Quenched-Domain Mode Oscillation in Waveguide Circuits

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Abstract—**An experimental study of pulsed transferred-electron oscillators operating in waveguide circuits is reported. Quenched-domain mode operation is shown to be present for short (less than 20-μm) supercritically doped GaAs samples operating at X-band frequencies (8.2–12.4 GHz) and for bias-voltage values up to six times threshold value. Analysis of the waveguide circuit shows that the chip sees a parallel resonant circuit, and theoretical computations with a sinusoidal voltage waveform are shown to be reasonably accurate. The load admittance presented to the chip is experimentally evaluated.**

I. Introduction

The experimental study reported here was made to determine the operating characteristics and modes of oscillation of short (active layers less than 20-μm) GaAs transferred-electron (TE) devices in X-band (8.2–12.4-GHz) waveguide circuits. It is of particular interest to determine if the quenched-domain mode (Q mode) could be obtained easily in waveguide circuits. All previous studies of the quenched-domain mode in this laboratory were conducted in a coaxial circuit [1].

Although several mode studies have been reported [2], [3], short devices have not been studied in detail. The present study is limited to pulsed operation in order to eliminate the complicated effects of increased lattice temperature. A bias pulse length of 0.5 μs and a repetition rate of 1 kHz were used for most of the tests.

Q-mode oscillators should be useful because the total RF power output is expected to decrease only slighly from the transit-time frequency to twice the transit-time frequency. Operation in waveguide circuits is desirable because low-frequency oscillations can be eliminated and widely tunable designs are possible. This mode is distinguished by the presence of high-field domains in the active layer of n-GaAs and by an RF voltage amplitude sufficient to "quench" these domains. That is the RF voltage is at least of value $V_B - V_{TH}$, i.e., the bias voltage minus the threshold voltage.

II. GaAs Material Characteristics

The TE devices tested in this study were made from wafers whose characteristics are given in Table I. The values given are only approximate, as some variation of threshold voltage and current existed for devices made from the same wafer. The donor density is constant within ± 5 percent for wafer A and is believed to be graded in wafer C. The devices were obtained from several device manufacturers and all wafers were grown epitaxially. Wafer A has an active n-layer grown by liquid epitaxy, a metallic (alloyed) cathode contact, and the devices were made by mesa construction. The other wafers have active layers grown by vapor-phase epitaxy, have an n$^+$ cathode contact on the active layer, and the devices were made as cleaved, square chips.

TABLE I
Approximate Parameters of GaAs Wafers Used for Fabrication of the TE Devices

Wafer Designation	n_0 (1/cm^3)	l (μm)	μ_H (m^2/V·s)	Area (cm^2)	Construction
A	2.0×10^{15}	15	0.78	1.1×10^{-4}	n-n$^+$
B	1.5×10^{15}	13	0.68	4.1×10^{-4}	n$^+$-n-n$^+$
C	2.2×10^{15}	18	0.60	4.1×10^{-4}	n$^+$-n-n$^+$
D	0.5×10^{15}	14	0.70	6.2×10^{-4}	n$^+$-n-n$^+$

The substrate is operated with positive bias for all devices. Wafers B and D would operate with reversed-bias polarity, but RF power output was less. For bias voltage well below the threshold value, the current–voltage relationships were linear and symmetric for all wafers, indicating that the contacts are all ohmic.

Tests of devices made from wafer D will not be described in detail as the devices are subcritically doped. A current drop was only observed for large RF level and the frequency of oscillation always decreased for increase of bias voltage. These results are typical of low n_0l product devices which do not form high-field domains.

III. Initial RF Tests of Wafer A

Initial tests were made on fifteen devices from wafer A with the crystals mounted in device packages of a size comparable to the 1N21 crystal packages (ceramic length$\approx$0.3 in). The devices were operated in a modified waveguide (WR-90) crystal mount with an external RF tuner. The RF power output and the dc–RF conversion efficiency were large only in a small frequency band approximately 2 GHz wide whose center varied from diode to diode. For bias-voltage values of approximately three and four times the threshold value, the RF frequency spectrum was stable and symmetric, and the RF pulse in time was single frequency and a duration the same as the bias pulse. Stable oscillation resulted in a current drop of 20–30 percent at four times threshold voltage and a dc–RF conversion efficiency of approximately 6 percent. The same device was operated in a coaxial resonant circuit and the maximum RF power output occurred at 6.5 GHz, which is the expected transit-time frequency.

In the bias range of useful operation, the RF power output and frequency always increase with an increase of bias voltage. The RF voltage values were estimated to be approximately equal to $V_B - V_{TH}$ (bias voltage minus threshold voltage) at frequencies of large RF output power. This condition is necessary for both Q mode and LSA mode.

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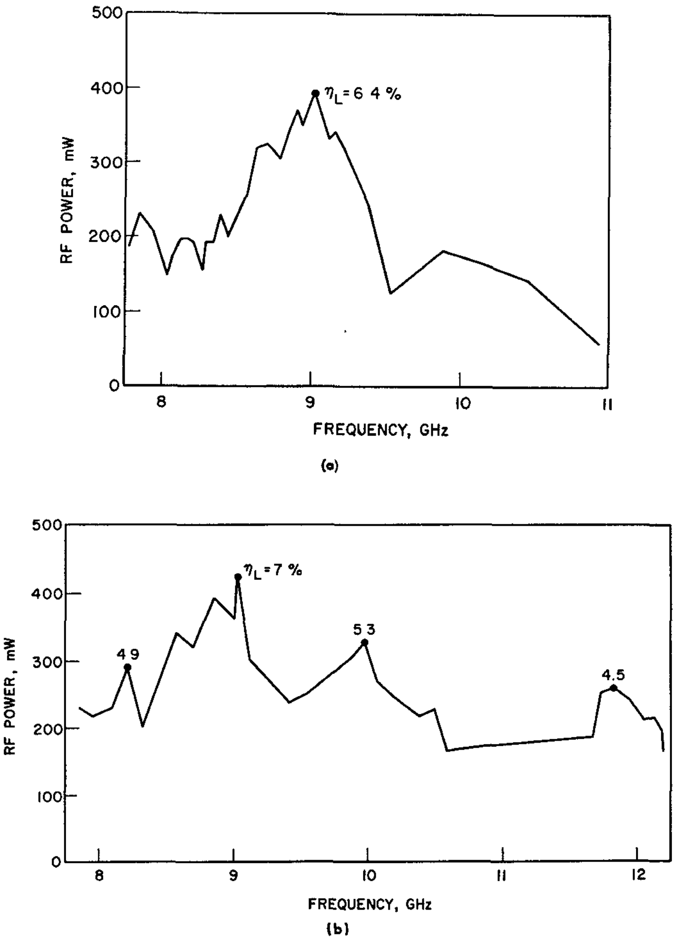

Fig. 1. Pulsed RF power output as a function of frequency for TE device A-24 in the waveguide circuit. (a) Matched load condition. (b) Slide-screw tuner used to maximize output power ($V_B=15$ V, η_L = load efficiency).

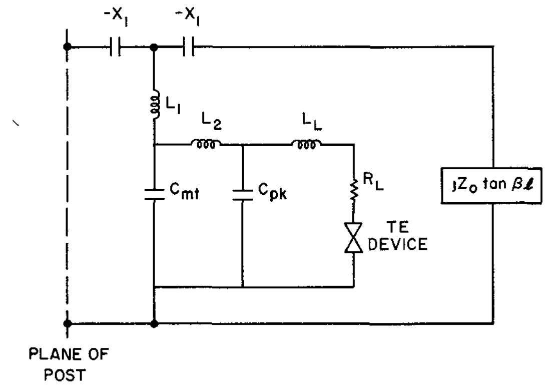

Fig. 2. Broad-band equivalent circuit for the waveguide (WR-90) TE device mount ($X_1=20$ Ω, $L_1=1.5+F(\text{GHz})-10)/16$ nH, $C_{mt}=0.05$ pF, $L_2=0.20$ nH, $C_{pk}=0.35$ pF, $\beta=2\pi/\lambda_g$, $L_L=0.52$ nH, $R_L=0.6$ Ω).

IV. Tests of Type A Devices in Small Device Packages

Five devices from wafer A were thermocompression bonded into small varactor cartridges (Type A-921, Ceramics International Division, Fansteel, Inc.) for RF tests. The length of the ceramic on these packages is only 0.033 in, and the lead wire is of 0.001-in diameter. The data reported here are for device A-24 which produced the best output power; otherwise, it is a typical device. The low-field resistance is 5.6 Ω and threshold conditions are $V_{TH}=4.45$ V and $I_{TH}=0.525$ A. Operation in a very lossy waveguide circuit produced transit-time oscillation at approximately 7 GHz for 10–20-V bias.

The measured RF power output as a function of frequency for this device is presented in Fig. 1(a) for the case in which an external tuner is not used. Fig. 1(b) shows the RF power versus frequency relation for operation with an external RF tuner. The frequency range over which significant power is obtained is much broader than for any of the devices in the larger package. Similar data were obtained at 20-V bias.

V. Waveguide Circuit Analysis

The waveguide resonant circuits used in these experiments utilize full-height X-band waveguide (WR-90). The small device package is centered in one broad wall and contracted by a 0.10-in-diameter inductive post which supplies bias power and is RF bypassed at the other wall. The circuit to be described in this section merely has an adjustable short plane about one-half waveguide wavelength from the inductive post.

RF impedance measurements have been made in 0.5-GHz steps for the full X-band range for the waveguide with 1) a full-height post, 2) an open-circuited package, and 3) a short-circuited package. For some of these measurements the sliding short circuit was removed and a matched load was placed behind the inductive post. The impedance for a matched load was taken as $(2b/a)(\lambda_g/\lambda)\eta$, where b/a is the waveguide height-to-width ratio, λ_g/λ is the ratio of waveguide wavelength to free-space wavelength, and η is the characteristic impedance of free space. Analysis of these data permitted the development of the equivalent circuit shown in Fig. 2, which is accurate over the full X-band range. The capacitive reactance X_1 and inductance L_1 are associated with the post. The calculations [4] predict that the capacitive reactance X_1 is nearly constant and that the post inductance L_1 increases slightly with frequency increase from 8 to 12 GHz. The impedance measurements verified this behavior. C_{pk} is the package capacitance, and L_L and R_L are the lead inductance and resistance, respectively.

The simpler equivalent circuit of Tsai *et al.* [5] was not adequate to explain the measured impedance data. However, Howes [6] has presented an equivalent circuit similar to Fig. 2.

The electrical position of the RF short was measured accurately in relation to its physical position, which is known within ±0.001 in. Impedance measurements made for the cavity using a short-circuited package and a packaged TE device (without bias) gave equal values of lead inductance. The lead inductance shows a small increase (approximately 20 percent) with frequency over the frequency band. Its average value is 0.52 nH. The lead resistance for the short-circuited device was constant and equal to 0.6 Ω.

Fig. 3 shows the calculated load conductance and susceptance presented to the TE chip for a shorting plane position 2.354 cm from the post and for three different waveguide load conditions. Z_L is the waveguide load at the plane of the post. The load conditions are Z_0 (matched load), $\frac{1}{2}Z_0$, and $2Z_0$. Oscillation with capacitive TE devices made from wafer A occurs at approximately 9.01 GHz, which is closest to the frequency of a parallel circuit resonance.

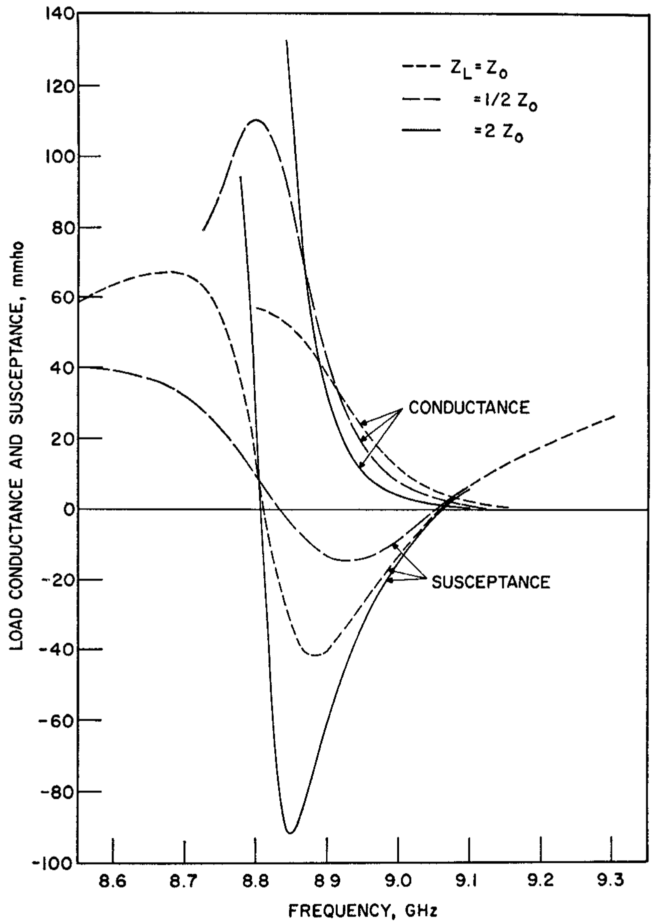

Fig. 3. Theoretical load admittance presented to the TE chip in the waveguide circuit as a function of frequency. Short plane is 2.354 cm from the post.

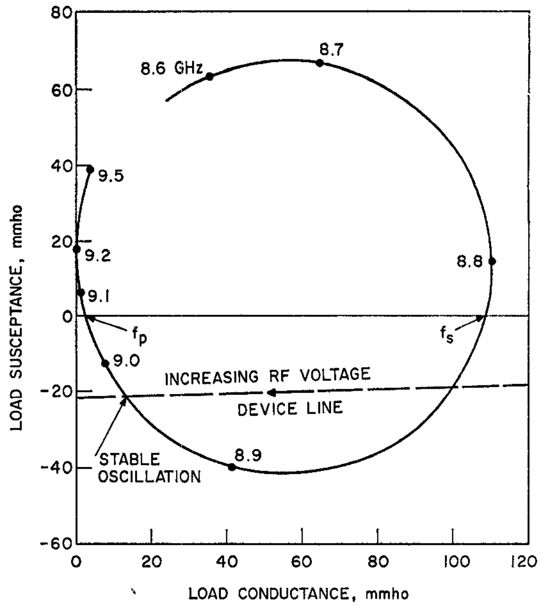

Fig. 4. Theoretical load admittance presented to the TE chip in a small package for the waveguide circuit with the short plane 2.354 cm from the post ($f_s = 8.8125$ GHz, $f_p = 9.062$ GHz).

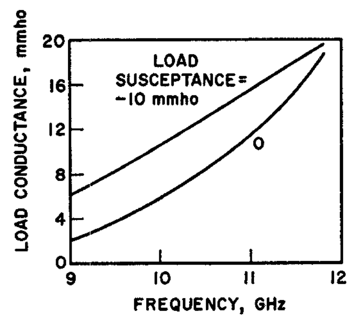

Fig. 5. Load conductance as a function of resonant frequency for matched waveguide circuit for two values of load susceptance.

From Fig. 3 the R_0/Q_0 factor for the matched circuit is found to be 0.994 Ω. The external Q factor Q_x was evaluated for the matched load condition and found to be 200. A load conductance estimate is $(R_0/Q_0 \cdot Q_x)^{-1} \approx 0.005$ mho (at 9.01 GHz), which is only a little different from the actual load conductance of 0.006 mho. However, the load conductance changes so rapidly with frequency that this method of calculation is not always reliable.

Fig. 4 shows the admittance plane plot of the load admittance for a matched waveguide load. Notice that the total circuit exhibits one series resonance (at 8.8125 GHz) and one parallel resonance (at 9.062 GHz) in X band. The dashed line represents the negative of the admittance of a fictitious TE device as the device RF voltage is varied for a fixed value of voltage bias. As Slater [7] has shown, the left-hand intersection of the device and circuit lines (near parallel resonance) is the only stable oscillation condition.

Fig. 5 shows how the load conductance increases with increase of the circuit resonant frequency (by decrease of the post to short-plane separation). This is the reason for the poor TE device performance without a tuner at high frequencies.

VI. TE Device Admittance Results

For the matched load condition, the device admittance can be calculated from the frequency of oscillation using the known equivalent circuit. The RF voltage can be obtained from the RF power value and the real load presented to the device. It is found that operation above 9.25 GHz occurs with an RF voltage less than $V_B - V_{TH}$ for 15-V bias. Thus increased device loading with frequency not only reduces the RF power but also reduces the RF voltage.

The load presented to the TE device (chip) with and without a tuner is shown in Fig. 6 for 15-V bias. For measurements with a tuner, the actual waveguide load admittance was measured at each frequency after adjustment of the tuner for best RF power output and the load presented to the TE device was evaluated.

Notice that the load susceptance is nearly the same for the two cases in Fig. 6, but that the load conductance is significantly lower at high frequencies for operation with the tuner. The RF voltage differences existing in the two cases mainly affect the device conductance and not the susceptance.

Fig. 7 shows the TE device admittance, RF voltage, and capacitance (C_D) as a function of frequency for optimum load (with tuner) at 15-V bias. The device capacitance is calculated from the susceptance and shows a tendency to increase with frequency. The low-field device capacitance is estimated from the area and thickness to be approximately 0.05 pF. Note that C_D is between two and four times this value.

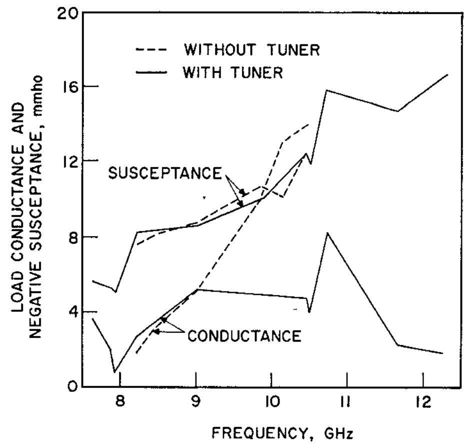

Fig. 6. Load conductance and susceptance versus frequency for TE device A-24 for the matched load case and for the best RF power condition using an RF tuner in load ($V_B = 15$ V).

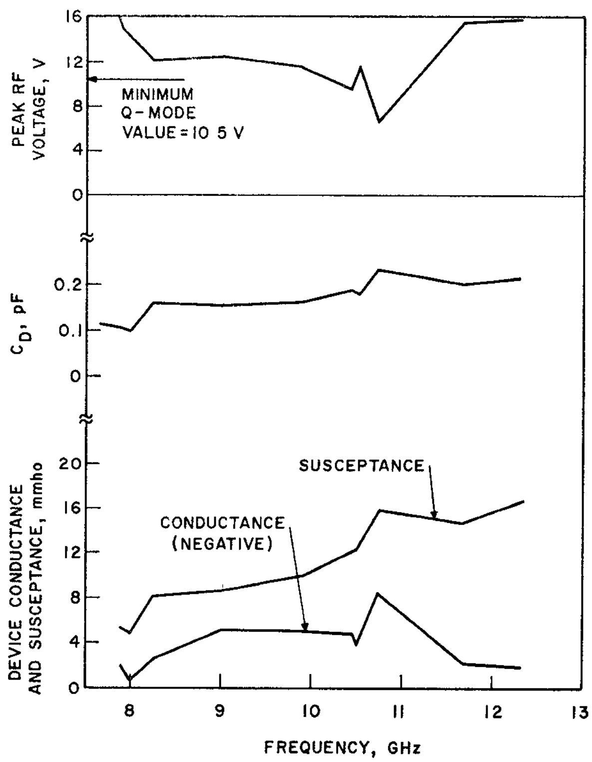

Fig. 7. Admittance, capacitance, and RF voltage as a function of RF frequency for TE device A-24 in the waveguide circuit with a tuner for best output power ($V_B = 15$ V).

The RF voltages are larger than for the matched load case but are still lower than expected for Q-mode operation over a portion of X band in both cases. The data obtained for 20-V bias were similar.

Fig. 8 shows the TE device characteristics near 9 GHz for various values of bias. A decrease in conductance and susceptance clearly occurs for increasing bias values. Furthermore, the RF voltage is sufficient for Q-mode operation at each value of bias.

These results are generally consistent with Q-mode behavior in a parallel resonant circuit. However, it does appear that for heavy circuit loading the fundamental RF voltage is

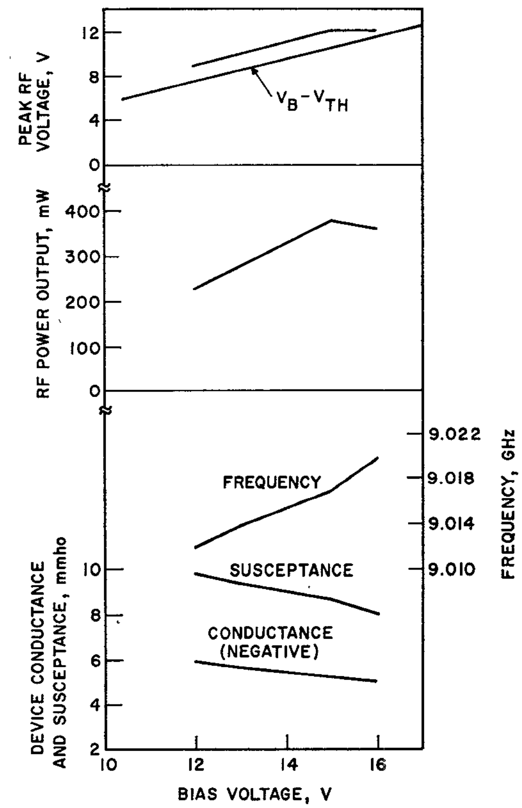

Fig. 8. RF power output, frequency of oscillation, RF voltage, load conductance, and susceptance as a function of bias voltage for TE device A-24 in the waveguide circuit without a tuner and for a post-to-short distance of 2.354 cm.

not always as large as required theoretically for the Q mode. In these cases, domain quenching must still occur, presumably by harmonic voltages. When the device loading is adjusted for maximum power output, the RF voltage is of the proper value and the Q-mode analysis should be accurate [1].

VII. Tests of High-Power Pulsed Devices

Six devices were tested from wafers B and C. The device manufacturer had thermocompression bonded the devices into small varactor packages (type S4).

The devices made from wafer B have a threshold voltage approximately equal to 5.5 V for either polarity and exhibit a strong current drop above threshold. Wafer C devices had higher threshold voltages and smaller current drops. The maximum permissible bias is limited by the device switching to a high-current state. The voltage at which this occurs varies greatly with adjustment of the RF circuit.

Fig. 9 presents efficiency data for the best device (wafer B) in a waveguide circuit. A slide-screw tuner and the bias voltage are adjusted to achieve maximum efficiency at each frequency. Oscillation can be obtained over the full X-band range, and operation within a small frequency region at approximately 9.5 GHz allowed better RF efficiency than obtained elsewhere. Fig. 10 shows a typical measurement of RF voltage, RF power, and frequency as a function of bias voltage. The RF voltage was estimated and is seen to be larger than $V_B - V_{TH}$, as it was in all other tests on this device. The RF circuit was not adjusted during this test so it is clear that the device has a positive tuning characteristic. This mode

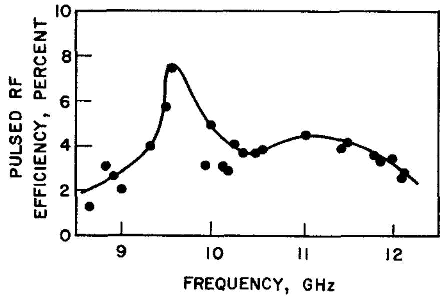

Fig. 9. Maximum pulsed RF efficiency as a function of frequency for device B-12 in a waveguide circuit with pulsed bias voltage 20–25 V ($V_{TH}=5.4$ V).

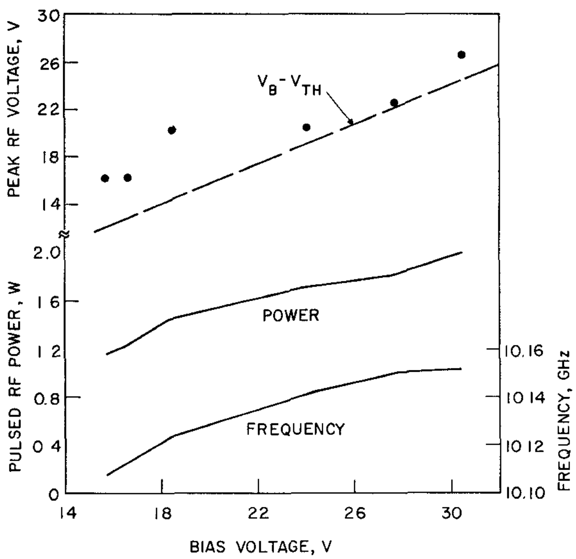

Fig. 10. Pulsed RF power, frequency, and estimated RF voltage as a function of bias voltage for device B-12 in a waveguide circuit ($V_{TH}=5.4$ V, $I_{TH}=1.8$ A, $I_{test}\approx 1.2$ A, $R_{\text{low field}}=2.2\ \Omega$).

could be obtained at bias values down to 12 V or approximately twice the threshold voltage. A minimum of 1 W of RF power could be obtained from this device throughout X band with the same type of power and frequency relation shown in Fig. 10.

The RF voltage values were estimated from the measured value of Q_x, R_0/Q_0, and P_x, the output power. Q_x was determined from the frequency-pulling range for a slide-screw tuner of known mismatch, and R_0/Q_0 was determined from measurement of the RF cavity tuning rate using a device package containing a varactor. The external load conductance is $(R_0/Q_0\cdot Q_x)^{-1}$ and the peak RF voltage [1] is approximately equal to $\sqrt{2P_x(R_0/Q_0)Q_x}$.

As shown in Fig. 11, oscillation with good efficiency was obtainable only near the transit-time frequency (approximately 5 GHz) and near twice the transit-time frequency (approximately 9.5 GHz) for the same device in a coaxial circuit. The wide tuning range in X band observed in the waveguide circuit could not be achieved in the coaxial circuit. However, the RF power and frequency again increased with increase of bias for all frequencies in the coaxial circuit.

The same tests were made on other devices constructed from wafers B and C and the results were similar to the data presented. The RF conversion efficiency for wafer C devices

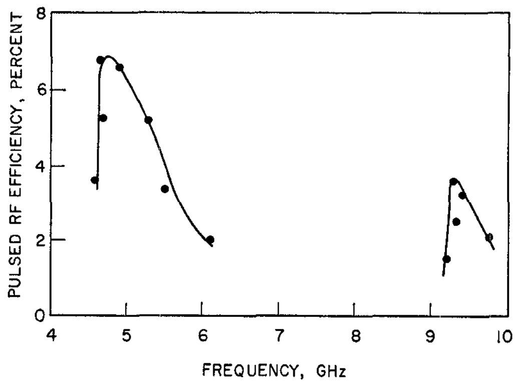

Fig. 11. Maximum pulsed RF efficiency as a function of frequency for device B-12 in a coaxial circuit for pulsed bias voltage from 14–25 V.

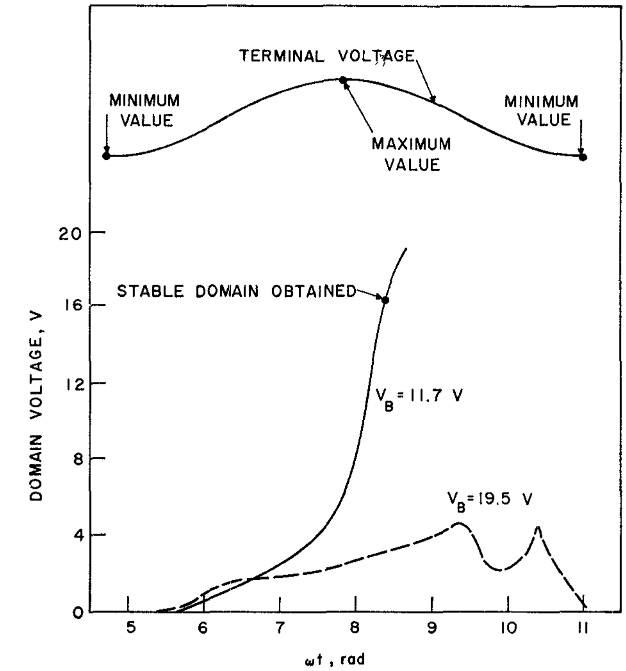

Fig. 12. Relative terminal voltage and values of domain voltage as a function of time from LSA program for a GaAs crystal ($l=13$ μm, $n_0=1.5\times 10^{15}$/cm³, $f=10$ GHz) with large bias voltage ($V_B=19.5$ V, $V_{ac}=17.5$ V) and with small bias voltage ($V_B=11.7$ V, $V_{ac}=9.75$ V). A donor variation of 1 percent was assumed.

was everywhere less but more uniform over the X-band frequency range.

Fig. 12 shows the domain growth process if the LSA mode [8] is assumed to be present for the parameters of wafer B with a donor density variation of 1 percent. Notice that a mature domain is formed at low bias (11.7 V). At the larger bias value (19.5 V), the domain voltage remains small enough so that LSA operation is achieved as long as the donor variation is not increased above 1 percent. Since wafers B and C are known to have donor variation of at least 10 percent, the LSA mode cannot exist.

Q-mode computations [1] do agree reasonably well with the experimental results. For the parameters of device B-12, RF power of 2.2 W at 13 percent efficiency was predicted at $V_B=20$ V for 10-GHz operation. For $V_B=30$ V, the RF power increases to 2.6 W but the efficiency drops to approximately 10 percent. Assuming some RF power loss to the resonant circuit (and lead resistance), these predictions agree reasonably well with the observed performance at 9.6 GHz: 1.8 W, 8.8 percent at $V_B = 18$ V and 2.2 W, 7.1 percent at 28 V.

VIII. Conclusions

It is concluded that Q-mode operation is present in the experimental devices that are supercritically doped for the following reasons. The RF voltage is sufficient for domain quenching or LSA operation, but the LSA mode has been shown not to exist in these devices. All of the experimental characteristics are consistent with the Q mode; for example, power and frequency increase monotonically with bias-voltage increase. The oscillators are widely tunable above the transit time frequency. The chip RF capacitance is two to four times the low-field capacitance. Finally, the RF power and efficiency calculated by Q-mode theory for operation at 20 and 30-V bias are in good agreement with the values measured experimentally, if some RF circuit loss is assumed to be present. Thus experiments with pulse-biased highly doped TE devices have verified Q-mode operation in X-band waveguide circuits over a bias-voltage range from approximately twice to approximately six times the threshold value. A broad frequency range of operation was obtained for proper RF circuit loading of chips in small device packages. The highest efficiency was obtained with GaAs material of n^+-n-n^+ construction which exhibited a current drop of greater than 30 percent.

Analysis of the post-type waveguide circuit showed that oscillation occurs with TE devices when the post to short-plane separation is about one-half waveguide wavelength. When viewed from the chip, the circuit is operating slightly below the frequency of parallel circuit resonance. Circuit RF loading effects were shown to be important for wide bandwidth operation of quenched-domain mode oscillators.

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On the Optimum Design of Tapered Waveguide Transitions

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***Abstract*—It has been found experimentally that the conventional optimization of waveguide tapers for the interconnection of circular waveguides with different diameters fails if the ratio in the diameters becomes too large. With the aid of an accurate numerical analysis program, the reason for the failure was found to be the reconversion from the unwanted mode to the main mode, which is neglected in all known synthesis procedures. The performance of tapers can be considerably improved by the implementation of other design equations and establishing new design criteria. This results in somewhat longer tapers. Various tapers were designed according to these procedures for a maximum of −40-dB H_{02}-mode level between 40 and 110 GHz, and preliminary measurements on fabricated units substantiate the improvement. It is further shown that the mode conversion at cutoff does not exhibit any singularity.**

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I. Introduction

Inhomogeneous TEM transmission lines and waveguides in the form of tapered matching sections or tapered waveguide transitions are frequently used for very broad-band applications [1]–[3]. For the intended use of millimeter waves as a transmission medium in communication systems, tapered circular waveguide transitions are needed which do not generate spurious modes above a tolerable level [4]. The problem in the design of these inhomogeneous waveguides is basically that of specifying a distributed inhomogeneity for minimum mode conversion and/or mini-