GPO PRICE $ ____________

CFSTI PRICE(S) $ ____________

Hard copy (HC) $2.00

Microfiche (MF) 1.50

ff 653 July 65

TECHNICAL REPORT NUMBER 3

AN INVESTIGATION OF A VARACTOR-DIODE MODULATOR

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May 8, 1964

N66 30851 (ACCESSION NUMBER)
(THRU) 1
50 (PAGES)
(CODE) 10
CR-76357 (NASA CR OR TMX OR AD NUMBER)
(CATEGORY)
FACILITY FORM 602

CONTRACT NAS8-5231

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PREFACE

It is the purpose of this study to present the basic theory of switching devices, with emphasis on voltage-controlled semiconductor switches. In particular, a single varactor diode is used as the switching element in a coaxial modulator.

The principles given are basic and can be applied to switches for all frequencies. The first two chapters are concerned with the fundamental operation of a varactor diode. There are followed by two chapters on modulators used as switching devices. The experimental results and conclusions are presented in the final two chapters.

Although this study was confined to the use of a single varactor diode as the switching element, the results can be generalized to include several cascaded diodes to yield an enhanced isolation and power handling capability.

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AN INVESTIGATION OF A VARACTOR-DIODE MODULATOR

J. W. Cook and H. M. Summer

I. INTRODUCTION

The feasibility of utilizing a solid-state-semiconductor diode as a modulating and switching device for radio-frequency energy has been recognized since semiconductors were first used as detectors. As early as 1956, a solid-state microwave modulator was developed using germanium point-contact diodes.[1] The theory of attenuation which emerged from initial studies remained intact until the advent of the varactor diode, a semiconductor-junction diode with a nonlinear reverse-bias capacitance. The introduction of this device into the microwave electronics field led to the design of modulators with greater power handling capability, greater isolation properties, and eventually to the development of new theories.

It was noted as early as 1943, that microwave semiconductor p-n diodes exhibited a nonlinear junction capacitance which varied as a function of applied voltage.[2] At that time, this phenomenon was considered a necessary nuisance and due to the small value of the depletion-layer capacitance its effect was negligible except at very high frequencies. With the rapid advance of the state-of-the-microwave-art, the varactor diode was developed and this "nuisance" effect achieved great importance.

The present state-of-the-art hosts sophisticated design of modulators employing several diodes as switching elements at X-band frequencies.[3] In particular, these designs deal with methods of achieving high isolation and large power-handling capability by employing special transmission line schemes which are feasible only because of the high operating frequency.

This investigation deals specifically with the design of a low-frequency modulator employing a single varactor diode as the switching element. The important criteria, to be considered are: the maximum power limitations of the modulator in each mode of operation, and a large-signal analysis which describes the effects of variations in time average capacitance on the modulator characteristics.

II. VARIABLE CAPACITANCE DIODES

General Considerations

The two most important types of varactor diodes are; (1) the abrupt-junction varactor diode and (2) the graded-junction varactor diode. In this investigation only the p-n abrupt-junction diode will be considered. The complete equivalent circuit for such a device operating in the reverse-bias region is shown[4] in Figure 1. The most important element is the nonlinear capacitance C(v). The shunt conductance of the junction is represented by g; L_o is the lead inductance; C_o is the case capacitance; and R_s represents the finite series resistance of the bulk semiconductor. It is desirable to operate at frequencies such that L_o may be neglected. If the d-c leakage is also negligible and if C_o is lumped with the other circuit capacitance, the equivalent circuit reduces to that shown by Figure 2.

The nonlinear depletion-region capacitance C(v) of an abrupt-junction diode operating in the reverse-bias region is defined[5] by

$$C(V_o) = K\left[(V_o + \phi)\right]^{-\gamma}, \; -\phi \leq V_o \leq V_b \tag{1}$$

where

$C(V_o)$ = the variational capacitance,

V_b = the reverse breakdown voltage of the diode,

V_o = $V_{dc} + v(t)$,

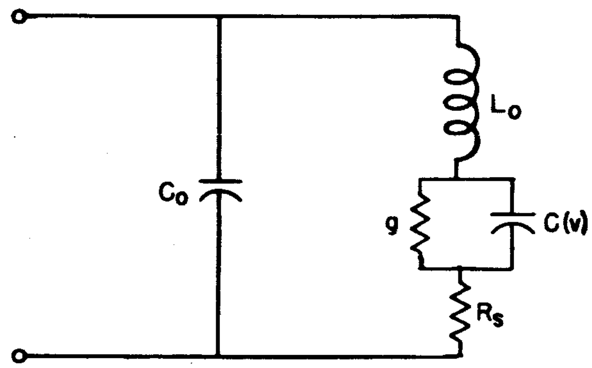


FIGURE 1 THE EXACT EQUIVALENT CIRCUIT OF A VARACTOR DIODE

(a) (b)

FIGURE 2 (a) THE APPROXIMATE EQUIVALENT CIRCUIT OF A VARACTOR DIODE
(b) THE SCHEMATIC REPRESENTATION OF A VARACTOR DIODE

V_{dc} = the d-c bias voltage,

$v(t)$ = the dynamic variation of voltage about V_{dc},

Φ = the contact potential of the diode,

γ = the characteristic exponent of the diode,

and

K = a constant of proportionality.

The characteristic exponent depends only upon the impurities present within the semiconductor material and is either .33 for the graded-junction diode or .5 for the abrupt-junction diode. The ideal charge distributions for the abrupt- and graded-junction diodes are shown in Figures 3a and 3b respectively. A typical capacitance-voltage characteristic obtained for a Pacific Semiconductor Inc. type PC-116 abrupt-junction varactor diode, using a method suggested by Chiffy and Gurley[6], is shown in Figure 4.

The original figure of merit of varactor quality was proposed by Uhlir[7] and was termed the cutoff frequency, f_c of the diode. This parameter which is defined as the largest frequency at which the quality factor "Q" can exceed 1.0 is given by

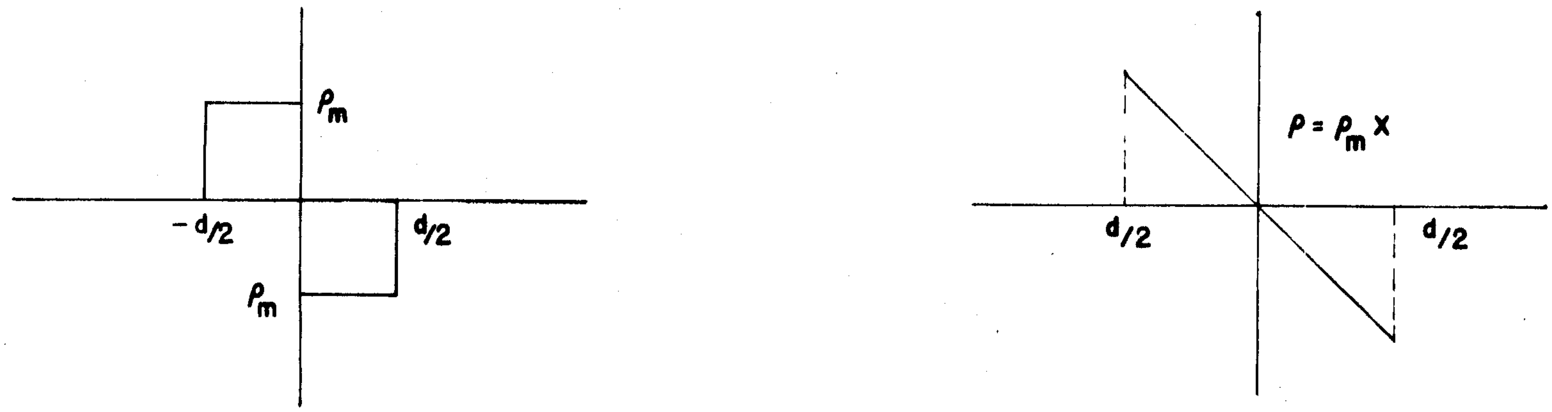
$$f_c = 1/(2\pi R_s C_{min}) \tag{2}$$

where

R_s = the series resistance,

and

C_{min} = the capacitance at reverse breakdown.



a   b

FIGURE 3 CHARGE DISTRIBUTION FOR (a) AN ABRUPT–JUNCTION VARACTOR DIODE AND (b) A GRADED–JUNCTION VARACTOR DIODE

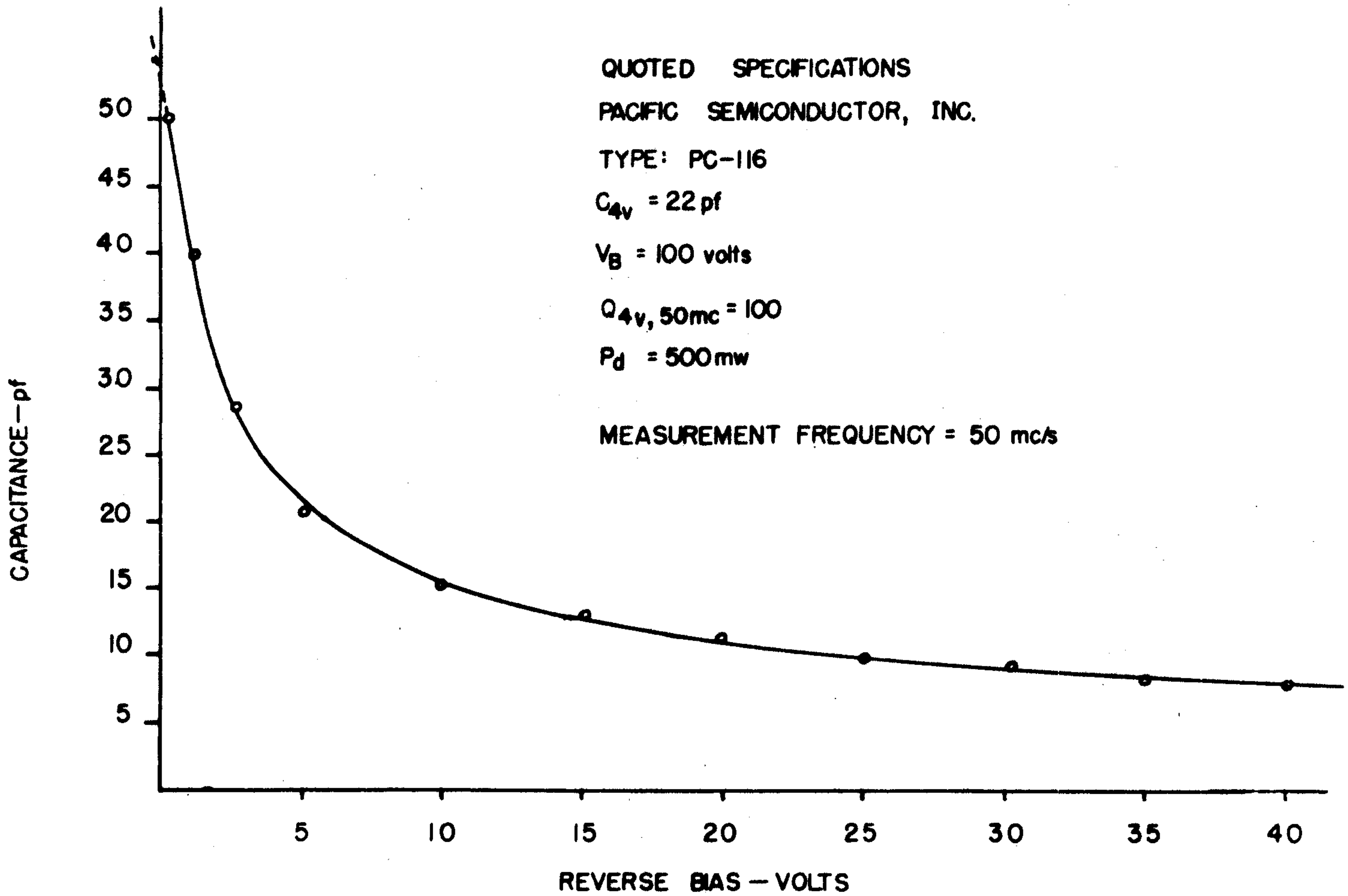


FIGURE 4 CAPACITANCE—BIAS VOLTAGE CHARACTERISTIC OF A P.S.I. TYPE PC-116 VARACTOR DIODE

Large-Signal Phenomena

In using varactor diodes as tuning elements in resonant circuits, a consideration of primary importance arises from the nonlinear capacitance-voltage relationship. An input signal is shown superimposed upon the capacitance-voltage characteristic curve of a typical varactor diode in Figure 5. If the voltage excursions about the quiescent point are quite small, the varactor capacitance is essentially constant and equal to C_q. Hence, for the small-signal case a tuned circuit incorporating the varactor has the usual amplitude-frequency response of a linear resonator. At larger input amplitudes, however, the effective capacitance increases to a value such as C_q' shown in Figure 5 and the resonant frequency decreases. For even larger excursions, the amplitude-frequency curve becomes distorted and ultimately exhibits "fold-over"[5], a condition under which the amplitude of oscillation is a multiple-valued function of the frequency. A plot illustrating "fold-over" is shown in Figure 6. Experimentally, this effect is observed as the so-called "jump phenomenon", a discontinuity in the amplitude of oscillation as the power input is increased under fixed-bias conditions.

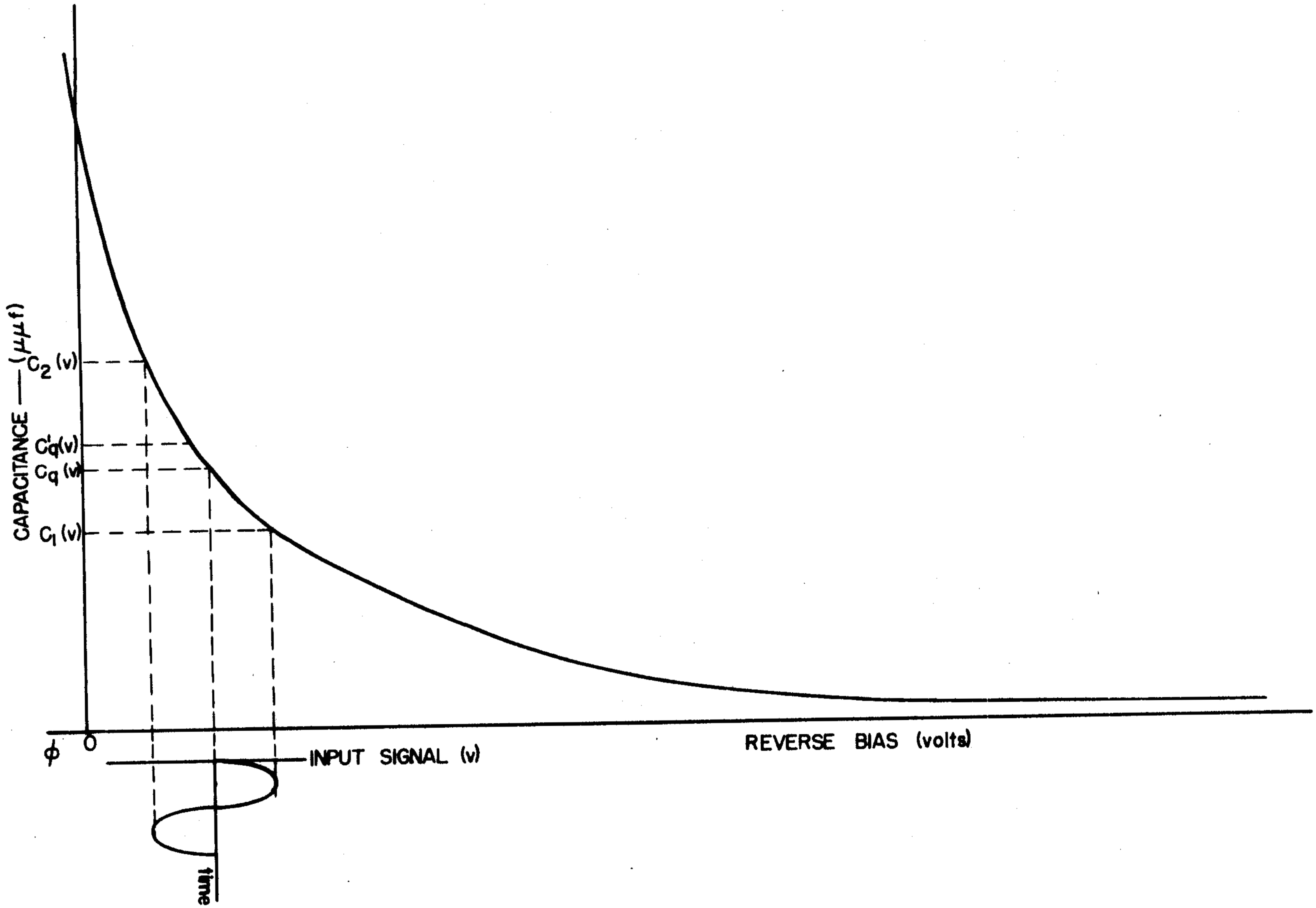


FIGURE 5 ILLUSTRATION OF VARIATION OF CAPACITANCE WITH SIGNAL LEVEL

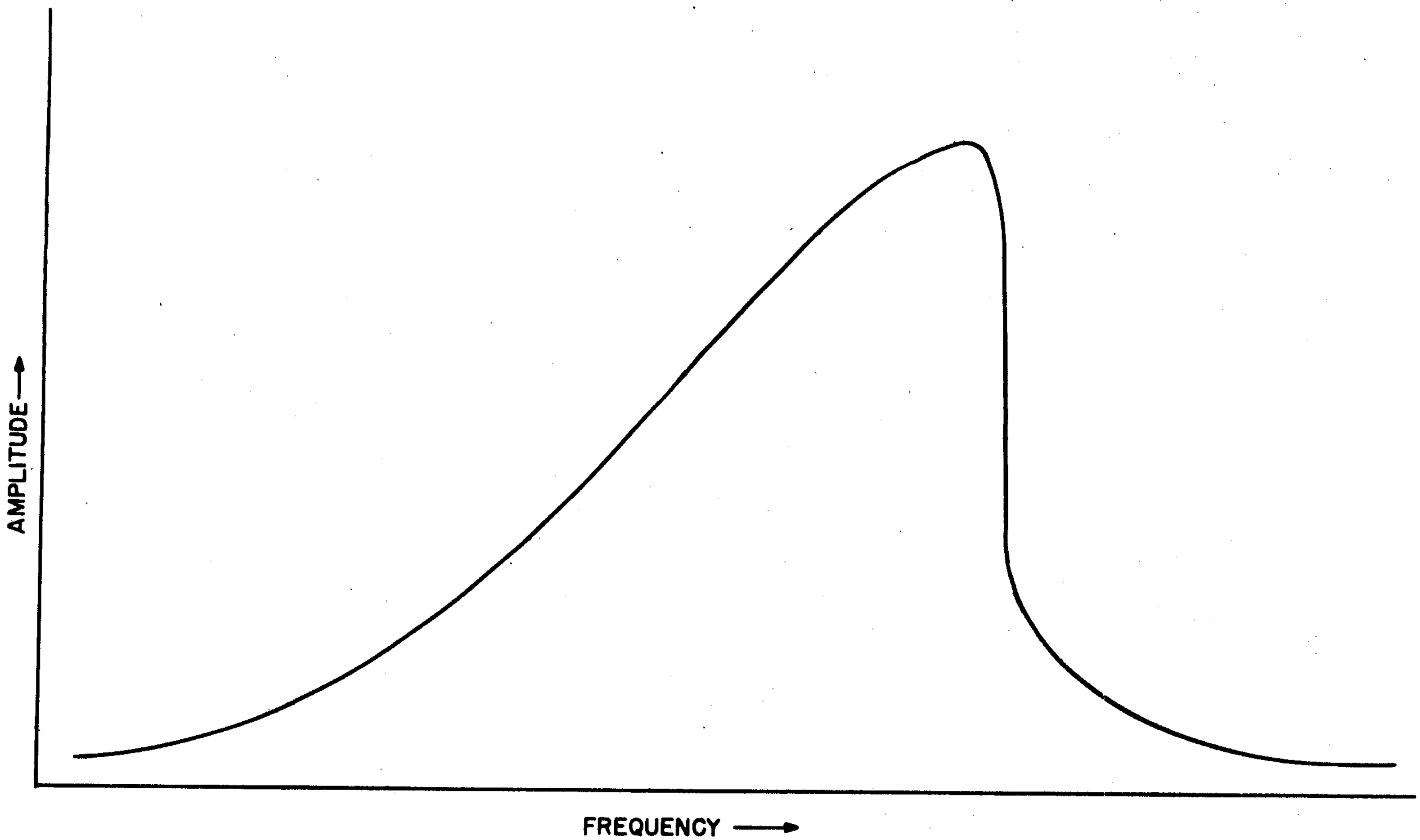


FIGURE 6 ILLUSTRATION OF FOLDOVER IN THE HIGH-FREQUENCY DIRECTION

III. THE VARACTOR-DIODE MODULATOR

Theoretical Analyses

The basic property of a varactor diode which allows its application in a switch is the change of impedance with a change of d-c bias applied across the diode. Several analyses of the diode switch have been developed on the assumption that the diode is a perfect attenuator and hence, most of the r-f power is either absorbed in the load or transmitted to the load with no power reflected[8].

A recent analysis has been presented by Garver[8] which distinguishes between the use of a perfect attenuator and a varactor diode as a switching device. This method yields an exact solution for the attenuation as a function of the various diode parameters.

The Shunt Modulator

The a-c equivalent circuit of a shunt type varactor-diode modulator is shown in Figure 7. The modulator is connected between the center conductor and outer conductor of a transmission line with a characteristic impedance of Z_o and is located one-half wavelength from the generator. The switch has two modes of operation: (1) the series-resonant mode, and (2) the parallel-resonant mode. Hereafter, the two modes of operation will be referred to as mode one and mode two respectively. It is necessary when discussing the operation of a diode switch to define certain terms. The attenuation of a device inserted in a transmission line is the ratio in decibels of the power incident to the power transmitted to the load. Attenuation in the series mode of operation, and attenuation in the parallel mode of operation are known as isolation and insertion loss, respectively.

In the equivalent circuit of a shunt-connected diode modulator as shown in Figure 7, E is the peak amplitude of the sinusoidal voltage developed across the system by the current source I. The load admittance is Y_o and the admittance of the modulator is Y_m. The power dissipated in the load Y_o which is assumed to be purely resistive is

$$P_L = \tfrac{1}{2} E^2 Y_o \quad . \tag{3}$$

If the modulator admittance $Y_m = G + jB$ where G is the conductance, and B is the susceptance of the modulator, then the power transmitted past the modulator to the load is

$$P_t = \tfrac{1}{2} I^2 Y_o / \left[(G + 2Y_o)^2 + (B)^2 \right] . \tag{4}$$

The incident power to the diode modulator is the power in the forward traveling wave moving toward the load. This is obtained by setting $Y_m = 0$. Thus,

$$P_i = I^2/(8Y_o) \; . \tag{5}$$

The logarithm of the ratio of equations (4) and (5) yields the expression for the attenuation of the varactor-diode modulator given by

$$\alpha = 10 \log_{10} \tfrac{1}{4} \left[(G/Y_o + 2)^2 + (B/Y_o)^2 \right] \tag{6}$$

where

α = the attenuation in decibels,

G = the conductance of the modulator,

B = the susceptance of the modulator,

and

Y_o = the characteristic admittance of the transmission line.

In order to yield the desired modulation characteristic, the diode has to be switched between two resonant modes. To achieve this result, tuning elements are added in series and in shunt with the diode. The result is shown in Figure 8a. The r-f admittance of the modulator circuit is operated alternately in series and shunt resonance, mode one and mode two respectively, by varying the capacitance of the diode with a control voltage.

From the equivalent circuit shown in Figure 8a the conductance and the susceptance of the modulator can be represented as

$$G = R_s / \left[R_s^{\,2} + (X_{L1} - X_C)^2 \right] , \tag{7}$$

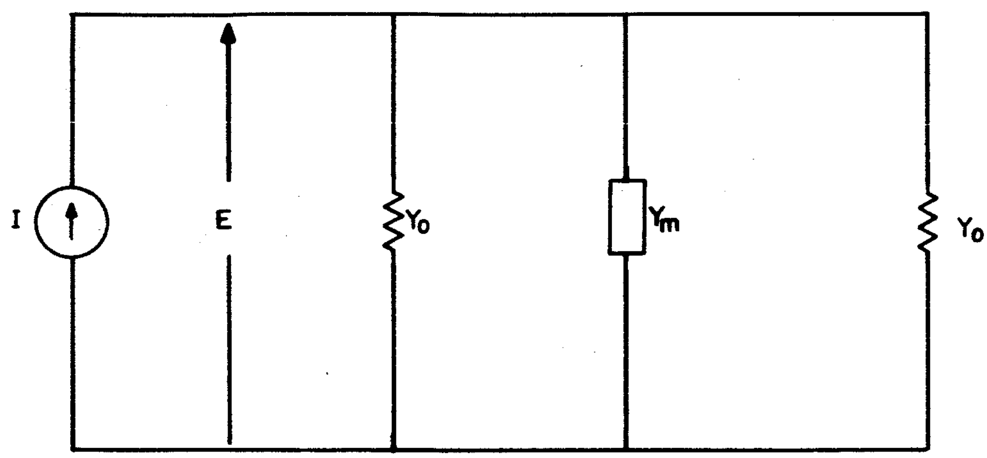


FIGURE 7 BASIC SHUNT DIODE MODULATOR AND ASSOCIATED TRANSMISSION LINE

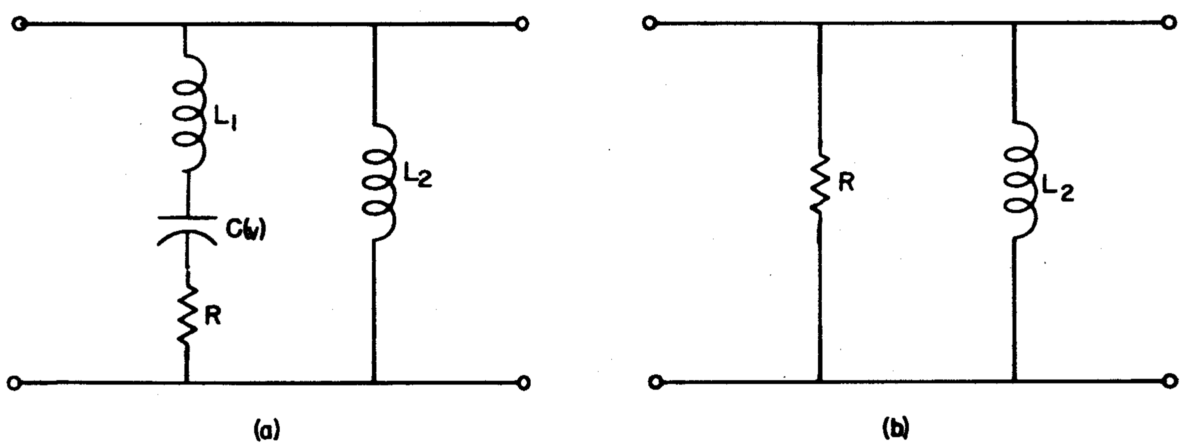


FIGURE 8 THE EQUIVALENT CIRCUIT OF A VARACTOR DIODE MODULATOR FOR
(a) MODE 2 OPERATION AND
(b) MODE 1 OPERATION

and

$$B = -\frac{R_s^2 + \left[X_{L1} - X_C\right]\left[(X_{L1} - X_C) + X_{L2}\right]}{X_{L2}\left[R_s^2 + (X_{L1} - X_C)^2\right]}, \tag{8}$$

respectively, where

X_{L1} = the inductive reactance of L_1,

X_{L2} = the inductive reactance of L_2,

and

X_C = the capacitive reactance of C(v).

From equation (8), it can be seen that the susceptance B is zero if

$$R_s^2 = -\left[\left[X_{L1} - X_C\right]\left[(X_{L1} - X_C) + X_{L2}\right]\right]. \tag{9}$$

Equation (9) is a quadratic in X_C with solutions

$$X_C = (X_{L1} + X_{L2}/2) \pm \sqrt{(X_{L2}/2)^2 - R_s^2}\ . \tag{10}$$

Equation (10) gives two values, X_{C1} and X_{C2}, for which the susceptance B equals zero. In particular, if $R_s \ll X_{L2}$ and $X_{L2} \gg 1$, approximate values are given by

$$X_{C1} = X_{L1} , \tag{11}$$

and

$$X_{C2} = X_{L1} + X_{L2} . \tag{12}$$

When $X_{C1} = X_{L1}$ the circuit is series resonant and maximum attenuation is obtained. When $X_{C2} = X_{L1} + X_{L2}$, the modulator impedance is a maximum and minimum attenuation is obtained.

With a low-level input signal and no d-c bias applied to the diode, $C_1(v)$ and L_1 form a series-resonant circuit. In this mode of operation, the r-f signal is either absorbed in R_s, or reflected, and maximum attenuation results. With the restriction that $X_{L1} = X_{C1}$ and $X_{L2} \gg 1$, equations (7) and (8) reduce to

$$G_1 = 1/R_s , \tag{13}$$

and

$$B_1 = - 1/X_{L2} \cong 0, \text{ respectively,} \tag{14}$$

where

G_1 = the conductance associated with $C_1(v)$,

and

B_1 = the susceptance associated with $C_1(v)$.

When reverse bias is applied to the diode, the capacitance C(v) is reduced, the series-resonant circuit is detuned, and parallel resonance occurs between $C_2(v)$, L_2, and L_1 as stated by equation (12). This mode of operation results in minimum attenuation and maximum transmission of r-f energy. For mode-two operation the conductance of the modulator remains unaltered and is given by equation (7). The susceptance, subject to the condition that $X_{L1} - X_{C2} = -X_{L2}$, becomes

$$B_2 = - \frac{R_s^2}{X_{L2}\left[R_s^2 + (X_{L1} - X_{C2})^2\right]} \cong 0 \, , \tag{15}$$

where

B_2 = the susceptance associated with $C_2(v)$.

The isolation of a shunt-diode switch results from the low impedance of a series resonance between L_1 and $C_1(v)$. From equation (6), isolations above 10 decibels in a transmission line with a characteristic impedance of Z_o, subject to the restriction that $(G/Y_o + 2) < B/Y_o$, are given approximately by

$$\alpha_1 = 20 \log_{10} Z_o/(2\ |X|\ ) \tag{16}$$

where

X = the series reactance which shunts the transmission line.

For a 50 ohm transmission line equation (16) becomes

$$\alpha_1 = 20 \log_{10} 25/\ |X|\ . \tag{17}$$

The solution for X in equation (16) yields:

$$X = Z_o/2 \times 10^{-\alpha_1/20} = 2\pi f L_1 - 1/2\pi f C(v)\ . \tag{18}$$

The solution for frequency in equation (18) is

$$f = \frac{X}{(4\pi L_1)} \pm \frac{1}{2\pi} \sqrt{X^2/(4L_1{}^2) + 1/\left[L_1\ C(v)\right]}\ . \tag{19}$$

For a given value of X, the isolation increases as the series-resonant frequency is approached and then decreases to the same isolation at - X. The bandwidth is the difference between the frequencies that correspond to X and - X. Hence, the substitution of - X for X in equation (19), followed by the subtraction of the corresponding frequencies for a given attenuation, yields the expression for bandwidth

$$\Delta f = f_1 - f_2 = X/(2\pi L_1) = \frac{Z_o \times 10^{-\alpha_1/20}}{2(2\pi L_1)} \quad . \tag{20}$$

It is seen that the bandwidth depends upon the characteristic impedance of the transmission line, the series inductance shunting the line, and the attenuation.

When operating in the second mode, the insertion loss of the modulator results from lossy elements within the parallel-resonant circuit and is due to low susceptance and negligible conductance. This insertion loss is usually more broadband than the isolation, and will be considered to be frequency independent.

The average power that a shunt modulator can isolate is determined by making Y_m as shown in Figure 7 equal to $1/R_s$. The incident power is

$$P_i = P_d + P_L \tag{21}$$

where

P_d = the power dissipated in the diode,

and

P_L = the power dissipated in the load.

Equation (21) may be rewritten as

$$P_i = P_d\ (1 + P_L/P_d) = P_d\ (1 + R_s/Z_o) \quad . \tag{22}$$

In a 50-ohm transmission line, a 2.5 ohm diode with a power dissipation rating of 500 milliwatts should be able to isolate approximately 525 milliwatts.

The maximum power-handling capability of the diode in the reverse-bias region is determined approximately by assuming that the quiescent bias equals one half of the breakdown voltage of the diode, and that the peak input signal amplitude is such that neither forward nor reverse conduction occurs. Hence,

$$P_i \text{ max} = (V_{rms})^2/Z_o = V_{br}^2/(8Z_o) \quad . \tag{23}$$

The maximum power that a varactor diode with a reverse breakdown voltage of 100 volts in a 50-ohm transmission line can dissipate is 250 milliwatts.

It should be noted that equations (22), (23), and similar expressions found in the literature, treat the varactor diode as a linear element and might lead one to expect optimum performance from the modulator at input power levels dictated by these equations. However, these expressions in no way relate isolation and insertion loss to the power input. In fact, it would seem as if the isolation and the insertion loss would vary as the inverse of the power input and directly as the power input respectively.

IV. GENERAL DESIGN CONSIDERATIONS

The first and most important component to be chosen in a varactor-diode modulator is the varactor diode. It is this element which controls such important factors as maximum isolation, minimum insertion loss, maximum power handling capability, and the desired bandwidth. An examination of the expressions relating the above-mentioned factors to various diode parameters such as; (1) the figure-of-merit (Q) of the diode at a specified operating voltage and frequency, (2) the capacitance at a specified operating voltage, (3) the maximum reverse-breakdown voltage, and (4) the ratio of maximum to minimum capacitance, will allow the design engineer to optimize the modulator from an examination of the electrical specifications of the diode.

An examination of equations (6) and (13) reveals that the maximum isolation of the modulator is a direct function of the conductance of the diode. To obtain a satisfactory isolation it is necessary that the series resistance of the varactor diode be small. For example, equation (6) shows that to obtain an isolation greater than 20 decibels in a 50-ohm line, the series resistance R_s must be less than 2.5 ohms. Larger values of R_s will correspondingly yield less isolation. Hence, the choice of a varactor diode with a small value of R_s (high Q) will yield maximum isolation.

In order to obtain a low insertion loss, equation (6) dictates that both the conductance (G) and susceptance (B) of the varactor diode be small. Since the insertion loss occurs when the modulator is at parallel resonance, equations (7) and (8) supply the necessary restrictions. It is evident from these equations that since a small value of R_s was chosen to yield sufficient isolation, a large value of X_{L2} must fulfill the condition for low insertion loss. Assume for example that the maximum allowable insertion loss is 0.5 decibel. If the inductance L_2 is chosen such that $X_{L2} \gg R_s$, Figure 9, then the insertion loss will be less than the allowable maximum. A closer examination of the modulator in the second mode of operation yields the information that $X_{L2} = X_{C2} - X_{L1}$, or a large value of X_{C2} will require a large value of X_{L2}. Since X_{L1} is set by the maximum value of capacitance at zero bias, it is necessary that the capacitance ratio be as large as possible to obtain a low insertion loss. Typical capacitance ratios are 3:1, 4:1, and 5:1.

The bandwidth in the isolation mode of operation is determined by the series shunting inductance L_1, the characteristic impedance of the transmission line, and the desired attenuation at the frequencies corresponding to the extremes of the bandwidth, for example, 10 decibels. It can be seen from equation (20) that the bandwidth is inversely proportional to L_1. Hence, a compromise must be reached between acceptable bandwidth and insertion loss.

The maximum power that a diode can safely dissipate in the series-resonant condition is a function of the series resistance, R_s, the characteristic impedance of the transmission line, and the rated power dissipation of the diode given by the manufacturers specifications.

The maximum power which can be dissipated in the parallel-resonant mode is determined primarily by the reverse-breakdown voltage of the diode.

To summarize, optimum modulator performance in a 50-ohm transmission line in which it is desired to transmit at least 500 milliwatts is obtained if a diode is chosen according to the following specifications:

1. Series resistance, $R_s \leq 2.5$ ohms.
2. Capacitance ratio, $C_{max}/C_{min} \geq 3.0$.
3. Power dissipation rating, $\geq$ 500 milliwatts.
4. Reverse breakdown voltage, $\geq$ 100 volts.

Once a diode is chosen, the next step is to choose L_1 such that $X_{L1} = X_{C1}$, where X_{C1} is the reactance of the varactor at zero bias. The maximum isolation is then calculated using equations (6), (13), and (14). The 10-decibel isolation bandwidth can then be calculated from equation (20). Next, choose the value of C_2 which occurs at one-half the reverse breakdown voltage of the varactor. This choice approximately maximizes the power input when the modulator is parallel resonant and also insures a low insertion loss. The value of L_2 is then calculated from the relation that $X_{C2} - X_{L1} = X_{L2}$.

The schematic diagram for the varactor-diode modulator is shown in Figure 9 along with calculated design parameters. The measured characteristics of a Pacific Semiconductor, Inc. type PC-116 varactor diode are presented with associated circuit constants in Table I. The actual modulator construction is shown in Figures 11 and 12.

TABLE I

MEASURED AND CALCULATED DESIGN

DATA FOR THE SHUNT MODULATOR

EMPLOYING A PSI PC-116 VARACTOR

DIODE AT 114.5 MEGACYCLES PER SECOND

PARAMETER	CALCULATED OR MEASURED VALUE
R_s*	2.5 ohms
C_{-11}**	14 picofarads
C_o*	48 picofarads
V_{br}*	- 100 volts
L_1	0.04 microhenry
L_2	0.096 microhenry
α_1	21 decibels
α_2	0.15 decibel
$\Delta f/f_o$ x 100%	15.5%

*Measured diode characteristics.

**The capacitance at -11 volts.

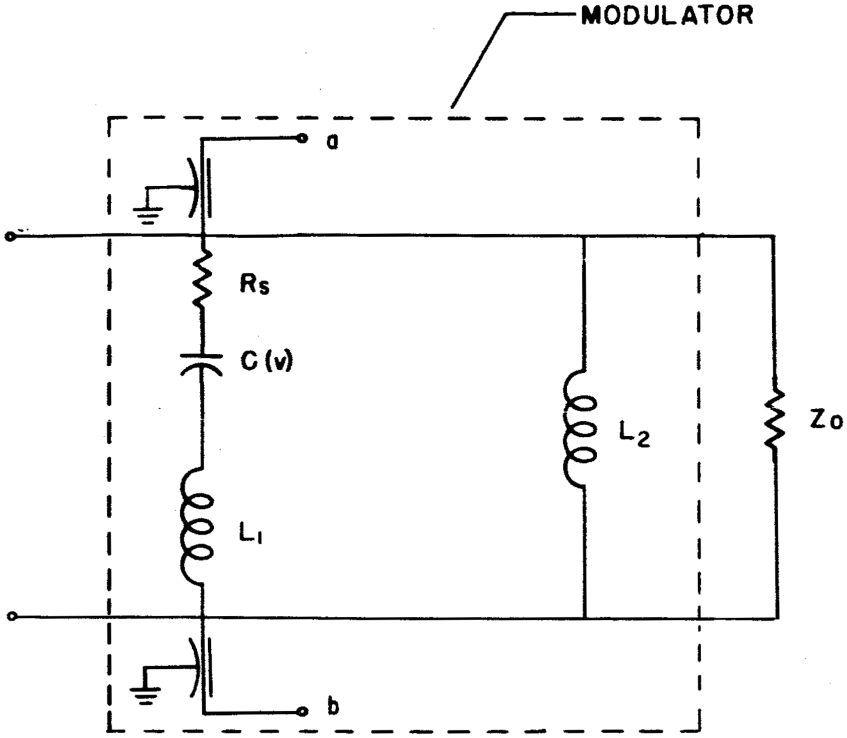


FIGURE 9 SCHEMATIC DIAGRAM OF VARACTOR – DIODE MODULATOR AND ASSOCIATED TRANSMISSION LINE

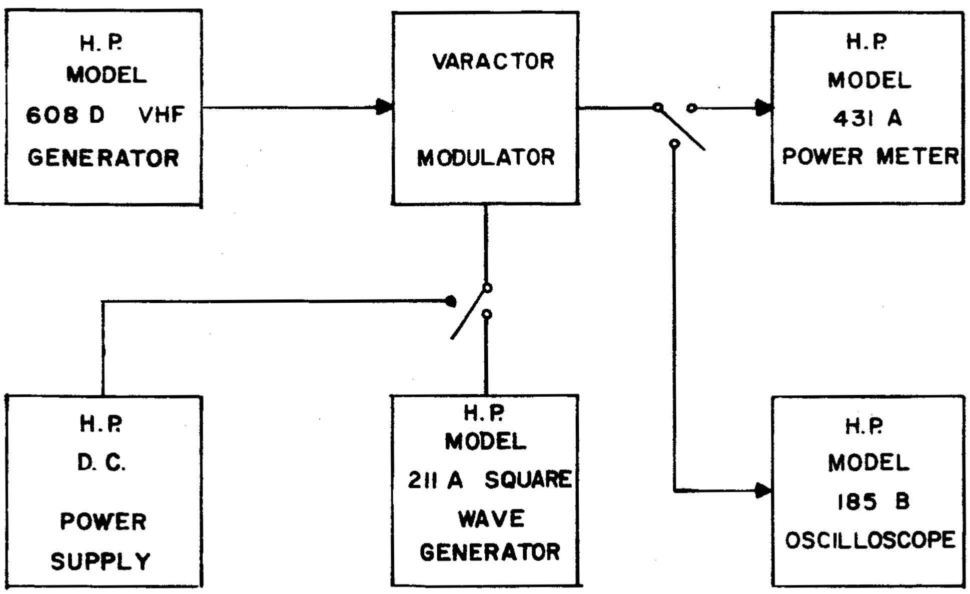


FIGURE 10 EQUIPMENT ARRANGEMENT USED TO OBTAIN EXPERIMENTAL DATA

V. EXPERIMENTAL RESULTS

The varactor-diode modulator shown in Figure 9 was designed and constructed according to the design procedure set forth in Chapter IV. The general circuit arrangement employed to obtain the experimental results is shown in Figure 10.

The isolation and insertion loss was measured as a function of frequency at a constant power input of 1.00 milliwatt using the equipment arrangement shown in Figure 10. Mode-one operation occurred at a bias voltage of + 0.08 volts and mode-two operation occurred at a reverse-bias voltage of - 11.0 volts. Equipment availability dictated that parallel resonance should occur at a reverse bias of - 11.0 volts rather than at - 50 volts (one half of the reverse breakdown voltage of the diode). The experimental results are shown in Figure 13. A theoretical maximum isolation was calculated from Equation (6) to be 21 decibels; the measured isolation was 23 decibels. The increase of two decibels in isolation may very well be attributed to the fact that the modulator was designed to provide maximum isolation at a d-c bias voltage of 0 volts whereas the optimum bias voltage was measured experimentally to be + 0.08 volts. This slight amount of forward bias could decrease the value of R_s below the measured 2.5 ohms, thereby increasing the maximum isolation. The maximum value of insertion loss at the center frequency was measured to be 0.08 decibels. The corresponding calculated value was approximately 0.10 decibels. To allow the comparison of the experimental results with theoretical expectations, equation (6) is plotted as a function of frequency and included in Figure 13. Figures 14 and 15 were obtained by sweeping the modulator with a Jerrold Model 900B Sweep Signal Generator. Figure 14 shows the attenuation characteristic of the modulator at a d-c bias voltage of + 0.08 volts. Figure 15 shows the same characteristic for a reverse bias of - 11.0 volts. In each case, the center frequency is 114.5 megacycles per second.

In mode-one when the modulator is series resonant, the isolation results from the low impedance of a series resonance between an inductor (L) and a capacitor (C). The bandwidth of the modulator in this mode of operation is inversely proportional to the Q of the series resonant circuit which is approximately 12.0. The measured 3 decibel bandwidth in Figure 13 agrees with computed data.

The insertion loss in the second mode of operation, is considerably more broadband than the isolation as depicted by Figure 13. The Q of the parallel resonant circuit is considerably lower than the Q of the series resonant circuit because of the heavy load imposed by Z_O. The calculated Q is approximately 0.7.

The maximum power that a varactor diode can safely dissipate in the series-resonant mode is given by equation (22). This expression does not necessarily give the amount of power the modulator can effectively isolate (above a specified minimum level), but specifies the maximum input power to the diode. The varactor-diode modulator analyzed in this study was designed using small-signal analysis and all previous results were obtained with an input power of one milliwatt. Equation (23) is

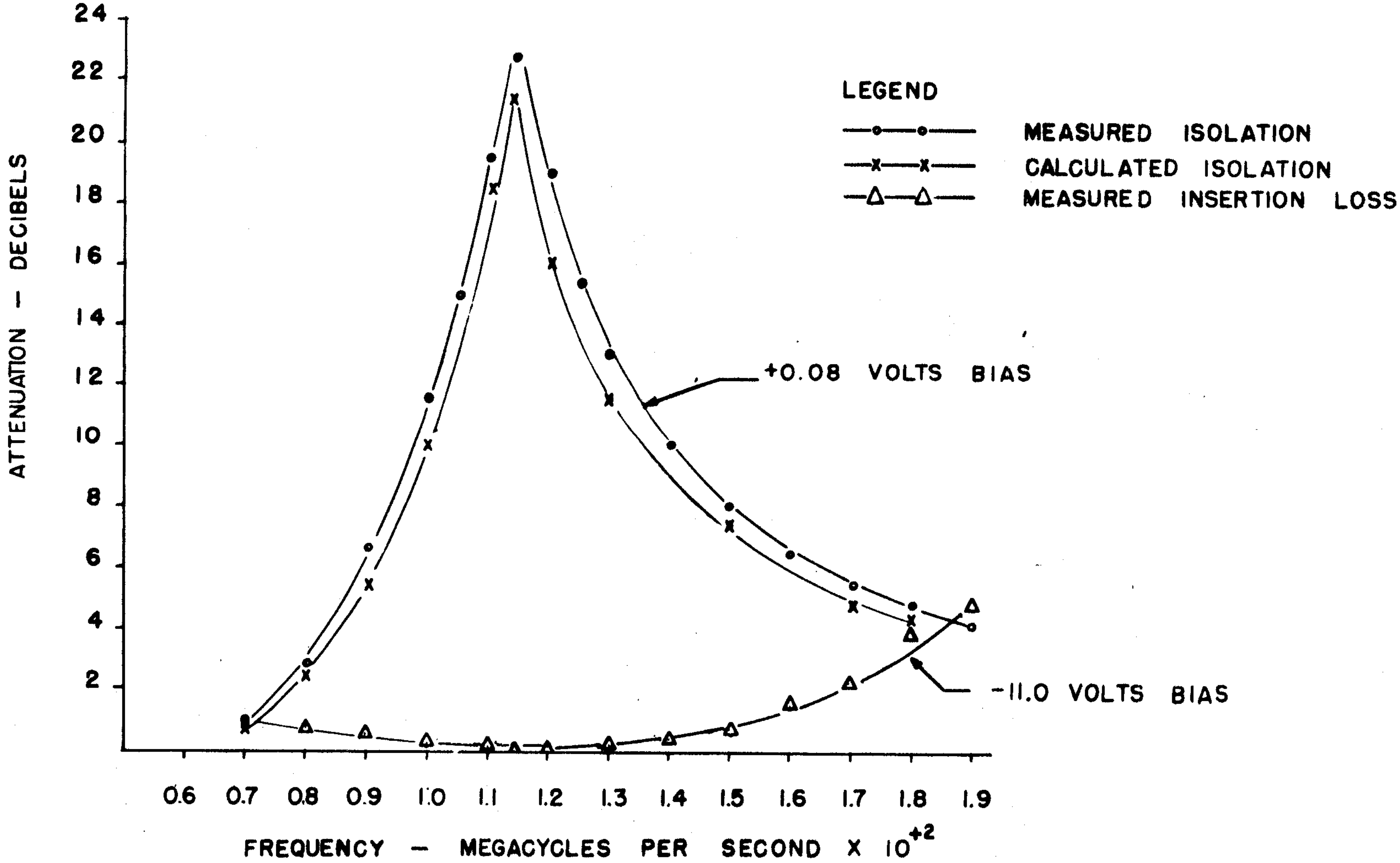


FIGURE 13 VARIATION OF ISOLATION AND INSERTION LOSS AS A FUNCTION OF FREQUENCY

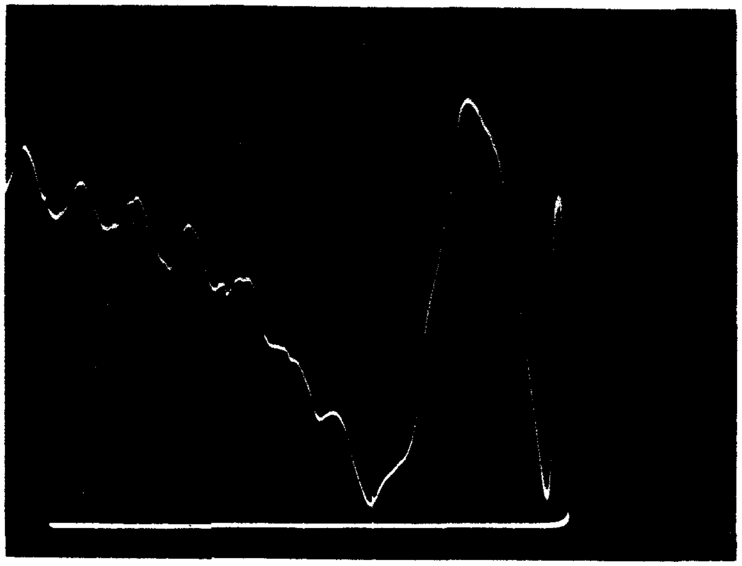


FIGURE 14 ISOLATION CHARACTERISTICS FOR MODE ONE SWITCHING

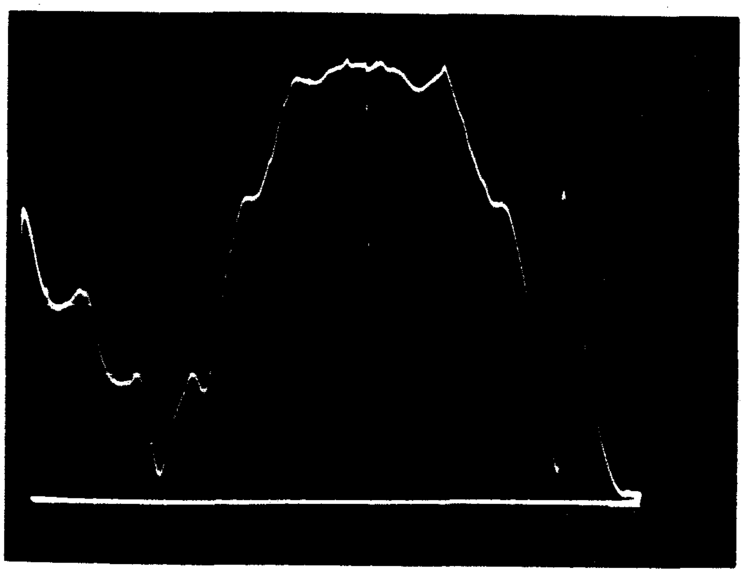


FIGURE 15 INSERTION LOSS CHARACTERISTICS FOR MODE TWO SWITCHING

an expression for the maximum power the varactor modulator can safely handle in the reverse-bias region.

CW power up to two watts was obtained from a Gonset "Communicator III" transmitter, at a frequency of 114.5 megacycles per second, and utilized to examine the operation of the modulator at higher power inputs. At power levels of two milliwatts the modulator behaved as in the low-level tests but, at higher incident power, the isolation decreased rapidly. It can be seen from Figure 16 that the isolation decreased from 22 decibels to 9 decibels as the input power was increased from one milliwatt to 100 milliwatts.

This effect can be attributed to the change in r-f impedance with increasing input power and is explained as follows. The forward contact potential of the varactor diode used in the modulator was measured to be + 0.5 volts and the d-c bias voltage across the varactor which corresponded to the maximum isolation condition was + 0.08 volts. Therefore, the amount of a-c voltage required to drive the diode into the forward conduction region is 0.42 volts peak. The calculated amount of power required to drive the diode into forward conduction is 2.02 milliwatts which corresponds to the breaking point shown in Figure 16. As the diode conducts in the forward direction, the time average capacitance increases, the reactance decreases, the series resonant circuit becomes detuned, and the maximum isolation decreases. Hence, forward conduction and decreased isolation occur at input power levels greater than two milliwatts.

It can be seen that the insertion loss incurred when the modulator

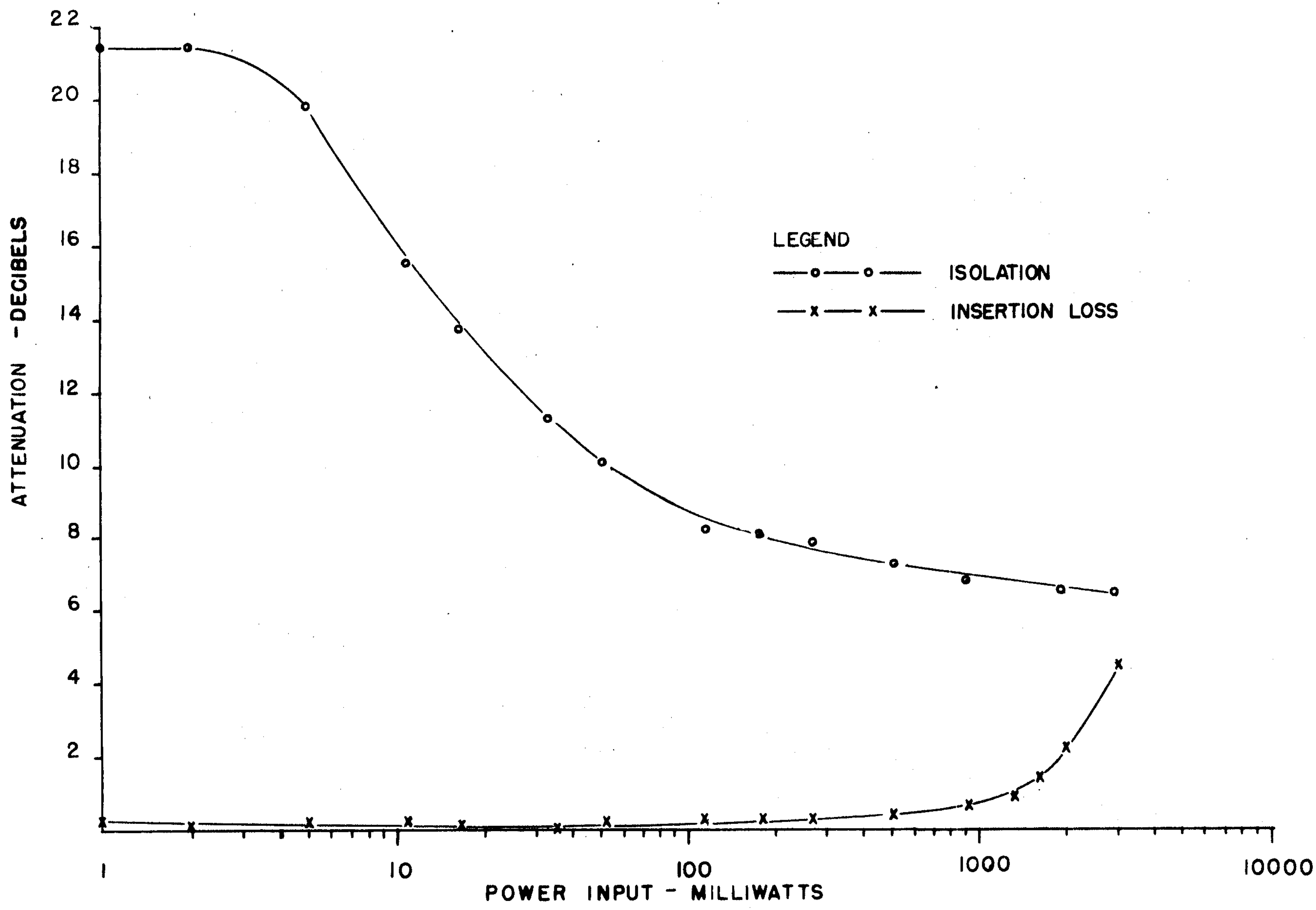


FIGURE 16 MEASURED ISOLATION AND INSERTION LOSS AS A FUNCTION OF INPUT POWER

is parallel resonant remains below one decibel at incident powers up to 1.2 watts then begins to increase rapidly. Essentially, this gradual and then rapid increase of insertion loss with increasing input power is due to the large amount of signal required to cause forward conduction and change the time average capacitance. The peak voltage required to drive the diode into forward conduction is 11.5 volts. The amount of power required to drive the diode into the conduction region was calculated to be 1.3 watts. From Figure 16 it can be seen that the curve begins to break at approximately 1.1 watts.

The modulation of the r-f energy was accomplished with the equipment arrangement shown in Figure 10. A Hewlett Packard Model 211 A Square Wave Generator was used to switch the modulator between the series and parallel mode at a frequency of 200 cycles per second and the resulting wave form was displayed on a Hewlett Packard Model 185B Sampling Oscilloscope. The result is shown in Figure 17. It is worth while to note at this point that the time required for the modulation voltage to alternate between 0 volts and - 11 volts is five milliseconds. The time required for the r-f energy to vary one complete cycle is approximately 8.75 nanoseconds. Hence, $5.72 \times 10^{+5}$ cycles of r-f energy should be contained within each modulation pulse. However, Figure 17 shows only four cycles of r-f energy within each modulation pulse. This particular phenomenon may be attributed to an inherent characteristic of the sampling oscilloscope which is that the apparent speed of the beam across the face of the cathode ray tube has no relation to its time scale in seconds per centimeter. The beam may take ten seconds to go across the face of the tube and yet the time scale could be thirty nanoseconds per centimeter. The amount of information displayed on the tube during the beam sweep time is a function of many items including the sampling rate.

To further illustrate the effect of forward conduction on the operation of the modulator, displays of the modulated r-f waveform were obtained for power inputs of one milliwatt and seven milliwatts and are shown in Figures 17 and 18 respectively.

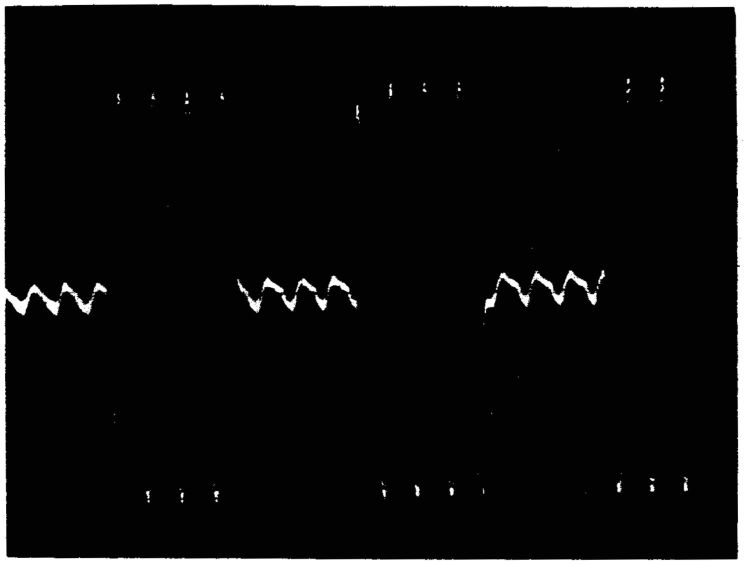


FIGURE 17 MODULATED WAVEFORM FOR AN INPUT POWER OF ONE MILLIWATT

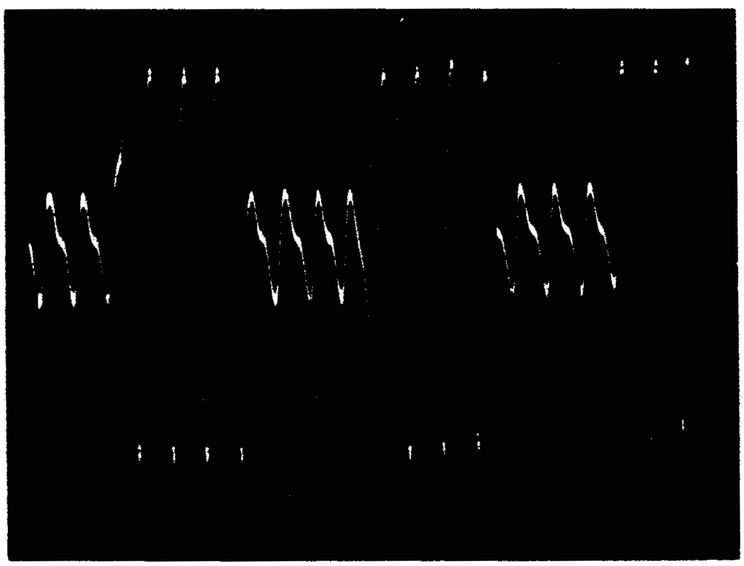


FIGURE 18 MODULATED WAVEFORM FOR AN INPUT POWER OF SEVEN MILLIWATTS

VI. CONCLUSIONS

A simple theory for the design of varactor-diode modulators has been developed and, in general, the results can be applied to any shunt modulator configuration employing a varactor diode as a tuning element. A procedure for selecting a varactor diode to yield optimum modulator performance is also discussed. The measured isolation, insertion loss, and bandwidth agree reasonably well with computed values.

The varactor modulator analyzed in this study provides an isolation in excess of 20 decibels and an insertion loss of less than 0.15 decibels at the design frequency of 114.5 megacycles per second. The insertion loss is less than one decibel over a bandwidth of 80 megacycles per second and the 3 decibel bandwidth in the isolation mode is approximately 8%. The maximum power input to the modulator when operating in mode-one is the power required to force the diode into the forward conduction region; any further increase in power results in decreased isolation. The operation of the modulator is completely independent of its position in the transmission line from the load and is dependent on its distance from the generator. The modulator requires no special techniques to achieve isolation greater than 20 decibels. The modulator has the added advantage that the characteristics between individual diodes can be compensated by adjusting the tuning elements. One of the principal disadvantages of this modulator is its low power-handling capability.

The varactor modulator could be very useful as an amplitude limiter for low-power transmitters. It is also ideal for applications in aerospace missiles because of the desired simplicity, compactness, light weight, and extremely low modulator power consumption.[10]

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REFERENCES

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