Millimeter-Wave Pretuned Modules

GÉRARD CACHIER, JACQUES ESPAIGNOL, AND JEAN STÉVANCE

***Abstract*—This paper presents pretuned modules for millimeter-wave devices. A general description of the device, some new developments of the technology, and the different techniques used for its characterization are covered. It is shown how the bias-to-waveguide coupling is made, and how these techniques lead to an oscillator which radiates directly into free space.**

I. Introduction

MORE THAN 20 years ago, the first backward-wave oscillators appeared at millimeter and submillimeter frequencies [1]. More recently, the solid-state devices —Si IMPATT's, GaAs TED's, and even InP TED's— made possible small-size oscillators delivering moderate amounts of power. Although significant progress has been realized in the past few years on the performance of these solid-state sources, it is a fact that many users do not seem to consider yet the millimeter waves as being a practical way to meet their needs. One of the reasons is the relatively high cost of these devices.

In order to overcome this problem, several proposals emerged a few years ago for developing low-cost circuit techniques at millimeter frequencies. These can be classified into three different and unequal classes: the stripline, the fin line, and the dielectric waveguide [2]. All these techniques have specific advantages due to the integration possibilities they offer. However, we will present here a different approach; we are concerned rather with the possibility of using solid-state devices (IMPATT sources) in millimeter circuits. Therefore, we concentrated on the circuit at the level of the chip, where everything is most important and most difficult to make. We designed the necessary circuitry to be held in one module. With respect to that problem, interconnecting the different devices seems rather easy, as will be shown in Section V.

We will first review the basic principles of the module and its design. A detailed description of the technology will then be given, as well as a method of characterizing directly at millimeter frequencies, which has proven to be very useful. We will present a simple mount with a bias feed and show how to use it to make an oscillator which radiates directly into free space.

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II. The Pretuned Module

A. General Description

The pretuned module (PTM) [3] is a rugged oscillator with both internal matching and radiating elements. Its equivalent circuit (Fig. 1) is designed to

1) match the diode reactance ($X_c + X_d = 0$);
2) prematch the diode resistance ($R_c \sim 1\ \Omega$);
3) help couple the RF energy to remote utilizations (and optimize the external tuning range for the module).

The third point requires the circuit to have a transverse dimension comparable with the wavelength [4], in order for the radiation conductance to have a significant value. This turns out to be an advantage for the PTM since one of the objectives at millimeter frequencies is to be able to make the circuits as big as possible. Theoretically, the best coupling element to the small diode is the biconical antenna [4], but the technological constraints make a flat structure preferable. The radial waveguide [5] has, therefore, been chosen as a more appropriate structure (Fig. 2). A similar radial waveguide is used in the well established "cap" circuit [6], [7]. We will use a radial waveguide slightly longer than $\lambda/4$, which will yield both the desired reactance ($\sim 10\ \Omega$), and a low resistance ($\sim 1\ \Omega$).

B. Theory [1]

When the negative resistance diode oscillates, we have

$$Y_d + Y_c = 0$$

where the diode admittance Y_d and the circuit admittance Y_c are taken at the chip terminals. For a given diode, the frequency of oscillation and the output power are governed by the complex Y_c and the circuit efficiency η. We will see that both are mostly determined by the geometry and the materials located near the chip.

To compute Y_c, we divide the electromagnetic system into two separate regions [8]: inside and outside the radial waveguide. The region outside is treated as a radiation problem [9]. The result of this calculation is introduced as a boundary condition for the radial waveguide by a terminal admittance Y_T equal to the radiation admittance. On the other hand, the radial waveguide (height b), working

[1] In this paragraph, we will use the admittance instead of the impedance as it is suggested by the radiation calculations of the disk antenna.

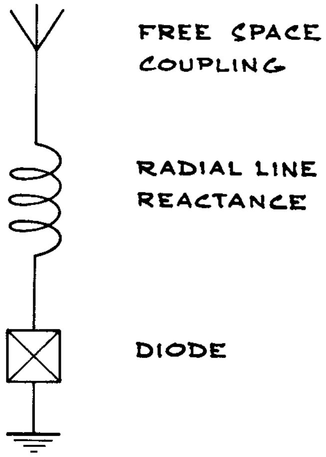

Fig. 1. Equivalent circuit for the pretuned module.

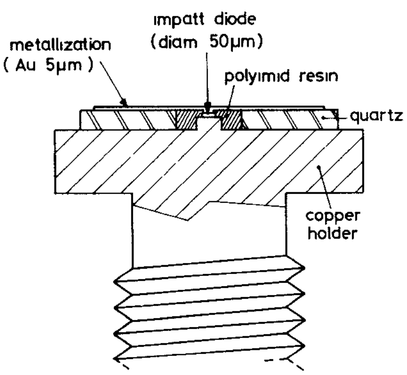

Fig. 2. Schematic cross section of the PTM, emphasizing the position and the relative dimensions of the IMPATT diode and the dielectric supported radial waveguide.

in its fundamental e-type mode, is described by introducing a radius (r) dependant characteristic admittance Y_0 [4]

$$Y_0=\sqrt{\frac{\epsilon}{\mu}}\ \frac{2\pi r}{b}.$$

Finally, the whole circuit (inside and outside the cap) can be calculated, yielding the theoretical value of Y_c.

This calculation is very useful for predicting the performance of the oscillators and understanding how the circuit works [8]. Let us here review some of the results.

1) The most significant part of the circuit susceptance results from the propagation inside in the open-ended radial waveguide. The calculation involves the height b and the phase constant in the cap taken radially. It will be used to predict the frequency of oscillation with a good accuracy.

2) This radial waveguide operates also as a transformer between a change in the terminal admittance δY_T and the resulting change in the circuit admittance δY_c.

3) The terminal admittance Y_T is similar for a module radiating directly into the free space and for a module radiating in a standard rectangular waveguide matched at both its ends.

4) As an experimental proof, we introduce the module in a rectangular waveguide as will be explained later (Fig. 4) and measure the admittance as a function of the position of the back short at a fixed frequency (see Section IV). We observe a "pretuned" characteristic (Fig. 8): the susceptance is predetermined with good accuracy, and the conductance is adjusted by the short within the desired area. The different points lie on a circle (congruent transformation).

5) The circuit losses, reducing the overall efficiency, are mostly of ohmic origin, and are located in the vicinity of the diode.

6) The design of the module depends on the value of the dielectric constant of the material supporting the radial line. It turns out from the propagation equations that the scaling has to be done by multiplying the phase constant in the radial direction by $\sqrt{\epsilon}$, and by keeping the height constant. There is, therefore, a reduction of the diameter of the cap for the higher ϵ. This in turn affects the terminal admittance Y_T by a significant reduction of the conductance G_T.

III. Technology

Typical dimensions for designing a module are a height of 0.03λ (λ = free-space wavelength), and a diameter of 0.3λ for fused quartz as a constituting material ($\epsilon=4$). We can, therefore, see that the module has a low profile and a large metallized top surface. This will ensure both mechanical ruggedness and easy handling.

The technology used to realize the module makes it very different from a conventional package: the process resembles that of a monolithic integrated circuit, although the device is not monolithic in a true sense. The main reason for this difference is the necessary mesa passivation, i.e., the imbedding of the diode in a massive dielectric environment. This passivation should not deteriorate the dc characteristics (leakage current, voltage handling). The passivating material has to stand elevated temperatures (300°C or more) also, and it should not be too lossy at millimeter frequencies. We have chosen a polyimid resin (Rhodeftal 200) which has all the desired characteristics ($\tan\delta<10^{-2}$). The mesa passivation entails a problem with the top contact, since the diode is usually completely immersed in the dielectric material. There are several ways to turn around this problem. One of them is the thermocompression bonding of a gold ball on the top of the chip.

Fig. 3 describes the different steps involved in the processing, using a p^+-p-n-n^+ silicon epitaxial wafer as a starting material. First is the thermocompression bonding of the silicon chip to the heat sink, with the epilayers down for best power dissipation. Next is the bonding of a small (200-μm) gold ball on the top of the chip, which will be used for the back contact to the diode. In addition to the simplification of the back contact, this process has the advantage of making it easier and more flexible to lap the quartz (see below). The silicon chip is then trim etched to realize the diode. After that, the polyimid resin is used to passivate the diode between the gold ball and the thermal heat sink.

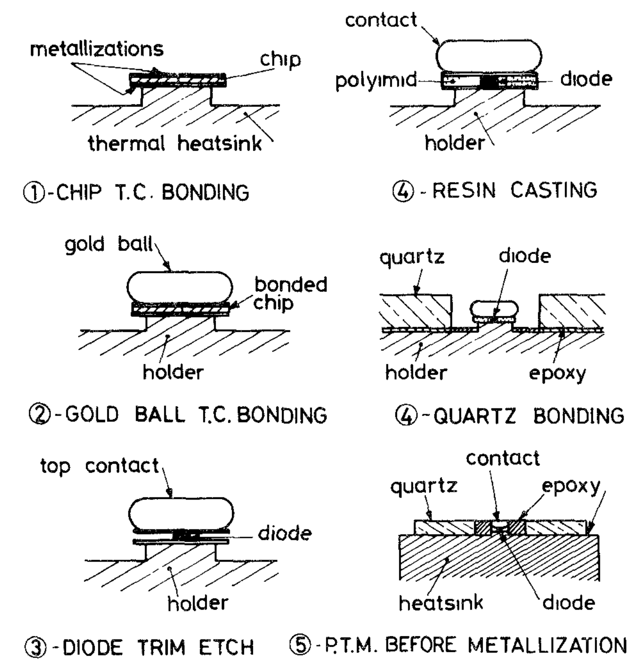

Fig. 3. Main processing steps for the realization of individual modules.

Next, the circuit is realized out of a quartz ring made by truncating a capillary tube. Typical outer diameter of the ring is 2.5 mm, and inner diameter of the order of 500 μm. This quartz ring is bonded to the thermal holder by a thin ($\sim$15-μm) layer of high-temperature epoxy. After that, the remaining hole is filled up with epoxy. Finally, the whole structure is lapped to make a planar structure. Then the top surface is metallized (Ti–Pt–Au or Cr–Pt–Au) in order to make a radial waveguide around the diode.

IV. Measurements and Characterization

A. Performance

For easy characterization, the module is inserted horizontally in a standard rectangular waveguide as indicated on Fig. 4. The large metallized top surface allows easy bias with a thin wire which crosses the waveguide walls through alumina rings. This wire is used to control the insertion of the module in the waveguide, so that the surface of the thermal holder lies in the plane of the bottom wall.

The output power is measured on each module. It is always in the 100-mW range, with a best result higher than 400 mW at a frequency of the order of 45 GHz. The most critical parameter for the power is the diode itself, with important factors as the epitaxial material and the quality of the metallizations. In order to know the influence of the dimensions of the module, we used the technique described previously to realize successively with the same diode different modules of increasing diameters and decreasing heights. The results of Fig. 5 show that some dependance on the dimensions of the circuit does exist, but is not extremely critical. The frequencies were between 40 and 50 GHz.

The frequency of oscillation is an important parameter to be controlled for practical use. Fig. 6 shows that the

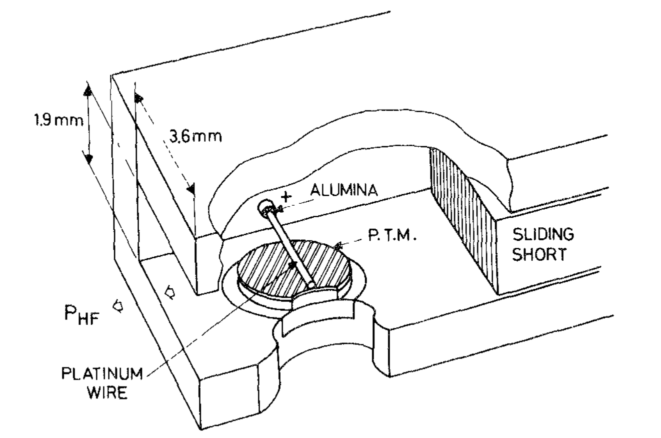

Fig. 4. PTM in a standard waveguide mount. Bias is taken with a thin wire crossing the waveguide walls through alumina rings.

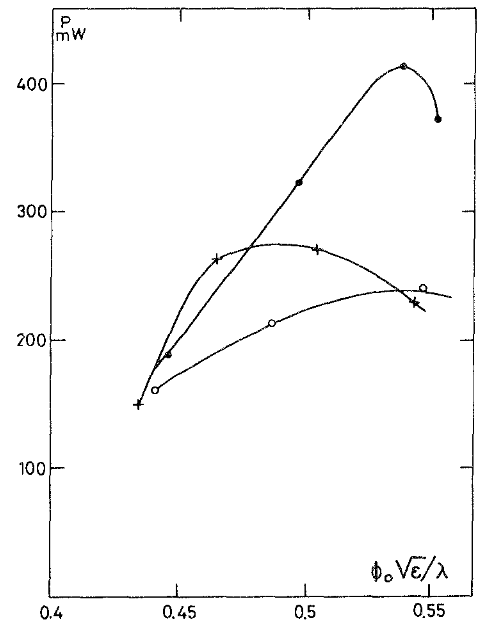

Fig. 5. Power delivered by several individual modules made with the same diode, by successively increasing the diameter Φ_0 of the radial waveguide. There is a smooth optimum for a diameter slightly higher than 0.5λ.

behavior of the module as a function of the position of the backshort is sound, and one can, therefore, expect a good predictability of the frequency. On the other hand, it is possible to use the frequency of oscillation to recompute the susceptance of the circuit and, hence, to check the accuracy of the theoretical model [8]. This is done by assuming that the susceptance of the diode is close to its susceptance just before the breakdown, which can be extrapolated from low-frequency (1-MHz) capacitance measurements.

The comparison between the theory and the measurements has been made on more than 20 oscillators (those used for power measurements). In order to facilitate the calculation, the theoretical model involving Bessel functions has been approximated by an analytical formula. This comparison has shown a frequency error of less than 6 percent. This discrepancy is not random, but looks

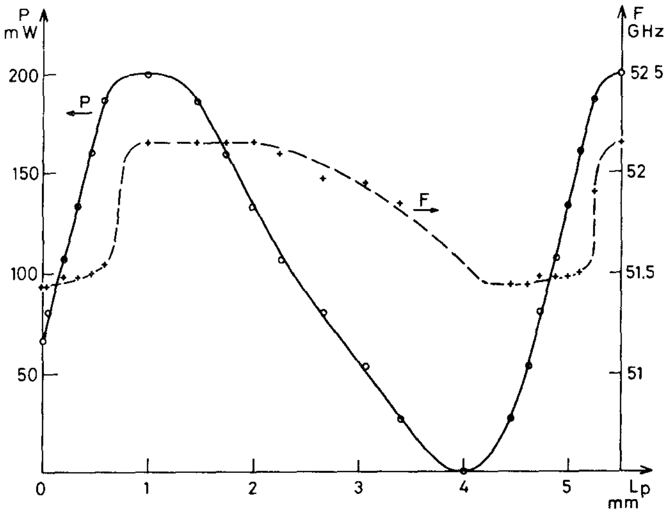

Fig. 6. Backshort tuning of the PTM. The behavior is smooth, and repeats itself after half-guide wavelength.

rather like a sum of systematic deviations related to one step or another in the process. In particular, those related to the diode itself look most important. Also of importance, is the gold ball bonded on the top of the diode, which brings a capacitance in parallel with the diode ($C_p \sim 0.1$ pF), and lowers the frequency of oscillation by about 10 percent.

B. Characterization

In order to get more information on the circuit, we use a separate method to measure directly the impedance Z_c and the efficiency η of the circuit. In this method [10], the diode is biased below the breakdown—as a varactor diode.

The capacitance C of the diode is known from low-frequency measurements, and the susceptance of the diode in the millimeter range can be reasonably approximated by $C\omega$. It can be tuned simply by changing the bias voltage. The diode is used, therefore, as a microwave standard for the imbedding impedance.

As a new feature in this method, the millimeter signal used for device testing is amplitude modulated at a frequency of 700 kHz. We monitor the IF signal demodulated by the diode. Its amplitude V_{IF} is related to the unknown imbedding impedance Z_c by

$$V_{\mathrm{IF}} = \frac{8\sqrt{2}}{\pi} \frac{\eta P}{\omega^2} \omega_{\mathrm{IF}} Z_{\mathrm{IF}} \frac{d(1/C)}{dv} \cdot \frac{1}{R_c} \frac{1}{1+((X_c - X_d)/R_c)^2} \qquad (1)$$

where P is the power used for device testing. V_{IF} is a function of the varactor bias, mainly through the reactance mismatch $X_c - X_d$. It comes through a maximum V_M when

$$X_c = X_d.$$

This yields the reactance of the circuit X_c with a good approximation. Now, let us consider two points of equal amplitude V_{IF} on each side of the resonance (which is observed by changing the varactor bias). The resistance of the circuit R_c is

$$R_c = \frac{1}{2} \delta X (a-1)^{-1/2}$$

where δX is the change of reactance between the two points, and $a = V_M / V_{\mathrm{IF}}$ (>1).

By careful amplitude calibration, one gets the efficiency η of the millimeter circuit, since all the parameters of (1) can be measured. Fig. 7 shows the general diagram of the setup used for this measurement.

This circuit characterization has been carried out on a large number of devices, by using a wide-band carcinotron as an oscillator. The S/N ratio was usually ranging between 20 and 30 dB. In order to facilitate the calculations, a programmable pocket calculator with printer was used to convert the measurements into impedance (or admittance) data. Fig. 8 shows an example of the admittance data obtained with back short tuning. The measured points lie on a circle as expected. This method has proven to be very useful to check the microwave properties of the circuit, as the surface resistance of the radial waveguide at millimeter frequencies.

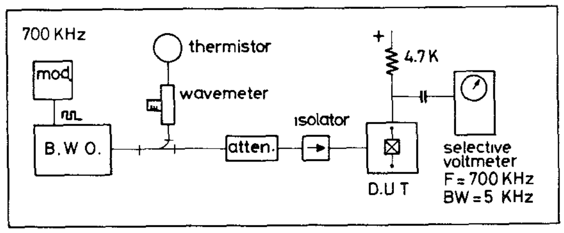

Fig. 7. General diagram of the setup used for characterizing the module by the modulation-transmission method.

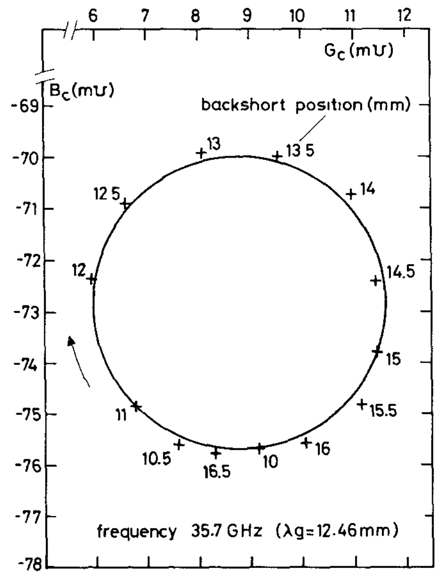

Fig. 8. Admittance measured as a function of the position of the back short. The diameter of the circle is a measure of the coupling to the waveguide. Its vertical position is related to the frequency of oscillation.

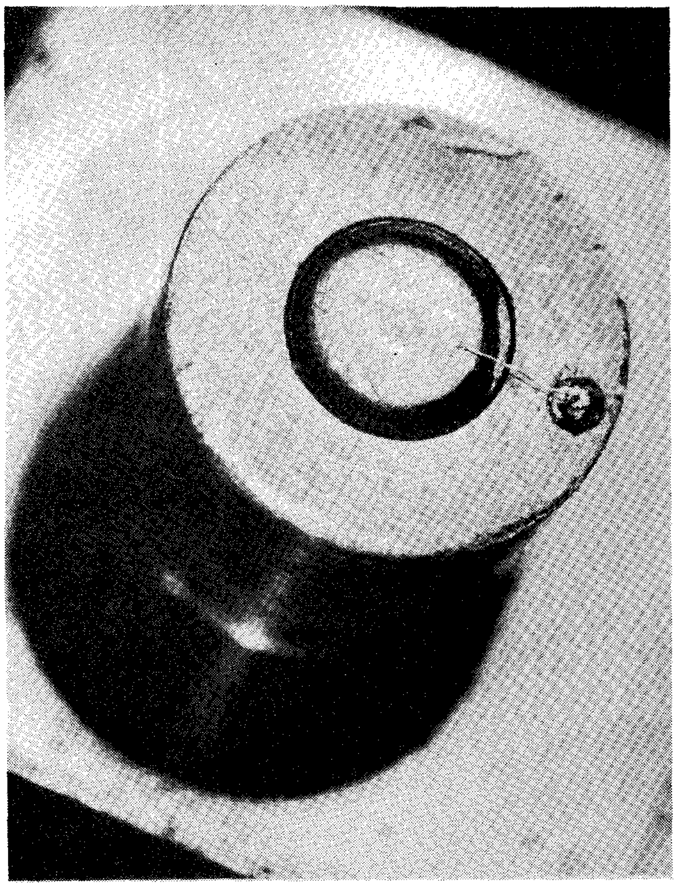

Fig. 9. Photomicrograph of a PTM mount with incorporated bias feed. The quartz ring is 2.5 mm in diameter, and the TC bonded gold wire on the right is 25 μm thick.

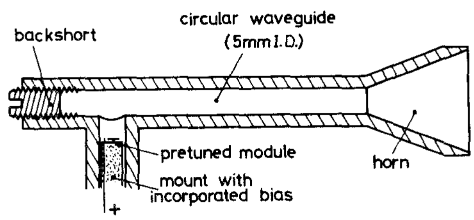

Fig. 10. Schematic cross section of a millimeter oscillator radiating directly into the free space, which is made from a PTM mount with incorporated bias. The positions of the module and of the back short are both adjusted for best power performance.

V. Source with Direct Radiation into Free Space

Most of the novel applications of millimeter waves involve free-space propagation. It is, therefore, important to have a technique which allows simple coupling of the device to a given aerial. In many cases, this aerial has to be highly directional as, for instance, in detection applications.

The pretuned module can be used in different configurations to solve this problem. One of them (Fig. 9) consists of having a small (about 1-cm^3) holder, with the bias feed incorporated to it. Eventually, a thermal heat sink can be attached to the holder. The dc bias is applied with an insulated wire crossing the thermal dissipator. A thin ($\sim$30-μm) gold wire is then bonded on top of the PTM as shown on the photograph. Due to its pretuned nature the module oscillates as such, and has an omnidirectional radiation diagram.

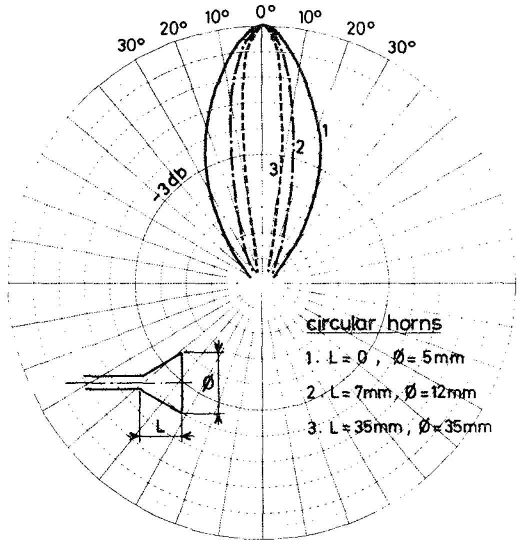

Fig. 11. Radiation diagrams of oscillators with different horns.

This source is small and simple, but cannot be used directly as an illuminator for a high-gain antenna. To do this, it is necessary to have a mode-filtering structure which provides the clean point source necessary for making a Gaussian beam. A simple way to do this is a hollow brass tube, 5-mm ID, acting as a TE_{11}-mode circular waveguide.

This tube is also a link between the PTM and the aerial. Looking down, we have either a horn (as on Fig. 10), or a Cassegrain antenna. Looking up is the PTM inserted in another tube, crossing the first one at a right angle. The mode filtering tube itself is terminated by a screw, which is a simple back short. The combination of this screw and of the position of the PTM allows enough adjustment for compensating for the VSWR at the open end and for optimizing the output power. Although this power cannot be measured directly, it can be estimated from the integration of the radiation diagram (Fig. 11). It is the same as with a conventional waveguide, and the radiation diagram is smooth.

VI. Conclusion

We have shown that the pretuned module is relatively simple to make, easy to use, and works well in a variety of simplified millimeter structures for free-space applications. These results look extremely encouraging as a first step.

We consider new developments in the future. As a matter of fact, the technologies are by no means settled definitively, and significant improvements can be expected. Some more flexibility can be obtained by external frequency adjustment. Also, the modular technique lends itself quite well to electronic tuning with a varactor diode, which will help make it even more practical.

Acknowledgment

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A Dual-Diode 73-GHz Gunn Oscillator

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***Abstract*—It is shown that two resonant cap structures can be mounted in a common waveguide to combine power from two Gunn diodes. Approximately 80-mW power was obtained at 73 GHz.**

I. Introduction

For the design of low-noise parametric amplifiers operating in the X band and above, pump frequencies of over 70 GHz are desirable. Single Gunn devices capable of generating adequate pump power—typically, 50 mW or greater—are not readily available at the time of this writing. Although greater than 50-mW power can be obtained from IMPATT devices, these are usually not suitable because of higher noise. Therefore, schemes employing more than one device are of interest. This paper reports one such scheme.

II. The Scheme

Resonant cap structures have been used for microwave generation with Gunn or IMPATT devices in the past [1]–[3]. In such an arrangement a diode is mounted on one wide wall of the waveguide. A disk is placed on top of the diode. Bias is applied through a post mounted on the disk. A moveable short is needed on one end for tuning. In the scheme described here, two such structures are placed in a waveguide, instead of one as shown in Fig. 1. Microwave power output approximately equal to twice the power obtainable from a single diode can be achieved in this way. In a somewhat similar arrangement, Stevens *et al.* have used post-mounted Gunn diodes separated $\sim\lambda g/2$ apart in a waveguide to combine power at 16 GHz [4].

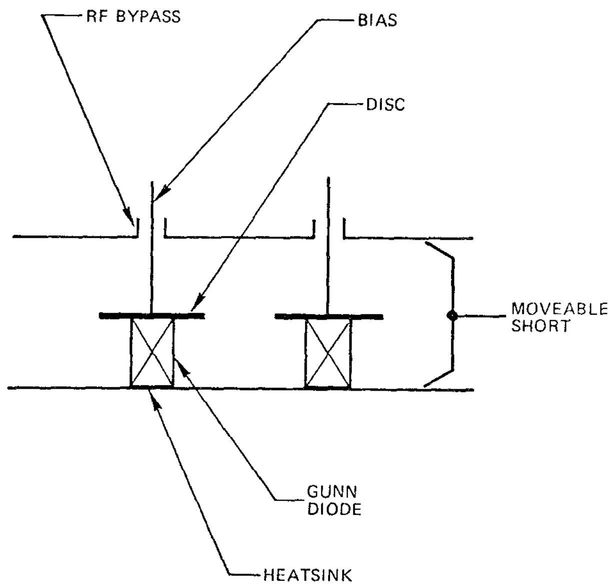

Fig. 1. Dual-diode Gunn oscillator.

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