Analysis and Design of an X-Band Actively Compensated IMPATT Diode Amplifier

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Abstract—**This paper shows how broad-banding of an IMPATT diode amplifier can be achieved using a circuit technique known as active reactance compensation. Theoretical analysis and experimental results both show that the gain-bandwidth products of an uncompensated IMPATT amplifier improves from $G^{1/2}B=k$ to $G^{1/4}B=2k$ (where k is a constant) for the same amplifier actively conpensated. The measured 3-dB bandwidth of 230 MHz for a 9.0-GHz amplifier having a gain of 10 dB is improved to 780 MHz at the same gain.**

I. Introduction

Although much progress has been made with the power capability of IMPATT amplifiers [1]–[6], the bandwidth achievable with these devices is still insufficient for many system applications, especially in the lower power preamplifier stages of a cascaded arrangement.

The purpose of this paper is to report the application of a new circuit technique to the problem of broad-banding IMPATT diode amplifiers. This circuit technique is called active reactance compensation. The technique of passive compensation previously applied to improve the gain-bandwidth product of parametric amplifiers [7] is well known. Humphreys [8] has calculated the form of the parametric amplifier gain-bandwidth product using the passive compensation technique and shown that under a high-gain condition, the gain-bandwidth product improves from

$$G^{1/2}BW=k$$

to

$$G^{1/4}BW=k.$$

Aitchison and Williams [9], [10] suggested a further extension of the technique of passive reactance compensation by using a duplicate parametric amplifier for the compensatory circuit instead of a passive circuit. Since the compensatory circuit is an active circuit, the technique is called active reactance compensation. The latter report that a 4.0-GHz parametric amplifier bandwidth of 75 MHz at a gain of 20 dB becomes greater than 500 MHz at the same gain. This result exceeds slightly the $G^{1/4}BW=2k$ prediction. Downing and Watson [11] have reported a theoretical analysis and experimental results on a 4.0-GHz actively compensated parametric amplifier. They assumed a high-gain approximation and showed that the gain-bandwidth product of an uncompensated parametric amplifier improved from $G^{1/2}BW=k$ to only $G^{1/4}BW=0.84k$ on actively compensating. This improvement factor is considerably smaller than that predicted by Aitchison and Williams. Their experimental work produced an 1.69 times improvement factor.

In this paper we derive expressions for the gain-bandwidth products of the uncompensated and actively compensated IMPATT amplifiers. Experimental results for uncompensated and actively compensated 9.0-GHz coaxial IMPATT diode amplifiers are also given and compared with theoretical predictions.

II. Principle of Active Reactance Compensation

Fig. 1(a) shows a shunt resonant circuit with negative conductance producing an amplified version of the signal applied to port 1 in the load at port 3. The susceptance variation with frequency for this circuit is shown in Fig. 1(a) as B_p. The addition of a series resonant circuit at the same frequency in the plane AA produces a susceptance frequency curve shown as B_s. Since this has a region of negative slope, it is possible to arrange that the total susceptance of the resultant circuit has a zero dB/df at the center frequency as shown in Fig. 1(a). This technique is known as passive compensation.

Under a given RF and dc condition, the input admittance of an IMPATT amplifier can also be considered a shunt circuit resonant at a center frequency f_0. A second amplifier placed a quarter-wavelength away will appear as a series resonant circuit in series with the first amplifier, as shown in Fig. 1(b). The two amplifiers can either be connected in the series configuration (Fig. 1(b)) as reported previously [12] or alternatively in a parallel configuration as shown in Fig. 1(c). Both arrangements are examples of active reactance compensation, since the two amplifiers compensate each other.

The gain of an IMPATT amplifier changes owing to the effect of change in both the input susceptance and input negative conductance of the amplifier with frequency. The technique of active compensation reduces these changes by using the admittance changes in one amplifier to compensate for the admittance changes in the second

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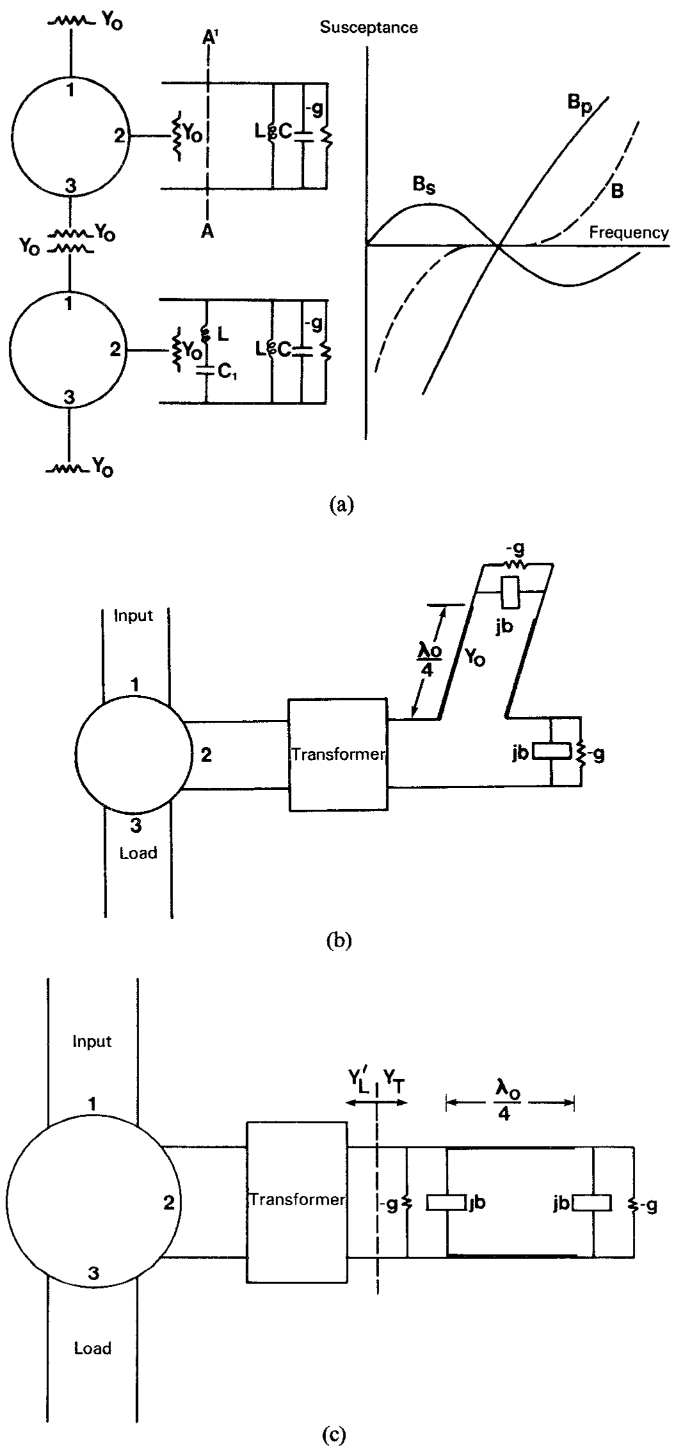


Fig. 1. (a) Susceptance-frequency curves for parallel resonant circuit B_p, series resonant circuit B_s, and the combined circuit B. (b) Schematic of a series actively compensated amplifier. (c) Schematic arrangement of a parallel-coupled actively compensated IMPATT amplifier.

amplifier and vice versa. In this paper the parallel configuration is considered (Fig. 1(c)) in which the two IMPATT amplifiers are placed at the ends of a quarter-wavelength line of characteristic admittance Y_0. Thus both the input susceptance and input conductance variations with frequency of the first amplifier are compensated by the transformed variations of the second amplifier and vice versa. The admittance Y_0 of the quarter-wavelength line is computed for optimum bandwidth.

III. Theoretical Analysis

In the analysis presented in this paper we shall derive the circuit design conditions for, and bandwidth resulting from, a ripple-free gain-frequency response for the IMPATT amplifier. Under a constant RF input and dc condition, both the input conductance and input susceptance of the IMPATT amplifier can be expressed by polynomial functions of Δf, the frequency deviation from the band center frequency f_0, given by

$$f = f_0 \pm \Delta f.$$

A. Input Admittance of the Uncompensated Amplifier

The input admittance of an IMPATT amplifier, operating at a center frequency f_0, can be written as

$$Y_{\text{in}} = -g(\Delta f) + jb(\Delta f) \tag{1}$$

where $g(\Delta f)$ and $b(\Delta f)$ are the input conductance and input susceptance of the amplifier, respectively.

The measured values of the input conductance and input susceptance of an uncompensated 9.0-GHz IMPATT diode amplifier are shown in Fig. 2 as functions of frequency. From this it can be seen that linear variation of both the input conductance and input susceptance is a good approximation, since in a practical uncompensated amplifier, Δf is of the order of $f_0/50$. Thus the input admittance of the uncompensated IMPATT amplifier can be written as

$$Y_{\text{in}} = -(g_0 - K_1 \Delta f) + jK_2 \Delta f \tag{2}$$

where g_0 is the input conductance at the band center frequency, $K_1 = \delta g/\delta f$ and $K_2 = \delta b/\delta f$.

B. Gain-Bandwidth Product of the Uncompensated Amplifier

The power gain $G(\Delta f)$ of the reflection IMPATT diode amplifier is given by

$$G(\Delta f) = \left| \frac{Y_L - Y_{\text{in}}^*}{Y_L + Y_{\text{in}}} \right|^2 \tag{3}$$

where Y_L is the load admittance presented to the active device, and Y_{in} is the input admittance of the amplifier (Fig. 3).

We now assume that the load Y_L is real and independent of frequency and is equal to g_L. Using 2, the power gain $G(\Delta f)$ is given by

$$G(\Delta f) = \frac{[g_L + (g_0 - K_1 \Delta f)]^2 + (K_2 \Delta f)^2}{[g_L - (g_0 - K_1 \Delta f)]^2 + (K_2 \Delta f)^2}. \tag{4}$$

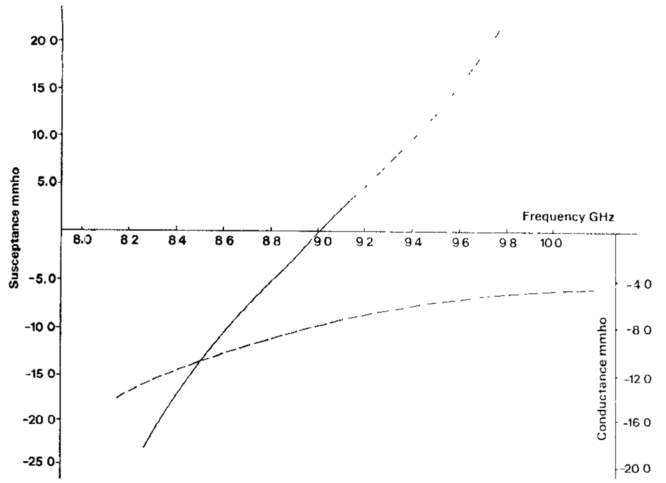


Fig. 2. Measured input conductance (-----) and susceptance (——) as functions of frequency for the encapsulated diode amplifier.

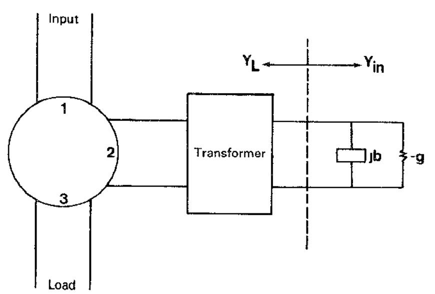


Fig. 3. An uncompensated IMPATT amplifier.

The band center frequency gain G is

$$G=\frac{[g_L+g_0]^2}{[g_L-g_0]^2}. \quad (5)$$

This can be written as

$$\frac{G^{1/2}+1}{G^{1/2}-1}=\frac{g_L}{g_0}. \quad (6)$$

We define the amplifier bandwidth such that at frequency $f=f_0\pm\Delta f$, the power gain $G(\Delta f)$ is

$$G(\Delta f)=G/2. \quad (7)$$

Substituting (5) into (4) and using (7) we get

$$[(g_L+g_0-K_1\Delta f)]^2+(K_2\Delta f)^2 = \frac{G}{2}\left[(g_L-g_0+K_1\Delta f)^2+(K_2\Delta f)^2\right]. \quad (8)$$

Assuming that $G\gg 2$, the above expression becomes

$$\Delta f^2\cdot\frac{K_2^2}{2}\left(1+\frac{K_1^2}{K_2^2}\right)G+\Delta f\cdot 2g_0K_1G^{1/2}-2g_0^2=0.$$

The solution of the above equation is given by

$$G^{1/2}\Delta f=\frac{-2g_0K_1\pm 2K_2g_0\sqrt{1+2\left(\frac{K_1}{K_2}\right)^2}}{K_2\left[1+\left(\frac{K_1}{K_2}\right)^2\right]}.$$

From the experimental investigation of our 9.0-GHz IMPATT amplifier (see Fig. 2) we find that $K_2\gg K_1$; therefore, the gain-bandwidth product of the uncompensated amplifier is

$$G^{1/2}BW=\frac{4g_0}{K_2}\left[1-\frac{K_1}{K_2\left\{1+\left(\frac{K_1}{K_2}\right)^2\right\}}\right]. \quad (9)$$

Thus the gain-bandwidth product of the uncompensated amplifier is a function of the input conductance at the band center frequency, the rate of change of input conductance with frequency, and rate of change of input susceptance with frequency. If the input conductance $g(\Delta f)$ is frequency independent, we get

$$G^{1/2}BW=\frac{4g_0}{K_2}. \quad (10)$$

C. Input Admittance of the Actively Compensated Amplifier

In the parallel coupled configuration of the actively compensated amplifier (Fig. 1(c)), the two identical amplifiers are placed at the ends of a quarter-wavelength line of characteristic admittance Y_0. The total admittance of the actively compensated amplifier is

$$Y_T=-g\left(1+\frac{Y_0^2}{g^2+b^2}\right)+jb\left(1-\frac{Y_0^2}{g^2+b^2}\right). \quad (11)$$

For a ripple-free gain-frequency response, it can be shown that for optimum bandwidth the characteristic admittance Y_0 of the quarter-wavelength line is equal to the modulus of the input conductance of the uncompensated amplifier at the band center frequency, i.e., $Y_0=g_0$. Thus (11) becomes

$$Y_T=-g\left(1+\frac{g_0^2}{g^2+b^2}\right)+jb\left(1-\frac{g_0^2}{g^2+b^2}\right) = -g_T+jb_T. \quad (12)$$

From the above equation, the total input conductance g_T of the actively compensated amplifier is

$$g_T=g\left(1+\frac{g_0^2}{g^2+b^2}\right)=g+\frac{gg_0^2}{g^2+b^2}. \quad (13)$$

We expand the expression for g_T in terms of frequency deviation Δf by writing g and b as polynomial functions of Δf.

$$g=g_0+\frac{\delta g}{\delta f}\Delta f+\frac{1}{2!}\frac{\delta^2 g}{\delta f^2}\Delta f^2+\frac{1}{3!}\frac{\delta^3 g}{\delta f^3}\Delta f^3+\cdots. \quad (14)$$

Similarly,

$$b=0+\frac{\delta b}{\delta f}\Delta f+\frac{1}{2!}\frac{\delta^2 b}{\delta f^2}\Delta f^2+\frac{1}{3!}\frac{\delta^3 b}{\delta f^3}\Delta f^3+\cdots. \quad (15)$$

Expanding (13) gives

$$g_T=g+g_0+\frac{\delta}{\delta f}\left(\frac{gg_0^2}{g^2+b^2}\right)\Delta f+\frac{1}{2!}\frac{\delta^2}{\delta f^2}\left(\frac{gg_0^2}{g^2+b^2}\right)\Delta f^2+\frac{1}{3!}\frac{\delta^3}{\delta f^3}\left(\frac{gg_0^2}{g^2+b^2}\right)\Delta f^3+\cdots \quad (16)$$

so that

$$g_T=2g_0+\frac{(K_1\Delta f)^2}{g_0}-\frac{(K_2\Delta f)^2}{g_0}\cdot\left[1-\left(\frac{K_2\Delta f}{g_0}\right)^2+\frac{2}{5}\left(\frac{K_2\Delta f}{g_0}\right)^4+\cdots\right]. \quad (17)$$

Similarly, (12) becomes

$$b_T=b-\frac{bg_0^2}{g^2+b^2}-\frac{\delta}{\delta f}\left(\frac{bg_0^2}{g^2+b^2}\right)\Delta f-\frac{1}{2!}\frac{\delta^2}{\delta f^2}\left(\frac{bg_0^2}{g^2+b^2}\right)\Delta f^2+\cdots. \quad (18)$$

Substituting (14) and (15) into (18) gives

$$b_T=-\frac{2}{g_0}K_1K_2(\Delta f)^2+\frac{1}{g_0^2}(K_2\Delta f)^3+\cdots. \quad (19)$$

D. Gain-Bandwidth Product of the Actively Compensated Amplifier

The power gain $G'(\Delta f)$ of the actively compensated amplifier is given by

$$G'(\Delta f)=\left|\frac{Y_L'-Y_T^*}{Y_L'+Y_T}\right|$$

where Y_L' is the actively compensated amplifier load which is assumed to be real and independent of frequency, i.e., $Y_L'=g_L'$. This can be written as

$$G'(\Delta f)=\frac{(g_L'+g_T)^2+b_T^2}{(g_L'-g_T)^2+b_T^2}. \quad (20)$$

The band center gain of the two amplifiers is assumed to be identical. Therefore,

$$\frac{(g_L'+g_T)^2}{(g_L'-g_T)^2}=\frac{(g_L+g_0)^2}{(g_L-g_0)^2}. \quad (21)$$

Since $g_T=2g_0$, we have $g_L'=2g_L$. Substituting for g_T and b_T from (17) and (19) into (20), $G'(\Delta f)$ becomes

$$G'(\Delta f)=\left[2g_L+2g_0\left\{1+\frac{(K_1\Delta f)^2}{2g_0^2}-\frac{(K_2\Delta f)^2}{2g_0^2}\cdot\left(1-\frac{(K_2\Delta f)^2}{g_0^2}+\frac{2}{5}\frac{(K_2\Delta f)^4}{g_0^4}\right)\right\}\right]^2+\left[\frac{(K_2\Delta f)^3}{g_0^2}-\frac{2K_1K_2(\Delta f)^2}{g_0}\right]^2\Bigg/\left[2g_L-2g_0\left\{1+\frac{(K_1\Delta f)^2}{2g_0^2}-\frac{(K_2\Delta f)^2}{2g_0^2}\cdot\left(1-\frac{(K_2\Delta f)^2}{g_0^2}+\frac{2}{5}\frac{(K_2\Delta f)^4}{g_0^4}\right)\right\}\right]^2+\left[\frac{(K_2\Delta f)^3}{g_0^2}-\frac{2K_1K_2(\Delta f)^2}{g_0}\right]^2.$$

The magnitude of the compensated input susceptance is much smaller than the magnitude of the input conductance and can be neglected. Thus

$$G'(\Delta f)=\left[2g_L+2g_0\left\{1+\frac{(K_1\Delta f)^2}{2g_0^2}-\frac{(K_2\Delta f)^2}{2g_0^2}\cdot\left(1-\frac{(K_2\Delta f)^2}{g_0^2}+\frac{2}{5}\frac{(K_2\Delta f)^4}{g_0^4}\right)\right\}\right]^2\Bigg/\left[2g_L-2g_0\left\{1+\frac{(K_1\Delta f)^2}{2g_0^2}-\frac{(K_2\Delta f)^2}{2g_0^2}\cdot\left(1-\frac{(K_2\Delta f)^2}{g_0^2}+\frac{2}{5}\frac{(K_2\Delta f)^4}{g_0^4}\right)\right\}\right]^2. \quad (22)$$

We define the amplifier bandwidth by (7), so (22) becomes

$$\frac{G^{1/2}}{\sqrt{2}}\left(\frac{2g_L}{2g_0}-1\right)-\left(\frac{2g_L}{2g_0}+1\right)=\left[\frac{(K_1\Delta f)^2}{2g_0^2}-\frac{(K_2\Delta f)^2}{2g_0^2}\left(1-\frac{(K_2\Delta f)^2}{g_0^2}+\frac{2}{5}\frac{(K_2\Delta f)^4}{g_0^4}\right)\right]\cdot\left[\frac{G^{1/2}}{\sqrt{2}}+1\right].$$

We write ν for the factor

$$1-\frac{(K_2\Delta f)^2}{g_0^2}+\frac{2}{5}\frac{(K_2\Delta f)^4}{g_0^4}.$$

Substituting from (6) and assuming that $G^{1/2} \gg 2$, we get

$$G^{1/2}\Delta f^2 = \frac{4g_0^2(\sqrt{2}-1)}{\left[K_2^2\left(\nu - \frac{K_1^2}{K_2^2}\right)\right]}.$$

Therefore, the gain-bandwidth product of the actively compensated amplifier is given by

$$G^{1/4}BW_c = \frac{4g_0(\sqrt{2}-1)^{1/2}}{K_2\left[\nu - \frac{K_1^2}{K_2^2}\right]^{1/2}} \tag{23}$$

where BW_c is the compensated bandwidth. Combining (23) with (9) we get

$$G^{1/4}BW_c = 1.74G^{1/2}BW \tag{24}$$

where we have substituted an experimental value for ν obtained from Fig. 2. If $K_1 = 0$, (23) becomes

$$G^{1/4}BW_c = \frac{4g_0}{K_2\nu^{1/2}}(\sqrt{2}-1)^{1/2}. \tag{25}$$

Using (10) gives

$$G^{1/4}BW_c = 1.15G^{1/2}BW. \tag{26}$$

This shows an improvement factor of 1.15 in the appropriate gain-bandwidth product. This is smaller than when the input conductance is frequency dependent (see (24)).

IV. Experimental IMPATT Diode Amplifier with Active Reactance Compensation

A. Circuit

A 9.0-GHz coaxial amplifier was constructed using an encapsulated single-chip ML 4704 IMPATT diode. The IMPATT is a p^+-n-n^+ device, encapsulated in a standard S_4 package. The diode was end mounted in the coaxial circuit in such a way that the package was recessed partially into the ground plane, thereby reducing the mounted package inductance considerably. The impedance of the partially recessed diode was measured using the network analyzer at an RF input level of +10 dBm and is shown in Fig. 4. The amplifier was resonated at 9.0 GHz by adding a line resonator to the recessed diode. A maximally flat two-section transformer was used for impedance matching to 50 Ω.

To construct an actively compensated amplifier, a second identical amplifier was coupled to the first amplifier through a quarter-wavelength line of impedance Z_0. In theory, this could be done either in a series coupled form or a parallel coupled form. In a practical coaxial structure the parallel coupled form is easier and this was used.

In our experimental actively compensated arrangement two identical amplifiers were coupled in parallel through a 42-Ω quarter-wavelength line at their outputs. The characteristic impedance was selected for optimum bandwidth, and this impedance was approximately equal to the modulus of the input resistance of the uncompensated amplifier at the band center frequency. Fig. 5 shows a cross section of the arrangement. The first amplifier was a two-section transformer coupled circuit. The second amplifier had adjustable capacitance coupling so that an optimum gain-frequency response of the actively compensated circuit could be obtained.

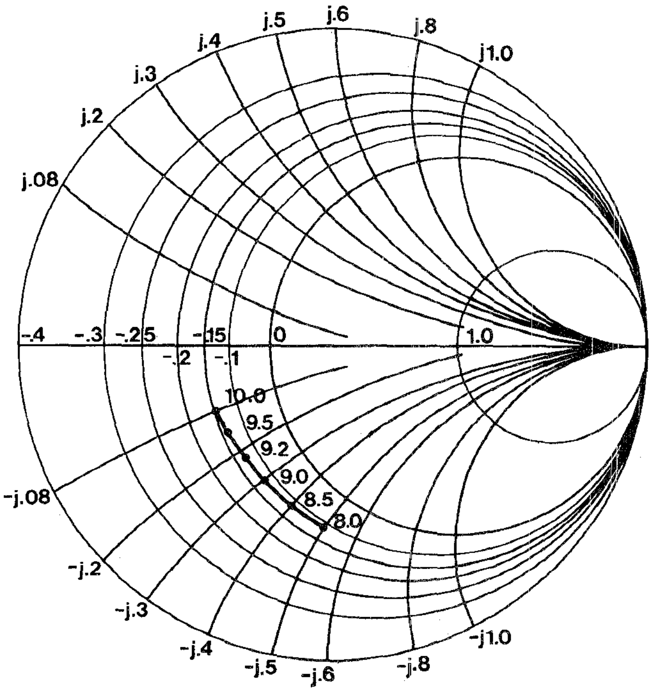


Fig. 4. Measured impedance of the encapsulated diode (ML4704) as a function of frequency, at a constant RF input (00 mW) and dc current.

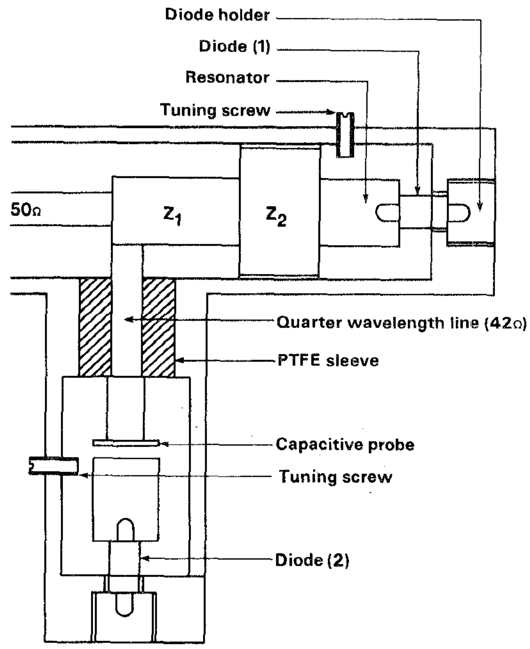
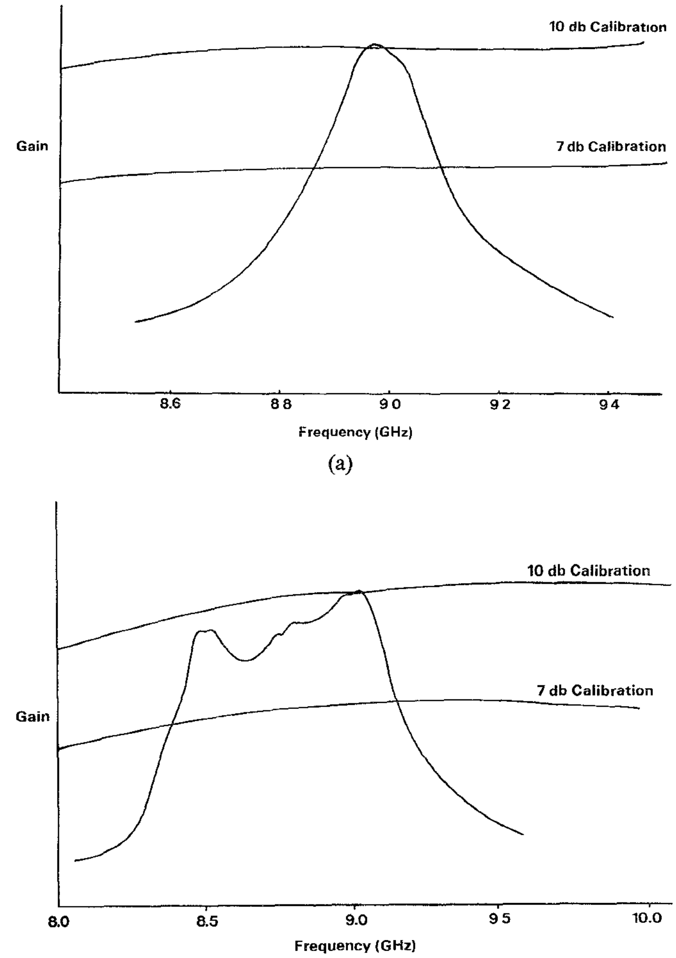
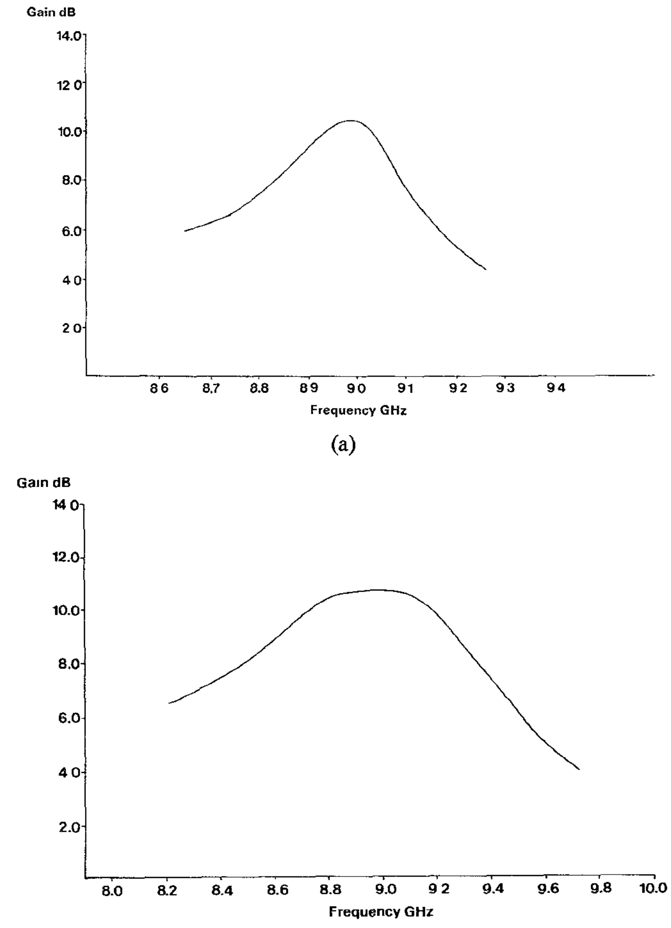


Fig. 5. Cross section of a coaxial IMPATT diode amplifier.



(b)

Fig. 6. (a) Measured gain-frequency response of the uncompensated amplifier. (b) Measured gain-frequency response of the actively compensated amplifier.



(b)

Fig. 7. (a) Computed gain-frequency response of the uncompensated amplifier. (b) Computed gain-frequency response of the actively compensated amplifier.

B. Results

The measured gain-frequency response of the 9.0-GHz uncompensated amplifier at an RF input power level of +10 dBm is shown in Fig. 6(a). At 10-dB gain, the measured 3-dB bandwidth is 230 MHz. Fig. 6(b) shows the measured gain-frequency response of the same amplifier actively compensated at the same gain, showing that the 3-dB bandwidth improves to 780 MHz. The gain-frequency response was computed, using the measured impedance values of the diode, shown in Fig. 4. At 10-dB gain, the computed gain-frequency response of the uncompensated amplifier (Fig. 7(a)) and the actively compensated amplifier (Fig. 7(b)) show that the 3-dB bandwidth improves from 290 to 960 MHz. The measured bandwidth improvement of 3.4 times is found to be slightly larger than the computed bandwidth improvement of 3.3 times, although the measured gain-bandwidth products of both the uncompensated and actively compensated amplifier are smaller than the computed values. The difference may be due to the computation assumption that the load impedance is real and constant over the operating bandwidth of the amplifier. Calculated values of the ratio $(G^{1/4}BW_c)/(G^{1/2}BW)$ from measurements and computation are 1.9 and 1.78, respectively. The analytical improvement factor is 1.74, which is in good agreement with the computed value.

Table I compares the analytical, computed and experimental performance.

TABLE I
ANALYTICAL, COMPUTED, AND EXPERIMENTAL GAIN-BANDWIDTH PRODUCT IMPROVEMENT

	Analytical	Computed	Experimental
$\frac{G^{\frac{1}{4}}\,BW_c}{G^{\frac{1}{2}}\,BW}$	1.75	1.78	1.9
$\frac{BW_c}{BW}$ at 10dB gain	3.1	3.3	3.4

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 broad-band 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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