Radiation Modes of Open Microstrip with Applications

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***Abstract*—The well known bound modes of open microstrip do not constitute a complete spectrum, for, continuous radiation and localized (reactive) modes are excited at discontinuities in microstrip circuits and antennas. This part of the spectrum has not been investigated before so that, up to date, radiation problems in microstrip are being investigated by nonmodal methods, such as the moment method. We derive here for the first time the complete spectrum of open microstrip, including one or more bound modes and a continuum, and demonstrate its straightforward application to a practical problem such as the excitation by a cylindrical probe of finite radius. Application of Lorentz's reciprocity principle is now possible in complete analogy to the problem of excitation of a close waveguide by a probe. Mode patterns, the equivalent circuit of a via-hole and its radiation pattern are characterized as a practical application of the foregoing theory.**

I. Introduction

UNLIKE closed waveguides that comprise an infinite, numerable spectrum of discrete modes, open waveguides comprise, possibly, a few discrete modes and a continuous spectrum besides.

If the cross section of the guide is one-dimensional or two-dimensional and separable, the complete spectrum can be found according to classical procedures [1], [2]. If instead the guide cross-section is two-dimensional and nonseparable, either closed or open, application of transverse resonance in the spectral [3] or in the space domain [4], or in the equivalent network form [5], [6] can always, in principle, yield its discrete spectrum.

Much less attention seems to have been paid so far to the more difficult question of the continuum of a nonseparable open cross-section [7], [8] such as that of open microstrip.

The nonavailability of this part of the spectrum has consequences in the study of discontinuity and radiation problems in open microstrip in as much as it prevents the adoption of modal techniques typical of closed waveguide.

For istance, the Spectral Domain Integral Equation (SDIE) approach does retain the open nature of the structure and has been applied to many practical components, [9]–[13]. Its effectiveness for many problems, however, is constrained by the necessity to model the fields over large interaction regions. This is a significant limitation, so that the fields at some distance before and after the discontinuity are often considered to consist solely of incident, reflected and transmitted fundamental modes. This assumption is valid for fairly well spaced planar circuits, however substrate modes and radiation fields only decay as the square root of the distance and, if significantly excited by the discontinuity, will invalidate this assumption. Moreover, it is cumbersome to extend these techniques to non planar discontinuities, e.g. via hole, probes [14].

By contrast, the excitation of modes by a probe in a close guide provides a classical example of the application of Lorentz' theorem [15]. By analogy, if the complete spectrum of open microstrip were available in usable form, the above class of problems would be amenable to a very efficient solution. It is, therefore, evident that a viable procedure for obtaining the continuous spectrum of open microstrip, now lacking, would be advantageous for the study of a wide class of problems, arising from either discontinuities in the line or when the latter is being operated as an antenna element.

It is the purpose of the present contribution to develop, in full hybrid form, the complete spectrum of open microstrip. Once this is developed, we demonstrate its application to the practical, nontrivial problem of excitation of microstrip by means of a probe of finite dimensions connecting ground plane and strip conductor (via-hole). Owing to the symmetry of the geometry, it is possible to derive in quasianalytical form the modal amplitude of each component of the complete spectrum of the line excited by the probe by application of Lorentz' theorem in a manner completely analogous to that used in determining the amplitude of each discrete mode in close waveguide.

II. Analysis, Complete Spectrum

Considering the microstrip cross-section of Fig. 1, we observe that this differs in two ways from the problem of a grounded dielectric slab guide

1) the fields are hybrid;
2) the geometry is essentially nonseparable due to the strip edges at $x = \pm a/2$.

As a consequence of 2), it is evident that any field in the cross-section has to be expanded as a Fourier integral in $k_x, (k_x^2 + k_y^2 = k_t^2)$, where two different Fourier components are coupled by diffraction at the strip edges (transverse diffraction).

The response of the guide to a source involves excitation of the discrete (bound) modes of the microstrip as well as of a continuum of modes, analogous in concept to those of the slab guide, but more complex in form because of the hybrid nature of the field and of the effect of transverse diffraction, that is k_x-mixing due to the effect of the strip. Such continuous modes

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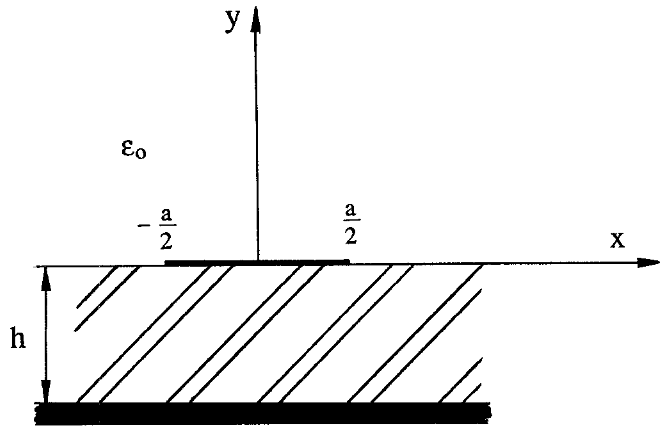

Fig. 1. Geometry of the microstrip line: cross section.

can be seen as wave packets in k_x, each packet corresponding to a fixed value of k_t which individually satisfy boundary and edge conditions on the strip, radiation at infinity, continuity of the transverse fields at the interfaces between dielectrics.

For a given k_t, there is still the possibility of degeneracy, that is the existence of a number of different strip currents and field configurations satisfying all boundary and edge conditions, pertaining to the same value of k_t, which are labeled by the discrete index μ.

Wave packets corresponding to different values of k_t and of the discrete index μ are mutually orthogonal over the cross-section, as defined below

$$\int_S \mathbf{E}_\mu(k_t)\mathbf{x}\mathbf{H}_\nu(k'_t)\cdot\mathbf{z}\,dS = \delta_{\mu\nu}\delta(k_t - k'_t) \tag{1}$$

where

$$k_t^2 = k_0^2 - \beta^2 \quad \text{being} \quad 0 < k_t < \infty.$$

In order for all k_x-components of a wave packet to travel down the guide with the same phase constant β, the phase shift α must be the same for all, i.e. a characteristic function of the packet $\alpha_\nu(k_t)$.

For each wave packet, labeled by k_t, ν, the y-directed LSM and LSE hertzian potentials in air and substrate regions, can be expressed as Fourier integrals in k_x with separate x and y dependence; for instance, for the even parity LSM potential in air we have

$$\Psi_\nu(x,y,k_t) = \frac{1}{j\omega\varepsilon_0}\int_0^\infty \tilde{I}_\nu(k_x,k_t)\Phi(k_x,x)\chi(y,k_x,k_t)\,dk_x \tag{2}$$

where

$$\Phi(k_x,x) = \sqrt{\frac{2}{\pi}}\cos(k_x x) \tag{3}$$

$$\chi(y,k_x,k_t) = \begin{cases} \sqrt{\frac{2}{\pi}}\cos(k_y y + \alpha_\nu) & : \quad 0 \le k_x \le k_t \\ \sqrt{\frac{2}{\pi}}\cos(\alpha_\nu)e^{-\gamma y} & : \quad k_t \le k_x \end{cases} \tag{4}$$

and

$$k_y^2 = k_t^2 - k_x^2; \quad \gamma^2 = k_x^2 - k_t^2; \quad q^2 = (\varepsilon_r - 1)k_0^2 + k_t^2 - k_x^2. \tag{5}$$

Analogous expressions hold for the substrate and for the LSE potential. We note the presence of a phase shift $\alpha_\nu(k_t)$ as yet undetermined in the y-dependence that can be seen physically as the phase delay of a wave in the y-direction impinging on the microstrip from the upper half plane. All fields, in fact, are evaluated from (2), once $\alpha_\nu(k_t)$ is determined. The purpose of the analysis is to set up an eigenvalue equation for α_ν; the eigenvector corresponding to the eigenvalue $\alpha_\nu(k_t)$ is a current distribution $\mathbf{J}_\nu = (J_{\nu x}, J_{\nu z})$ on the microstrip consistent with the potential (2).

By imposing the continuity of the tangential electric fields at the air-dielectric interface and the discontinuity of the transverse magnetic field on the strip given by

$$\mathbf{y}\times(\mathbf{H}_\nu^+ - \mathbf{H}_\nu^-) = \mathbf{J}_\nu \tag{6}$$

where $\mathbf{H}_\nu^+, \mathbf{H}_\nu^-$ are magnetic fields at $y = 0$ in air and in the substrate respectively, it is possible to evaluate the spectral amplitudes $\tilde{I}_\nu(k_x,k_t), \tilde{V}_\nu(k_x,k_t)$ which appear in the expressions of the potentials as combinations, via β and k_x, of the F-transforms of the currents on the strip; these are omitted for sake of brevity.

The condition of vanishing tangential electric field on the conductor

$$\mathbf{E}_\nu \bullet \mathbf{J}_\nu = 0 \tag{7}$$

yields then the sought eigenvalue equation for $\alpha_\nu(k_t)$.

Although the algebra of the theoretical development is complicated by the hybrid nature of the field, the practical evaluation of the spectrum α_ν for values sampled in the range $0 \le k_t < \infty$ can be quickly achieved using a Ritz-Galerkin technique and assuming $\mathbf{J}_\nu$ as a test current distribution.

In particular, the following expansions for the currents on the strip are used:

$$J_x = w_x(x)\sum_{n=1}^{N} I_{xn}f_{xn}(x) \tag{8a}$$

$$J_z = w_z(x)\sum_{n=1}^{N} I_{zn}f_{zn}(x) \tag{8b}$$

where

$$w_x(x) = \sqrt{1 - \left(\frac{2x}{a}\right)^2} \tag{9a}$$

$$w_z(x) = \frac{1}{\sqrt{1 - \left(\frac{2x}{a}\right)^2}} \tag{9b}$$

are the appropriate weight functions in order to take into account the behavior of the currents on the edge of the strip, I_{xn} and I_{zn} are the unknown amplitudes of each component, $f_{xn}(x), f_{zn}(x)$ the Chebyshev polynomials. In terms of (8) above, it is possible to reduce (7) to the following matrix equation:

$$\begin{bmatrix} Z_{xx} & Z_{xz} \\ Z_{zx} & Z_{zz} \end{bmatrix}\begin{bmatrix} I_x \\ jI_z \end{bmatrix} = 0 \tag{10}$$

where the Z-blocks are real matrices of the type

$$Z = P(\alpha_\nu) + Q \tag{11}$$

the term $P(\alpha_\nu)$ is obtained by integration in the spectral domain over the propagating part of the spectrum ($\beta < k_0$) while Q, independent of α_ν, is relative to the non propagating components of the spectrum; the expressions of the matrix elements are detailed in the Appendix.

The condition of vanishing determinant of the matrix $[Z]$ allows the evaluation of α_ν.

Owing to symmetry ($Z_{xz} = Z_{zx}$), it is possible to recover the following condition of orthogonality between eigenvectors corresponding to different eigenvalues

$$[I_\nu]_t[R][I_\mu] = 0 \tag{12}$$

where the matrix $R_{\nu\mu} = R(\alpha_\nu, \alpha_\mu)$ is defined by

$$(\operatorname{cotg}\alpha_\nu - \operatorname{cotg}\alpha_\mu)R(\alpha_\nu, \alpha_\mu) = P(\alpha_\nu) - P(\alpha_\mu). \tag{13}$$

For a given value of k_t, (12) is a statement of the orthogonality between modes corresponding to different values of the discrete index in (1), implying that two such modes do not exchange power on the strip. An interpolating curve for α_ν in the proper range of k_t is sufficient for the accurate description of the field.

III. Microstrip Discontinuity

The complete spectrum derived in the previous section allows us to recast radiation and discontinuity problems in microstrip on the same formal basis as problems involving mode excitation in classical waveguide where Lorentz' theorem can be applied.

As an application of the foregoing theory, we will consider the problem of a coaxial feed or the germane one of a via-hole in microstrip and demonstrate how knowledge of the complete spectrum allows us to retrace the conceptual steps of the closed waveguide case [15, Sections 5–6], leading now to a simple analytical formula for the far field and a variational expression of the shunt impedance modeling the post.

A. Mode Excitation

The configuration under study is shown in Fig. 2 and consists of a symmetrically placed conducting post connecting ground plane and strip conductor; it is assumed the via to be represented by an equivalent flat post having the same perimeter, and the current $\mathbf{J}$ on the surface of the post to be uniform and y-directed

$$\mathbf{J} = J_0\mathbf{y} \text{ (A/m)}. \tag{14}$$

Application of Lorentz' reciprocity principle yields the amplitudes of the forward, $C_0^+, C^+(k_t)$, and backward, $C_0^-, C^-(k_t)$, waves of the fundamental mode and of each wave packet of the continuous spectrum

$$-2C^+(k_t) = \int_{\text{Probe Surface}} \mathbf{E}^-(k_t)\cdot\mathbf{J}\,dS \tag{15a}$$

$$-2C^-(k_t) = \int_{\text{Probe Surface}} \mathbf{E}^+(k_t)\cdot\mathbf{J}\,dS. \tag{15b}$$

$\mathbf{E}^\pm(k_t)$ represents the total forward/backward travelling continuous E-field and can be easily obtained by (2); in this particular case only the E_y component of the substrate field is involved, giving

$$\begin{aligned} C^+(k_t) = C^-(k_t) &= -\langle \mathbf{J}, \mathbf{e}(k_t)\rangle \\ &= -\frac{1}{2}\int_0^\infty \tilde{E}_y^s(k_x, k_t)X(k_x, k_t)dk_x. \end{aligned} \tag{16}$$

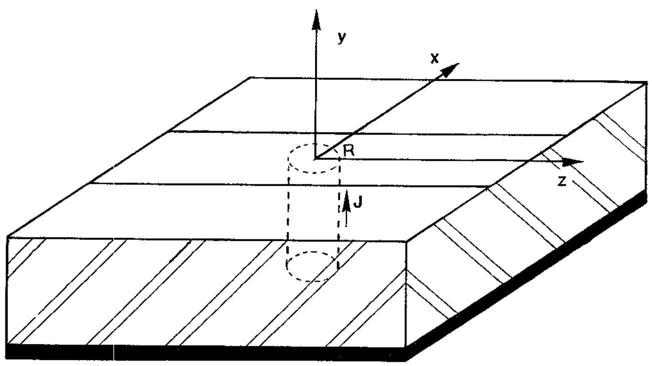

Fig. 2. Conducting post connecting ground plane and strip conductor: geometry.

$\tilde{E}_y^s(k_x, k_t)$ is a Fourier component of the modal distributions and $X(k_x, k_t)$ is the coupling coefficient, given by

$$X(k_x, k_t) = \frac{2}{\pi}wh\frac{\sin\left(\frac{k_x w}{2}\right)}{\frac{k_x w}{2}}\frac{\tan qh}{qh}. \tag{17}$$

The same formal expressions (15), (16) hold for the coefficients of the fundamental mode not reported here for brevity: it is enough to replace the field expressions of the continuum by those relative to the fundamental mode.

B. Variational Formulation for the Post Impedance

Retracing the conceptual steps of the closed waveguide case, [15], the total field in the guide is the sum of an incident and scattered field and must vanish on the perfecly conducting strip.

The resulting integral equation for the current is

$$(1+R)\mathbf{e}_0 = \int_0^\infty dk_t\langle \mathbf{J}, \mathbf{e}(k_t)\rangle \mathbf{e}(k_t) \tag{18}$$

where R is the reflection coefficient of the fundamental mode and $\mathbf{J}$ is the unknown current distribution on the post.

An expression for the shunt impedance Z modeling the post in the line can then be obtained by expressing R in terms of Z, yielding

$$\frac{Z}{Z_0} = \int_0^\infty dk_t\left[\frac{\langle \mathbf{J}; \mathbf{e}(k_t)\rangle}{\langle \mathbf{J}; \mathbf{e}_0\rangle}\right]^2 \tag{19}$$

Z_0 being the characteristic impedance of the line.

C. Radiated Field

Knowledge of the current on the post, upon application of the stationary phase method, allows us to recover the far field of the via in presence of the microstrip line. This is

$$E_x = -j\frac{\pi}{2}k_0C(k_t)e^{-j\alpha_\nu(k_t)}F_x(k_x)\sin\theta\cos\phi\frac{e^{-jk_0r}}{r} \tag{20a}$$

$$E_y = j\frac{\pi}{2}k_0C(k_t)e^{-j\alpha_\nu(k_t)}F_y(k_x)\sin\phi\cos\theta\frac{e^{-jk_0r}}{r} \tag{20b}$$

$$E_z = -j\frac{\pi}{2}k_0C(k_t)e^{-j\alpha_\nu(k_t)}F_z(k_x)\sin\phi\cos\theta\frac{e^{-jk_0r}}{r} \tag{20c}$$

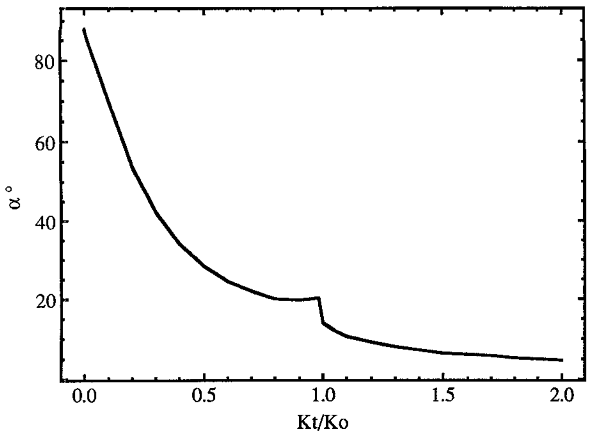

Fig. 3. First eigenvalue α (degree) as a function of the transverse wavenumber.

where

$$k_t = k_0 \sin(\theta), \quad k_x = k_0 \sin\theta \cos\phi.$$

$C(k_t)$ is the wave packet amplitude as given by (16) and functions F are linear combinations of the amplitudes of the potentials. Far field components can be evaluated by changing to polar coordinates

$$\begin{bmatrix} E_\theta \\ E_\phi \end{bmatrix} = \begin{bmatrix} \cos\theta\cos\phi & \cos\theta\sin\phi & -\sin\theta \\ -\sin\phi & \cos\phi & 0 \end{bmatrix} \begin{bmatrix} E_x \\ E_y \\ E_z \end{bmatrix}. \quad (21)$$

It is noted that the dependences on elevation (θ), azimuthal angle (ϕ), and distance (r) are separated in the above formula. Consequently, the far field is seen to be a product of the microstrip mode pattern $F(\theta, \phi)$ and of that of the probe pattern.

IV. Results

A. Field Characterization

A numerical program has been developed to solve (10) and a microstrip line with the following characteristics has been analyzed: strip width $a = 2.8$ mm, substrate thickness $h = 1.5$ mm, substrate relative permittivity $\varepsilon_r = 4.5$, negligible strip thickness, which give a strip characteristic impedance $Z_0 = 50\ \Omega$.

In Fig. 3 the eigenvalue α as a function of the transverse wavenumber is shown. It is noted the presence of a discontinuous derivative at $k_t = k_0$ arising from the wave number (1).

We derived the modal field distribution for $k_t/k_0 = 0.6$: Fig. 4 shows the amplitude of the E_y electric field component close to the strip, for $I_{z0} = 1$ A/m: it is easily verified that this component satisfies its boundary conditions at the air-dielectric interface and at the strip edges. The same holds true for the remaining components not reported here.

Another and more severe check has been performed on a derived quantity such as the charge density on the strip: the surface charge density can be easily evaluated from (8) as well as from the discontinuity of the E_y field component at $y = 0$.

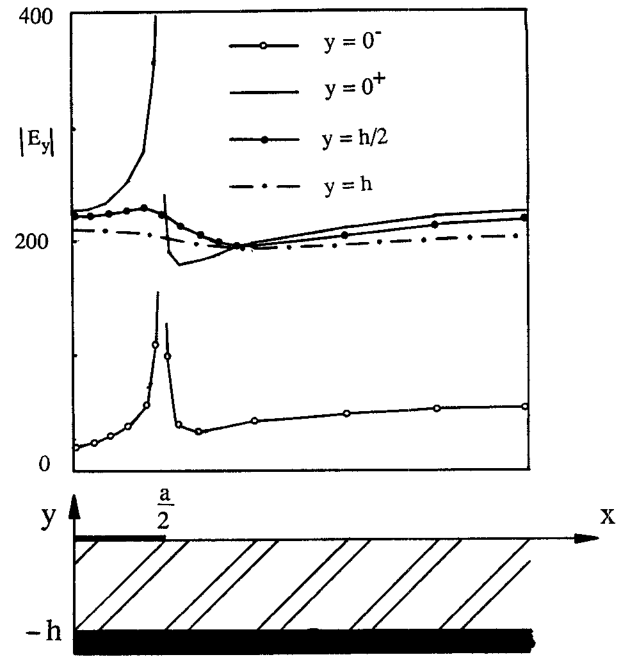

Fig. 4. Modulus of the E_y field component close to the strip.

At the centre of the strip, the two results differ by about 9.5 percent; the agreement is good, also considering that

1) the test is relative to the E_y field component about which no condition has been enforced;
2) just one expansion function for each current has been used.

Finally, Fig. 5 shows the distributions of the amplitudes of the transverse electric field over the cross-section of the radiation mode: it is evident the different nature of the radiation mode with respect to that of the bound mode.

Whereas the latter is seen to correspond to a quasistatic field pattern, the former distribution shows clearly the transverse standing wave character of the continuous mode. Even in this case all boundary and edge conditions are satisfied, however the field amplitude behaves in the proper fashion of a radiation field in both transverse directions.

B. Application to a Via-Hole Ground

The analysis developed in Section III has been applied to simulate a via-hole on GaAs produced and measured by GEC Marconi and represented in Fig. 6: it is assumed that the flat post equivalent to the actual cone shaped via has the same perimeter at its mid height section. Although from a rigorous viewpoint the quantities appearing in (19) need to be computed for each frequency, nonetheless, for practical purposes, it is sufficient to evaluate the L, R elements at midband and then these values can be used also at different frequencies with a modest error.

A prediction of the transmission coefficient derived by means of the present variational formulation is compared in Fig. 7 with measurements and with results obtained by a 3-D mode matching technique [16].

It is also noted that, while the mode-matching simulation of [16] has been developed for a boxed structure, the present one has been developed for an open microstrip. As a consequence, the formulation of [16] naturally takes into account

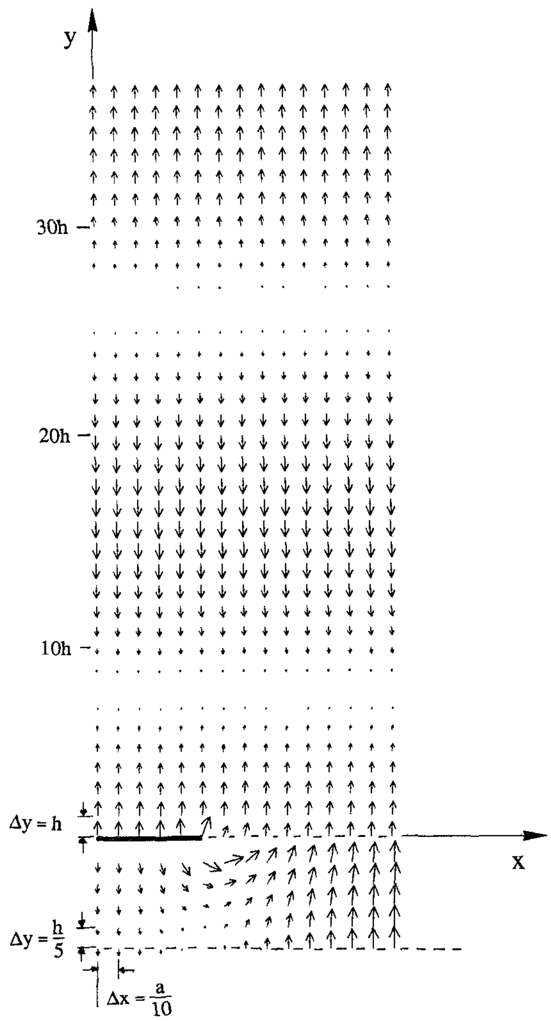

Fig. 5. Near field pattern over the cross-section of the radiation mode ($\frac{k_t}{k_0} = 0.6$).

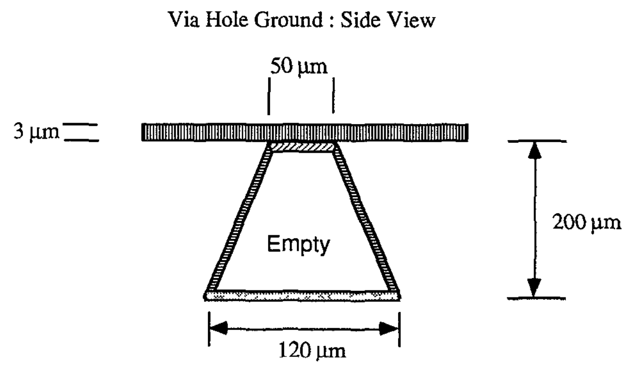

Fig. 6. Scheme of the microstrip via hole ground on GaAs ($\varepsilon_r = 12.9$). The microstrip width is $a = 570$ μm.

package interactions, while the rigorous equivalent circuit here introduced directly accounts for radiative losses.

The results of our equivalent circuit is also compared in the same figure with theoretical data computed by a commercial software package [17]. A more detailed discussion of the equivalent radiation resistance, reported in Fig. 8, and inductance of the post can be found in [18], [19].

Finally we computed the far field excited by the finite post, as given by (21), with amplitudes derived as described in

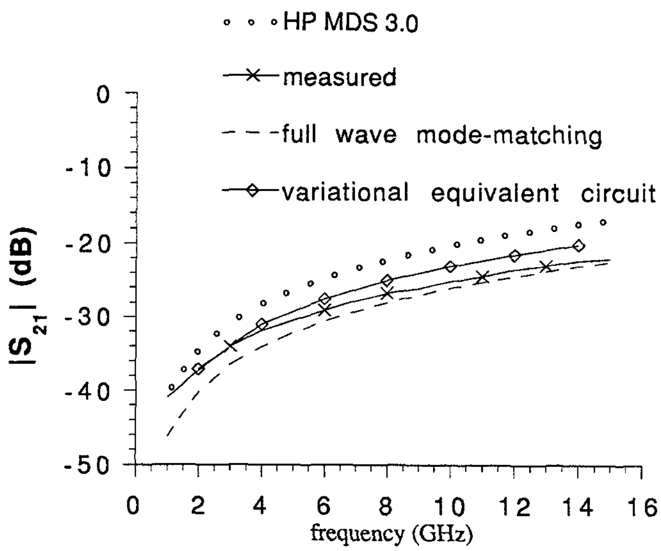

Fig. 7. Comparison of measured data with our equivalent circuit for a via hole on GaAs. Also shown are full wave results relative to the boxed case [16] and to a commercial model of the via.

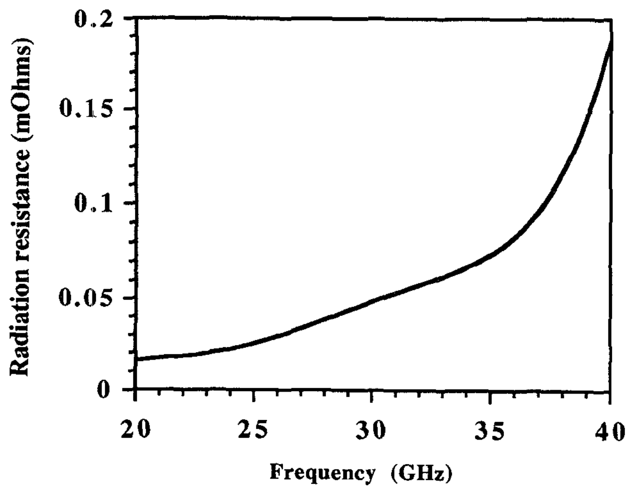

Fig. 8. Radiation resistance of the post as a function of frequency.

Section III; calculations for the longitudinal plane are plotted in Fig. 9. This symmetrical radiation pattern is characterized by the presence of two main lobes at $\theta \cong 10°, 75°$.

On the plane x-z the radiation is also concentrated around the forward direction, as in a leaky wave antenna. We expect therefore considerable interaction due to this mode with neighboring circuit elements in line with the strip, whereas radiation from a probe without the strip would be essentially omnidirectional.

The computer program runs on a μVAX 3600; most of CPU time (about 3 hours) is required to generate the eigenvalue curve $\alpha(k_t)$: this is due to the fact that the integral (19) for the evaluation of the impedance converges for a high value of the upper limit of integration ($k_t \sim 80 - 100$ k_0). Once this data is filed, the calculation of the equivalent circuit requires a few seconds per frequency point.

V. Conclusion

For the first time, we have derived the complete spectrum of open microstrip. This knowledge allows one to solve

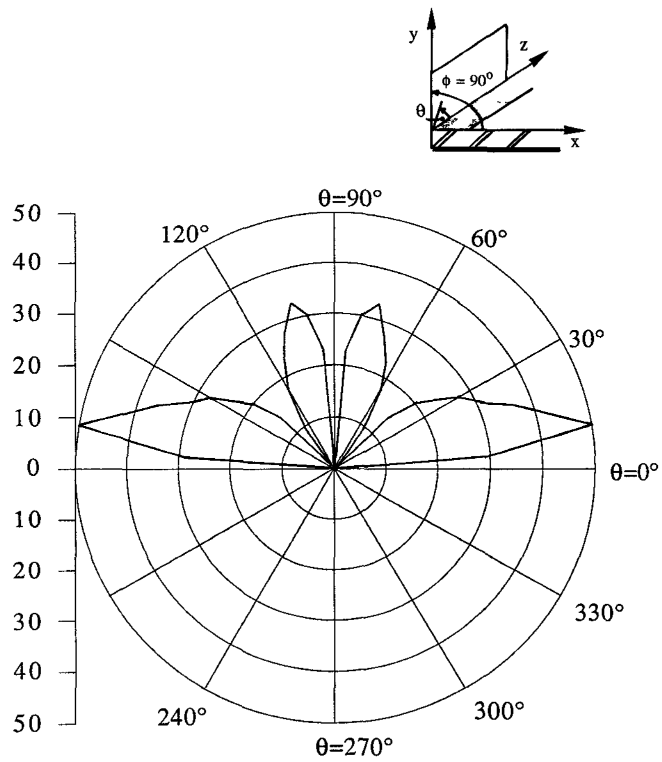

Fig. 9. Radiation pattern: $E_\theta(\phi = 90°)$ field.

microstrip discontinuities and radiation problems in a manner conceptually analogous to that of determining modal excitation by a source in a classical waveguide.

A nontrivial application is given to the case of a probe of finite thickness connecting strip and ground plane, i.e. either a coaxial excitation of microstrip or a short circuit realized by a post. Numerical results are presented for the equivalent circuit of the post and compared with experiment and other existing data showing moreover how far the presence of the strip influences the radiation pattern of the probe.

Appendix

After obtaining the electric fields from potentials (2), and using the expansion (8) for the currents, the condition (7) leads to the following expressions of the Z blocks of (10), according to the notation of (11)

$$P^{mn}_{x_i x_j}(\alpha_\nu) = \int_0^{k_t} dk_x \langle w_{x_i}, f_{x_i n}, \Phi_{p_i}\rangle A_{x_i x_j} \langle w_{x_j}, f_{x_j m}, \Phi_{p_j}\rangle \quad i = 1,2;\ j = 1,2 \quad \text{(A1)}$$

$$Q^{mn}_{x_i x_j} = \int_{k_t}^{\infty} dk_x \langle w_{x_i}, f_{x_i n}, \Phi_{p_i}\rangle a_{x_i x_j} \langle w_{x_j}, f_{x_j m}, \Phi_{p_j}\rangle \quad i = 1,2;\ j = 1,2 \quad \text{(A2)}$$

where

$$x_1 = x,\ x_2 = z,\ p_1 = h,\ p_2 = e;$$

and

$$A_{xx} = \frac{k_y \tan\alpha_\nu(\varepsilon_r k_0^2 - k_x^2) - q\tan qh(k_0^2 - k_x^2)}{D_{\mathrm{TE}} D_{\mathrm{TM}}} \quad \text{(A3)}$$

$$a_{xx} = \frac{\gamma(\varepsilon_r k_0^2 - k_x^2) - q\tan qh(k_0^2 - k_x^2)}{d_{\mathrm{TE}} d_{\mathrm{TM}}} \quad \text{(A4)}$$

$$A_{xz} = \frac{\beta k_x(k_y \tan\alpha_\nu - q\tan qh)}{D_{\mathrm{TE}} D_{\mathrm{TM}}} = A_{zx} \quad \text{(A5)}$$

$$a_{xz} = \frac{\beta k_x(\gamma - q\tan qh)}{d_{\mathrm{TE}} d_{\mathrm{TM}}} = a_{zx} \quad \text{(A6)}$$

$$A_{zz} = \frac{k_y \tan\alpha_\nu(\varepsilon_r k_0^2 - \beta^2) - q\tan qh(k_0^2 - \beta^2)}{D_{\mathrm{TE}} D_{\mathrm{TM}}} \quad \text{(A7)}$$

$$a_{zz} = \frac{\gamma(\varepsilon_r k_0^2 - \beta^2) - q\tan qh(k_0^2 - \beta^2)}{d_{\mathrm{TE}} d_{\mathrm{TM}}}. \quad \text{(A8)}$$

Functions $D_{\mathrm{TE}}, D_{\mathrm{TM}}, d_{\mathrm{TE}}, d_{\mathrm{TM}}$ are given below

$$D_{\mathrm{TE}} = -k_y \cot\alpha_\nu(k_t) + q\cot qh \quad \text{(A9)}$$

$$D_{\mathrm{TM}} = \varepsilon_r k_y \tan\alpha_\nu(k_t) - q\tan qh \quad \text{(A10)}$$

$$d_{\mathrm{TE}} = \gamma + q\cot qh \quad \text{(A11)}$$

$$d_{\mathrm{TM}} = \varepsilon_r\gamma - q\tan qh. \quad \text{(A12)}$$

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