RADIATION MODES OF SLOTLINE WITH APPLICATION TO MILLIMETRIC CIRCUITS

T.Rozzi, G. Gerini, M. Righi

Dipartimento di Elettronica e Automatica, Università di Ancona
Via delle Brecce Bianche-60100 ANCONA-ITALY

Abstract – **Bound modes of slotline have been extensively investigated [1]. Due to its open nature, slotline also features a continuum of radiation modes both in air and into the dielectric substrate that are excited by discontinuities such as steps, bends, short circuits, etc.**
Moreover, transverse metal strips across the slot form array patterns in slotline antennas, coupling strongly to radiation modes. Such fields propagate without attenuation a long way away from their source and are difficult to handle by standard numerical techniques.

This part of the spectrum has never been investigated before. This contribution deals with the continuum of slotline in full hybrid form: its simple application to the problem of transverse metal strip is then demonstrated providing details of radiation patterns. The cascade of strips, including interaction via the radiation field, is being currently investigated.

I. INTRODUCTION

Although slotline is often used as an open transmission media, even for antenna feeds, its radiation modes pose a considerable problem and have escaped attention so far.

In absence of information over the full spectrum, non-modal [2,3], computer-intensive techniques are currently used, often under various effort-reducing approximations, in order to study discontinuities problems and radiation patterns in open planar circuits in the microwave range. Such techniques work effectively in situations where energy can be considered as being carried by just the bound modes at a reasonably short range (in terms of wavelength) away from the source of excitation; their effectiveness deteriorates with the square of the linear dimensions of the region where the near field or current have to be tested.

Radiation in air and into the substrate, however, is particularly important in the millimetric regime both as a pervasive parasitic effect in presence of discontinuities and as a design feature, e.g. beam-forming in planar antennas.

Modal techniques, with modes usable in practice, offer in principle, many theoretical and practical advantages, including orthogonality, the implicit satisfaction of the guide boundary conditions and the availability of Lorentz's theorem, among others. The continuum of a grounded slab is, in fact, well known. The problem in finding the continuum of slotline is due to the non-separability of its cross-section, involving transverse diffraction. Constructing a continuum, in this case, consists in building wave packets, each characterized by a transverse wavenumber K_t ($0 \leq K_t < \infty$), a transverse field distribution, labelled by a discrete index ν, each packet individually satisfying all dielectric boundary and singular edge conditions imposed by the metallization in the near field as well as the radiation condition in the far field.

Moreover, each packet is orthogonal to all others packets and to the bound modes in the standard manner :

$$\iint_{Cr.\ Sect.} (\mathbf{E}_\nu(x,y;K_t) \times \mathbf{H}^*_\mu(x,y;K_t')) \, dS = \delta_{\nu\mu} \, \delta(K_t - K_t') \quad (1)$$

Consequently all the modal properties of closed waveguide are recovered. For instance, it is straightforward to compute the excitation of slotline by a given source and its radiation.

As a significant example, we analyze the scattering by a transverse metal strip across the slot.

II. OUTLINE OF THEORY

We derived the continuous modes, just as discrete, bound modes, by using y-directed electric and magnetic Hertzian potentials, whose expressions are chosen to satisfy the correct boundary conditions for the fields. In particular we represented each wave packet in the form of a Fourier integral over Kx and we introduced a phase shift $\alpha_\nu(K_t)$ in the expression of potentials for the air region. This angle has the physical interpretation of phase shift that all Fourier components of the same packet undergo upon incidence on the aperture discontinuity at y=0 when considering transverse propagation in the y-direction. By imposing the continuity of the transverse fields on the aperture (see fig. 1a,b) we obtain and a generalized eigenvalue equation whose eigenvalue is $\cot g\alpha_\nu(K_t)$.

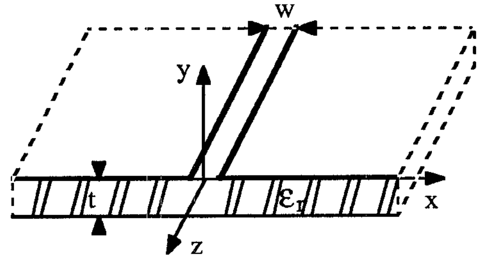

Fig.1a) Slotline

K

CH3141-9/92/0000-0503$01.00 © 1992 IEEE

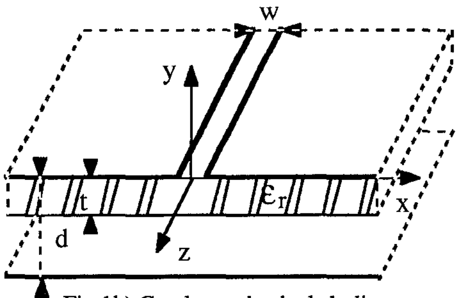

Fig.1b) Conductor-backed slotline

For a given real value of the transverse wavenumber K_t, the continuous modes satisfy the following equation:

$$\begin{bmatrix} Y^d_{11}(K_t)-Y^a_{11}(K_t) & Y^d_{12}(K_t)-Y^a_{12}(K_t) \\ Y^d_{21}(K_t)-Y^a_{21}(K_t) & Y^d_{22}(K_t)-Y^a_{22}(K_t) \end{bmatrix} \begin{bmatrix} E_{x0} \\ jw\frac{\partial}{\partial x}E_{z0} \end{bmatrix} = \underline{0} \tag{2}$$

where $Y^{d;a}_{ij}$ are the integral admittance operator for the dielectric region and the air region respectively,that relate the tangential components of the electric field with those of the magnetic field at the interface; E_{x0}, $jw\frac{\partial}{\partial x}E_{z0}$ denote the distribution of field on the aperture and are the eigenfunctions corresponding to K_t and ν. In particular in the air region we have:

$$Y^a_{ij} = B + G\ \text{cotg}\alpha \tag{3}$$

where B is the non-radiative part of the integral transverse admittance operator, while G is the radiative one, in particular they can be represented in standard manner as in [4] :

G=\i(0,K_t,\b($Y^{TM}(K_y)\cos^2(\theta)+Y^{TE}(K_y)\sin^2(\theta))\Phi(x,K_x)\Phi(x',K_x)dK_x$)

(4)

B has the same expression but it is integrated over the interval $K_t \div \infty$.

$$Y^{TM}(K_y) = \frac{\omega\varepsilon_0}{K_y} \quad (5a) \qquad Y^{TE}(K_y) = \frac{K_y}{\omega\mu_0} \quad (5b)$$

are the admittance of TM and TE plane waves and

$$\cos(\theta) = \frac{K_x}{\sqrt{K_x^2+\beta^2}} \quad (6a) \qquad \sin(\theta) = \frac{\beta}{\sqrt{K_x^2+\beta^2}}$$

(6b)

are the relations that define the angle of rotation about the direction of propagation to resolve a hybrid field into pure TM and TE components.
Then (2) can be written in the following general form:

$$\underline{\underline{A}}\cdot\underline{v} = \lambda\ \underline{\underline{C}}\cdot\underline{v} \tag{7}$$

The numerical solution of (2) is carried out in the space domain by the Galerkin's method. We use $\frac{\partial}{\partial x}E_{z0}$ instead of E_{z0} because it satisfies the same singular edge condition of E_{x0} and then they can be represented with the same set of expanding functions. The slot edges give a discontinuity of a square root type and then we model the interface electric field by means of Chebyshev polynomials orthonormal over the aperture ($|x| < W/2$) with the weight function:

$$W(x) = \left(1 - \left(\frac{2x}{w}\right)^2\right)^{-1/2} \tag{8}$$

that correctly represents the singularity. From the field distribution at the interface, it is possible to deduce by standards methods its distribution over the guide cross section, including the far field.

III. Results on the radiation modes

The continuum of both slotline and conductor-backed slotline (fig. 1 a,b) have been investigated. In the first case, in particular, two positions of the source are possible, either above or below the ground plane.

Eigenvalues (modes) of (2) can be classified in either of two types:

1) finite cotg α ($\alpha < 180^o$)

2) infinite cotg α , or very large anyhow ($\alpha \approx 180^o$)

The former represent waves penetrating into and diffracted by the slot, whereas the latter characterize waves not dissimilar to those supported by an infinite ground plane with or without dielectric slab.

Fig. 2,3 report cotg α (K_t) for varying K_t for a slot of dimensions w=2 mm and 1 mm, respectively. It is noted that the number of finite eigenvalues (slot modes) increases with increasing K_t more rapidly for a broader slot.

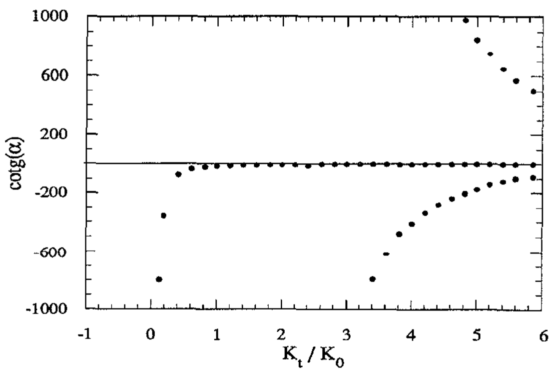

Fig. 2 Eigenvalue behaviour with K_t
[w=2mm]

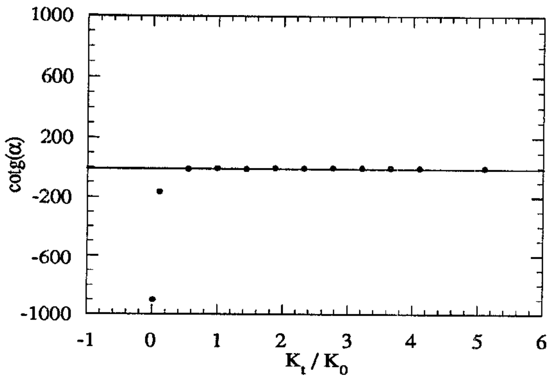

Fig. 3 Eigenvalue behaviour with K_t
[w=1mm]

To each eigenvalue is associated an interface field showing, expectedly, larger transverse variations for broader slots, as shown in fig.4 and 5a÷c.

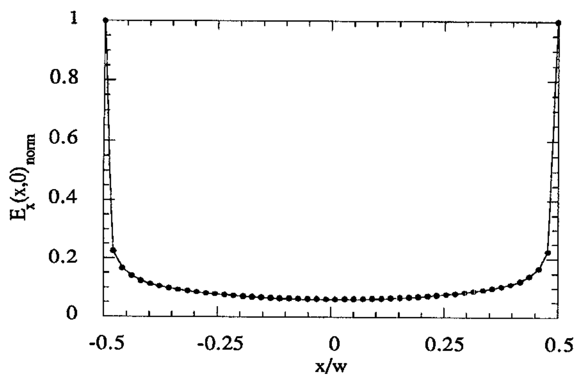

Fig. 4) Interface configuration for Ex
[w=1mm ; cotg(α_1)=-2.28]

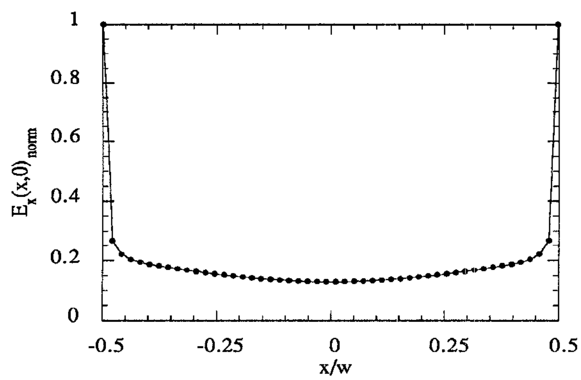

Fig. 5a) Interface configuration for Ex
[w=2mm ; cotg(α_1)=129.5]

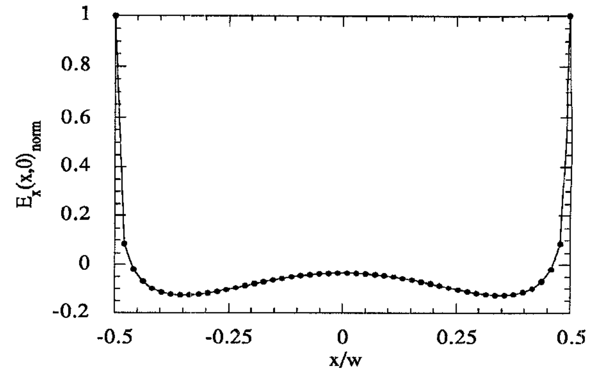

Fig. 5b) Interface configuration for Ex
[w=2mm ; cotg(α_2)=-27.9]

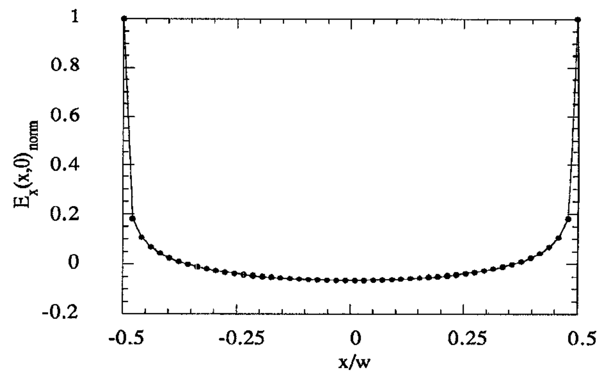

Fig. 5c) Interface configuration for Ex
[w=2mm ; cotg(α_3)=-0.71]

The second type of modes (air modes) carries very little power with respect to the first type.

IV. APPLICATION TO TRANSVERSE STRIP

As a simple application we determined the far field radiation pattern of a transverse metal strip at the air-dielectric interface. We imposed a simple current behaviour on the strip, that satisfies the correct edge conditions at the strip ends ($z = \pm t/2$):

$$J_x(x,z) = J_0 \left(1 - \left(\frac{2z}{t}\right)^2\right)^{-1/2} \quad (9)$$

We obtain the excitation coefficient of an individual wave packet, $A_\nu^{\pm}(K_t)$, by the current source (9) by means of Lorentz's theorem:

$$A_\nu^{\pm}(K_t) = -\frac{1}{2\,N(K_t)} \iint_S E_{x\nu}(x\,,y\,;K_t)\cdot J(x\,,z\,)\, e^{\pm j\beta z}\, dS \quad (10)$$

where $E_{xv}(x , y ; K_t)$ is the wave packet, and $N(K_t)$ is the orthonormalization coefficient obtained imposing (1).
The total radiated field is then obtained as the superposition of all packets by integrating over K_t:

$$E_x^{\pm}(x,y,z)= \int_0^{\infty} A_v^{\pm}(K_t)\, E_{xv}(x , y ; K_t)\, e^{\pm j\beta z}\, dK_t \qquad (11)$$

The analytical evaluation of this integral is efficiently carried out by means of the stationary phase method.

In fig. 6-7 are shown the far field distributions in the longitudinal plane for different slot widths and for different positions of the ground plane. In fig. 8 is shown the far field distribution in the transverse plane for w=2mm, d=5mm and f=10GHz. We can note that variations of the slot width (we have considered width greater than those common used) do not affect the shape of the radiation pattern both in longitudinal and transverse plane, on the contrary the variation of the ground plane position sensibly affects the radiation pattern in the longitudinal section.

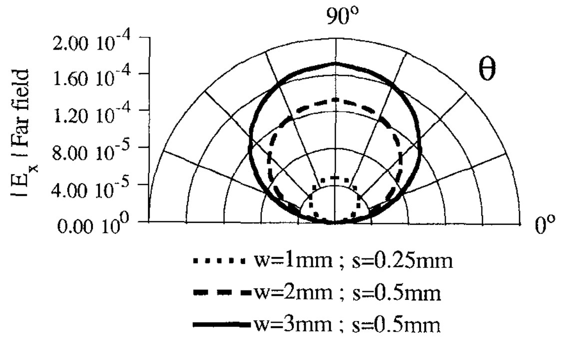

Fig. 6) Far field pattern of Ex [$0^o \le \theta \le 180^o$ $\Phi = 90^o$]
s is the strip width, d=5mm

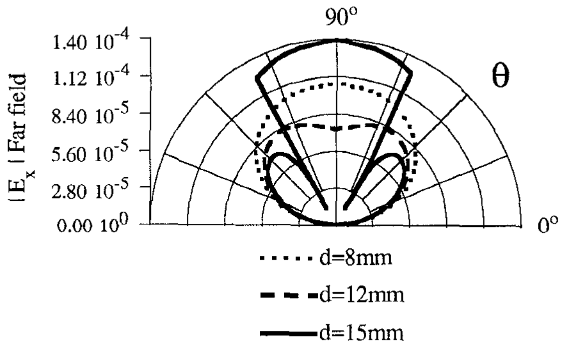

Fig. 7) Far field pattern of Ex [$0^o \le \theta \le 180^o$ $\Phi = 90^o$]
w=2mm, s=0.5mm

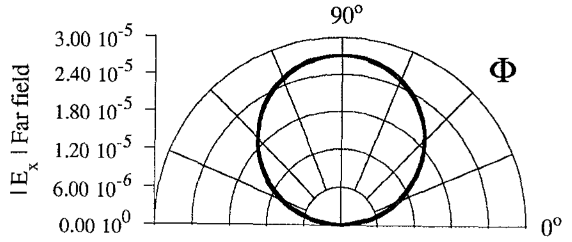

Fig. 8) Far field pattern of Ex [$0^o \le \Phi \le 180^o$ $\theta = 90^o$]
w=5mm, d=5mm, s=0.5mm

V. Conclusion

In this contribution, we have derived the radiation spectrum of slotline in full hybrid form. This part of the spectrum had not been investigated before.

Its application to the practical problem of scattering by a transverse strip across the slot is demonstrated, producing detailed near field and and far field distributions with a minimum of computational effort.

References

[1] T.Rozzi, F.Moglie, A.Morini, E.Marchionna, M.Politi: "Hybrid modes, substrate leakage, and losses of slotline at millimeter-wave frequencies" IEEE Trans. Microwave Theory and Tech., vol 38, pp.1069-1078, Aug. 1990

[2] Hung-Yu Yang, N.G.Alexopoulos: "A dynamic model for microstrip-slotline transition and related structures" IEEE Trans. Microwave Theory and Tech., vol 36, pp.286-292, Feb. 1988

[3] T.Rozzi, A.Morini, A.Pallotta, F.Moglie: "A modified dynamic model for planar microwave circuits" IEEE Trans. Microwave Theory and Tech., Dec. 1990

[4] T. Itoh: "Dispersion characteristic of generalized printed transmission lines" IEEE Trans. Microwave Theory and Tech., vol.28, pp733-736, July 1980