GaAs only. (b) Predicted and measured effective dielectric constant versus frequency for line on GaAs only.

topologies. A comparison of crucial material constants with both the experimental and the theoretical results was also carried out. As an example Fig. 6 shows the measured and theoretical effective dielectric constants (ϵ_{eff}) of the T1 structure over the 2–20 GHz frequency range. Here it can be seen that the indicated dielectric constant (ϵ_r) of GaAs lies in the range 12.88 ± 0.025 . Similarly, the results from the more complex multidielectric structures established the dielectric constant of polyimide as being in the range 3.4 ± 0.2 . The effects on line loss characteristics of both surface roughness and line metal resistivities were also calculated from comparisons of measured and theoretical results. Fig. 7 presents the line loss sensitivity to variations in metal resistivity for the case of a 26Ω impedance ring resonator (T3).

It is useful to compare these results with established theories. For the single-dielectric microstrip on GaAs, Chu [14] has shown excellent agreement in effective dielectric

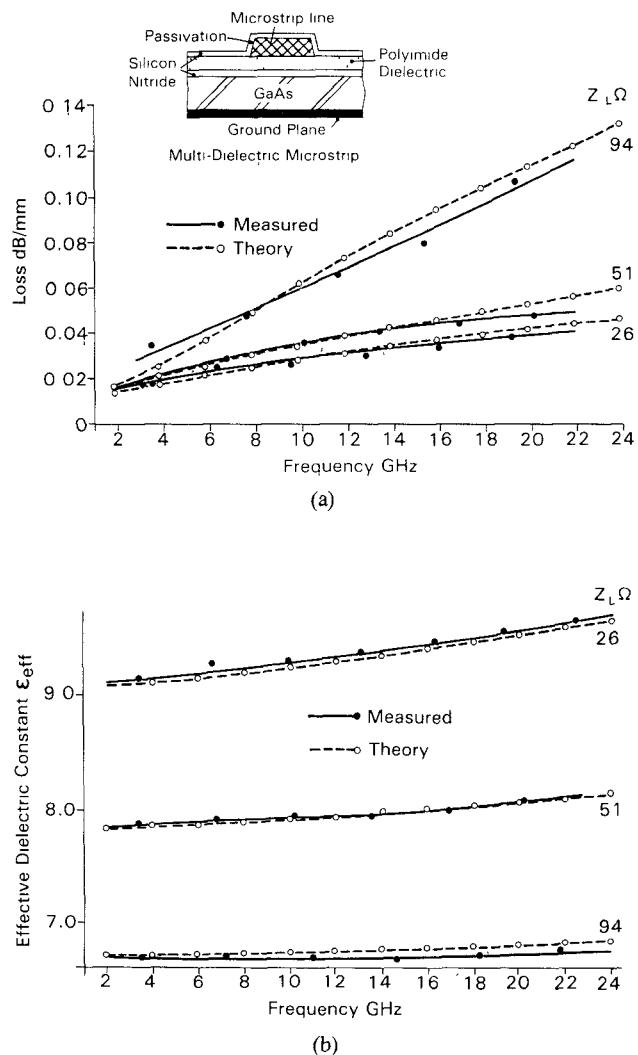


Fig. 5. (a) Predicted and measured loss versus frequency of MMIC microstrip lines. (b) Predicted and measured effective dielectric constant versus frequency of MMIC microstrip line.

constants using the model derived by Getsinger [15]. This has also been confirmed in this work. However, in the case of a multiple-dielectric microstrip it is a more complex situation. In Fig. 8 it is shown that the Getsinger model can still be used by essentially converting the multiple-dielectric case into an equivalent single-dielectric case. There exists a specific value of the equivalent single dielectric ($\epsilon_{r,net}$) such that excellent agreement exists between the model and practice. (In this case $\epsilon_{r,net} = 11.4$ for a 53Ω line.) The value of $\epsilon_{r,net}$ will vary with the impedance value selected and can be fitted to a simple expression; it can then be used to modify the Getsinger model for the multidielectric case. This exercise is useful when fits are required with commercial or existing software which only handle single-dielectric microstrip cases. However, it must be borne in mind that the value for $\epsilon_{r,net}$ can only be determined with reference to a theory or measurement technique which can accurately model the multidielectric case. Another example where fits with commercial software can be required is in the case of loss variation with frequency. Here the approach described in this paper can

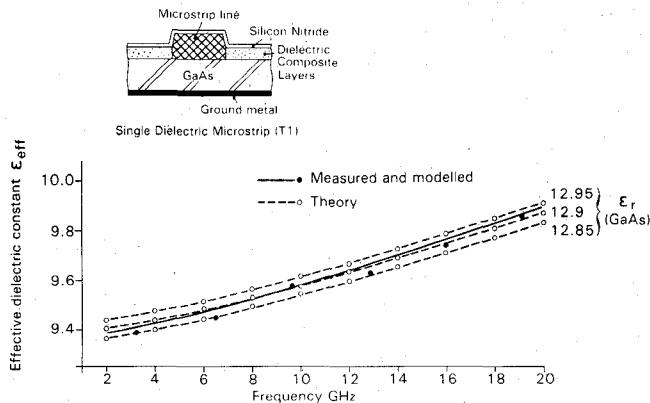


Fig. 6. Effective dielectric constant sensitivity to ϵ_r for a 25Ω line on GaAs only.

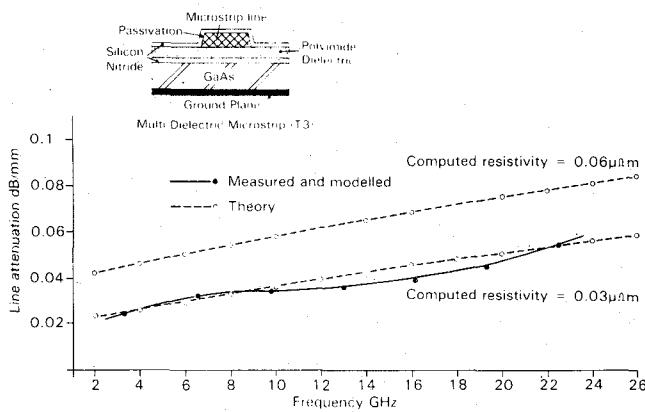


Fig. 7. Line attenuation sensitivity to metal resistivity for a 26Ω MMIC line.

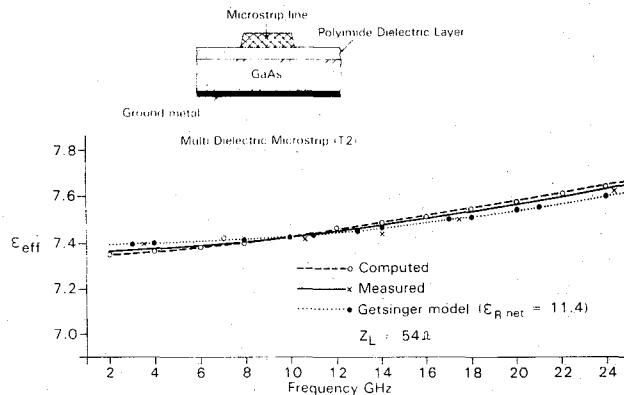


Fig. 8. Comparison of Getsinger model with measured and modeled multiedielectric MMIC microstrip versus frequency.

be used to calculate and then fit line losses for such programs as Super Compact™ [13] or Touchstone. In these cases the losses are defined as having a (freq.)^{1/2} dependence related to a loss value set at the lower band edge.

IV. MMIC MICROSTRIP DESIGN DATA BASE

The above detailed measurement and modeling work on several batches of ring resonators clearly established the accuracy and applicability of the improved analysis

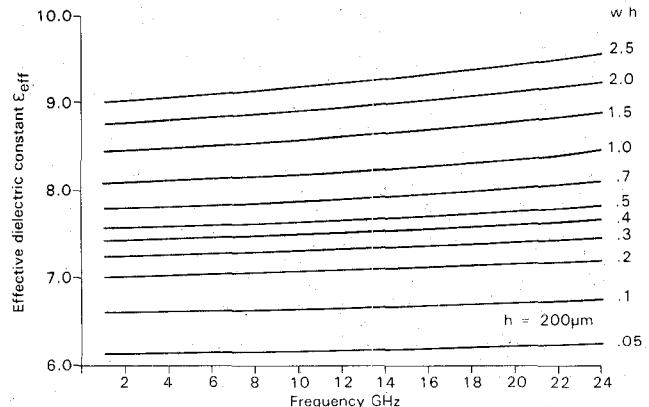


Fig. 9. Effective dielectric constant of MMIC microstrip lines as a function of frequency with w/h values.

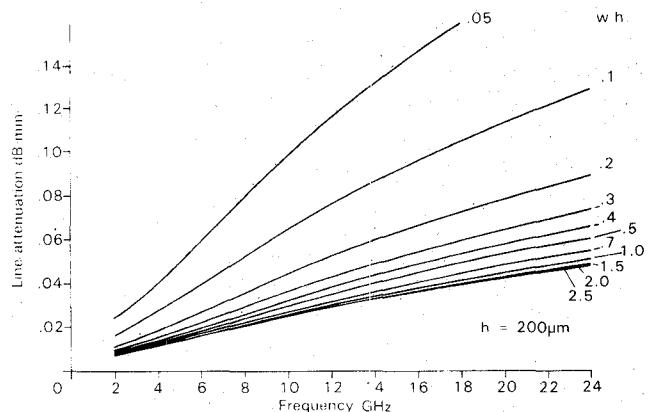


Fig. 10. Line attenuation of MMIC microstrip lines as a function of frequency and w/h values.

package. This CAD package, combined with further experimental work, was then used to develop a comprehensive design data base for the realization of high-accuracy, low-loss microstrip lines on GaAs MMIC's. Examples of the depth of information produced are presented in Figs. 9 and 10, which show the losses and effective dielectric constants exhibited by transmission lines realized on a commercially available GaAs IC foundry process.

The advantage of the combined practical and theoretical approach described is that a high degree of accuracy can be achieved over a wide range of microstrip line impedances and structures without the need for microwave measurements on a large number of lines. Instead, only a limited number of ring resonators have to be characterized in order to fix the required material constant values for the input file of the program MCLINES. At this stage MCLINES analysis runs can be used to "fill in" the data base for all the line impedances and structures required. Further, known foundry process variables can be programmed in order to establish the sensitivity of line loss and dielectric constants to process and mask manufacturing deviations. In this way the MMIC designer can now account for changes in metal resistivities, surface roughness, and physical dimensions, as well as dielectric constants and loss tangents. It is strongly suggested that this

depth of theoretical and practical characterization data is required for the accurate design and control of all MMIC circuit components.

V. CONCLUSIONS

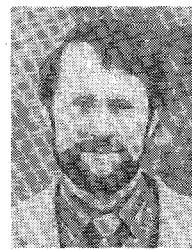
Using extensive experimental assessments, a powerful hybrid mode CAD package has been validated for multidi-electric microstrip in MMIC's to 24 GHz. RF loss and phase velocities have been related closely to physical and material parameters, resulting in a generalized MMIC microstrip design data base. As such this represents a significant contribution to CAD for MMIC's.

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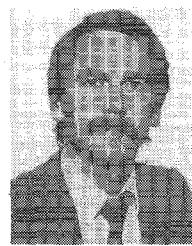
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Hugh J. Finlay was born in Belfast, N. Ireland. He received the B.Sc. degree (with honors) in electrical engineering and the M.Sc. degree in microwave communications and solid-state technology, both at The Queen's University of Belfast, N. Ireland.

He joined Plessey Electronics Systems Research Ltd., Hampshire, U.K., and worked on Gunn, IMPATT, and TRAPATT oscillators for systems use. He was also involved in the design and modeling of MIC components. As Principal Engineer he contributed to the development of solid-state short-range radar systems and integrated MIC subassemblies for airborne use. He developed the company's first GaAs FET VCO's and demonstrated their use in a complete microwave communication system. In 1977 he joined Plessey Research Caswell Ltd., Northants, England, where he designed a number of GaAs FET low-noise amplifiers for the company's product line. His research activities included GaAs FET oscillators, and dual gate GaAs FET QPSK modulators and discriminators. He carried out design studies for satellite switching matrices using GaAs MMIC switches, amplifiers, and passive networks. In 1983 he was Team Leader for MMIC component characterization and modeling. Later, in 1984-85, he was involved in characterization and modeling of MMIC components for the F1 foundry process. He applied new microwave software to solve specific MMIC design problems in conjunction with the University of Duisburg, W. Germany, and contributed to the F1 foundry design guide. He also developed and researched new circuit designs for high-packing-density MMIC's. In 1986 he joined Amplica Inc., Newbury Park, CA, and has been responsible for the design of RF subassemblies, power amplifiers, and subsystems using MMIC switches, FET amplifiers, and limiters. His current activities include switching amplifiers and LNA's for the company's product line.

Mr. Finlay has published and presented over 20 technical papers on MIC, MMIC designs, and RF subsystems. In 1986 he received (jointly with coauthors) a certificate of recognition for the best paper presented in a session of the 1986 IEEE MTT Symposium. Also in 1986, he received two awards from the Plessey Innovation Awards Scheme. He has filed ten U.K. patents on various microwave circuits and is a member of the IEE London.



Rolf H. Jansen (M'75-SM'84) received the M.S. (1972), and Ph.D. (1975) degrees, both in electrical engineering, from the University of Aachen (RWTH). In his theses he treated large-signal bipolar transistor modeling and the hybrid-mode analysis of arbitrarily shaped microstrip structures, respectively. He continued his research work at the RWTH Aachen Microwave Laboratory as a Senior Research Engineer (1976-1979), where he was mainly engaged in the characterization of MIC components and the CAD of

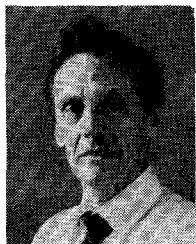
microwave circuits. He was also in charge of the thin-film technology of the microwave lab and, since 1977, worked as a research associate for radio communication at Standard Elektrik Lorenz AG (SEL) in Pforzheim, West Germany.

In 1979, he became Professor of Electrical Engineering at the University of Duisburg, near Düsseldorf/Cologne, and did teaching and research on such topics as electromagnetic theory, microwave techniques and CAD, measurement techniques, and modeling. His university career was supplemented by a one-year leave 1981/1982 as a full-time scientist with SEL Pforzheim, and by a variety of software and hardware projects for the communication industry since 1976. He developed, introduced, and tested the first layout-oriented general-purpose microwave CAD package in a West Germany production-oriented industry environment.

Dr. Jansen is the author of 60 technical papers in the field of microwave CAD and related topics and the recipient of the outstanding publications award in 1979 from the German Society of Radio Engineers. Presently, with a preparatory phase since the end of 1984, he is engaged in the development of a novel engineering CAD workstation for GaAs MMIC's with Plessey Research Caswell, GB, following completely new

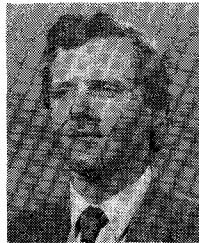
design concepts. He is cofounder of MCAD Software and Design Corp. in Aachen and owner of another small microwave company. He is on the editorial board of the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES and is a member of two MTT technical committees. He served as the West Germany MTT Chapter Chairman for the period 6/85 to 5/87 and is one of the two Distinguished Microwave Lecturers appointed by the MTT-Society for the year 1987/88.

fier, VCO designs, and he is currently studying novel techniques for high density circuit layouts on GaAs MMICs. Mr. Jenkins is coauthor of 12 publications covering both SAW and GaAs IC technologies.



John A. Jenkins was born in Gloucestershire, England. After leaving the Forest of Dean Mining and Technical School he completed a five-year apprenticeship with the A.E.I. Company, Rugby, as an electrical fitter on large electrical machines. He obtained an H.N.C. in electrical engineering with endorsements in electronics and applied electronics at the East Warwickshire College.

He continued at A.E.I. working on sound and vibration analysis of large electrical machines before moving to the Electrical Laboratory of Associated Engineering Company in 1960. He joined Plessey Research Caswell in 1973, where he worked initially on surface acoustic wave devices. In 1982 he started work on GaAs IC devices, assessing coupled line effects in closely spaced MMIC switch matrices. Other GaAs work has included low-noise ampli-



Ian G. Eddison was born in Harrogate, England, in 1952. He received the B.Sc. degree (with honors) in electronic engineering (1973) and the Ph.D. in microwave engineering (1978) from the University of Leeds, England.

In 1976 he joined Plessey Research Caswell Ltd., where he worked on millimeter-wave Gunn oscillators. From 1976 until 1981 his research work covered a wide range of mm-wave oscillator projects, from GaAs and InP Gunn devices to Si IMPATT devices. In 1982 he was appointed Research Leader for Plessey's GaAs IC circuit design area, becoming Group Leader in 1985. During this period he was involved in the development of Plessey's microwave wafer prober, the characterization and development of Plessey's F1 foundry process nonlinear IC design, and higher packing density circuit design techniques.

Dr. Eddison has published over 20 papers and has been issued several U.K. patents. In 1986 he was appointed Manager of Plessey's GaAs IC technology R&D department, where he is currently responsible for the GaA IC process R&D line.