Hybrid Mode Analysis of Microstrip Lines on Anisotropic Substrates

ABDEL-MONIEM A. EL-SHERBINY, MEMBER, IEEE

Abstract —A rigorous hybrid mode analysis is applied to microstrip lines on anisotropic substrates to determine its high-frequency performance. The analysis is based on a general formulation of the problem of planar transmission lines on multilayered substrates with uniaxial anisotropy, of which the microstrip line is a special case. Exact solution is obtained using a functional equation technique which was previously developed and applied to microstrip and bilateral finlines. The results were used to check the validity of the concept of equivalent isotropic substrate, suggested by some authors to simplify the calculation of the parameters of these lines. Certain approximations are introduced to allow the efficient calculation of the characteristics of microstrips on anisotropic substrates at relatively high frequencies or for wide strips. Numerical results are given for some values of the parameters of microstrip lines on sapphire and include the regions of excitation of higher modes.

I. Introduction

CRYSTALLINE substrates, e.g., sapphire, are used in microstrip lines intended for use in some applications. They have certain advantages over ceramics which include: lower losses, higher homogeneity, and lower variation of electrical parameters from specimen to specimen. However, being anisotropic, they lead to electrical performance of lines which differs somewhat from that of lines on isotropic substrates typically used.

Although microstrip lines were extensively investigated by a large number of authors and their theory is fairly well established, the case of anisotropic substrates received much less attention and at the present time there is no standard method for their calculation.

The quasi-static characteristics of microstrip lines on anisotropic substrates have been investigated by several authors [1]–[5]. They studied mainly the static parameters of the lines, obtained from the solution of electrostatic problems. Some authors suggested that the effect of anisotropy can be accounted for by the introduction of some equivalent isotropic medium with some effective parameters. Thus with uniaxially anisotropic substrates cut with the optical axis perpendicular to the plane of the substrate, it was assumed [4] that microstrip lines with substrate thickness d and relative dielectric constants in the directions parallel and perpendicular to the axis ε_z and ε_t, respectively, will behave as if they were isotropic with relative dielectric constant ε_e and effective substrate thickness d_e given by

$$d_e = d\sqrt{\varepsilon_t/\varepsilon_z} \qquad \varepsilon_e = \sqrt{\varepsilon_t \varepsilon_z}\,.$$

Width-dependent equivalent parameters were introduced by Edwards and Owens [1], [2] and were shown to give sufficiently accurate dispersion characteristics when used with existing wave theories of microstrip lines on isotropic substrates.

The purpose of this paper is to present a rigorous treatment of microstrip lines on anisotropic substrates and to check the validity of the previously introduced assumptions over an extended range of frequencies.

Manuscript received April 12, 1981; revised July 28, 1981.
The author is at 8832 Midbury Drive, Huntington Beach, CA 92646, on leave of absence from the Department of Electronics and Computer Engineering, Faculty of Engineering, Ains Sham University, Cairo, Egypt.

II. Formulation of the Problem of a Planar Transmission Line in a Multilayered Anisotropic Medium

Referring to the configuration shown in Fig. 1, comprising a planar transmission line embedded in a multilayered uniaxially anisotropic medium confined between two planar shields, which may be either conducting or magnetic walls. The permittivity tensor of the media of the layers is assumed to be of the form

$$\hat{\varepsilon} = \varepsilon_0 \begin{pmatrix} \varepsilon_t & 0 & 0 \\ 0 & \varepsilon_t & 0 \\ 0 & 0 & \varepsilon_z \end{pmatrix}.$$

The Cartesian coordinates are taken such that the z-axis is perpendicular to the plane of the line, and the magnetic properties are described by the scalar permeability $\mu_0\mu_r$. It can be easily seen that this configuration includes as special cases: microstrip lines when regions I, II are uniform and the shields are conductors or slot and suspended strip, as well as coplanar lines when region I is uniform (air) and region II is a dielectric layer-air structure. Moreover, it can include symmetrical structures such as bilateral finlines (more correctly bilateral slot-lines) when one of the shields is a magnetic wall. The dependence of the fields on time t and the longitudinal coordinate y is taken in the form $e^{-i(\omega t-\gamma y)}$, where γ is a real propagation constant and ω is the angular frequency. Field and current components will be expressed through their Fourier transforms

$$f(x,z) = \int \tilde{f}(\alpha, z) e^{-i\alpha x}\, d\alpha$$

where f represents any of the envolved functions and α is a complex parameter.

In the dielectric regions all field components can be

0018-9480/81/1200-1261$00.75 ©1981 IEEE

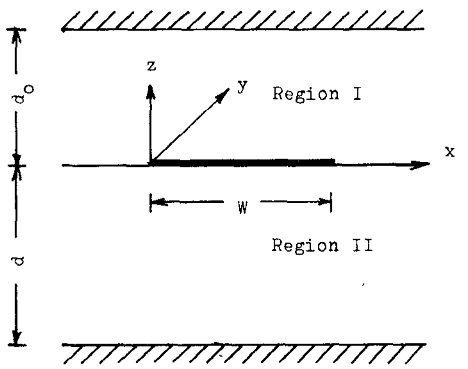

Fig. 1 General configuration of a planar transmission line in a multilayered anisotropic structure. Regions I, II are layered anisotropic media.

expressed through E_z, H_z. In terms of the transforms the following relations hold:

$$\frac{i}{\kappa^2}\frac{\partial \tilde{E}_z}{\partial z} = -\alpha \tilde{E}_x + \gamma \tilde{E}_y$$
$$-\omega\mu_0\mu_z \tilde{H}_r = \gamma \tilde{E}_x + \alpha \tilde{E}_y \quad (1)$$
$$i\frac{\partial \tilde{H}_z}{\partial z} = -\alpha \tilde{H}_x + \gamma \tilde{H}_y$$
$$\omega\varepsilon_0\varepsilon_z \tilde{E}_z = \gamma \tilde{H}_x + \alpha \tilde{H}_y . \quad (2)$$

$\tilde{E}_z$ and $\tilde{H}_z$ satisfy one-dimensional Helmholtz equations

$$\left[\frac{d^2}{dz^2} - \kappa^2(\alpha^2+\gamma^2-k_0^2\varepsilon_z\mu_r)\right]\tilde{E}_z = 0$$
$$\left[\frac{d^2}{dz^2} - (\alpha^2+\gamma^2-k_0^2\varepsilon_t\mu_r)\right]\tilde{H}_z = 0 \quad (3)$$

where

$$\kappa^2 = \varepsilon_t/\varepsilon_z \qquad k_0^2 = \omega^2\varepsilon_0\mu_0 .$$

Relations (1) and (2) determine the boundary conditions which must be satisfied by $\tilde{E}_z$, $\tilde{H}_z$.

1) Over the shields either $\tilde{H}_z = 0, \partial\tilde{E}_z/\partial z = 0$ for conductors, or $\partial\tilde{H}_z/\partial z = 0$, $\tilde{E}_z = 0$ for magnetic walls.

2) On the interfaces between the dielectric layers (excluding the plane $z=0$ of the transmission line) the quantities

$$(1/\kappa^2)\partial\tilde{E}_z/\partial z,\ \mu_r\tilde{H}_z,\ \partial\tilde{H}_z/\partial z,\ \varepsilon_z\tilde{E}_z$$

are continuous.

3) Over the plane of the line $z=0$, $(1/\kappa^2)\partial\tilde{E}_z/\partial z$ and $\mu_r\tilde{H}_z$ are continuous while the discontinuities in $\partial\tilde{H}_z/\partial z$ and $\varepsilon_z\tilde{E}_z$ are given by

$$\omega\varepsilon_0[\varepsilon_z\tilde{E}_z]_{z=\pm 0} = -\alpha\tilde{J}_x(\alpha) + \gamma\tilde{J}_y(\alpha)$$
$$-i[\partial\tilde{H}_z/\partial z]_{z=\pm 0} = \gamma\tilde{J}_x(\alpha) + \alpha\tilde{J}_y(\alpha)$$

where J_x, J_y are the total surface current density vector components flowing on the thin transmission-line conductors. Introducing the notations

$$U_1(\alpha) = \omega\varepsilon_0[\varepsilon_z\tilde{E}_z]_{z=\pm 0} = -\alpha\tilde{J}_x + \gamma\tilde{J}_y$$
$$U_2(\alpha) = -i[\partial\tilde{H}_z/\partial z]_{z=\pm 0} = \gamma\tilde{J}_x + \alpha\tilde{J}_y$$
$$F_1(\alpha) = [(i/\kappa^2)\partial\tilde{E}_z/\partial z]_{z=0} = [-\alpha\tilde{E}_x + \gamma\tilde{E}_y]_{z=0}$$

and

$$F_2(\alpha) = -\omega\mu_0[\mu_r\tilde{H}_z]_{z=0} = [\gamma\tilde{E}_x + \alpha\tilde{E}_y]_{z=0}.$$

The general relation between the introduced functions F_1, F_2, U_1, U_2 can be written in the form

$$U_1 = i\omega\varepsilon_0\chi_1 F_1$$
$$i\omega\mu_0 U_2 = \chi_2 F_2 . \quad (4)$$

Functions χ_1, χ_2 are determined from the solution of the boundary value problems expressed by (3) and boundary conditions on $\tilde{E}_z$ and $\tilde{H}_z$. They are the Fourier transform of the inverse Green's functions of the structure. For the shielded structure considered χ_1, χ_2 are meromorphic functions of the complex variable α. Their poles on the upper half plane $\operatorname{Im}\alpha > 0$ are denoted by α_n, β_n, respectively, while the zeros are denoted by ν_n, σ_n.

Equations (4) relate the tangential components of the electric field to the exciting surface current sources. Functions U_1, U_2, F_1, F_2 have certain properties arising from the boundary conditions imposed on surface currents and tangential fields. Functions satisfying (4) and the boundary conditions give the complete solution of the problem of planar transmission lines on anisotropic substrates. Relations (4) have been derived and used to analyze the problems of microstrip and bilateral finlines on isotropic substrates [6]. It is shown that the theory can be extended to include the more general case of lines on layered anisotropic media.

III. Microstrip Line on Anisotropic Substrate

This is a special case of the configuration of Fig. 1, when region I is an air medium $\varepsilon_t = \varepsilon_z = 1$, while region II is an anisotropic homogeneous medium and the strip conductor of width W lies on the plane $z=0$ on the interval $0 < x < W$. The upper shield is at a distance $z = d_0$, while the base conductor coincides with the plane $z = -d$, where d is the substrate thickness. Functions χ_1, χ_2 have the form

$$\chi_1(\alpha) = \frac{\coth R_0 d_0}{R_0} + \varepsilon_e \frac{\coth R_z d_e}{R_z}$$
$$\chi_2(\alpha) = R_0 \coth R_0 d_0 + \frac{1}{\mu_r} R_t \coth R_t d$$

where

$$R_0 = \sqrt{\alpha^2+\gamma^2-k_0^2} \qquad R_z = \sqrt{\alpha^2+\gamma^2-k_0^2\mu_r\varepsilon_z}$$
$$R_t = \sqrt{\alpha^2+\gamma^2-k_0^2\mu_r\varepsilon_t}, \qquad k_0^2 = \omega^2\varepsilon_0\mu_0 .$$

Following the method described in [6], functions F_1, F_2 are represented in the form

$$F_1(\alpha) = F_1^-(\alpha) \pm e^{i\alpha W} F_1^-(-\alpha)$$
$$F_2(\alpha) = F_2^-(\alpha) \mp e^{i\alpha W} F_2^-(-\alpha)$$

where upper signs refer to modes with symmetrical distribution of longitudinal current with respect to the strip center, while lower signs refer to antisymmetrical modes. F_1^-, F_2^- are functions, analytic on the lower half plane $\operatorname{Im}\alpha < 0$. For symmetrical modes the problem reduces to the solution of the following sets of simultaneous algebraic equations:

$$A_n = 1 + \sum_{m=0}^{\infty} \frac{\xi_m}{\alpha_n + \alpha_m} A_m, \qquad n = 0, 1, 2, \cdots$$

$$B_n = 1 + \sum_{m=1}^{\infty} \frac{\zeta_m}{\beta_n + \beta_m} B_m, \qquad n = 1, 2, \cdots \tag{5}$$

with coefficients

$$\xi_n = \frac{\operatorname{Res} \chi_1^-(\alpha_n)}{\chi_1^-(-\alpha_n)} e^{i\alpha_n W} \qquad \zeta_n = -\frac{\operatorname{Res} \chi_2^-(\beta_n)}{\chi_2^-(-\beta_n)} e^{i\beta_n W}$$

$\chi_1^-(\alpha), \chi_2^-(\alpha)$ are the minus functions resulting from the factorization of χ_1, χ_2. Functions F_1^-, F_2^- are given by

$$F_1^-(\alpha) = \frac{P}{\chi_1^-} \left\{ 1 - \sum_{n=0}^{\infty} \frac{\xi_n}{\alpha - \alpha_n} A_n \right\}$$

$$F_2^-(\alpha) = \frac{Q}{\chi_2^-} \left\{ 1 - \sum_{n=1}^{\infty} \frac{\zeta_n}{\alpha - \beta_n} B_n \right\}. \tag{6}$$

P, Q are some constants.

The propagation constant γ is calculated from the conditions

$$F_1^-(\pm i\gamma) \pm iF_2^-(\pm i\gamma) = 0. \tag{7}$$

Expressions (6) together with (7) give the complete solution to the problem. Once functions F_1^- and F_2^- are determined, the wave impedance which is defined here as the ratio of the quasi-static voltage at the center of the strip to the total longitudinal current, can be calculated as in the case of isotropic substrate, from the expression

$$Z_0 = \frac{i}{2\alpha_0} \frac{\gamma F_1^-(-\alpha_0)}{\omega\varepsilon_0 F_1^-(0)\chi_1(0)} e^{i\alpha_0 W/2}$$

where α_0 is the lowest order pole of χ_1, given by

$$\alpha_0 = \sqrt{k_0^2 \mu_r \varepsilon_z - \gamma^2}.$$

IV. Wide Strip/High Frequency Approximation and Higher Order Modes

The nature of propagation in microstrip lines has been discussed in [6], where it was interpreted as a multiple reflection process of the plane TEM wave between the strip and the base conductor from the strip edges. Conditions of total reflection from the edges exist in the anisotropic case when γ lies in the interval

$$\bar{\nu}_0^2 < \gamma^2 < k_0^2 \mu_r \varepsilon_z$$

where $\bar{\nu}_0$ is the zero-order root of χ_1 when $\gamma = 0$. In this case all zeros and poles of χ_1, χ_2, except for α_0, are imaginary and the coefficients ξ_n, ζ_n are exponentially decaying. Systems (5) acquire a simple form when

$$e^{-W\sqrt{\left(\frac{\pi}{d_e}\right)^2 + \gamma^2 - k_0^2 \varepsilon_z \mu_r}} \ll 1$$

$$\frac{1}{\varepsilon_z} e^{-W\sqrt{\gamma^2 - k_0^2}} \ll 1 \tag{8}$$

which mean that higher order evanescent waveguide modes in the regions below and above the strip, excited by the reflection of the TEM wave in the region between the strip and the substrate, will hardly reach the other edge of the strip due to high attenuation. In this case all coefficients ξ_n, ζ_n except ξ_0 will be negligibly small and dispersion equation (7) can be written in the real form

$$1 + k_0^2 A^2 \frac{\cos(\phi/2 + \psi)}{\cos(\phi/2 - \psi)} = 0 \tag{9}$$

where

$$\psi = \tan^{-1}(\gamma/\alpha_0) = \sin^{-1}\left(\gamma/k_0\sqrt{\mu_r \varepsilon_z}\right)$$

$$\phi = \alpha_0 W - 2\arg\chi_1^+(\alpha_0)$$

$$A = \left| \frac{\chi_1^+(i\gamma)}{\chi_2^+(i\gamma)} \right|.$$

Explicit expressions for A, $\arg\chi_1^+$ can be obtained through the application of the theory of factorization of meromorphic functions in infinite products [8]:

$$A = \prod_n \frac{\left(1 + \frac{\gamma}{\nu_n'}\right)\left(1 + \frac{\gamma}{\beta_n'}\right)}{\left(1 + \frac{\gamma}{\alpha_n'}\right)\left(1 + \frac{\gamma}{\sigma_n'}\right)} \cdot \frac{\alpha_0}{k_0}\sqrt{|\chi_1(0)/\mu_r \varepsilon_z \chi_2(0)|}$$

$$\arg\chi_1^+(\alpha_0) = \sum_n \tan^{-1}(\alpha_0/\alpha_n') - \tan^{-1}(\alpha_0/\nu_n') - \frac{\alpha_0}{\pi}\{d_0 \ln(1 + d_e/d_0) + d_e \ln(1 + d_0/d_e)\}$$

where

$$\alpha_n = i\alpha_n'$$
$$\beta_n = i\beta_n'$$
$$\nu_n = i\nu_n'$$
$$\sigma_n = i\sigma_n'.$$

For the same approximations the line impedance is given by

$$Z_0 = \frac{60\pi\gamma}{k_0 \cos(\phi/2)} \sqrt{-d_e/\varepsilon_e \chi_1(0)}. \tag{10}$$

Conditions (8) prevail for lines with large stripwidth to substrate thickness ratio and/or at high enough frequencies. It is difficult to determine the exact regions where the approximate expressions hold. However, if the propagation constant γ is known or can be estimated conditions (8) can be checked and if found valid the impedance Z_0 can be computed from (10). Expressions (9), (10) can be used, within the range of their validity, to calculate the width of the strip for a given impedance and frequency. Thus for a specified Z_0 the angle ϕ is determined from (10) and substituted in (9) and the resulting equation solved for γ. Once γ is calculated the width can be determined from the

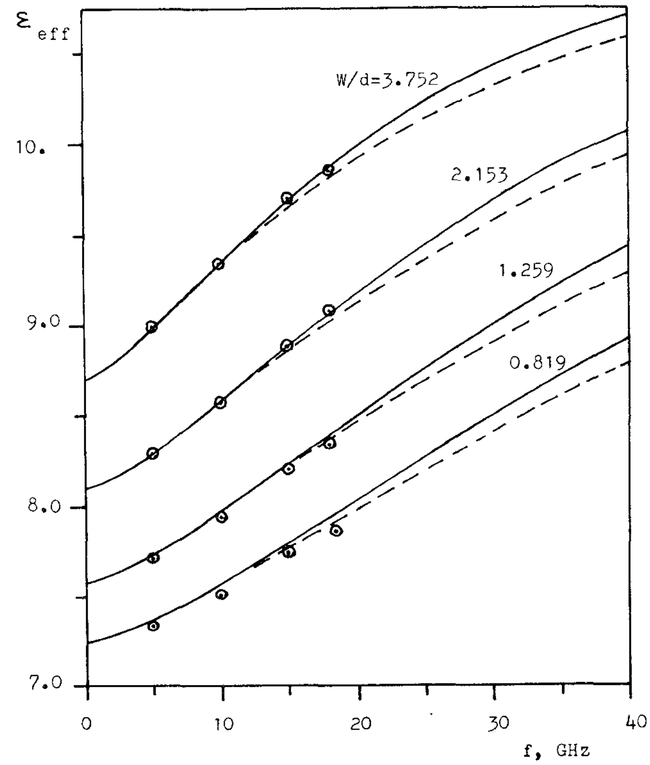

Fig. 2. Dispersion characteristics of the lines investigated in [1], [2]. ——— present theory, ---- present theory regarding sapphire as an isotropic dielectric with equivalent ε given by [1], [2], ⊙ experimental results

expression

$$W=\frac{\phi}{\alpha_0}+\frac{2}{\alpha_0}\arg\chi_1^+(\alpha_0).$$

Approximate expression (9) is even more accurate for the determination of the characteristics of higher even modes. The reason is that these modes will exist only at high enough frequencies that conditions (8) will be automatically satisfied. For the calculation of odd modes, including the first higher mode in the microstrip line, expression (9) must be modified to

$$1+k_0^2A^2\frac{\sin(\phi/2+\psi)}{\sin(\phi/2-\psi)}=0. \tag{11}$$

Expressions (9), (11) can be derived also using the transverse resonance method applied in [7] to open microstrips.

V. Numerical Results and Conclusions

Comparing the expressions for χ_1, χ_2 with their expressions in the isotropic case, i.e., when $\varepsilon_z=\varepsilon_t=\varepsilon_r$ we notice that except for the root R_z function χ_1 keeps its form. It was shown in [6] that the TM-to-z component of the field, which is described by χ_1, dominates at low frequencies. This leads to the conclusion that the equivalent isotropic parameters can be introduced in the manner suggested by Horno [4] and are expected to give accurate results in the quasi-static limit.

However, in the anisotropic case the high frequency limit of the effective dielectric constant of the microstrip line ε_{eff} is always ε_z irrespective of the line dimensions and it approaches this limit faster as the strip width increases. This situation favors the use of width-dependent parameters if the microstrip dispersion is to be taken into account.

Computations have been performed using the present theory to calculate the high-frequency performance of microstrip lines on sapphire substrates with $\varepsilon_t=9.4$, $\varepsilon_z=11.6$ and with dimensions of the lines investigated by Edwards and Owens [1], [2]. The results shown on Fig. 2 are in good agreement with the experimental measurements given by the same authors.

Calculations were also carried out using the equivalent isotropic parameters suggested in [1], [2], [4]. They were compared with the results obtained from the exact theory, leading to the following conclusions.

1) Values of the effective dielectric constant ε_{eff} and the line impedance Z_0 calculated using the isotropic parameters suggested by Horno differ by about 4 percent at 5 GHz up to about 9 percent at 40 GHz from the results of the theory. Therefore, they can be used mostly for rough design calculations.

2) The width-dependent parameters suggested by Edwards and Owens lead to values of ε_{eff} and Z_0 agreeing well with the theory up to 25 GHz and differ by less than 2 percent up to 40 GHz over the range of frequencies and geometries considered. Therefore, this equivalent isotropic dielectric constant is expected to give accurate results over

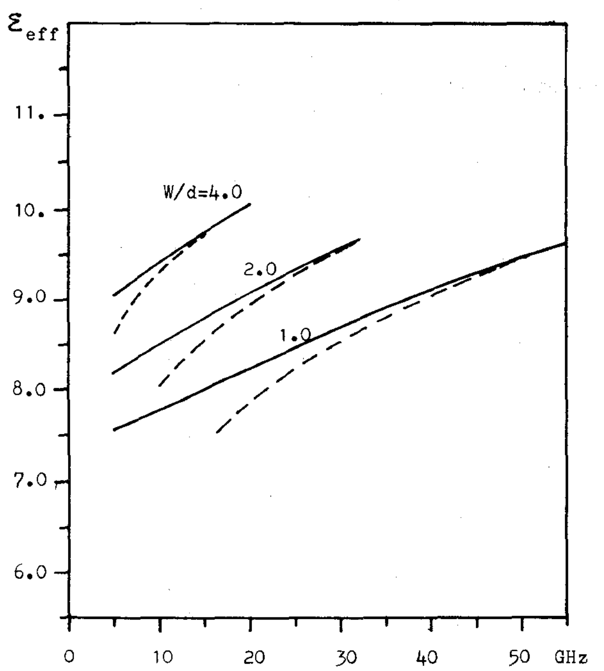

Fig. 3. Dispersion characteristics of microstrip on 0.5-mm sapphire substrate. ——— exact, ---- approximate results. The curves are terminated at the points of excitation of the first higher mode.

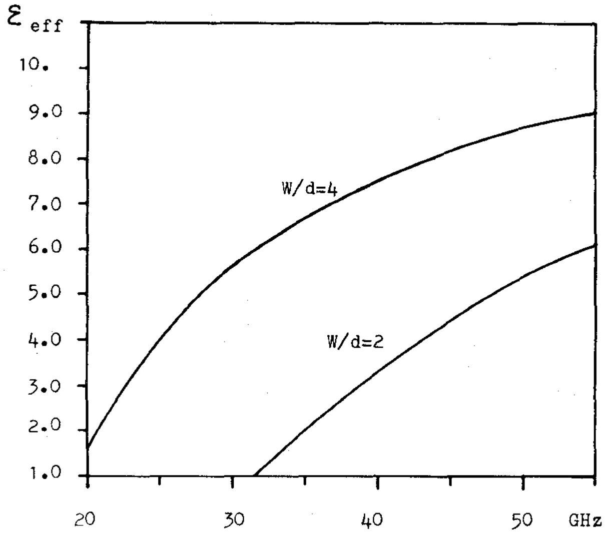

Fig. 5. Dispersion and cutoff characteristics of the first higher odd mode in microstrip line on 0.5-mm sapphire.

wide frequency range when used with simplified theory or model accounting for dispersion in microstrip with isotropic substrates for CAD purposes.

3) When the microstrip is to be used at higher frequencies or when higher accuracy is required rigorous theory has to be used for the determination of its performance.

To check the validity of the approximations introduced, lines with strip width to substrate thickness ratios $W/d = 1,2,4$ on 0.5-mm thick sapphire were also calculated and the results obtained from the exact and approximate equations are shown on Figs. 3 and 4. The dispersion characteristics of the first higher (odd) mode are also given on Fig. 5 where the results of the exact and approximate expressions were found nearly coincident starting from the cutoff frequency.

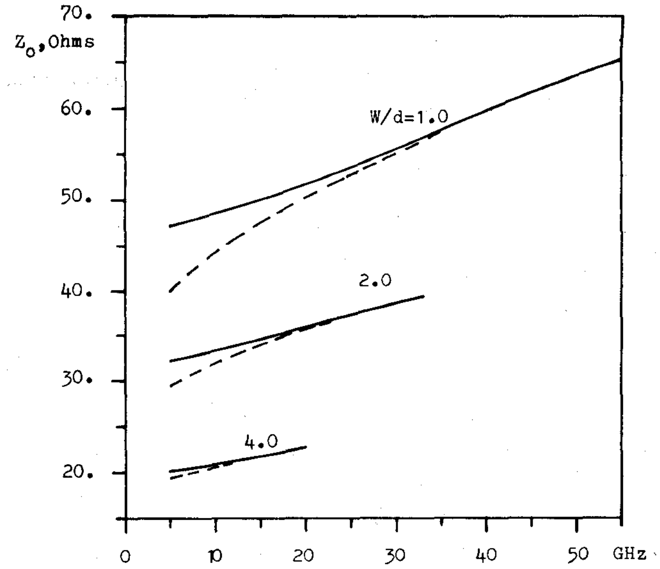

Fig. 4. Impedance of microstrip line on sapphire. ——— exact, ---- approximate results.

References

[1] R. P. Owens, J. E. Aitken, and T. C. Edwards, "Quasi-static characteristics of microstrip on an anisotropic sapphire substrate," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, Aug. 1976.

[2] T. C. Edwards and R. P. Owens, "2–18 GHz dispersion measurements on 10–100 Ω microstrip lines on sapphire," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, Aug. 1976.

[3] N. G. Alexopoulos and C. M. Krowne, "Characteristics of single and coupled microstrips on anisotropic substrates," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, June 1978.

[4] M. Horno, "Calculation of quasi-static characteristics of microstrip on anisotropic substrate using mapping method," presented at the IEEE MTT-S Int. Microwave Symp., Washington, DC, May 28–30, 1980.

[5] M. El-Said and A. A. Ahmed, "Microstrip analysis on anisotropic and/or inhomogeneous substrate with finite element method," presented at the IEEE MTT-S Int. Microwave Symp., Washington, DC, May 28–30, 1980.

[6] A-M. A. El-Sherbiny, "Exact analysis of shielded microstrip lines and bilateral finlines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, July 1981.

[7] A. T. Fialkovskiy, "Theory of higher order modes in a nonsymmetrical stripline," *Radioengineering and Electron Physics*, vol. 21, no. 4, 1976.

[8] R. Mittra and S. W. Lee, *Analytical Techniques in the Theory of Guided Waves*. New York: Macmillan, 1971.

Abdel-Moniem A. El-Sherbiny (M'78) received the B.S.E.E. degree from Cairo University, Cairo, Egypt, in 1960 and the Ph.D. degree from Leningrad State University, Leningrad, U.S.S.R., in 1968.

After graduation he worked as an Instructor in the Military Technical College, Cairo, Egypt, during the academic year 1960–1961. From 1963 to 1967 he worked towards the Ph.D. degree in Leningrad State University, developing slow wave periodic structures for high power microwave devices. He joined Ain Shams University, Cairo, in 1968 first as a lecturer, then Assistant Professor and Professor. During this period he performed and supervised research and development of new microwave and millimeter-wave components and transmission media as well as providing consultation to several projects. Recent publications deal with the millimeter-wave performance of planar transmission lines including microstrip, fin and slot lines on isotropic and anisotropic substrates. Since March 1981 he is on a leave of absence from Ain Shams University, teaching engineering electromagnetics at University of California, Irvine, and persuing research.

Chip Level IMPATT Combining at 40 GHz

CHARLES T. RUCKER, SENIOR MEMBER, IEEE, JOHN W. AMOSS, SENIOR MEMBER, IEEE, AND GERALD N. HILL

Abstract—**Results with series and series–parallel connections of CW 40-GHz IMPATT diodes on diamond are discussed. The effects of device and circuit losses on the efficiency are treated. Device loss associated with the stabilizing capacitors appears likely as the major factor limiting the combining efficiency. Maximum combining efficiency of 82 percent has been demonstrated for two diodes connected in series. The multichip geometries utilize Raytheon gallium arsenide CW double-drift diode chips and are essentially scaled versions of successful *X*-band geometries previously reported [1] by the authors.**

I. Introduction

The authors have reported successful work with series and series–parallel connected IMPATT devices at X-band frequencies [1]; it is desirable to extend this technique to millimeter frequencies if possible. Reported herein is work aimed at scaling the X-band multichip geometries and techniques to 40 GHz. Current results show the approach to be feasible with up to 82-percent combining efficiency; however, high combining efficiency (η_c) has not been routinely obtained. A possible explanation, related to capacitor loss, for some of the low η_c values is summarized. The effects of circuit losses are also treated. The function of the capacitors and criteria for their selection have been discussed in [1].

II. Geometry Fabrication

Fig. 1 is a sketch showing a typical successful X-band CW device and its nominally equivalent 40-GHz counterpart. The devices are not exactly to scale. Also shown, in Fig. 2, is an SEM photomicrograph of a typical two-chip 40-GHz device.

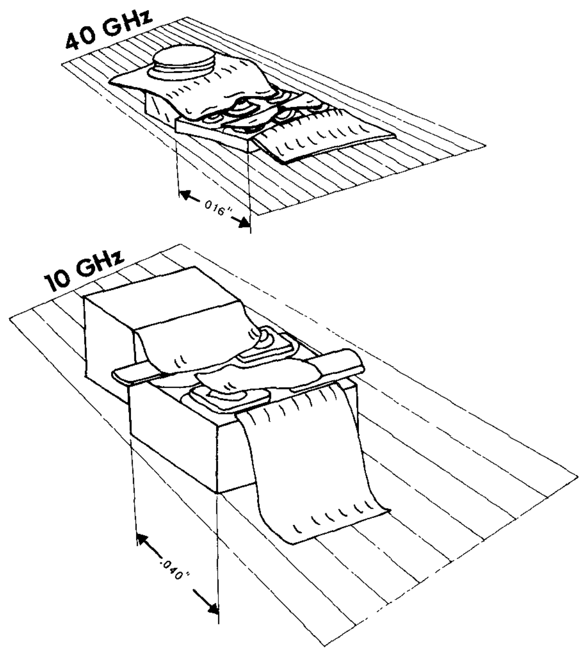

Fig. 1. Multichip IMPATT geometries, 10 GHz and 40 GHz.

In general, all dimensions, including diamond heat sink size, lead lengths, etc., have been held quite close to scale. The initial intent was to utilize diamond heatsinks between 0.010 and 0.013 in square for the 40-GHz work. This dimension was increased to 0.016 in in order to ease assembly problems and to accomodate diode and capacitor dimensions somewhat larger than anticipated. No identifiable spurious modes related to the increased size of the

Manuscript received April 13, 1981; revised August 10, 1981. This work was supported by the Avionics Laboratory, Air Force Wright Aeronautical Laboratory, Wright Patterson Air Force Base, OH.

The authors are with the Engineering Experiment Station, Georgia Institute of Technology, Atlanta, GA 30332.

0018-9480/81/1200-1266$00.75 ©1981 IEEE