TABLE III
RESULTS OF I-BAND SERIES COMBINING EXPERIMENTS WITH STACKED DEVICES

Diode	No. of Mesas	Net Cj (pF)	Breakdown Voltage (Volts)	Output Power (Watts)	Frequency (GHz)	Efficiency (%)	Pulse Width (usec)
42	2	0.75 @ −20v	57	22.4	7.74	17.9	0.5
21	2	0.56 @ −20v	58	20	8.21	24	1.5
20	2	0.86 @ −20v	48	14.3	8.36	16.5	0.2
26	3	0.51 @ −30v	74	30	8.05	16	0.2

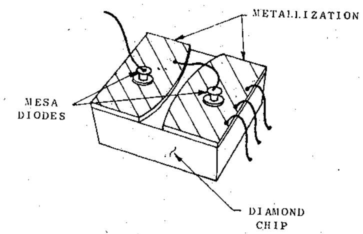

Fig. 2. Initial diamond heat sunk series pair.

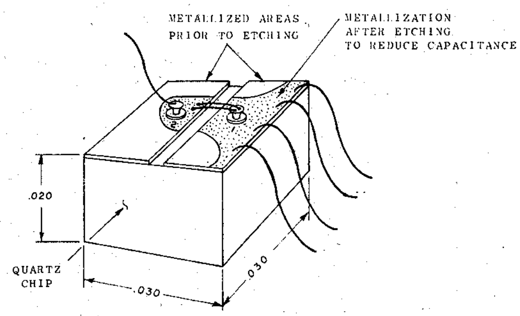

Fig. 3. Diode configuration for reduced capacitance and dielectric loss.

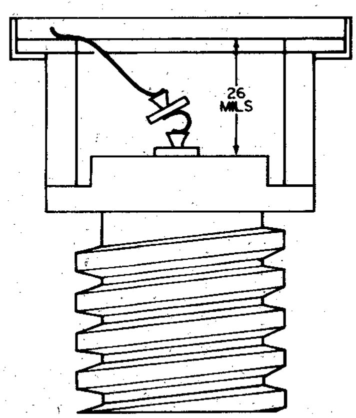

Fig. 4. I-band suspended series combining.

parallel. This approach was utilized in an effort to minimize the parasitic interconnect inductance. The top of the diamond was first etched with a "finger" protruding from one half of the metallization into the other half. The diode which was to be connected to ground was then mounted on the "finger" and a wide, low inductance gold strap (mesh or ribbon) was used between points A–A to contact the diode from both sides (Fig. 5). Numerous diode pairs were assembled in this configuration; none showed efficiency better than 12 percent. Geometric and heat-sinking restrictions did not allow further reduction in diode spacing.

As a final test of the techniques being used to fabricate the various test diodes discussed, series interdigital IMPATT diodes were fabricated. These diodes performed well, as expected from Josenhans' work [5].

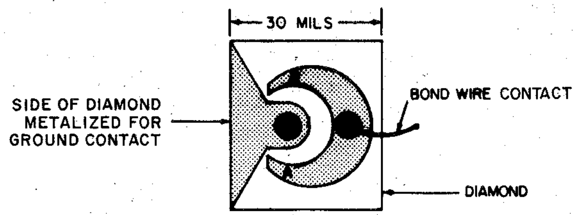

Fig. 5. Interdigital series pair. A wire mesh connecting points A to the top of the left diode is not shown.

CONCLUSIONS

The series connection of TRAPATT diodes for high efficiency operation has been shown to be feasible from approximately 0.5 to 8 GHz. In the lower frequency experiments, higher power outputs than would be expected from scaling considerations were often observed, accompanied by higher operating currents. This is probably due to the increased impedance associated with the series connection and is in agreement with the common experience that, all other things being equal, diode efficiency decreases with increasing area (decreasing impedance). At the lower frequencies, successful operation can be obtained with realistic parallel heat sunk structures suitable for long pulse or CW operation. This capability is yet to be demonstrated at I band.

ACKNOWLEDGMENT

The authors wish to thank M. L. Bark of the Naval Research Laboratory for his assistance in fabricating the devices used at 2 GHz, W. C. Jenkins of the Naval Research Laboratory for building the circuit for the 0.5-GHz measurements, and C. Bozler and G. N. Hill of Sperry Microwave Components for fabrication of the 8-GHz diodes.

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Characterization of Microwave Oscillator and Amplifier Circuits Using an IMPATT Diode Biased Below Breakdown

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***Abstract*—A convenient laboratory technique is described to measure the internal circuit loss, the load conductance, and equivalent circuit susceptances of microwave diode oscillators and amplifiers using an IMPATT diode biased just below its breakdown voltage.**

In the characterization of the admittance of IMPATT diodes at X-band frequencies or above, two difficulties are commonly encountered. Transformation of measurable admittances through the

Manuscript received February 4, 1974; revised April 8, 1974. This work was supported in part by an S. R. C. Research Grant and in part by the Ministry of Defense. Crown Copyright, published by permission of Her Majesty's Stationery Office.
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package and local circuit causes inaccuracy in the derived values of admittance at the diode terminals while unknown losses within the circuit mount cause inaccuracies in the estimates of diode conductance in addition to undesirable losses of RF power [1]. Our objective is to describe a simple technique to measure: 1) the equivalent susceptance directly across the terminals of the active semiconductor region (the diode terminals within its package); 2) the external load conductance referred to the diode terminals; and 3) the internal losses referred to the diode terminals at a frequency very close to the oscillator frequency. The technique has the advantages of speed and ease of operation and does not require the circuit disturbance often necessary with network analyzer techniques. It is applicable only when the resonance of the diode with its surrounding circuitry has a single-tuned form and any other spurious resonances are sufficiently displaced in frequency to allow the circuit to be represented by a simple lumped-equivalent parallel LC circuit [2], [3], as shown in Fig. 1. A further implied limitation in this representation which is usually met in oscillators but may be suspect in wide-band amplifiers is that the energy stored in the susceptive components must be much larger than that dissipated per cycle in the conductive components. Under these conditions the conductance is stationary near to the operating frequency and the susceptive circuit seen from the diode terminals can be simply described over the frequency range of interest by a susceptance of the form $B_0 + \{\partial B/\partial\omega\}_0\,\Delta\omega$ where the subscript refers to the center frequency. B_0 and $\{\partial B/\partial\omega\}_0$ can then be completely specified by a circuit containing only two components (an inductance L and capacitance C) connected in parallel across the diode susceptance. The requirement for low loss allows the loss to be expressed as simple parallel conductors [2], [3]. G_D is the diode conductance and G_C represents the internal circuit loss.

The resonant frequency of the circuit in Fig. 1 is ω_0 given by

$$\omega_0 = \frac{1}{[L(C_D + C)]^{1/2}}. \tag{1}$$

In practice the IMPATT is biased below breakdown and microwave power is injected from an external source (effectively from G_L in Fig. 1). The magnitude of the power reflection coefficient $|\rho|$ is given by

$$|\rho| = \frac{|G_L - (G_D + G_C)|^2 + (B_D + B_C)^2}{|G_L + (G_D + G_C)|^2 + (B_D + B_C)^2} \tag{2}$$

where B_D is the diode susceptance and B_C is the circuit susceptance. On resonance $|\rho| = \rho_0$ given by

$$\rho_0 = \left|\frac{G_L - (G_D + G_C)}{G_L + (G_D + G_C)}\right|^2. \tag{3}$$

Equation (2) shows that $|\rho| \simeq 1$ at frequencies away from resonance when the low-loss approximation mentioned in the preceding is valid. When the frequency is swept through resonance there is an absorption dip in the reflected power down to the value given by (3).

If $\Delta\omega$ is the full bandwidth between frequencies where $|\rho| = n\rho_0$, analysis of the circuit of Fig. 1 gives without approximation beyond those stated originally

$$G_L - (G_D + G_C) = \pm\left(\frac{1 - n\rho_0}{(n-1)}\right)^{1/2} \cdot \Delta\omega(C_D + C). \tag{4}$$

Usually, it is convenient to take $n = 2$ (3-dB points).

Using (3) and (4)

$$G_L = \frac{[1 \pm (\rho_0)^{1/2}]}{2(\rho_0)^{1/2}} \cdot \left(\frac{(1 - n\rho_0)}{(n-1)}\right)^{1/2} \cdot \Delta\omega \cdot (C_D + C) \tag{5}$$

$$(G_D + G_C) = \frac{[(1 \mp (\rho_0)^{1/2}]}{2(\rho_0)^{1/2}} \cdot \Delta\omega \cdot (C_D + C). \tag{6}$$

In passing it may be noted that the loaded Q factor Q_L of the circuit is

$$Q_L = \frac{\omega_0}{\Delta\omega} \frac{[\rho_0(n-1)]^{1/2}}{(1 - n\rho_0)^{1/2}}. \tag{7}$$

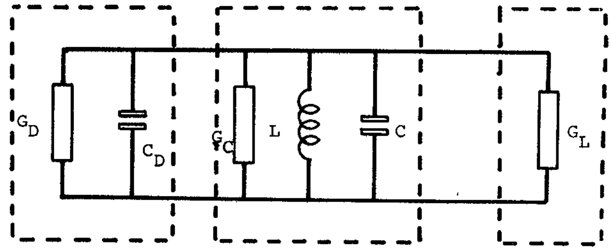

Fig. 1. Equivalent circuit of a singly resonant IMPATT and its microwave cavity.

There is an ambiguity between (5) and (6) because the same power reflection coefficient will be obtained for both one ratio of $G_D + G_C$ and G_L and for its inverse. The ambiguity was resolved in our experiments by noting the change of ρ_0 with bias voltage. In some cases ρ_0 increased with increasing voltage while it decreased in other cases. As the bias voltage was increased from a value below that necessary to completely deplete the diode active region we know that the losses must decrease as more of the device was depleted of free charge. Referring to (3), ρ_0 would then only increase with increasing bias voltage if $G_L > G_D + G_C$. Conversely, ρ_0 would decrease with increasing bias voltage if $G_L < G_D + G_C$.

Measurement of $(C_D + C)$ in addition to the directly measurable quantities ρ_0, ω_0, and $\Delta\omega$, will allow estimation of the quantities in (5) and (6). $(C_D + C)$ may be measured by making small variations in diode capacity ΔC_D via the bias voltage V_B so that the resonance frequency is altered by $\Delta\omega_B$:

$$(C_D + C) = -\frac{\omega_0}{2} \cdot \frac{\{\Delta C_D/\Delta V_B\}}{\{\Delta\omega_B/\Delta V_B\}}. \tag{8}$$

It is this variation of the diode capacity within the package that fixes the equivalent circuit at the diode active region terminals. The relationship between C_D and V_B may be measured at lower frequencies using accurate bridge techniques. Determination of the package reactance separately from the diode reactance is not critical for the relationship of V_B and C_D because it is only the slope of the relationship which has importance. The inductance L may be obtained from $(C_D + C)$ using (1) so that the reactive circuit components are determined. The modification of the resonant frequency when the diode is biased into useful operation above breakdown allows determination of the dependence of C_D on working conditions when dynamic susceptance effects become important [4].

Measurement of $(G_D + G_C)$, G_L, C, L, and Q_L using the preceding relationships was made with the equipment illustrated in Fig. 2. The bias voltage was as close to breakdown as possible without negative resistance effects distorting the absorption line shape and the injected RF voltage was small. The display of reflected power against frequency carried an added bonus. Any spurious resonance whose frequencies were not under the control of the diode bias voltage could immediately be identified. Measurements were confined to conditions where the resonance did not change its shape with small changes of bias voltage (which changes only one susceptive element) so that we knew the singly resonant condition assumed in the preceding was satisfied. Only the center frequency

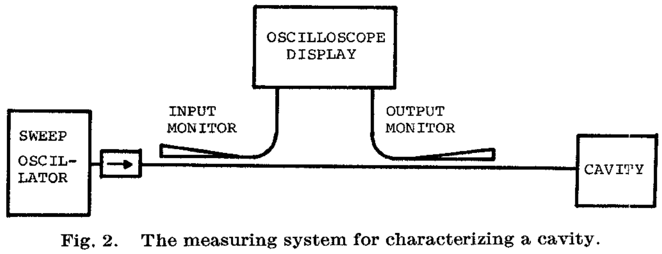

Fig. 2. The measuring system for characterizing a cavity.

TABLE I

Approx. Height	$\omega_0/2\pi$	C_D+C (pF)		L (pH)		G_L (mS)		G_D+G_C (mS)		$G_L+(G_D+G_C)$ (mS)		ρ_0		Q_L	
of Resonant Cap (mm)	(GHz)	SC	ML	SC	ML	SC	ML	SC	ML	SC	ML	SC	ML	SC	ML
1	9.54	3.8	4.1	73	68	2.0	4.6	1.7	2.1	3.7	6.7	0.007	0.13	61	36
2	9.50	3.2	3.2	89	89	1.7	3.7	1.3	1.7	3.0	5.4	0.021	0.14	64	36
3	9.45	2.7	2.7	105	105	1.5	2.9	1.0	1.3	2.5	4.2	0.036	0.15	64	38
4	9.28	1.6	1.4	179	206	1.3	2.7	0.7	1.0	2.0	3.7	0.096	0.21	46	22

was bias dependent. In practice, measurements satisfied this criterion if these spurious resonances were separated from the required resonance over a frequency range equal to several times $\Delta\omega$ or $\Delta\omega_B$. This range had to include the oscillation frequencies of the IMPATT's when biased above breakdown otherwise the derived circuit values would be irrelevant to the operation of the IMPATT diode when it was oscillating. The difference of oscillation frequency and the absorption frequency below breakdown was small enough to easily satisfy this criterion in our measurements, because our circuits had a large dominance of the equivalent circuit susceptance over the variable part of the diode's dynamic susceptance. Several types of cavity were used and results are shown for a waveguide cap circuit [5], [6] in Table I. Care has to be taken with this circuit because there is a resonance between the diode and the circuitry associated with the mounting post and radial cap and a further resonance in the waveguide between the sliding short circuit and the post. The results in Table I were taken for two conditions which were singly resonant. In one (marked SC) the short-circuit was an odd integral number of quarter wavelengths from the post so that it reflected a high impedance at the post plane. This was necessary both to avoid interaction between the two resonances and to give a known circuit condition for comparison with the next case. The positioning of the short circuit did not appreciably alter the results for small movements. In the second condition (marked ML) the sliding short circuit was replaced by a matched load. The objective of this was to increase the effective G_L by a factor of 2 because of the effective source conductance at the post terminals is half the characteristic conductance in this latter case. This was done because we had no other means of independently checking the measurements. It can be seen that this behavior is shown in Table I for various heights of the radial cap. A further point to note is the comparability of the useful load conductance G_L and internal loss conductance $G_D + G_C$. The ratio of these two is the ratio of the useful and lost power in the active oscillator. When the diode is operating as an oscillator the modulus of its negative conductance G_N must be equal to the sum of the positive conductances $G_L + (G_D + G_C)$. From (5) and (6),

$$G_N = G_L + (G_D + G_C) = \frac{(1 - n\rho_0)^{1/2}}{[\rho_0(n-1)]^{1/2}} \cdot \Delta\omega \cdot (C_D + C). \quad (9)$$

It may be argued that G_D is not the loss conductance present in the diode when it is biased into useful operation so invalidating the equality with G_N in (9). However, many diodes we have tested have an absorption bandwidth $\Delta\omega$ which becomes substantially independent of bias voltage just below breakdown implying that the residual losses are not caused by undepleted active regions of the diode and may still be present at the higher bias voltages of useful operation.

Once the cavity equivalent circuit parameters have been obtained it is possible to gain further information about the diode susceptance when it is oscillating or amplifying. An example is shown in Fig. 3 where the bias current induced variation of the diode susceptance ΔB_S at constant load impedance has been obtained from the bias dependence of the oscillation frequency $\Delta\omega_S$ where

$$\Delta B_S = -2(C_D + C) \cdot \Delta\omega_S.$$

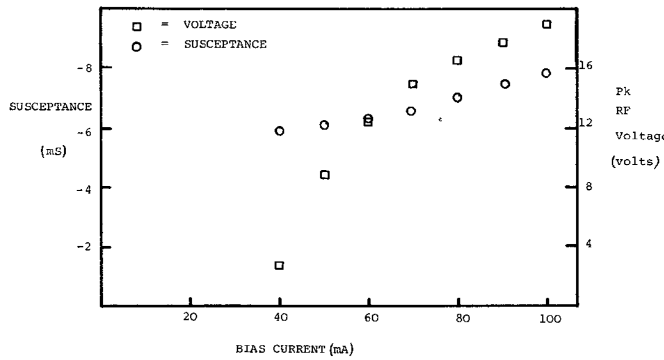

Fig. 3. Bias current variation of diode susceptance and RF voltage at $G_N = 2.7$ mS, $\omega_0 = 2\pi \times 10^{10}$ rad/s^{-1} in a coaxial cavity.

In conclusion we have developed a simple technique to characterize the diode and circuit of an IMPATT oscillator without mechanically interferring with the diode or its circuit. With some reduction of convenience and accuracy, we have also found the technique useful in characterizing the bandwidth-limiting and loss characteristics of circuits used for transferred-electron reflection amplifiers by substituting an IMPATT diode in the appropriate circuit.

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Very Large Impedance Steps in Microstrip

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Abstract—**Experimental determination of an equivalent circuit for very large impedance steps in microstrip is described. The equivalent circuit is shown to be valid in the frequency range from 1 to 2 GHz on 0.0635-cm-thick alumina substrate, although the experi-**

Manuscript received December 10, 1973; revised March 11, 1974.
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