this MIC mixer. A discrete-component IF amplifier is built into the mixer housing. IF matching is incorporated into the amplifier circuitry. An even more compact mixer/IF preamplifier can be configured by utilizing a subminiature flat pack or TO-cased IF amplifier. As has been noted, an overall double-sideband noise figure of 6 dB maximum has been obtained with this MIC mixer preamplifier (including a 2.5-dB IF contribution). The diodes are self-biased, with a total LO drive of +13 dBm. The measured LO-to-RF isolation is 15 dB/min, and the VSWR is typically 1.5:1.

This mixer is attractive in terms of both size and cost. The LO and signal waveguides with the required cross-sectional area for mating to standard flanges account for most of the volume. By utilizing coax inputs [4] or integrating more front-end components, the full miniaturization potential of K_a-band MIC construction will be realized. Low cost is achieved because the substrate is readily etched and machined, and the housing does not require tight tolerances.

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REFERENCES

[1] W. J. Getsinger, "Microstrip dispersion model," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 34–39, Jan. 1973.
[2] T. H. Oxley *et al.*, "Hybrid microwave integrated circuits for millimeter wavelengths," in *1972 IEEE-GMTT Int. Microwave Symp. Digest of Technical Papers*, pp. 224–226.
[3] D. L. Saul, "Wideband microstrip components and the IFM discriminator," in *1974 Millimeter Waves Conf. Digest*, vol. 2, pp. 4D-1–D4-12, March 1974.
[4] M. A. Maury, Jr., and W. A. Wambach, "A new 40 GHz connector," in *1974 Millimeter Waves Conf. Digest*, vol. 2, pp. D9-1–D9-9, March 1974.

Dielectric Rod Antennas for Millimeter-Wave Integrated Circuits

YIH SHIAU

***Abstract*—The design of dielectric rod antennas for millimeter-wave integrated-circuit applications is described. The experimental investigation was initially performed for scaled models at *Ku* band and then developed at *V* band. A moderately high-gain alumina dielectric rod antenna that is entirely compatible with insular integrated circuits has been designed and tested. The antenna has been fabricated and integrated, as one of the system components, into short-range *V*-band transmitter and receiver modules. The measured gain was found to be 15.2 dB. Radiation characteristics are discussed.**

I. INTRODUCTION

The investigations of microwave integrated circuits employing dielectric insular waveguides are finding broad usage in short-range communication systems at millimeter-wave frequencies [1], [2]. The millimeter-wave insular line integrated circuits (MILIC) technology offers many advantages such as low production costs, high reliability, compact size, and high target resolution, etc. [1], [3]. However, the design of the transmitter and receiver units using MILIC technology has indicated an ever-increasing need for a special antenna design. The desired antenna, which must be compatible with integrated-circuit philosophy, should result in a relatively high gain for the transmitter and receiver. The antenna structure, when integrated into the system, should also have the benefits of small size and economical solution to the problem.

The subject addressed in this short paper, in addition to providing a brief description of the theoretical design data, focuses on the experimental design of the millimeter-wave dielectric rod antenna. As a result, an alumina dielectric rod antenna with a gain of 15.2 dB has been successfully designed and fabricated as an easily replaceable component for the transmitter and receiver modules [4].

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The author is with the IIT Research Institute, Chicago, IL 60616.

II. ANTENNA DESIGN DATA

A satisfactory explanation of the operation of dielectric rod antennas can be obtained by establishing analogies between the end-fire array and the dielectric rod. A very good treatment of this theory of electromagnetic wave propagation along dielectric rods has been given by Kiely [5]. Since the theory has been utilized and proven to be very useful for engineering applications at the microwave frequency range by many investigators [5]–[8], it is used here to examine certain features of the radiation characteristics and the gain of a dielectric rod at *Ku* and *V* band.

From the end-fire array model for dielectric rod antennas, it has been found that the directivity of the dielectric rod increases as $\gamma = \lambda_0/\lambda$ increases, where λ and λ_0 are the wavelength in the dielectric and in free space, respectively. However, a value of γ as high as 1.1 can only be used to give optimum radiation characteristics [5]. This optimum is obtained from the assumption that no standing waves exist along the dielectric rod. This implies that the rod has to be matched to free space at the end of the antenna. Simple design formulas for the diameter of the dielectric rod near the feed $d_{\max}$ and the diameter of the rod at the free end $d_{\min}$ can now be obtained. These useful formulas, theoretically determined from the conditions $\gamma = 1.1$ and $\gamma = 1$, are [5]

$$d_{\max} \approx \frac{\lambda_0}{\sqrt{\pi(\varepsilon_r - 1)}} \tag{1}$$

and

$$d_{\min} \approx \frac{\lambda_0}{\sqrt{2.5\pi(\varepsilon_r - 1)}}. \tag{2}$$

From an antenna designer's point of view, the two simple relations, given by (1) and (2), may be used as working formulas in the design of dielectric rod antennas.

III. EXPERIMENTAL DESIGN

Based on the design data given by (1) and (2), a total of 20 different antennas identified as having substantial merits were constructed and tested. Experimental measurements of radiation pattern and antenna gain were first performed at *Ku* band. One antenna configuration investigated is shown in Fig. 1(a). Using the HI-K707L material with a relative dielectric constant of about 9.8 as the dielectric rod medium, a cross-sectional dimension of approximately 160 mil at the feed end was used for the test antenna. The dielectric rod was then progressively and uniformly tapered to $d_{\min}$ in both planes at a distance L from the feed point. The value of $d_{\min}$ was designed to be about 110 mil. Pattern and gain measurements for the antenna were taken for various lengths L. In Fig. 1, results for three of those variations are shown. The corresponding gains measured were found to increase from about 14.5 dB for the $L = 5\lambda_0$ case to approximately 16.2 dB for $L = 8\lambda_0$. The measurement for the gain

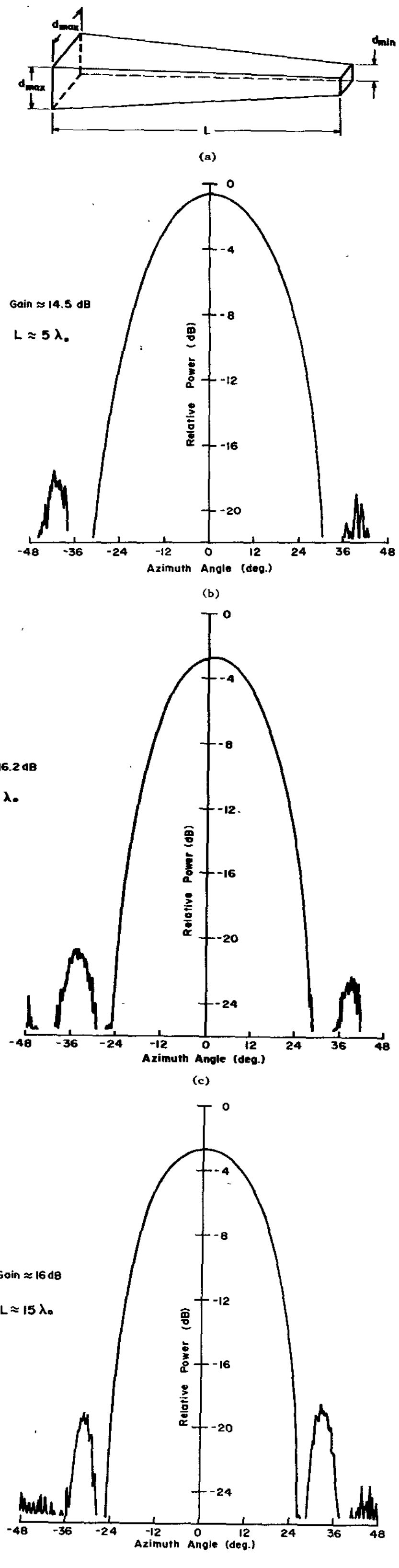

Fig. 1. Measured patterns at *Ku* band for square antenna cross section.

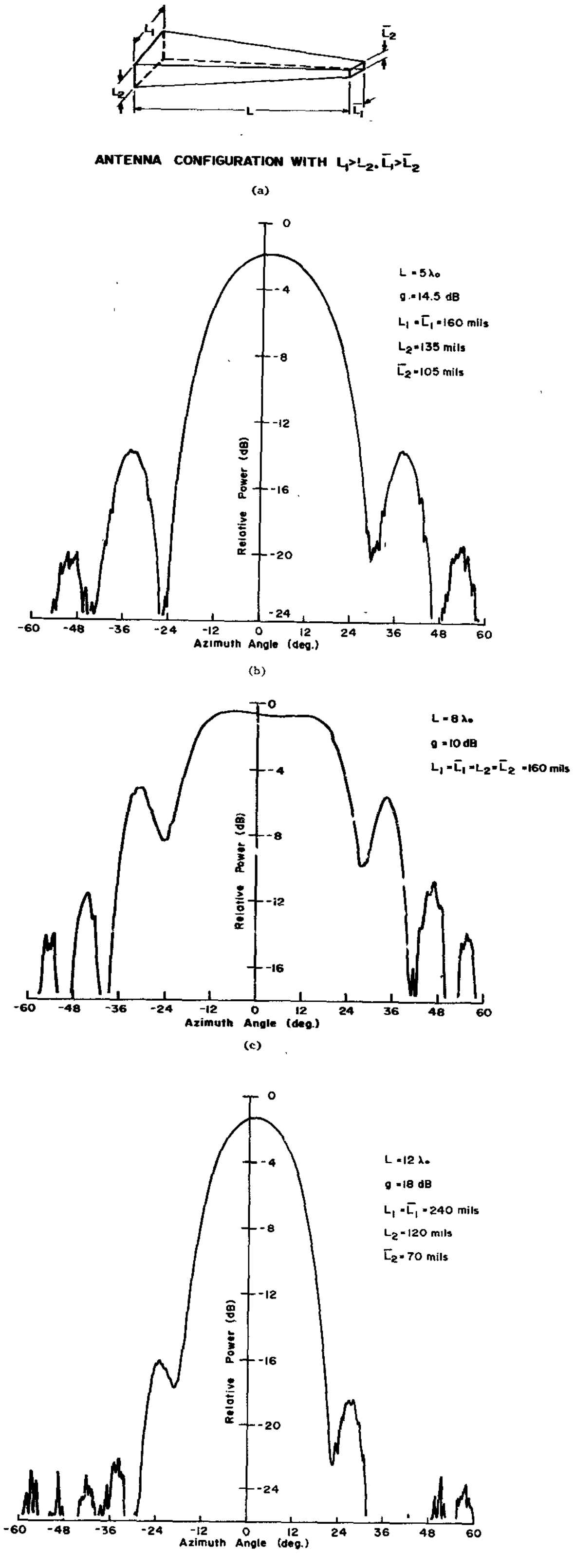

Fig. 2. Measured patterns at *Ku* band for rectangular antenna cross section.

of the $L = 15\lambda_0$ antenna did not reveal any increase in gain as compared to that of $L = 8\lambda_0$.

For the purpose of improving the antenna gain, the *Ku*-band antenna configuration was then made to have a wider width in the top and bottom walls of the dielectric rod. Some of the measured patterns and antenna gains are graphically presented in Fig. 2. The maximum gain that was achieved was about 18 dB. This antenna had a cross-sectional dimension of about 240 by 120 mil at the feed end, which was then gradually tapered to about 240 by 70 mil over a length of approximately 12 wave-

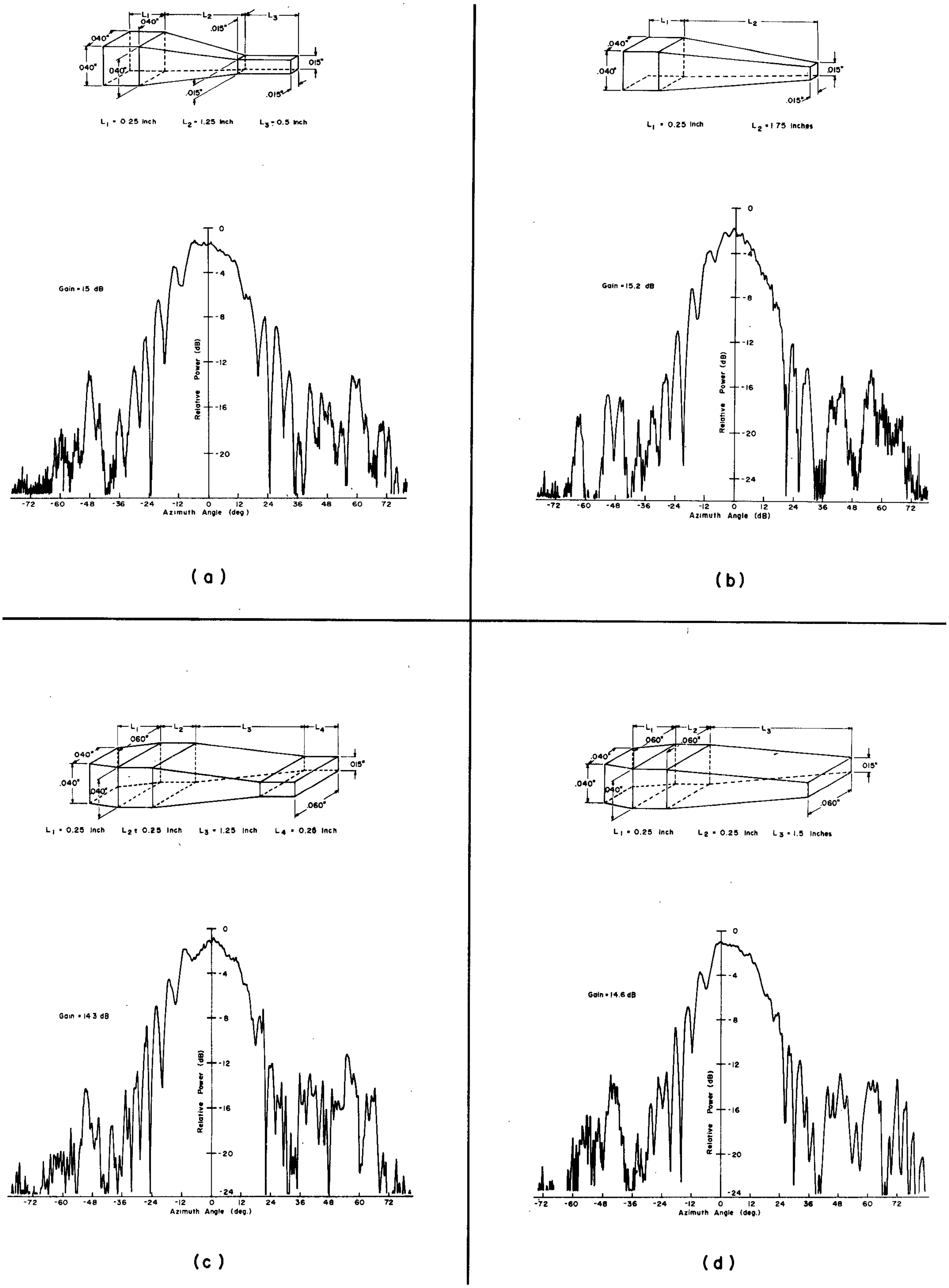

Fig. 3. Antenna configurations and measured patterns at V band.

lengths. Its corresponding radiation pattern is shown in Fig. 2(d). Increasing the antenna length to about 18 wavelengths did not reveal any significant enhancement in antenna gain.

Based on the results of the experiments conducted at *Ku* band, four different antenna structures were scaled and constructed at *V* band. In designing these *V*-band antennas, special attention was given to the possible interface problem between the dielectric insular waveguide and the antenna. The cross-sectional dimension in the interface region was made to have a transition as smooth as possible, such as to minimize any possible mismatch between the dielectric waveguide and the antenna. The various configurations are shown in Fig. 3 together with their pattern and gain measurements. The highest gain obtained is 15.2 dB corresponding to the measured radiation pattern of Fig. 3(b).

In general, the empirical results agree well with the design formulas given by (1) and (2). It was found that for a given relative dielectric constant, the optimum radiation characteristics can be obtained if the cross-sectional dimension $d_{\max}$ is properly chosen. For HI-K707L material, a dimension of $d_{\max}/\lambda_0 \approx 0.2$ was observed to provide the maximum gain. The gain of a dielectric rod is initially increased by increasing the antenna length. However, the relationship is not linear. For sufficiently long length, a considerable increase in antenna length will not result in a corresponding increase in the gain of the antenna. Instead, a slight decrease in antenna gain had been noticed after about 14 wavelengths in some antenna configurations investigated. The maximum gain realized was about 18 dB at *Ku* band and 15.2 at *V* band.

IV. Concluding Remarks

The effort to design a millimeter-wave antenna suitable for millimeter-wave integrated-circuit applications has been described. The results of these measurements were evaluated. The antenna of Fig. 3(b) was selected as the most suitable configuration for this MILIC application. This antenna was fabricated and integrated directly into the transmitter and the receiver modules as an easily replaceable component [9]. The antenna is bonded to a T-shaped fixture which is fastened to the main plate with two screws. The insular waveguide on the antenna fixture and main plate are butted together to form a continuous energy path. Since the dielectric antenna protrudes out of the transmitter case, a protective polystyrene radome has been provided in order to prevent breakage. The addition of a single radome adds approximately 0.2-dB loss to the overall system. The receiver module [9] used the same dielectric rod antenna. The receiver antenna is also replaceable in case of damage and is protected by a polystyrene radome.

To provide a higher gain for MILIC technology, the consideration of possible alternative techniques with regard to millimeter-wave integrated-circuit applications and requirements is being considered. A detailed discussion on the empirical results described previously and the limitations of the existing concepts of utilizing a simple dielectric rod for MILIC applications has been reported elsewhere [4].

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References

[1] E. A. J. Marcatili, "Dielectric rectangular waveguide and directional coupler for integrated optics," *Bell System Technical Journal*, vol. 48, no. 7, pp. 2079–2102, Sept. 1969.
[2] R. M. Knox and P. P. Toulios, "Integrated circuits for the millimeter through optical frequency range," *Proc. of the Symposium on Submillimeter Waves*, New York, March 31–April 2, 1970.
[3] P. P. Toulios and R. M. Knox, *Image Line Integrated Circuits for System Applications at Millimeter Wavelengths*, U.S. Army Electronics Command, Contract ECOM-73-0217-F, July 1974.
[4] B. J. Levin and J. E. Kietzer, *Hybrid Millimeter-Wave Integrated Circuits*, U.S. Army Electronics Command, Contract ECOM-74-0577-F, Oct. 1975.
[5] D. G. Kiely, *Dielectric Aerials*. London, England: Methuen and Co., Ltd., 1952.
[6] G. E. Mueller and W. H. Tyrrell, "Polyrod antennas," *Bell System Technical Journal*, vol. 26, p. 837.
[7] F. E. Butterfield, "Dielectric sheet radiators," *IRE Trans. Antennas Propag.*, p. 152, Oct. 1954.
[8] G. C. Southworth, "Principles and applications of waveguide transmission," *Bell Laboratories Series D*, Van Nostrand Co., Inc., Princeton, NJ, 1950.
[9] J. E. Kietzer, A. R. Kaurs, and B. J. Levin, "A *V*-band communication transmitter and receiver system using dielectric waveguide integrated circuits," this issue, pp. 797–803.

E-Type Modes in Cylindrical Dielectric Waveguides with Periodic Boundary Perturbations

ERCAN TOPUZ, MEMBER, IEEE

Abstract—**A general method for the analysis of the effects of periodic boundary perturbations on *E*-type modes of dielectric waveguides (DW's) is presented. The method is applied to hollow cylindrical dielectric waveguide (HCDW), and numerical solutions for surface wave and radiation mode intensities are given.**

The effects of periodic boundary perturbations on *E*-type modes of dielectric waveguides (DW's) have not been investigated. Existing asymptotic approximations [1]–[4] are only valid in the optical region, where core-cladding refractive index differences are orders of magnitude smaller than in the millimeter region. In the present short paper we extend the work of Marcuse and Derosier [5], [6] for *H*-type modes of DW to include *E*-type modes and, taking hollow cylindrical dielectric waveguide (HCDW) as a specific example, give approximate solutions of surface wave and radiation intensities.

HCDW is shown in Fig. 1 with sinusoidally perturbed boundaries. The modes of the unperturbed guide are well known [7]. *E*-type modes exist only in case of cylindrical symmetry, and with $H_\phi = h_\phi \exp \pm j\beta z$ we have for surface wave modes

$$h_{\phi m} = \begin{cases} j\omega\varepsilon_0/\Gamma_m C_{1m} I_1(\Gamma_m r), & r < a_1 \\ j\omega\varepsilon_0\varepsilon_r/\Omega_m [C_{2m} J_1(\Omega_m r) + D_{2m} Y_1(\Omega_m r)], & a_1 < r < a_2 \\ -j\omega\varepsilon_0/\Gamma_m D_{3m} K_1(\Gamma_m r), & r > a_2 \end{cases}$$

and for radiation modes

$$h_\phi = \begin{cases} j\omega\varepsilon_0/\overline{\Omega}_1 A_1 J_1(\overline{\Omega}_1 r), & r < a_1 \\ j\omega\varepsilon_0\varepsilon_r/\overline{\Omega}_2 \{A_2 J_1(\overline{\Omega}_2 r) + B_2 Y_1(\overline{\Omega}_2 r)\}, & a_1 < r < a_2 \\ j\omega\varepsilon_0/\overline{\Omega}_3 \{A_3 J_1(\overline{\Omega}_3 r) + B_3 Y_1(\overline{\Omega}_3 r)\}, & r > a_2 \end{cases} \quad (2)$$

where J_1 and Y_1 are Bessel functions, I_1, K_1 are modified Bessel functions of the first and second kind and of the first order, Γ and Ω are radial propagation constants given as

$$\Gamma_m = \sqrt{\beta_m^2 - k_0^2} \qquad \Omega_m = \sqrt{\varepsilon_r k_0^2 - \beta_m^2}$$
$$\overline{\Omega}_3 = \overline{\Omega}_1 = \sqrt{k_0^2 - \beta^2} \qquad \overline{\Omega}_2 = \sqrt{\varepsilon_r k_0^2 - \beta^2} \quad (3)$$

$A,B,C,D,$ are integration constants, and $\exp(j\omega t)$ time dependence is assumed. Normalizing these modes to carry the same

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The author is with the ITU Electrical Engineering Factory, Gumussuyu, Istanbul, Turkey.