Measurement of the Large-Signal Characteristics of Microwave Solid-State Devices Using an Injection-Locking Technique

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Abstract—**A method for the measurement of the dynamic admittance and susceptance of "negative-resistance" diodes is described. The device under test is allowed to oscillate in a microwave cavity, and operated as an injection-locked oscillator. Injected locking signals of the same order as the free-running output power of the oscillator are used. The dynamic conductance and susceptance of the device are obtained from the phase and amplitude response of the system.**

INTRODUCTION

Solid-state devices used for the generation and subsequent amplification of power at microwave frequencies are usually operated under large-signal conditions, in order that reasonable power outputs may be obtained. The proper design of the circuits associated with such a device requires that the large-signal device parameters be known.

Conventional test equipments, such as network analyzers, may be used to make large-signal measurements directly. In the case of the Gunn diode, or the tunnel diode, oscillations may occur if the device is placed directly in a coax or waveguide mount.

Such oscillations are usually restricted by including a resistance in parallel with the diode in the mount. It may not always be possible to stabilize Gunn devices in this manner.

In the case of the IMPATT diode, the "negative resistance" is low ($\sim$1–2 Ω) and appears in series with a large reactive component. In order that acceptable measurement accuracy be achieved, an impedance transforming section may have to be included between the diode and the measuring system.

The injection-locking technique described allows the diode to be operated in a microwave cavity under typical operating conditions while the measurements are carried out.

Fig. 1 shows a typical injection-locking system as used with a microwave "negative-resistance" device. The locking and output signals are separated by a circulator. The conditions for locking in such a system have been considered by Quine [1] who used the simplified equivalent circuit of Fig. 2.

The active device is represented by a voltage dependent conductance in parallel with a susceptance due to C_0. C_0 is assumed to be constant. This assumption is reasonable for small-signal operation, but is certainly not true for large-signal conditions. However, the assumption enables some useful predictions to be made of the general properties of locked oscillators and, as will be seen later, the changes that occur in the susceptance in practice, under large-signal conditions, can be derived from the experimentally measured locking characteristics. The cavity and its coupling system are represented by the inductor and transformer, respectively. When locking occurs, only one frequency can exist in the system, and thus the impedance presented to the load by the diode and cavity, via the coupling system, must satisfy the normal transmission line voltage equations. Thus the impedance "seen" by the diode and the resonant circuit is given by

$$G_{\mathrm{IN}} + jB_{\mathrm{IN}} = \frac{G_0}{n^2}\left[\frac{1 - a^2}{1 + a^2 + 2a\cos\Theta}\right] + j\,\frac{G_0}{n^2}\left[\frac{2a\sin\Theta}{1 + a^2 + 2a\cos\Theta}\right] \tag{1}$$

the terms being defined in Fig. 2. In the steady-state condition $(G_{\mathrm{IN}} - G_D) = 0$ and the total susceptance must be zero.

If the difference between the angular frequency of the locking signal (ω_L) and the angular frequency of the free-running oscillator

Manuscript received May 2, 1974; revised August 28, 1974. This work was supported by the Science Research Council.
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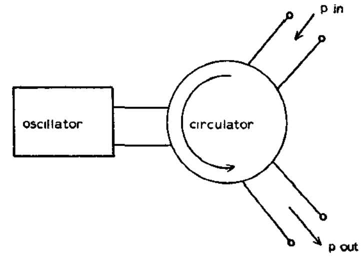

Fig. 1. Typical injection-locked system.

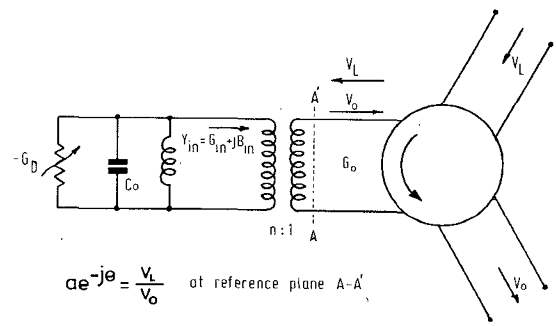

Fig. 2. Equivalent circuit used by Quine.

(ω_0) is small, then a simplified expression may be used for the change of susceptance with frequency for the diode and its resonant circuit. Making this assumption and equating the imaginary parts of (1) gives [2]

$$\frac{G_0}{n^2}\left[\frac{2a\sin\Theta}{1 + a^2 + 2a\cos\Theta}\right] = -2\omega_0\frac{C_0\Delta_\omega}{\omega_0} \tag{2}$$

where

$$\Delta_\omega = \omega_L - \omega_0 = 2\pi(F_L - F_0).$$

In most conventional injection-locking systems, the locking power is very small compared with the free-running output power and hence $a \ll 1.0$. Quine uses this assumption to obtain the expression

$$\Theta \simeq \sin^{-1}\left[-Q\cdot\ \frac{\Delta_\omega}{\omega_0}\cdot\ \frac{1}{a}\right] \tag{3}$$

where

$$Q = \frac{\omega_0 C_0 n^2}{G_0}$$

which is similar to that obtained by Adler [3] for a feedback oscillator. In the present case, "a" takes values up to unity (locking signal power = output power) and an exact solution then yields [2]

$$\Theta = \sin^{-1}\left[\frac{k}{(1+k^2)^{1/2}}\cdot\frac{1}{a}\right] + \sin^{-1}\left[\frac{k}{(1+k^2)^{1/2}}\right] \tag{4}$$

where

$$k = \frac{2C_0 n^2\Delta\omega}{G_0} = 2\cdot Q\cdot(\Delta\omega/\omega_0)$$

which is similar to that obtained by Paciorek [4] for a feedback oscillator.

The curves of Θ against "a" can be more conveniently expressed as Θ versus power gain in decibels. The theoretical curves for the large-signal case take the form shown in Fig. 3. The curves for the small-signal case (Quine's theory) are similar but the maximum value of Θ is restricted to $\pm 90°$.

The theoretical curves are symmetrical and their shape is independent of the sign of ΔF for both the small-signal and large-signal case.

The RF voltage across the diode terminals will increase, as the magnitude of the locking signal is increased from a low value until, when the locking gain becomes zero, the diode presents zero conductance to the circuit and contributes no power to the output. Observation of the oscillator output power as the locking power is being varied and also the phase and gain of the system enables the relative change in RF voltage across the device to be evaluated. Equation (1) shows that at any time, the conductance and susceptance of the diode are related to the coupling ratio of the oscillator, and to the phase and gain of the system. Provided that these can be established the diode conductance and susceptance may be evaluated. If one is interested only in the relative changes that occur in these parameters due, say, to a change of dc bias, or temperature, then the coupling ratio need not be evaluated. The gain may be measured directly (using a network analyzer) and the relative phase measured between the two ports of the circulator. In order that the measured phase values may be substituted into (1), some correction must be introduced to account for the finite length of any measuring system. The change in the diode conductance may be found from the real part of (1).

In practice, the change in the RF voltage across the diode also produces a change in the diode susceptance. The change in the total circuit susceptance (that of diode plus cavity) may be found from the imaginary part of (1).

The result of the susceptance change is to make the large-signal phase-gain curves asymmetrical. This asymmetry is noticeable even in most small-signal locking experiments.

The reason for this asymmetry is that the change in diode susceptance as the device RF voltage is increased is independent of the sign of the frequency deviation $F_L - F_0$. The susceptance of the device is influenced only by the magnitude of the RF voltage at its terminals. The change in susceptance can be regarded as causing a change in the F_0 of the oscillator. Thus for a given small-signal value of frequency deviation, the large-signal value is effectively altered and if, for example, it is decreased in the case of a positive initial deviation, then it is increased for a negative initial deviation, with the resultant assymetry in the phase-gain curves.

The circuit separating the diode from the load has been represented by an ideal transformer with a parallel susceptance due to the cavity. In practice, the coupling circuit is distributed and may contain other resonant elements. The phase-frequency response of such a circuit may be nonlinear and hence distort the phase-gain curves. The presence of resonances other than the main cavity resonance may make the oscillator output noisy or introduce instability. The design of a cavity to be used for making the phase-gain measurements should thus be relatively simple and free from such irregularities.

The simple theory outlined in the preceding predicts the general form of the phase-gain curves obtained when the locking signal is varied from a low level to a level which is comparable to the free-running oscillator output level. In the next section the experimental measurements necessary to obtain the phase-gain curves and hence the diode admittance values will be described.

MEASUREMENT TECHNIQUES

Measurements are made using the equipment shown in Fig. 4. A circulator is used to separate the locking and output signals, and a microwave network analyzer is used to compare their amplitudes and phases. The locking power is varied and the power output of the system, together with the phase-amplitude characteristics, recorded. This is repeated for a number of locking frequencies close to the natural frequency of the oscillator (i.e., within 5 percent). The measured values are substituted into (1) to give the normalized conductance (or susceptance) as a function of normalized RF voltage. The normalizing factor for each parameter is the relevant free-running value.

The information may be put in absolute terms if one point on the graph is fixed. In the present case the diode conductance and susceptance are measured for the free-running condition. The oscillator cavity is made with a detachable diode mount, which takes the form of a precision 7-mm connector. The diode under test is located at the mating plane of the connector and the admittance measured at this point, looking into the cavity, is the negative of the free-running admittance of the diode. The diode admittance together with the free-running output power may be used to find the RF voltage at the diode, for the free-running case. Thus the normalizing factors for conductance, susceptance, and voltage are known.

The phase values which are substituted into (1) are referred to the line of symmetry of the phase-gain curves. The values of phase obtained directly from the network analyzer are displaced from the line by an arbitrary amount due to differences in the electrical line lengths in the reference and test channels of the analyzer. The phase offset is obtained from a study of the small-signal region of the phase-gain characteristics, i.e., where the gain is greater than 10 dB and preferably $\sim$20 dB.

If the data include a curve for $\Delta F = 0$, then a line representing the phase offset is asymptotic to this curve in the high gain region. If the curve for $\Delta F = 0$ is not included, then the phase offset may be estimated from a line of symmetry for the curves in the high gain region. Some idea of this procedure can be obtained by examining Fig. 6. Here a curve for $\Delta F = 0$ has been included and this provides a line of symmetry for the high gain region. In this figure the phase offset is approximately 100°.

A correction is also required for the change of phase that occurs with changing frequency, again due to the difference in line lengths. This is facilitated by making a calibration of phase against frequency, around the operating frequency, with a short circuit replacing the test cavity.

This method of diode characterization may be automated to a reasonable degree. A data logger is used to record the information

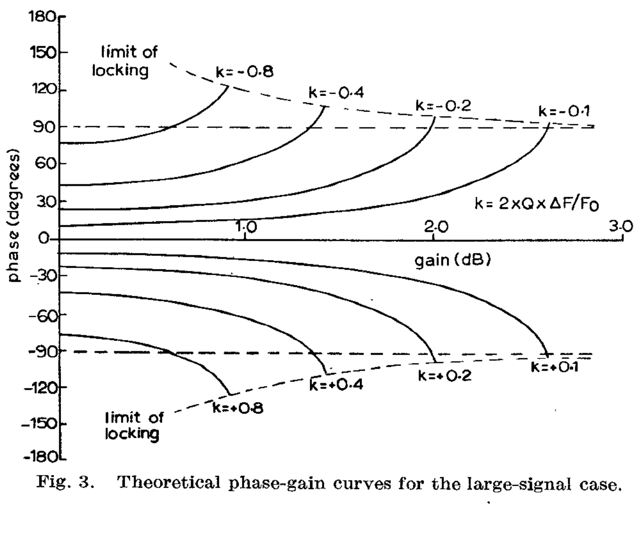

Fig. 3. Theoretical phase-gain curves for the large-signal case.

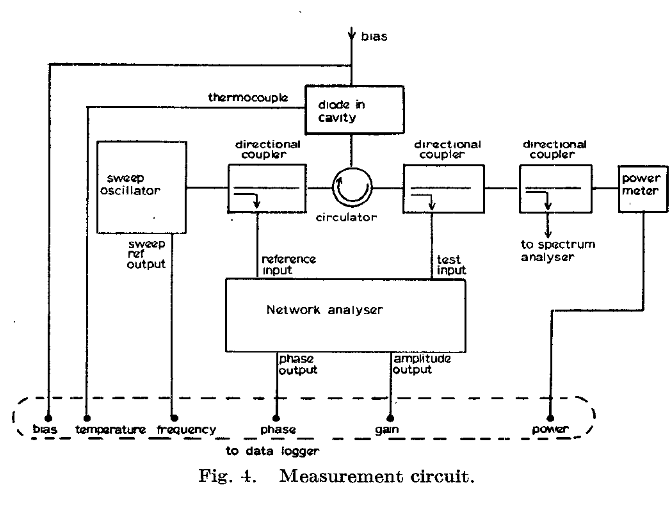

Fig. 4. Measurement circuit.

which is then processed by a simple computer program to directly give plots of diode conductance and susceptance as functions of RF voltage swing.

MEASUREMENT ON GUNN DIODES

The test cavity used for the Gunn diode measurements is shown in Fig. 5.

The resonant structure is a length of 50-Ω 7-mm line which is approximately $\lambda/2$ long at the free-running frequency. RF power is coupled into and out of the cavity via a capacitive probe. The probe insertion is variable which allows the coupling to the diode to be adjusted. The diode is mounted in a copper end piece which is water cooled to maintain frequency stability. The dc bias is applied through a low-pass filter.

The Gunn diodes used were a low power longitudinal type with structures as shown. The diode was operated at 9.2 GHz and the coupling to the diode was adjusted for the maximum free-running power output of 25 mW. Typical phase-gain curves for this diode are shown in Fig. 6. Fig. 7 shows the conductance and susceptance as

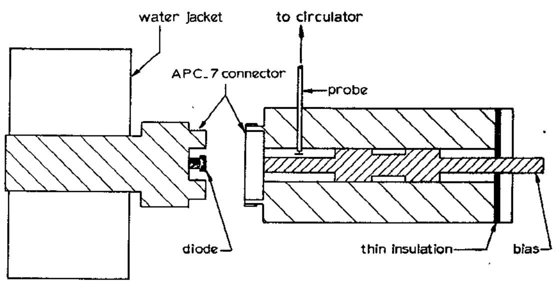

Fig. 5. Test cavity for Gunn diodes.

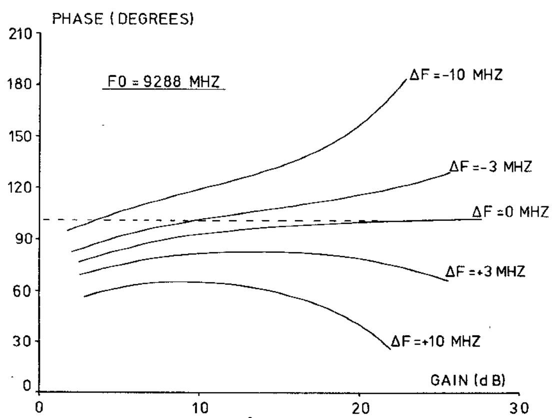

Fig. 6. Typical phase-gain curves obtained for a Gunn diode.

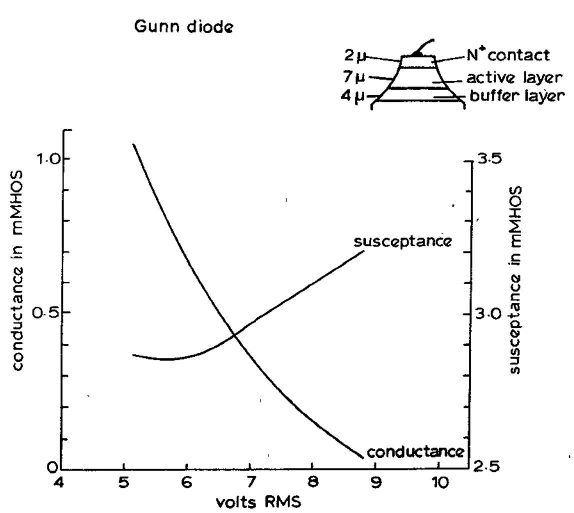

Fig. 7. Dynamic conductance and susceptance for Gunn diode.

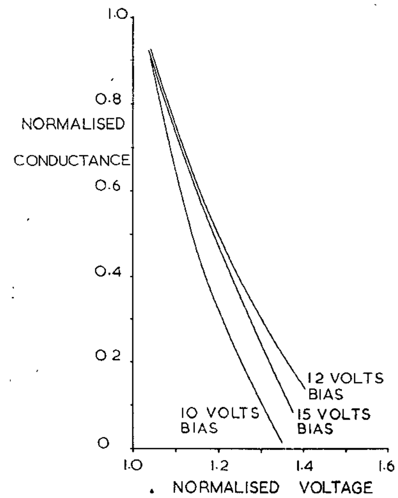

Fig. 8. Variation of conductance with dc bias for Gunn diode.

functions of RF voltage for the semiconductor chip. The effects of the diode package have been allowed for using the package equivalent derived by Owens and Cawsey [5].

This equivalent circuit, for the S-4 type package, contains five elements whose values represent the average of a number of measurements. While this circuit is not strictly accurate in the present case, due to variations in bonding, etc., it indicates the effect that the package has on the device conductance and susceptance. The package has the same effect as an ideal transformer of turns ratio approximately 3.2 to 1, the conductance at the chip being some ten times less than that at the terminals.

The effects of dc bias changes on the diode conductance have also been investigated. Fig. 8 shows a plot of normalized conductance for three different dc bias voltages.

Theoretical considerations of the effect of bias voltage on the dynamic conductance show that, as the dc bias is increased beyond the threshold value, so the fall of conductance with increasing RF voltage is reduced [6]. This is indicated in the experimental curves until the 12-V bias point is reached. Further increase in dc bias causes the curve to fall back again, probably due to the limit of thermal dissipation of the diode being approached. The diode used had a threshold voltage of 4.5 V.

The dependence of the diode conductance (and susceptance) upon diode temperature has also been examined. A thermal isolation piece, consisting of a short brass cylinder, having very thin walls, was placed between the diode heatsink and the main body of the cavity. The temperature of the diode heatsink was raised and the cavity body was cooled by means of a cold water jacket. The effects of the rise in temperature were thus restricted to the diode and its immediate surroundings, leaving the cavity resonant frequency and the output coupling ratio relatively unaffected. The dynamic conductance as a function of RF voltage was evaluated for four heatsink temperatures in the range 15°–50°C. The results are shown in Fig. 9 where the curves are all normalized with respect to the 15°C case.

MEASUREMENTS ON IMPATT DIODES

The test cavity used for the IMPATT diode measurements is shown in Fig. 10.

This is a slug-tuned cavity, again with an APC-7 connector located at the diode mounting plane. The diode was mounted in a similar fashion to the Gunn diode and the holder was also water cooled. Bias was applied through a bias-tee which was connected as shown. The bias-tee is thus considered as part of the cavity and its imperfections are included in the transformer representing the coupling between diode and load.

A Hewlett-Packard IMPATT diode type 5082-0432 was used for initial measurements. This is a silicon IMPATT diode with a typical output of 100 mW in the frequency range of 8–12 GHz. The breakdown voltage is 78 V, and at breakdown the junction capacity is 0.2 pF. The device was operated at 25-mA bias current and the cavity was adjusted to give 30-mW output at 8.3 GHz. Conductance and susceptance characteristics are shown in Fig. 11. These values refer to the package terminals and are not transformed to the diode chip.

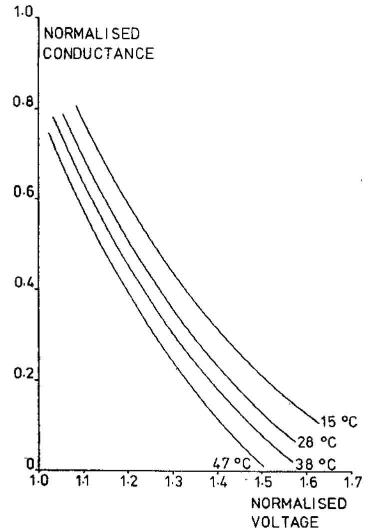

Fig. 9. Variation of conductance with temperature for Gunn diode.

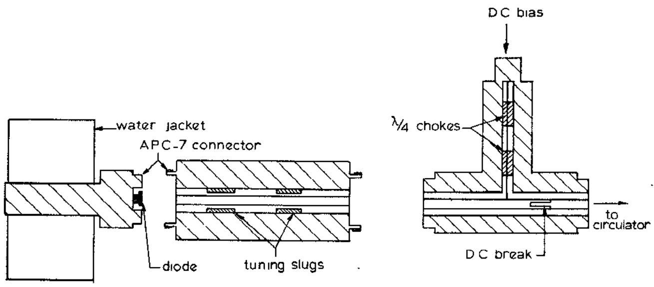

Fig. 10. Test cavity for IMPATT diodes.

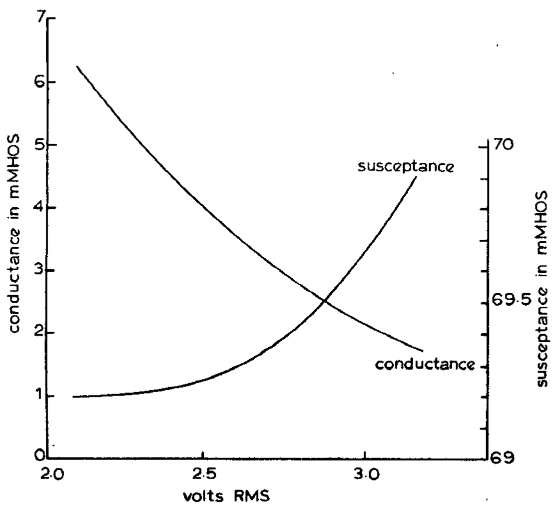

Fig. 11. Dynamic conductance and susceptance for IMPATT diode.

DISCUSSION

This method of device characterization has several advantages over other methods.

The device under test is operated as an oscillator near to its normal working point. This eliminates problems of device stabilization which are encountered with some other methods.

The variation in the locking signal amplitude appears to the device as a change in the complex admittance which it "sees." Thus the method is comparable with other techniques in which the load admittance is physically altered. Advantages of the present method are that the frequency remains independent of the load condition, and that the RF voltage at the diode is controllable. The method is relatively fast and amenable to automation.

The relative measurements can all be made without physically disturbing the cavity. This is important in the case of small changes in the diode admittance, e.g., when the dc bias is changed, since they may often be obscured by the results of mechanical disturbance.

With this method, results can only be obtained over a limited range of diode RF voltage. However, this range is in the region in which the diode is usually operated.

The accuracy of this method of diode characterization is affected by several factors, the most important being the following.

1) Inaccuracies in the measurement of the free-running diode admittance. These are due to the physical disturbance of the cavity required to make the measurement and to the subsequent alteration of the field pattern near the diode site.

2) Errors in the relative measurement of conductance against voltage (and susceptance against voltage). These are mainly due to difficulty in estimating a line of symmetry for the phase-gain curves, which is required for the 0° phase reference.

3) Changes produced in the voltage and current waveforms at the diode, due to the presence of the large locking signal. The voltage and current waveforms in a Gunn diode oscillator have been examined under large-signal injection-locking conditions. The harmonic content which is of the order of 10 percent is not greatly influenced by the introduction of a locking signal.

An error is introduced in the calibration procedure when the test cavity is replaced by a short circuit. The error arises from the difference in the electrical length between the cavity and the short circuit, and, for the frequency deviations used, is less than 3 percent.

A correction may be made from a calibration taken with a short circuit placed at the diode site. This calibration may also be used to correct for nonlinearities in the phase-frequency response of the cavity coupling system.

Typical measured values for the deviation from a linear phase-frequency response account for an error of 1 percent.

The overall measurement accuracy is estimated to be better than ± 20 percent.

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Designing Microstrip Matching Networks for Microwave-Transistor Power Amplifiers

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***Abstract*—A computer-aided design procedure is formulated for *L*-band-transistor power amplifiers. The procedure incorporates a precision measurement technique for transistor impedances, a model for large step discontinuities in microstrip, and an optimization routine for the direct realization of broad-band matching circuits.**

Manuscript received May 7, 1974; revised July 1, 1974.
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