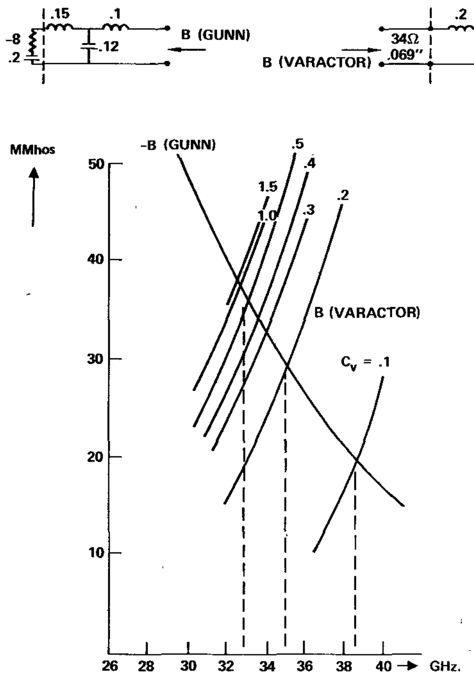

Fig. 3. Small-signal susceptance of Gunn diode and of ribbon leaded varactor diode with admittance inverter. Intersection of curves indicates theoretical resonance (with no additional components).

in the 38-GHz range. This was the same power that Varian obtained at 37.5 GHz using their standard waveguide mount. It was, therefore, apparent that experimentally a condition of approximate "best match" was obtained on microstrip. An equivalent circuit, neglecting discontinuities, is shown in Fig. 2. On the basis of this model, there are two frequencies within the 26–40-GHz range at which the impedance Z_x is purely resistive. At these frequencies (33.5 and 37.2 GHz) the circuit is capable of oscillating if the diode negative resistance can be made equal to the equivalent resistive load. It was possible to adjust the diode position and bias so that oscillations occurred at either 38 or 42 GHz, somewhat higher than predicted by the small-signal model.

In Fig. 3 a line is plotted showing the small-signal susceptance of the packaged Gunn diode. The package leads are responsible for a resonance somewhere below 26 GHz, above which the device appears inductive. In order to resonate the diode, it is necessary to provide a capacitive susceptance in parallel with the package. Varactor chip capacitors at millimeter-wave frequencies appear inductive when even the smallest bonding leads are used. To capacitively resonate the diode, an impedance inverter was used between the varactor bonding wire and the Gunn diode. The inverter consists of a quarter-wavelength distributed line, the impedance of which is optimized to give the largest frequency changes with change in varactor capacitance. In Fig. 3 the susceptance of varactor plus impedance inverter is plotted versus frequency for various values of varactor capacitance. The crossover points, where the susceptance is opposite and equal to that of the Gunn diode, represent the frequency of oscillation. This is true if the load appears resistive at the diode package when matched by the transformers. Varactor capacitance changes from 0.1 to 0.5 pF are within possibility representing a change of 5 GHz. More likely is a change between 0.2 and 0.5 pF, resulting in a 2-GHz change. Frequency-dependent matching components, such as the distributed transformers, may further decrease the tuning range.

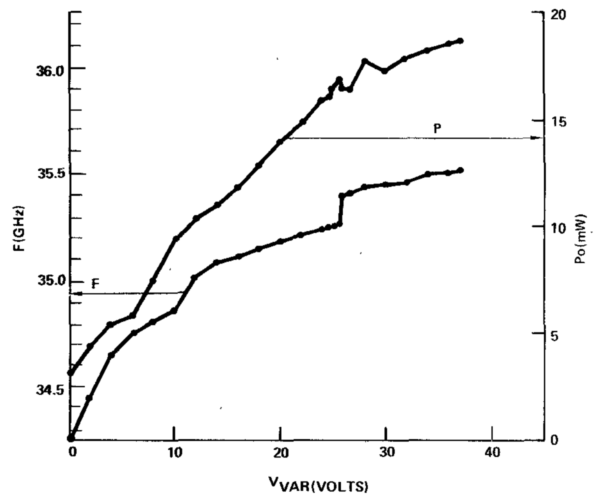

Fig. 4. Frequency and power output versus varactor reverse voltage.

Experimentally, as shown in Fig. 4, a 1.2-GHz tuning range was achieved near 35 GHz. With increasing bias voltage, the frequency increased. Since an increase in bias voltage on the varactor represents a decrease in capacitance, the admittance inverter functioned as described.

It should be apparent from Fig. 3 that if the external parasitic elements of the Gunn diode were reduced, the diode susceptance curve would tend to level, resulting in spreading of the crossover points and an increase in tuning range. This accounts for the fact that even at much lower frequencies wide-band solid-state oscillators almost all utilize unpackaged diodes.

IV. Conclusions

A millimeter-wave varactor-tuned Gunn oscillator was constructed on low-dielectric-constant microstrip. Output powers between 5 and 19 mW were obtained from a 40-mW diode over a 1-GHz bandwidth. It was shown how a 2–5-GHz bandwidth might be attainable using packaged diodes and how a further increase in tuning range would require diodes in chip form.

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A Low-Noise Millimeter MIC Mixer

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***Abstract*—This short paper describes a K_a-band integrated circuit mixer which, when integrated with a 2.5-dB-noise-figure IF amplifier, yields a 6-dB double-sideband noise figure over the band 36.5–38.5 GHz. A readily machinable low dielectric constant substrate has been utilized.**

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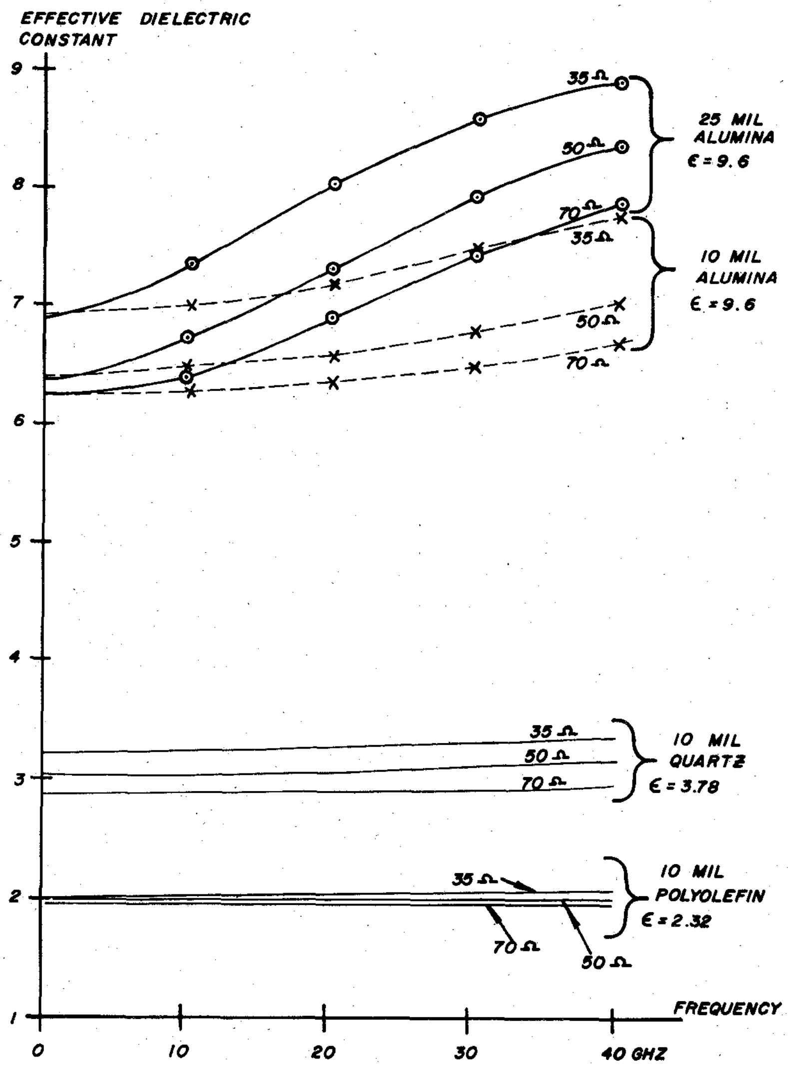

Fig. 1. Effective dielectric constant versus frequency for microstrip with impedance and substrate material as parameters.

A K_a-band balanced mixer has been successfully implemented in microstrip using irradiated polyolefin as the substrate material. The mixer, a balanced design utilizing a K_a-band "rat-race" signal/LO coupler, was fabricated on a $\frac{1}{2} \times \frac{1}{2}$-in 10-mil (0.010)-in-thick substrate utilizing high-quality beam-lead GaAs Schottky diodes, and was closely integrated with a discrete-component, 500–1000 MHz, 2.5-dB-noise-figure IF preamplifier. The resultant measured double-sideband noise figure for a LO setting of 37.5 GHz was 6 dB (this translates to a 5-dB double-sideband noise figure with a 1.5-dB-noise-figure IF amplifier), which compares quite favorably with any mixer available today. Moreover, the diodes used are relatively inexpensive, making this an attractive mixer where both cost and size are factors.

In choosing the optimum substrate material for K_a band, a number of factors must be considered. Alumina, the usual choice at X band and below, exhibits considerable dispersion [1] at K_a band, as shown in Fig. 1. Reducing the substrate thickness to reduce dispersion results in excessively narrow line-widths, thus impacting both line loss and reproducibility. Quartz, which has been applied to K_a-band designs with success [2], exhibits reasonable dispersion and yields practical linewidths. However, it is fragile and not suitable for all environments. Uniform low-loss low-dielectric-constant material has been shown to be an effective substrate material at K_a band [3]. The results reported herein demonstrate that it can be applied to the realization of low-noise front-end components.

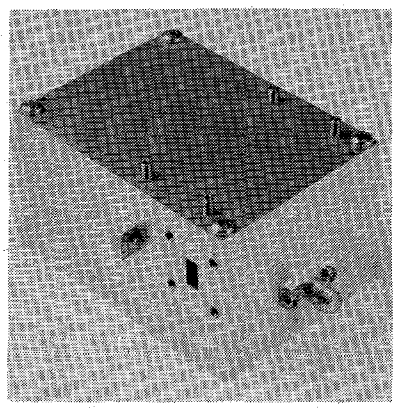

Fig. 2. Photograph of a mixer preamplifier utilizing a K_a-band MIC mixer.

The mixer substrate is mounted in a compact metallic housing, and the microstrip signal and LO ports are probe-coupled to corresponding waveguide-entry sections machined therein. Diode matching, for both the LO and signal, is accomplished with stubs that are printed along with the "rat-race," dc return, and IF low-pass filtering. Fig. 2. shows a mixer preamplifier utilizing 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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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].

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

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