ties of magnetic materials at high frequencies. In 1962 he transferred to the High Frequency Impedance Standards Section where he did research and development work on inductance standards, impedance bridges, inductive voltage dividers, attenuators, and directional couplers. In 1972 his emphasis shifted to developing Josephson junction detectors for precision rf attenuation measurements. Since 1974 he and a co-worker, Glenn Engen, have been responsible for advancing the theory and application of the six-port concept to RF and microwave measurements. He and Engen received the Department of Commerce Gold Medal Award in 1976 for their development of the six-port concept.

Mr. Hoer is the author or coauthor of 36 technical papers and holds two patents.

Propagation Parameters of Coupled Microstrip-Like Transmission Lines for Millimeter-Wave Applications

SHIBAN K. KOUL AND BHARATHI BHAT

Abstract—**A variational expression is derived for the propagation parameters of coupled microstrip-like transmission lines for millimeter-wave applications using the "transverse transmission line" method. Numerical results are presented for the coupled inverted microstrip lines, and for the coupled suspended microstrip lines. The effects of the top and sidewalls and also of the finite thickness of strip conductors on the even- and odd-mode impendances are studied. The use of a dielectric overlay in equalizing the even- and odd-mode phase velocities is investigated.**

I. Introduction

MICROSTRIP-LIKE transmission lines, which incorporate an air gap between the dielectric substrate and the ground plane, such as the inverted microstrip and the suspended microstrip, are known to offer less circuit losses and less stringent dimensional tolerances compared with the conventional microstrip lines [1]–[3]. The same advantages accrue in the case of coupled microstrip-like transmission lines shown in Fig. 1 and its two special cases, namely; the coupled inverted microstrip lines and coupled suspended microstrip lines which result when $h_2=0$ and $h_3=0$, respectively. These structures, therefore, find applications in the design of filters and couplers at millimeter-wave frequencies. Smith [4] has analyzed the even- and odd-mode capacitances of the coupled lines on a suspended substrate based upon conformal transformation. Mirshekar-Syahkal and Davies [5] have analyzed shielded multilayer dielectrics with arbitrary coplanar conductors using spectral-domain approach.

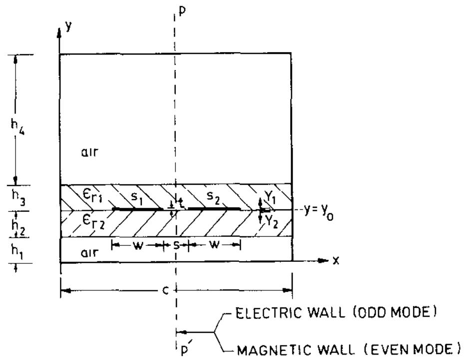

Fig. 1 Coupled sandwiched microstrip structure.

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The purpose of this paper is to present a comprehensive analysis of coupled microstrip-like transmission lines having multilayer dielectrics. The analysis uses variational technique combined with "transverse transmission line" method. This is an extremely simple technique compared with other methods reported in the literature [4]–[6]. Variational expressions for the even- and odd-mode capacitances for the general shielded sandwiched structure (Fig. 1) are derived. The even- and odd-mode impedances (Z_{0e} and Z_{0o}) and phase velocities (v_{phe} and v_{pho}) of the coupled inverted microstrip lines, and coupled suspended microstrip lines are analyzed in detail. The effects of the top and sidewalls and also of the finite thickness of strip conductors on the impedance characteristics are reported. The use of a dielectric overlay in equalizing $v_{ph\,e}$ and $v_{ph\,o}$ is investigated.

II. Formulation of Green's Function Using Transverse Transmission Line Method

Using the symmetry with respect to $x = c/2$, the coupled sandwiched structure shown in Fig. 1 can be analyzed using the even- and odd-mode method. It suffices to consider only half the structure and the problem reduces to that of finding the capacitance of strip conductor S_1 with respect to ground with appropriate boundary conditions applied at the walls $x=0$ and $c/2$ for the even- and odd-mode excitations.

The Green's function $G(x, y/x_0, y_0)$ due to a unit charge located at (x_0, y_0) satisfies the Poisson's differential equation

$$\nabla^2 G(x, y/x_0, y_0) = -\frac{1}{\epsilon}\delta(x-x_0)\delta(y-y_0) \qquad (1)$$

where $\delta(x-x_0)$ and $\delta(y-y_0)$ are Dirac's delta functions. In this problem, the Green's function can be expressed as

$$G(x, y/x_0, y_0) = \sum_n G_n(x)G_n(y). \qquad (2)$$

In order to satisfy the boundary conditions, namely $G=0$ at $x=0$ and $\partial G/\partial x = 0$ at $x=c/2$ for the even-mode excitation and $G=0$ at $x=0$ and $c/2$ for the odd-mode excitation, $G_n(x)$ takes the form

$$G_n(x) = \sin(\beta_n x) \qquad (3)$$

where

$$\beta_n = \begin{cases} \dfrac{(2n+1)\pi}{c}, & \text{for even mode} \\ \dfrac{(2n\pi)}{c}, & \text{for odd mode.} \end{cases} \qquad (4)$$

Using (2) and (3) in (1), we obtain

$$\left[\frac{d^2}{dy^2} - \beta_n^2\right] G_n(y) = -\frac{4}{c\epsilon}\sin(\beta_n x_0)\delta(y-y_0). \qquad (5)$$

In order to solve for $G_n(y)$, we use the "transverse transmission line" method outlined by Crampagne *et al.* [7]. For a transmission line with a current source of intensity I_s located at $y = y_0$, the voltage V along the line satisfies the following differential equation:

$$\frac{d^2V}{dy^2} - \gamma^2 V = -\left(\frac{\gamma}{Y_c}\right) I_s \delta(y-y_0) \qquad (6)$$

where γ is the propagation constant and Y_c is the characteristic admittance of the line. Setting $Y_c = \epsilon$, and comparing (5) and (6), we get the following analogous relations:

$$V \equiv G_n(y) \qquad \gamma \equiv \beta_n \qquad I_s \equiv \frac{4}{\beta_n c}\sin\beta_n x_0. \qquad (7)$$

Since the voltage on the transmission line at $y = y_0$ is given by $V = I_s/Y$, we get

$$G_n(y) = \frac{4}{\beta_n c Y}\sin\beta_n x_0 \qquad (8)$$

where Y is the admittance at the charge plane $y = y_0$. Using (3), (4), and (8) in (2), the Green's function at $y = y_0$ can be expressed as

$$G_n(x, y_0/x_0, y_0)\binom{0e}{0o} = \sum_{n\binom{\text{odd}}{\text{even}}} \frac{4}{n\pi Y} \cdot \sin\frac{n\pi x_0}{c}\sin\frac{n\pi x}{c} \qquad (9)$$

where the subscripts, $0e$ and $0o$, refer to the even- and odd-mode, respectively.

III. Variational Expressions for Even- and Odd-Mode Capacitances

The variational expression for the capacitance of TEM-mode transmission line is given in [8]. The even- and odd-mode capacitances for the coupled microstrip-like transmission lines can be expressed as

$$C_{\substack{0e\\0o}} = \frac{\left[\int_{S_1} f(x)\binom{0e}{0o}dx\right]^2}{\int_{S_1}\int_{S_1} G(x, y_0/x_0, y_0)\binom{0e}{0o} f(x)\binom{0e}{0o} f(x_0)\binom{0e}{0o} dx\,dx_0} \qquad (10)$$

where $f(x)_{0e}$ and $f(x)_{0o}$ are the even- and odd-mode charge distributions on the strip conductor. They are assumed to be

$$f(x)\binom{0e}{0o} = \begin{cases} \dfrac{1}{w}\left[1 + A\binom{0e}{0o}\left|\dfrac{2}{w}\left(x - \dfrac{c-s-w}{2}\right)\right|^3\right], \\ \left(\dfrac{c-s}{2} - w\right) \leqslant x \leqslant \left(\dfrac{c-s}{2}\right) \\ 0, \quad \text{otherwise.} \end{cases} \qquad (11)$$

A_{0e} and A_{0o} are the constants for the even- and odd-mode excitations, respectively. Substituting (9) and (11) in (10), and simplifying, we obtain

$$C\binom{0e}{0o} = \frac{\left[1 + \left(A\binom{0e}{0o}/4\right)\right]^2}{\sum_{n\binom{\text{odd}}{\text{even}}} g_n\left(L_n + A\binom{0e}{0o}M_n\right)^2} \qquad (12)$$

where

$$M_n=\left(\frac{2c}{n\pi w}\right)^3 \sin\left\{\frac{n\pi w}{2c}\left(\frac{c-s}{w}-1\right)\right\}$$
$$\cdot\left[3\left\{\left(\frac{n\pi w}{2c}\right)^2-2\right\}\cos\left(\frac{n\pi w}{2c}\right)\right.$$
$$\left.+\left(\frac{n\pi w}{2c}\right)\left\{\left(\frac{n\pi w}{2c}\right)^2-6\right\}\sin\left(\frac{n\pi w}{2c}\right)+6\right] \tag{13a}$$

$$L_n=\sin\left\{\frac{n\pi w}{2c}\left(\frac{c-s}{w}-1\right)\right\}\sin\left(\frac{n\pi w}{2c}\right) \tag{13b}$$

$$g_n=\frac{4}{n\pi Y}\left(\frac{2c}{n\pi w}\right)^2 \tag{13c}$$

$$A\binom{0e}{0o}=-\frac{\sum_{n\binom{\text{odd}}{\text{even}}}(4M_n-L_n)L_n g_n}{\sum_{n\binom{\text{odd}}{\text{even}}}(4M_n-L_n)M_n g_n}. \tag{14}$$

The constants A_{0e} and A_{0o} are obtained by maximizing C_{0e} and C_{0o}, respectively.

Combining (12) with standard formulas [8], the even- and odd-mode characteristic impedances and phase velocities can be computed. Expressions (12)–(14) are general and can be applied to parallel-coupled microstrip-like transmission lines with multilayered dielectrics in a shielded configuration. It is only necessary to substitute the appropriate expression for Y at the charge plane depending on the geometry of the structure.

Referring to Fig. 1, the admittance Y at the charge plane is given by

$$Y=Y_1+Y_2 \tag{15}$$

where Y_1 and Y_2 are the admittances at the plane $y=h_1+h_2$ looking in the positive and negative y directions, respectively. Using the transmission line formula to obtain Y_1 and Y_2, we get

$$Y|_{y=h_1+h_2}=\epsilon_0\left[\epsilon_{r1}\left\{\frac{\coth\frac{n\pi h_4}{c}\coth\frac{n\pi h_3}{c}+\epsilon_{r1}}{\epsilon_{r1}\coth\frac{n\pi h_3}{c}+\coth\frac{n\pi h_4}{c}}\right\}\right.$$
$$\left.+\epsilon_{r2}\left\{\frac{\coth\frac{n\pi h_1}{c}\coth\frac{n\pi h_2}{c}+\epsilon_{r2}}{\epsilon_{r2}\coth\frac{n\pi h_2}{c}+\coth\frac{n\pi h_1}{c}}\right\}\right]. \tag{16}$$

IV. Numerical Results and Discussion

The effects of the shielding sidewalls and of the top wall on the even- and odd-mode impedances of the coupled sandwiched microstrip line are depicted in Fig. 2. The variations in the optimum values of A_{0e} and A_{0o} are also shown in the same figure. For typical representative parameters chosen ($\epsilon_{r1}=\epsilon_{r2}=3.78$, $w/h_1=1.0$, $s/h_1=0.2$, and $h_2/h_1=h_3/h_1=0.508$), it is seen that for fixed value

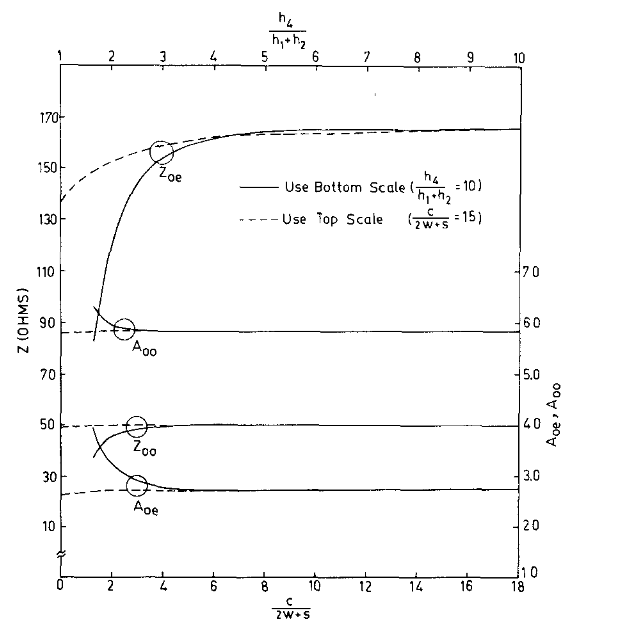

Fig. 2 Effect of sidewalls and top wall on the even- and odd-mode impedances and optimum A_{0e} and A_{0o} of the coupled sandwiched microstrip, $w/h_1=1, s/h_1=0.2, h_2/h_1=h_3/h_1=0.508, t/b=0, \epsilon_{r1}=\epsilon_{r2}=3.78$.

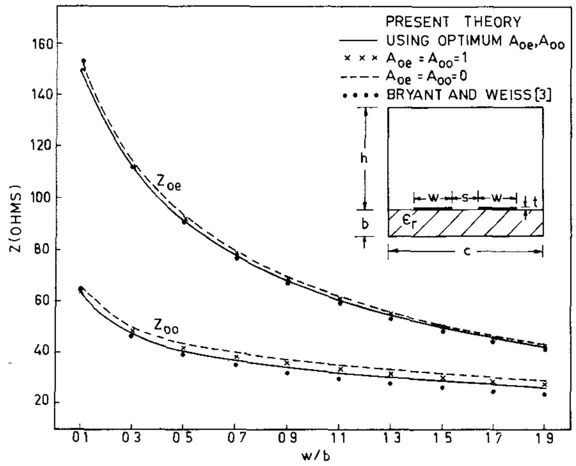

Fig. 3 Calculated even- and odd-mode impedances of coupled microstrip and comparison with Bryant and Weiss [6], $s/b=0.2, t/b=0, \epsilon_r=10$.

of $h_4/(h_1+h_2)=10$, the effect of sidewalls becomes negligible when $c/(2w+s)>10$. For a fixed value of $c/(2w+s)=15$, the effect of the top wall becomes negligible when $h_4/(h_1+h_2)\geqslant 10$. In all the succeeding computations, the parameters $c/(2w+s)$ and $h_4/(h_1+h_2)$ are fixed at 15 and 10, respectively, so that the sidewalls and top wall have negligible effect on the field configuration.

Numerical results of coupled microstrip line (Fig. 3) computed by setting $h_1=h_3=0, h_2=b, h_4=h$, and $\epsilon_{r2}=\epsilon_r$ are found to be in good agreement with the results of Bryant and Weiss [6]. Superimposed on the graph are the results computed with much simpler assumed charge distributions. It can be seen that Z_{0e} and Z_{0o} obtained by setting $A_{0e}=A_{0o}=1$ are in good agreement with the values obtained by using the optimum values of A_{0e} and A_{0o}. On the other hand, with $A_{0e}=A_{0o}=0$, values of Z_{0e} are

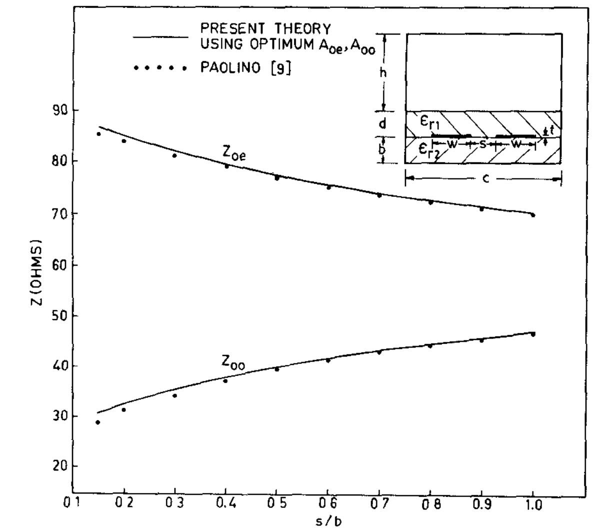

Fig. 4 Calculated even- and odd-mode impedances of coupled microstrip with overlay and comparison with Paolino [9], $w/b=0.4$, $d/b=1.0$, $h/b=\infty$, $t/b=0$, $\epsilon_{r1}=10.1$, $\epsilon_{r2}=10.1$.

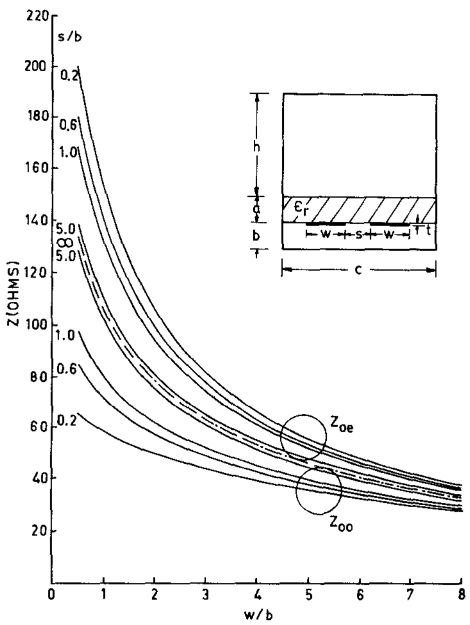

Fig. 5 Even- and odd-mode impedances of coupled inverted microstrip, $a/b=0.508$, $t/b=0$, $\epsilon_r=3.78$.

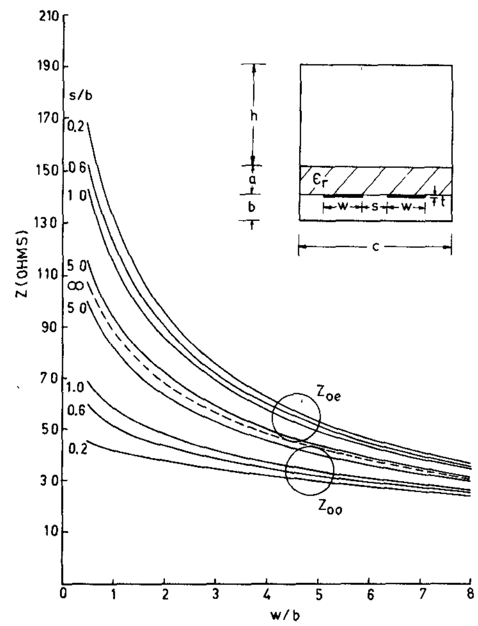

Fig. 6 Even- and odd-mode impedances of coupled inverted microstrip, $a/b=0.64$, $t/b=0$, $\epsilon_r=9.6$.

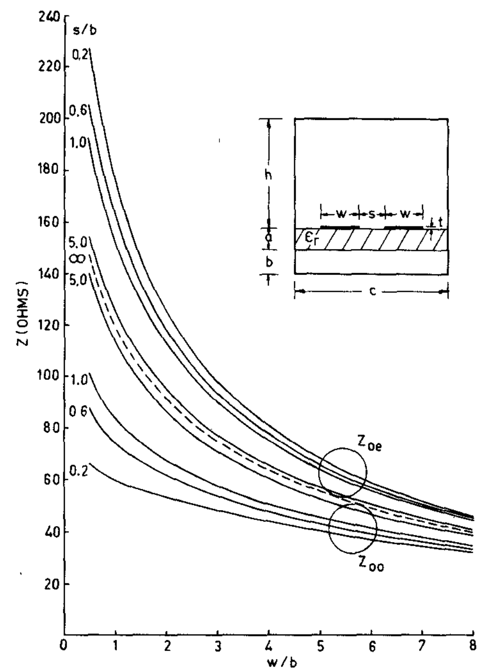

Fig. 7 Even- and odd-mode impedances of coupled suspended microstrip, $a/b=0.508$, $t/b=0$, $\epsilon_r=3.78$.

within 2 percent, while Z_{0o} differs by 3–11 percent, from the results obtained with optimum A_{0e} and A_{0o}. Numerical results reported in all the subsequent figures use optimum values of A_{0e} and A_{0o} given by (14). Computed results of Z_{0e} and Z_{0o} for the coupled microstrip with overlay obtained by setting $h_1=0$ (Fig. 4) agree well with those of Paolino [9].

Computations of propagation parameters for the coupled inverted microstrip are carried out by setting $h_1=b$, $h_2=0$, $h_3=a$, $h_4=h_1$, and $\epsilon_{r1}=\epsilon_r$, and for the coupled suspended microstrip by setting $h_1=b$, $h_2=a$, $h_3=0$, $h_4=h$, and $\epsilon_{r2}=\epsilon_r$. For the coupled inverted microstrip, the variations of Z_{0e} and Z_{0o} as a function of w/b are plotted in Fig. 5 for $a/b=0.508$ and $\epsilon_r=3.78$ and in Fig. 6 for $a/b=0.64$ and $\epsilon_r=9.6$. The corresponding variations for the coupled suspended microstrip are plotted in Fig. 7 for $a/b=0.508$ and $\epsilon_r=3.78$, and in Fig. 8 for $a/b=0.64$ and $\epsilon_r=9.6$. It is found that for a given set of values (s/b, Z_{0e} and Z_{0o}), the strip conductor, in both the configurations, is nearly 2 to 3 times wider than that of coupled microstrip. These structures, therefore, have useful applications at millimeter-wave frequencies.

Fig. 9 shows the variation of the ratio v_{phe}/v_{pho} as a function of w/b with ϵ_r as the parameter for the coupled inverted microstrip lines, and coupled suspended microstrip lines. For both these configurations, this ratio is

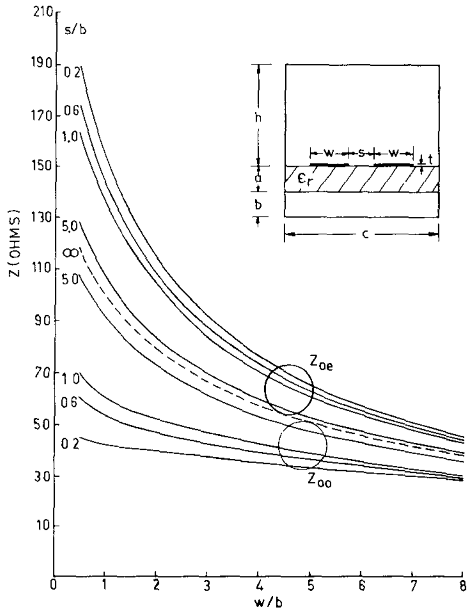

Fig. 8 Even- and odd-mode impedances of coupled suspended microstrip, $a/b = 0.64, t/b = 0, \epsilon_r = 9.6$.

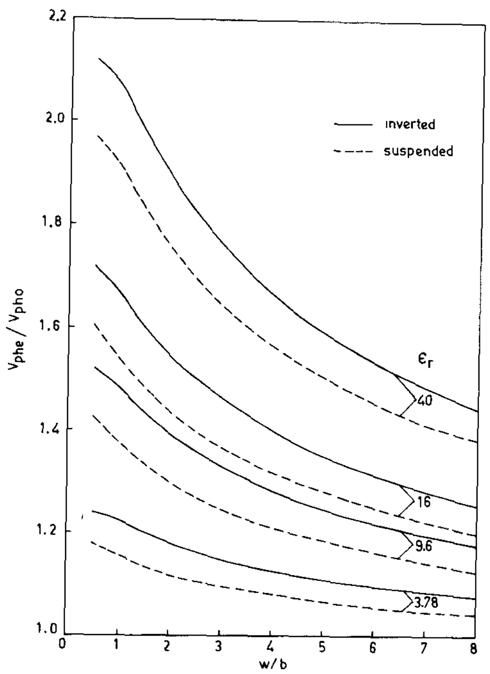

Fig. 9 Ratio of even- and odd-mode phase velocities versus w/b for coupled inverted microstrip lines, and coupled suspended microstrip lines, $a/b = 0.5, s/b = 0.2, t/b = 0$.

found to be greater than that for coupled microstrip. For example, in a coupled inverted microstrip having $a/b = 0.5$, $s/b = 0.2$, and $w/b = 1$, the v_{phe}/v_{pho} ratio is equal to 1.48 for $\epsilon_r = 9.6$. For coupler applications, such large differences in v_{phe} and v_{pho} lead to poor directivity. Since this is essentially due to odd-mode loading, equalization of the phase velocities can be achieved by perturbing only the even-mode fields. This can be implemented by introducing a dielectric overlay on the bottom ground plane. However, this will increase the conductor loss, slightly, due to the

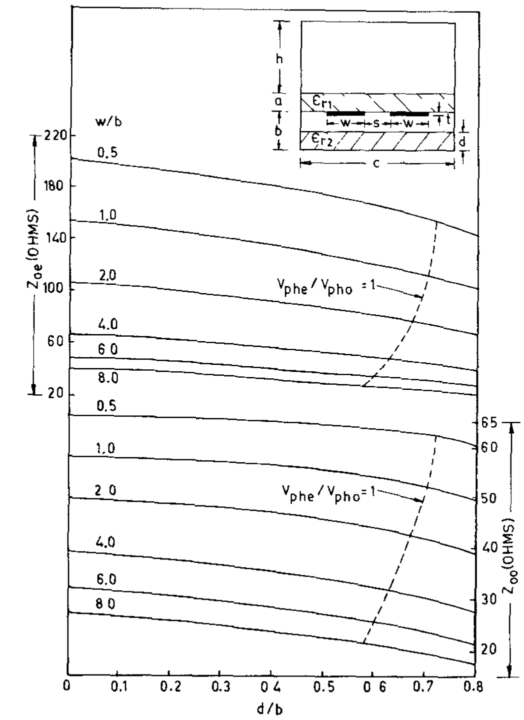

Fig. 10 Effect of dielectric overlay on impedance characteristics of coupled inverted microstrip, $a/b = 0.508, s/b = 0.2, t/b = 0, \epsilon_{r1} = 3.78, \epsilon_{r2} = 9.6$.

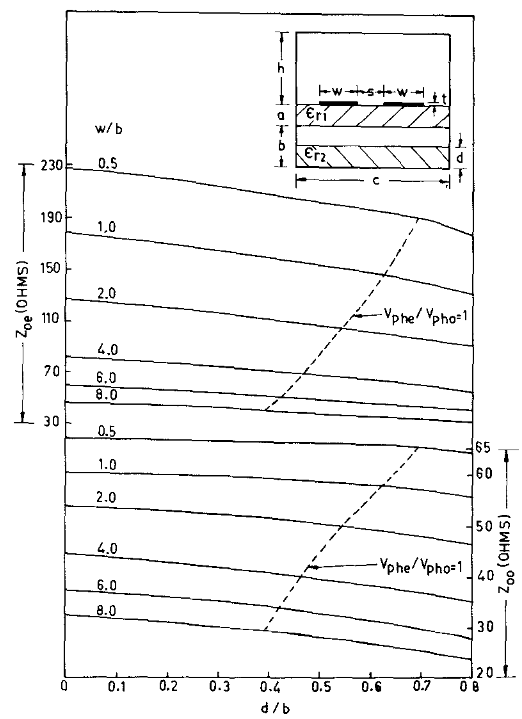

Fig. 11 Effect of dielectric overlay on impedance characteristics of coupled suspended microstrip, $a/b = 0.508, s/b = 0.2, t/b = 0, \epsilon_{r1} = 3.78, \epsilon_{r2} = 9.6$.

ground plane. Figs. 10 and 11 show the effect of this overlay on Z_{0e} and Z_{0o} of the coupled inverted microstrip lines, and coupled suspended microstrip lines, respectively. As expected, increasing overlay thickness d decreases Z_{0o} rather slowly, while Z_{0e} decreases rapidly. The dotted lines

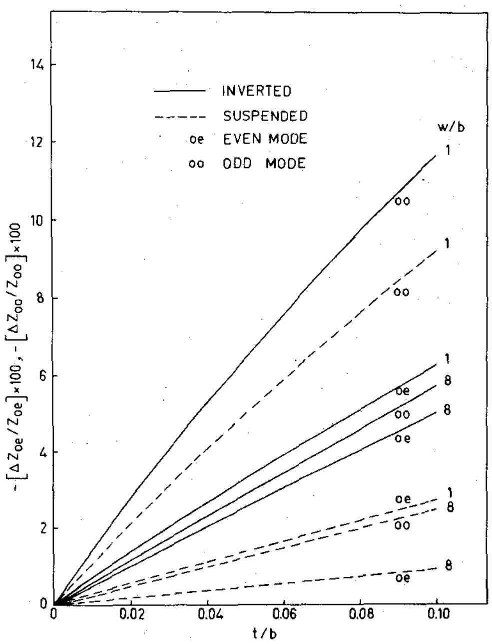

Fig. 12 Effect of finite thickness of strip conductor on the even- and odd-mode impedances for coupled inverted microstrip lines and coupled suspended microstrip lines, $a/b=0.508$, $a/b=0.2$, $\epsilon_r=3.78$.

in both of the figures indicate the contours along which $v_{phe}=v_{pho}$.

The effect of finite thickness t of the strip conductors can be taken into account by considering two layers of charge, one at the charge plane $y=y_0$ and the other at $y=y_0\pm t$, where the positive sign is for the coupled suspended microstrip lines and the negative sign is for the coupled inverted microstrip lines. The Green's function in the expression for capacitance is replaced by the average value of the Green's functions at these two planes. With this, g_n in (12) and (14) is replaced by $g_n t_n$ where

$$t_n=\frac{1}{2}\left\{1+\frac{\sinh\dfrac{n\pi(p-t)}{c}}{\sinh\dfrac{n\pi p}{c}}\right\} \tag{17}$$

$p=b$ for coupled inverted microstrip, and $p=h$ for coupled suspended microstrip.

Fig. 12 shows the percentage change in the values of Z_{0e} and Z_{0o} of the coupled inverted microstrip lines, and coupled suspended microstrip lines as a function of t/b with w/b as the parameter. The wider the strip conductor, the smaller the change in Z_{0e} and Z_{0o} with an increase in t/b. The change in Z_{0o} is much greater than the change observed in Z_{0e}. This effect is similar to that observed in coupled microstrip. However, for the representative parameters chosen, it is seen that for t/b up to 0.015, the percentage decrease in Z_{0e} and Z_{0o} is within 2 percent for both configurations.

V. Conclusions

The even- and odd-mode impedance and phase velocity characteristics of the coupled inverted microstrip lines and coupled suspended microstrip lines are analyzed using the simple "transverse transmission line" method combined with the variational technique. For standard dielectric substrates ($\epsilon_r=3.78, 9.6$) and for a given set of values of s/b, Z_{0e}, and Z_{0o}, values of w/b in the case of the coupled inverted microstrip lines, and coupled suspended microstrip lines are nearly 2 to 3 times larger than those of coupled microstrip lines; thus permitting their use at millimeter-wave frequencies. The difference between the even- and odd-mode phase velocities in both these configurations is found to be greater than that in a coupled microstrip line having comparable impedance levels. Equalization of phase velocities can be achieved by using a dielectric overlay on the bottom ground plane which perturbs essentially the even-mode fields.

Expressions for C_{0e} and C_{0o} derived in this paper are general and can be applied to a class of symmetrically coupled microstrip-like transmission lines with multilayered dielectrics in shielded configuration, so long as appropriate expressions for the admittance Y at the charge plane are substituted.

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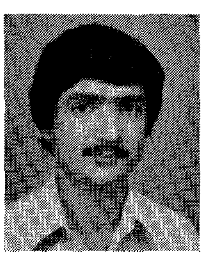

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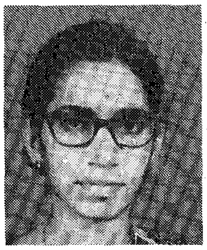
Mr. Koul was awarded a gold medal by the Institution of Engineers, India, for securing first position among all the disciplines of engineering in the B.E. degree course.



Bharathi Bhat received the B.E. degree in electrical communication engineering and the M.E. degree in electronics from the Indian Institute of Science, Bangalore, India, in 1963 and 1965, respectively. She continued her graduate studies at Harvard University, Cambridge, MA, and received the M.S. and Ph.D. degrees in applied physics in 1967 and 1971, respectively.

From 1971 to 1972, she worked as a Post-Doctoral Research Fellow in the Division of Engineering and Applied Physics, Harvard University. In 1973, she joined the Indian Institute of Technology (I.I.T.), New Delhi, India, as Assistant Professor. Since 1977, she has been a Professor and is currently the Head of the Centre for Applied Research in Electronics (C.A.R.E.) at I.I.T.. She is the Leader of the Microwave Group in C.A.R.E. and has been engaged in research and developmental projects in the areas of microwave- and millimeter-wave integrated circuits and antennas.

Dr. Bhat is a Fellow of the Institution of Electronics and Telecommunication Engineers (IETE), India. She is presently the Honorary Editor of *IETE Journal* for the Electromagnetics Section.

Patent Abstracts

These Patent Abstracts of recently issued patents are intended to provide the minimum information necessary for readers to determine if they are interested in examining the patent in more detail. Complete copies of patents are available for a small fee by writing: U.S. Patent and Trademark Office, Box 9, Washington, DC, 20231.

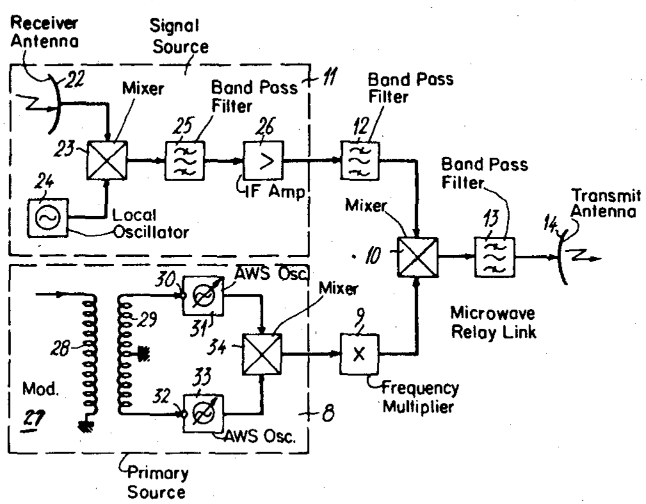
4,271,504 Jun. 2, 1981

Frequency Modulators for Use in Microwave Links

Inventors: Pierre C. Brossard; Jeannine L. G. Henaff.
Filed: Nov. 17, 1978.

Abstract—Frequency modulators for use in microwave link transmission systems include acoustic surface wave (ASW) oscillators. Each frequency modulator comprises two voltage control quadripole ASW oscillators. Control inputs of the two ASW oscillators receive modulation signal in opposite phase. Outputs of ASW oscillators are mixed in a mixer for delivering an IF signal. Thus frequency deviation is larger than in a conventional arrangement. Moreover, surface wave oscillators make it possible to directly insert a service channel signal in a microwave link repeater, without demodulation and modulation operations.

8 Claims, 5 Drawing Figures



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Microwave Receiver

Inventor: Akira Takayama.
Assignee: Alps Electric Co., Ltd.
Filed: Sep. 17, 1979.

Abstract—A microwave receiver is constituted by an antenna having a reflector and a primary radiator, and a converter having an unbalanced input terminal. A core conductor projected from the unbalanced terminal of the converter is extended inside the reflector through a bore formed therein so as to function as the primary radiator of the antenna. The converter is fixed directly to the wall of the reflector.

4 Claims, 3 Drawing Figures