and to P. Martin, K. Broome, and M. Appleby for their assistance.

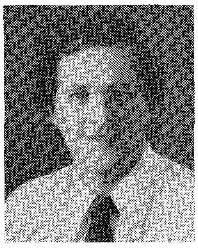
References

[1] J. J. Whalen, M. Thorn, and M. C. Calcatera, "Microwave nanosecond pulse burn out properties of GaAs MESFETS," in *MTT-S Int. Microwave Symp. Proc.*, Orlando, FL, May 1979.

[2] J. J. Whalen, M. C. Calcatera, and M. L. Thorn, "Microwave nanosecond pulse burn out properties of GaAs MESFETS," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 1026–1031, Dec. 1979.

[3] T. Suzuki, M. Otsubo, T. Ishii, and K. Shirahata, "Study on reliability of low noise GaAs MESFET's," in *Ninth European Microwave Conf. Proc.* pp. 331–337, Brighton, Sept. 1979.

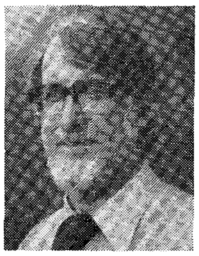
[4] J. Arnold, "Ruggedised GaAs FET amplifiers for military applications," in *Proc. Military Microwave Conf.*, London, pp. 651–656, Oct. 1980.



David S. James (M'71) was born in Bradford-on-Avon, England, on January 24, 1945. He recieved the B.Sc. and Ph.D. degrees in electonics engineering from the University College of North Wales, Bangor, U.K.

From 1970 to 1977, he was employed by the Department of Communications, Communications Research Centre, Ottawa, Ont., Canada. He is now with Ferranti Ltd., Manchester, U.K. His work involves the development of passive and solid-state microwave circuits, especially low-noise satellite subsystems.

Dr. James was Chairman of the Ottawa X-MTT Chapter and is a member of the A. V.S. and the IEE (UK).



Leslie Dormer was born in Manchester, England, on November 23, 1930. He was educated up to Higher National Certificate level in chemistry at Stockport College and John Dalton College, Manchester.

After five years in Textile Research, he became a Fuel Technologist working for Hawker Siddeley Dynamics from 1962–1967. In February 1967, he joined Ferranti Ltd. as a Chemist in the semiconductor facility, in which position he developed an interest in microwave parametric amplifier production. During the early years in the semiconductor facility he was awarded patents for metallization processes. In April 1973, he became Chief Chemist for Ferranti Electronics Ltd. and has since become actively involved in sponsored research into problems associated with semiconductor devices as used in solid-state microwave equipment.

Horn Image-Guide Leaky-Wave Antenna

TRANG N. TRINH, STUDENT MEMBER, IEEE, RAJ MITTRA, FELLOW, IEEE, AND ROY J. PALETA, JR.

***Abstract*—A novel structure for a frequency-scanning millimeter-wave antenna is described. The antenna is constructed by embedding a dielectric leaky-wave antenna in a long trough with metal flares attached along both sides. The optimum flare angle for achieving maximum gain is theoretically predicted. The design of the leaky-wave antenna, which is comprised of metallic-strip perturbations on top of the dielectric guide, is also discussed.**

Manuscript received April 13, 1981; revised August 6, 1981.

T. N. Trinh and R. Mittra are with the Department of Electrical Engineering, University of Illinois, Urbana, IL 61801.

R. J. Paleta, Jr., is with the Harris Corporation, T.S.T.O. 24/2455, Melbourne, FL 32901.

I. Introduction

RECENT TRENDS toward the use of low-cost, high-resolution antennas for short-range communications have encouraged the design and development of different millimeter-wave antennas [1]–[4]. The leaky-wave antenna

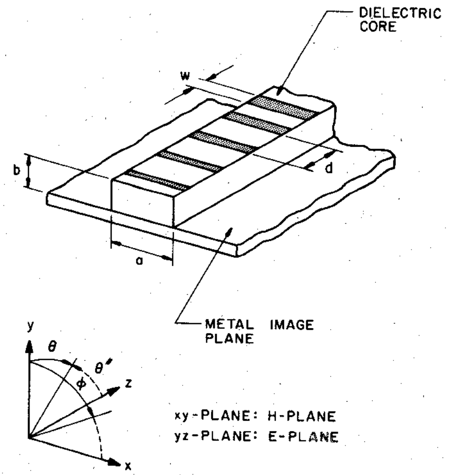


Fig. 1. Image-guide leaky-wave antenna and the coordinate system.

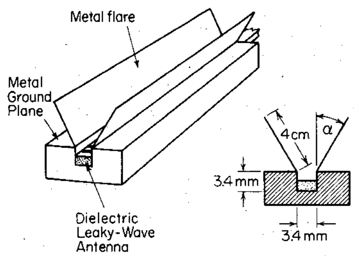


Fig. 2. Horn image-guide leaky-wave antenna.

is well-suited for these applications due to its relatively simple design, low cost, light weight, and compatibility with the dielectric-based millimeter-wave integrated circuits. Another major advantage of this structure is its ability to electronically scan the beam, simply by varying the operating frequency. Thus this type of antenna structure is capable of replacing many of the mechanically scanned, less-reliable gimbled structures which are too slow for many applications.

The image-guide leaky-wave antenna, like most traveling wave antennas, is long (in terms of wavelengths) in the longitudinal or the z-direction and has a cross section in the order of a wavelength (see Fig. 1). The dielectric guide with periodic perturbations on top is mounted on a metal ground plane. The purpose of this ground plane is to assure that all the power would be radiated from the top surface only. Without this ground plane, an equal amount of power would be radiated from both the top and the bottom of the dielectric guide. These antennas provide good radiation characteristics in the E-plane which is the plane containing the longitudinal axis of the antenna. However, the radiation patterns of these antennas are usually too wide for many applications in the plane transverse to the axis.

To reduce the beamwidth in the transverse plane, we propose a structure in which the dielectric guide with periodic perturbations on top is embedded in a rectangular trough with a metal flare attached along each side, as shown in Fig. 2. With this arrangement, the antenna behaves like a linear array in the longitudinal or (E-) plane, while the radiation pattern resembles that of a horn in the transverse (or H-) plane.

II. Design Criteria

A. Linear Array in the E-plane

In an image-guide leaky-wave antenna, the propagation energy in the guide is scattered by the metallic strips on top of the dielectric guide. These perturbations cause the guided-wave energy to radiate off the dielectric guiding structure or "leaked" into space. Each of the strips behaves like an element of a linear array. Hence, the key factors in the design of a leaky-wave antenna structure are the operating frequency, the perturbation spacing, and the guide wavelength. For periodic strip spacing, the angular direction θ_n of the leaky-wave beam is given by

$$\theta'_n = \sin^{-1}(\lambda_0/\lambda_g + n\lambda_0/d) \tag{1}$$

where λ_0 is the free-space wavelength, d is the perturbation spacing between the centers of two adjacent strips, and n is the index of the space harmonic $(0, \pm 1, \pm 2, \cdots)$ and is most likely to be -1, and λ_g is the guide wavelength. For slight perturbation, the real part of the longitudinal propagation constant of the structure has been approximated by that of the unperturbed structure. The antenna is frequency scannable if $|\lambda_0/\lambda_g + n\lambda_0/d| < 1$. For a broadside array, d should be made equal to the guided wavelength λ_g.

At each strip, it is desirable that the rate of energy radiated be small. Thus a portion of the guided-wave energy will propagate through the entire antenna length, which implies a larger effective aperture. But equally important, the guided energy should be completely radiated by these strips so that there is no energy left at the antenna truncation which would radiate endfire. Experiments show that the rate of energy radiated along the dielectric leaky-wave antenna structure is strongly influenced by the width of the perturbation strips on the top surface of the dielectric guide.

For very narrow strip widths, the radiation from each element is so small that a very long antenna has to be constructed to radiate all energy in a effective manner. Our experiments have shown that, for an array with periodically spaced elements, there is always a noticeable amount of residual energy radiated endfire even if 50 strips of width of less than 0.2 λ_g were used. This indicates that not all of the guided-wave energy has been radiated by the metal strips.

On the other hand, if the metal strips are too wide (>0.5 λ_g), the bulk of radiated energy is produced by the first few strips and, consequently, the effective aperture is very small. Also, for large strip widths, the sidelobes are very high, due probably to the large mismatch at the first strip. An experimental study of the gain characteristics of the antenna revealed that the highest gain was obtained when the strip width was approximately $0.4\lambda_g$ [5].

However, it is well known that if the attenuation of the traveling wave in the z-direction due to the energy leakage at the leaky-wave region of the antenna is a constant (dB/m), the radiation pattern is not symmetric about the main beam direction. Consequently, the energy transfer from the leaky beam to the surface wave, and vice versa, is not optimum [6]. A suitable modification of the geometry in the leaky-wave region of the antenna should be made to obtain a slowly varying leaky rate to increase the conversion efficiency [7]. If the variation of the strip width is correct, all the surface energy will convert completely into a symmetric leaky beam (and vice versa). This can be done by tapering the size of the periodic scatterers [6]. Experimental results show that the radiation characteristics in the E-plane are much improved when the strip width is continuously tapered. By linearly tapering the strip width (see Fig. 1), the rate of energy radiated into space is lowered. The effective aperture is enlarged and the beamwidth is narrowed. The larger strips at the end of the antenna radiate efficiently and assure that only negligible energy is left in the endfire (axial) direction. Since the mechanisms of perturbation and radiation are not clear, the widths of these metallic strips were experimentally investigated only. The width of the mth strip is found from the following empirical relation:

$$W_m = \begin{cases} (0.15+0.015(m-1))\lambda_g, & m \leq 18 \\ 0.4\lambda_g, & m > 18 \end{cases} \tag{2}$$

where λ_g is the guide wavelength. Near-field power distribution plots have shown that little or no radiation is produced by the first few strips, and their roles appear to be limited to providing an impedance match in the transition between the surface-wave and the leaky-wave regions.

The sidelobe levels for an antenna with tapered strips were much lower than that of a uniform strip width structure. However, if the tapering rate of the width of these metallic strips is slower than that given in (2), i.e., more than 18 small strips with tapering width were used, the gain and sidelobe characteristics of the antenna remain relatively unchanged. More exotic distributions for the strip width, e.g., the logarithmic or exponential, were not investigated in this study.

B. Flare Horn in the H-plane

It is well known that the overall gain of a conventional image-guide leaky-wave antenna structure is relatively low. This is due to the fact that the beamwidth in the transverse or H-plane is extremely wide; for instance, the null-to-null beamwidth of these antennas is almost 180°. This is understandable since the aperture in the transverse plane is on the order of a wavelength only.

To enlarge the aperture in the H-plane, we propose a new structure in which the dielectric guide with metal strip perturbations on top is embedded in a metal trough of rectangular cross section and with a metal flare attached to each side. The resulting configuration is shown in Fig. 2. The bottom of the trough assures that only a single beam is radiated from the antenna structure. The bottom plane redirects the energy which would have been radiated from the bottom surface of the dielectric guide back up and out the upper surface at exactly the same angle as the radiation from the upper surface [3].

The width of the trough was chosen such that it could easily interface with the truncated metal waveguide which was used to feed the antenna structure. The dielectric guide was designed to fit snugly against the sidewalls of the metal trough thus eliminating the need for adhesive materials which were used to bond the dielectric guide to the ground plane. The depth of the trough is such that the propagation constant of the structure can be determined using the dielectric-loaded trough guide model (see the Appendix). The height of the dielectric guide is then chosen to provide single-mode operation.

Since the radiation in the E-plane is fixed by the distributions of the silver strips, the overall gain of the antenna depends only on the radiation pattern in the H-plane and is thus a function of the flare angle and the length of the flared metal horn. With H_x as the major guided magnetic field component in the dielectric guide, the current element on each metal strip is z-oriented, and the principal aperture field of the metal flare is E_z. Assuming that the structure is infinitely long in the z-direction allows the metal flares to be modeled as an H-plane sectoral horn [8], [9]. For a conventional slotted metal waveguide, the use of longitudinal long metal flares has been known as a Slorn [10].

III. CONSTRUCTION

Under the guidelines and constraints described in Section II, a dielectric waveguide of dimensions 3.4×1.4 mm was constructed. The trough depth was arbitrarily chosen to be 3.4 mm to simplify the constructions. It was observed that as long as the trough was not too shallow, the radiation patterns of the structure were almost independent of the trough depth. At 81.5 GHz, the guide wavelength was calculated to be 2.70 mm ($\varepsilon_r = 2.47$).

The length of the flare horn was chosen to be 4 cm in this design. Fig. 3 shows the comparison between the measured relative gain of the constructed antenna as a function of the flare angle and that of an H-plane sectoral horn. The results of this figure have confirmed the assumption that the H-plane sectoral horn is a reasonable model to predict the flare angle for maximum gain in the H-plane.

For the strip width distribution chosen according to (2) and the flare angle selected from Fig. 3 for maximum gain, the radiation patterns in both the E- and H-planes of the horn dielectric guide leaky-wave antenna were measured and plotted in Fig. 4. The number of silver strips on top of the dielectric guide was 32. However, it was found that the length of the antenna was overdesigned since little or no power was detected after the 27th strip. The antenna can be truncated at this point with very little sacrifice of the overall gain. The half-power beamwidth (HPBW) in the E- and H-planes are 4°, 13°, respectively. If the attenuation of the guided wave due to radiation at each strip is small, the

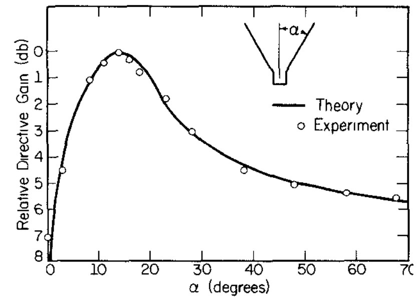


Fig. 3. Relative gains of the constructed antenna and an H-plane sectoral horn versus flare angle.

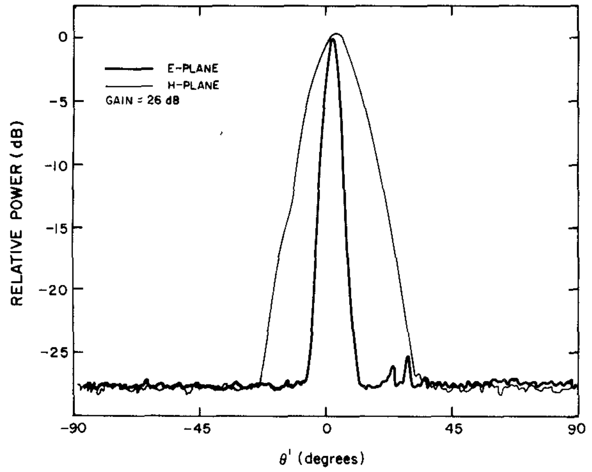


Fig. 4. E- and H-plane patterns of the constructed antenna.

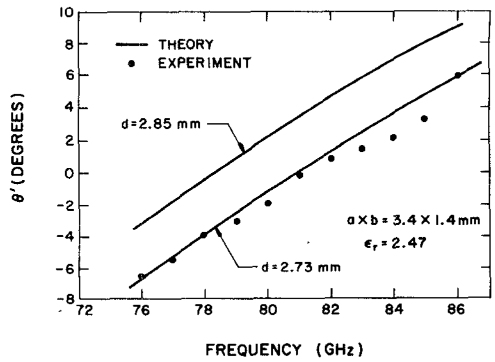


Fig. 5. Computed and measured main beam directions.

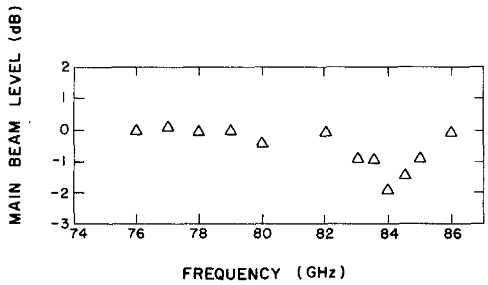


Fig. 6. Main beam level versus frequency.

radiation pattern in the E-plane can be approximated by that of a cophasally excited linear array [4]. For a broadside array, the HPBW is then approximated by

$$\text{HPBW} \simeq \frac{\lambda_0}{L} = \frac{\lambda_0}{d(m-1)} \tag{3}$$

where λ_0 is the free-space wavelength, L is the aperture length, d is the perturbation spacing between two adjacent strips, and m is the number of strips. For $m=26$, $d=2.73$ mm, and $f=81.5$ GHz, the HPBW in the E-plane calculated from (3) is 3°. This beamwidth is slightly narrower than the measured value. However, it should be noted that the first few strips in our antenna array are merely an impedance match, and contribute very little to the radiation.

The antenna was designed for broadside radiation. The consequent experimental beam angle deviated slightly from the normal direction. This is interesting since the strip width determines the rate of energy radiated off the guide but does not change the angle of radiation significantly. The $E-H$ tuner was employed in front of the feed end to ensure good matching. The return loss of the structure was measured to be about 15 dB. The overall gain of the antenna is the product of the gains in the E- and H-planes and was measured to be 26 dB. The sidelobe levels were at least 25 dB below the main lobe. Since there is no endfire radiation detected, it is assumed that most of the radiated energy is produced by the strip elements.

The angular beam direction was scanned by varying the operating frequency. Fig. 5 shows the measured and computed main beam directions as a function of the frequency. In our experiments, the frequency was varied from 76 to 86 GHz. The theoretical beam angle was calculated from (1). At each frequency, a new guide wavelength has to be determined (see the Appendix). As mentioned earlier, if the perturbation spacing is made equal to λ_g, the main beam will be at the broadside direction. The entire angular scanning range, however, can be shifted more positively or negatively by changing the perturbation spacing between the adjacent strips.

The main beam level of the antenna is relatively stable throughout the operating frequency range except at around 84 GHz where it falls about 2 dB below the reference level (Fig. 6).

IV. Conclusions

A horn dielectric guide leaky-wave antenna structure has been described. Both E- and H-plane beamwidths were much narrower than for a conventional leaky-wave design. Optimum designs of the perturbation strips on top of the dielectric waveguide and the flare angles of the horn have been determined. If the horn flare angle is not made prohibitively large, the antenna aperture can still be flush mounted by recessing the actual dielectric antenna into the supporting structure, such as the outer hull of an aircraft.

Appendix

Calculations of the Propagation Constants

It is advantageous to keep the dielectric waveguide in contact with the metallic sidewalls of the trough in order to ease the supporting problems and to eliminate the need for adhesive materials which were used to bond the dielectric

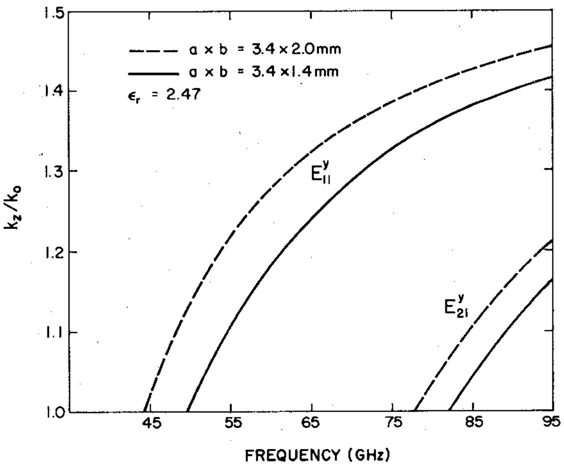


Fig. 7. Dispersion characteristics of the guiding structure.

guide to the ground plane. This study was primarily conducted using this configuration and, therefore, only the propagation constants for this structure are presented. When there is a gap between the sidewalls, the expressions for the propagation constants can be found elsewhere [3], [4].

For the E_{11}^y mode, the field's variation in the x-direction must satisfy the boundary conditions at the metal sidewalls resulting in

$$k_x = p\pi/a, \qquad p = 1, 2, \cdots \tag{A1}$$

where a is the width of the trough and $p = 1$ for the fundamental mode.

Matching the tangential fields at the dielectric–air interface yields the following eigenvalue equation for the k_y:

$$k_y b = q\pi/2 - \tan^{-1}(\eta k_y/\varepsilon_r), \qquad q = 1, 2, \cdots \tag{A2}$$

where

$$\eta = \left[(\varepsilon_r - 1)k_0^2 - k_y^2\right]^{-1/2}. \tag{A3}$$

η is the field decay coefficient outside of the dielectric, k_0 is the wavenumber of free space, b is the height of the dielectric guide, and $q = 1$ for the fundamental mode.

The longitudinal propagation constant is obtained as follows:

$$k_z = 2\pi/\lambda_g = \left(\varepsilon_r k_0^2 - k_x^2 - k_y^2\right)^{1/2} \tag{A4}$$

where λ_g is the guide wavelength.

The dispersion curve for this guiding structure is shown in Fig. 7. For a fixed trough width, the range of the single-mode operation can be varied by changing the height of the dielectric guide. The multisignal complexity of the antenna can be avoided by restricting the operating frequency to the single-mode region of the dispersion curve.

References

[1] Y. Shiau, "Dielectric rod antennas for millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 869–872, Nov. 1976.

[2] K. L. Klohn, H. Jacobs and E. Freibergs, "Silicon waveguide frequency scanning linear array antenna," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 764–773, Oct. 1978.

[3] R. E. Horn, H. Jacobs, E. Freibergs, and K. L. Klon, "Electronic modulated beam-steerable silicon waveguide array antenna," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 647–653 June 1980.

[4] T. Itoh and B. Adelseck, "Trapped image guide for millimeter-wave circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 1433–1436, Dec. 1980.

[5] S. Ray, Master's Thesis, Department of Electrical Engineering, University of Illinois at Urbana-Champaign, to be published.

[6] T. Tamir and H. L. Bertoni, "Unified theory of optical-beam couplers," in *Dig. Tech. Papers, Topical Meet. on Integrated Optics-Guided Waves, Materials and Devices*, (Las Vegas, NV, 1972).

[7] R. E. Collin and F. J. Zucker, *Antenna Theory*, vol. 2. New York: McGraw-Hill, 1969, ch. 19, 20.

[8] W. L. Weeks, *Antenna Engineering*. New York: McGraw-Hill, 1968, ch. 6.

[9] S. A. Schelkunoff and H. T. Friis, *Antennas: Theory and Practice*. New York: Wiley, 1952, ch. 16.

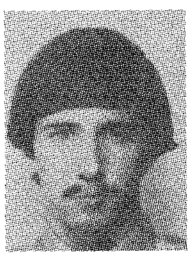
[10] C. H. Walter, *Traveling Wave Antennas*. New York: McGraw-Hill, 1965, ch. 8.

Trang N. Trinh (S'81) was born in Nha Trang, Vietnam, on November 28, 1956. He received the B.S. degree in electrical engineering from Wilkes College, Wilkes Barre, PA, in 1978, and the M.S. degree in electrical engineering from the University of Illinois, Urbana, IL, in 1980.

He is presently studying toward the Ph.D. degree at the University of Illinois, where he has been working on the millimeter-wave integrated circuits and components.

Mr. Trinh is a member of Sigma Xi.

Raj Mittra (S '54–M '57–SM '69–F '71) is Professor of Electrical Engineering and Associate Director of the Electromagnetics Laboratory at the University of Illinois, Urbana, IL. He is a Past-President of AP-S, is currently serving as an editor of IEEE Transactions on Antennas and Propagation and an active contributor to the many activities of the Society. He serves as a Consultant to several industrial and governmental organizations, including the NASA Jet Propulsion Laboratory of the California Institute of Technology. His professional interests include the areas of analytical and computerated electromagnetics, satellite antennas, integrated circuits, coherent optics, transient problems, radar scattering, and the like.



Roy J. Paleta, Jr., was born in Chicago, IL on April 6, 1957. He received the B.S. and M.S. degrees in electrical engineering from the University of Illinois, Champaign-Urbana, in 1980 and 1981, respectively.

Since graduation he has been employed by the Harris Corporation, Melbourne, FL, developing automatic test procedures for testing advanced satellite communications systems.