RECENT PROGRESS IN COMPUTING CAPABILITIES OF MILLIMETRE WAVE INTEGRATED PASSIVE COMPONENTS

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SUMMARY

In this paper, several computational schemes, applicable to passive components to be used in millimetre wave circuitry, are discussed.

These schemes lie on concepts and techniques which are in most-cases, well known but rarely experimented on integrated components due to the importance of the numerical effort to produce.

INTRODUCTION

Millimetre wave passive integrated components can be etched either in the so-called "E-plane" or "H-plane" of a metallic box with a rectangular cross section.

As far as the "E-plane" components are concerned, the inside dimensions of the rectangular cross section are those of the standard rectangular waveguide (Waveguide Technology).

Metallic boxes, supporting the "H-plane" components, have inside dimensions much more important and ensure, in most cases, their conservation (Microstrip Technology).

By assembling the passive components together with active components (Diodes, Transistors) one elaborates the whole or a part of millimetre wave subsystems (Front-ends, Amplifiers, Oscillators, Mixers ...) of future Satellite or Radio link communication systems.

Most of passive components are achieved by juxtaposition of several transmission media. These transmission media can be constituted by strip and/or slot arrangements which are etched on one or two sides of a lossless dielectric substrate as shown in Fig. 1.

One can distinguish between "H-plane" transmission media and "E-plane" transmission media following the slow wave or the fast wave nature of the propagating fundamental mode. Therefore, as described in Fig. 2-a, the Fineline is a "E-plane" transmission medium ($\lambda_g/\lambda_o>1$) when the slotline (Fig. 2-b) is a "H-plane" transmission medium ($\lambda_g/\lambda_o<1$).

This above distinction between transmission media specifies the excitation procedure that must necessarily induce a field interaction between propagating waves of the same nature by virtue of basic coupling laws. As a result, "E-plane" transmission media are better excited with waveguide transition when coaxial transitions are well suitable for excitation of "H-plane" transmission media.

The whole or partial modeling of a given passive component lies on the availability of an exact solution of fullwave analyses of transmission media on one hand, of elementary discontinuities (which are introduced between them (voluntarily or unvoluntarily) on the other hand.

Such a modeling appears finally as the combination of very complex analyses which, in most cases today, is still either non-existent or in a rough sketch state.

Beside the necessary modeling, it is often requested a synthesis accompanied by an optimisation of the component specifications. So, facing these formidable problems which obsess the every day life of circuit designers, the computer appears as an irreplaceable robot partner.

This paper discusses some recent new capabilities of standard numerical techniques applicable to "H-plane" transmission media and also uniaxial and/or multiaxial elementary discontinuities of both types.

SPECTRAL DOMAIN TECHNIQUE FOR "H-PLANE" TRANSMISSION MEDIA

An "H-plane" transmission medium can be seen as an engineering model of an open transmission medium. The Spectral Domain Technique is capable to describe numerically both continuous and discrete modes in such a waveguide. The discrete mode description in (1) is a direct adaptation of that originally applied on discrete modes in "E-plane" transmission medium (2). Roughly speaking, the continuous modes in open transmission media could approximated as the limiting form of fast discrete modes in "H-plane" transmission media if one ignores the obvious difference between boundary conditions at large distance of the substrate.

A key point of the description of continuous modes lies in the specification of the integral representation of their longitudinal field outside the substrate written as

$$\begin{Bmatrix} E_z(x,y) \\ H_z(x,y) \end{Bmatrix} = \frac{1}{2\pi} \int_{-\infty}^{+\infty} \begin{Bmatrix} \tilde{E}_z(\alpha,y) \\ \tilde{H}_z(\alpha,y) \end{Bmatrix} e^{-j\alpha x} \, d\alpha$$

where the spectral densities $\tilde{E}_z(\alpha,y)$ and $\tilde{H}_z(\alpha,y)$ are expressed as the superposition of two spectral plane waves progressing in y directions

$$\begin{Bmatrix} \tilde{E}_z(\alpha,y) \\ \tilde{H}_z(\alpha,y) \end{Bmatrix} = \begin{Bmatrix} A(\alpha) \\ A'(\alpha) \end{Bmatrix} e^{-j\gamma y} + \begin{Bmatrix} B(\alpha) \\ B'(\alpha) \end{Bmatrix} e^{j\gamma y}$$

with the wavenumber γ defined as

$$\gamma = \pm\ (\rho^2 - \alpha^2)^{1/2}$$

As shown in Fig. 3, the integration path in the spectral domain can remain the real axis of the complex α-plane because 1) it does not cross over the the branch cuts of the double-valued γ function and 2) it belongs to the Riemann sheet where the radiation condition ($I_m\ \gamma \geqslant 0$) is satisfied.

The continuous modes are specified by $\rho^2 > 0$ so that the integration path sweeps for $0 < |\alpha| < \rho$ the visible range ($\gamma^2 > 0$) and for $\rho \leqslant |\alpha| < \infty$ the invisible range ($\gamma^2 \leqslant 0$) of the two-dimensional plane wave representation (3).

Another key point of the description lies in the computation of the complex infinite power flow of an individual continuous mode. It indicates a degeneracy (order 4 in a slotline) and suggests a possible description of their modal fields by hybrid diffracted waves emanating from TE or TM filamentary sources at large distance of the transmission medium in y directions as illustrated in Fig. 4.

Up these points, the formulation becomes almost identical for discrete and continuous modes. The case of continuous modes remains peculiar in that the boundary conditions must be apply separately in visible and invisible ranges.

Some results on slotline without and with substrate are reported in (4) and (5) respectively.

It is clear that this recent extension of the Spectral Domain Technique gives now a lot of perspectives for computing "H-plane" type discontinuities.

COMBINATION OF THE SPECTRAL DOMAIN TECHNIQUE WITH THE MODAL ANALYSYS FOR COMPUTING UNIAXIAL "E-PLANE" DISCONTINUITIES

The basic idea, is to solve the excitation problems in each port reference plane by the Spectral Domain Technique and after the diffraction problems in both junction and ports, up to the reference planes, by the Modal Analysis (6).

The Modal Analysis starts from functional equations involving the tangential fields in the two ports 1 and 2 in the transverse junction plane z =0 (see Fig. 5). On S_A which belongs to both cross sections S_1 and S_2, they are written as

$$\vec{E}_T^1(x,y) = \vec{E}_T^2(x,y)$$

$$\vec{H}_T^1(x,y) = \vec{H}_T^2(x,y)$$

and on S_C which belongs only to the cross section S_1, they are written as

$$\vec{E}_T^2(x,y) = 0$$

$$\vec{H}_T^2(x,y) + \vec{J}_T^1(x,y) \times \vec{u} = 0$$

where $\vec{J}_T^1$ (x,y) is the current surface on S_C.

The Totality principlc generates incident and reflected modal amplitudes toward and from the junction plane.

Orthogonality transforms the above functional equations in a linear set from which the generalized scattering matrix is derived.

It is finally shown that coefficients of the linear set can be efficiently computed directly in the spectral domain.

A lot of results, including an experimental validation on finline impedance step and tapering finline to waveguide transition can be found in (7) (8) and (9).

Let note that this scheme can be readily reproduced on "H-plane" type uniaxial discontinuities by adding the continuous mode description in the excitation problems as already done on more academic configurations(10) (11).

COMBINATION OF THE SPECTRAL DOMAIN TECHNIQUE WITH THE FINITE ELEMENT METHOD FOR COMPUTING MULTIAXIAL "E-PLANE" DISCONTINUITIES

A typical example of multiaxial "E-plane" discontinuity is a finline Te junction which plays an important role in the design of SPDT devices (Fig.6).

Here, the basic idea, which has been already revealed in (12) is to solve the excitation problems in each port reference plane by the Spectral Domain Technique and after the diffraction problems in both the junction and ports, up to the reference planes, by the Finite Element Method.

By noting V, the entire volume including the junction and ports up to the reference planes, the electric field at any point of a given region i is derived from the minimization of the functional

$$F(E_i) = \int_V (\vec{\nabla} \times \vec{E}_i)(\vec{\nabla} \times \vec{E}_i) - \varepsilon_{ri}\omega^2\mu_o\varepsilon_o\vec{E}_i\vec{E}_i\ dV$$

Determination of the scattering matrix is then done as follows. One prescribes the transverse electric field in the port j at the reference plane J_j very close to the junction as

$$\vec{E}_{Tj} = (a_j + b_j)\ \vec{e}_{Tj}^{\,1}\ e^{j\phi_j} + \ldots.$$
.. evanescent modes

Orthogonality in port j yields

$$a_j + b_j = \int_{S_j} (\vec{E}_{Tj} \times \vec{h}_j^{1*})\ e^{-j\phi_j}\ \vec{u}_j\ dS$$

where $\vec{E}_{Tj}$ is computed by the Finite Element Method and $\vec{h}_{Tj}^{1*}$ by the Spectral Domain Technique.

In the port j but now in a reference plane P_j far away from the junction, the transverse electric field is prescribed as

$$\vec{E}_{Tj} = A_j^1\ \vec{e}_{Tj}^{\,1}$$

with

$$A_j^1 = (a_j\ e^{j\beta_j^1 l_j} + b_j\ e^{-j\beta_j^1 l_j})e^{j\phi_j}$$

where β_j are derived again from the Spectral Domain Technique.

From above relations, in the port j, it becomes possible to compute the reflection coefficient ρ_j at the reference plane P_j by

$$\frac{A_j^1}{e^{j\beta_j l_j} + \rho_j e^{-j\beta_j l_j}} = \frac{1 + \rho_j}{\int_{S_j} (\vec{E}_{Tj} \times \vec{h}_{Tj}^{1*}) \, \vec{u}_j \, dS}$$

where $A_j^1 = 1$ if the mode 1 is normalized.

So, making a copy of this derivation for other port one can identify the S scattering matrix elements from the determinantal equation

$$\begin{vmatrix} (S_{11} - \rho_1) & S_{12} & S_{13} \\ S_{21} & (S_{22} - \rho_2) & S_{23} \\ S_{31} & S_{32} & (S_{33} - \rho_3) \end{vmatrix} = 0$$

Preliminary results on finline impedance step computed by this numerical scheme are presented in a regular session of this conference (13).

INTEGRAL EQUATION TECHNIQUES FOR COMPUTING "H-PLANE" DISCONTINUITIES

Such full wave analyses have been recently resuscitaded by printed antenna studies especially those dealing with electromagnetically coupled microstrip dipoles (14).

On the general microstrip configuration with two dielectric substrates outlined in Fig. 7, an interal equation for current and associates charge densities can be written as

$$\vec{e}_z \times [\sum_{j=1,2} \iint_{S_j} (j\omega \bar{\bar{G}}_{A_{ij}} (\vec{r}_i/\vec{r}_j) . \vec{J}_{S_j}(\vec{r}_j) + G_{V_{ij}} (\vec{r}_i/\vec{r}_j) . \rho_{S_j}(\vec{r}_j)) \, dS' + Z_S \, \vec{J}_{S_i} (\vec{r}_i)] = \vec{e}_z \times \vec{E}_i^{(e)} (\vec{r}_i)$$

where $\bar{\bar{G}}_{A_{ij}}$ is the diadic Green's function of the vector potential $\vec{A}$ on conductor i due to source j

$G_{V_{ij}}$ is the Green's function of the scalar potential V

$\vec{J}_{S_j}$ is the surface current density on conductor j

ρ_{S_j} is the associated surface charge density

$\vec{r}_i$ locates the observation point (i = 1 upper conductor ; i = 2 bottom conductor)

$\vec{r}_j$ locates the source point (j = 1 upper conductor ; j = 2 bottom conductor)

Z_S is the surface impedance (ohmic losses)

$\vec{E}_i^{(e)}$ is the excitation electric field.

The Green's functions relevant to this problem are given by Sommerfeld-type integrals, which require special integration techniques, and can be found in (15) or (16).

The unknowns $\vec{J}_S$ and ρ_S are related through the continuity equation, so that the only unknown quantity is $\vec{J}_S$, which is determined by Moment's method.

By assuming that upper and bottom conductors are strips parallel this to the x direction this technique allows to derive the scattering matrix of their uniaxial junction (Fig. 8).

A gap generator excites one strip while the second is loated by the charge ρ_L representing its abrupt interruption. Then, the calculated current forms standing waves of a TEM-loke mode far from the strip coupling region and far from the generator on the excited port. By transmission line concepts one can then compute the reflection coefficient at any plane of the TEM zone as

$$\rho_1 = S_{11} + \frac{S_{12} S_{21}}{(1/\rho_L - S_{22})}$$

The main advantage of this scheme lies in its versatility since it can be also applied to multiaxial configurations. Disadvantages lie in large CPU time and storage requirement especially for the necessary accurate computation of Green's functions. In any case this technique has good potentialities for the future.

CONCLUSION

In this paper, four numerical schemes able to solve accurately full wave analyses of "E-plane" and "H-plane" type discontinuities are discussed. All the presented schemes are exact in the numerical sense, thus applicable to the modeling of any passive components in millimetre wave circuitry.

ACKNOWLEDGMENTS

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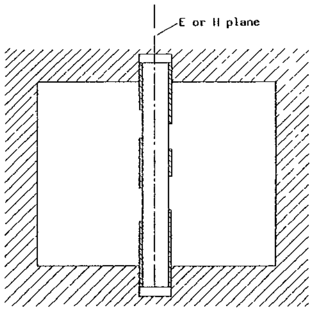
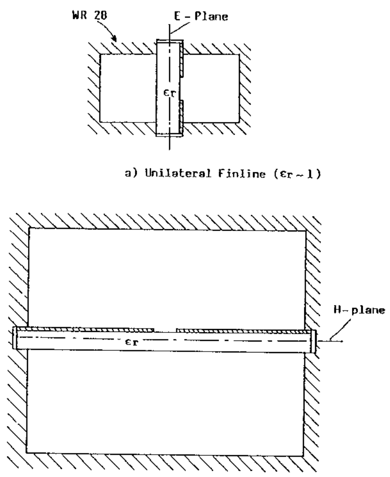


FIG 1. General configuration of a millimetre wave transmission medium.



a) Unilateral Finline (εr ~ 1)

b) (Slot line εr ~ 10)

FIG 2 Examples of transmission media

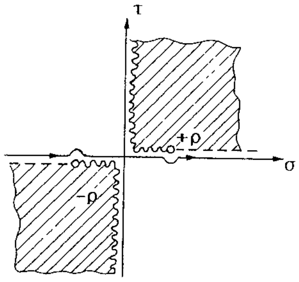


FIG 3. Integration path of the integral representation for continuous modes.

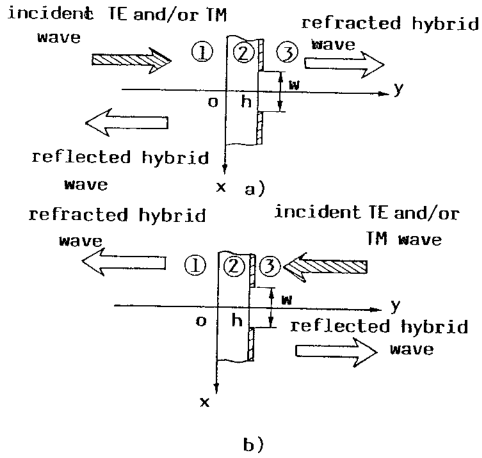


FIG 4. Generation of continuous modes by diffraction in the spectral domain.

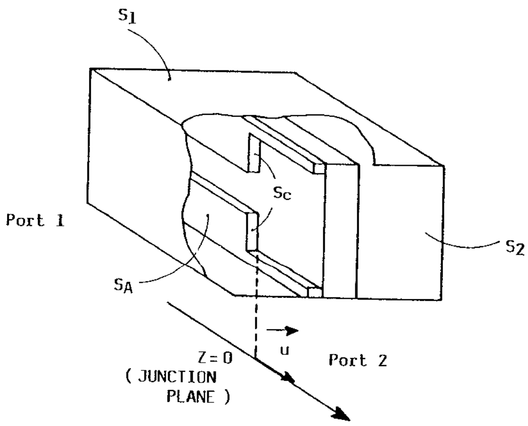


FIG 5. Impedance step in finline.

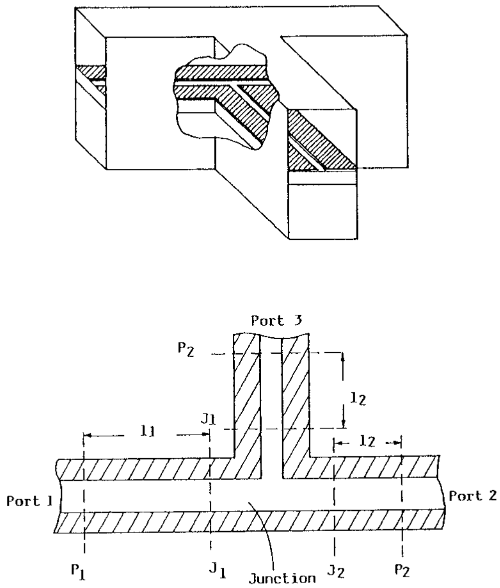


FIG 6. Te-junction in finline. Perspective and top views

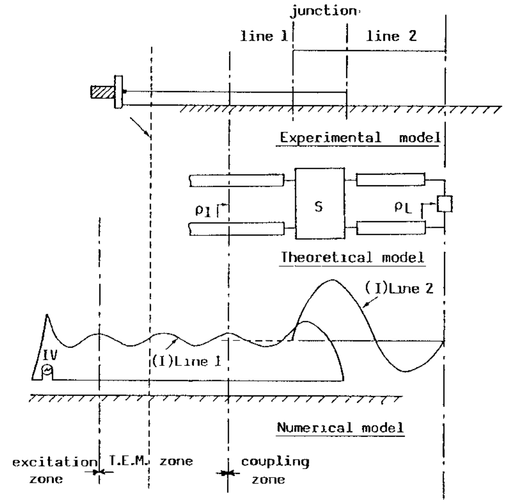


FIG 8. Integral Equation Technique scheme for " H-plane " discontinuities.

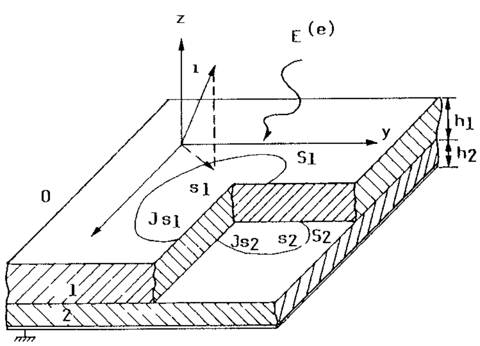


FIG 7. General microstrip configuration with two dielectric substrates.