Experimental results obtained at an input power level of 50 mW show a similar improvement in the gain-bandwidth product—the 3-dB bandwidth of 320 MHz at 7-dB gain becomes 1030 MHz on compensation at the same gain.

V. Conclusions

It has been shown both theoretically and experimentally that the circuit technique of active reactance compensation is useful for broad-banding IMPATT amplifiers. The results obtained are in agreement with those predicted previously for parametric amplifiers.

There is reasonable agreement between the computed bandwidths and the bandwidths calculated from the analytical expressions derived for the gain bandwidth products both for the uncompensated and actively compensated amplifiers. The gain bandwidth product for the experimental amplifier is slightly higher than the corresponding theoretical and computed figures. It is suggested that this may be due to the 1-dB gain ripple which is observed in practice.

Acknowledgment

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Circuit Behavior and Impedance Characteristics of Broad-Band TRAPATT-Mode Amplifiers

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Abstract—**The characteristics of TRAPATT-mode high-efficiency oscillators and broad-band amplifiers are reviewed. It is concluded that a broad-band amplifier like a high-efficiency oscillator should have capacitive circuit impedances and is distinguished from a high-efficiency oscillator principally by the number of important harmonics employed. A smaller number of harmonics for the amplifier can lead to broader bandwidth but lower efficiency. The relative merits of experimental amplifer circuits are discussed. It is shown that coaxial-line circuits employing diode packages with large lead inductances are characterized by harmonic impedances which can have large values over broad frequency bands. However, it is also shown that the device waveforms in this case are excesssively degraded and relatively low-efficiency results. On the other hand, coupled microstrip circuits with a low-impedance diode mount can provide broadband low impedances at both fundamental and third harmonic and have exhibited better performance.**

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I. Introduction

THE THEORY of TRAPATT-mode oscillations in avalanche diodes has been well established by Clorfeine and others [1]–[3]. TRAPATT-mode operation has been obtained with both high-efficiency oscillator and broad-band stable-amplifier circuits. Evans [4], [5] analyzed high-efficiency oscillator circuits and showed that these circuits rely on a large number of harmonics to

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achieve high efficiency. The work of Evans and Snapp [6], Yanai [7], Carroll [8], and Mackintosh [9] provides a clear description of the characteristics of the narrow-band TRAPATT-mode high-efficiency oscillator. Yanai [7] also discussed starting conditions and showed clearly that the IMPATT mode is not required for TRAPATT starting.

Several authors have discussed broad-band TRAPATT-mode amplifiers [10]–[12]. However, a clear discussion of the characteristics of these amplifiers has not been presented. In the present paper a brief review of high-efficiency oscillators is presented first. It is argued that an oscillator and a broad-band amplifier do not represent fundamentally different modes of operation, and that their circuit behavior and impedance characteristics can be similar. Measured and calculated impedance characteristics of amplifiers employing coaxial-line and coupled-microstrip circuits are presented. These and previously published impedance data are discussed in terms of amplifier performance.

II. High-Efficiency Oscillators

Fig. 1 shows the configuration of the high-efficiency oscillator circuit which is now generally known as the Evans circuit and is believed to be capable of higher power and efficiency than any other TRAPATT circuit that has been proposed. The basic operation has been described as follows by Evans [4], [5]. The rapid drop in diode voltage caused by the avalanche is transmitted through the package and mount network and down the transmission line. The wavefront reflected at the low-pass filter with reversed polarity arrives back at the diode after the trapped plasma has been drained from the diode and initiates a new avalanche. Good performance with the Evans circuit results when the effects of the package and mount are small and the connecting line has low dispersion. Under these conditions, the returning wavefront contains a large number of harmonics with nearly optimum phasing for fast rise time. This is a requirement for high power and efficiency but also leads to narrow bandwidths.

In the case of a high-efficiency oscillator, measured values of the circuit impedance presented to the diode chip [5], [6], [13], and [14] are capacitive at the fundamental frequency F_1 and lower harmonics as might be expected from the inductive nature of the rapidly switching avalanche diode. More specifically, the measured circuit reactances are capacitive up to a frequency $F_p = n_p F_1$ and can be inductive above F_p. A simple approximate model can be used to calculate a circuit reactance characteristic which is in qualitative agreement with the measured reactance data at the harmonic frequencies. In particular, if effects of the package and mount are small, experimental values of the length l of the connecting line in Fig. 1 are nearly one-half wavelength $\lambda_1/2$ at F_1, i.e., $l = \lambda_1/2 - \Delta$ where Δ is small. Assuming that the low-pass filter presents an effective short circuit at all harmonics F_n, then the harmonic impedances presented by the circuit at the

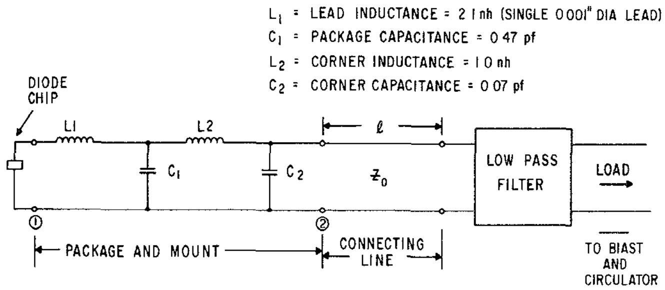

Fig. 1. Equivalent circuit for coaxial TRAPATT oscillator (or broad-band amplifier).

diode are pure reactances X_n given by $Z_0 \tan(2n\pi l/\lambda_1) = -Z_0 \tan(2n\pi\Delta/\lambda_1)$. Thus the line reactance has a pole for $n = n_p = \lambda_1/4\Delta$ and is capacitive for $n < n_p$. At the lower harmonics the diode reactance determined from the above expression is approximately that of a constant inductance $= Z_0\Delta/v$ where v is the velocity of propagation of the connecting line. Carroll [8] has suggested that the circuit acts approximately as a negative inductance, i.e., as a capacitive reactance the magnitude of which increases with increasing frequency.

The optimum value of Z_0 represents a compromise between production of and recovery from the highly conducting state. Thus the charging time constant Z_0C_0 (where C_0 is equal to the depletion layer capacitance near breakdown plus any other capacitance near the diode) must be short enough to permit a fast increase in voltage to produce a strong plasma. The requirement that the recovery time be about half [2] of the fundamental period $1/F_1$ sets a lower limit on Z_0. Thus, since recovery time is roughly proportional to Z_0 [4], too small a value of Z_0 causes the recovery time to be too short, and an efficient waveform is not obtained. Values of the product $Z_0C_0F_1$ of about 0.07 (i.e., a time constant about 7 percent of the fundamental period) have yielded good results in experimental oscillators, provided package and mount effects are small. In practice, discontinuity effects associated with the mount may become significant if Z_0 is made small enough to accommodate a large area diode. In this case, a lumped capacitor very close to the diode may be helpful [4].

It was natural to try to deduce an impedance characteristic for broad-band amplifiers from the above simple model for high-efficiency oscillators. It was reasoned that the broad-band amplifier impedance function should be similar to that for the high-efficiency oscillator but with a smaller number of important harmonics, i.e., harmonics below a certain frequency F_p which have capacitive circuit reactance. It was further reasoned that for $F > F_p$ the circuit reactance of each harmonic should be stabilized, i.e., have high values over a broad frequency band. Thus, with fewer important harmonics, the bandwidth would be greater because the diode waveforms would be less sharp and could be maintained more easily over a given bandwidth. However, it also seemed clear that the increase in bandwidth could only be obtained with a reduction in efficiency. Obviously, for high efficiency, an amplifier circuit should provide broad-band capacitive reactances at as many harmonics as possible.

Note that in this harmonic analysis the high-efficiency oscillator and broad-band amplifier are not considered fundamentally different modes of operation. In particular, it is emphasized that time-delayed triggering (TDT) [10], [13], and [15] occurs in a broad-band amplifier as well as in a high-efficiency oscillator. Thus, based on the harmonic analysis, one expects that amplifier operation with any prescribed bandwidth (within certain limits) can be obtained by replacing a part (or all) of the nondispersive connecting line of the classical Evans circuit with a dispersive lumped-element line which passes only the lower harmonics and cuts off at the higher harmonics. Qualitatively, the bandwidth expected depends on the amount of dispersion introduced.

III. Coaxial-Line Amplifier Circuits

The configuration of Fig. 1 can also represent the equivalent circuit for one type of broad-band amplifier. In this case, limitation of the number of important harmonics can be accomplished by employing a package with a relatively large lead inductance which can be modeled as a length of high-impedance line. Therefore, it is clear that the effect of a large value of L_1 on the efficiency is approximately equivalent to employing too large a value of Z_0 as discussed in Section II. The effect of L_1 on the circuit impedance can be further understood by noting that the frequencies at which a reactance pole occurs are independent of L_1, whereas the frequencies at which a reactance zero occurs are lowered by L_1. As a result the reactance slope dX/dF occurring on the high-frequency side of a pole increases with increasing values of L_1. The value of L_1 affects primarily the bandwidth which hopefully is increased because a smaller number of important harmonics are required for the diode waveforms.

Stable amplifier operation (i.e., zero or very low output power in the absence of an input signal) or free-running oscillation can be obtained by adjustment of only the resistance presented to the diode at the power-extraction frequency (e.g., the fundamental or second harmonic). Experiments show that this is true even for amplifiers designed with large L_1. For example, in an S-band experiment, the impedance of the diode was about 5 Ω for the free-running oscillator and about 20 Ω for the broad-band amplifier, these values of resistance being consistent with the measured value of 6-dB gain.[1] Furthermore, bias conditions, added power, and efficiency were the same in both cases. Since only the cold circuit impedance at the fundamental was changed in these experiments, the results imply that the oscillator and amplifier states were identical [16], i.e., harmonic impedances and device waveforms were the same in both cases. These data also show that the TRAPATT amplifier is open-circuit stable, whereas IMPATT amplifiers have been observed to be short-circuit stable. Finally, it should be noted that a free-running pulsed oscillator with large L_1 usually exhibited considerable interpulse starting jitter and required very critical tuning. The jitter was very much less in the case of an amplifier with large L_1. Thus the free-running oscillator and driven amplifier differ essentially only in terms of starting conditions and not under steady-state conditions.

Fig. 2 shows measured and calculated impedance data at the diode chip[2] position for an experimental coaxial-line broad-band amplifier having the equivalent circuit of Fig. 1. These data were obtained by transforming the measured (or calculated) impedance values at reference plane 2 to reference plane 1 through the package and mount network, the element values of which had been measured previously. For these tests, the diode chips were mounted in relatively large packages (International Ceramics A819 or A924). A single 0.001-in-diam gold wire lead between diode chip and package rim resulted in an inductance $L_1=2.1$ nH, while a pair of 0.001-in leads to diametrically opposite points on the package rim resulted in $L_1=1.2$ nH. Tuning slugs (not shown in Fig. 1) were employed beyond the low-pass filter to optimize the impedance at the fundamental frequency. The impedance data in Fig. 2 are for the slugs optimized at 2.8 GHz; however, because of the low-pass filter, the impedance at the harmonics is essentially independent of the slug positions. These slugs could be adjusted to obtain either free-running oscillator or stable-amplifier operation as discussed above.

The point-by-point data in Fig. 2 show output power obtained with a constant input power of 11 W by adjusting the tuning slugs at each frequency. Note that the power curves in Fig. 2 represent power at the fundamental frequency only. The power curves in the harmonic regions are included to show how harmonic impedance affects fundamental power. Therefore, the power data in Fig. 2 define the fundamental frequency band (in this case, 2.65–3.075 GHz) over which acceptable harmonic reactances occur for TRAPATT amplification. No adjustment of the tuning slugs could result in TRAPATT amplification outside this 14.85-percent bandwidth. It was concluded that instantaneous bandwidth, being at most only about 8 percent, was limited principally by the reactance slope at the fundamental. Note that the reactance slope dX/dF at F_1 is much greater than the value $4\pi L_1$, associated with a single-tuned series resonance of L_1 and a simple capacitor.

The data in Fig. 2 show that a low-impedance series resonance is obtained near the fundamental frequency, as in the case of the high-efficiency oscillator. However, in contrast with the oscillator case, high values of impedance

[1]Similar results were also reported in [11] for a coupled-line circuit.

[2]n^+pp^+ silicon diodes were employed having breakdown voltages between 90 and 100 V, depletion layer widths of 3 μm and active layers of 1.4×10^{-4} cm^2. The corresponding depletion layer capacitance at breakdown was approximately 1.2 pF. Typical operation was with dc pulse bias of 60 V and 3 A. Bias pulse widths of 0.4 μs were employed.

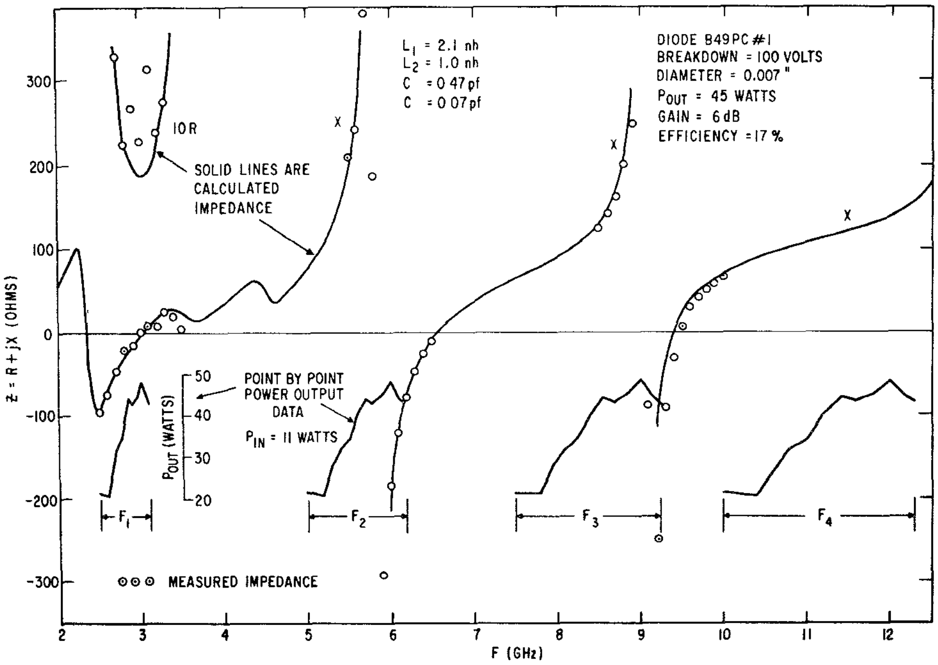

Fig. 2. Impedance characteristics for coaxial-line amplifier.

are obtained within the harmonic bands. Recalling that the impedance at the fundamental was optimized at each frequency, it is seen that the data indicate that TRAPATT amplification is limited at the low end of the band because the circuit provides too low an inductive reactance at the harmonics. With the diodes employed in these tests, a value of inductive reactance greater than 75 Ω appears to be necessary for TRAPATT amplification. TRAPATT amplification also appears to be limited at the high end of the band because the circuit provides too low a reactance at the harmonics. In particular, it appears that the high end of the amplification band is limited because the reactance at F_3 passes through zero and takes on a small positive value. However, because of the high reactance slopes near the high end of the second- and third-harmonic bands, a slight change in frequency can cause a very large change in reactance. For this reason, these particular data are not considered reliable enough to establish whether TRAPATT amplification is limited because the circuit provides too low a capacitive reactance or because the circuit reactance passes through zero and takes on a small inductive value. However, it will be established later that a low-capacitive circuit reactance is desirable at the third harmonic and perhaps even at the second harmonic. It will be clear that a low value of inductive circuit reactance is not desirable.

Results similar to those in Fig. 2 were obtained with several experimental amplifier designs having a wide range of values for L_1, L_2, C_1, C_2, Z_0, and l [14]. Bandwidths as high as 10 percent and efficiencies as high as 20 percent were occasionally obtained. In each case, the results of the point-by-point output power measurements coupled with impedance measurements consistently showed broad-band TRAPATT amplification over a range of fundamental frequencies having high values of reactance at the harmonics.

Measured amplifier impedance data reported previously [13] for the configuration of Fig. 1 also show very high reactance at the third (−1207-Ω) harmonic as well as at the fourth (−418-Ω) and fifth (−177-Ω) harmonic, and a relatively lower reactance at the second (−35-Ω) harmonic. The reactance slope at the second harmonic in [13] is about the same as in Fig. 2 of the present paper. Furthermore, amplifier gain and efficiency reported in [13] are essentially the same as reported in [14]. Finally, it should be noted that no reliable data have ever been presented showing that the configuration of Fig. 1 with large L_1 can provide a low-capacitive reactance at the second or third harmonic over a broad frequency band.

It is seen, therefore, that reasonably broad-band TRAPATT amplification has been obtained by at least two groups of workers using the configuration of Fig. 1 with high harmonic impedances. These broad-band results can be understood by noting that when the harmonic impedances are higher than a certain value[3], even a rapid and relatively high fractional variation of the harmonic impedances with frequency will have little effect on amplification at the fundamental frequency.

Some simple conclusions can be drawn when the circuit provides a high impedance at each harmonic. First, it is clear that under these conditions, the diode impedance must also be high, implying that plasma inductance L is nearly in parallel resonance with the depletion layer

[3]This value of impedance at harmonic F_n is perhaps several times the magnitude of the reactance $1/(2\pi F_n C_0)$ of the depletion-layer capacitance C_0 at breakdown.

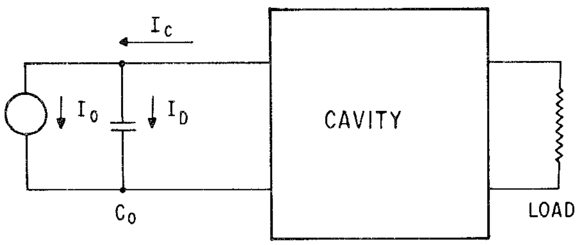

Fig. 3. Approximate representation of TRAPATT diode.

capacitance C_0 at each harmonic. In this case, L varies inversely with the square of the harmonic number n, in sharp contrast with the simple model of the Evans circuit which is characterized by a nearly constant L. Next, consider the approximate equivalent circuit shown in Fig. 3[4] in which the diode is represented by a capacitor C_0 and an impulse generator having current I_0 equal to the diode conduction current [4], [8]. If the cavity reactances are high at the harmonics compared to the reactance of C_0, it follows that the cavity current I_c at the harmonics is small compared to the displacement current I_D flowing in C_0. On the other hand, the cavity current at the fundamental can be high due to the relatively low impedance at F_1. In this case, it is concluded that the total diode current has relatively low harmonic content, having a nearly sinusoidal waveform corresponding to the fundamental frequency. Moreover, as in the relaxation mode [17], [18], the diode voltage waveform [8] can be expected to take on a roughly triangular shape (as a result of the periodic charging and discharging of the depletion capacitance) with a fundamental sine-wave component superimposed. Therefore, the voltage and current waveforms obtained with high harmonic reactances differ significantly from the sharply spiked waveforms of the Evans oscillator circuit. On this basis, it seems reasonable to expect inherently lower efficiency for an amplifier having high harmonic reactances. It should be noted, however, that the diode voltage waveform for the broad-band amplifier must also have a trigger portion near the end of the period in order to initiate a new avalanche. Therefore, as stated previously, time-delayed triggering is not prevented by the dispersion introduced by L_1; rather, it is only made less effective, i.e., the trigger having reduced rate of rise results in a weaker plasma and correspondingly reduced efficiency.

Computer calculations by Parker [19] have shown that a steady-state TRAPATT-mode solution for the device/circuit interaction can indeed be obtained with a direct current plus only a fundamental frequency current component. Moreover, these calculations show that the efficiency under these conditions is lower than with a larger number of harmonics, in agreement with the experimental results of Evans [4].

It seems clear that high harmonic impedances are not optimum for high efficiency. Instead, it is clear that this merely represents a simple broad-band harmonic condition readily obtainable in practice by using a large lead inductance in the configuration of Fig. 1. A better impedance characteristic obtainable with coupled microstrip circuits will be discussed in the next section.

IV. Coupled-Microstrip Amplifier Circuits

Fig. 4 shows calculated values of the impedance $R+jX$ for a four-port network comprised of a pair of idealized coupled lines having two ports with open-circuit terminations. This configuration is easier to fabricate in microstrip form than others requiring short-circuit terminations. The calculations were based on the four-port impedance matrix of the coupled microstriplines derived from a consideration of the even and odd characteristic modes [20]. For simplicity, the effects of frequency dispersion, open-end capacitances, and the difference between even- and odd-mode propagation constants were neglected in the preliminary calculations of Fig. 4. Note that with these approximations, the impedance characteristics are exact periodic functions of frequency. However, in spite of these approximations, the calculations provide a useful qualitative description of the impedance characteristics of a pair of coupled microstriplines.

It was desired to obtain relatively high capacitive reactance at the second harmonic[5] and relatively low capacitive reactances at the fundamental and third harmonics. Under these conditions, substantial current flow can occur at the fundamental and third harmonic, and improved device waveforms were expected.[6] The data in Fig. 4 show that relatively low capacitive reactances can be obtained within a fundamental frequency band lying just below 3 GHz and within a corresponding third-harmonic band just below 9 GHz. However, the reactance X_2 at the second harmonic F_2 is capacitive only for $F_2 > 6$ GHz. Therefore, it would be desirable to move the reactance pole at 6 GHz downward in frequency relative to the reactance zeros at 3 and 9 GHz. Fortunately, this can be accomplished simply by adding a shunt capacitor at port 2. This tends to lower significantly the frequencies at which poles occur but has much less effect near frequencies at which reactance zeros occur. However, it was clear that in order to take full advantage of this possibility, one must employ a diode mount having ultralow series lead inductance. This is necessary, because a series inductance

[4]It has been suggested that this equivalent circuit should also include a plasma inductance in shunt with C_0. However, this would only represent additional diode conduction current and would be redundant, since the current generator is a completely general representation of the conduction current. Furthermore, as noted above, the plasma inductance for a broad-band amplifier can have grossly different frequency variation than for a narrow-band Evans circuit.

[5]At this point it was felt that in order to produce a satisfactory voltage waveform, the broad-band amplifier circuit having a smaller number of important harmonics should provide a relatively larger reactance at the second harmonic than is provided by the high-efficiency Evans circuit characterized by a negative inductance [8].

[6]The reader is referred to the discussion of TRAPATT waveforms given by Clorfeine [1].

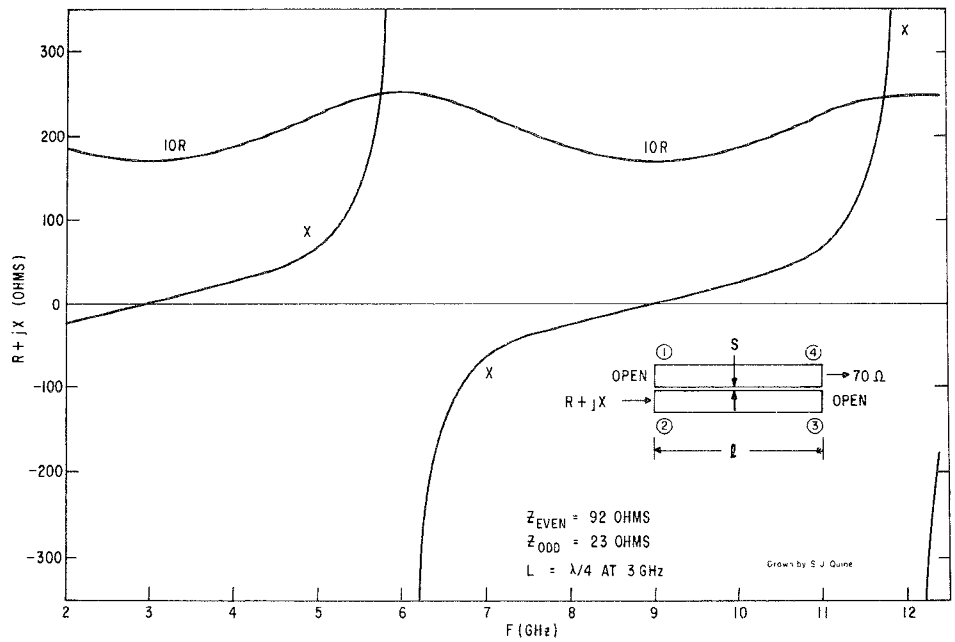

Fig. 4. Calculated impedance for idealized quarter-wave coupled lines.

at port 2 has the effect of lowering the frequencies at which a reactance zero occurs with no effect on the frequency at which a pole occurs. The shunt capacitance required to move the reactance pole to a lower frequency can be provided by an open stub at port 2. If the length of this stub is made approximately a quarter wavelength at the third harmonic, considerable reduction should be expected in the value of the series resistance R and in the power dissipated at the third harmonic. With open circuits on ports 1 and 3, the transmission between ports 2 and 4 is inherently low in the second-harmonic region for all values of the gap S. Thus, with the open stub at port 2, the gap S has a significant effect on the impedance at the chip only in the fundamental region and affects primarily the resistive component.

Detailed calculations [14] have verified that a practical coupled-microstrip circuit (with properly adjusted parameters) can provide the desired impedance characteristics at the diode chip. These calculations employed approximate calculated values for the elements of the equivalent circuit of the diode mount and the open-end capacitances at ports 1 and 3. The calculated values of the even- and odd-mode impedance and propagation constants of the coupled microstriplines [20], [21] were corrected for the small effects of dispersion [22]–[24]. Fig. 5 shows a typical set of calculated impedance values.

Tests were performed with coupled-microstrip amplifier circuits fabricated on 0.062-in-thick teflon-glass substrates having a relative dielectric constant of 2.54. The diodes employed were similar to those employed with the coaxial circuits. However, in order to obtain as small a diode lead inductance as possible, the diode chip was soldered to a copper stud which was then threaded through a hole in the microstrip ground plate until the chip contacted the microstrip conductor. The circuit was fixed tuned except for a small conducting tab placed over the diode (at port 2 of Fig. 4) to obtain the effect of the shunt capacitance discussed in the preceding paragraph. Results were poor without the tab adjusted properly, indicating agreement with the theory. Although not fully optimized, the coupled-microstrip circuits provided results which were generally better than obtained with the coaxial-line circuits. The highest efficiency was 23 percent and was obtained with 53-W output and 6.0-dB gain at 2.6 GHz. The bandwidth in this case was approximately 10 percent, but bias-circuit oscillations[7] occurred at several points within the amplification band. Also significant was the fact that the noise in the detected RF output power waveforms was considerably lower with the coupled-microstrip circuits, indicating more stable TRAPATT operation. These results tend to support the conclusion that a low impedance at the third harmonic can result in improved performance.

Amplifiers employing coupled-microstrip circuits with power extraction at the second harmonic were described in [11], [25], [26], and [28]. Relatively large-area diodes were employed in [25] and [26], and the measured circuit impedances presented to the diode at the second, third, and fourth harmonics were considerably larger than the reactance X_0 of the depletion layer capacitance. Under these conditions, the reactive voltage waveforms for these second-harmonic-extraction amplifiers must be similar to those for the fundamental-extraction amplifiers discussed in Section III of this paper, and relatively low efficiency can also be expected. It was suggested [25], [26] that the inductive harmonic impedances employed were required in order to prevent premature avalanche. In contrast to these results, oscillators of the Evans type using harmonic

[7]The bias oscillation problem is discussed in [25], [29], and [30].

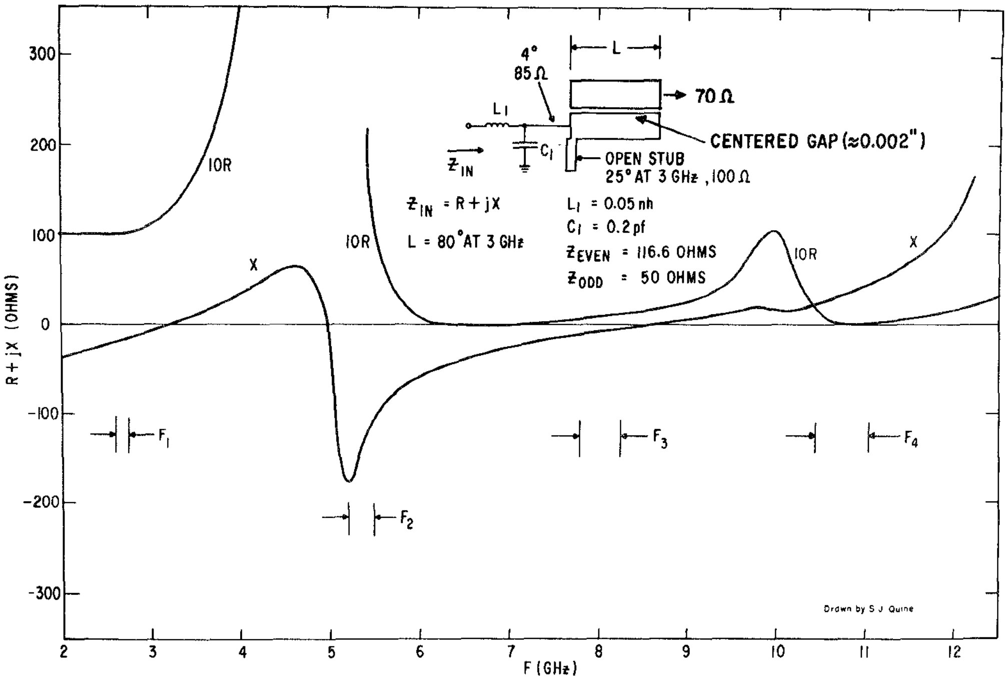

Fig. 5. Calculated circuit impedance at chip for practical coupled-microstrip amplifier.

extraction with small-area diodes and having relatively high efficiency (35 percent with fundamental power of 9.66 W at 1.01 GHz, 24 percent with second-harmonic power of 5.78 W at 3.02 GHz) have been reported [27].

A very flexible lumped-element broad-band amplifier circuit which produced excellent results with second-harmonic extraction is described in [28]. This circuit provided relatively independent tuning of the fundamental and of the second and third harmonics. The data in Fig. 4 of [28] showing high inductive impedance at the third harmonic were obtained with an amplifier having 15.8-percent efficiency[8]. An efficiency of over 28 percent is also reported for this same circuit in [28] but with different tuning conditions.

In summary, it is noted that all circuits which have been reported with power extraction at the second harmonic and which have employed high values of harmonic impedances have yielded efficiencies of less than 20 percent. The impedance characteristics of circuits for second-harmonic extraction which have yielded efficiencies greater than 20 percent ([27], [28]) have not been reported.

Broad-band amplifiers employing coupled-microstrip circuits similar to those discussed in the present paper are described in [29] and [30]. Power was extracted at the fundamental frequency of 3.3 GHz, with efficiencies usually exceeding 25 percent and as high as 30 percent. These are among the highest efficiencies reported to date for broad-band TRAPATT amplifiers. Measured circuit impedances in the third-harmonic region for these fixed-tuned amplifiers were relatively low and inductive (being comparable to X_0). However, the measured impedance values were referenced at the package output, and it is not clear what impedance values were presented to the chip.

The above results obtained with the coupled-microstrip circuits indicate that a low impedance is desirable at the third harmonic. However, these data cannot be considered accurate enough to also establish whether the third-harmonic impedance should be capacitive or inductive. However, the impedance calculations do show that the coupled-microstrip circuit parameters can be adjusted to obtain either a low capacitive or low inductive reactance at the third harmonic, i.e., whichever is optimum.

The results of computer calculations such as those by Parker can help resolve the question concerning the sign of the third-harmonic reactance. Using a semianalytical solution for a p-i-n diode, Parker [31] showed that capacitive circuit reactances at F_1 and F_3 result in high efficiency. The circuit impedance at F_3 [31, Table V] is low, being on the same order as the impedance at F_1. Note that the impedance at F_2 is infinite. Parker's calculations indicate the importance, in general, of employing a large number of harmonics to obtain high efficiency and also indicate that some degree of power dissipation at the harmonics may improve the efficiency of power extraction at the fundamental. Although Parker's steady-state solutions were obtained specifically for oscillators, it can be correctly argued that the same steady-state solutions apply for amplifiers [16]. Trew [32] has presented device waveforms for a device current comprised only of fundamental and third-harmonic components. Fourier analysis of the waveforms in Fig. 6 of [32] reveals that the circuit reactance is capacitive at F_1 but inductive at F_3. Furthermore, the impedance at F_3 is approximately three times the

[8]The writer is indebted to A. Clorfeine and A. Rosen for providing this information.

value at F_1. Significantly, the efficiency is low. The computer calculations, therefore, tend to support the conclusion that a low-capacitive circuit impedance is desirable at F_3.

V. Conclusions

It was shown that an amplifier employing a package with large lead inductance can have wide bandwidths for the harmonic impedances. However, arguments were presented indicating that the high harmonic impedances characteristic of these circuits are not optimum but result in relatively low efficiency.

Design considerations for coupled-microstrip circuits were also discussed. These circuits can provide a low-capacitive reactance at the third harmonic which may yield a better current waveform. Furthermore, the slow variation of impedance with frequency can result in broader amplification bandwidths.

It is concluded that high efficiency with a broad-band amplifier can result when the circuit provides capacitive reactances at the fundamental and at several harmonic frequencies. This suggests that an optimum broad-band amplifier circuit may be one which provides these capacitive reactances over the widest bandwidth for the largest number of harmonics.

The above conclusions concerning efficiency were based on the results of experiments with amplifiers employing fundamental power extraction. However, there is some basis for believing that these same conclusions also apply for amplifiers employing second-harmonic power extraction.

TRAPATT amplifiers are still seriously proposed for use as microwave power sources for systems employing phased-array antennas and are, in fact, considered by many as the best solution economically for *S*-band radar applications [33], [34]. Significant progress has been reported recently by the Naval Research Laboratory [35] on the development of devices capable of achieving wide pulse and high-duty factor operation with high-efficiency broad-band circuits [28], [30]. It may be feasible to overcome the pulse width and duty factor limitations by cooling the TRAPATT device. In this connection, A. Rosen [36] has suggested that Peltier cooling techniques are especially convenient in experimental studies in that the TRAPATT circuit can be uncovered and adjusted for optimum performance under actual cooled conditions.

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Experimental Attenuation of Rectangular Waveguides at Millimeter Wavelengths

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***Abstract*—The experimental values of attenuation of commercially available rectangular waveguides were determined at frequencies between 25 and 200 GHz with emphasis on high accuracy. They were compared with the theoretical values computed from the dc conductivities, taking into consideration temperature effects, work hardening, size effects, surface roughness, and a room-temperature anomaly of the skin effect. A new way to express the excess attenuation due to these effects was formulated.**

Excess ratios of attenuation of coin-silver waveguides were found to be well below the values used in engineering in the past. They can satisfactorily be explained by surface roughness. The normalized excess attenuations of copper guides are higher than those of guides made of silver but lower than cited in the literature.

I. Introduction

SEARCHING for reliable data on the attenuation of rectangular waveguides for the millimeter-wave region frequently becomes a frustrating endeavor. One usually finds that the theoretical values listed in handbooks [1], [2] are considerably lower than those observed in practice. Also, these values are valid for pure silver whereas the commercially available waveguides are made primarily of coin silver (90-percent Ag, 10-percent Cu), which yields theoretical values of attenuation increased by a factor somewhere around 1.25. Catalogs of companies selling waveguides usually contain the same values [3], [4] or data reprinted from a book on plasma diagnostics [5]–[7] with experimental attenuation values increased by about a factor of 2 above the theoretical values. Results of rather thorough investigations were published by Benson [8] with emphasis on guide materials and surface preparations such as chemical polishing and annealing. His results confirm the previously observed increases of the attenuation and include data for the range of 20 to 140 GHz. The increases are due to excess losses, which the author attributes to surface roughness and minute surface irregularities. The data have an uncertainty of about ± 7 percent and some of them show variations of up to ± 20 percent. Some of Benson's results are in disagreement with those published earlier [9]. The review of these and other publications made it apparent that additional research to furnish more reliable data was highly desirable.

While studying the surface characteristics of metals, particularly copper, the author found [10] that a major contribution to the excess losses of copper at millimeter wavelengths is caused by an anomaly of the skin effect at room temperature. This, in turn, causes increased values of attenuation. The increases are larger than those caused by the effects of surface roughness and work hardening. In order to correlate the results of the original surface-resistance experiments with waveguide data and to generate reliable data, careful experiments were conducted. These experiments resulted in accurate values of the attenuation of standard rectangular waveguides in the millimeter-wave region. Subsequently, the obtained data were evaluated in great detail. The results of these efforts are presented in this paper.

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