TOWARDS AN UNIFIED EFFICIENT ALGORITHM FOR CHARACTERIZING THE PLANAR PERIODIC WAVEGUIDES

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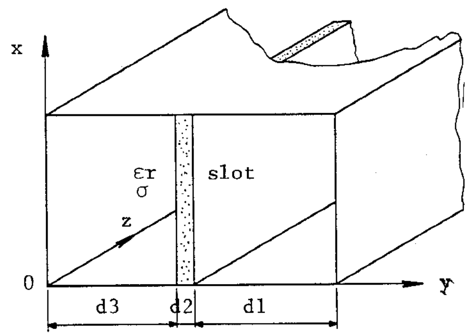
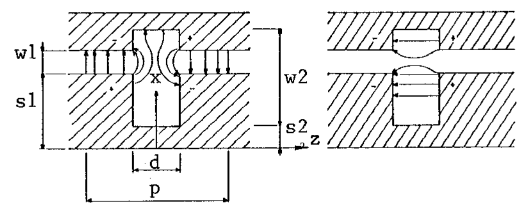
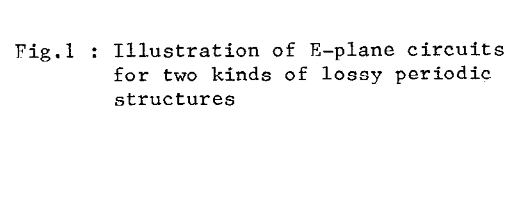
ABSTRACT

A new efficient algorithm (modified three dimensional spectral domain solution with "modal spectrum" in propagation direction) applied to a variety of planar waveguides with periodic stubs is achieved. In this paper, slow-wave propagation characteristics of both symmetrically and unsymmetrically loaded periodic structures with lossy dielectric layer such as finline and coplanar waveguide (CPW) are investigated. Many important features like pass-band and stop-band related to cut-off and resonant frequencies are discussed in detail based on numerical computations which are compared with measured results obtained by transmission line experimental procedure.





(a) : Periodic finline with arbitrary located stubs



(b) : Periodic CPW line with symmetric loaded stubs

Fig.1 : Illustration of E-plane circuits for two kinds of lossy periodic structures

INTRODUCTION

Coplanar waveguide (CPW) and finline M.I.S (Metal-Insulator-Semiconductor) structures are proposed and analysed by several authors [1∿5] for realizing phase shifters, delay lines and electronically variable filters.
They allow reducing significantly the component dimensions due to the slow-wave propagation. Monolithic technology can be used to realize high compact microwave and millimeter wave circuits.
However, the question about an efficient slow-wave mode excitation will arise wherein.
It is well known that periodically loaded transmission line could generate the slowing-down of propagation in a certain band.
In addition, the periodic structures have received considerable attention for their application to wider-bandwith couplers [6] and high-quality filters. Recently, the network analytical method has been employed to investigate theoretically the passband qand shopband properties of periodically loaded striplines and finlines [7]. Besides, a new efficient hybrid solution to these structures has been reported together with experimental structures. [8,9].
In this paper, a modified three dimensional spectral domain approach is presented to analyze the characteristics of planar periodically loaded structure. It should be noted that a new concept called "Modal spectrum" with respect to harmonic waves in propagation direction is introduced in this analysis, which leads to a considerable alleviation of the analytical formulation and of numerical computation of eigenproblems.
Slow-wave properties as well as passband and stop-band dehaviours related to the cut-off and resonant frequencies are discussed on numerical results, for both symmetrically and unsymmetrically loaded structures.

THEORETICAL FORMULATION

In the following, the principle of the modified three dimensional spectral domain approach will be illustrated for two kinds of lossy periodic structures shown in figure 1. The loss consideration of dielectric layer is to some extent because the periodic conductor strips could be placed on the lossy semiconductor substrate in the case of interconnexion with other monolithic chips. Here, it is assumed that metallizations have vanishing thickness and that the periodic stubs extend to infinity in the $\pm z$ directions.

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The electromagnetic field in each homogeneous region is described by two scalar potentials : ψ_e (LSM) and ψ_n (LSE) satisfying the Helmholtz equation and the boundary conditions at the shielding and symmetry walls. We can express the LSE/LSM potentials in each region by using the Fourier Transform in the x-direction and the Floquet harmonic representation in the z-direction : $\beta n = \beta o + 2n\pi/p$, where βo is the dominant mode propagation constant and $2n\pi/p$ are higher-order harmonics due tc periodic stubs, which can readily be considered as "modal spectrum" in the Fourier sense, i.e. the harmonic wave variation ir guided propagation direction could be regarded as natural Fourier development in half periodic cell [P/2].

Taking this fact into account, we can derive a set of comprehensive form as follows : ($\xi_n = 2n\pi/p$)

$$\psi^e_{(i)} = 1/\sqrt{\alpha^2 + \beta_n^2}\,[\sin \alpha x.\cos \xi_n z - j.\sin \alpha x \sin \xi_n z]\,\phi^e_{(i)}(y)\,e^{-j\beta oz} \tag{1}$$

$$\psi^h_{(i)} = 1/\sqrt{\alpha^2 + \beta_n^2}\,[\cos \alpha x.\cos \xi_n z - j.\cos \alpha x \sin \xi_n z]\,\phi^h_{(i)}(y)\,e^{-j\beta oz}$$

$i = 1, 2, 3...$

This formulation explains a very interesting physical phenomenon called "higher-order resonant harmonic decoupling" (see figure 2, which means that even-odd harmonics in z-variation could readily be separated regadless of dominant harmonic propagation in periodic field representation. This procedure provides us with a powerful two-dimensional (x-z) Fourier Transform tool. It should be pointec out that even and odd harmonics correspond to fictious magnetic anc electric walls respectively. Of course, the dominant harmonic component β_0 may always occur unless at cut-off frequency.

As previously mentionned, the two scalar potentials $\psi^e_{(i)}$ and $\psi^h_{(i)}$ must satisfy the Helmotz equation. So the modal LSM/LSE inhomogeneous transmission line equation (to y) can be set in the Fourier domain. All succeeding steps [10] appear to be similar to Itoh's immitance procedure.

Nevertheless, the field components can be expanded in the concrete analytical form along our analysis.

We easily obtain a set of algebraic coupled equations by straightfoward way :

$$\begin{vmatrix} \tilde{Y}_{xx}(\alpha, \xi_n, \beta_0) & \tilde{Y}_{xz}(\alpha, \xi_n, \beta_0) \\ \tilde{Y}_{zx}(\alpha, \xi_n, \beta_0) & \tilde{Y}_{zz}(\alpha, \xi_n, \beta_0) \end{vmatrix} \cdot \begin{vmatrix} \tilde{E}_x \\ \tilde{E}_z \end{vmatrix} = \begin{vmatrix} \tilde{J}_x \\ \tilde{J}_z \end{vmatrix} \tag{2}$$

Unlike the network analysis method [7] for the investigation of periodic planar line, the unknown aperture field can efficiently be divided into two directional field components (x-z) and also, the field quantities are directly expressed in terms of Fourier series, thus it is more convenient to apply the Galerkin's procedure in this method toward an unified efficient algorithm.

Up to this stage, we have to select carefully the basis functions according to the bidimensional boundary conditions.

According to the field polarization (see figure 1) in the aperture, the basis functions to be used for the TE mode differ from those foı the T.M mode. In any case, we can define "guided" basis functions

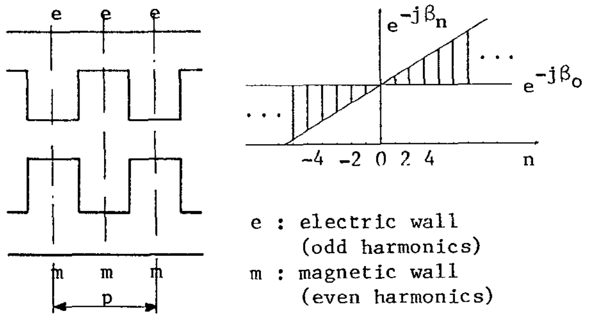

Fig.2 : Linear superposition of even and odd harmonics in single periodic cell.

in longitudinal region and "stored" basis functions in transverse region by taking Z- harmonic coupling into account.

In this paper, we set the basis functions in two subregions in terms of the multiplication of f(x) and g(z) :

$$E_x = \sum_i^{N_{x1}} \sum_j^{N_{z1}} a^I_{ij}\, f^I_{xi}(x)\, g^I_{xj}(z) + \sum_i^{N_{x2}} \sum_j^{N_{z2}} a^{II}_{ij}\, f^{II}_{xi}(x)\, g^{II}_{xj}(z)$$

$$E_z = \sum_i^{N_{x1}} \sum_j^{N_{z1}} b^I_{ij}\, f^I_{zi}(x)\, g^I_{zj}(z) + \sum_i^{N_{x1}} \sum_j^{N_{z1}} b^{II}_{ij}\, f^{II}_{zi}(x)\, g^{II}_{zj}(z) \tag{3}$$

where the superscripts I, II denote the subregion represented in the aperture, and a_{ij}, b_{ij} are the unknown coefficients.

Taking the Fourier development and inner products, a non-trivial solution for the propagation constant can be obtained by setting the determinant of coefficient matrix equal to zero.

Being ready to apply the Galerkin's Technique, such basis functions can be taken as a set of completely orthonormal series. In this paper, the familiar triangular function form based on Itoh's argument (no edge term for g(z)) are used for high speed computation without spurious solutions. The choice of basis functions depends not only on the bounday conditions but also on the propagation mode behaviour. (The TE mode and/or TEM mode for single and coupled slots are dominant along these structures. It is noted that TM mode should occur only in the resonant state). This consideration can ensure both magnetic walls at $z = \pm P/2$ and $S_1 < x < S_1 + W_1$. The fast convergence behaviour can be observed by using a few basis function number for most of the cases ($N_{x1,2} = N_{z1,2} = 2$ for spectral term $m = 20 \sim 30$ and modal spectrum $n = 5$)

NUMERICAL RESULTS :

Numerical results given in the following are obtained for d_1 = 8 mm, d_2 = 0,66 mm, d_3 = 14,2 mm in WR-90 waveguide.

A main principle for obtaining a slowwave is to store the electric and magnetic energy separately in space. Thereby the M.I.S (transverse space operation) structures and periodic structures (longitudinal space operation) are employed to generate the slowing-down of propagation in a certain frequency range.

In this paper, the slowwave phenomenon observed in the pass band by both experimental and theoretical analysis could be explained as the coupling of ± higher order mode in each periodic cell (electric and magnetic energy to be concentrated respectively in smaller slo (w_1) and larger slot (W_2)).

At and beyond resonant frequency point, all periodic cells can effectively be regarded as cascade coupled cavities where the stub plays a significant role.

Figure 3 illustrates dispersion characteristics of periodic finline with arbitrary located stubs. The comparison between measured and calculated results shows a very good agreement over the passband range, which valids this method. It can be seen that the passband is limited by two points : cut-off and resonant frequencies due tho shielded waveguide and periodic stubs. The former seems to be constant (approximatively equal to that of corresponding uniform structure). Indeed the influence of periodic stubs becomes negligible near the cut-off point.

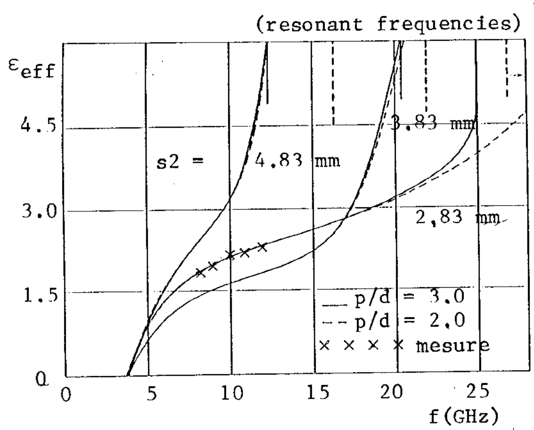

Fig.3 : Dispersion characteristics of periodic finline with arbitrary located stubs.
wl = 0.5, w2 = 4.5, sl = 4.83 (mm)
p×d = 3×1 (mm), ε_r = 2.22

On the other hand, by moving the stubs form S_2 = 2,83 mm (symmetric case) to S_2 = 4,83 mm (offset case) the resonant frequency goes down considerably.

Another interesting phenomenon is that the resonant frequency can effectively be changed by adjusting the period length without varying the dispersion characteristics over the passband range unless the frequency is in the shadow of resonance. The resonance phenomenon arises in two cases :

$$(1) \quad \begin{matrix} S_1 - S_2 \\ \text{and/or} \\ w_2 - w_1 - s_1 \end{matrix} = C\,(2k-1)/4\,.\lambda \qquad (4)$$

$$(2) \quad p = n\,\lambda g/2 \quad (k, n = 1, 2, 3......)$$

The coefficient C is determined by geometric conditions. It can easily be seen that passbands and stopbands will occur periodically with the frequency.

From results plotted on figure 4 several applications can be considered. For example, using long stubs (w_2/w_1 ↑), very narrow selective filters can be obtained because the resonant frequency closes the cutoff frequency.

Another possibility : for small values of w_2/w_1 (and small S_1) dispersion of odd mode tends to that of even mode all over the passband, so, broad bandwidth directional couplers become realizable.

Figure 5 shows the substrate losses as a function of frequency. In general losses are always in accordance with the propagation constant (increasing with frequency). However, for odd modes, a minimum can be observed near cut-off frequency. It is remarked

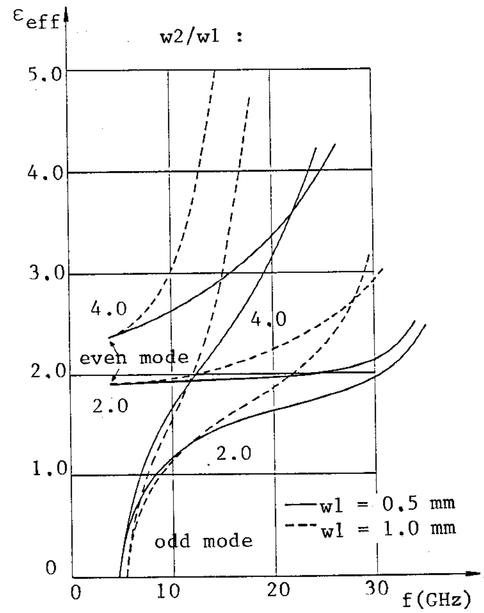

Fig.4 : Mode behaviours of the CPW structure as a function of frequency
wl = 0.5 mm, sl = 4.08 mm, ε_{eff} = 2.22
p × d = 3.0 × 1.0 (mm)

that optimal propagation (minimal losses) approaches ε_{eff} = 1.

CONCLUSION :

A new concept called "modal spectrum" in propagation direction has been introduced and successfully applied in the theoretical analysis. It makes possible to use directly the three dimensional spectral domain approach in both symmetrically and unsymmetrically loaded periodic structures. Several examples based on this unified algorithm show the slow-wave phenomenon as well as passband and stopband behaviour related to the cut-off and resonant frequencies. The dielectric losses can be involded.

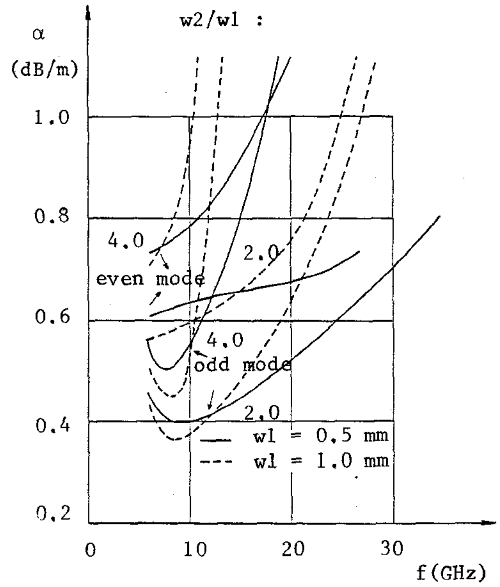

Fig.5 : Propagation loss of the CPW structure as a function of frequency for different ratio w2/w1. (σ = 0.001 1/Ωm) (the same conditions as Fig.4)

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