Computer-Aided Design Models for Millimeter-Wave Finlines and Suspended-Substrate Microstrip Lines

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Abstract —This paper presents closed-form expressions that model the electrical characteristics of finlines and suspended-substrate microstrip lines over a wide range of structural parameters for millimeter-wave applications. The expressions for finlines are within ± 0.8 percent for the cutoff wavelength and within ± 2 percent for the phase constant and the characteristic impedance (based on the power-voltage definition for a small slot $(d/b \leqslant 0.2)$) of spectral-domain results. The expressions for suspended and inverted microstrips are within ± 1 percent. The equations are vital to CAD of millimeter-wave planar circuits using these media.

I. Introduction

CONVENTIONAL microstrip techniques can be applied to millimeter-wave circuits by mere scaling of the linear dimensions. However, such an approach is invariably associated with critical tolerances and very narrow structural dimensions which are incompatible with hybrid devices. This led Schneider to the proposal of suspended-substrate microstrip lines [1] and Meier to that of finlines [2].

With increasing activities in the millimeter-wave field, more attention has been paid to the E-plane transmission-line technique, among various alternatives, for integration of millimeter-wave circuits. During the last couple of years, various suspended-substrate lines have been combined efficiently to form versatile mixed waveguide integrated circuits mounted in the E-plane of a metal waveguide housing.

With growing interest from the component designer, there has been increasing activity concerning the theoretical foundations of the new media. Initially, research was carried out towards experimental determination of the phase constant, field distribution, and characteristic impedance [1], [2]. During the past several years, a number of papers have been published by several authors on the theoretical determination of these parameters. For example, some approximate field solutions to the finline problem were proposed [3]–[5], but the spectral-domain technique [6]–[9] subsequently made them obsolete. This technique is an extension of the space-domain analysis [10], [11]. The finline cutoff wavelength has also been determined using the transmission-line matrix analysis (TLM) [12] and the finite-element method [13]. However, these methods neglect the effects of finite metallization thickness and supporting grooves. Such effects have been taken into consideration by Beyer and Wolff [14] by Saad and Schuneman [15] using the mode-matching technique, and also by Kitazawa and Mittra [16] using the network analytical method. So far the most generalized approach to the problem of finline propagation characteristics has been reported by Vahldieck [17]. This particular hybrid-mode approach takes into account the effects of multilayered dielectric, finite metallization thickness, and substrate holding grooves. All these rigorous analyses lead to complicated and time-consuming computer programming. Besides the rigorous analyses described above, the propagation constant in finlines has been approximated by various methods [18]–[20]. Such approximations are of poor accuracy and need solutions of transcendental equations.

Feeling the need for closed-form solutions, Sharma and Hoefer [21] developed purely empirical expressions valid only for particular permitivity values and small fingaps. Keeping all these factors in mind, we have developed closed-form equations [22]–[25] semiempirically by the least-squares curve fitting to numerical results obtained by rigorous numerical techniques [12], [13]. Such equations are applicable over a complete useful range of finline parameters with adequate accuracy for all practical purposes.

The electrical characteristics in suspended and inverted microstrips have been analyzed by various authors using the variational technique in the spectral domain [26], [27], and also using the Green's function approach [28]. Unlike the conventional microstrip [29], there are no simple and accurate models for such lines.

In this paper, closed-form equations are developed for dispersion in bilateral and unilateral finlines by using equivalent susceptances of waveguide T-junctions, and for the characteristic impedances by curve fitting to the spectral-domain results. Equations for the phase velocities and the characteristic impedance in suspended-substrate microstrip lines were obtained by least-square curve fitting to spectral-domain results. In addition, closed-form synthesis equations are presented for the microstrip lines.

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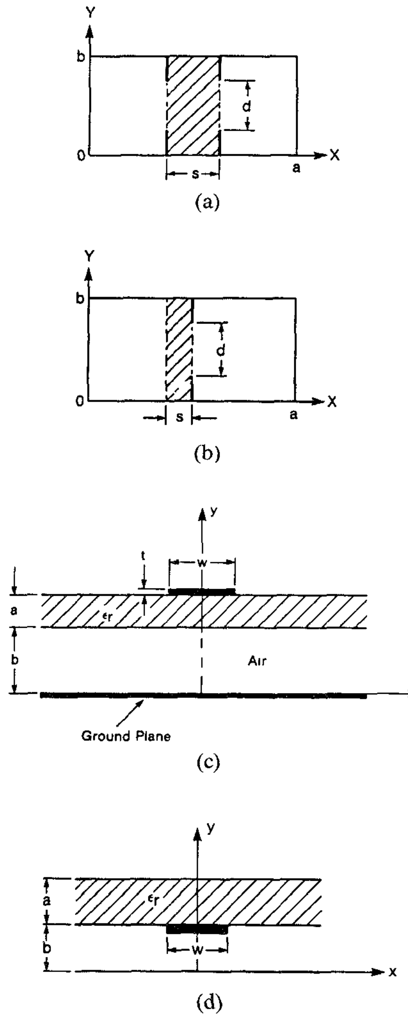


Fig. 1. (a) Bilateral finline. (b) Unilateral finline. (c) Suspended microstrip line. (d) Inverted microstrip line.

II. The Derivations of the Expression for Wave Propagation in Finlines

The guided wavelength in finlines shown in Fig. 1(a) and (b) can be written as

$$\lambda_g = \lambda\sqrt{\epsilon_e(f)} \tag{1}$$

where λ is the free-space wavelength and $\epsilon_e(f)$, the frequency-dependent effective dielectric constant of the finline, is given by [2]

$$\epsilon_e(f) = k_e - (\lambda/\lambda_{ca})^2 \tag{2}$$

where k_e is the equivalent dielectric constant at frequency f corresponding to the wavelength λ, and λ_{ca} is the cutoff wavelength of a finned waveguide of identical dimensions and completely filled with air. It has been shown [30] that for moderate ϵ_r ($\epsilon_r \leqslant 2.5$) and all practical thin substrates, one can make a first-order approximation by equating k_e to its value k_c at cutoff frequency. Otherwise, k_e must be considered frequency-dependent and has the general form

$$k_e = k_c f(d/b, s/a, b/\lambda, \epsilon_r) \tag{3}$$

$$k_c = (\lambda_{cd}/\lambda_{ca})^2. \tag{4}$$

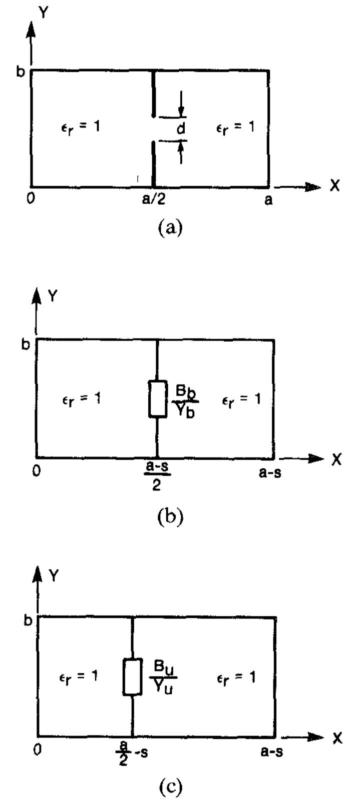


Fig. 2. (a) Finned waveguide. (b) Equivalent circuit for bilateral finline. (c) Equivalent circuit for unilateral finline.

λ_{cd} is the cutoff wavelength of the finline. The function $f(d/b, s/a, b/\lambda, \epsilon_r)$ can be obtained from the dispersion model of Pramanick and Bhartia [25]. Having defined the effective dielectric constant $\epsilon_e(f)$, the characteristic impedance of the finline is defined as [2]

$$Z_0 = Z_{0\infty}/\sqrt{\epsilon_e(f)} \tag{5}$$

where $Z_{0\infty}$ is a function of the cross-sectional dimensions of the finline and a slowly varying function of frequency.

A. Derivation of Expression for the Cutoff Wavelength in Bilateral Finlines

The cutoff wavelength in an air-filled finned waveguide shown in Fig. 2(a) is given by [31]

$$\frac{b}{\lambda_{ca}} = \frac{b}{2a}\left[1+\left(\frac{4}{\pi}\right)\left(\frac{b}{a}\right)\left(1+0.2\sqrt{\frac{b}{a}}\right)x\right]^{-1/2} \tag{6}$$

where

$$x = \ln\csc\left(\frac{\pi}{2}\frac{d}{b}\right) \tag{7}$$

and the corresponding normalized susceptance of the fingap is given by

$$\frac{B_a}{Y_a} = \frac{4b}{\lambda_{ca}}x. \tag{8}$$

The bilateral finline in Fig. 1(a) can be replaced by a waveguide of major dimension of the cross section $(a-s)$ and the minor dimension b, with a susceptance of B_b/Y_b

across the center as shown in Fig. 2(b). Using Marcuvitz's [32] formula for the equivalent network of a waveguide T-junction, the susceptance can be modeled as

$$\frac{B_b}{Y_b} = \frac{4b}{\lambda_{cd}}[x + \epsilon_r G_d] \tag{9}$$

where the term G_d accounts for the effect of the substrate and is given by

$$G_d = \eta_d \arctan(1/\eta_d) + \ln[1+\eta_d^2]^{1/2} \tag{10}$$

$$\eta_d = (s/a)/(b/a)(d/b). \tag{11}$$

Therefore, as with (6), the normalized cutoff wavelength in the bilateral finline will be given by

$$\frac{b}{\lambda_{cd}} = \frac{b}{2(a-s)}\left[1+\left(\frac{4}{\pi}\right)\left(\frac{b}{a-s}\right)\left(1+0.2\sqrt{\frac{b}{a-s}}\right)x_b\right]^{-1/2} \tag{12}$$

where x_b is obtained by comparing (8) and (9) as

$$x_b = x + \epsilon_r G_d. \tag{13}$$

The normalized cutoff wavelength in the air-filled bilateral finline b/λ_{ca} is obtained by using $\epsilon_r = 1$ in the above equations.

B. Cutoff Wavelength In Unilateral Fin Line

The unilateral finline in Fig. 1(b) can be replaced by a waveguide of cross-sectional dimensions $a-s$ and b and asymmetrically loaded with an equivalent susceptance B_u/Y_u as shown in Fig. 2(c).

Using Marcuvitz's [32] formula for the equivalent network of a waveguide T-junction, the susceptance can be modeled as

$$\frac{B_u}{Y_u} = \frac{4b}{\lambda_{cd}}[2x + \epsilon_r(G_a + G_d)] \tag{14}$$

where

$$G_a = \eta_a \arctan(1/\eta_a) + \ln[1+\eta_a^2]^{1/2} \tag{15}$$

$$\eta_a = \eta_d\left(\frac{d}{b}\right). \tag{16}$$

Therefore, as with (6), the normalized cutoff wavelength b/λ_{cd} in unilateral finline can be written as

$$\frac{b}{\lambda_{cd}} = \frac{b}{2(a-s)}\left[1+\left(\frac{4}{\pi}\right)\left(\frac{b}{a-s}\right)\left(1+0.2\sqrt{\frac{b}{a-s}}\right)Kx_u\right]^{-1/2} \tag{17}$$

where

$$x_u = 2\ln\csc\left(\frac{\pi}{2}\frac{d}{b}\right) + \epsilon_r[G_a + G_d]$$

and the factor K accounts for the asymmetry in the structure. It has been found empirically that K should have the following form:

$$K = 1 - \frac{s}{a}F(s/a)(1.231 - 0.0769\epsilon_r) \tag{18}$$

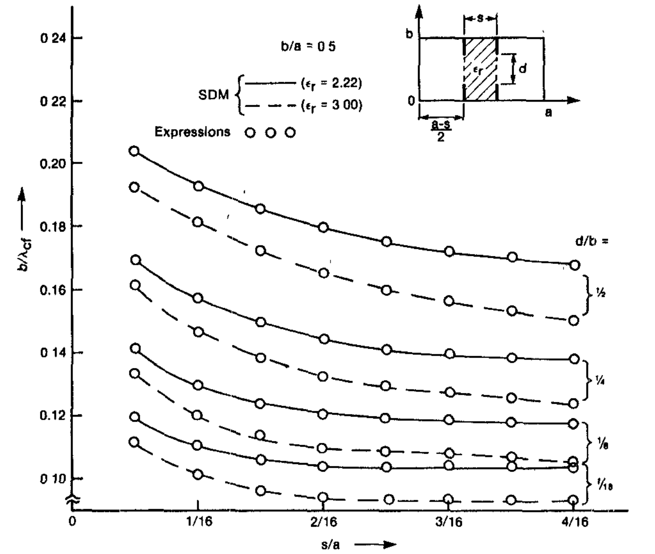


Fig. 3. Normalized cutoff frequency (b/λ_{cd}) in bilateral finline as a function of substrate thickness.

where

$$\begin{aligned} F(s/a) &= 1.9454r^3 - 12.504r^2 + 31.524r \\ &\quad - 25.1223 \qquad \text{for } d/b \leqslant 0.5 \\ &= 2.588r^3 - 17.06r^2 + 42.451r \\ &\quad - 33.934 \qquad \text{for } 0.5 \leqslant d/b \leqslant 0.75 \\ &= 3.47r^3 - 23.77r^2 + 59.285r \\ &\quad - 48.05 \qquad \text{for } 0.75 \leqslant d/b \leqslant 1.00 \end{aligned}$$

with $r = \ln(a/s)$ for $s/a \leqslant 1/8$ and $2.2 \leqslant \epsilon_r \leqslant 3.8$. The normalized cutoff wavelength b/λ_{ca} of the air-filled unilateral finline is obtained from (6).

The accuracy of these new expressions for the cutoff wavelengths in bilateral and unilateral finline is of the order of ± 0.8 percent over the range $2.2 \leqslant \epsilon_r \leqslant 3.8$, $1/64 \leqslant s/a \leqslant 1/8$, $d/b \leqslant 1.00$ and $0 \leqslant b/a \leqslant 1$ of structural parameters. This virtually includes the complete range of finline use. Figs. 3 and 4 display the results graphically. This inspires confidence in the above results.

C. Characteristic Impedance

Finline being a non-TEM line, the characteristic impedance is not uniquely defined. According to Jansen's arguments [33] and the experimental findings of Willing and Spielman [34], with chip resistors connected across the fin gap, the power-voltage definition appears well confirmed. In this definition, voltage is defined as the line integral of the electric field between the fins along the shortest path over the substrate surface, and power is defined as the total average power flowing through the line. Such a definition is particularly suitable for matching beam lead devices with small fin gaps. Based on this and curve fitting to spectral-domain results [6], the following expression has been obtained for the characteristic impedance of bilateral finline and has an accuracy of ± 2 percent for $d/b \leqslant 0.25$:

$$Z_0 = \frac{240\pi^2(p\bar{x} + q)(b/(a-s))}{(0.385\bar{x} + 1.7621)^2\sqrt{\epsilon_e(f)}} \tag{19}$$

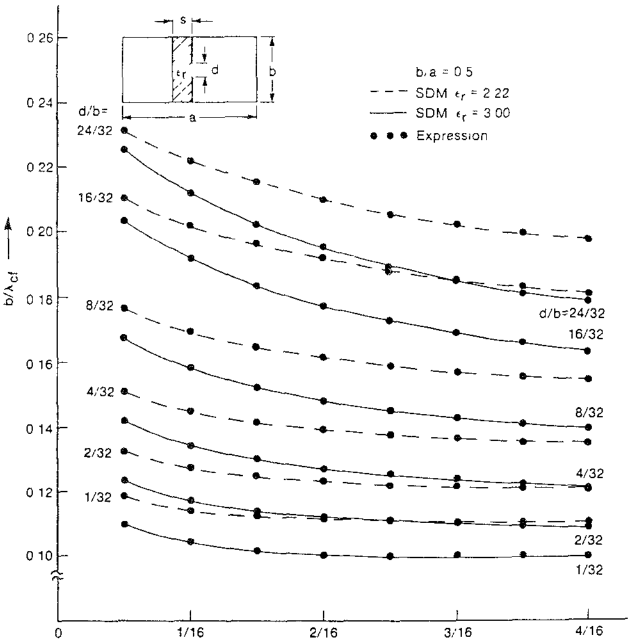


Fig 4. Normalized cutoff frequency (b/λ_{cd}) in unilateral finline as a function of substrate thickness.

where

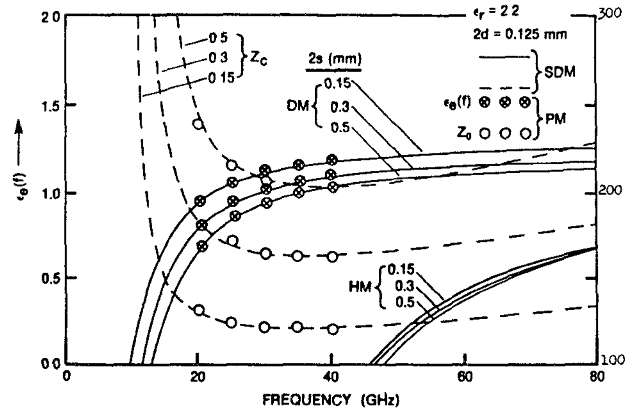
$$\bar{x} = x_b - G_d(\epsilon_r - 1)$$

$$p = 0.097(b/\lambda)^2 + 0.01(b/\lambda) + 0.04095$$

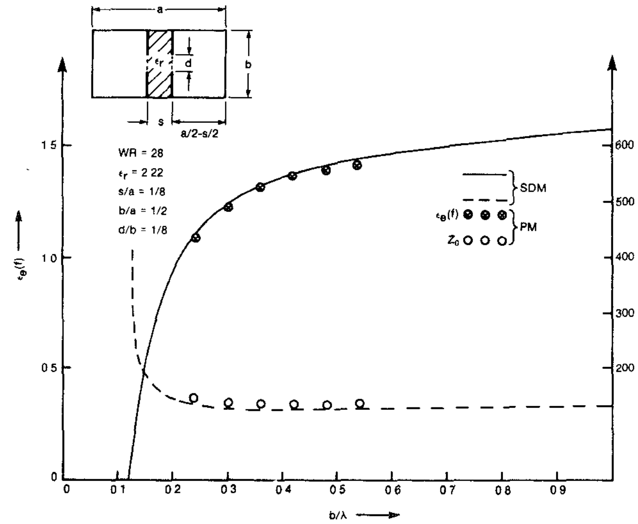
$$q = 0.0031(b/\lambda) + 0.89.$$

A similar equation for unilateral finlines developed by the authors has been presented in [22].

The computed propagation constant and the characteristic impedance for the bilateral finline are shown in Fig. 5(a) and (b). The agreement is within ± 2 percent for $0.2 \leqslant b/\lambda \leqslant 0.6$, $s/a \leqslant 1/20$, and $d/b \leqslant 0.2$.



(a)



(b)

Fig. 5. (a) Dispersion in bilateral finline $s/a = 1/56$. (b) Dispersion in bilateral finline $s/a = 1/8$.

III. Derivation of Expressions for Propagation Characteristics in Suspended and Inverted Microstrips

A. Analysis

The models of suspended-substrate microstrips are based on the equation for the impedance Z_0 of the microstrip in an homogeneous medium, and the static TEM effective dielectric constant $\epsilon_e(0)$ at zero frequency. Z_0 is given within ± 0.03 percent by [35]

$$Z_0 = \frac{\eta_0}{2\pi} \ln\left[\frac{f(u)}{u} + \left(1 + \left(\frac{2}{u}\right)^2\right)^{1/2}\right] \tag{20}$$

where

$$f(u) = 6 + (2\pi - 6)\exp\left[-\left(\frac{30.666}{u}\right)^{0.7528}\right]$$

where $u = w/(a+b)$ for suspended microstrip (Fig. 1(c)), $u = w/b$ for inverted microstrip (Fig. 1(d)), and $\eta_0 = 120\pi$ Ω.

The characteristic impedance of the microstrip is given by

$$Z = Z_0/\sqrt{\epsilon_e(0)}. \tag{21}$$

We assume a general form of the equation for the effective dielectric constant $\epsilon_e(0)$ as

$$f_1\{\epsilon_e(0)\} = 1 + q(w/b, a/b)\{f_2(\epsilon_r) - 1\}. \tag{22}$$

By least-squares curve fitting to results obtained by the spectral-domain method [26], the unknown functions f_1 and f_2 for the suspended microstrip are determined to be

$$f_1(\epsilon_e(0)) = 1/\sqrt{\epsilon_e(0)} \tag{23}$$

$$f_2(\epsilon_r) = \frac{1}{\sqrt{\epsilon_r}} \tag{24}$$

and

$$q(w/b, a/b) = \frac{a}{b}\left[a_1 - b_1 \ln(w/b)\right] \tag{25}$$

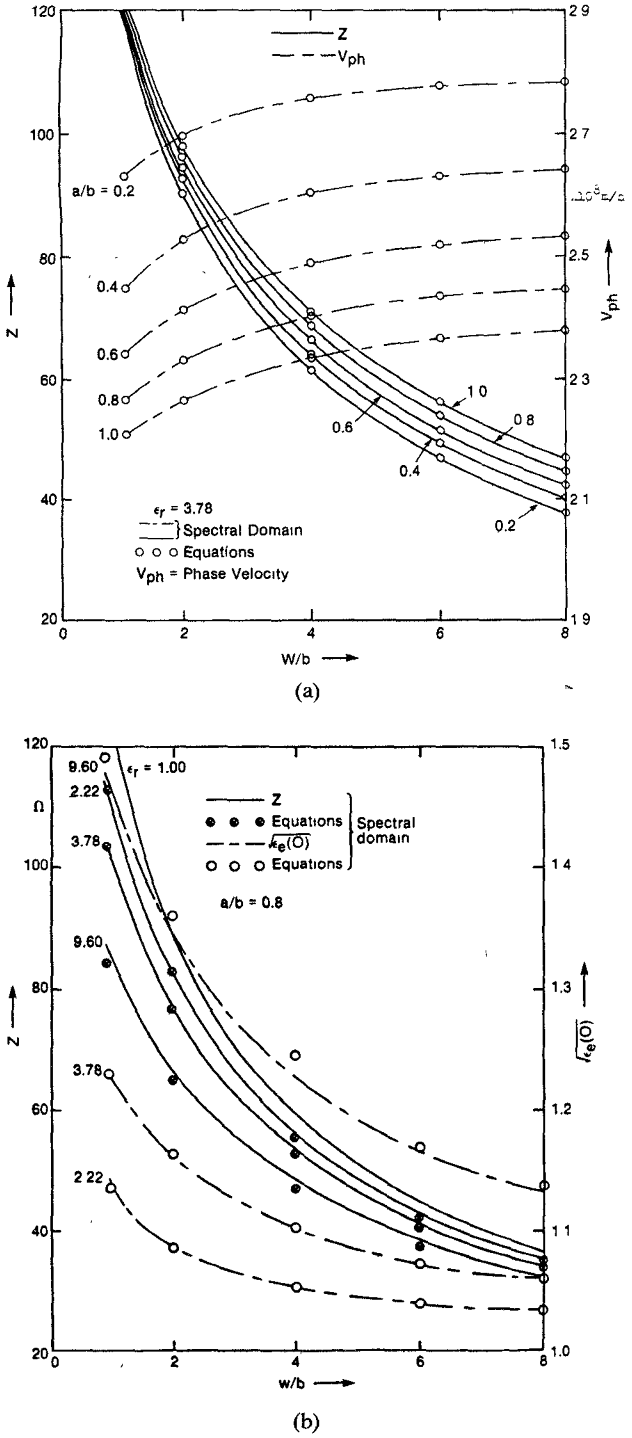


Fig. 6. (a) Characteristics of suspended microstrip. (b) Characteristics of inverted microstrip.

where

$$a_1 = [0.8621 - 0.1251 \ln(a/b)]^4$$
$$b_1 = [0.4986 - 0.1397 \ln(a/b)]^4.$$

For inverted microstrip, the functions are determined to be

$$f_1[\epsilon_e(0)] = \sqrt{\epsilon_e(0)} \qquad (26)$$
$$f_2(\epsilon_r) = \sqrt{\epsilon_r} \qquad (27)$$

while $q(w/b, a/b)$ has the same form as in the case of suspended microstrip line, and a_1 and b_1 are defined as

$$a_1 = [0.5173 - 0.1515 \ln(a/b)]^2$$
$$b_1 = [0.3092 - 0.1047 \ln(a/b)]^2.$$

The above analysis equations are accurate to within ± 1 percent for $1 \leqslant w/b \leqslant 8$, $0.2 \leqslant a/b \leqslant 1$, and $\epsilon_r \leqslant 6.00$.

B. Synthesis

In order to fabricate the appropriate suspended-substrate microstrip, one requires the strip width w given the dimensions a and b, and the substrate dielectric constant ϵ_r for a specified characteristic impedance Z at a specified operating frequency f. Under the assumption that the quasistatic mode f does not play any role (dispersion is neglected), the following synthesis equations are presented:

$$\frac{w}{b} = \left(1 + \frac{a}{b}\right) \exp\left[\frac{1}{2}\left\{-p + \left(p^2 - 4q\right)^{1/2}\right\}\right] \qquad (28)$$

where

$$p = \frac{m}{n} + \frac{A}{B}$$
$$q = \frac{Am - Z/60}{Bn}$$
$$A = 1 + \left(\frac{a}{b}\right)\left[a_1 - b_1 \ln\left(1 + \frac{a}{b}\right)\right]\left[\frac{1}{\sqrt{\epsilon_r}} - 1\right]$$
$$B = \left(\frac{a}{b}\right) b_1 \left(1 - \frac{1}{\sqrt{\epsilon_r}}\right)$$
$$\left.\begin{array}{l} m = 2.1330 \\ n = -0.9054 \end{array}\right\}, \quad \text{for } 0.5 \leqslant \frac{w}{a+b} \leqslant 3$$
$$\left.\begin{array}{l} m = 1.8620 \\ n = -0.6420 \end{array}\right\} \quad \text{for } 3 \leqslant \frac{w}{a+b} \leqslant 6.667.$$

For suspended microstrip line, (28) offers an accuracy of ± 1.5 percent. The accuracy can be further improved by using the more accurate analysis equation. For inverted microstrip

$$\frac{w}{b} = \exp\left[\frac{1}{2}\left\{-p - \left(p^2 - 4q\right)^{1/2}\right\}\right] \qquad (29)$$

where

$$p = 8.9374 \bar{B} \bar{Z} - 8.665$$
$$q = 18.767 - 8.9374 \bar{A} \bar{Z}$$
$$\bar{A} = 1 + \frac{a}{b}\left(\sqrt{\epsilon_r} - 1\right) a_1$$
$$\bar{B} = \frac{a}{b}\left(\sqrt{\epsilon_r} - 1\right) b_1$$
$$\bar{Z} = Z/60.$$

The above equation offers an accuracy of ± 1 percent.

C. Application to Anisotropic Substrate

The above equations are also valid with anisotropic substrates by defining an isotropic equivalence of the anisotropic substrate where the equivalent substrate thickness $\bar{a}$ and the equivalent substrate dielectric constant $\bar{\epsilon}_r$ are given by [35]

$$\bar{a} = a\left(\epsilon_x/\epsilon_y\right)^{1/2} \tag{30}$$

$$\bar{\epsilon}_r = \left(\epsilon_x\epsilon_y\right)^{1/2}. \tag{31}$$

The method requires that one of the principal axes of the substrate be parallel to the substrate (x axis) and the other normal to the substrate (y axis). The strip dimensions remain unchanged. The procedure for utilizing the above model then proceeds through calculation of the mode capacitance C_m

$$C_m = 1/\left(\bar{V}_m \bar{Z}_m\right) \tag{32}$$

where $\bar{V}_m$ is the mode phase velocity and Z_m the mode impedance of the equivalent isotropic substrate. The effective dielectric constant and the impedance of the anisotropic microstrip is then computed from

$$\epsilon_e(0) = C_m/C_0$$

$$Z_a = 1/\left(c\sqrt{C_m C_0}\right) \tag{33}$$

where c is the velocity of EM wave in free space and C_0 is the mode capacitance of the microstrip.

IV. Conclusion

Novel analytical expressions have been reported for the propagation characteristics of suspended-substrate finlines and microstrip lines. They are given in a form easily implemented on a desk-top computer. In terms of accuracy and computational effort, the described equations represent a considerable improvement to present techniques for millimeter-wave integrated-circuit design using these transmission media. The equations are primarily set up for use in computer-aided millimeter-wave circuit optimization routines, and are fully compatible with the needs and trends of modern computer-aided circuit design.

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