$\sim$700 Hz/$\sqrt{\text{kHz}}$ at the IF output of a hyperbolic-sine-law mixer. Considering our probable error of 10–20 percent on each of those measurements, the data are consistent with the hypothesized doubling of local oscillator FM noise by the hyperbolic-sine-law mixer.

Appendix
Effect of Diode Imbalance on Mixer Noise

An expression relating diode imbalance to hyperbolic-sine-law mixer noise can be derived in a straightforward fashion. Assume that the diodes have identical saturation current but slightly different α parameters. The diode current is given by

$$I(v) = I_s(e^{\alpha v} - e^{-(\alpha+\delta\alpha)v}) \tag{A-1}$$

where $\delta\alpha/\alpha \ll 1$. If V_{LO} is much greater than the signal and noise voltages, we can obtain an expansion of (A-1) similar to (10). From this we find the components of the total current at dc (which is due to diode imbalance) and ω_{LO}, as well as the contributions of local-oscillator noise at $\omega_{LO} \pm \omega_{IF}$ and $2\omega_{LO} \pm \omega_{IF}$. Using the notation of Sections III and IV we have

$$\frac{I_{dc}}{I_{LO}} = \frac{\delta\alpha}{\alpha} \cdot \frac{\alpha V_{LO}}{8} \tag{A-2}$$

and

$$\frac{P_1}{P_2} = \left[\frac{(\delta\alpha/\alpha)\cdot\alpha V_{LO}\cdot I_0(\alpha V_{LO})}{4I_2(\alpha V_{LO})}\right]^2 \cdot \frac{n^2(\omega_{LO} \pm \omega_{IF})}{n^2(2\omega_{LO} \pm \omega_{IF})}. \tag{A-3}$$

Equations (A-2) and (A-3) can be combined and simplified by noting that $I_0(\alpha V_{LO}) \approx I_2(\alpha V_{LO})$, and assuming that $n^2(\omega)$ falls off roughly like $(\omega_{LO} - \omega)^{-2}$ away from the carrier. Then

$$\frac{P_1}{P_2} \simeq \left[2\frac{I_{dc}}{I_{LO}} \cdot \frac{\omega_{LO}}{\omega_{IF}}\right]^2. \tag{A-4}$$

Equation (A-4) shows that when $\omega_{IF} \ll \omega_{LO}$, extremely good diode balance is required to ensure $P_1 \ll P_2$.

A calculation assuming identical α parameters but slightly different values of the saturation current also yields the result given in (A-4).

Acknowledgment

The authors wish to thank W. W. Snell, Jr., who provided them with the low-noise 60-GHz hyperbolic-sine-law mixer, and also L. J. Greenstein, V. K. Prabhu, and H. E. Rowe for offering helpful insights and advice.

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The Ridged-Waveguide-Cavity Gunn Oscillator for Wide-Band Tuning

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***Abstract*—A method of construction of the Gunn oscillator for wide-band frequency tuning has been developed using the ridged-waveguide cavity. The ridged waveguide can be designed to provide the dominant mode with a wide bandwidth, and also to provide the higher order cutoff modes whose resonance is a limiting factor of the tuning range for oscillators of conventional design with reduced stored energies to permit wide-band tuning. A prototype oscillator having a packaged *X*-band diode demonstrated an 8–18-GHz tuning range.**

Introduction

In the course of investigating a transmission line that would be suitable for use with semiconductor devices at high microwave frequencies (from the *X*-band to the millimeter wavelengths where control of parasitics of the circuit and the package becomes all the more important for wide-band applications) the ridged-waveguide-cavity Gunn oscillator was developed. The use of the Gunn diode offers a convenient means of testing properties of the microwave circuit over a very wide frequency range because of its capability for wide-band operation [1], [2]. This short paper presents the structure and the performance characteristics of the ridged-waveguide-cavity Gunn oscillator operating at frequencies from 8 to 18 GHz.

Theory

Fig. 1 illustrates the oscillator structure where the ridges extend uniformly all the way to the shorting plunger. The field in the waveguide excited by the packaged diode may be represented by a sum of the normal modes of the ridged waveguide [3], of which only the dominant TE_{10} mode propagates while all other modes evanesce at frequencies of interest. The bandwidth of TE_{10} single-mode transmission as well as its characteristic impedance is controlled by the combination of a, b, w, and g, referring to the figure. The amplitudes of the evanescent modes are kept small by making the difference small between the cross section in the plane at the diode center and that of the waveguide. In addition, the diode placed at the center of the symmetric waveguide decouples from a certain class of asymmetric modes. The evanescent modes are made to evanesce quickly by choosing a and b small, thereby reducing the reactive energies associated with these modes, so long as they are consistent with the bandwidth and impedance requirements. The geometry of the ridged waveguide has the needed flexibility to meet these requirements.

Experiment

Several units of the oscillator were built and tested using packaged *X*-band Gunn diodes and ridged waveguides having various cross sectional shapes and dimensions. A typical tuning

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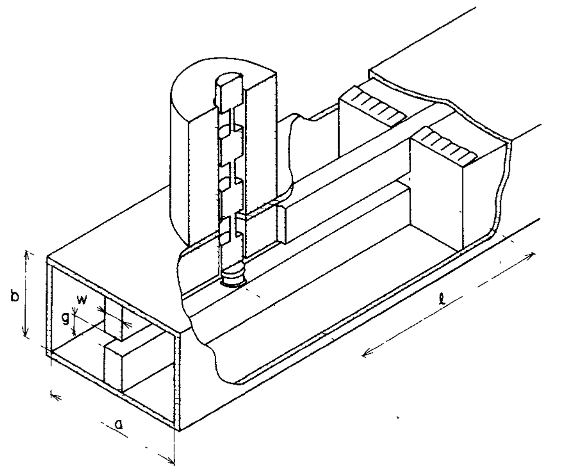

Fig. 1. Structure of the ridged-waveguide-cavity Gunn oscillator.

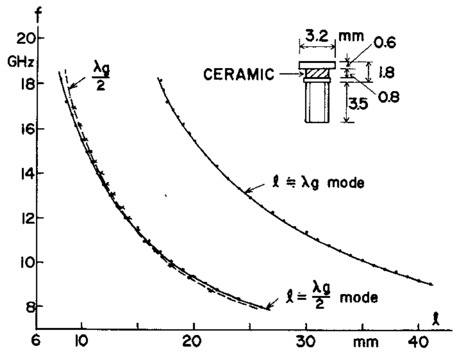

Fig. 2. Typical tuning curves of the ridged-waveguide-cavity Gunn oscillator. Waveguide dimensions are: $a = 15.8$ mm, $b = 7.9$ mm, $w = 3.4$ mm, and $g = 1.8$ mm. Package dimensions of the Gunn diode; GD-311C (Nippon Electric Co.) are shown in the inset. Ordinate: oscillation frequency given in gigahertz. Abscissa: distance between the diode center and the shorting plunger given in millimeters.

TABLE I

TE Hybrid Mode	: Cutoff Freq.	: $Y_{0\infty}$
TE_{10}	: 5.45 GHz	: 8.5 mmho
TE_{20}	: 21.75 GHz	: ——
TE_{30}	: 25.24 GHz	: ——
TE Trough Mode :		:
TE_{10}	: 19.18 GHz	: ——

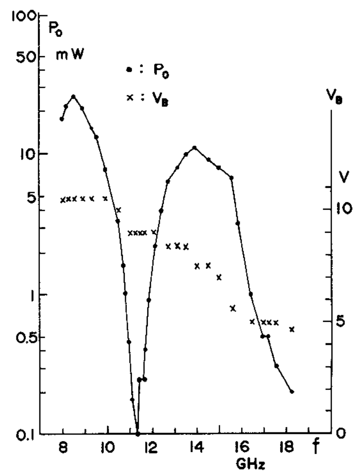

Fig. 3. A typical output power profile of the oscillator. Ordinate: output power given in milliwatts. Abscissa: oscillation frequency given in gigahertz.

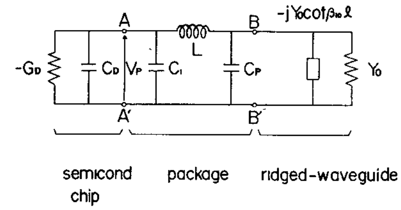

Fig. 4. An equivalent circuit of the ridged-waveguide-cavity Gunn oscillator.

curve obtained with the GD-311C diode (Nippon Electric Co.) mounted in a waveguide of $a = 15.8$ mm, $b = 7.9$ mm, $w = 3.4$ mm, and $g = 1.8$ mm is shown in Fig. 2 where l is the distance given in millimeters between the center of the diode and the short-circuit plunger, and the broken curve is the $\lambda_g/2$ reference curve. Absence of saturation from the tuning curve [4]–[7] indicates that the oscillation frequency was controlled through the TE_{10} mode over an 8–18-GHz range, demonstrating the validity of the method. The cutoff frequencies of the first four modes and the characteristic admittance of the dominant mode calculated on the computer are given in Table I. The output power profile corresponding to this tuning curve is presented in Fig. 3, where the dc bias voltage was also adjusted for the maximum output power at each frequency of measurement as indicated in the figure. Referring back to Fig. 2, both the $\lambda_g/2$ mode and the λ_g modes of oscillation were possible at l from 17 to 26 millimeter. Mode selection was made by the adjustment of dc bias voltage, as seen from Fig. 3. No uncontrolled mode jumpings were encountered in this frequency range.

It is possible to describe the oscillator performance by a simple equivalent circuit shown in Fig. 4 where the widely accepted representation is used for the packaged Gunn diode [6], [8]. Because the diode shunts directly across the waveguide, the series inductance due to TE_{n0} modes ($n = 2,3,4,\cdots$) is practically zero. Effects of all other cutoff modes can be accounted for by the package ceramic capacitance, C_p. Since the equivalent circuit involves only one series element, subsequent circuit calculations are simple. Writing the node equation at the AA' port of Fig. 4 and solving its imaginary part for $Y_0 \cot \beta_{10} l$, one obtains an expression for the tuning curve

$$Y_0 \cot \beta_{10} l = \omega C_p - (1/\omega L) \times \frac{2[1 - (\omega/\omega_s)^2] - 1 - \sqrt{1 - (2Y_0\omega L)^2[1 - (\omega/\omega_s)^2]^2}}{2[1 - (\omega/\omega_s)^2]} \tag{1}$$

where $\omega_s = [L(C_D + C_i)]^{-1/2}$ and Y_0 and β_{10} are standard symbols. Under the condition of (1), the load conductance at AA', G_A, is given by

$$G_A = \frac{4Y_0[1 - (\omega/\omega_s)^2]^2}{(1 + \sqrt{1 - (2Y_0\omega L)^2[1 - (\omega/\omega_s)^2]^2})^2 + (2Y_0\omega L)^2[1 - (\omega/\omega_s)^2]^2}. \tag{2}$$

The series resonance of L and $C_D + C_i$ requires an infinite susceptance at the diode terminals, BB', at $\omega = \omega_s$ and hence requires the tuning curve to intersect with the $\lambda_g/2$ reference curve and the output power profile to dip, as expected from and evidenced by (1) and (2) and Figs. 2 and 3, respectively. If desired, characterization of the packaged diode under actual operating conditions is possible from the measured tuning curve and output power profile with the aid of (1) and (2). Methods of obtaining the output power at ω_s are currently under investigation.

Conclusion

A new method of construction of the Gunn oscillator for wide-band frequency tuning has been developed using the ridged-waveguide cavity. The ridged waveguide can be designed to provide the dominant mode with a wide bandwidth and to provide the cutoff modes with reduced reactive energies to permit wide-band operation of the oscillator. An experimental oscillator was tunable continuously from 8 to 18 GHz, demonstrating the validity of the method. The approach presented in this paper will be applicable to designs of various broad-band microwave circuits involving small devices other than the Gunn diode.

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Design and Performance of a 60–90-GHz Broad-Band Mixer

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***Abstract*—This short paper describes a practical design and performance of 60–90-GHz broad-band mixers. Conversion loss was less than 11 dB in the 60–90-GHz region, and the conversion-loss deviation could be less than about 1 dB throughout the 30-GHz band with fixed circuit parameters. In this design, a new construction technique, using a wave absorber in place of a filter in the IF circuit, was employed.**

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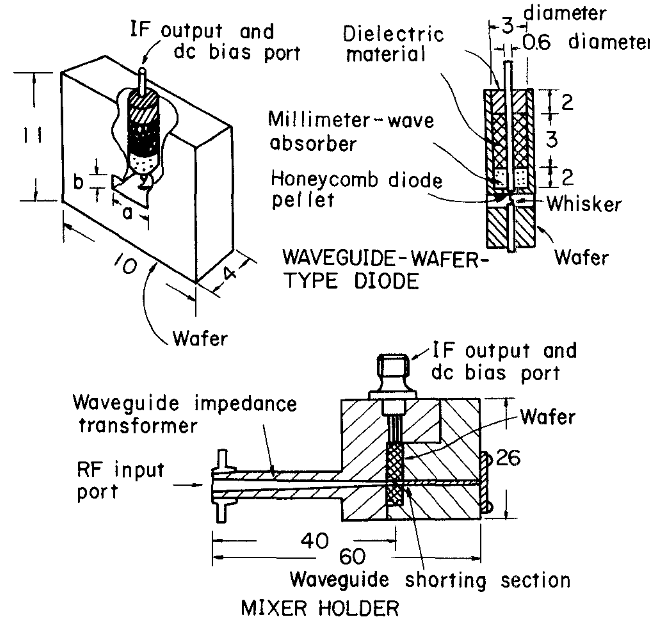

Fig. 1. Construction of a 60–90-GHz broad-band mixer. All measurements in millimeters.

I. Introduction

When referring to a broad-band mixer (receiving frequency converter), the following two cases are pertinent: 1) a mixer characterized by a broad IF bandwidth with a constant local-oscillator frequency; and 2) a mixer characterized by a broad RF bandwidth with a constant IF. The usefulness of the previous two mixer cases depends on the purposes of their applications. For example, the former is necessary for a millimeter-wave communications system, and for this application millimeter-wave mixers with a 1-GHz IF band have been reported [1]–[3]. For application to millimeter-wave measuring equipment, such as a selective levelmeter or a spectrum analyzer, however, the latter is important in order to cover the millimeter-wave band broadly and increase the sensitivity of the equipment by narrow-band signal receiving. In this case, mixers with a millimeter-wave band above 10 GHz are necessary. However, no investigations about them have been reported.

This short paper represents a design and performance of broad-band mixers characterized by a broad RF bandwidth for application to millimeter-wave measuring equipment. A broad-band mixer characterized by a broad IF bandwidth could also be obtained by the same method [4].

II. Design of a 60–90-GHz Broad-Band Mixer

A. Mixer Diode

In the millimeter-wave region, because of the large parasitic reactances, it is difficult to obtain broad-band characteristics by means of a pill-type diode package, which is generally used in the microwave region. For this reason, the "waveguide-wafer-type diode," as shown in Fig. 1, is widely used [1]–[3]. In this short paper, an image impedance-matched waveguide-wafer-type mixer using a GaAs Schottky-barrier diode with a junction diameter of 3 μm is presented. Computed values of the diode impedance (Z_d) for matching to the local oscillator were (49–j62) Ω at 60 GHz and (37–j49) Ω at 90 GHz (Table I) [5].

B. RF Circuit

A whisker 20 μm in diameter and 320 μm in length, was employed. The waveguide width a is 3.099 mm. Thus the equivalent inductance of the whisker is about 0.3 nH [6]. In this case,