EFFICIENT ANALYSIS OF FINLINE WITH FINITE METALLIZATION THICKNESS

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ABSTRACT

The transverse resonance method is used for deriving dispersion characteristics of fundamental and higher order modes in finline with arbitrary slot widths and locations with regard to the finite thickness of metallization. In addition to the propagation constant, the characteristic impedance is calculated on the basis of the power-voltage definition.

INTRODUCTION

Finline structures have found frequent applications in millimeter-wave components. The problem of finline propagation characteristics has been treated by many exact and approximate methods. The most general, rigorous and systematic one is the mode matching technique, which has been used by Vahldieck (1) to show the influence of metallization thickness and mounting grooves on these characteristics. The analysis of finlines with finite mettalization thickness utilizing the network analytical method for electromagnetic fields has been presented in (2). The spectral domain technique for an idealized finline structures (zero metallization thickness and absence of mounting grooves) has been described, e.g., in (3)-(5). The idealized structure of this transmission line has also been analyzed in terms of the singular integral equation technique (6).

All the four rigorous methods are based on the hybrid mode formulation and they are complicated, computer time consuming and cumbersome. On the other hand, the need for the design of finline circuits in a lucid and tractable manner has caused the development of closed-form design equations like, e.g., in (7),(8). However, the design equations are appropriated for the idealized finline structure and accurate enough only for a definite range of its parameters. Similar disadvantades can be found in the analysis reported in (9) and based on the transverse resonance method (TRM) which was previously used for slotline by Cohn (10).

Until recently there have been no efficient analysis of finline with finite metallization thickness, and less complicated than the above mentioned rigorous mathods. The purpose of this paper is to present such a formulation of TRM, which should partly fill in this gap. The proposed method used for the lossless unilateral finline shown in Fig.1 will be presented here as an example illustrating the problem. Slot and dielectric layers are arbitrarily located in waveguide housing.

BASIS OF SOLUTION

The dispersion characteristics of modes in finline can be obtained by solving the boundary value problem of the half - wavelength finline resonator. Hence, the length of resonator in the x direction (Fig.1), equal to π/β_x (β_x-phase constant for fundamental or higher order mode in finline), is an unknown function of the resonant frequency f. The analysis is limited to such modes in finline which are excited (in the x direction) by an incident TE_{10}-wave of the empty waveguide.

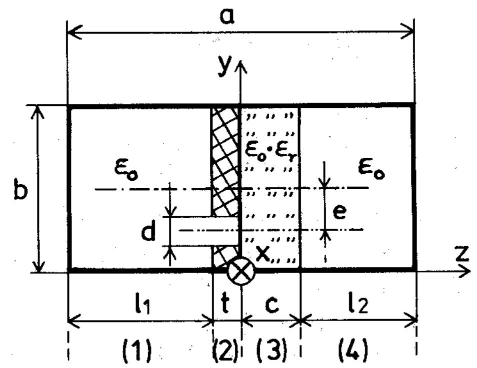

Fig.1: Unilateral finline.

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The electromagnetic fields in the four regions shown in Fig.1 are superpositions of waveguide modes "propagated" in the z direction, which have π variation in the x direction when the half-wavelength resonator is considered. Therefore, the full set of modes satisfying the boundary conditions is as follows: TE_{1n} (to z axis) for n an integer $n \geq 0$ and TM_{1n} for $n \geq 1$. The coupling of TE- and TM-modes is automatically involved by assumption that only the y component of the E-field exists in the slot at z=0 (T_0 plane). Additionally, the analysis is simplified by assuming that only TE_{10}-mode is present in the region (2). The amplitude of each mode in each region must be such, that when the full set of modes is superimposed, the boundary and continuity conditions in the planes: T_2 and T_0 will be met, and all fields components in plane T_3 will be matched. That is an equivalent condition that transverse resonance will occur in the circuit presented in Fig.2. The boundary value problem has been expressed by the following real equation:

$$B_t = B_+ + B_- = 0 \qquad (1)$$

where B_t is the total susceptance in the plane T_0, and B_+, B_- are susceptances in this plane looking in the +z and -z directions, respectively.

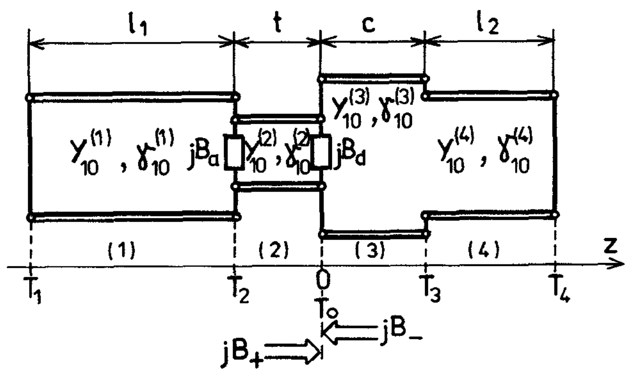

Fig.2: Equivalent circuit for transverse section of half-wavelength unilateral finline resonator.

The B_t is a function of f and β_x for the given set of parameters: ε_r, a, b, c, d, e, t, l_1 and l_2 which are indicated in Fig.1. The roots of equation (1), at a certain frequency, are phase constants of the fundamental and higher order modes in finline. It is self-evident that the above equation has no solution when $\beta_x = \beta_0$ (β_0-free space phase constant). However, since propagation characteristics in finline are monotonic, their values, at the discontinuity points of equation (1), can be calculated with ordered accuracy.

FORMULAS FOR ELEMENTS OF THE RESONATOR EQUIVALENT CIRCUIT

Components B_+ and B_- of the resonance equation are easily derived by using the conventional circuit theory, and the elements of the circuit in Fig.2 are analytically determined below. This circuit consists of the segments of TE_{10}-mode air- or dielectric-filled waveguide, which are described by propagation constants $\gamma_{10}^{(i)}$ i=1,2,3,4 and characteristic admittances $Y_{10}^{(i)}$ based on the power-voltage definition. These parameters are defined as follows:

$$\left(\gamma_{10}^{(i)}\right)^2 = \begin{cases} \varepsilon_r \cdot \gamma_0^2 - \gamma_x^2 & \text{region (3)} \\ \gamma_0^2 - \gamma_x^2 & \text{otherwise} \end{cases}$$

$$Y_{10}^{(i)} = \begin{cases} \frac{1}{2\eta_0} \cdot \frac{b}{d} \cdot \frac{\gamma_y \cdot \gamma_{10}^{(i)}}{\gamma_x \cdot \gamma_0} & \text{region (2)} \\ \frac{1}{2\eta_0} \cdot \frac{\gamma_y \cdot \gamma_{10}^{(i)}}{\gamma_x \cdot \gamma_0} & \text{otherwise} \end{cases}$$

where $\gamma_0 = j \cdot \beta_0$, $\gamma_x = j \cdot \beta_x$, $\eta_0 = \sqrt{\mu_0 / \varepsilon_0}$, $\gamma_y = j\beta_y = j\frac{\pi}{b}$.

The susceptance B_a represents all higher order modes (TE_{1n}, TM_{1n} for $n \geq 1$) in the T_2 plane from region (1). These modes in regions (3) and (4) are equalized in the plane T_0 by the susceptance B_d. Both equivalent susceptances have been derived by using a properly modified variational technique, which has been applied previously by Collin (11) for the capacitive aperture in waveguide.

According to the earlier assumptions, there is only the y component of the E-field in the plane T_2. The electromagnetic field in region (1) is TE to x axis and can be constructed as the superposition of TEx_{1n} modes. Fig.3 shows the network representation of these hybrid modes for $n \geq 1$.

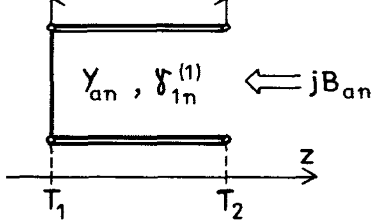

Fig.3: Equivalent network for higher order waveguide modes in region (1).

Their characteristic admittances based on the power-voltage definition are as follows:

$$Y_{an} = Y_{10}^{(1)} \cdot \gamma_{10}^{(1)} / \gamma_{1n}^{(1)}$$

and the mode propagation constants are expressed as

$$\left(\gamma_{1n}^{(1)}\right)^2 = \gamma_o^2 - \gamma_x^2 - \left(n \cdot \gamma_y\right)^2 .$$

The E-field in the slot at plane T_2 is constant along the y axis, therefore susceptance B_a is described by the equation:

$$B_a = B_{aG} + 2 \cdot \sum_{n=1} \left[B_{an} - \frac{1}{n} \cdot Y_{10}^{(1)} \cdot \frac{\gamma_{10}^{(1)}}{\gamma_y} \right] \cdot P_{n0}^2 \qquad (2)$$

with $B_{aG} = Y_{10}^{(1)} \cdot \frac{\gamma_{10}^{(1)}}{\gamma_y} \cdot (-\ln \alpha_2)$

where $\alpha_2 = \sin\left(\frac{\pi}{2} \cdot \frac{d}{b}\right) \cdot \sin\left[\frac{\pi}{2} \cdot \left(1 - 2 \cdot \frac{e}{b}\right)\right]$,

and B_{an} is the input susceptance of the TEx_{1n}-mode as in Fig.3. Coefficients P_{n0} can be obtained from the tables of Chebyshev polynomials.

The formula for B_d is as follows:

$$B_d = B_{dG} + 2 \cdot \sum_{n=1} \left[B_{dn} - \frac{1}{n} \cdot Y_{10}^{(3)} \cdot \frac{\gamma_{10}^{(3)}}{\gamma_y} \right] \cdot P_{n0}^2 \qquad (3)$$

with $B_{dG} = Y_{10}^{(3)} \cdot \frac{\gamma_{10}^{(3)}}{\gamma_y} \cdot (-\ln \alpha_2)$.

Susceptance B_{dn} involves the coupling between TE_{1n}(to z) and TM_{1n}(to z) waves in regions (3) and (4) and is determined as follows:

$$B_{dn} = \left[\left(n \cdot \gamma_y\right)^2 \cdot B_{dn}^{TM} + \gamma_x^2 \cdot B_{dn}^{TE} \right] \Big/ \left[\left(n \cdot \gamma_y\right)^2 + \gamma_x^2 \right]$$

where B_{dn}^{TM} and B_{dn}^{TE} are input susceptances at T_0 as in Fig.4 for TM_{1n} and TE_{1n} modes, respectively. Propagation constants and characteristic admittances of the transmission lines in Fig.4 are equal:

$$\left(\gamma_{1n}^{(3)}\right)^2 = \varepsilon_r \cdot \gamma_o^2 - \gamma_x^2 - \left(n \cdot \gamma_y\right)^2 \; ; \; \gamma_{1n}^{(4)} = \gamma_{1n}^{(1)} \; ;$$

$$\left(Y_{1n}^{(3)}\right)^{TM} = Y_{10}^{(3)} \cdot \varepsilon_r \cdot \gamma_o^2 \Big/ \left(\gamma_{10}^{(3)} \cdot \gamma_{1n}^{(3)}\right) \; ;$$

$$\left(Y_{1n}^{(4)}\right)^{TM} = Y_{10}^{(4)} \cdot \gamma_o^2 \Big/ \left(\gamma_{10}^{(4)} \cdot \gamma_{1n}^{(4)}\right) \; ;$$

$$\left(Y_{1n}^{(i)}\right)^{TE} = Y_{10}^{(i)} \cdot \gamma_{1n}^{(i)} \Big/ \gamma_{10}^{(i)} \qquad \text{with } i=3,4 .$$

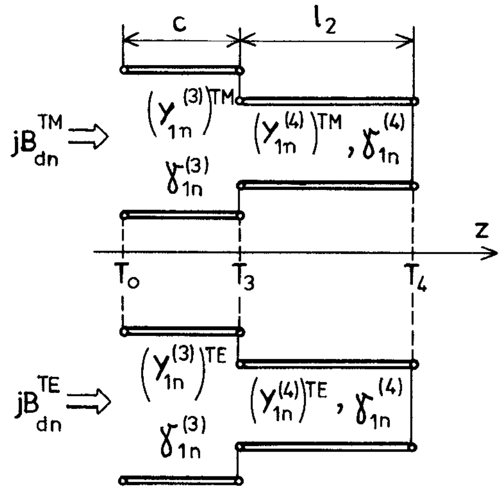

Fig.4: Network representation for higher order modes in regions (3) and (4).

The above formulas allow us to construct the resonance condition (1).

The characteristic impedance of finline on the basis of power-voltage definition can be calculated by using the procedure given in (10).

NUMERICAL RESULTS

The bisection method has been used to solve equation (1). The first twelve terms of

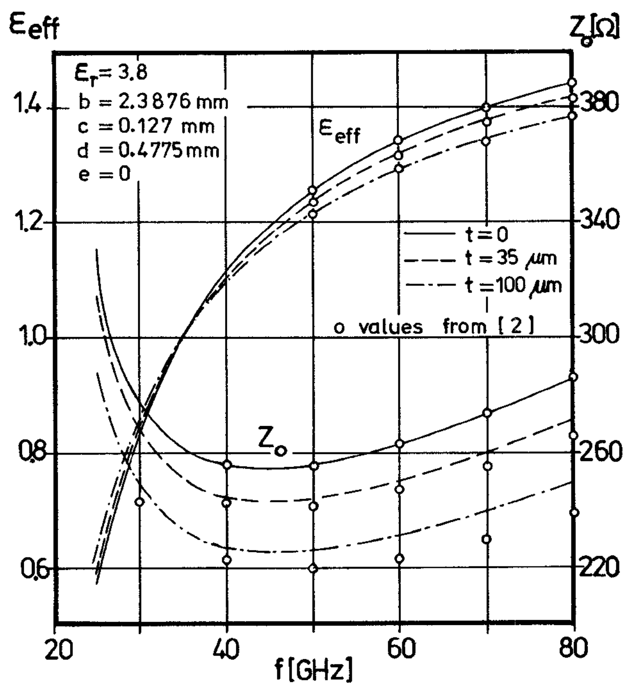

Fig.5: Characteristics of unilateral finline (fundamental mode).

the series in equations (2) and (3) have been taken into account.

Fig.5 shows frequency dependence of the effect of the metallization thickness on the effective dielectric constant ε_{eff}, and on the characteristic impedance Z_0 in a unilateral finline. The published data (2) are also plotted in Fig.5 and close agreement between the two sets of results is seen for effective dielectric constant. The results for Z_0 (for t=0) presented in Fig.5 show good agreement. Comparing the characteristic impedance for the case of finite metallization thickness to results published in (2) yields deviations of less than 3 per cent for t=35µm and less than 6 per cent for t=100µm.

The comparison of the numerical results obtained by this method (TRM) and by the singular integral equation technique (SIET) (12), for propagation constants of fundamental and higher order modes in unilateral finline with zero metallization thickness, is presented in Table 1. The agreement is very good.

Table 1 - The propagation constants (β_x) of the first three modes in a unilateral finline. parameters: ε_r=3.75 , a=1.65mm, b=0.825mm, c=0.11mm, d=0.3mm, l_1=0.77mm, e=t=0 .

f [GHz]	mode no.	β_x [rad/mm]	
		SIET	TRM
60	1	0.17908	0.17894
120	1	2.85901	2.85913
180	1	4.79879	4.79910
	2	0.63598	0.63617
230	1	6.45114	6.45071
	2	3.12472	3.12480
	3	2.05299	2.05570
250	1	7.13004	7.12945
	2	3.76650	3.76664
	3	2.97324	2.97567

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