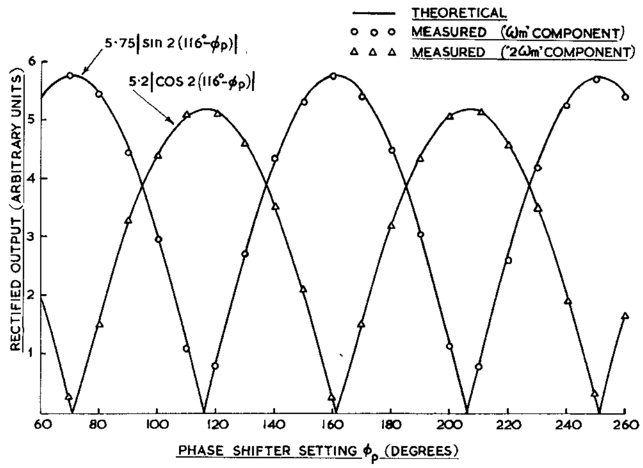

Fig. 2. Variations of the fundamental and second-harmonic components of the detected signal with phase-shifter setting.

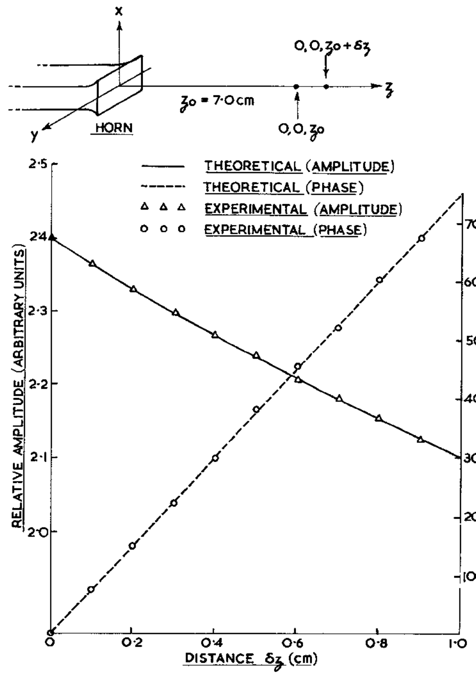

Fig. 3. Far-field amplitude and phase of a pyramidal horn.

A. Dependence of the Fundamental and Second-Harmonic Components of the Detected Signal on the Phase-Shifter Setting

Before measuring the field, a preliminary experiment was performed to verify that the rectified outputs of the two selective amplifiers depend on the phase-shifter setting according to (3) and (4). These equations form the basis of the simultaneous determination of amplitude and phase. In this test, the system was operated manually and the dipole was located in the far field of the horn.

The observations are presented in Fig. 2. It is particularly important to note that the maxima of each component are equal, because this is an indication that: 1) the differential amplifier was properly adjusted, and 2) the antenna was correctly matched to the phase shifter. Subsequent field measurements were therefore free from errors due to the back-scattered unmodulated signal [9].

B. Measurements of Amplitude and Phase

The far field of the horn was explored by measuring both the amplitude and phase of the x component of the electric field. The measurements were made at discrete points along the z axis, using the automatic system. The results of the measurements are shown in Fig. 3. The theoretical curves are those for a distant point-source located in the throat of the horn.

VII. Conclusions

An automatic system which simultaneously measures amplitude and phase distributions of millimeter-wave fields has been described. The principle of the system is that of the vibrating-dipole technique. The high accuracy of the system has been demonstrated by measurements in a far field of known configuration.

The simultaneous measurement of amplitude and phase in a homodyne detection system which uses only one hybrid junction is possible because the wave reradiated by the vibrating dipole is phase modulated. The amplitudes of the fundamental and second-harmonic components of the detected signal vary as the sine and cosine, respectively, of the same phase angle. All the other modulated-scattering techniques employ double-sideband amplitude modulation with carrier (DSBWC) of the scattered wave [8]. The simultaneous measurement of amplitude and phase in a homodyne system with DSBWC is possible [10], but requires two hybrid junctions and also a 3-dB coupler.

To take full advantage of the automatic system the phase shifter employed should have a sufficiently large differential phase-shift. Our phase shifter was of the dielectric-vane type, with maximum differential phase-shift of only 180°. Therefore automatic field measurements were restricted to such regions in space in which phase changes were within 180°. To continue measurements beyond this region it was necessary to switch off the servo system and to manually adjust the phase shifter to a new convenient setting.

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High-Frequency Gunn Oscillators

THOMAS G. RUTTAN

***Abstract*—Recent work to achieve high output power of "fundamental mode" Gunn-effect oscillators at frequencies ranging from 25 to 71 GHz is described. Ambient powers of 370 mW at 6.7-percent efficiency at 25 GHz, 260 mW at 4.5-percent efficiency at 38 GHz 150 mW at 4-percent efficiency at 54 GHz, and 30 mW at 1-percent efficiency at 71 GHz were obtained from single-diode structures. Combining two diodes in a push–pull circuit yielded 400 mW at 3.5-percent efficiency at 32 GHz and 260 mW at 4.0-percent efficiency at 42 GHz. This represents some of the highest powers and efficiencies reported to date from millimeter-wave Gunn-effect oscillators.**

I. Introduction

Millimeter-wave Gunn-effect oscillators have been in existence since the early days of Gunn-effect work [1]–[5], but have only been capable of low output power for some local oscillator applica-

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tions or laboratory studies. These Gunn-effect devices were not able to compete with IMPATT's or other millimeter-wave devices for such higher power applications as parametric amplifier pumps or transmitters. Over the last several years work has been done to develop high-frequency (25–71 GHz) Gunn oscillators with high output power. This paper describes the devices and circuits used in this work and presents the results of this effort.

II. Devices

The Gunn devices used in this work are made from Varian solution grown liquid phase n-type epitaxial GaAs. The active layer is typically 2.0–2.5 μm thick for devices used at 30 GHz and above, and 3–3.5 μm for devices used from 25–30 GHz. The active layer is grown on a tin-doped GaAs substrate. Typical doping levels were 4.0–4.5 $\times$ 10^{15} carriers/cm^3 for low *Ka*-band devices, and ranging to 5.5–6.0 $\times$ 10^{15} carriers/cm^3 for 40 GHz and above. Recently, we have tried a wafer-doped 6.5–7.0 $\times$ 10^{15} carriers/cm^2 that works rather well at 65–70 GHz, giving about 1-percent efficiency.

The cathode and anode ohmic contacts used on these devices consist of a gold–germanium nickel metal layer alloyed onto the epitaxial and substrate sides. Most of the devices used have the cathode ultrasonically compression bonded to the diode-package heat-sink stud. Recently, we have tested some devices with a 1–2-mil gold plating on the cathode side that is then soldered to the diode-package stud. This method is very promising for obtaining better heat sinking of the device as well as eliminating bonding damage to the chip, thus improving power output and efficiency of the devices. An n+ epitaxial regrown cathode contact has been tried at these frequencies, but we found much lower efficiency and power output than with direct metallization contacts, as discussed by Caldwell and Rosztoczy [6].

The diode package used here is a standard commercially available Varian N34 package, consisting of a gold-plated copper heat-sink stud on a 3-48 screw with a 0.115-in diam flange. This package has worked quite well throughout much of the millimeter frequency range of interest, and has also been discussed previously by Fank and Day [4].

III. Circuit Description

Several circuits are utilized at these high frequencies, depending upon the requirements and application of the oscillator. One of these is a single Gunn-diode waveguide circuit consisting of a half-wave resonant cavity with the Gunn diode placed at one end of the cavity. The dc bias for the Gunn diode is obtained from an inductive post in contact with the diode and RF isolation by standard choking techniques. The output power is inductively coupled through an iris, and the frequency is tunable by means of a dielectric rod inserted into the cavity to perturb the electric fields.

A double Gunn-diode push–pull waveguide circuit has also been developed that allows one to add the powers of the individual diodes. Proper coupling in this push–pull mode makes it possible to obtain slightly more than the combined maximum power that each diode is capable of in a single-diode circuit. One also has the convenience of parallel dc device operation and series RF operation. The physical configuration of this circuit has been discussed by the author previously [7].

The coaxial line circuit is used at millimeter-wave frequencies up to about 50 GHz. It is a one-half wavelength resonant structure with the Gunn device placed at a low impedance point at one end of the cavity. The center conductor rests on the Gunn diode and brings in the dc bias. RF isolation for the bias network is obtained with a choke similar to those used in the waveguide circuits discussed previously. The power is magnetically coupled through an iris slot at one end of the cavity to a waveguide. If this cavity is mechanically tuned, a dielectric rod is inserted through a hole in the center of the cavity where the electric fields are maximum.

A radial line circuit has recently been used for 60–70-GHz oscillators. A radial line circuit was chosen because the common Gunn-diode circuits such as waveguide and coaxial resonators become too small to work with and difficult to optimize at frequencies above 60 GHz. In this circuit the radial resonator rests on the Gunn diode and is placed in a section of standard waveguide with a sliding short for optimization. The resonant frequency is determined by the diameter of the radial resonator and the coupling by the gap from the resonator to the waveguide wall. Other parameters of this

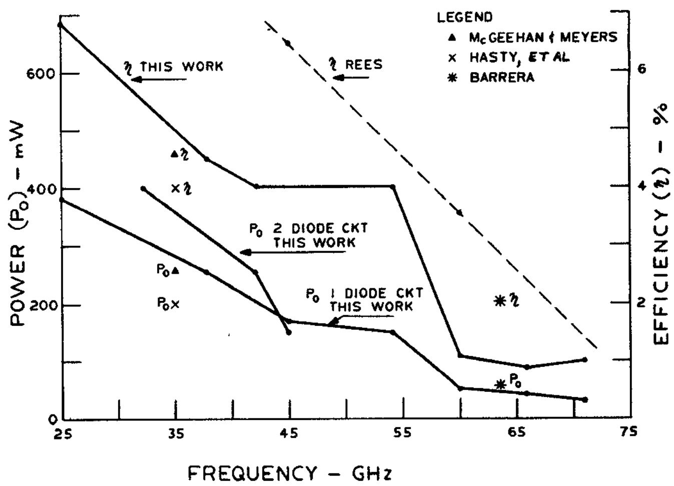

Fig. 1. Summary of power and efficiency versus frequency.

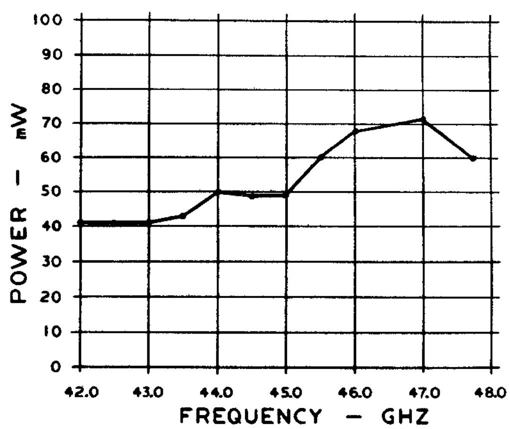

Fig. 2. Broad-band tuning curve of a waveguide Gunn oscillator.

circuit have been discussed, employing an IMPATT diode, by Groves and Lewis [8]. This circuit is very promising for high-frequency applications due to its simplicity and performance. The external Q is low (typically about 100 for a tightly coupled circuit), however, in comparison with a waveguide circuit. All of the power levels reported in this work at and above 65 GHz have been obtained in this circuit.

IV. Results

The work reported in this paper is primarily concerned with achieving high- or maximum-power output of Gunn oscillators at millimeter frequencies with either narrow or broad-band tuning. Consequently, most of the results are with near critically coupled circuits.

A summary of our best power and efficiencies to date appears in Fig. 1. The highest efficiency at any point on this graph may not correspond to the highest power at that point, although many did. These results were obtained using the various circuits described earlier. This figure also compares the results of this work with the empirical results of Hasty [9], McGeehan [10], and the LSA work of Barrera [11], and the LSA efficiency calculation of Rees [12]. The efficiencies of this work compare favorably with Rees' calculations in the mid-50-GHz range and near 70 GHz while falling somewhat low at the other frequencies of interest. It can be seen from Fig. 1 that the efficiency of this work decreases much slower with increasing frequency above 60 GHz, thus indicating that the high-frequency limit (where $\eta = 0$) is probably higher than that calculated by Rees.

Figs. 2 and 3 give some idea of the broad-band capabilities of these devices. Fig. 2 is a tuning curve of a single-diode waveguide oscillator decoupled approximately 2–3 dB. This circuit had the tuning rod inserted from the cavity wall opposite the coupling iris. The

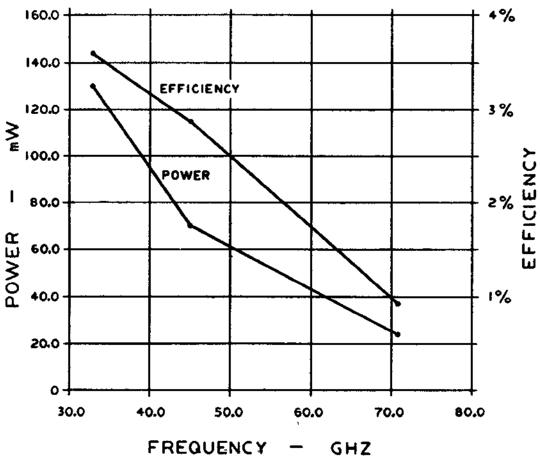

Fig. 3. Broad-band performance of a single Gunn diode.

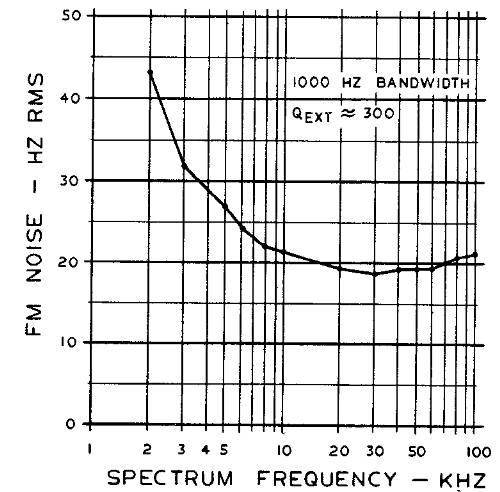

Fig. 4. FM noise of a waveguide Gunn oscillator.

data in Fig. 3, taken in three circuits of different frequencies, show the broad-band capabilities of our high-frequency Gunn diodes. A sample of the diodes used in this work has exhibited RF power over greater than an octave frequency bandwidth.

Fig. 4 displays FM noise measurement (in a 1000-Hz bandwidth) of a near critically coupled waveguide circuit at 28.24 GHz with 185-mW power output. The diode was a metallic-contact mesa diode and was running at about 3.5-percent efficiency. The FM noise measured on 35-GHz oscillators of approximately the same Q_{ext} and output power have yielded nearly identical data to that in Fig. 3. As of this writing no FM noise measurement has been made above this frequency, but based on the fact that the measured Q_{ext} of higher frequency (e.g., 40–50 GHz) near critically coupled oscillators are nearly identical to the oscillator of Fig. 4, one would expect the FM noise to be approximately the same as the data in Fig. 4. The FM noise of the double-diode push–pull circuit should be close to its single-diode waveguide counterpart, since the Q_{ext} are similar.

Acknowledgment

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Application of Quarter-Wave Transformers for Precise Measurement of Complex Microwave Conductivity of Semiconductors

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***Abstract*—The utility of a quarter-wave transformer for precise measurement of complex microwave conductivity of semiconductors has been demonstrated. It has been shown for a chosen conductivity of 9 Ω·cm that the improvement in measurement accuracy is nearly by a factor of 3 over the conventional reflection measurement using a Teflon transformer.**

I. Introduction

Wide applications of semiconductors in microwave solid-state devices call for an accurate determination of the complex microwave conductivity of these materials. Of the various methods so far used to this end, the conventional technique of the measurement of complex transmission or reflection coefficient of a waveguide completely filled with the sample plays a significant role. The accuracy attainable through these methods using commercially available precision standards for attenuation and phase shift has been estimated by Datta and Nag [1]. It has been shown that, because of lack of accuracy in commercial standards, the potential accuracy of these methods is rather poor, especially in reflection-type measurement where one has to measure a high VSWR (nearly 20 dB) and a small phase angle (nearly 10°). An improvement in the accuracy of the experimental results is expected when reflection measurements are taken with a quarter-wave transformer inserted between the sample-filled section and the empty guide, since with this arrangement the phase of the reflection coefficient in the input guide increases with a simultaneous decrease of its magnitude. This short paper presents the experimental result of complex microwave conductivity in X band at room temperature on 9 Ω·cm p-type silicon obtained through this modified technique and describes, with reference to these data, the utility of a quarter-wave transformer in semiconductor parameter diagnosis.

II. Theory

For the two configurations shown in Fig. 1, the propagation constant Γ_s for the sample-filled section may be obtained solving the following transcendental relations, respectively,

$$m_1 \angle -\phi_1 = \frac{1 + \rho_1 \angle -\theta_1}{1 - \rho_1 \angle -\theta_1} = \frac{\Gamma_1}{\Gamma_s} \tanh \Gamma_s l \tag{1}$$

$$m_2 \angle -\phi = \frac{1 + \rho_2 \angle -\theta}{1 - \rho_2 \angle -\theta} = \frac{\Gamma_1}{\Gamma_d} \frac{(\Gamma_d/\Gamma_s) \tanh \Gamma_s l + Z}{1 + Z(\Gamma_d/\Gamma_s) \tanh \Gamma_s l} \tag{2}$$

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