A Novel Exponential Power Combiner/Divider

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***Abstract*—A simple and novel theoretical analysis for an exponential two-way power combiner/divider is developed. The proposed power combiner/divider's parameters, such as input impedance and reflection and transmission coefficients, are determined. The capacitance effect due to the finite thickness of the coupled-line conductor for both uniform and nonuniform transmission lines is found. The properties of the combiner/divider have been tested experimentally.**

I. Introduction

SINCE THE exponential transmission line (ETL) inherits a wide-frequency-band capability comparable to either a microstrip line or a slotline, it lends itself to considerable applications in such microwave circuits as resonators, filters, matching transformers, matching sections, and directional couplers [1]–[3].

This paper analyzes the proposed exponential two-way power combiner/divider in terms of the even and odd modes and calculates the capacitive effect in the coupled lines, which is attributed to the finite thickness of the coupled conductors. This effect of the finite thickness of the coupled conductors will modify the odd-mode velocity and hence its impedance and will provide more accurate analytical data for microwave circuit designers.

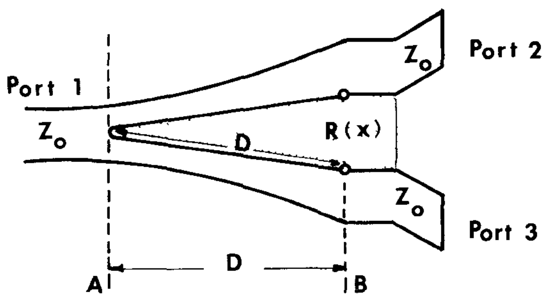
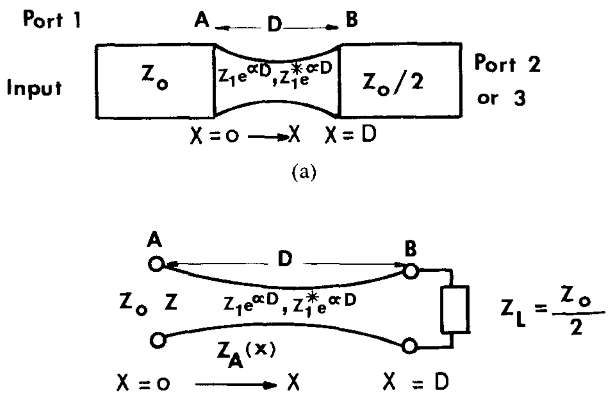


Fig. 1. The two-way exponential power combiner/divider.



(a)

(b)

Fig. 2. Equivalent circuit of ETL power divider (input at port 1).

II. The Exponential Two-Way Power Combiner/Divider

When the proposed combiner/divider, shown in Fig. 1, is used as a power divider (with input signal applied at port 1), the resistance across the proposed network need not be considered. Fig. 1 therefore is reduced to Fig. 2(a) or 2(b). In this case only one mode of propagation exists and this mode has the even-mode property.

The input impedance and the reflection coefficient of the two-way exponential power divider (ETPD) at point A may be found as

$$Z_{1n_{A_d}} = \frac{Z_1^* e^{\alpha D}\left[Z_L(Z_1 + Z_1^*) + j(Z_L(Z_1 - Z_1^*) + 2Z_1 Z_1^*)\tan\beta D\right]}{Z_1^*(Z_1 + Z_1^*) + j\left[(2Z_L Z_1 + Z_1)(Z_1 - Z_1^*)\tan\beta D\right]} \tag{1}$$

$$\rho_{A_d} = \frac{Z_1^*(Z_1 + Z_1^*)(Z_L e^{\alpha D} - Z_0) + j\left[Z^*(Z_1 - Z_1^*)(Z_L e^{\alpha D} + Z_0) + Z_1^*(2Z_1 Z_1^* e^{\alpha D} - 2Z_0 Z_L)\right]\tan\beta D}{Z_1^*(Z_1 + Z_1^*)(Z_L e^{\alpha D} + Z_0) + j\left[Z_1^*(Z_1 - Z_1^*)(Z_L e^{\alpha D} - Z_0) + 2Z_1^*(Z_1 Z_1^* e^{\alpha D} + 2Z_0 Z_L)\right]\tan\beta D}. \tag{2}$$

Here α is the taper rate, which is normally constant, and its value is related to the free-space phase constant β_0 by: $B_0 > (\alpha/2)$; D is the electrical length of the exponential transmission line (ETL), Z_1 is the characteristic impedance of the exponential coupled lines of the divider of length D at $x = 0$ (see Fig. 1); and Z_1^* is the complex conjugate of Z_1. Z_1 and Z_1^* are related to the impedance of the coupled lines at point x by the following relations:

$$Z_D(x) = Z_1 e^{\alpha D} \tag{3}$$

$$Z_D^*(x) = Z_1^* e^{\alpha D}. \tag{4}$$

Z_L is the load impedance, in this case equal to $(Z_0/2)$. Usually $Z_0 = \sqrt{(L_0/C_0)}\, e^{j\theta_0}$, $Z_0^* = /(L_0/C_0)e^{-j\theta}$, $\theta_0 =$

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$\tan^{-1}(-\alpha/2\beta)$, and Z_0^* is the complex conjugate of the local characteristic impedance at $D=0$. The ETL phase constant is represented by [4]

$$\beta=\sqrt{\beta_0^2-\left(\frac{\alpha}{2}\right)^2}. \tag{5}$$

The ratio of voltages at points B and A of the ETL may be expressed as

$$\frac{V_B}{V_A}=\frac{e^{\alpha D}(1+\tan^2\beta)^{1/2}}{1+j\left[\left(\frac{Z_1-Z_1^*}{Z_1+Z_1^*}\right)+\frac{2Z_1Z_1^*}{(Z_1+Z_1^*)Z_Le^{-D}}\right]\tan\beta D}. \tag{6}$$

Usually the transmission coefficient of the exponential transmission line may be represented as

$$\tau_{A_d}=(1+\rho_{A_d})\frac{V_B}{V_A}. \tag{7}$$

From (6) and (7), the power divider transmission coefficient is computed as

$$\tau_{A_d}=\frac{2Z_L(Z_1+Z_1^*)(1+\tan^2\beta D)^{1/2}}{(Z_L+Z_0\bar{e}^{\alpha D})(Z_1+Z_1^*)+j[Z_L(Z_1-Z_1^*)+2Z_1Z_1^*+2Z_0Z_L\bar{e}^{\alpha D}-Z_0Z_L(Z_1-Z_1^*)]\tan\beta D}. \tag{8}$$

III. THE ETL TWO-WAY POWER COMBINER/DIVIDER AS A COMBINER

When the ETL two-way power combiner/divider is used as a combiner, the resistance across the combiner forms a nonconverting array; therefore the modal analysis can be used and the following simple nondegenerate modes are possible:

mode A	1	1
mode B	1	−1.

A. Mode A

Mode A is an even mode; therefore as far as mode A is concerned, the resistance across the combiner need not be considered and the equivalent circuit of the combiner, when the input signal is applied at ports 2 and 3, is as shown in Fig. 3(a) and (b).

The input impedance of mode A at point B is given by

$$Z_{1n_{B\text{mode(A)}}}=\frac{Z_L\bar{e}^{\alpha_e D}(Z_2+Z_2^*)+j[Z_L(Z_2-Z_2^*)\bar{e}^{\alpha_e D}+2Z_2Z_2^*]\tan\beta D}{(Z_2+Z_2^*)+j[(Z_2-Z_2^*)+2Z_L\bar{e}^{\alpha_e D}]\tan\beta D} \tag{9}$$

where Z_2 is the characteristic impedance of mode A at point $X=0$, Z_2^* is the complex conjugate of Z_2, also at $X=0$, and α_e is the mode A taper rate (see Section VI).

The reflection coefficient of mode A may be written as

$$\rho_{B_{\text{mode (A)}}}=\frac{(Z_2+Z_2^*)(Z_L\bar{e}^{\alpha_e D}-Z_0)+j[(Z_2-Z_2^*)(Z_L\bar{e}^{\alpha_e D}-Z_0)+2(Z_2Z_2^*-Z_LZ_0\bar{e}^{\alpha_e D}]\tan\beta D}{(Z_2+Z_2^*)(Z_L\bar{e}^{\alpha D}+Z_0)+j[(Z_L-Z_2^*)(Z_L\bar{e}^{\alpha_e D}+Z_0)+2Z_2Z_2^*+2Z_LZ_0\bar{e}^{\alpha_e D}]\tan\beta D}. \tag{10a}$$

The transmission coefficient of mode A of the proposed combiner may be expressed as

$$\tau_c=\frac{e^{\alpha D}(1+\tan^2\beta D)^{1/2}}{\left(1+\frac{Z_0}{Z_L\bar{e}^{\alpha_e D}}\right)+j\left[\left(\frac{Z_2-Z_2^*}{Z_2+Z_2^*}\right)\left(1+\frac{Z_0}{Z_L\bar{e}^{\alpha_e D}}\right)+\frac{2Z_2Z_2^*}{Z_L\bar{e}^{\alpha_e D}(Z_2)+Z_2^*}+\frac{2Z_0}{(Z_2+Z_2^*)}\right]\tan\beta D}. \tag{10b}$$

B. Mode B

Since mode B has odd symmetry, the equivalent circuit of the combiner, when the input signal is applied at ports 2 and 3, is as shown in Fig. 4.

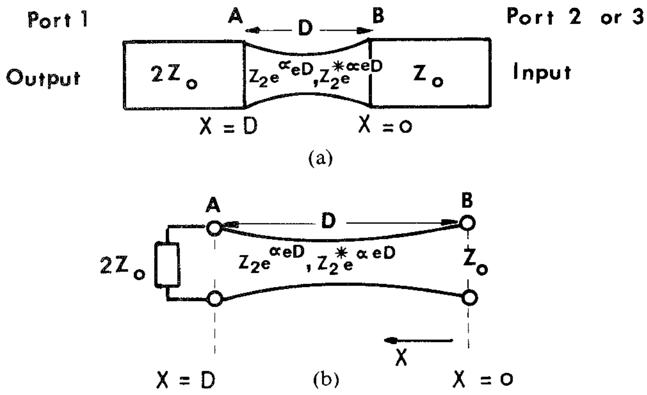

Fig. 3. Equivalent circuit of mode A.

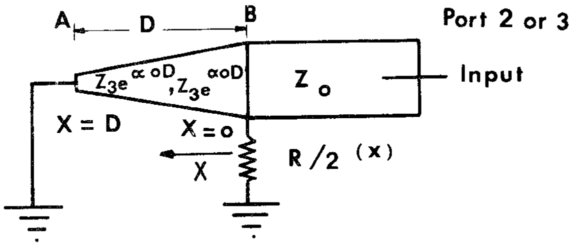

Fig. 4. Equivalent circuit of mode B.

The input impedance and reflection coefficient of mode B may be given, respectively, as

$$Z_{\mathrm{in}_{B_{\mathrm{mode\,(B)}}}}=\frac{j(2Z_3Z_3^*R\tan\beta D)}{R(Z_3+Z_3^*)+j\left[R(Z_3^*-Z_3)+4Z_3Z_3^*\right]\tan\beta D} \tag{11}$$

$$\rho_{B_{\mathrm{mode\,(B)}}}=\frac{-Z_0R(Z_3+Z_3^*)+j\left[2RZ_3Z_3^*-RZ_0(Z_3^*-Z_3)-4Z_3Z_3^*Z_0\right]\tan\beta D}{Z_0R(Z_3+Z_3^*)+j\left[2RZ_3Z_3^*+RZ_0(Z_3^*-Z_3)+4Z_3Z_3^*Z_0\right]\tan\beta D} \tag{12}$$

where Z_3 is the characteristic impedance of mode B at point $x=0$ (see Section VI) and Z_3^* is the complex conjugate of Z_3.

IV. The Parameters of the Combiner/Divider for a Uniform Transmission Line

For a uniform transmission line (UTL), $\alpha_e=\alpha_o=\alpha=0$, where α_e and α_o are the even- and odd-mode taper rate, respectively. For the uniform power combiner/divider, the characteristic impedances of the divider and the combiner are, respectively,

$$Z_1=\frac{Z_0}{\sqrt{2}}\ \text{and}\ Z_3=Z_3^*=Z_2=Z_2^*=\sqrt{2}\,Z_0.$$

$Z_0=Z_0^*$ and the load impedance $Z_L=2Z_0$ for the combiner case and $Z_0/2$ for the divider case. The divider's parameters may be found as

$$Z_{1n_{A_d}}=\frac{Z_0\left(1+j\sqrt{2}\tan\beta D\right)}{\sqrt{2}\left(\sqrt{2}+j\tan\beta D\right)} \tag{13}$$

$$\rho_{A_d}=\frac{1}{3+j\sqrt{2}\tan\beta D} \tag{14}$$

$$\tau_{A_d}=\frac{2\left(1+\tan^2\beta D\right)^{1/2}}{3+j2\sqrt{2}\tan\beta D} \tag{15}$$

while for mode A of the combiner, the input impedance and the reflection and transmission coefficients may be found as

$$Z_{\mathrm{in}_{B_{\mathrm{mode\,(A)}}}}=\frac{\sqrt{2}\,Z_0(2+j/2\tan\beta D)}{\sqrt{2}+j2\tan\beta D} \tag{16}$$

$$\rho_{B_{\mathrm{mode\,(A)}}}=\frac{1}{3+j2\sqrt{2}\tan\beta D} \tag{17}$$

$$\tau_{B_{\mathrm{mode\,(A)}}}=\frac{4\left(1+\tan^2\beta D\right)^{1/2}}{3+j2\sqrt{2}\tan\beta D}. \tag{18}$$

For mode B, the input impedance and reflection coefficient may be written as

$$Z_{\mathrm{in}_{B_{\mathrm{mode\,(B)}}}}=\frac{j\sqrt{2}\,(R\tan\beta D)Z_0}{R+j2\sqrt{2}\,Z_0\tan\beta D} \tag{19}$$

$$\beta_{B_{\mathrm{mode\,(B)}}}=\frac{-R+j\left[2(R-2Z_0)\tan\beta D\right]}{R+j\left[2(R+2Z_0)\tan\beta D\right]}. \tag{20}$$

E. The Effect of the Finite Thickness on the Coupled Lines

To arrive at more accurate practical designs for any coupled lines, that is, including the ETL structure, the capacitance effect which is caused by the finite thickness of the coupled line conductors must be taken into consideration. This capacitance causes an increase in the odd-mode capacitance and hence its impedance.

The capacitance due to the finite thickness of the coupled lines for both UTL and ETL may be computed. For the UTL,

$$C_t=\frac{tD\xi_0}{S} \tag{21}$$

where t is the thickness of the conductor, S is the separation between the conductors, D is the parallel conductor length, and $\xi_0=8.8419\times10^{-12}$ F/M. For the ETL exponential coupled-line conductor, this capacitance may be written as

$$C_t=\frac{tD\xi_0}{Se^{\alpha D}} \tag{22}$$

where $Se^{+\alpha D}$ is the separation between the coupled-line conductor. Usually the odd-mode capacitance is given [5] by

$$C_{\mathrm{odd}}=C_p+C_e+C_{ga}+C_{gd}. \tag{23}$$

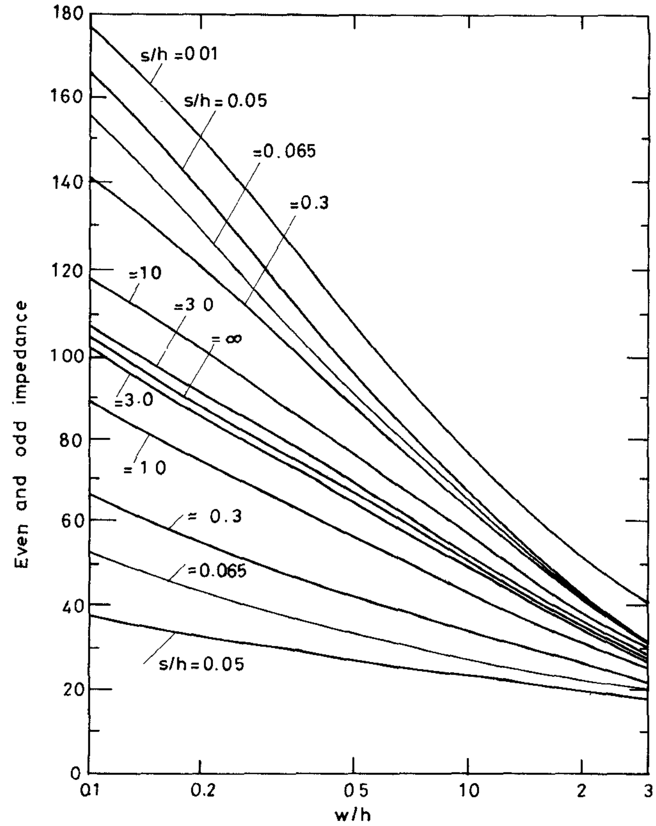

Fig. 5. Even- and odd-mode impedances versus w/h with s/h as parameter.

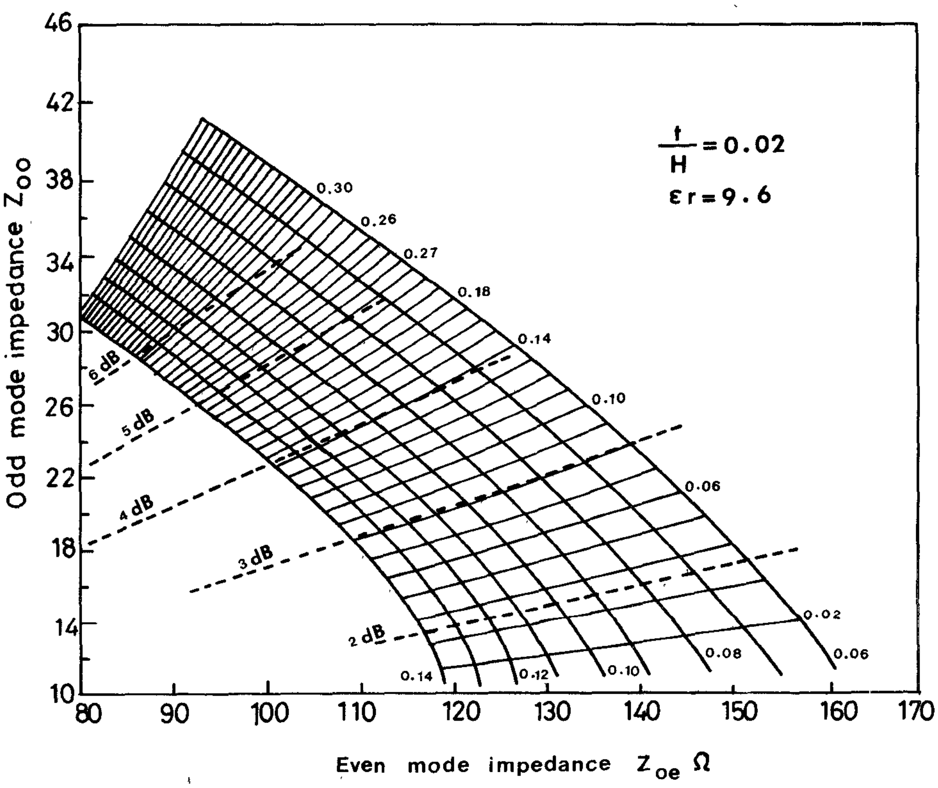

Fig. 6. Coupler design curves for alumina.

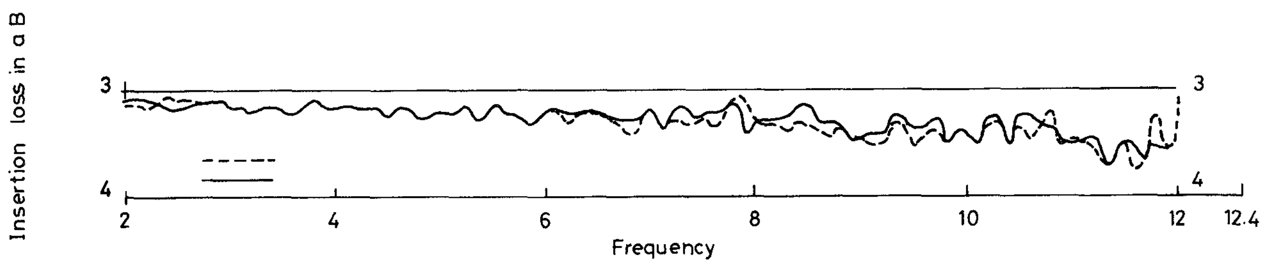

Fig. 7. Insertion loss of the combiner/divider versus frequency, in GHz.

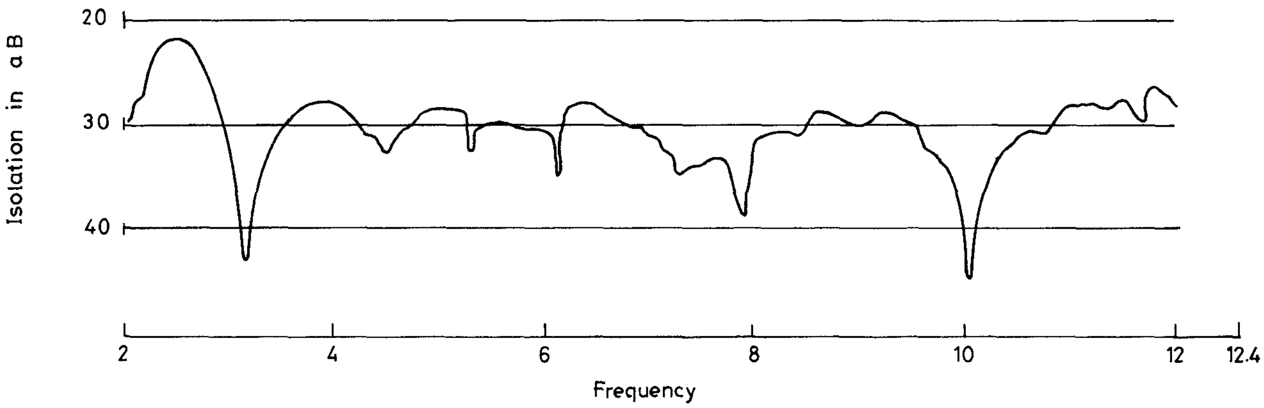

Fig. 8. Isolation of the combiner/divider versus frequency, in GHz.

All the previous terms in the above equation are defined in [5].

The modified odd-mode capacitance, therefore, can be expressed as

$$C_{\text{odd}} = C_p + C_e + C_{ga} + C_{gd} + C_t. \tag{24}$$

The modified odd-mode impedance is found as

$$Z_{\text{odd}} = \frac{1}{V\left[\left(C_p + C_{gd} + C_p\right)\left(C_{ga} + 2C_t\right)\right]^{1/2}} \tag{25}$$

where V is the odd-mode phase velocity.

VI. Microstrip Exponential Two-Way Power Combiner/Divider

For physical realization, the proposed combiner/divider's parameter must be known, i.e., Z_2, Z_3, D, α_e and α_o. The exponential length D is divided into small sections of equal intervals. $Z_A(x)$ and $Z_B(x)$ at each section of x are calculated according to

$$Z_A(x) = Z_2 e^{\alpha_e x} \tag{26}$$

$$Z_B(x) = Z_3 e^{\alpha_o x}. \tag{27}$$

The dimensions of w/h and S/h corresponding to the known values of $Z_A(x)$ and $Z_B(x)$ at each section of x are computed by using the graphs shown in Figs. 5 and 6. Only the initial values of mode A and mode B impedances at $x = 0$ are known. Therefore the spaces S between the exponential coupled lines, the thickness h of the substrate, and the width w of the exponential lines are assumed. The final shapes of the combiner/divider are obtained by constructing the corresponding points of x, w, and S and then connecting them by smooth curves as shown in Fig. 1. Mode A and mode B impedances at $x = 0$ are $Z_2 = 92.3\ \Omega$ and $Z_3 = 27.1\ \Omega$. The corresponding values of w and S are 0.328 mm and 0.009 mm, respectively. The value of the even-mode taper rate is taken as 2.0 m^{-1}.

VII. The Experimental Results

The proposed power combiner/divider is fabricated on alumina substrate with a dielectric constant of $\xi_r = 9.7$, a thickness of 0.635 mm, and a central frequency of operation of 5 GHz. A distributed resistance of value $R(x) = \sum_{N=1}^{100}(400 - 100N)$ is built on the combiner. The insertion loss of the combiner/divider is shown as a function of frequency in Fig. 7. The input signal is applied at port 1 and the output signals are measured at port 2 and then at port 3.

The combiner/divider behaves quite well, the maximum imbalance being of the order of 0.1 dB at all frequencies of operation (2 to 12.4 GHz). The isolation between ports 2 and 3 is reasonable, about 30 dB across the frequency band except for a slight ripple at 2.8 GHz (see Fig. 8). This could be improved by using several distributed resistances of different values. The return loss of ports 1, 2, and 3 of the combiner/divider is shown in Fig. 9. It is clear from the graph that the return loss of port 1 is better than 28 dB for the frequency range 3–10 GHz. It is also better than 24 dB for the frequency range 10–12.4 GHz. The return loss of port 2 or 3, however, is better than 27 dB for the frequency range 3–8 GHz and better than 25 dB for the frequency range 8–12.4 GHz.

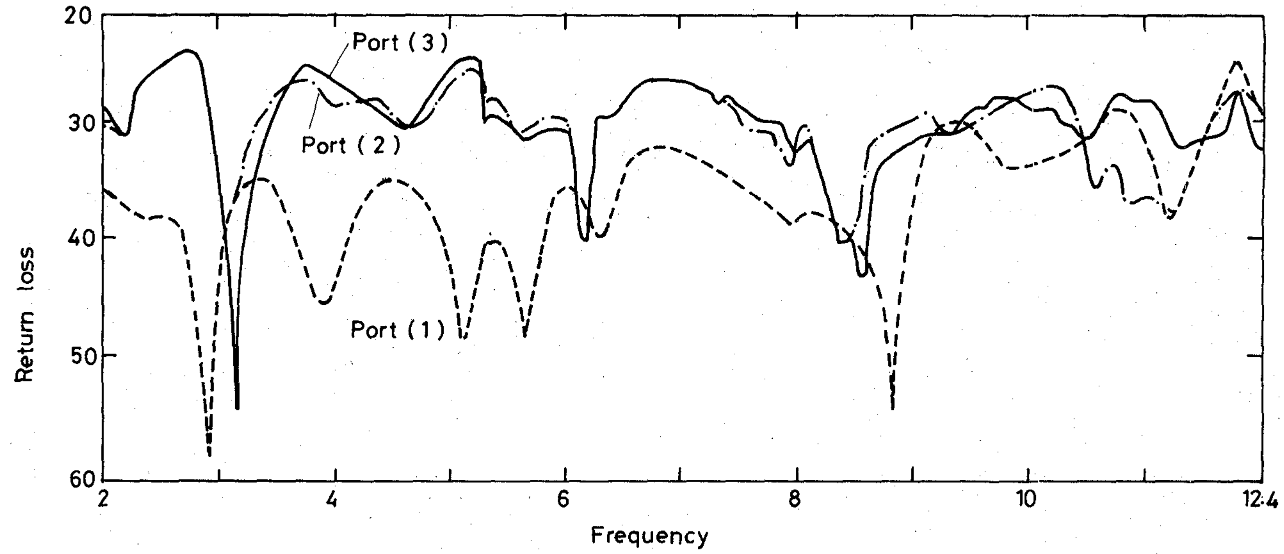

Fig. 9. Return loss of the combiner/divider versus frequency, in GHz.

VIII. Conclusion

By taking into consideration the increase in odd-mode capacitance of the coupled lines which is caused by the finite thickness of the coupled-lines conductor, more accurate design data can be generated for both uniform and nonuniform coupled lines. The proposed power combiner/divider was successively analyzed in terms of the even- and the odd-mode concept. The performance of the exponential power combiner/divider can be affected by the distributed resistance across the combiner. The experimental results show the capability of using this network as a combiner/divider successively.

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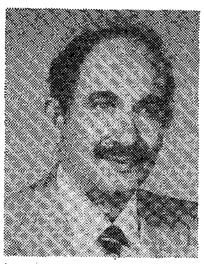

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