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THE MEASUREMENT OF HARMONICS PRODUCED BY A PIN DIODE IN A MICROWAVE SWITCHING APPLICATION

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ABSTRACT

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The design and construction of a microwave switch utilizing a PIN diode in a strip transmission line configuration are discussed.

Methods for measuring harmonics produced by the switch, in both switching states, are presented and the results of the measurements of second- and third-harmonic signals are given.

Factors influencing harmonic production are discussed.

CHAPTER I

INTRODUCTION

During recent years there has been an increasing interest in the development of semiconductor devices which may be used to control power flow in microwave transmission lines. Typical applications are in discrete step phase shifters, amplitude modulators, remotely controlled attenuators, antenna beam selectors for beam forming matrix, and duplexers. A number of designs are currently available which make use of the property of a suitable semiconductor diode that its microwave impedance is a function of the applied d.c. bias, and these designs offer the attractive features of small size, low power consumption and electronic control. This report is concerned with experimentally determining the harmonic power generated by a PIN diode in a typical microwave switching application.

Efficient switching is possible only if the diode exhibits a large variation of impedance between two suitable bias conditions. The two types of diodes which have been employed most successfully in the past are the varactor diode in which the junction capacitance is a function of reverse voltage and the PIN silicon junction diode in which the diode resistance is a function of the forward current.

The main feature which distinguishes the PIN junction diode from other types of diode is the semiconductor wafer material. This material has a low impurity concentration, and its behavior is substantially the same as that of the instrinsic semiconductor, i.e., the PIN diode is better described as a variable resistor than as a conventional diode. Its normal use is at a sufficiently high frequency that it does not rectify the applied signal and thus does not produce extremely large harmonics. This characteristic of the PIN diode depends upon the minority carrier lifetime being much longer than the period of the controlled signal.

The diode consists of a lightly doped, high-resistivity "I" region between N+ and P+ type layers. The "I" region consists of a depletion layer (swept free of carriers by built-in potential) and a lightly doped high-resistivity N type layer. High breakdown voltages can be obtained due to the "I" region width and high resistivity. The high breakdown voltage permits the control of RF power levels formerly unobtainable with semiconductor elements.

The power-handling capabilities of PIN diodes are determined by three factors, the ratio of the power controlled to the power dissipated, the power actually dissipated by the diode resistance, and the reverse breakdown voltage of the diode. The ratio of the power controlled to the power dissipated depends upon the losses existing in the two bias states used to perform the switching operation. The power dissipation is, of course, limited by the thermal resistance of the semiconductor, its cartridge, and the microwave diode mount. The peak-power rating is determined by either the diode breakdown voltage or the temperature rise during the pulse. A breakdown voltage of 700 volts corresponds to a peak RF power level of 1.2 kw for a 50 ohm line and a peak-to-peak voltage excursion equal to the breakdown voltage. In actual practice RF voltage excursions into the forward direction with applied reverse bias are possible without changing the diode characteristics; however, an increase in the harmonic power generated by the diode will occur.

The use of PIN diodes as semiconductor elements in microwave electronics has developed due to the two important properties of these diodes mentioned above, the ability to operate effectively in a high power environment, and the fact that very little harmonic power will be generated.

CHAPTER II

THEORY OF OPERATION

General Switching Theory

The requirement for a means to control in an efficient manner the transmission of relatively high microwave energy has existed since the development of radar. This requirement developed when it was recognized that one antenna system must be shared by a transmitter and a receiver to produce an effective radar system. In recent years the introduction of electronically scanned antenna arrays for use with radar and communication systems has greatly increased this requirement. One of the most basic control elements is the microwave switch, from which several more complex devices can be constructed. This chapter describes the work required to develop a semiconductor diode switch capable of controlling microwave energy on a transmission line operating in the TEM mode of propagation.

When considering a diode switch incorporated in a microwave transmission line it is necessary to define certain terms. RF power incident on an ideal attenuating device is either absorbed in or transmitted past the device, with no power reflected. The attenuation α of the device is defined as the ratio in decibels of the incident power to the transmitted power. The attenuation written in equation form is

$$\alpha = 10 \log_{10} \frac{P_i}{P_t} . \qquad (2\text{-}1)$$

If the attenuation of the device can be changed from some low value to some high value the device is called a switch. The insertion loss is defined as the minimum value of attenuation, and the isolation is defined as the maximum value of attenuation.

When a diode is inserted in shunt with a transmission line, RF power incident on the diode is reflected by, absorbed in, and transmitted past the diode. A diode switching element is different from an attenuating device in that most of the incident RF power not transmitted past the element is reflected rather than absorbed. In fact, in an ideal diode switch, the incident power is either completely transmitted or completely reflected. The definitions of attenuation, insertion loss, and isolation are the same for the diode switch as those for the absorption switch defined above.

From the discussion above it is apparent that an ideal diode switch is a single pole single throw element, and as such will produce reflection when in the "switch open" condition. The reflected energy must then be dissipated in the RF generator or some isolating device. To provide efficient switching operation it is necessary to incorporate the diode switch in a suitable microwave circuit. The circuit arrangement discussed below illustrates one of the applications of the diode switch.

Figure 1 is an RF schematic depicting the operation of two diodes and two 3 db couplers in a single pole double throw switching circuit. Energy entering the input port is divided equally between the arms A and B. The energy at B is retarded 90° in phase with respect to the energy at A. The diode switches in arms A and B are located the same distance from both couplers. If the diodes are in the low attenuation

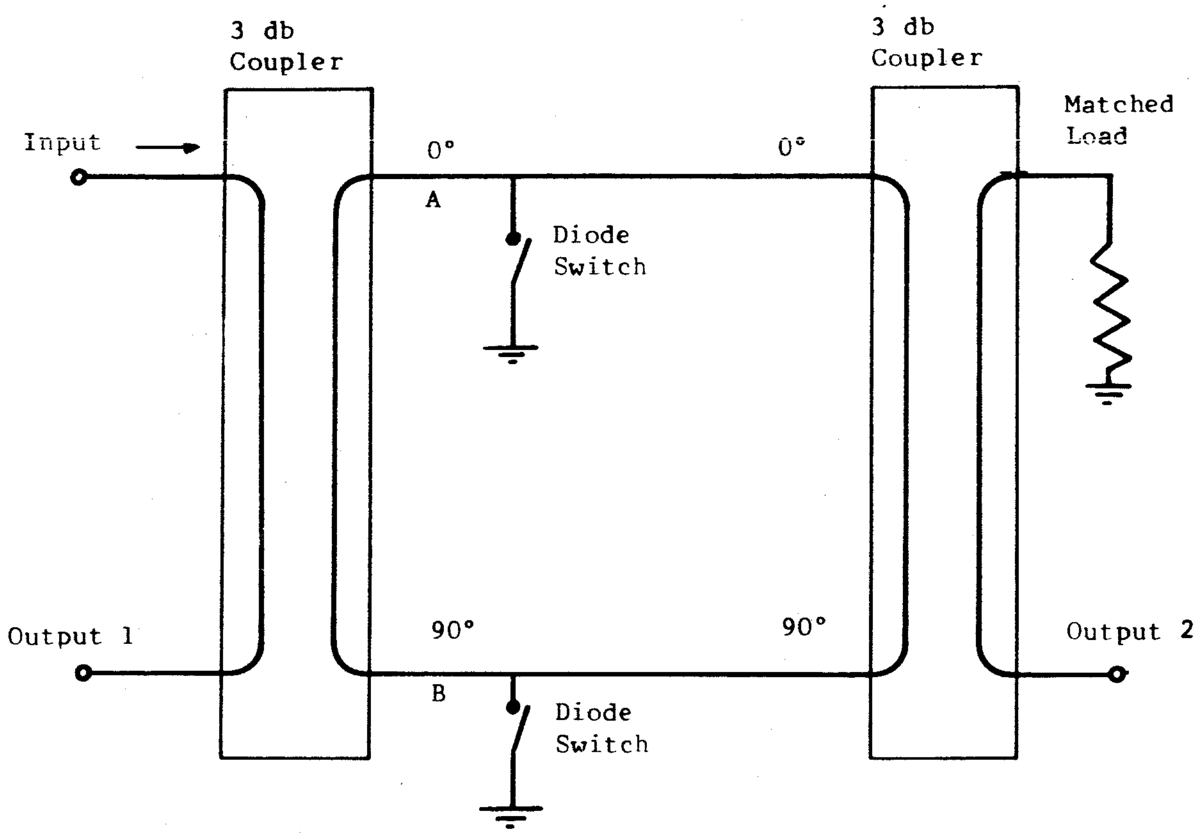


Fig. 1. The RF Schematic of a single pole double throw switch.

state energy will be transmitted past the diodes to the second coupler, where the energy in each arm will divide equally between the matched load and output port 2. However, the phase relation between the energy in arm A and arm B is such that energy cancellation will occur at the matched load, and the sum of the energies in arms A and B will be realized at output port 2. The other state of the switch will be when the diodes are reflecting the incident energy. In this state the phase is such that energy cancellation will occur at the input (no reflected energy on the input line) and energy combination will occur at output port 1. In neither state is energy reflected back on the input line nor is energy dissipated in the matched load. Theoretically this switch is a lossless device. As mentioned above this is one of many applications that require a microwave switch as the basic element.

A semiconductor diode, in shunt across the center and the outer conductor of a TEM wave transmission line, will provide attenuation. To derive an expression for the attenuation as a function of the impedance offered by the diode it will be necessary to use the equivalent circuit shown in Fig. 2. If the diode is considered to be of zero thickness and in a bilaterally matched transmission line, then the equivalent circuit will be valid for most calculations.[1] In the equivalent circuit I is the peak amplitude of the sinusoidal current source which is assumed to have an output admittance of Y_0, Y represents the diode admittance and L ($=Y_0$) is the matched load behind the switch. V is the peak amplitude of the resulting sinusoidal voltage. The power P_L dissipated in the load L is given by

$$P_L = \frac{1}{2} VV^* Y_0 \qquad (2\text{-}2)$$

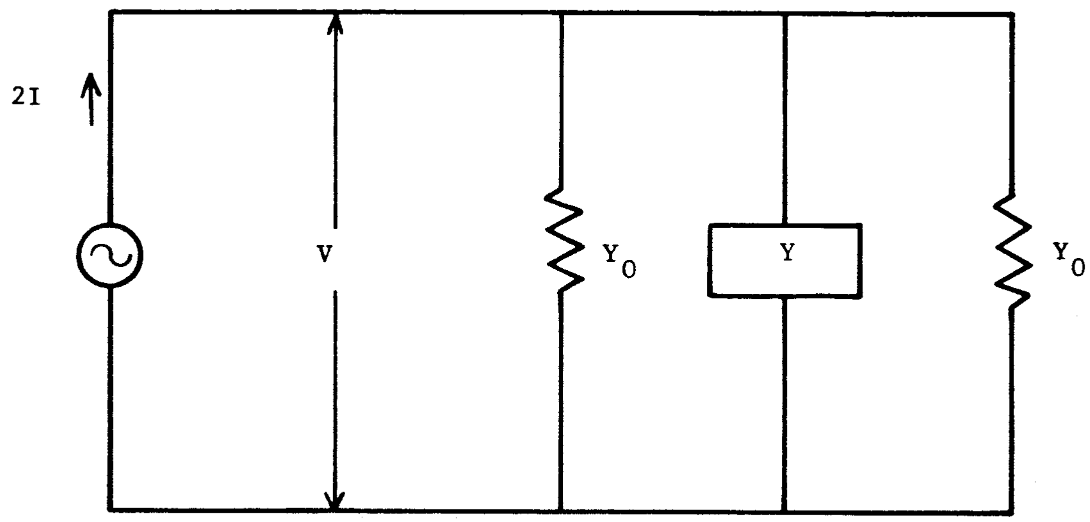


Fig. 2. The equivalent circuit of a diode with admittance Y in shunt in a transmission line of characteristic admittance Y_0.

For an admittance of $Y = G + jB$ the voltage V will be given by

$$V = \frac{2\,I}{G + jB + 2\,Y_0}$$

from which the power delivered to the load can be obtained as

$$P_L = \frac{2\,|I|^2\,Y_0}{|G + jB + 2Y_0|^2} \; . \tag{2-3}$$

This is also the power transmitted (P_t) past the diode. The power incident (P_i) on the diode is the power in the forward traveling wave going toward L. This is obtained from the equivalent circuit by setting Y equal to zero. Thus

$$P_i = \frac{|I|^2}{2\,Y_0} \tag{2-4}$$

and from Eq. 2-1 the attenuation will be

$$\alpha = 10\,\log_{10}\,\frac{|G + jB + 2\,Y_0|^2}{4\,Y_0^{\,2}} \tag{2-5}$$

which can be rearranged to yield

$$\alpha = 10\,\log_{10}\,\frac{(G/Y_0 + 2)^2 + (B/Y_0)^2}{4} \tag{2-6}$$

The quantities G/Y_0 and B/Y_0 are simply the normalized conductance and susceptance of the admittance shunting the transmission line. It should be restated that this analysis applies to an admittance (or impedance) shunting a matched transmission line and that the element producing the admittance must be small compared to the operating wavelength.

Smith Chart Contours of Constant Attenuation

As in most transmission line work, the Smith Chart has proven invaluable in the design of microwave switching circuits. In particular a Smith Chart showing contours of constant attenuation for a shunt impedance is extremely useful.[2] From Eq. 2-5 it is seen that contours of constant attenuation will be represented by

$$\left| G/Y_0 + 2 + jB/Y_0 \right|^2 = \text{constant.} \qquad (2\text{-}7)$$

The normalized admittance is related to the reflection coefficient by the equation

$$Y/Y_0 = G/Y_0 + j\,B/Y_0 = \frac{1 - P \exp j\theta}{1 + P \exp j\theta} \qquad (2\text{-}8)$$

where P is the magnitude and θ is the phase angle of the reflection coefficient. Throughout the rest of this thesis normalized admittance will be written with small letters, i.e.,

$$y = g + jb\,.$$

From Eq. 2-7 put

$$k = \left| g + 2 + jb \right|$$

then using Eq. 2-8

$$k = \left| 2 + \frac{1 - P \exp j\theta}{1 + P \exp j\theta} \right|$$

which reduces to

$$k = \left| \frac{3 + P \exp j\theta}{1 + P \exp j\theta} \right| .$$

Squaring k yields

$$k^2 = \frac{9 + 6P\cos\theta + P^2}{1 + 2P\cos\theta + P^2}$$

or

$$k^2 + 2k^2 P\cos\theta + k^2P^2 = 9 + 6P\cos\theta + P^2$$

and

$$P^2 - 2P \cos\theta \; \frac{3 - k^2}{k^2 - 1} + \frac{k^2 - 9}{k^2 - 1} = 0. \qquad (2\text{-}9)$$

From analytical geometry it is known that the equation in plane polar coordinates (P, θ) of a circle of radius r with the center at (b, 0) is given by

$$P^2 + b^2 - 2Pb\cos\theta - r^2 = 0. \qquad (2\text{-}10)$$

A comparison of Eq. 2-9 with 2-10 yields

$$b = \frac{3 - k^2}{k^2 - 1}$$

and

$$b^2 - r^2 = \frac{k^2 - 9}{k^2 - 1}$$

from which

$$r = \frac{2K}{k^2 - 1}.$$

Equation 2-9 represents a system of circles with origins at polar coordinates

$$P = \frac{3 - k^2}{k^2 - 1} \quad ; \quad \theta = 0$$

and with radii of

$$r = \frac{2K}{k^2 - 1}.$$

Although the above calculations are for a shunting admittance, the system of circles can be plotted for a shunt impedance by locating the centers at

$$P = \frac{3 - k^2}{k^2 - 1} \qquad \theta = 180° .$$

Figure 3 shows contours of constant attenuation, calculated from Eq. 2-9 for a shunt impedance, plotted on a Smith Chart.

Diode Power Handling Capability

The analysis of the maximum RF power capability of the diode switch requires a knowledge of the diode parameters. The RF equivalent circuit of a PIN diode and some typical values are shown in Fig. 4. The variable resistance R_i is a function of the d-c bias. The variation of R_i with bias can be described in the following terms.

With zero bias the bulk resistance of the I region will be between 8 and 12 K-ohms. This value depends upon the junction area, I region width, and resistivity. Under reverse bias, a depletion region develops and the dissipative losses associated with this region will be less than that of the I region. As reverse bias is increased, the depletion layer widens, and the losses that are associated with the I layer capacitance will decrease which corresponds to an increase in R_i. With about 50 volts reverse bias R_i will have increased to about 5 times its zero bias value. Under forward bias the reverse effect is observed. Conductivity modulation (produced by the bias current) in the I layer will cause R_i to drop rapidly with forward current. At a bias current of 100 ma, R_i will have a value of about 0.20 ohms.

One of two factors determine the maximum incident power the diode can switch. These factors are the reverse breakdown voltage (voltage required to produce approximately a 10 μa reverse current), and the maximum safe operating temperature of the semiconductor junction. Since in a switching configuration the diode will be in either of two bias states it is to be expected that the state with the lowest maximum

IMPEDANCE OR ADMITTANCE COORDINATES

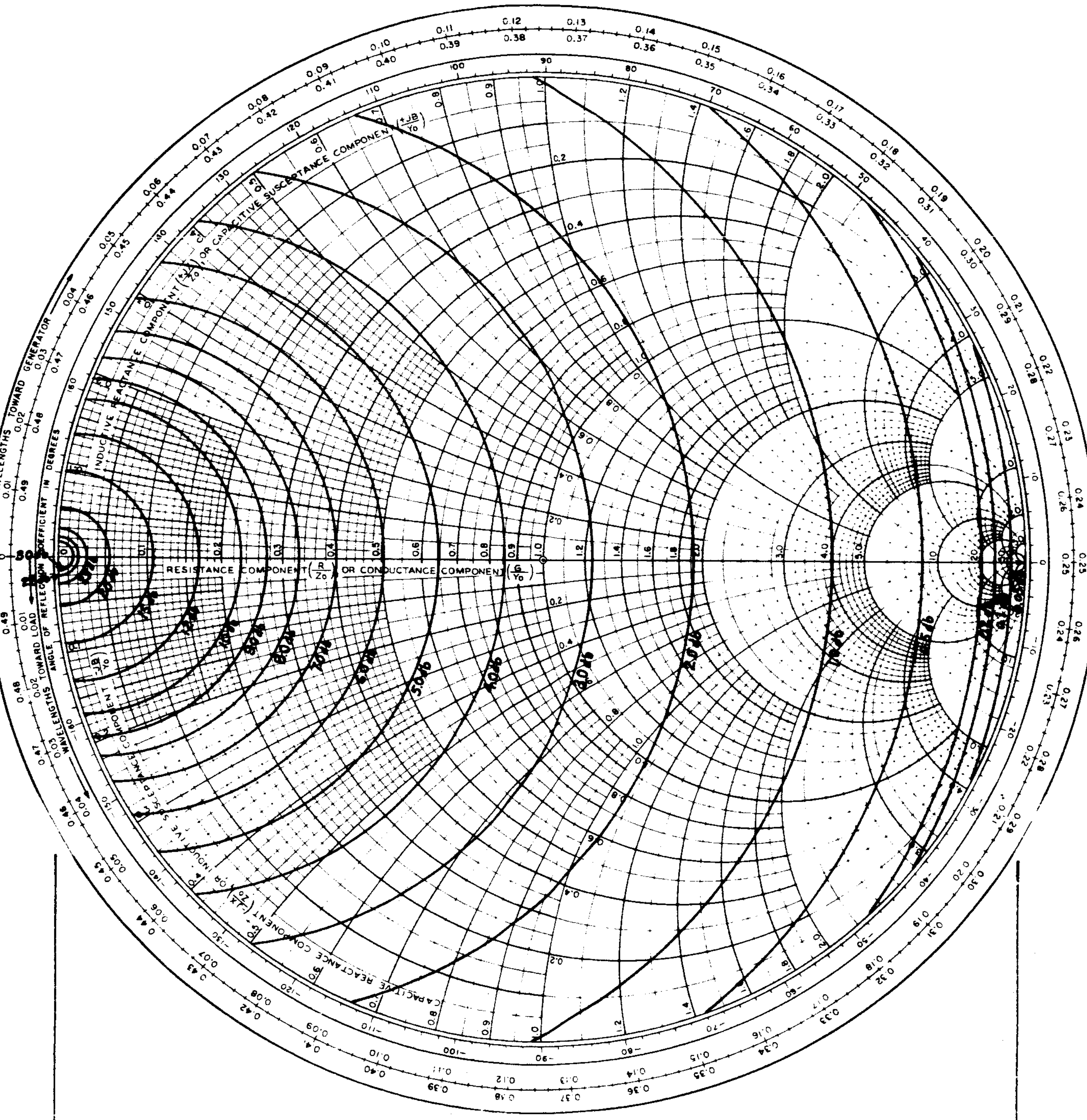
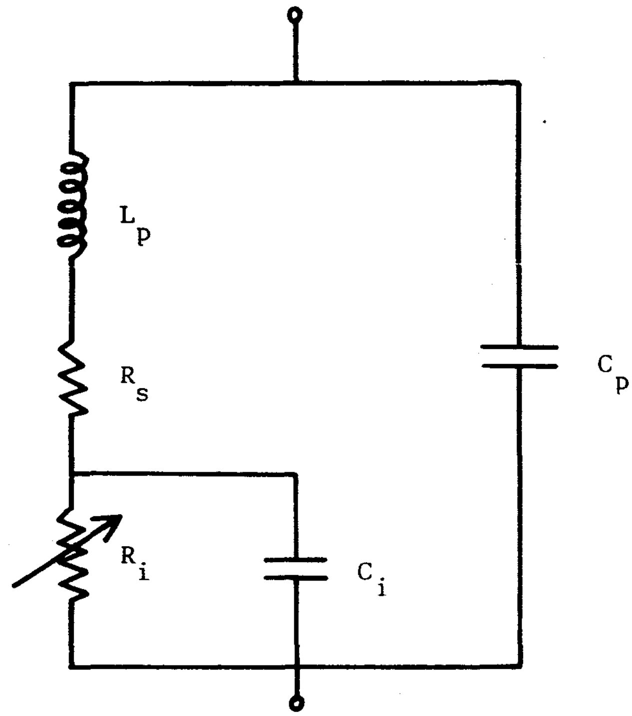


Fig. 3. The Smith Chart with Contours of Constant Attenuation Plotted for a Shunt Impedance.



L_p	- package inductance	$\approx$	2 nh
C_p	- package capacitance	$\approx$	0.3 pf
C_i	- capacitance across the I layer	$\approx$	.05 pf
R_s	- contact resistance	$\approx$	0.5 Ω
R_i	- resistance of the I layer		(see text)
V_b	- reverse breakdown voltage	$\approx$	700 volts

Fig. 4. Overall equivalent circuit of a PIN diode.

power limit will determine the maximum incident power that can be handled by the switch.

The reverse breakdown voltage of the diode limits the peak power the diode can control in the reverse bias state. Under forward bias the average incident power is limited by the maximum power that can be dissipated by the semiconductor junction.

To obtain the peak power which a diode in shunt with a transmission line can withstand when reverse biased, reference is made to Fig. 2, in which Y is replaced by C_b (an equivalent capacitance for the diode under reverse bias). The maximum RF voltage is just V, as shown in Fig. 2, from which

$$\overline{P}_L = \frac{1}{2} V^2 Y_0$$

but

$$V = \frac{V_b}{2}$$

for an RF voltage excursion from zero bias to V_b (assuming the diode is d-c reverse biased at $V_b/2$). Then

$$\overline{P}_L = \frac{V_b^{\,2}}{8Z_0} = \overline{P}_{iP} \quad . \qquad (2\text{-}11)$$

Equation 2-11 will yield the maximum peak power available at a matched load, this will also be the maximum incident peak power ($\overline{P}_{ip}$) if the power reflected by the diode is small ($\frac{1}{j\omega C_b} >> Z_0$) .

The maximum average power a shunting diode can switch is determined by making Y equal to $1/(R_s + R_i)$ in Fig. 2. As previously mentioned, the maximum operating temperature of the junction (around 150°C) and the thermal characteristics of the semiconductor material, its cartridge, and microwave circuit mount determine the power that can be safely dissipated by the diode. This power will be labeled $\overline{P}_d$. From Fig. 2 with

$$Y = \frac{1}{R_s + R_i} = \frac{1}{R_t}$$

and

$$V = \frac{2I}{2Y_0 + 1/R_t} = \frac{2\,I\,R_t}{2Y_0\,R_t + 1}$$

then

$$\overline{P}_d = \frac{1}{2}\,\frac{V^2}{R_t} = \frac{2I^2\,R_t}{(2Y_0R_t + 1)^2}\,. \qquad (2\text{-}12)$$

Solving Eq. 2-4 for $|I|^2$ and substituting into Eq. 2-12 yields

$$\overline{P}_{ia} = \frac{(2\,Y_0\,R_t + 1)^2}{4\,Y_0\,R_t}\,\overline{P}_d\,. \qquad (2\text{-}13)$$

For a typical diode R_t is less than 1 ohm and the transmission line characteristic admittance is less than 0.1 mho, thus the squared term can be assumed equal to unity so that

$$\overline{P}_{ia} = \frac{Z_0}{4\,R_t}\,\overline{P}_d\,. \qquad (2\text{-}14)$$

Equations 2-11 and 2-14 define the maximum peak power and the maximum average power that can be switched without exceeding the safe operating limits of the diode. In practical applications the peak power limit is somewhat higher since RF voltage swing into the forward bias region can occur without appreciable rectification, however some degrading of the switching characteristic will be noted (an increase in the insertion loss and the harmonics generated).

To determine the highest CW power that a diode can switch requires that the transmission characteristic impedance be varied until the peak power rating ($\bar{P}_{ip}$) and the average power rating ($\bar{P}_{ia}$) of the diode switch are equal. From Eq. 2-11 and 2-14 set

$$\bar{P}_{ip} = \bar{P}_{ia}$$

which is

$$\frac{V_b^{\,2}}{8Z_0} = \frac{Z_0}{4\,R_t}\,\bar{P}_d$$

or

$$Z_0 = V_b\sqrt{\frac{R_t}{2\,\bar{P}_d}} \qquad (2\text{-}15)$$

and

$$P_{CW} = \frac{V_b}{8\sqrt{R_t/2\,\bar{P}_d}} \qquad (2\text{-}16)$$

A typical PIN diode with

$R_t = 1$ ohm $\qquad \bar{P}_d = 5$ watts

$V_b = 500$ volts

can switch 200 watts of CW power in a 160-ohm transmission line.

A diode switching element operating in a pulsed power environment can be designed to handle a maximum peak power by varying the transmission line characteristic impedance until the average power rating ($\bar{P}_{ia}$) of the diode is equal to D times the peak power rating ($\bar{P}_{ip}$). The quantity D is the duty cycle of the pulsed power. Proceeding as above let

$$\bar{P}_{ia} = \bar{D}\,P_{ip}$$

or

$$\frac{D\,V_b^{\,2}}{8\,Z_0} = \frac{Z_0\,\bar{P}_d}{4\,R_t}$$

then

$$Z_0 = V_b\sqrt{\frac{D\ R_t}{2\ \overline{P}_d}} \tag{2-17}$$

and

$$P_{pp} = \frac{V_b}{8\sqrt{\frac{D\ R_t}{2\ \overline{P}_d}}}\ . \tag{2-18}$$

Consider a commercially available PIN diode with the following parameters

R_t = 1 ohm $\qquad$ $\overline{P}_d$ = 2.5 watts $\qquad$ V_b = 1000 volts,

operating in a pulsed power environment having a duty cycle of 10^{-3}. This diode will be capable of switching 9 K-watts of RF power in a 14-ohm transmission line.

Some Design Considerations

A number of factors must be considered in the design of a satisfactory diode switch. The three most important are 1) the required isolation and maximum insertion loss, 2) the maximum peak and average power to be switched, and 3) the operating bandwidth. These factors are functions of the characteristic impedance of the transmission line, and thus the selection of the line characteristic impedance must be based on a consideration of these factors. As an example, it was shown above that a PIN diode can switch 9 K-watts of peak power with a duty cycle of 10^{-3} in a 14-ohm transmission line. However, the normalized shunt resistance will be

$$r = 1/14 = .071$$

which corresponds to (from Fig. 3) a maximum isolation of 18 db. In a 50-ohm transmission line the same diode can achieve 28 db isolation at a maximum peak power of 2.5 K watts.

It is recognized from Fig. 3 that the maximum isolation obtainable with an impedance in shunt across a transmission line occurs when the resistive component of the impedance is realized across the line. The equivalent circuit of the PIN diode (Fig. 4) includes reactive elements that possess sufficient reactance at microwave frequencies to impair efficient switching operation. The finite sizes of the diode and the microwave mount also introduce reactive effects. To realize only the resistive component across the line requires a tuning element to effectively cancel the reactance. The tuning element can consist of appropriate lengths of open or shunted transmission line, or it can consist of lumped capacitance and inductance incorporated in the diode mounting structure. In either case the behavior of the tuning element with frequency will determine the effective operating bandwidth of the switch. Figure 3 clearly shows the effect of a reactive component of the shunt impedance on the isolation and the insertion loss. The insertion loss is seen to be much less dependent on reactance than is isolation, hence it is to be expected that the bandwidth in the "switch on" condition will be much greater than in the "switch off" condition.

CHAPTER III

ELECTRICAL DESIGN REQUIREMENTS

The Transmission Line

In the range of a few hundred to several thousand megacycles, printed microwave components and lines possess a number of advantages over conventional coaxial or waveguide structures. First, they allow the laboratory design and fabrication of components without the requirement for a considerable amount of machine shop work and its associated time delay. Second, the printed version results in considerable space and weight saving in the frequency range mentioned. Finally the simplified construction of the printed version results in lower costs. Having considered the advantages listed above for the printed or strip transmission line, against the advantages of other types of TEM transmission lines, it was decided to construct the switch in a strip transmission line.

The flat-strip coaxial transmission line was first used by V. H. Rumsey and H. W. Jamison during World War II as a feed structure in an antenna system for the Navy. Robert Barnett, from the Air Force Cambridge Research Center, early in 1949 used this new type of flat-strip coaxial line for many experimental filters, directional couplers, matched loads, 3-db hybrids, and various other circuit elements. This work was reported at the Dayton IRE Meeting in 1951.[3]

The planar transmission system upon which the microwave printed circuit technique is based results from an evolution of the coaxial transmission system. This evolution process can be seen by referring to Fig. 5.

If the coaxial line is deformed in such a manner that both the center and outer conductors are square or rectangular in cross section and then if the side walls of the rectangular coaxial system are extended to infinity, the resultant "strip" transmission system while possessing all of the advantages of the coaxial system, now has a form factor which is adaptable to the printed circuit technique.[3]

The planar transmission system can be readily adapted to printed circuit techniques by using two sheets of solid dielectric as spacers between the outer conductors which are plated to the dielectric, the center conductor being sandwiched between these sheets. The center conductor can be printed on one of the dielectric sheets in a number of ways; however, experimental work can be readily accomplished by using thin metal foil with an adhesive coating on one side. The center conductor can be cut to the required width with a razor blade or other sharp instrument. Since the characteristic impedance of the system is a function of the center conductor width, it is readily adaptable to circuits requiring impedance changes.[3]

The characteristic impedance of a double ground plane printed strip (strip-line) is determined by the width and thickness of the center strip, the ground-plane to ground-plane spacing, and the dielectric constant of the material. A theoretical derivation of the equation used in determining the characteristic impedance of strip transmission lines is given in the paper by Barnett.[3] A choice of dielectric material necessarily determines the ground-plane spacing and the dielectric constant, leaving the strip width as the only variable. The equation

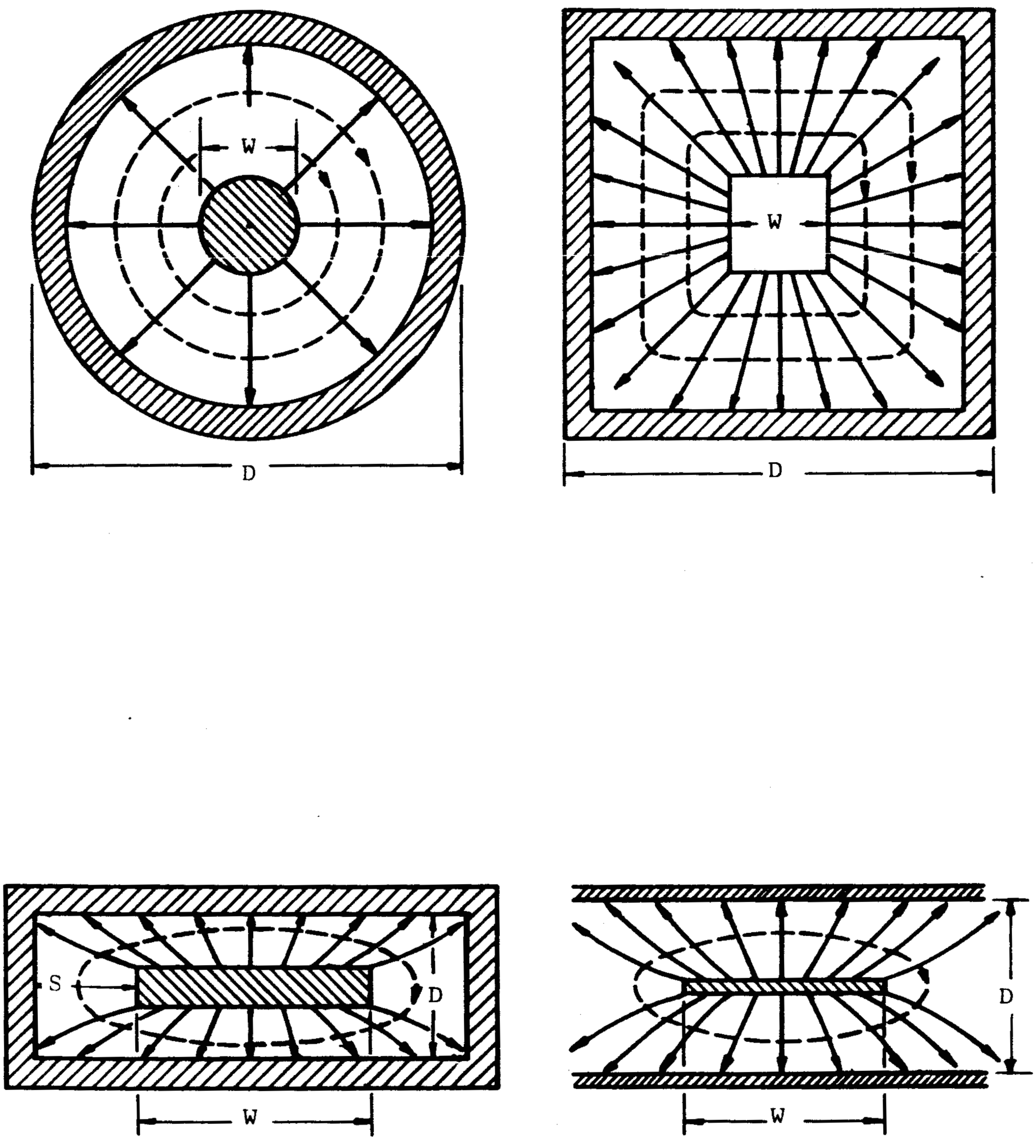


Fig. 5. The evolution of the flat strip transmission line.

relating characteristic impedance to strip width is furnished by the manufacturer of commercially available copper clad dielectrics.

In order to utilize standard coaxial line test equipment to measure the electrical characteristics of strip-line and strip-line components, it is necessary to develop a broadband coaxial line to strip-line adapter. In most practical applications care must be taken in effecting a transition from regular coaxial transmission line to the flat strip transmission line in order that higher order modes are not excited. For large strip line sizes, an adapter in which the strip-line and coaxial line lie along the same axis leads to simple adapters with excellent broadband characteristics.[4]

The Symmetrical Tee Junction

It will be of considerable importance to examine the equivalent circuit of one form of discontinuity in the balanced strip transmission line. This discontinuity, a strip-line tee junction, is required to tune out the reactive component of the diode and mount, and to provide a d-c bias return path. The manner in which the tee junction will be used to accomplish this will be discussed in Chapter IV.

Before discussing the strip-line tee junction, it is desirable to mention certain general qualitative considerations. A balanced strip transmission line denotes one in which the center conductor is equally spaced between the outer conductors. Discontinuities in balanced strip transmission line will possess purely reactive equivalent networks if, 1) the discontinuity is balanced, 2) the distance D (Fig. 5-d) is less than $\lambda/2$, and 3) the discontinuity structure contains no dissipative elements. If the discontinuity structure is unbalanced the radial transmission line dominant mode (similar to the TEM mode in a parallel plat transmission line) will be excited, and radiation will occur producing resistive contributions to the equivalent circuit. A small amount of unbalance can be tolerated and the radiation prevented by boxing in a small region surrounding the discontinuity (using shorting pins to maintain the outer conductors at the same potential) so that the unwanted excited mode becomes the dominant mode in rectangular waveguide, and by choosing the dimensions such that the mode is below cutoff.[5] Even a balanced discontinuity will excite a variety of higher radial line modes, the lowest of these being similar to the TM_1 mode in a parallel plate line. None of these modes will propagate if the length D (Fig. 5-d) is less than $\lambda/2$. Therefore, although the propagation of the dominant strip transmission line mode (TEM) does not impose any conditions on the outer conductor spacing, a practical system must have a separation distance D less than $\lambda/2$.

The use of the symmetric tee junction (shown in Fig. 6) is required in a wide variety of strip transmission line microwave circuits. The ability to realize accurate electrical design of these strip-line components requires precise knowledge of the parameters which represent the tee junction.

The equivalent circuit representation of the strip-line symmetrical tee junction used in Chapter IV of this report is one recommended by Franco and Oliner.[6] Figure 7(a) shows the top view of the center conductor at the tee junction with appropriate reference planes noted. Fig. 7(b) is the equivalent electrical circuit. The evaluation of the various circuit elements is accomplished by referring to graphs and equations given in the paper by Franco and Oliner.[6]

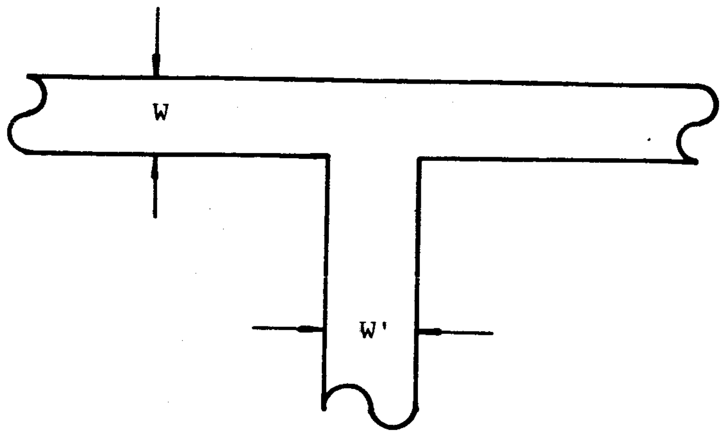
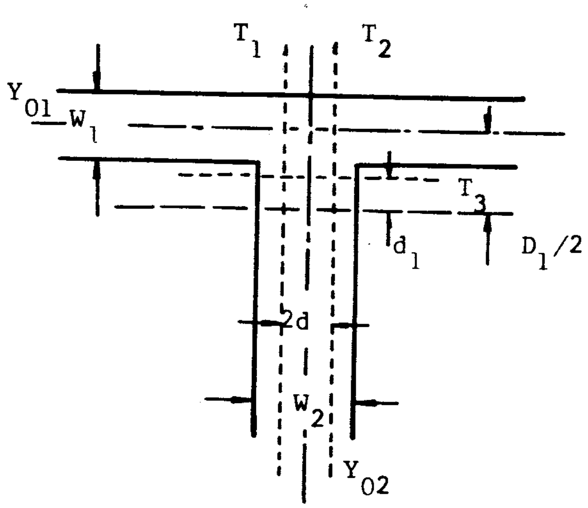
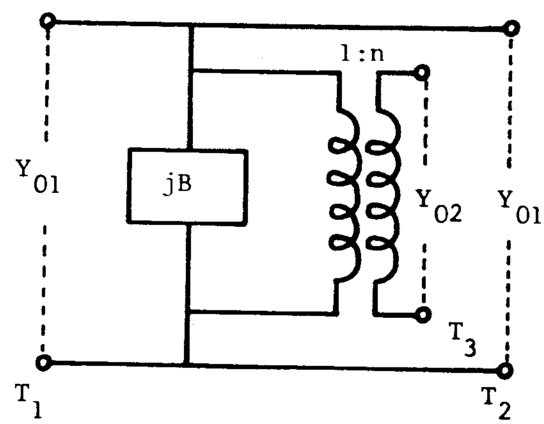


Fig. 6. The strip line symmetrical tee junction (top view of the center conductor).



(a) The tee junction with reference planes noted.



(b) The equivalent circuit

Fig. 7. The equivalent circuit for the strip line symmetrical tee junction.

At this point sufficient theory has been developed and general design criterions discussed to enable construction of the diode switch. A discussion of bias techniques and the diode mount structure is conspicuously absent. The reason is that to a large extent the effect of the diode mount and bias by-pass circuitry cannot be predicted mathematically. The switch design as discussed in the next chapter does not require an exact determination of the individual effects of the diode mount, but only an equivalent circuit representing the total effect. This equivalent circuit is determined by standard microwave measurements.

CHAPTER IV

THE PHYSICAL DESIGN

Material for the Strip Line Construction

The strip transmission line circuit was constructed using two copper clad Tellite 3B dielectric sheets manufactured by Tellite Corporation of Orange, New Jersey. The dielectric sheets can be purchased with one or both sides copper clad. The sheets purchased were .125" thick with a 2 ounce copper plating on one side. Some of the characteristics of this material listed in the manufacturer's specification are given below.

Item	Value
Dielectric Constant	2.32 $\pm$ .01 to 4.3 Gc
Dissipation Factor	.00015 to 10 Gc
Dielectric Strength	466 volts/mil.
Operating Temperature	
Continuous	220°F
Intermittent	500°F
Peel Strength	3 lb/inch

The ground plane spacing (D in Fig. 5-d) using this material is .25" neglecting the .002" thickness of the center conductor. The equation relating center conductor width to transmission line characteristic impedance is given by the manufacturer for Tellite 3B as

$$W = \frac{15.07}{Z_0} - .114 \text{ inches.}$$

The center conductor was cut from Permacel EE 3990 copper foil tape manufactured by Permacel, New Brunswick, New Jersey. The dielectric constant and dissipation factor of the adhesive portion of the tape had no noticeable effect on the circuit performance. The copper foil has a thickness of .002" and a weight of 1 oz/sq.ft.

A modified Keuffel and Esser Circuit Path Cutting Tool (56-1031) contained in the KE 56-1291 Stabiline Cut 'N' Strip Tool Kit was used to cut the copper foil tape to the correct width. The bevel of the cutting edge of the blade was increased to eliminate drag and a spacer was ground to obtain the desired spacing between the two cutting edges.

The coaxial-to-strip-transmission-line adapter developed for the switch is a modification of the one described by Graven.[7] The modification consists of eliminating the slot milled in the lower surface of the upper dielectric sheet. This was accomplished by reducing the center conductor pin to .010" rather than .031" and increasing the rectangular flange spacing to .265". The electrical connection from the center conductor of the adapter to the strip-line center conductor results from the pressure exerted by the flanges when the securing nuts are tightened. The tongue of the adapter center conductor forms a .005" channel in both the upper and lower dielectric sheets under this pressure. The input VSWR of a section of 50-ohm strip transmission line using two adapters (input and output) and a 50-ohm coaxial load was less than 1.02 over a 10% frequency range centered at 1.61 Gc.

Design Specification

The switch was designed as one phase of an effort to determine the harmonic generation by a PIN diode in a typical switching operation. With this purpose in mind the following design specifications were determined.

ITEM	VALUE
Peak Power	1 K-watt
Average Power	1 watt
Duty Cycle	10^{-3}
Pulse Length	1.4 μ sec.
Insertion Loss	$\leq$ .1 db
Isolation	$\geq$ 28 db
Frequency	1.61 Gc
Bandwidth	1%

To meet this specification a Microwave Associates MA-457162 PIN diode was selected to operate in a 50 ohm strip transmission line with a .25" ground plane spacing. The important parameters for this diode are listed below.

ITEM	VALUE
V_b	760 volts
R_t	.66 ohms at 100 ma bias current
C_t	.59 pf at -50 volts bias
θ_t	14° C/watt

The maximum peak power that can be switched is 1.44 K-watts (from Eq. 2-11). Using an equivalent thermal resistance for the diode and mount structure of 25°C/watt and a maximum safe diode junction operating temperature of 150°C, it is found that $\overline{P}_d$ (Eq. 2-14) is 5 watts. The maximum average power that can be switched by this diode is 95 watts (Eq. 2-14).

The isolation is determined by the use of Eq. 2-6. It is assumed that only the resistive component of the diode (reactance tuned out) is in shunt across the transmission line, Eq. 2-6 will yield

$$\alpha = 10 \log_{10} \frac{(50/.66 + 2)^2}{4} = 31.8 \text{ db.}$$

It is impossible to determine a theoretical value for the insertion loss (using Eq. 2-6) since the reactive effects associated with the diode and mounting structure are not known. However, it is expected that a tuning circuit will make possible an insertion loss of less than .1 db.

To measure accurately physical distances and to relate these distances to fractions of a wavelength in the strip transmission line, requires accurate calibration of the coaxial-to-strip-line adapter. The calibration procedure was designed to measure the length of line between a slotted line reference plane (short) and the edge of the strip-line dielectric sheet. This was accomplished by constructing a short on the strip-line a known distance from the edge of the dielectric sheet. The difference in position of the minimum produced by the reference plane short and that produced by the strip-line short is used to determine the effective length of the adapter. Let Y be the distance between the minimum points measured towards the generator, λ_a be the wavelength in air, and λ_d be the wavelength in the strip-line, then

$$\frac{Y}{\lambda_a} + \frac{X}{\lambda_d} = \frac{1}{2} .$$

The length X is the distance from the slotted line reference plane to the first minimum on the strip-line. The position of the first minimum on the strip-line is determined from the position of the strip-line short and λ_d. The procedure discussed above was performed for three positions of the strip-line short and the results averaged to obtain the final result. The effective length (in terms of length of strip-line) was determined to be 1.67". The values of λ_a and λ_d are 7.34" and 4.84" respectively. An impedance measurement made on the slotted line can be transformed to any point on the strip-line using this length and the distance from the point in question to the edge of the dielectric sheet.

The Diode Mount

The microwave diode mount must be designed to perform two functions; 1) the diode must be rigidly supported such that physical contact is maintained between one terminal of the diode and the center conductor of the strip-line, and 2) the mount must provide an RF short between the second diode terminal and one ground plane while maintaining an open circuit condition to the d-c bias supply. A cross-sectional view of the diode mount designed to perform these functions is shown in Fig. 8. The main body of the mount is a Gremar 5760 BNC female connector. The d-c isolation is achieved by the two mica capacitors and the brass disk. The brass disk is electrically connected to the second diode terminal and forms one plate of the capacitor as well as a heat sink and spacer for the diode. This by-pass capacitor maintains the second diode terminal at RF ground potential and prevents RF propagation along the bias line. The two brass screws, insulated from the brass plate, secure the mount to the transmission line and maintain the lower ground plane

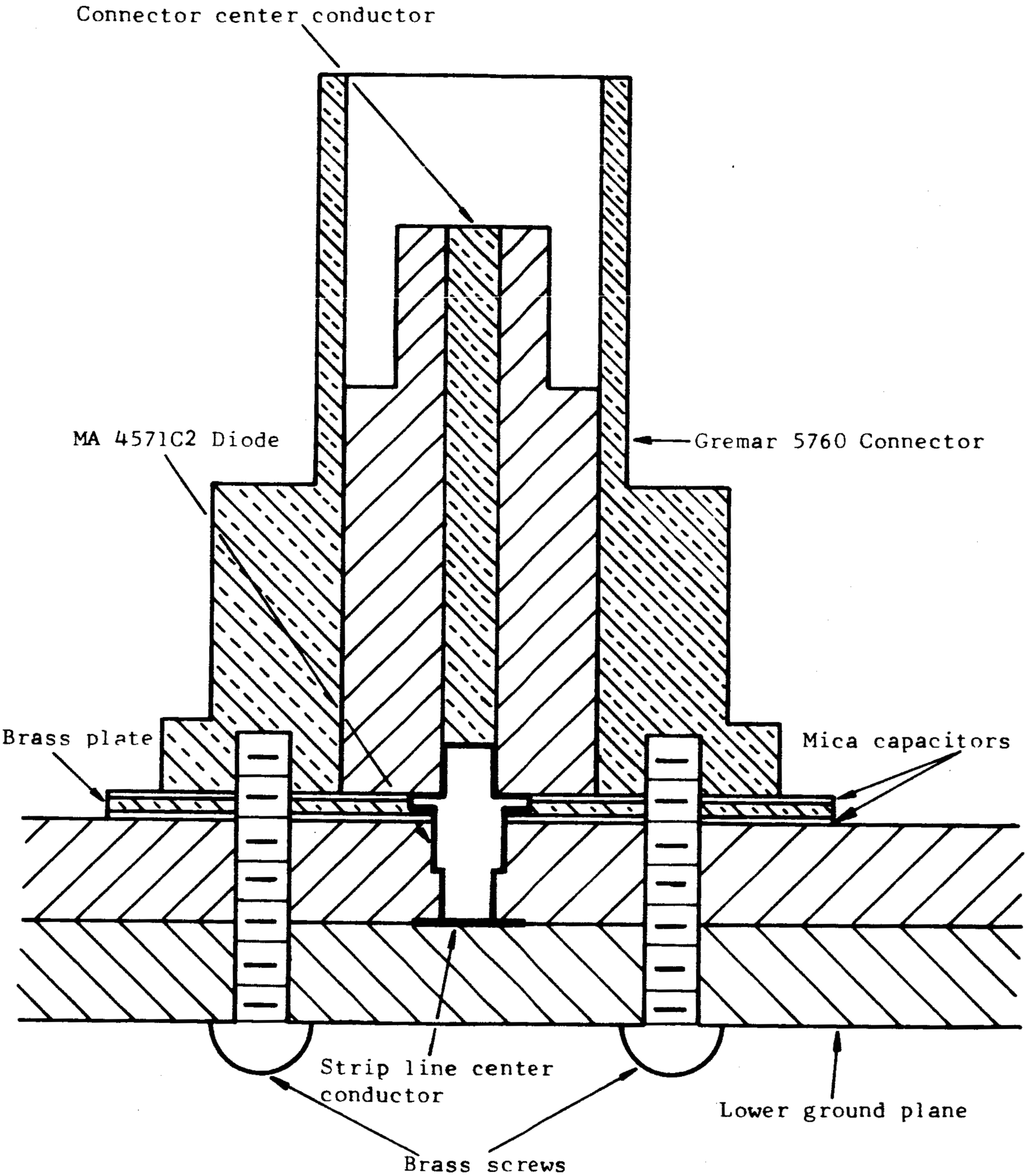


Fig. 8. A cross-sectional view of the diode mount.

and the Gremar connector at the same potential. The two screws and the diode ride in slots that allow movement perpendicular to the plane of the page. The slots are .150" in length and provide a "fine" tune of the diode and mount. The dc bias current flows in the connector center conductor through the diode to the center conductor of the strip transmission line. A quarter-wavelength shorted transmission line completes the return path for the bias current. Since the switch is to operate at essentially one frequency (narrow band), the quarter-wavelength line appears as an RF open circuit and does not impair the switching operation.

The equivalent impedance of the diode and mount in both the forward and reverse bias states was measured by incorporating the mount in a 50-ohm strip-line at a point 2.15" from the edge of the strip-line dielectric sheet.

The following measurements were made for a 50 ma forward bias current.

ITEM	VALUE
VSWR	16
λ_a	186.3 mm
Isolation	7.7 db
Minimum shift toward load	33 mm

The equivalent shunt impedance for the diode (150 ma forward bias) is obtained from this data by a conventional Smith Chart analysis plotted in Fig. 9. Point A is the impedance as measured at the reference plane of the slotted line. This impedance is transformed to the diode plane by considering the length of the coaxial adapter and the 2.15"

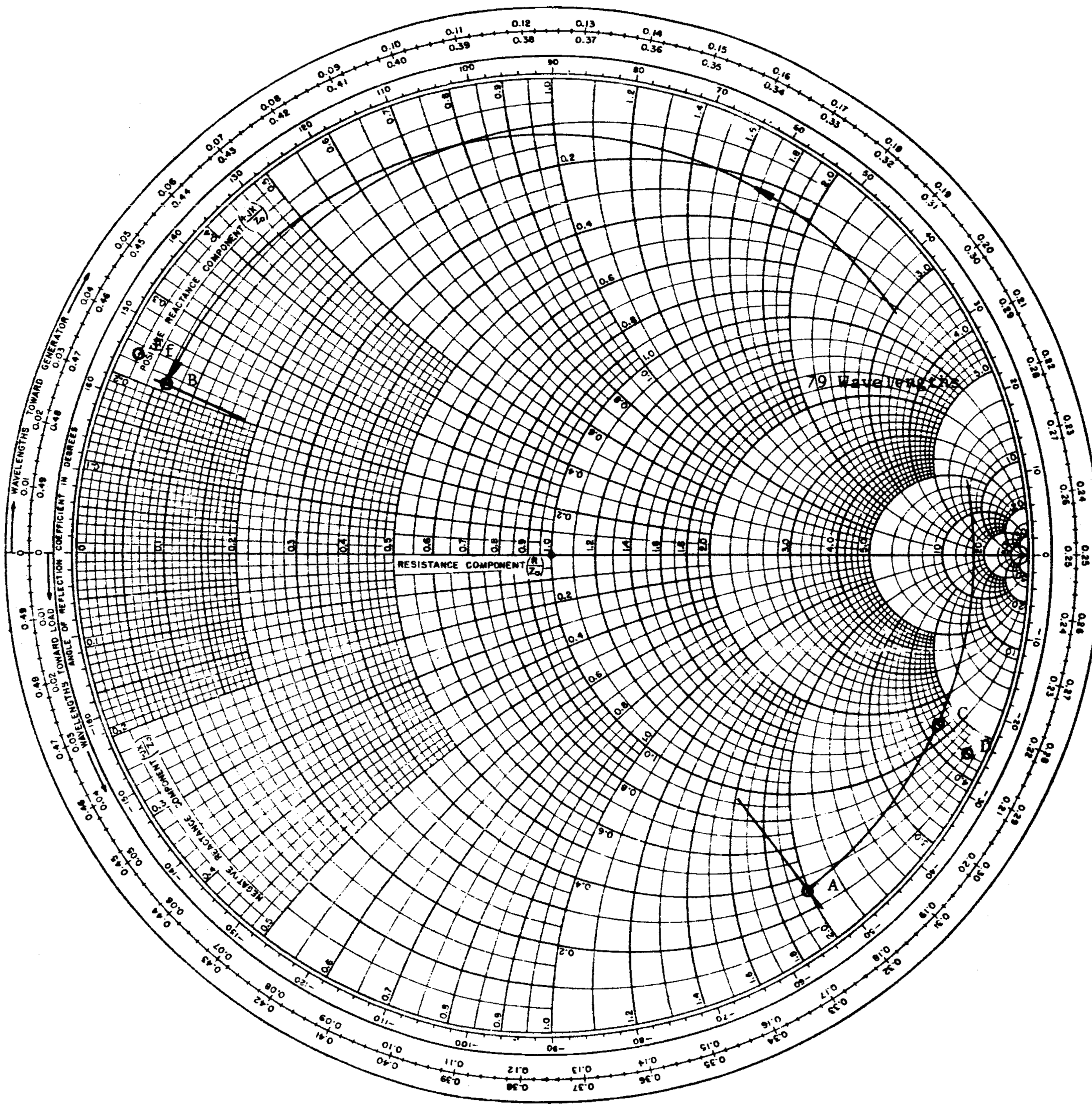


Fig. 9. Smith Chart Construction for the Forward Bias Impedance.

strip-line. This length is

$$2.15 + 1.67 = 3.82''$$

which represents

$$\frac{3.82}{4.84} = .79$$

wavelengths. Moving this distance from point A toward the load on a circle of constant VSWR (= 16) yields the point B. The line past the diode is terminated in a 50-ohm load. The impedance represented by point B is a combination of the matched load and the diode impedance. The equivalent shunt admittance at the diode plane is obtained by reflecting point B in the origin to obtain point C. The matched load admittance (1 mho) is subtracted from the admittance represented by point C to obtain the shunt admittance of the diode and mount (point D). Finally the equivalent shunt impedance presented by the diode and mount to the strip-line is obtained by reflecting point D in the origin to yield point E_f. A comparison of the isolation measured with that predicted from Fig. 3 for the point E_f (Fig. 9) shows close agreement.

The equivalent shunt impedance for the reverse bias state was obtained in a manner similar to that described above for the forward bias. The slotted line and power meter measurements for a reverse bias of 50 volts are given below.

ITEM	VALUE
VSWR	1.48
λ_a	186.3 mm
Insertion loss	.2 db
Minimum shift toward generator	34.1 mm

The Smith Chart analysis for the reverse bias state is shown in Fig. 10. Point A is plotted from the data listed above and transformed 0.79 wavelengths toward the load to obtain point B. Point B is reflected in the origin to yield the shunt admittance at point C. The load admittance of 1-mho is subtracted to obtain the shunt admittance of the diode and mount (point D). The shunt impedance is obtained at point E_r by reflecting point D in the origin. A quick check on this result is available by comparing the measured insertion loss with that given by Fig. 3 for point E_r (Fig. 10).

The information required to finalize the switch design is complete. The point E in Figs. 9 and 10 represents the shunt impedance (normalized) produced by the diode and mount in the forward and reverse bias states, respectively. To realize the maximum possible isolation that can be obtained, it is necessary to transform the impedance represented by the point E_f to the point A as shown in Fig. 11. This is accomplished by locating the diode at the end of a 50-ohm stub line connected to the main transmission line at a symmetrical tee junction. The physical length of this line must be such that 0.464 electrical wavelengths of transmission line will connect the diode to the tee junction. In the reverse bias state this length of line will transform the impedance at point E_r to that at point B. It is seen (Fig. 3) that point A will result in 28 db isolation and point B will result in .05 db insertion loss. A switch constructed under these conditions will meet the design specification.

Symmetrical Tee Junction Parameters

To determine the physical length of the diode stub line it becomes necessary to evaluate the parameters for the tee junction equivalent

IMPEDANCE OR ADMITTANCE COORDINATES

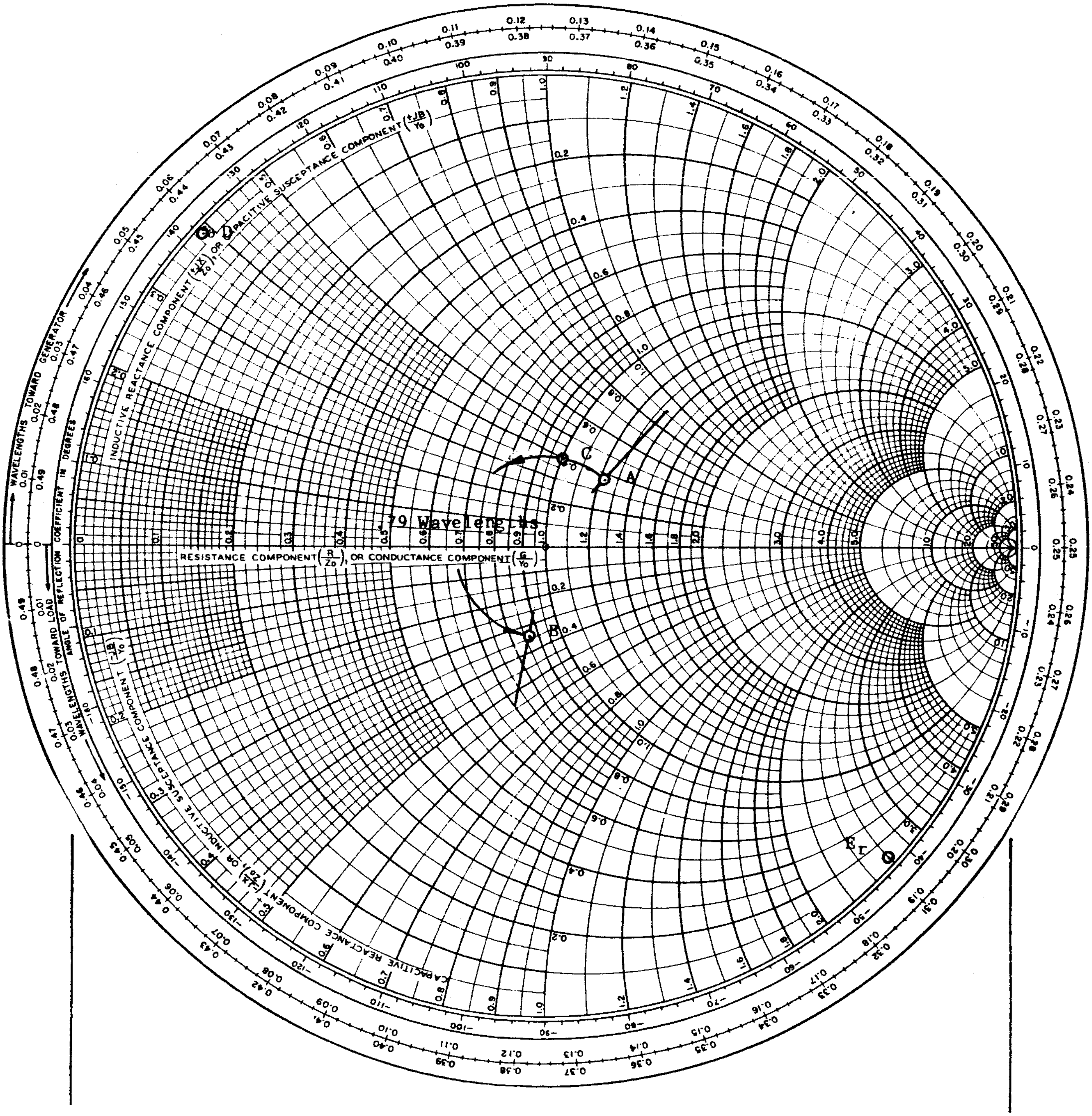


Fig. 10. Smith Chart Construction for the Reverse Bias Impedance.

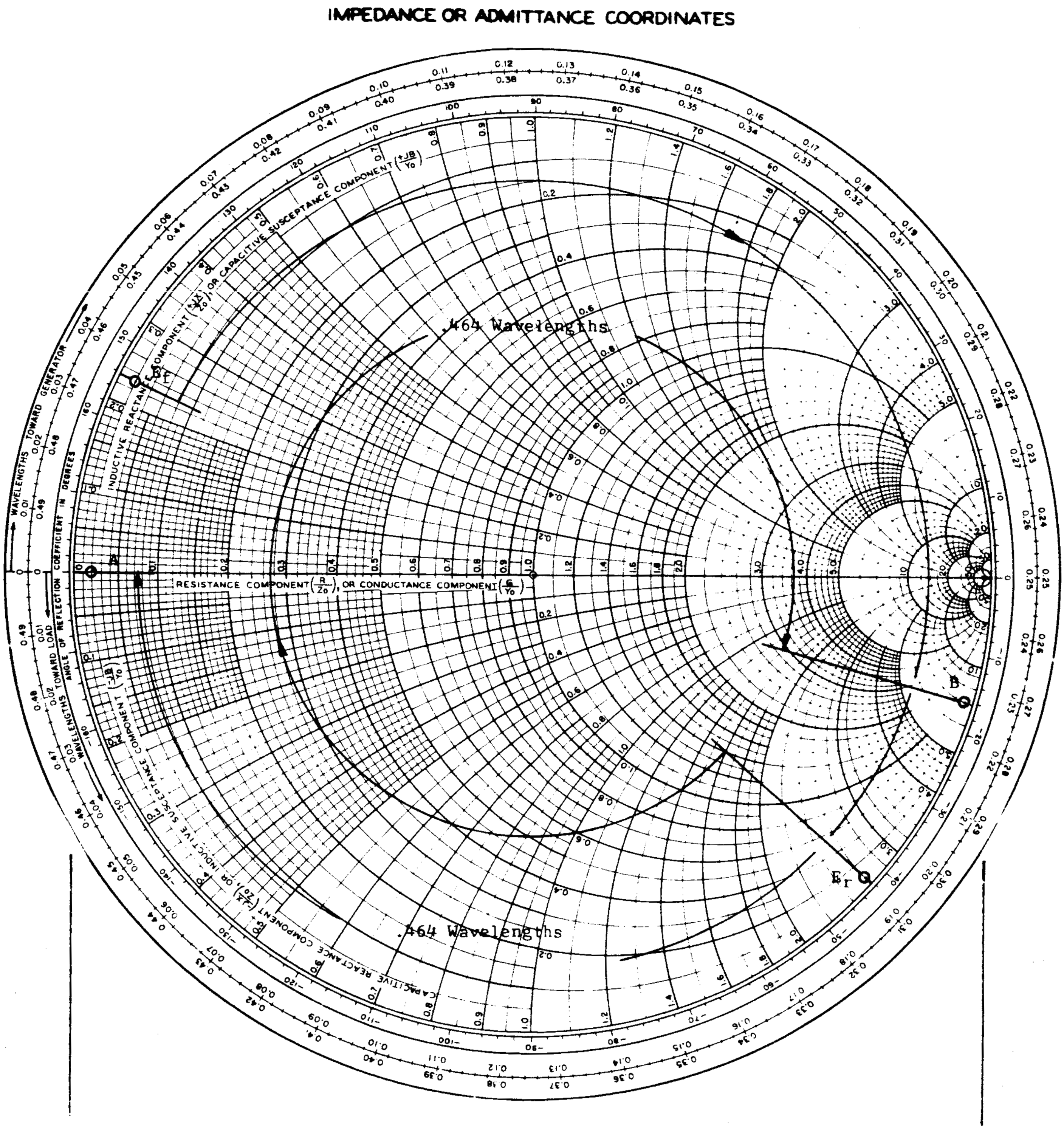


Fig. 11. Construction for Realizing the Maximum Isolation.

circuit. Of particular importance is the evaluation of the reference plane T_3 and the susceptance element B. The susceptance element is in shunt with the diode admittance transformed to the reference plane T_3 and will shift the point A (Fig. 11) away from the conductance axis. This could require a change in the 0.464 electrical wavelengths previously computed for the length of the diode stub line.

The evaluation of the tee junction parameters is a straightforward process outlined in considerable detail in the article by Franco and Oliner.[6] The stub line and the main transmission line have the same characteristic admittance and thus the same center conductor width. Listed below are the values of the parameters for the equivalent circuit of Fig. 7.

ITEM	VALUE
D'	.291"
d'	.033"
n	.996
d	.018"
B	0

The location of the reference plane T_3 is the only parameter required in determining the physical length of the diode stub line, since B = 0 and n is approximately 1. The length T_3, computed by

$$T_3 = D_1/2 - d' = .113" ,$$

is referenced to the center line of the main transmission line. The physical measurements are made with greater accuracy when referred to the edge of the center conductor. Thus, to provide an effective electrical stub line length of .464 wavelengths from T_3, requires a physical line length of

$$.464 \times 4.84 + .025 = 2.27''$$

measured from the center conductor edge.

The Physical Circuit

The physical layout (top view) of the microwave switch is shown in Fig. 12. The shorted stub line was constructed with the shorted plane at the edge of the dielectric sheet. This facilitated the assembly and dismantling of the switch; however, it produced a certain amount of ambiguity in the actual electrical location of the short. The correct length of the shorted stub line was determined experimentally. The length shown in Fig. 12 for the shorted line is 0.050" shorter than a quarter wavelength in the strip-line. The transmission line was assembled without the diode stub line and tested at 1.61 Gc/S. The input VSWR was less than 1.05 with the line terminated in a 50-ohm coaxial load. The switch was assembled as shown in Fig. 12 and tested with a low power (7 mw) microwave source. The following results were obtained.

ITEM	VALUE
Isolation	30 db at 120 ma
Insertion loss	.08 db at -150 volts
Input VSWR	1.17

As stated earlier, the PIN diode must have a minority carrier lifetime much longer than the period of the RF signal to be controlled in order that appreciable rectification of the signal will not occur. This means that fast pulse modulation of RF signals is difficult since the I layer stored charge requires some finite time to build up and decay.

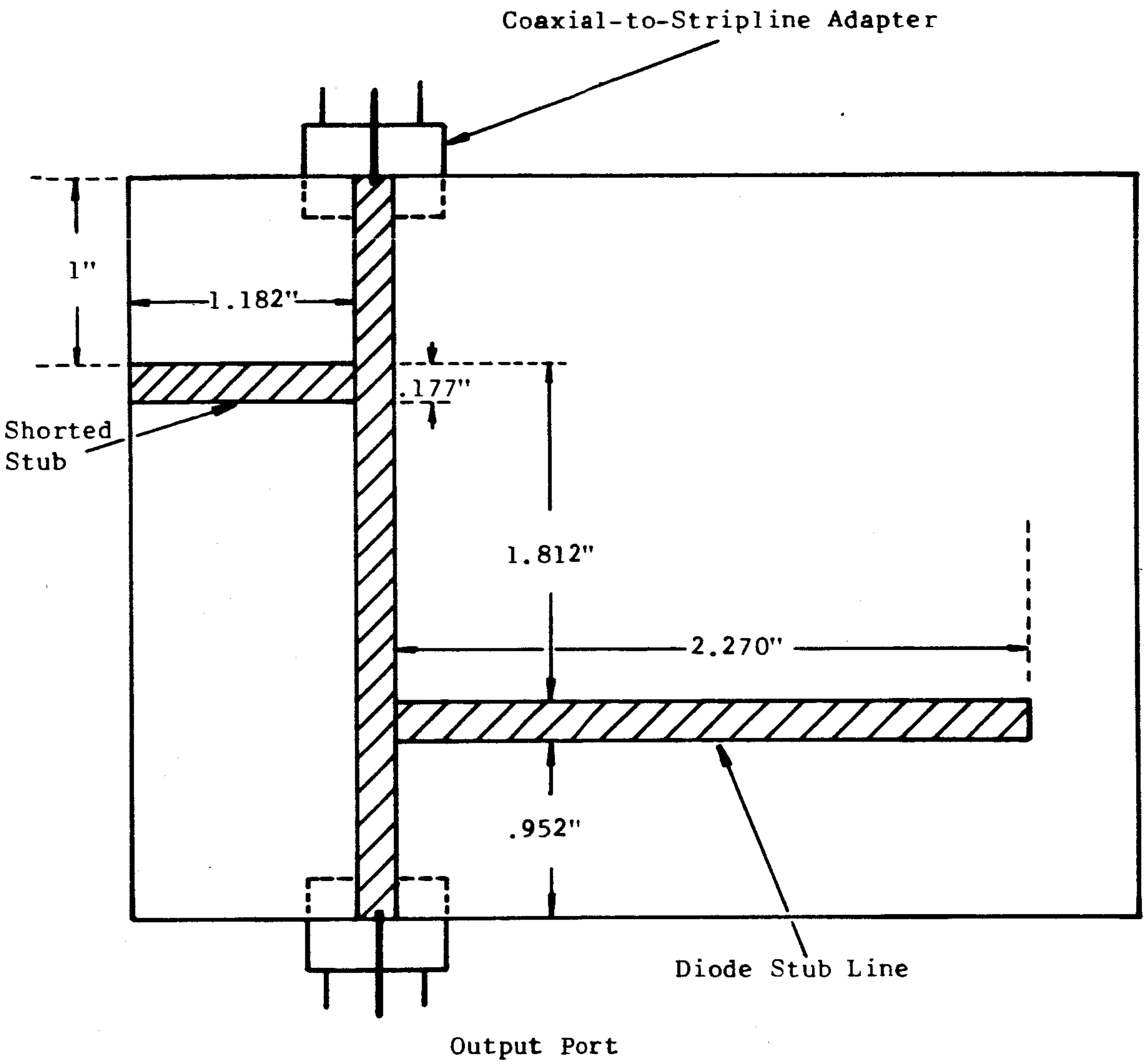


Fig. 12. Center conductor layout for the final switch design.

The time required for the charge to build up is of the order of 1 μsec., however, it is possible to reduce this time with special shaped current pulses. A test, consisting of a square wave applied to the bias circuit, was conducted to determine the time required to switch the RF energy. A low level CW signal was applied to the switch. A time of 1 μsec. was required to switch from the "RF OFF" condition to the "RF ON" condition. The time required to switch from the "RF ON" to the "RF OFF" condition was .5 μsec.

CHAPTER V

HIGH POWER PERFORMANCE

The High Power Source

A high-peak-power microwave source was designed and constructed to enable testing of the diode switch under the conditions listed in the specification. The power source consists of a cathode-pulsed planar triode in a coaxial cavity. The reentrant cavity, manufactured by Trak Microwave Corp., requires a GL 6897 planar triode as the active element and provides 7 K-watts of peak power at 1.61 Gc when properly pulsed. The input power requirements for the cavity oscillator are listed below.

ITEM	VALUE
Maximum Pulse Voltage	3500 volts
Maximum Pulse Current	5 amps
Filament Voltage	6.3 volts
Filament Current	1.05 amps

A six-section, 50-ohm, line-type pulse forming network was constructed to generate a 1.2 μsec. rectangular pulse. A variable voltage d-c supply charges the network through a charging inductor to a maximum of 2200 volts. The pulse network is discharged through a 5C22 hydrogen thyratron to produce an 1100 volt 1.2 μsec. pulse.

A transformer was designed to match the 50-ohm pulse forming network to the 700-ohm load resistance of the triode. In order to supply the triode filament current through the pulse transformer from a low-voltage-insulated filament transformer, two secondary windings that are isolated from each other with respect to direct current were provided. The two secondaries are identical, wound side by side; thus they are closely coupled as far as pulses are concerned.

The performance of the pulse forming network is displayed in Fig. 13(a), a photograph of the voltage waveform across a 50-ohm resistive load. The design pulse width of 1.2 μsec. is realized at the bottom (narrow portion) of the pulse, however, the effective rectangular pulse width appears to be nearer 1.5 μsec. Fig. 13(b) is a photograph of the voltage waveform applied to the cathode of the triode. This pulse is very similar to Fig. 13(a) which indicates that the transformer efficiently matches the 50-ohm pulse forming network to the 700-ohm triode load. Figure 13(b) shows the characteristic backswing produced by the transformer flux density returning to the remanent point.

The photograph shown in Fig. 13(c) is the spectrum obtained by pulsing the triode with the voltage waveform of Fig. 13(b). The photograph was taken on a Polarad Model TSA Spectrum Analyzer at a sweep rate of 566 kc/cm about a center frequency of 1.61 Gc. Note that the minima of the spectrum do not go to zero, indicating frequency modulation on the pulse; also the asymmetry of the spectrum, indicative of amplitude modulation, may be noted. The frequency separation of 1.35 mc between the minimum points indicates that the effective RF pulse width is 1.48 μsec. On the basis of this pulse width measurement the pulse repetition rate was adjusted to 670 pps to provide a 10^{-3} duty cycle.

The High Power Measurements

The test equipment required to measure the isolation provided by the switch is shown on Fig. 14. The measurement procedure was performed

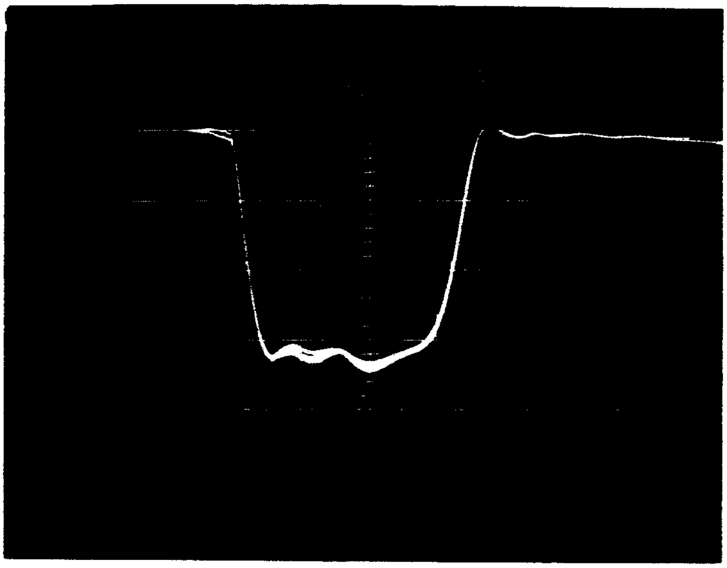


Fig. 13-a. Voltage waveform across a 50-ohm resistive load. Sweep speed 0.5 μsec./cm. Amplitude 1100 volts.

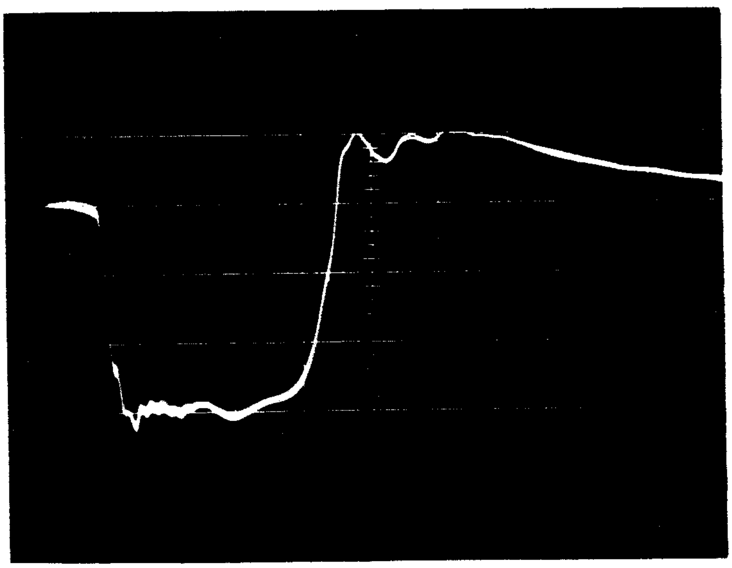


Fig. 13-b. Voltage waveform across the cathode of the triode. Sweep speed 0.5 μsec./cm. Amplitude 3000 volts.

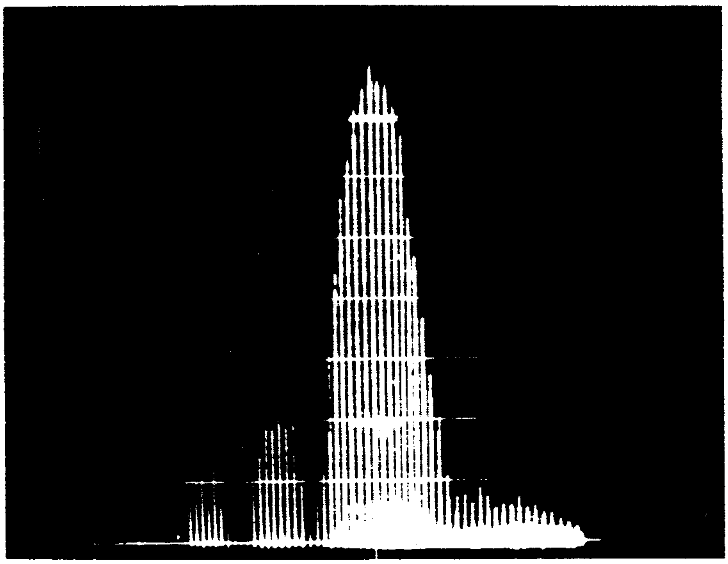


Fig. 13-c. Spectrum of the RF energy produced by the cavity oscillator.

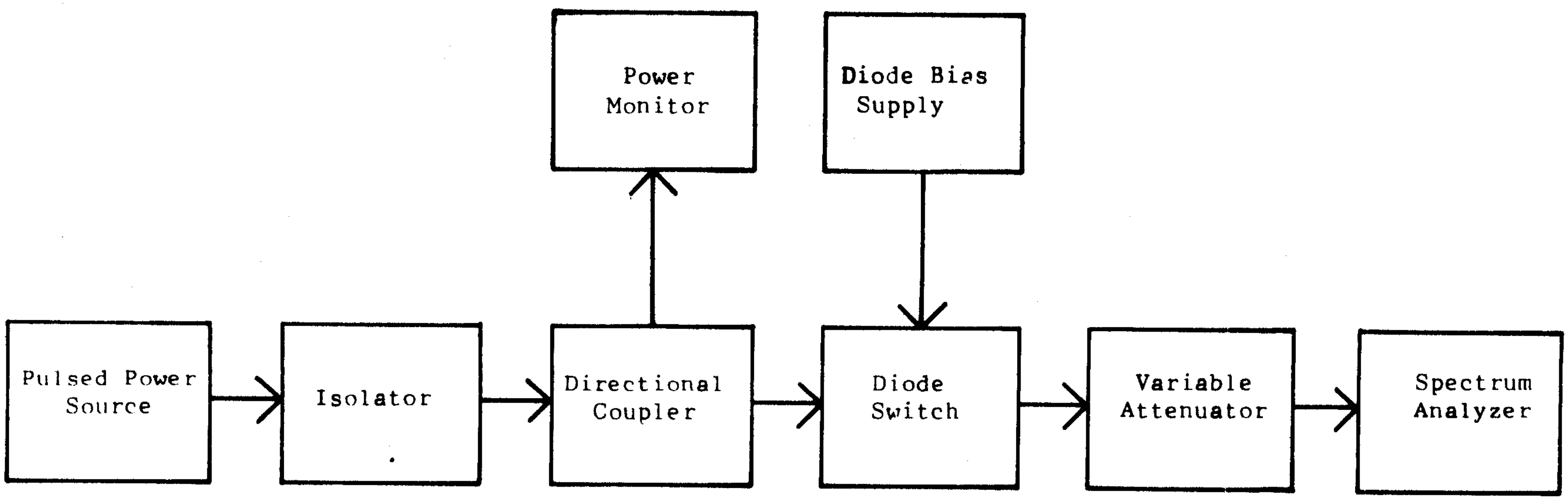


Fig. 14. Test setup for measuring switch isolation under high power operation.

in the following manner; for each value of incident power the switch was removed from the circuit and a reference level established on the spectrum analyzer, next the switch was inserted and the forward bias adjusted to the desired value, then the variable attenuator was adjusted to obtain the same reference on the spectrum analyzer. The difference in attenuation noted on the variable attenuator was recorded as the isolation.

The insertion loss, produced by the switch in the reverse bias state, was measured using the equipment indicated in Fig. 15. The variable attenuator was set to provide the desired incident power, as determined by the power monitor. The diode switch was inserted in the circuit and the reverse bias set on the bias supply. The difference between the readings of the power monitor was recorded as the insertion loss. The power monitor, consisting of a Hewlett-Packard 431C Power Meter and assorted attenuator pads, is capable of indicating power changes of .1 db or less when operating on the db scale between 0 db and -1 db. All measurements were made with the power level adjusted so that the power meter would indicate in this range. The directional coupler and spectrum analyzer were used to detect any change in the RF pulse that might indicate RF breakdown.

Results of the Measurements

The results of the isolation measurements are shown in graphical form in Fig. 16. The predicted effects of decreased RF resistance under increasing dc bias is apparent (isolation increases as dc bias increases). The measurements show an increase in the RF resistance as the pulse power is increased. The 2.5 db decrease in isolation, noted as the peak

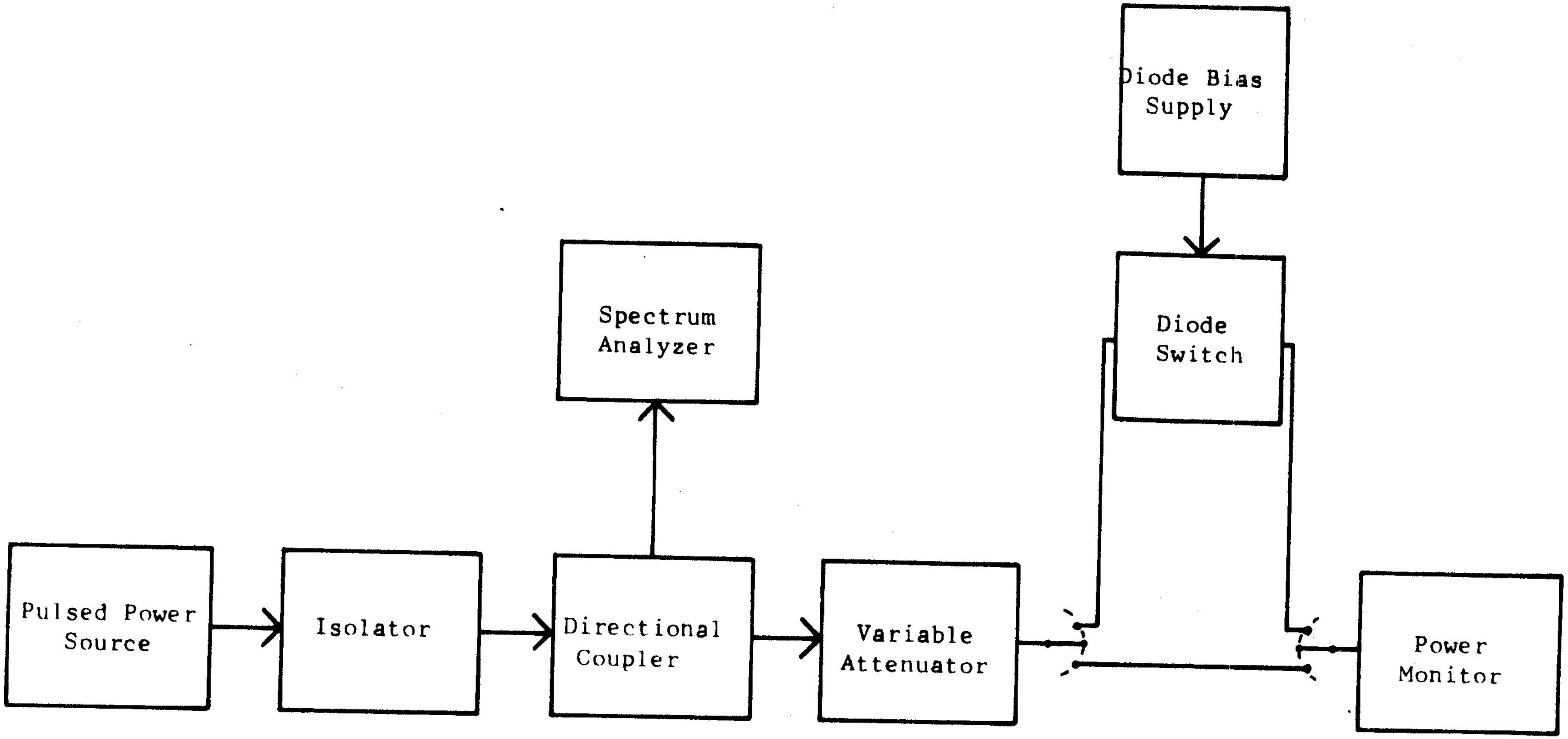


Fig. 15. Test setup for measuring switch insertion-loss under high power operation.

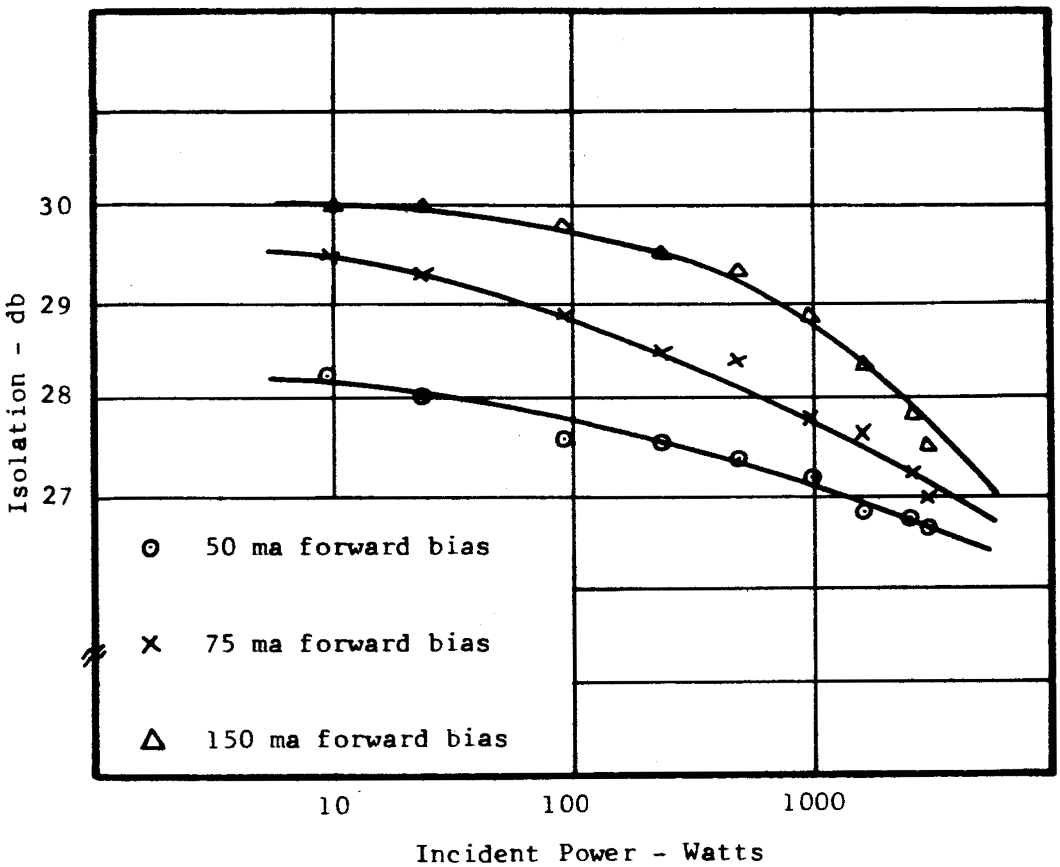


Fig. 16. Isolation vs incident power for three values of forward current.

power was increased from 10 watts to 3 K-watts for a 150 ma bias current, represents an increase (from .82 to 1.1 ohms) in the resistance shunting the transmission line. The switch design specification for isolation (28 db) was obtained with 150 ma bias current at 1 K-watt of incident peak power.

Figure 17 is a graph showing the variation in insertion loss as the peak power was increased. The results were obtained for three values of reverse bias voltage. The results indicate that large RF voltage excursion into the diode forward bias state is possible with little increase in the insertion loss. The insertion loss for a 50 volt bias was 0.3 db when operating under 3.5 K-watts of incident power. This corresponds to an RF voltage of 595 volts zero to peak, or an RF voltage excursion of 545 volts into the forward bias region. It was necessary to reverse-bias the diode at 375 volts to meet the design specification for insertion loss at 1 K-watt of incident power.

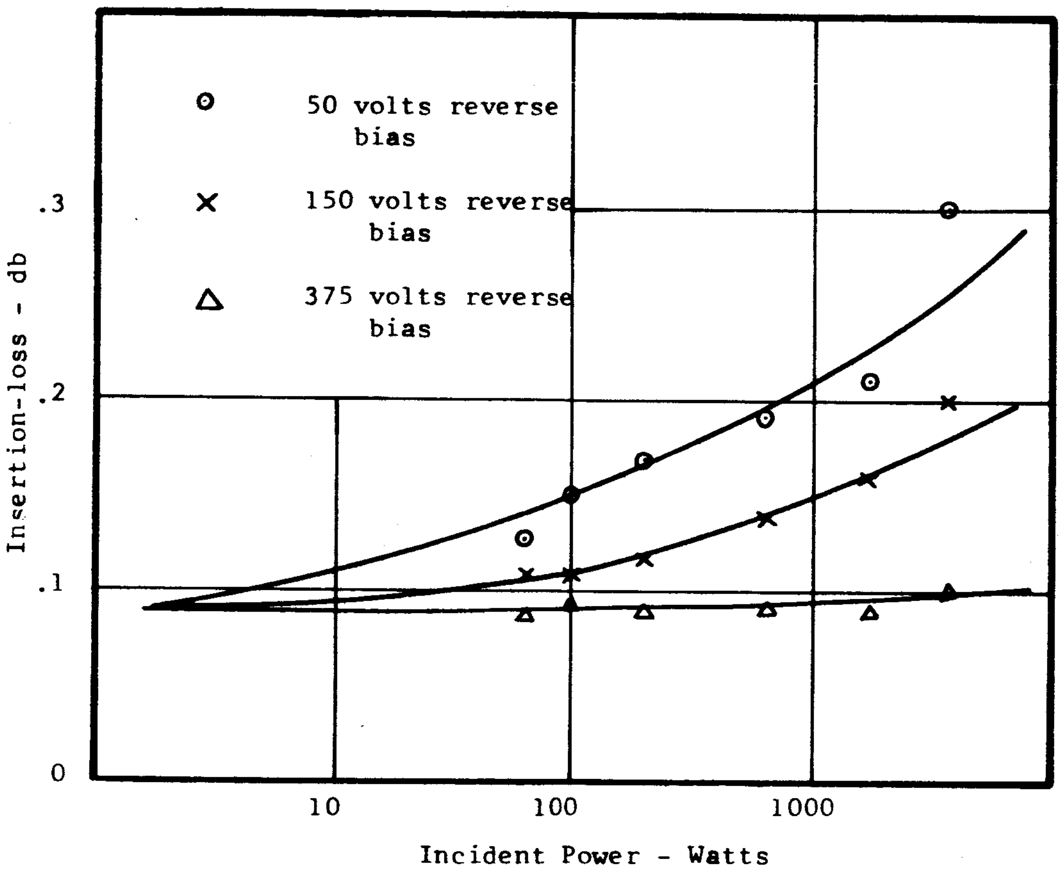


Fig. 17. Insertion-loss vs incident power for three values of reverse voltage.

CHAPTER VI

HARMONIC MEASUREMENTS

General Considerations

A primary factor influencing the design of the switch was the requirement that harmonic power measurements made external to the switch be directly related to the harmonics generated by the diode. The results of the measurements could then be plotted as harmonic power generated by the diode vs incident fundamental power at the diode.

The physical lengths of the switching circuit (fig. 12) were calculated to "match" the diode at the second harmonic frequency to the output port of the switch. The parameters for the symmetrical tee junction were calculated at the second harmonic wavelength and were found to be approximately the same as determined for the fundamental. The reference plane T_3 (fig. 7) has the same location for the second harmonic as for the fundamental and thus the electrical length of the shorted stub is unchanged. The matching was accomplished by recognizing that the shorted stub reflects a short to the junction of the main transmission line at the second harmonic wavelength. The length of line separating the diode stub line from the shorted stub line (1.812") is .75 of the second harmonic wavelength so that the short circuit is transformed into an open circuit at the diode stub line junction. Thus the second harmonic energy generated by the diode propagates down the stub line to the tee junction. Since one side of the junction is electrically open, the energy will be directed without reflection to the output port. The second harmonic power measured at the output port is equal in magnitude to the second harmonic power generated by the diode.

The symmetrical tee junction parameters calculated at the third harmonic were in close agreement with the fundamental frequency parameters. The shorted stub line presents an open circuit to the main transmission line at the third harmonic. If the main transmission line is matched the third harmonic energy generated by the diode will divide equally at the tee junction. Under these conditions a 2:1 SWR (at third harmonic) will exist on the diode stub line. If it is assumed that the diode impedance at the third harmonic is 50 ohms, the third harmonic power measured at the output effort of the switch will be 3.5db below the power generated by the diode.

It is to be expected that the harmonics generated by the diode in the forward bias state will be a function of the d.c. bias current and the incident RF (fundamental frequency) power at the diode. The d.c. bias current must maintain a sufficient charge in the intrinsic region of the diode to allow RF current flow. If the negative half cycle of the RF current depletes the stored charge in the intrinsic region the diode will be driven into reverse bias, creating a nonlinear action with resulting high harmonics. In the forward bias state the diode nearly short circuits the line so that short circuit RF current flows through the diode. From Fig. 2 it is seen that if $Y = \infty$ (short circuit) the current through Y must be 2I. Thus

$$I_{sc} = 2I \qquad (3\text{-}1)$$

where

$$I = \frac{2P}{Z_o} \qquad (3\text{-}2)$$

and P is the incident RF power. Since the RF current is sinusoidal, the instantaneous short circuit current can be expressed as

$$i = I_{sc} \sin \omega t$$

so that the instantaneous charge q is given by

$$q = I_{sc} \int_{o}^{t} \sin(\omega\tau)\, d\tau$$

$$= \frac{-I_{cs}}{\omega} \cos(\omega t)$$

The minimum charge required in the intrinsic region to prevent reverse bias is

$$Q = \frac{I_{cs}}{\omega} \qquad (3\text{-}3)$$

The charge furnished by the d.c. bias current is obtained from

$$Q = I_{dc}\tau \qquad (3\text{-}4)$$

where τ is the minority carrier lifetime. Equating equations 3-3 and 3-4 and using equations (3-1) and (3-2) with Z_o = 50-ohms yields

$$P_{dc} = 6.25 \left[\omega\tau\ I_{dc}\right]^2 \qquad (3\text{-}5)$$

where P_{dc} is the value of incident power required to produce charge depletion for a given d-c bias current I_{dc}. PIN diodes developed for relatively high power (1-10kw) switching applications will have a minority carrier lifetime τ in the range .2-2 μsecs. For a fundamental frequency of 1.61 Gc. equation 3-5 will reduce to

$$P_{dc} = 2550\ I_{dc}^2 \qquad \tau = 2x10^{-6} \text{ sec} \quad (3\text{-}6)$$

$$P_{dc} = 25.5\ I_{dc}^2 \qquad \tau = 2x10^{-7} \text{ sec} \quad (3\text{-}7)$$

where I_{dc} is expressed in milliamperes and P_{dc} in watts. Thus 10 ma bias current will be required if the incident power is 2.5 kw and τ = .2 μsecs. Equation 3-5 should be interpreted as a lower limit on the d-c bias current for a given incident power; however, it is to be expected that some nonlinear action will exist above this limit.

The diode switch operating in the reverse bias state will generate harmonics depending on the reverse bias voltage and the incident power. Equation 2-11 defines the incident RF power required to produce a peak RF voltage equal to one half the breakdown voltage. If $V_b/2$ in equation 2-11 is replaced by V_B the resulting equation

$$P_i = \frac{V_B^2}{2Z_o} \qquad V_B < \frac{V_b}{2} \qquad (3\text{-}8)$$

yields the incident power required to produce a peak RF voltage capable of driving the diode to zero bias for a given d-c reverse bias V_B. Incident power greater than P_i will cause some charge to be injected into the diode depletion region; however, the charge will be withdrawn during the negative half cycle. The operating frequency (1.61Gc) has a period ($\approx$.7 nsec) much less than the minority carrier lifetime so that the probability of carrier recombination is very small. This probability will increase with increasing incident power. The recombination process provides the nonlinear action responsible for harmonic generation in the reverse bias state. Thus for a given incident power, minimum harmonics will be generated when the d-c reverse bias is at a maximum. A practical maximum reverse bias exists since diode failure will occur when the sum of the d-c reverse bias voltage and the zero to peak RF voltage exceeds a voltage proportional to the diode breakdown voltage. The maximum reverse bias was arbitrarily set at $V_b/2$ for the experimental work of this report, and no diode failure occurred for incident powers up to 66.4 dbm.

Harmonic Measurement Technique

A block diagram of the equipment setup used to obtain the harmonic measurements is included as Fig. 18. The pulsed power source, previously described in Chapter II, generated second harmonic power 20db below the fundamental power. The third harmonic content of the pulsed power source

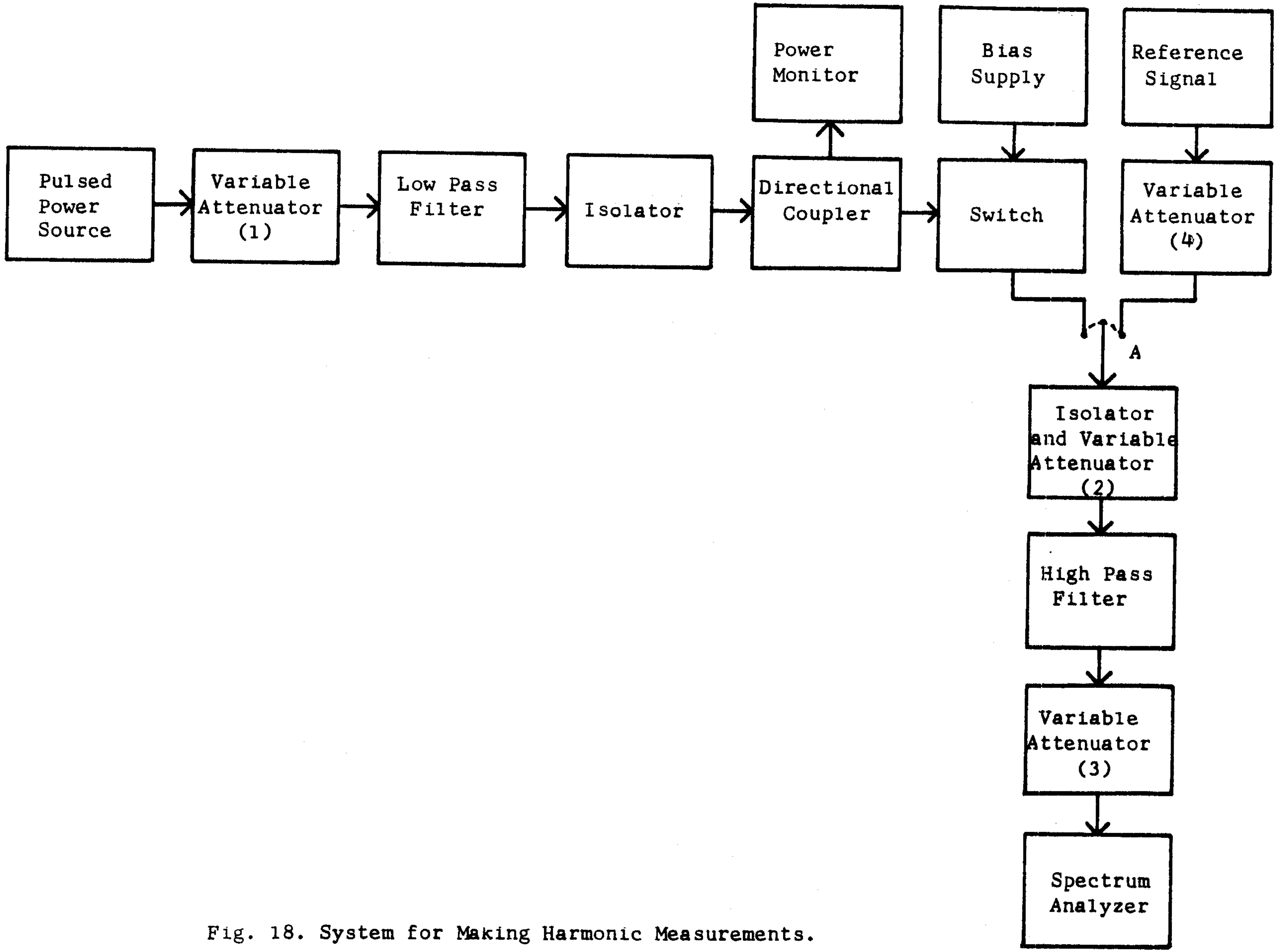


Fig. 18. System for Making Harmonic Measurements.

was approximately 30db below the fundamental power. The presence of harmonics in the RF source required the addition of a low pass filter to the RF circuit. The low pass filter, a Hewlett-Packard 360C, provided an isolation of 75db at the second and third harmonic. The test signal at the output port of the low pass filter contained harmonics at least 95db below the fundamental.

The switch operating in the forward bias state reflects the incident power. A PRD type 1210 ferrite isolator was used to absorb the reflected power, thus effectively matching the line as seen by the switch looking toward the generator. A directional coupler and power meter sampled the fundamental power incident on the diode switch.

The measuring of the harmonic power at the output port of the switch was accomplished by comparing it with a known reference signal. A high pass filter (waveguide below cutoff) prevented the fundamental power from reaching the spectrum analyzer which was used as the comparing instrument. The fundamental power reflected by the high pass filter was absorbed in a second PRD type 1210 isolator, thus preserving the matched line as seen by the switch looking toward the spectrum analyzer at the fundamental frequency. The five feet of RG-48U waveguide used as the high pass filter provided 400db of attenuation at the fundamental frequency. Variable attenuator (3) consisted of a Hewlett-Packard Model 5375A variable attenuator connected to the RG-48U waveguide. For third harmonic measurements the RG-48U was replaced with RG-50U waveguide to provide second harmonic as well as fundamental power attenuation. The Hewlett-Packard Model 5375A was replaced with a Model J382A variable attenuator.

The second and third harmonic reference signals used in the measuring process were CW signals obtained from a reflex klystron. A Sperry 2K41 klystron provided the second harmonic reference signal while the third harmonic reference signal was generated by a Varian 244G klystron.

The procedure followed to determine the harmonic power generated by the diode using the test setup of Fig. 18 is given below:

A specific fundamental power incident on the diode was obtained by adjusting attenuator 1 until the desired power level was indicated on the power meter. The diode switch was set at a maximum bias level (maximum reverse voltage or maximum forward current) in one of the two bias states. The output of the switch was connected at point A to the spectrum analyzer and attenuator (2) adjusted to give a suitable deflection on the spectrum analyzer CRT with attenuator (3) set at zero. Next, the line at A was disconnected from the switch and connected to the reference signal. Attenuator (4) was adjusted to obtain the same deflection on the spectrum analyzer CRT as was produced by the diode harmonic signal. The reference signal was disconnected at A and connected to a power meter. The power meter reading was recorded as P_A. The output port of the switch was reconnected to the spectrum analyzer at point A and the bias level decreased in steps. At each step the attenuation of attenuator (3) was increased to maintain a constant deflection on the analyzer CRT. The change in attenuation from step to step was recorded as the increase in generated harmonics accompanying the decrease in bias. This same procedure was followed for the diode biased in the other state, and for each new incident fundamental power.

The measurement process as previously described is based on the comparison of a CW reference signal with the diode generated harmonic signal (pulsed) using a spectrum analyzer as the comparing instrument. The sensitivity or gain of a spectrum analyzer is a function of the pulse width of the displayed signal, with the maximum sensitivity achieved when displaying a CW signal. The loss of sensitivity, due to a pulsed signal, was determined at the fundamental frequency by displaying a known pulsed signal (1 KW peak, 10^{-3} duty cycle) and comparing it with a CW signal required to produce the same display deflection. For the spectrum analyzer referred to in this report it was found that the loss in sensitivity for a pulsed signal of 1.48 μsec pulse width was 31.6db. The spectra of the second and third harmonics generated by the diode were identical with the fundamental spectrum (Fig. 13c) indicating that the harmonics had the same time domain pulse shape as the fundamental, and hence the same loss in sensitivity was experienced for the harmonics. The magnitude of the generated harmonics was obtained by adding 31.6db to P_A plus the attenuation of attenuator 4.

The theoretical study and its predictions of the behavior of the diode switch required that the transmission system be matched at the fundamental frequency and its harmonics. The ferrite isolators used to match the system produced a VSWR of 1.2 when connected to an open or short circuit. Whenever possible in the measuring procedure attenuator pads were used in conjunction with the isolators to provide additional reduction in the VSWR.

Results

Figure 19 depicts in graphical form the magnitude of the second harmonic power generated by the diode for different values of forward bias current and incident fundamental power. The curves of Fig. 19 have the form predicted by the discussion given at the beginning of the chapter, i.e., the harmonic power increased as the incident fundamental power increased and as the bias current was reduced. Although the negative slope of these curves increased as the bias current was reduced, a marked increase, indicating charge depletion of the "I" region, was not present. An evaluation of equation 3-6 for the maximum incident power (3470 watts peak) reveals that a bias current of 1.17ma is sufficient to prevent charge depletion. Equation 3-7 yields a bias current of 11.7ma for the same incident power. It should be pointed out that harmonic power measure-

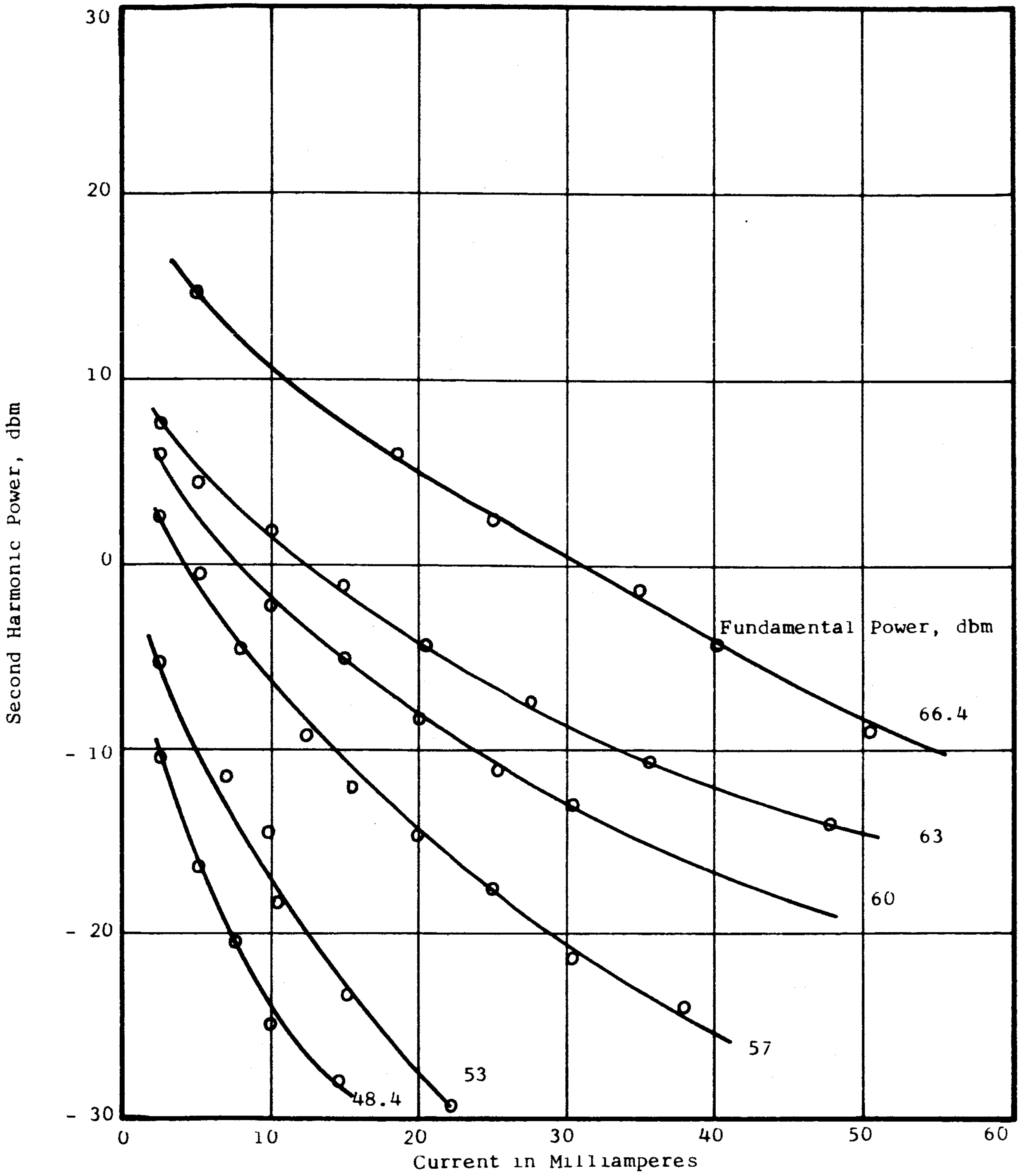


Fig. 19. Generated Second-Harmonic Power for Forward Bias.

ments were not made for bias currents greater than 50ma since the diode generated second harmonic was approaching the harmonic content of the test signal (within 10db).

The results of the second harmonic measurements for the diode biased in the reverse bias state are displayed in Fig. 20. The functional dependence of the generated harmonic on reverse bias and incident power is evident. Two knees appear in the middle four curves. The lower knee (higher reverse bias) occurs at a reverse bias voltage near V_B as predicted by equation 3-8. The curves for the higher incident power have a smaller slope at large reverse voltages indicating that some nonlinear effect is encountered on the negative half cycle of the RF voltage, probably in the form of increased reverse current through the diode. The second harmonic generated at the highest incident power remained fairly constant with reverse bias, after the initial decrease, indicating distortion of the fundamental signal on both the positive and negative half cycles. It is apparent that the incident power has reached a maximum and that any further increase would probably result in failure of the diode. In any respect, increasing the reverse bias voltage is no longer effective in reducing the generated harmonic.

Interference control requirements for airborne electrical and electronic equipment are specified in MIL-I-6181D. Article 4.3.3.2.1 of Military Specification MIL-I-6181D states that the peak power output of the second and third harmonics of the output fundamental frequency shall be at least 60db below that of the fundamental, and in no event greater than 1 watt. The peak power output of any harmonic above the third, must be at least 80db below that of the fundamental, and in no event greater than 10^{-2} watts.

The measurements performed at the third harmonic frequency show that the generated third harmonic is at least 7db below the second harmonic for

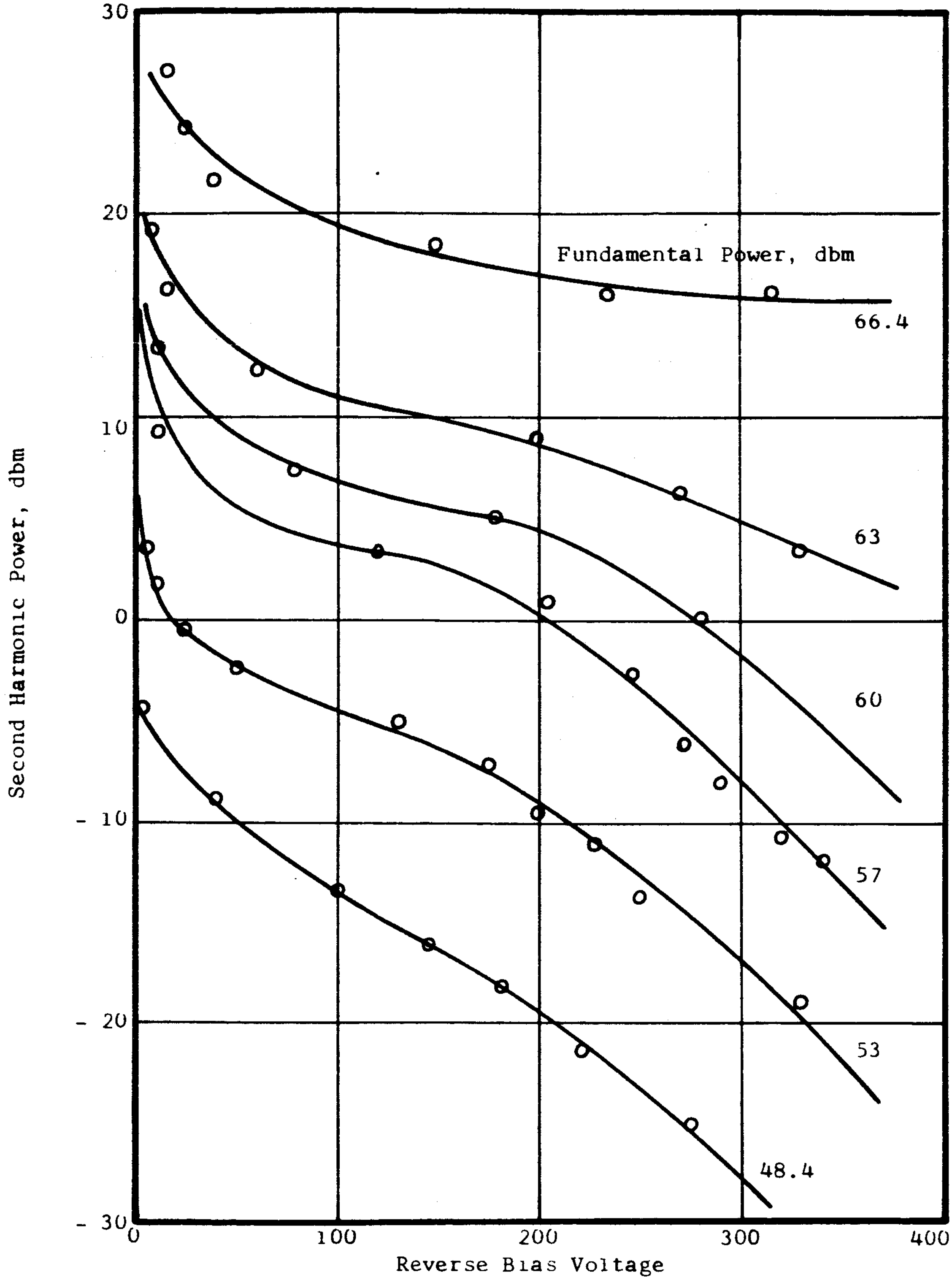


Fig. 20. Generated Second-Harmonic Power for Reverse Bias.

normal bias operation. The results of the third harmonic measurements with the diode forward biased are shown in Fig. 21 with the corresponding second harmonic results. Although only the results for two incident fundamental powers are shown, data taken at other incident power levels produced curves similar to thos shown in Fig. 21. It should be stated that the third harmonic data displayed in Fig. 21 and Fig. 22 was measured at point A (Fig. 18) and is at least 3db below that generated by the diode. Figure 22 depicts in graphical form the dependence of the generated third harmonic power on reverse bias voltage and fundamental power.

The third harmonic content of the output signal exceeds the second harmonic for low bias levels in both the forward and reverse bias states. However, for normal switching operation the forward bias current should exceed 50ma and the reverse bias voltage should exceed 200 volts. A preliminary search of the higher harmonic frequencies failed to disclose any signals within the sensitivity of the analyzer. This would place the higher order harmonics greater than 80db below the fundamental under normal bias conditions.

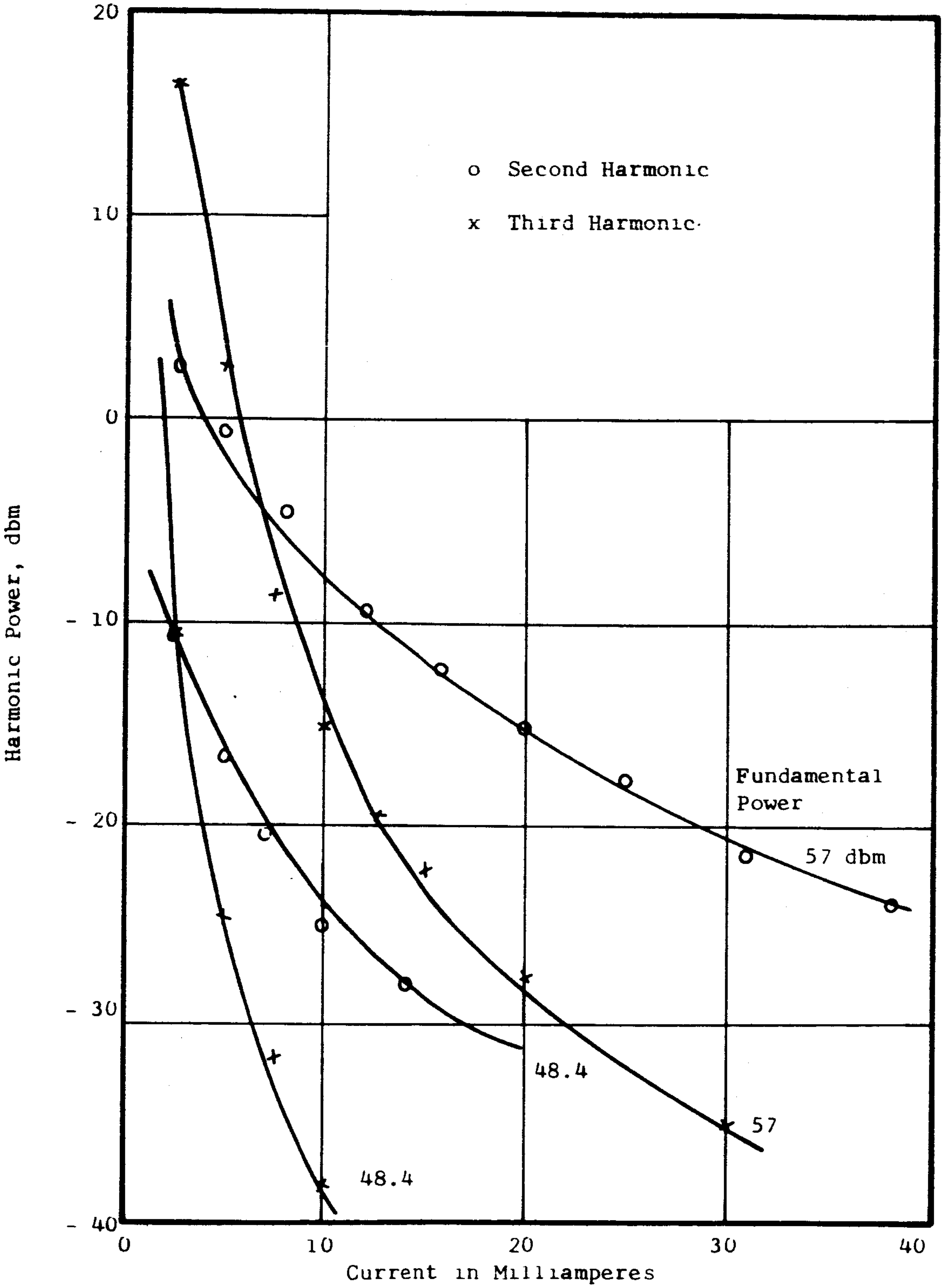


Fig. 21. Second and Third Harmonic Power for Forward Bias.

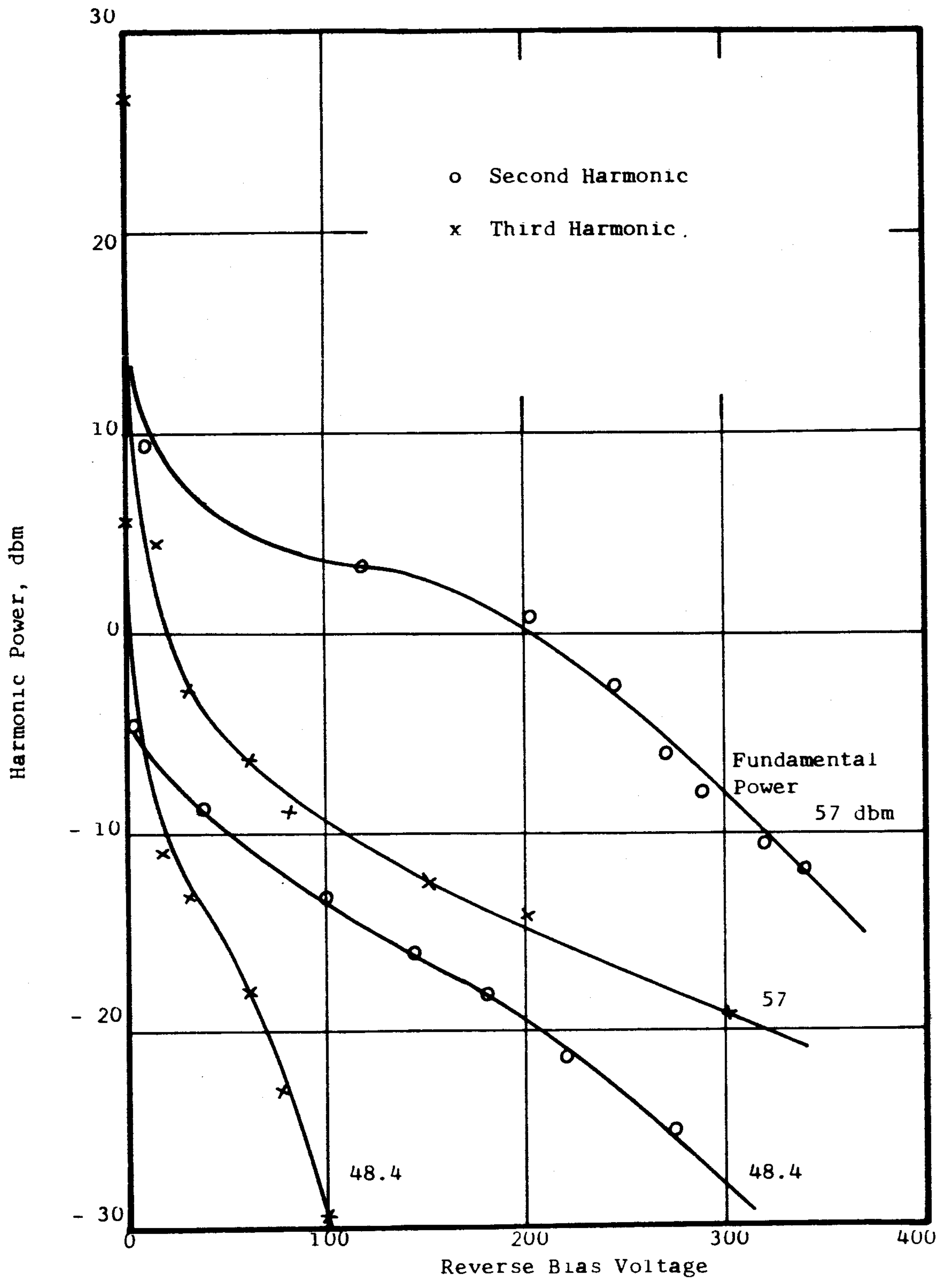


Fig. 22. Second and Third Harmonic Power for Reverse Bias.

CHAPTER VII

CONCLUSIONS

The area of microwave semiconductor electronics as applied to the switching of microwave energy at relatively high levels has been investigated in this report. The use of a PIN diode as the semiconductor element coupled with the theory of TEM transmission line switching has allowed the construction of a microwave switch capable of handling in excess of 3 K-watts of peak power.

The theory of operation was developed in a general manner requiring only that a switching element be available that could exhibit two different impedance states at microwave frequencies. The curves of constant attenuation plotted on a Smith Chart (Fig. 3) were used often in the design stage and proved invaluable in arriving at the final switch design. The final switch design, utilizing the diode located on a stub line, is but one of a number of possible designs to achieve a variable impedance shunting a transmission line. In this configuration the actual switching operation, resulting from a change in the diode bias, is produced by the change in equivalent impedance at the tee junction.

Since the primary concern of this study was harmonic generation by the diode, the choice of switching circuit was based on simplicity of the circuit analysis at the harmonic frequencies. The results of the harmonic measurements are displayed independently of the microwave switching circuit and represent the diode behavior for various fundamental frequency power levels and bias levels.

The power handling capabilities of the diode were discussed and some equations derived to predict the maximum power to be handled by a given diode. The maximum power derived, for the diode in the reverse bias state, was in terms of the incident power on the diode. The final switch design located the diode on a stub line, however, the length of the stub line (.464 wavelengths) and the reverse bias reflection coefficient of the diode were such that the RF voltage at the diode was the same as the voltage at the tee junction. Thus, Eq. 2-11 is still valid and yields the peak power required to produce a peak-to-peak RF voltage equal to the reverse breakdown voltage of the diode. This peak power limit is not a limit in the sense that the diode will be destroyed if the limit is exceeded but rather a point at which peak power in excess of this limit will result in increased insertion loss, and generated harmonics.

The experimental data obtained from the harmonic measurements support the statement that PIN diodes operating in typical switching applications produce little harmonic power. The data displayed in Fig. 19 shows that the requirements of MIL-I-6181D would be met for a forward bias current of 40ma or greater. Results of tests on switch isolation vs. forward bias current (Fig. 16) indicate that the bias current should be 100ma or greater. The harmonics generated in the reverse bias state exceed the specifications of MIL-I-6181D for incident powers greater than 60dbm. At this power level a reverse bias voltage near $V_b/2$ was required to meet the specification.

An analysis was performed to determine the optimum characteristic impedance for the transmission line. This optimum value permits the diode to operate at its maximum peak and average power level in a pulsed environment. However, it was pointed out that the resulting impedance would determine the maximum insertion loss for a given diode. The diode actually used in the switch should be capable of switching 95 watts of average power. The maximum average power in the test was 3 watts. The diode stub line could have been increased in width to produce a lower characteristic impedance. The maximum peak power ($\overline{P}_{ip}$) would have increased (Eq. 2-11), and the increase in insertion loss noted in Fig. 17 would have occurred at a high peak power, however, the decrease in characteristic impedance would have decreased the maximum isolation (Eq. 2-6) and prevented the attaining of the required 28db isolation.

The effect of reducing the characteristic impedance of the diode stub line is to reduce the peak-to-peak RF voltage excursion for a given incident power. Under these conditions the harmonics generated by the diode in the reverse bias state will be reduced. However, the increased short circuit RF current would require an increase in the forward bias current to maintain the same harmonic generation in the forward bias state.

In conclusion, this report has shown that it is feasible to use a PIN diode as the switching element in a high power microwave switch. The switch was tested at power levels to 3-K watts and the changes in switch performance noted. It was pointed out that the switching circuit developed for this study could have been altered (lower diode stub line characteristic impedance) to reduce the harmonics generated by the diode in the reverse bias state, but only at the expense of reduced isolation. PIN diodes are available with reverse breakdown voltages in excess of 1000 volts. The use of these diodes should enable the design of a switch capable of providing adequate isolation and at the same time meeting the requirements of MIL-I-6181D while operating at peak powers of 7-K watts or greater.

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