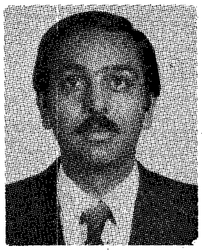
Department. He later became the Samuel Sachs Professor of Electrical Engineering and the Director of the Laboratory of Applied Electronic Services. He joined the University of California at San Diego, La Jolla, as a Professor of Electrical Engineering in 1979. His research ranges from surface-wave antennas, microwave masers, lasers, hypersonic plasma, and holography to integrated optics and the use of lidar for air-pollution monitoring.

Dr. Chang is a member of the American Physical Society, the American Optical Society, and the American Society of University Professors.



Siamak Forouhar (M'80) was born in Tehran, Iran, on August 25, 1950. He received the B.S. degree in electrical engineering from the Aryamehr Institute of Technology, Tehran, Iran, in 1972, and the M.S. degree in microwaves and modern optics from the University College, London, England, in 1975.

Since 1979, he has been working as a graduate student in the department of Electrical Engineering and Computer Sciences at the University of California at San Diego, La Jolla. He is working toward a Ph.D. degree with emphasis on the design and fabrication of grating lenses for integrated optical circuits.

Jean-Marc Delavaux (M'80) was born in Algiers, Algeria, on March 25, 1953. He received the degree of engineer from the Ecole Nationale Superieure d'Electronique, Toulouse, France, in 1978, and the M.S. degree in electrical Engineering from Washington University, St. Louis, in 1979.

Since 1979, he has been working as a graduate student in the department of Electrical Engineering and Computer Sciences at the University of California at San Diego, La Jolla. He is working toward a Ph.D. degree with emphasis on analysis and fabrication of grating lens structures.

Aperture Coupling Between Dielectric Image Lines

INDER J. BAHL, SENIOR MEMBER, IEEE, AND PRAKASH BHARTIA, SENIOR MEMBER, IEEE

Abstract—**Aperture coupling between dielectric image lines is used to develop a design technique for directional couplers at millimeter-wave frequencies. Expressions for coupling coefficients and directivity, employing coupling between image lines through apertures in the common ground plane are developed. The design procedure is illustrated by application to 10-, 20-, and 30-dB directional couplers in rectangular image lines with circular aperture coupling.**

I. Introduction

Dielectric image lines and their applications in active and passive devices for millimeter-wave integrated circuits have been reported in the literature [1]–[9].

Manuscript received December 16, 1980; revised April 27, 1981. This work was supported by the Natural Sciences and Engineering Research Council of Canada under Grant A-0001.

I. J. Bahl is with the Department of Electrical Engineering, University of Ottawa, Canada K1N 6N5.

P. Bhartia is with the Defence Electronics Division, Defence Research Establishment, Ottawa, Canada K1A 0Z4.

In this paper a design technique for directional couplers using dielectric image lines is given. Initially, a brief account of transmission-line properties of the dielectric image line is presented together with coupling between image lines through apertures in the common ground plane. Expressions are derived for the directivity and coupling coefficients and design curves are presented. Finally, the design procedure developed is illustrated by application to the design of 10-, 20-, and 30-dB directional couplers.

II. Image-Line Properties

The geometry of an image line, as shown in Fig. 1, comprises a rectangular dielectric slab of relative permittivity ϵ_r backed by a perfectly conducting ground plane. The main transverse field components of the E_{mn}^y modes are E_y and H_x. Omitting the $t-z$ dependence, $\exp[j(\omega t - k_z z)]$, where ω is the angular frequency and k_z is the propagation constant in the z direction; the field components inside and

0018-9480/81/0900-0891$00.75 ©1981 IEEE

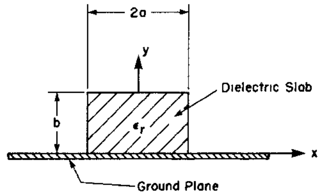


Fig. 1. Dielectric image-line configuration.

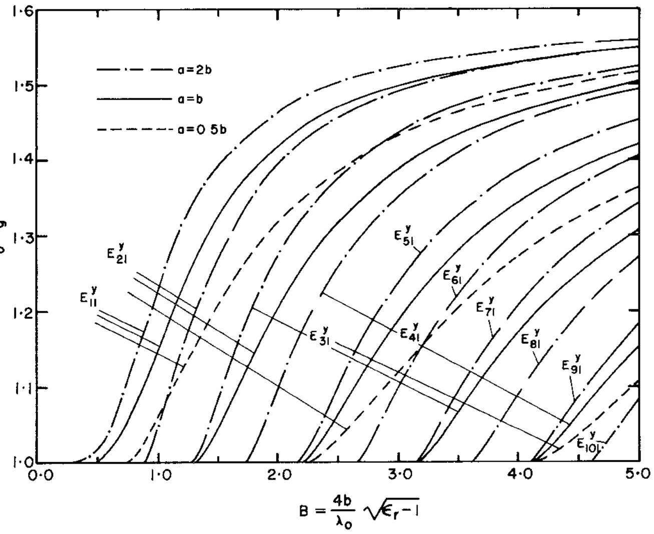


Fig. 2. Guide wavelength for various E^y_{m1} modes as a function of the normalized guide thickness B. $\epsilon_r = 2.5$.

outside the dielectric slab are given by [1]

$$E_y = E_0 \cos k_x x \cos k_y y, \qquad |x| \leq a,\ y \leq b \tag{1}$$

$$E_y = E_0 \cos k_x a \cos k_y y\, e^{-k_{x0}(|x|-a)}, \qquad |x| > a,\ y < b \tag{2}$$

$$E_y = E_0 \cos k_x x \cos k_y b\, e^{-k_{y0}(y-b)}, \qquad |x| < a,\ y > b \tag{3}$$

$$E_y = 0, \qquad |x| > a,\ y > b \tag{4}$$

$$H_x = -\frac{1}{\eta} E_y \tag{5}$$

where

$$\eta = \sqrt{\frac{\mu_0}{\epsilon_0}}\, \frac{1}{n_{\mathrm{eff}}} = \sqrt{\frac{\mu_0}{\epsilon_0}}\, \frac{k_0}{k_z} \tag{6}$$

and k_x, k_y, and k_{x0}, k_{y0} are the propagation constants in the x and y directions in the dielectric and air, respectively.

A typical set of dispersion curves for E^y_{mn} modes giving the ratio of free-space wavelength λ_0 to the guide wavelength λ_g as a function of normalized guide dimensions B $(=4b\sqrt{\epsilon_r - 1}/\lambda_0)$ is shown in Fig. 2. The relative permittivity of the dielectric slab is selected to be 2.5 and is "optimum" dielectric constant value for maximum bandwidth cosiderations [2].

III. Aperture Coupling

Dielectric image-line couplers so far reported in the literature employ a parallel coupled line approach. This type of coupler tends to be narrow band because the coupling is a result of the interference of two waveguide modes known as the even (symmetric) and odd (asymmetric) modes which propagate at different velocities. This difference in velocities also degrades the directivity characteristics. An alternative approach to the above is to use aperture coupling through the common ground plane, as shown in Fig. 3. This technique has been successfully used at microwave frequencies for closed waveguide directional couplers. The analysis, also valid for open waveguide structures is applied here to the dielectric image line neglecting the effect of radiation losses from the coupling holes or slots.

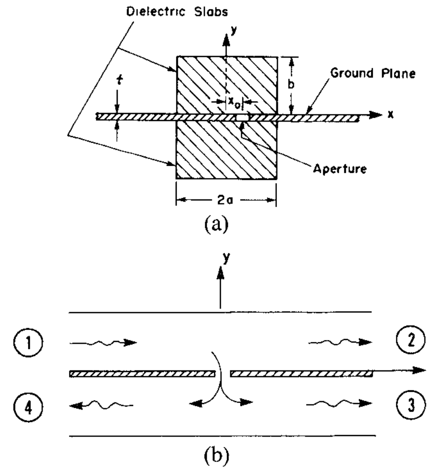


Fig. 3. (a) Aperture coupling configuration between dielectric image lines. (b) Illustration of the coupled waves through an aperture.

The electromagnetic coupling through small apertures in a common wall of rectangular waveguides has been well described in the literature [10]–[12]. The coupling from one line to the other may be calculated by evaluating the dipole moments associated with the aperture fields. Here for the sake of simplicity we assume that both dielectric guides are identical.

Consider two image lines having a common ground plane (Fig. 3) and coupled through an aperture of major dimension much less than the guide wavelength in one of the guides. Let α_e and α_m be the electric and magnetic polarizabilities of the aperture, respectively. The fields radiated by the electric dipole may be expressed as [13]

$$\bar{E}_e = \begin{cases} A_1 \bar{E}^+, & z > 0 \\ A_2 \bar{E}^-, & z < 0 \end{cases} \tag{7}$$

$$\bar{H}_e = \begin{cases} A_1 \bar{H}^+, & z > 0 \\ A_2 \bar{H}^-, & z < 0 \end{cases} \tag{8}$$

whereas that radiated by the magnetic dipole are

$$\bar{E}_m = \begin{cases} B_1 \bar{E}^+, & z > 0 \\ B_2 \bar{E}^-, & z < 0 \end{cases} \tag{9}$$

$$\bar{H}_m = \begin{cases} B_1 \bar{H}^+, & z > 0 \\ B_2 \bar{H}^-, & z < 0 \end{cases} \tag{10}$$

where the superscript + denotes propagation along the positive z direction and superscript − indicates propagation along the negative z direction. Considering only the E^y_{11} mode, the field components $\bar{E}^+$, $\bar{E}^-$, H^+, and $\bar{H}^-$ in the guide are given by (neglecting the H_z components)

$$\bar{E}^{+}=\hat{y}\cos k_x x\cos k_y y e^{-jk_z z}, \qquad z>0 \tag{11a}$$

$$\bar{E}^{-}=\hat{y}\cos k_x x\cos k_y y e^{jk_z z}, \qquad z<0 \tag{11b}$$

$$\bar{H}^{+}=-\frac{1}{\eta}\hat{x}\cos k_x x\cos k_y y e^{-jk_z z}, \qquad z>0 \tag{11c}$$

$$\bar{H}^{-}=\frac{1}{\eta}\hat{x}\cos k_x x\cos k_y y e^{jk_z z}, \qquad z<0. \tag{11d}$$

The expressions for A_1, A_2, B_1, and B_2 may be written as [10]

$$A_1=A_2=\frac{-j\omega}{P_n}\bar{E}^{-}\cdot\bar{P} \tag{12a}$$

$$B_1=-B_2=\frac{j\omega\mu_0}{P_n}\bar{H}^{-}\cdot\bar{M} \tag{12b}$$

where μ_0 is the free-space permeability and P_n is defined at $z=0$ as

$$P_n=2\int_0^b\int_{-a}^a\bar{E}^{+}\times\bar{H}^{+}\cdot\hat{z}\,dx\,dy$$
$$=\frac{1}{\eta}\left\{a+\frac{\sin(2k_x a)}{2k_x}\right\}\left\{b+\frac{\sin(2k_y b)}{2k_y}\right\}. \tag{13}$$

The electric and magnetic polarization vectors on the aperture which is located at $x=x_0$, $y=0$, $z=0$ are defined as [10]

$$\bar{P}=-\epsilon_0\epsilon_r\alpha_e\bar{E}=-\hat{y}\epsilon_0\epsilon_r\alpha_e\cos(k_x x_0) \tag{14}$$

where ϵ_0 is the free-space permittivity and

$$\bar{M}=-\alpha_m\bar{H}=\hat{x}\alpha_m\cos(k_x x_0)/\eta. \tag{15}$$

From (11), (12a), (13)–(15), the expressions for A_1, A_2, B_1, and B_2 on the aperture become

$$A_1=A_2=j\omega\epsilon_0\epsilon_r\alpha_e\eta$$
$$\cdot\frac{\cos^2(k_x x_0)}{[a+0.5\sin(2k_x a)/k_x][b+0.5\sin(2k_y b)/k_y]} \tag{16}$$

$$B_1=-B_2=\frac{j\omega\mu_0\alpha_m}{\eta}$$
$$\cdot\frac{\cos^2(k_x x_0)}{[a+0.5\sin(2k_x a)/k_x][b+0.5\sin(2k_y b)/k_y]}. \tag{17}$$

The total field radiated into the lower guide is readily evaluated from

$$\bar{E}=\begin{cases}(A_1+B_1)\bar{E}^{+}, & z>0\\(A_2+B_2)\bar{E}^{-}, & z<0\end{cases} \tag{18}$$

$$\bar{H}=\begin{cases}(A_1+B_1)\bar{H}^{+}, & z>0\\(A_2+B_2)\bar{H}^{-}, & z<0.\end{cases} \tag{19}$$

From (16)–(18), it can be shown that $A_1+B_1\neq 0$, which means that for the single hole case, there is always radiation in the forward direction, i.e., towards port (3) (Fig. 3 (b)). For no radiation in the backward direction, i.e., $A_2+B_2=0$, from (16), (17), we get

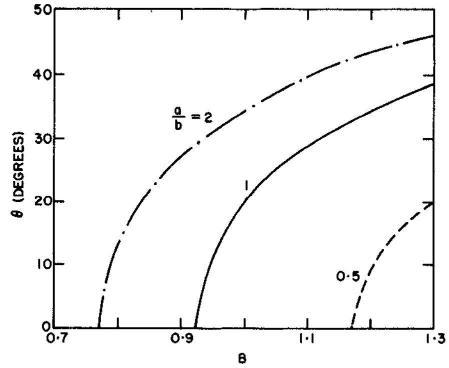


Fig. 4. Variation of the angle θ, between the axes of two image lines, versus B for no coupled waves in the backward direction $\epsilon_r=2.5$, E_{11}^y mode.

$$\frac{\epsilon_r}{n_{\text{eff}}^2}=\frac{\alpha_m}{\alpha_e}. \tag{20}$$

The expressions for electric and magnetic polarizabilities of commonly used aperture configurations are given in [13]. Thus it is possible to realize $A_2+B_2=0$ by selecting α_e and α_m. However, for a given aperture it is also possible to obtain $A_2+B_2=0$ by decreasing the value of the magnetic polarization by rotating the lower guide by an angle θ. Thus the magnetic dipole component is reduced by a factor $\cos\theta$, and (20) becomes

$$\frac{\epsilon_r}{n_{\text{eff}}^2}=\frac{\alpha_m}{\alpha_e}\cos\theta. \tag{21}$$

For a circular aperture

$$\theta=\cos^{-1}\left(\frac{\epsilon_r}{2n_{\text{eff}}^2}\right). \tag{22}$$

The variation of the angle θ as a function of B for various values of a/b ratio is shown in Fig. 4.

A. Coupling

The coupling coefficients for a circular aperture of radius r are defined by

$$C_f=-20\log|(A_1+B_1\cos\theta)|$$
$$=-20\log\left\{\frac{\frac{4}{3}\omega r^3\cos^2(k_x x_0)\left[\frac{1}{2}\epsilon_0\epsilon_r\eta+\mu_0\cos\theta/\eta\right]}{\left[a+\dfrac{\sin(2k_x a)}{2k_x}\right]\left[b+\dfrac{\sin(2k_y b)}{2k_y}\right]}\right\} \tag{23}$$

$$C_b=-20\log|(A_2+B_2\cos\theta)|$$
$$=-20\log\left\{\frac{\frac{4}{3}\omega r^3\cos^2(k_x x_0)\left|\frac{1}{2}\epsilon_0\epsilon_r\eta-\mu_0\cos\theta/\eta\right|}{\left[a+\dfrac{\sin(2k_x a)}{2k_x}\right]\left[b+\dfrac{\sin(2k_y b)}{2k_y}\right]}\right\} \tag{24}$$

where the subscripts b and f denote coupling in the back-

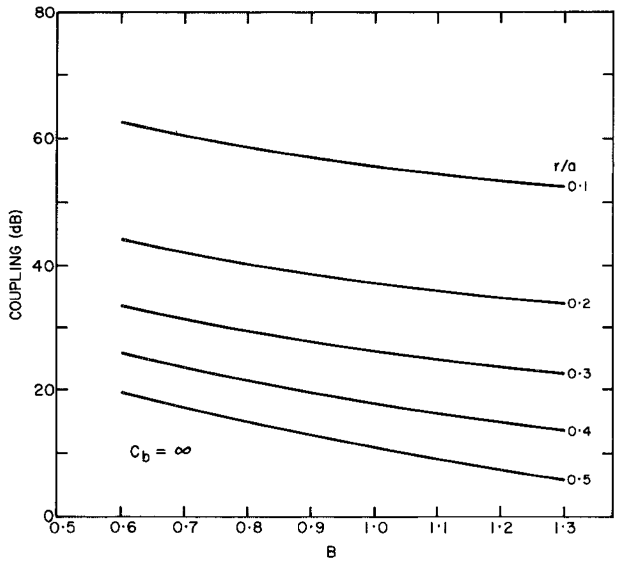


Fig. 5. Coupling in the forward direction through a circular hole as a function of B for no coupled waves in the backward direction. $\epsilon_r = 2.5$, $a/b = 1.0$, E_{11}^y mode.

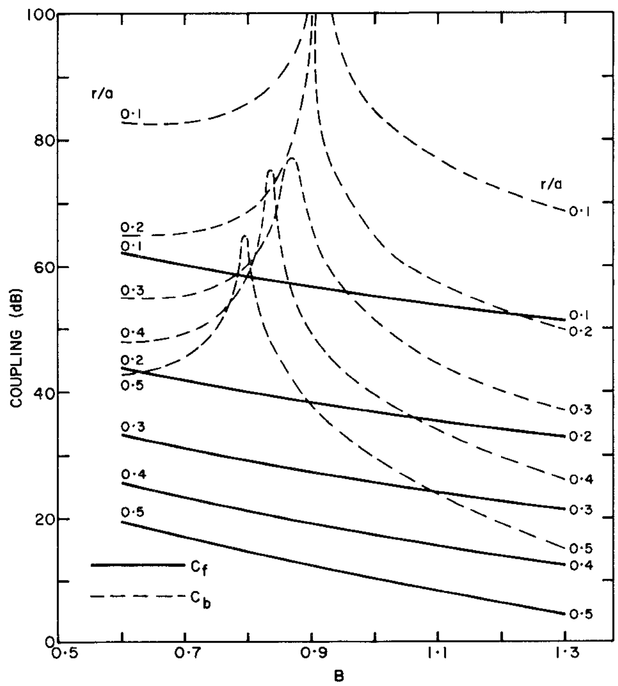


Fig. 6. Coupling through a circular hole versus B for $\epsilon_r = 2.5$, $a/b = 1.0$, and E_{11}^y mode.

ward and forward directions, respectively. If the angle θ is selected using (22), the coupling in the backward direction will be zero.

Calculated values of the coupling in the forward direction (when coupling in the backward direction is zero) for $\epsilon_r = 2.5$, $a/b = 1.0$, $f = 35$ GHz, $x_0 = 0$ and the circular aperture, as a function of B are shown in Fig. 5. For the same set of parameters and $\theta = 0$ (axes of the two lines are parallel), the variation of the coupling in both the directions is shown in Fig. 6. It may be noted from this figure that the coupling in the forward direction is stronger than in the backward direction. The coupling for the rectangular slot were also evaluated and found more or less to be same in both the directions. Fig. 7 depicts the variation of C_f with B.

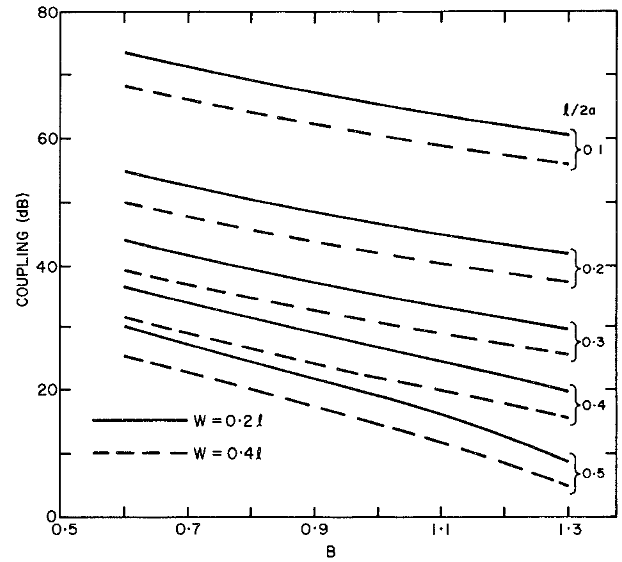


Fig. 7. Coupling through a slot as a function of B for $\epsilon_r = 2.5$, $a/b = 1.0$, and E_{11}^y mode.

B. Directivity

The directivity is given by

$$D = 20 \log \left| \frac{A_1 + B_1 \cos\theta}{A_2 + B_2 \cos\theta} \right|$$

$$= 20 \log \left| \frac{\frac{1}{2}\epsilon_0 \epsilon_r \eta + \mu_0 \cos\theta / \eta}{\frac{1}{2}\epsilon_0 \epsilon_r \eta - \mu_0 \cos\theta / \eta} \right|. \quad (25)$$

C. Effect of Ground Plane Thickness and Large Size Apertures

In the derivation of (23)–(25) it has been assumed that the thickness of the common ground plane t is negligible and the aperture size is small compared with its resonant dimensions. The correction factor due to the combined effects of ground plane thickness and large aperture size is given by [11], [12]

$$\mathrm{CF} = \frac{\exp\left\{(-2\pi t A/\lambda_c)\left[1 - \left(\frac{\lambda_c}{\lambda_0}\right)^2\right]^{1/2}\right\}}{1 - \left(\frac{\lambda_c}{\lambda_0}\right)^2}$$

where λ_0 is the operating wavelength and λ_c is the cutoff wavelength of the aperture considered as a waveguide operating in a mode appropriate to the particular form of excitation. The factor A accounts for the interaction of local fields on either side of the aperture and has been determined empirically [14] to be 3 for a narrow slot and unity for a circular hole. For negligible ground plane thickness or apertures filled with a dielectric, simple expressions for the cutoff wavelength of a rectangular slot and a circular hole are given in Table I. Thus the effect of large size aperture may be incorporated in (23)–(25) by

TABLE I
CUTOFF WAVELENGTHS FOR CIRCULAR AND RECTANGULAR APERTURES

Aperture	Cut off Mode	λ_c Magnetic coupling	λ_c Electric coupling
Circular hole ($2r$)	TE_{11}	$3.41\sqrt{\epsilon_r}\, r$	
	TM_{01}		$2.61\sqrt{\epsilon_r}\, r$
Rectangular slot (ℓ, W)	TE_{10}	$2\sqrt{\epsilon_r}\, \ell$	
	TM_{11}		$\frac{\sqrt{\epsilon_r}}{[(\pi/\ell)^2+(\pi/W)^2]^{\frac{1}{2}}}$

just multiplying A_1, A_2 by CF with λ_c related for electric coupling and B_1, B_2 by CF with λ_c related for magnetic coupling from Table I.

IV. DESIGN OF A DIRECTIONAL COUPLER

When a number of apertures spaced $\lambda_g/4$ apart are used in the common ground plane, the total amplitude of the backward coupled waves is zero and the structure gives rise to good directivity and tight coupling over the band of frequencies. For an N element coupler, the expressions for the coupling and directivity are given by

$$C=-20\log\left(\left|\sum_{n=1}^{N} C_{fn}\right|\right) \tag{26}$$

$$D=20\log\left\{\frac{\left|\sum_{n=1}^{N} C_{fn}\right|}{\left|\sum_{n=1}^{N} C_{bn}\exp\left(j2k_z\{n-1\}d\right)\right|}\right\} \tag{27}$$

where C_{fn} and C_{bn}, $n=1,2,\cdots,N$ are the coupling coefficients in the forward and backward directions, respectively, and d is the physical distance between two adjacent apertures.

If we express C_{fn} and C_{bn} as products of frequency-independent amplitude constants (c_n) and frequency-dependent factors (T_b, T_f):

$$C_{fn}=c_nT_f \quad C_{bn}=c_nT_b$$

then (26) and (27) become

$$C=-20\log\left(|T_f\sum_{n=1}^{N} c_n|\right) \tag{28}$$

$$D=-C-20\log|T_b|-20\log F \tag{29}$$

where F is known as the array factor and is given by

$$F=|\sum_{n=1}^{N} c_n\exp\left(j2k_z\{n-1\}d\right)|. \tag{30}$$

Design of a directional coupler requires an appropriate choice of the array factor. If a minimum value of the directivity over a given frequency band is required, then the array factor can be designed using Chebyshev polynomials [10], [11].

Consider an example of a five hole 30-dB coupler with minimum directivity of 30 dB. For $\epsilon_r=2.5$, $f_0=35$ GHz, $a/b=1.0$, and for the dominant E_{11}^y mode operation, the value of B (i.e., normalized thickness of the guide) can be selected depending upon the bandwidth requirement. Suppose we are interested in a frequency band 30–40 GHz, then $B=1.1$ will be a better choice as higher values of B give higher coupling. The calculated values of various parameters of the image line at 35 GHz are $a=b=1.925$ mm, $\lambda_g=7.184$ mm and spacing between the circular holes $d=\lambda_g/4=1.796$ mm. The radii of the holes for a Chebyshev taper are $r_1=r_5=0.195$ mm, $r_2=r_4=0.298$ mm, and $r_3=0.337$ mm.

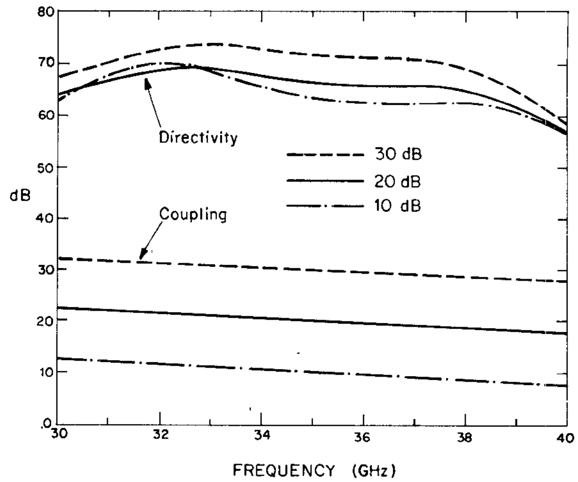


Fig. 8. Variations of the coupling and directivity of 10-, 20-, and 30-dB directional couplers with frequency.

The calculated performance of the directional coupler is shown in Fig. 8. Over the band (30–40 GHz), the variation in the coupling is less than 2.4 dB and the directivity is better than 59 dB. Since at millimeter wave frequencies, it is easy to accommodate a large number of holes, one can improve the performance further by selecting higher order Chebyshev polynomials.

Fig. 8 also depicts the performance characteristics of the 20- and 10-dB directional couplers designed using the same procedure to operate over the 30–40-GHz band. In this case, with all other dimensions remaining the same, $r_1=r_5=0.292$ mm, $r_2=r_4=0.43$ mm, and $r_3=0.484$ mm for the 20-dB coupler, and $r_1=r_5=0.438$ mm, $r_2=r_4=0.616$ mm, and $r_3=0.677$ mm for the 10-dB coupler.

V. CONCLUSION

This paper presents a design technique for directional couplers using dielectric image lines. Transmission-line properties of the dielectric image line together with coupling between image lines through circular holes and rectangular slots have been discussed. Analysis shows that as in the case of rectangular waveguide couplers, the coupling in the forward direction using a single aperture in the common ground plane of the image lines cannot be made zero. Finally, the design procedure developed is illustrated by application to the design of 10, 20, and 30-dB directional couplers.

REFERENCES

[1] R. M. Knox and P. P. Toulios, "Integrated circuits for the millimeter through optical frequency range," in *Proc. Symp. Submillimeter Waves*, (Polytechnic Press of Polytechnic Ins. of Brooklyn, NY), pp.

497–516, 1970.
[2] R. J. Collier and R. D. Birch, "The bandwidth of image guide," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 932–935, Aug. 1980.
[3] S. Shindo and T. Itanami, "Low-loss rectangular dielectric image line for millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 747–751, Oct. 1978.
[4] K. Solbach, "The measurement of the radiation losses in dielectric image line bends and the calculation of a minimum acceptable curvature radius," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 51–53, Jan. 1979.
[5] K. Solbach, "The calculation and the measurement of the coupling properties of dielectric image lines of rectangular cross section," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 54–58, Jan. 1979.
[6] J. A. Paul and Y. W. Chang, "Millimeter-wave image-guide integrated passive devices," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 751–754, Oct. 1978.
[7] B-S Song and T. Itoh, "A distributed feedback dielectric waveguide oscillator with a built-in leaky-wave antenna," in *IEEE MTT-S Int. Microwave Symp. Digest*, pp. 217–219, 1979.
[8] H. Jacobs and M. M. Chrepta, "Electronic phase shifter for millimeter-wave semiconductor dielectric integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 411–417, Apr. 1974.
[9] R. M. Knox, "Dielectric waveguide microwave integrated circuits—An overview," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 806–814, Nov. 1976.
[10] R. E. Collin, *Foundations for Microwave Engineering.* New York: McGraw-Hill, 1966, pp. 183–190.
[11] R. Levy, "Directional couplers," in *Advances in Microwaves*, vol. 1, L. Young, Ed, New York: Academic, 1966, pp. 164–191.
[12] R. Levy, "Analysis and synthesis of waveguide multiaperture directional couplers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 995–1006, Dec. 1968.
[13] M. Kumar and B. N. Das, "Coupled transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 7–10, Jan. 1977.
[14] S. B. Cohn, "Coupling by large apertures," *Proc. IRE*, vol. 40, pp. 696–699, June 1952.

Inder J. Bahl (M'80) was born in Sham Chaurasi, Punjab, India on January 27, 1944. He received the M.Sc. degree in physics and M.Sc. (Tech.) in electronics from the Birla Institute of Technology and Science, Pilani, India, in 1967 and 1969, respectively. In 1975, he received the Ph.D. degree in electrical engineering from the Indian Institute of Technology (IIT), Kanpur, India.

From 1969 to 1970, he worked in Tropo Scatter Communication Project, in the Department of Electrical Engineering at IIT, Kanpur. During 1971 to 1974, he was Senior Research Assistant in the same department. From 1974 to 1978, he was with the Advanced Centre for Electronic Systems, IIT, Kanpur as a Research Engineer, where he was engaged in research in p-i-n diode phase shifters, microwave integrated circuits, printed antennas, phased-array radar, and industrial applications of microwaves. In January 1979, he joined the Department of Electrical Engineering, University of Ottawa, Ottawa, Ont., Canada where he is presently a Research Associate. He is now working on microwaves in biological systems, microwave and millimeter-wave integrated circuits, microstrip antennas and millimeter-wave antennas. Dr. Bahl is the author of over 40 research papers in the areas of microwave theory and techniques as well as antennas. He published two books *Microstrip Lines and Slotlines* (Dedham, Artech House, 1979) and *Microstrip Antennas* (Artech House, 1980).

Prakash Bhartia (S'68–M'71–SM'76) was born in Calcultta, India in 1944. He received the B. Tech. (Hons) degree in electrical engineering from the Indian Institute of Technology, Bombay, in 1966, and the M.Sc. and Ph.D. degrees from the University of Manitoba, Winnipeg, Canada, in 1968 and 1971, respectively.

From 1971 to 1973 he was a research associate at the University of Manitoba and joined the faculty of Engineering at the University of Regina in 1973. In 1977 he moved to the Defence Research Establishment, Ottawa and was simultaneously appointed a Non-Resident Professor at the University of Ottawa. He has conducted research in both theoretical and applied electromagnetics, scattering, diffraction, radio, satellite and inertial navigation, etc., and is currently working in the areas of Electromagnetic Compatibility, Electromagnetic Pulse and Nuclear vulnerability and survivability of satellites. He has coauthored the book *Microstrip Antennas* (Dedham, Artech House, 1980).

Dr. Bhartia is a member of Institute of Electrical Engineers (England), International Microwave Power Institute, Professional Engineer in the Province of Saskatchewan and Commission B of URSI. He is on the Editorial Board of MTT and AP and an Associate Editor for the *Journal of Microwave Power.*