A Power Reflection Technique for Characterization of High Quality Varactor Diodes

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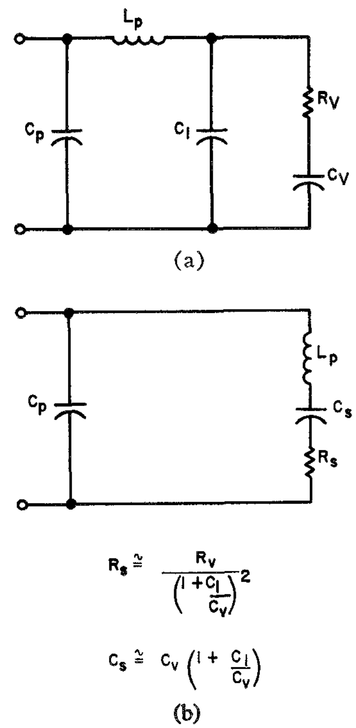
Abstract—High quality varactor diodes are currently characterized by either a "relative impedance" or a "transmission" technique. In view of the present limitations of these methods, a third method, a "reflection" technique, has been proposed. This method requires measuring the power reflected from a diode near its series resonant frequency. The diode is located as the termination of a low impedance coaxial line. The primary virtue of the power reflection technique is the possibility of accurate determination of the diode series resistance.

Both the reflection method and the transmission method offer the advantage of microwave characterization near the diode series resonant frequency. The power reflection technique is presently limited to diodes that series resonante below 24 GHz. The accuracy of the technique is limited by possible impedance transformations due to the diode mount and diode package.

I. Introduction

As THE QUALITY of microwave varactor diodes rapidly improves, a simple and accurate means of characterization has become increasingly important. A new method has been developed which is suitable for characterization of the highest quality varactor diodes. The commonly accepted varactor equivalent circuit is given in Fig. 1. The present methods for determining the series resistance of the equivalent circuit consist of either a "relative impedance" or a "transmission" technique. The methods of varactor characterization are summarized in Fig. 2.



$$R_s \cong \frac{R_V}{\left(1+\frac{C_l}{C_V}\right)^2}$$

$$C_s \cong C_V\left(1+\frac{C_l}{C_V}\right)$$

Fig. 1. Varactor equivalent circuits.

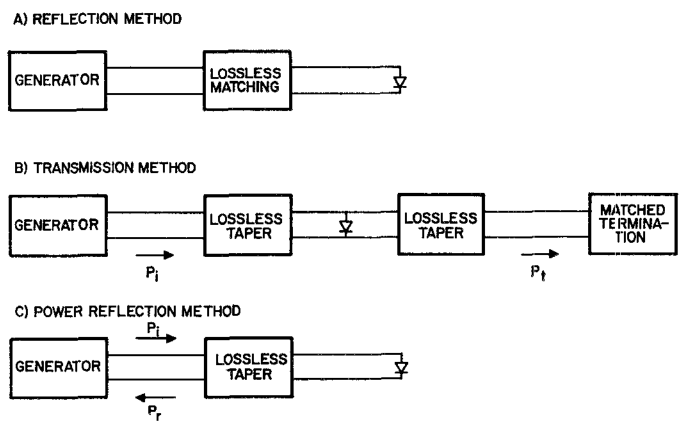


Fig. 2. Varactor characterization techniques.

The relative impedance method [1], [2] requires matching the diode impedance to a slotted line impedance with a lossless, tunable transformer. By varying the diode bias voltage and measuring the VSWR, the diode quality may be calculated by assuming a diode resistance which does not vary with bias voltage. Since epitaxial diodes have shown significant variation of resistance with bias voltage, this method of characterization is unsuitable for many high quality varactor diodes. An additional problem with this method is the large correction required for system loss.

A variation of the relative impedance method is the hyperbolic distance method [3]. Again the diode terminates a transmission line. However, a lossless tunable transformer is no longer required for matching the diode to the line. The diode quality is found by measuring the hyperbolic distance between two bias states. This method requires measurement of high VSWR, measurement of phase angle, correction for line losses, and the assumption of a constant diode series resistance.

Both the relative impedance method and the hyperbolic distance method require a correction for measurement losses. However, these losses will vary with the

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bias voltage in a manner which cannot be predicted. The position of the voltage standing wave pattern relative to the lossiest equipment positions will determine the unwanted system losses. The primary objections to the relative impedance and hyperbolic distance methods are the assumption of constant diode resistance and the large correction for measurement losses which cannot be accurately determined.

The transmission method [4], [5] requires mounting the diode in reduced-height waveguide and measuring the transmitted power near the series resonant frequency. The method consists of either holding the bias fixed and varying the frequency [4] or holding the frequency fixed and varying the bias [5]. Due to the necessity of mounting the diode in waveguide, this method is limited in the range of junction capacitance and package inductance which may be resonated. The diode mount must permit dc biasing without loss of RF energy. Also, in a waveguide measuring system the value of characteristic impedance does not always have an unambiguous value [6]. For diode packages with a thin single wire, the waveguide impedance is well defined [4]. However, all diode packages do not satisfy this condition. The transmission method has given higher cutoff frequencies when compared to the relative impedance method [3], [6].

In an attempt to improve upon the shortcomings of the existing methods, a new method of varactor characterization has been developed, the power reflection technique. The measurement requires locating the varactor diode as the termination of a low impedance section of coaxial line. The varactor parameters may be determined from measuring the VSWR, and therefore the reflected power, as a function of frequency near the series resonant frequency. This method eliminates the dc biasing problem found in the transmission method.

The power reflection technique could also be applied by varying the bias voltage and holding the frequency fixed. In this case, only the reactance of the junction capacitance is varied. For variable frequency measurements, the reactance of every reactive element in the circuit varies, whether or not it is adequately accounted for in the equivalent circuit. Fixed frequency measurements would simplify possible corrections due to the impedance transforming properties of the diode mounting configuration, which probably vary with frequency. However, fixed frequency measurements could only be applied to diodes with a constant series resistance.

It is beyond the scope of this paper to justify the varactor equivalent circuit of Fig. 1. Such a justification would require both VSWR and phase measurements taken at a fixed frequency and a knowledge of the impedance transformation of the diode mount. The primary purpose of this paper is the application of a simple method for accurate determination of the resistance which terminates a transmission line. This resistance is the diode series resistance (R_s) at the chosen bias voltage, if there is no impedance transformation caused by the diode mount and the diode package. Possible impedance transformations must be carefully investigated when measuring high quality diodes.

II. Power Reflection Characterization

The varactor parameters for the equivalent circuit of Fig. 1(b) may be calculated from the equations derived below. The return loss of the diode, (RL), and the diode resistance are given by the following relations.

$$\begin{aligned}(\mathrm{RL}) &= \left(\frac{P_{\mathrm{in}}}{P_r}\right)_0 = \frac{1}{|\rho_0|^2} = \left(\frac{R_s + Z_0}{R_s - Z_0}\right)^2 \\ &= \left(\frac{S_0 + 1}{S_0 - 1}\right)^2\end{aligned} \tag{1}$$

$$R_S = Z_0\left(\frac{\sqrt{\mathrm{RL}} - 1}{\sqrt{\mathrm{RL}} + 1}\right) = \frac{Z_0}{S_0} \tag{2}$$

where

ρ_0 = voltage reflection coefficient at resonance
Z_0 = characteristic impedance at the varactor
S_0 = VSWR at the resonant frequency.

These relations hold for an overcoupled line, $R_s < Z_0$, although a similar relation holds for an undercoupled line. The package capacitance C_p is assumed to have a negligible effect on the diode impedance near the series resonant frequency. Upon determining the resistance from (2) and the capacitance C_s from low-frequency bridge measurements at 1 MHz, the varactor cutoff frequency may be calculated [7]. It should be noted that the value of C_s must include the influence of C_1, the internal fringing capacitance [8]. The value of C_s is simply the difference between the total measured capacitance and the capacitance C_p, which is measured from empty packages.

The parameters L_p and C_s may be calculated from a knowledge of the 3 dB frequencies, which are determined when the power reflected from the load is twice that at resonance. At the 3 dB frequencies, f_1 and f_2, the following relations hold.

$$\frac{(\mathrm{RL})}{2} = \left(\frac{P_{\mathrm{in}}}{P_r}\right)_{1,2} = \frac{1}{|\rho_{1,2}|^2} = \frac{|Z_s + Z_0|^2}{|Z_s - Z_0|^2} \tag{3}$$

$$Z_s = R_s \pm jX_s \quad (\text{at } f_1 \text{ or } f_2). \tag{4}$$

Solving (3) gives the reactive impedance at the 3 dB frequencies.

$$X_s = \left(\frac{(\mathrm{RL})}{(\mathrm{RL}) - 2}\right)^{1/2} \frac{2R_s}{\sqrt{\mathrm{RL}} - 1}\,. \tag{5}$$

Notice that the reactive impedance does not equal the real impedance at the 3 dB frequencies since the present definition of 3 dB frequencies is not equivalent to the more conventional definition. Rearranging (1), (2), and (5), gives

$$X_s = R_s \frac{(S_0^2 - 1)}{(-S_0^2 + 6S_0 - 1)^{1/2}} = R_s V(S_0). \quad (6)$$

Further manipulation gives the remaining varactor parameters.

$$X_s = \omega_2 L_p - \frac{1}{\omega_2 C_s} = \frac{1}{\omega_1 C_s} - \frac{1}{\omega_2 C_s} = \frac{f_2 - f_1}{\omega_0 C_s f_0} \quad (7)$$

$$f_{c0} = f_0 Q = \frac{1}{2\pi R_s C_s} = V(S_0) \frac{f_0^2}{f_2 - f_1} \quad (8)$$

$$C_s = \frac{f_2 - f_1}{2\pi f_1 f_2 R_s V(S_0)} \quad (9)$$

$$L_p = \frac{1}{4\pi^2 f_1 f_2 C_s}. \quad (10)$$

A relation between the VSWR at the 3 dB frequencies and the VSWR at resonance may be found from (1) and (3).

$$S_{1,2} = \frac{S_0 + 1 + \sqrt{2}(S_0 - 1)}{S_0 + 1 - \sqrt{2}(S_0 - 1)}. \quad (11)$$

The functions $S_{1,2}$ and $V(S_0)$ have been plotted in Fig. 3. Since the function $V(S_0)$ has real values for $S_0 < 5.82$, the method must be modified for values of R_s less than $Z_0/5.82$.

For the case of an undercoupled line, $R_s > Z_0$, a similar derivation gives

$$R_s = Z_0 S_0 \quad (12)$$

$$f_{c0} = \frac{V(S_0) f_0^2}{S_0(f_2 - f_1)} \quad (13)$$

$$C_s = \frac{f_2 - f_1}{2\pi f_1 f_2 Z_0 V(S_0)}. \quad (14)$$

The coupling of the line may be determined by observing the shift of the voltage minimum when the diode is at the resonant frequency compared to an open circuited line. For critical coupling ($R_s = Z_0$), both sets of equations are valid.

The complete measurement setup is shown in Fig. 4. It should be noted that the VSWR has been reduced for better measurement accuracy by lowering the characteristic impedance at the varactor. The importance of low VSWR for the measurement of high quality varactor diodes cannot be overemphasized. Although the present setup uses a taper to lower the characteristic impedance, a superior technique would use a low impedance slotted line. With a low impedance slotted line, the taper losses would be eliminated.

Although the ideal measurement system would have negligible losses, the measurement of high quality varactors will be limited by the transformer losses. These losses may be estimated by inserting a shorted package

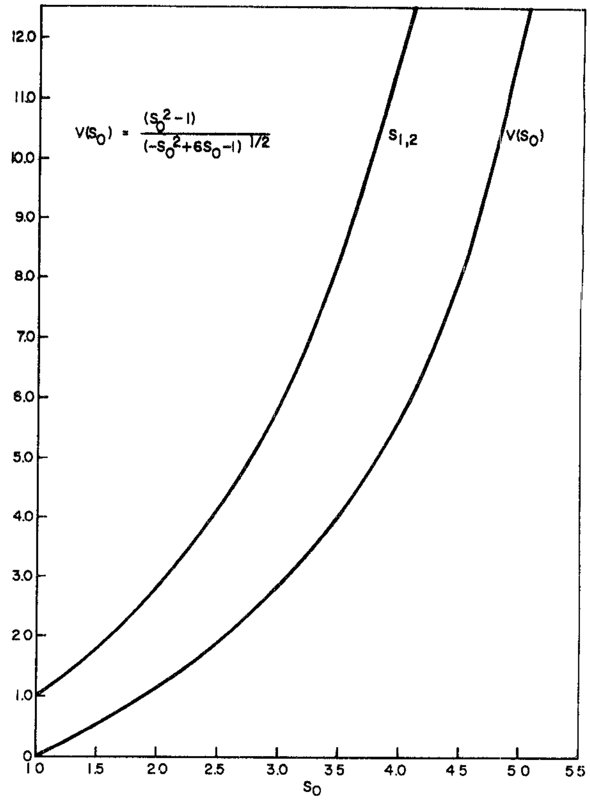

Fig. 3. VSWR at 3 dB frequency and $V(S_0)$ vs. S_0.

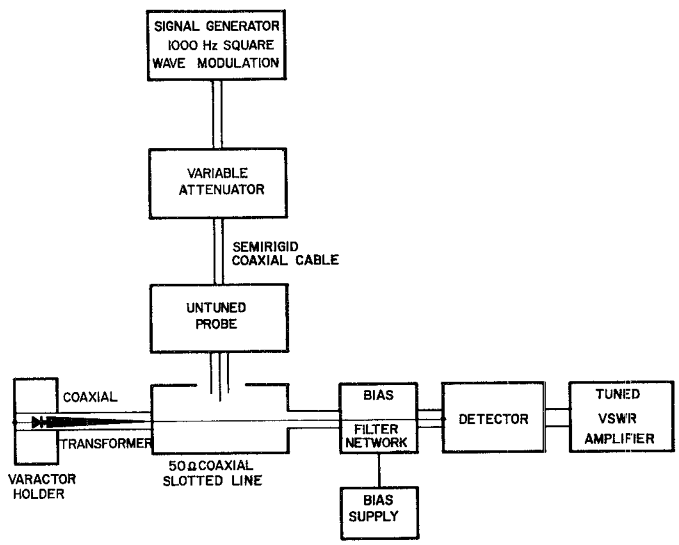

Fig. 4. Power reflection test arrangement.

in the holder and measuring the return loss as a function of frequency. Since the reflection coefficient of the taper will vary with frequency, an average value of taper return loss must be estimated. The variation of taper reflection coefficient with frequency has been given elsewhere [9]. After determining an average value for transformer loss, the diode return loss at the resonant frequency may be calculated from

$$(\mathrm{RL})_d = (\mathrm{RL})_m - (\mathrm{RL})_T \quad (15)$$

where

$(RL)_d$ = diode return loss in dB
$(RL)_m$ = measured return loss
$(RL)_T$ = taper return loss.

Since VSWR is directly related to return loss, the measured VSWR at resonance may be corrected for the transformer loss.

In summary, three variations of the power reflection technique have been presented for measurement of varactor cutoff frequency. Each method will find application for various diodes. The simplest method (Method 1 of Table I) consists of measuring R_s at the series resonant frequency and measuring C_s at 1 MHz. For diodes with moderately high cutoff frequencies, the cutoff frequency may be determined from the VSWR at resonance and the 3 dB frequencies (Method 2 of Table I). Finally, for the highest cutoff frequency diodes where taper losses are significant, the cutoff frequency may be calculated from measuring the value of R_s corrected for taper loss and measuring C_s at 1 MHz (Method 3 of Table I). Since the highest cutoff frequency diodes usually have a significant correction due to C_1, this capacitance must be included in the calculation for varactor cutoff frequency.

TABLE I

CALCULATED CUTOFF FREQUENCY (GHz)

Diode	C_s (pF)	Method 1 Measure R_s and C_s (1 MHz)	Method 2 Measure R_s, f_1, and f_2 Use (8) or (13)	Method 3 Measure R_s, $(RL)_T$, and C_s (1 MHz)
1	0.50	172	110	212
2	1.28	62	56	76
3	0.75	64	53	68
4	1.00	45	41	50

III. EXPERIMENTAL RESULTS

The transformer consisted of a tapered section which transforms 50 ohms to 7.2 ohms over a distance of 3.5 inches. The transformation consisted of a linear taper on the center conductor and an exponential taper on the outer conductor. The end of the center conductor was designed to hold the microwave "pill" package. A sketch and photograph of the varactor holder are given in Figs. 5 and 6. All parts were gold plated for minimum loss.

When a dummy silver package was inserted in the holder, the average VSWR was greater than 20 over the frequency range 4–18 GHz, which corresponds to a taper return loss of less than 0.9 dB. The variation of VSWR with frequency agreed reasonably well with the prediction for an exponential taper.

The value of C_1 was determined by calculating the difference between the empty package capacitance and the packaged diode with the bonding wire held in place but not bonded. This method gives only an approximate evaluation of C_1. The value of C_1 for two 0.3 mil wires was less than 0.01 pF, which compares with values of 0.15 pF [6], 0.04 pF [8], 0.13 pF [10], and 0.25 pF [11], reported by other investigators. Each individual packaging method should be measured for C_1.

Typical VSWR data for four diodes is given in Fig. 7. The cutoff frequencies calculated from the three methods described previously are tabulated in Table I. The calculation given by Method 2 was made from (8). The results of Table I indicate Method 3 should be used for characterization of high quality diodes. This method will give the most accurate measurement of series resistance and, therefore, cutoff frequency.

The primary source of error is the impedance transformation due to the diode mount and the diode package [8]. The transformation due to the mount has been minimized by using small diameter coaxial line and forming a short chamfer on the end of the center conductor. The impedance transformation of the mount and the package may cause the measured resistance to differ from the diode resistance. An accurate method of evaluating these impedance transformations is needed. The authors have assumed the impedance transformation of the mount is unity. The impedance transformation of the package is partially included by the correction for C_1 given in Fig. 1(b).

In measuring the highest cutoff frequency diodes which are available [7], one is limited with coaxial lines to a resonant frequency below 24 GHz. This limits the junction capacitance to values greater than about 0.20 pF. Due to the limitation of finite contact resistance, higher cutoff frequency can usually be obtained from building smaller capacitance diodes. For this reason, the value of cutoff frequency should be attributed to a particular size diode. It should be clear that diode theory predicts an arbitrarily high cutoff frequency can be achieved from designing low breakdown voltages, small junction size, and low contact resistance.

The measurement data for typical low capacitance GaAs Schottky barrier diodes are presented in Fig. 8. Since bias tee's have not been available for frequencies above 12 GHz, data is usually taken at zero bias. After correcting for a taper return loss of 0.9 dB and measuring the 1 MHz diode capacitance, the cutoff frequency may be calculated using Method 3. No correction was made for C_1 since its value is negligible for this package. For an assumption of constant series resistance, which appears to be a valid assumption for these diodes, the cutoff frequency may be calculated as a function of bias voltage by measuring the variation of diode capacitance at 1 MHz.

It should be emphasized that higher cutoff frequency varactors are available with smaller junction capacities. However, using the present technique, these diodes must be resonated by increasing the package inductance. Since the package inductance must be low for practical circuits, waveguide measurement techniques are required for characterization of the highest quality varactor diodes.

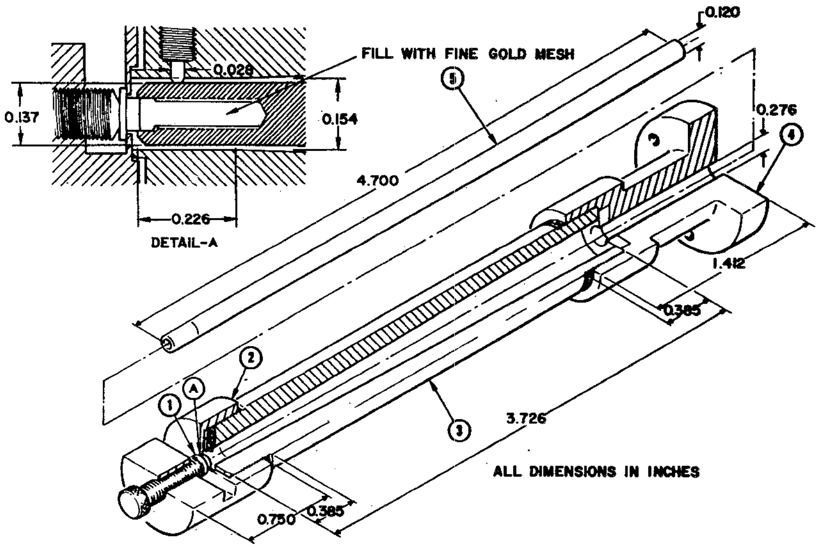

Fig. 5. Coaxial transformer assembly.

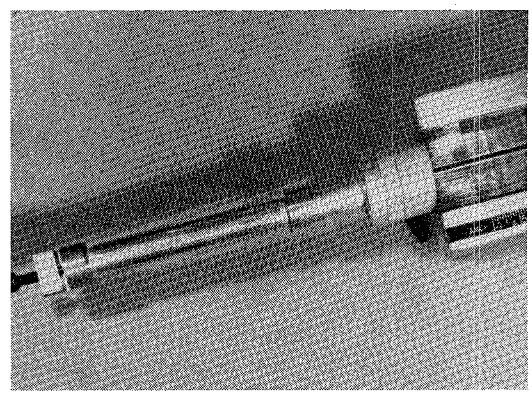

Fig. 6. Coaxial transformer assembly.

LEGEND TYPE	BIAS VOLT.	V_B VOLTS	f_{co} GHz	C_S (pF)
o GaAs SCHOTTKY	0	9.3	110	0.50
x GaAs MESA	0	12.2	56	1.28
Δ Si MESA	-6	112	53	0.75
□ Si SCHOTTKY	0	33	41	1.00

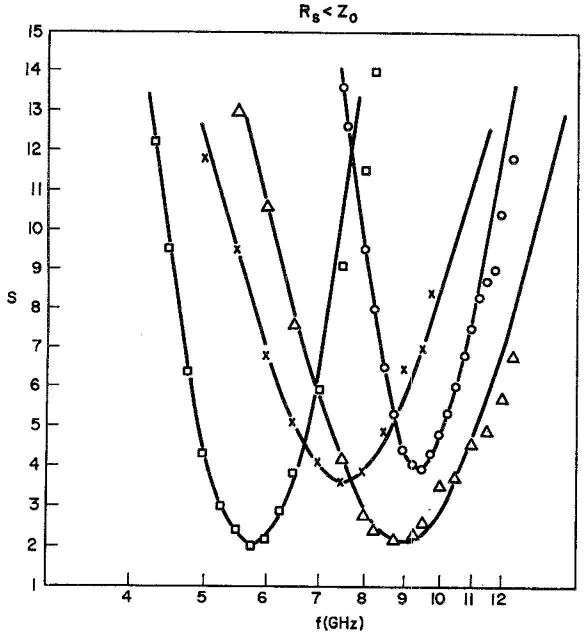

Fig. 7. Varactor data.

UNIT	BIAS (VOLTS)	C_S (pF)	UNCORRECTED f_{CO} (GHz)	CORRECTED f_{CO} (GHz)
o	0	0.287	385	522
o	-4	0.115	958	1,300
x	0	0.254	375	460
x	-6	0.149	639	785

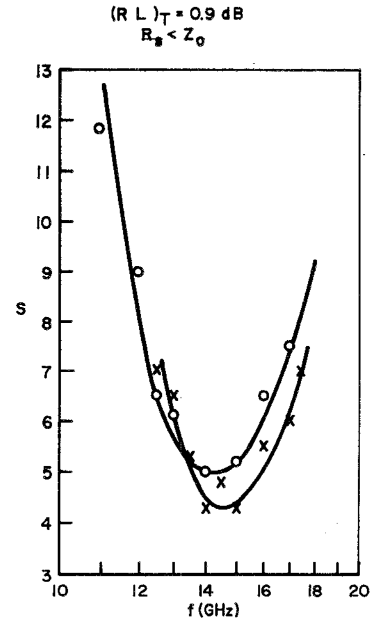

Fig. 8. Data for GaAs Schottky barrier diode.

IV. Conclusions

A new method has been proposed for measuring the parameters of high quality diodes. The distinguishing feature of this method is the measurement of the resistance which terminates a low impedance coaxial line. If the impedance transformation of the diode mount and package is unity, the measured resistance is the varactor series resistance. For low cutoff frequency diodes, all of the varactor parameters, L_p, C_s, R_s, and f_{cs}, may be calculated when the holder losses may be neglected. For the highest cutoff frequency diodes, corrections for losses must be made. A low impedance slotted line should lower the correction for loss. The power reflection technique has been used to measure zero-bias cutoff frequencies of 500 GHz on GaAs diodes with junction capacities of 0.26 pF. The power reflection technique is limited to diodes which series resonate below 24 GHz for present 50 ohm slotted lines.

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A Positive Resistance Up-Converter for Ultra-Low-Noise Amplification

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***Abstract*—An ultra-low-noise two-channel tunable amplifier system, operating in the 1.5 to 2.5 GHz frequency range, consisting of a cooled positive resistance parametric up-converter followed by a traveling-wave maser (TWM) and down-converter, has been developed. A theoretical analysis of the important operating and design parameters of the up-converter is presented and experimentally verified. Experimental data is given on the operation of the up-converter with the input and output ports reversed (down-converter), and is shown to correlate with the theoretical model.**

A brief discussion is presented on the TWM, and the spurious signal considerations which govern the choice of maser center frequency (up-converter output frequency). Finally, some preliminary system data is given showing the low noise performance of the overall cascaded amplifier integrated with a 4.2°K closed-cycle refrigerator.

I. INTRODUCTION

A. General

DURING the past few years the technology of ultra-low-noise amplifiers has been in a dynamic growth stage. The impetus to this growth has been the constantly increasing activity in applications such as satellite communications, deep space communications, telemetry, and radio astronomy.

Two amplifier configurations that have been actively competing for use in these systems are the traveling-wave maser (TWM) and the cooled one-port parametric amplifier. The selection of one of these amplifiers for a given application has created many thought-provoking questions and controversies concerning trade offs in operating characteristics, complexity, cost, and the like. There are so many parameters that enter into the selection that it is almost impossible to present a simple set of criteria that will yield a satisfactory rule for the optimum choice.

The advantages associated with the traveling-wave maser and the parametric amplifier are relatively well-known. For example, the maser provides:

1) the ultimate in low-noise performance,
2) excellent stability characteristics (by virtue of the relative insensitivity of maser gain to pump power and frequency variations),
3) unconditional gain stability (resulting from isolator elements integrally located in the maser structure),
4) excellent intermodulation distortion characteristics,
5) no burnout problems.

The parametric amplifier, on the other hand, has the well-known advantages of:

1) providing broad instantaneous bandwidths with relative ease (5 to 10 percent),
2) rapid and simple electronic tunability,

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