Microsemi-Watertown

THE PIN DIODE CIRCUIT DESIGNERS' HANDBOOK

The PIN Diode Circuit Designers' Handbook was written for the Microwave and RF Design Engineer. Microsemi Corp. has radically changed the presentation of this PIN diode applications engineering material to increase its usefulness to Microwave and RF Circuit Designers. A major part of this Handbook is devoted to the basic circuit applications of this unique device.

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Preface

This PIN Diode Circuit Designers' Handbook was written for the Microwave and RF Design Engineer. A major part of this Handbook is devoted to the basic circuit applications of this unique device. In each chapter, a circuit function is treated in detail followed by specific selected applications. For example, in Chapter 2, the common PIN diode switch configurations are presented, followed by sections comparing those features of PIN diode switch designs for unique to high power microwave switches and high power lower frequency (RF-band) switches.

There are many unique market applications, such as the Wireless Communications Market, where new network applications and system designs outpace the component technology needed to support them. Therefore, there are sections that discuss the unique circuit functional requirements appropriate to these newer market applications. Wireless Telecommunications power control circuits are discussed in terms of the role PIN diodes play in providing low distortion, low Bit-Error-Rate (BER) performance for RF Channel components, particularly in next generation multimedia systems such as PCS and UMTS. Additionally, the characteristics of high power HF Band switches are treated in detail as well as those of switches designed for Magnetic Resonance Imaging (MRI) systems.

An appendix on distortion in PIN diode Switches and Attenuators has been included, because of the increased importance of this parameter to RF Channel performance of Wireless Communications Systems. The subject of driver circuits for PIN diode switches and Attenuator circuits is always relevant to any practical component design, and thus has been included in a separate appendix.

PIN Diode Physics topics, such as PIN diode forward and reverse bias operating characteristics and equivalent circuits, stored charge and lifetime, distortion and non-linearity, and thermal impedance, are contained in specific appendices for supplementary and reference material.

We hope that the organization of this material will be found useful by circuit and system designers, for whom this Handbook was written.

Any comments, additions, or deletions would be appreciated.

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CHAPTER - 1

PIN DIODE GENERAL DESCRIPTION

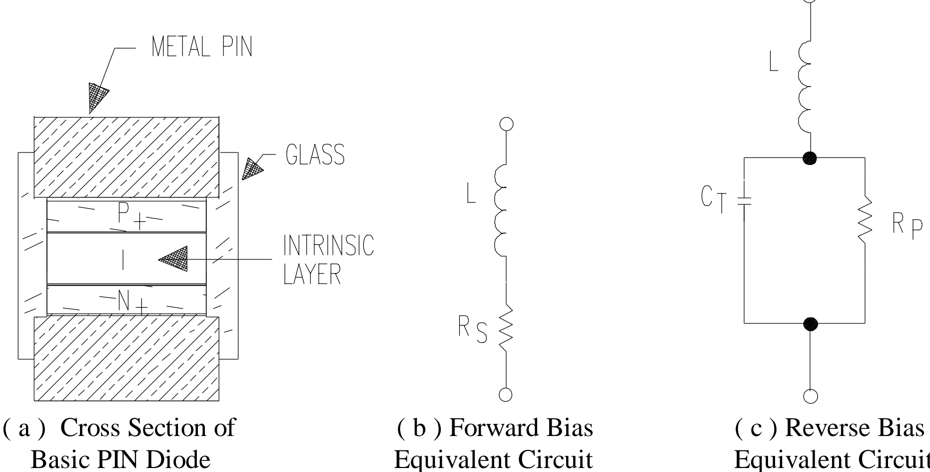
NOTES

PIN DIODE GENERAL DESCRIPTION

This chapter presents a general overview of PIN diode operating characteristics to form an adequate basis for the subsequent chapters on the various PIN diode functional circuits. Supplemental material on PIN Diode Physics is included in the Appendices section of the Handbook.

A microwave PIN diode is a semiconductor device that operates as a variable resistor at RF and Microwave frequencies. A PIN diode is a current controlled device in contrast to a varactor diode which is a voltage controlled device. Varactors diodes are design with thin epitaxial I-layers (for a high "Q" in the reverse bias) and little or no concern for carrier lifetime (Stored Charge).When the forward bias control current of the PIN diode is varied continuously, it can be used for attenuating, leveling, and amplitude modulating an RF signal. When the control current is switched on and off, or in discrete steps, the device can be used for switching, pulse modulating, and phase shifting an RF signal. The microwave PIN diode's small physical size compared to a wavelength, high switching speed, and low package parasitic reactances, make it an ideal component for use in miniature, broadband RF signal control circuits. In addition, the PIN diode has the ability to control large RF signal power while using much smaller levels of control power.

Microsemi PIN diodes offer a unique highly reliable package due to voidless construction, metallurically bonded pin structure, and an extremely rugged SOGO surface passivation. SOGO passivated devices may be driven into reverse voltage breakdown without the reverse voltage characteristic collapsing. Microsemi PIN diodes offer significant electrical and thermal advantages compared to PIN diodes manufactured by other suppliers. The Microsemi PIN diode is generally constructed using a PIN chip that has a thicker I-region, larger cross sectional area and longer carrier lifetime for the same basic electrical characteristics of series resistance (R_S), and capacitance (C_T). This results in PIN diodes that produce lower signal distortion at all frequencies and power levels as well as devices that are capable of handling greater average and peak power than those manufactured by conventional techniques. In addition, since there are no ribbons or wires within the Microsemi's package, large surge currents may be safely handled and the parasitic resistance and inductance are minimized.



(a) Cross Section of Basic PIN Diode

(b) Forward Bias Equivalent Circuit

(c) Reverse Bias Equivalent Circuit

Figure 1.1 PIN Diode and the Corresponding Equivalent Circuits

A drawing of a PIN diode chip is shown in Figure 1.1 (a). The performance characteristics of the PIN diode depend mainly on the chip geometry and the processed semiconductor material in the intrinsic or I - region, of the finished diode. When the diode is forward biased, holes and electrons are injected into the

I-region. This charge does not recombine instantaneously, but has a finite lifetime (τ) in the I-region. If the PIN diode is reverse biased , there is no stored charge in the I-region and the device behaves like a Capacitance (C_T) shunted by a parallel resistance (R_P). These equivalent circuit parameters are defined in the section below. If the d-c voltage across the PIN diode is zero, there remains some finite charge stored in the I-region, but it is not mobile. If operated at zero volts d-c, any PIN diode behaves as a somewhat lossy Capacitor. Some small d-c Voltage (called the "punch-through" Voltage) must be applied to the I-region to sweep out this remaining fixed charge. These ideas are developed farther in Appendix A.

RF ELECTRICAL EQUIVALENT CIRCUITS PARAMETERS OF THE PIN DIODE

FORWARD BIAS EQUIVALENT CIRCUIT

The equivalent circuit for the forward biased PIN diode, Figure 1.1 (b), consists of a series combination of the series resistance (R_s) and a small Inductance (L_s). R_s is a function of the Forward Bias Current (I_f) and this function is shown in Figure 1.2 for the UM 9552 PIN Attenuator Diode. L_s depends on the geometrical properties of the package such as metal pin length and diameter. L_s is a small parasitic element that has little effect on Microsemi PIN diode performance below 1 GHz

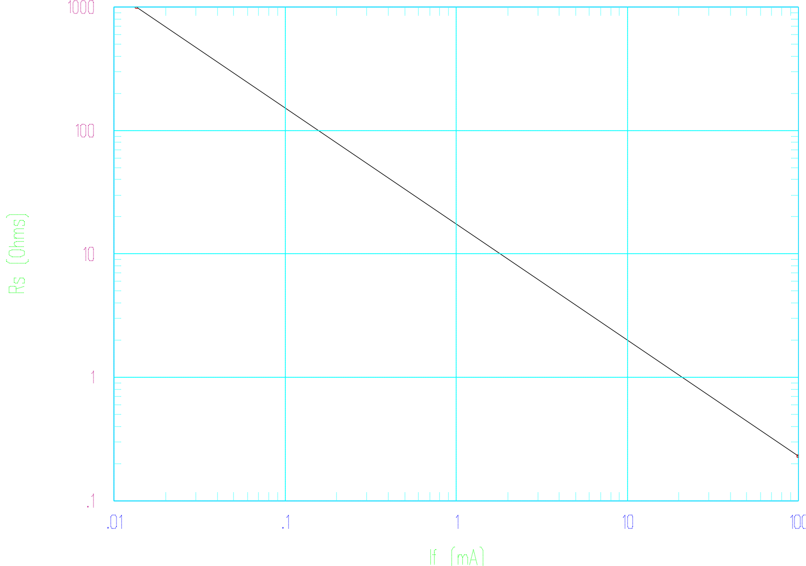


Figure 1.2. Typical Forward Biased Series Resistance vs Bias Current for the UM 9552 PIN Diode

The forward biased PIN diode is a Current Controlled Resistor, which is useful in low distortion Attenuator and Amplitude Modulator Applications. The R_s vs I_f relationship is described as:

$R_s = W^2 / (\mu_n + \mu_p)\, Q$ (Ohms) or $R_s = W^2 / (\mu_n + \mu_p)\, I_f \tau$ (Ohms)

where: $Q_s = I_f\,\tau$, W = I-region Width, I_f = Forward Bias Current, τ = Minority Carrier Lifetime

μ_n = Electron Mobility, μ_p = Hole Mobility

This equation is valid for frequencies higher than the transit time of the I-region: $f > 1300/W^2$ (f in MHz and W in microns). It also assumes that the RF signal does not modulate the stored charge (Appendix A). At lower frequencies, the PIN diode rectifies the RF signal (just as any pn-junction diode would).

REVERSED BIAS EQUIVALENT CIRCUIT

The Reverse Bias Equivalent Circuit consists of the PIN diode Capacitance (C_T), a shunt loss element, (R_p), and the parasitic Inductance (L_s). The defining equation for C_T is:

$$C_t = \varepsilon A / W$$

which is valid for frequencies above the dielectric relaxation frequency of the I-region, ie:

$$f > 1 / 2\pi\rho\varepsilon$$

where ε = dielectric constant of Silicon, A =Diode Junction Area, and ρ = Resistivity of Silicon.

Ct decreases somewhat from 0 Volts to the "Punch-Through" Voltage and remains constant for reverse bias Voltage (V_r) greater than the "Punch-Through" Voltage. The PIN diode's reverse bias Capacitance vs Voltage behavior is different than a pn-junction diode, which exhibits a continuously variable Capacitance vs Reverse Voltage out to the Breakdown Voltage (VBR). The reverse biased PIN diode is easier to Impedance match than the Varactor, because of its flat Ct vs Vr characteristic.

The shunt Loss (G_p) is maximum at 0 Volts and decreases to a fixed value as the reverse bias Voltage is increased. An upper cutoff frequency for the PIN diode could be defined as that frequency at which L_s resonates with the periodic average value of C_t.

LARGE SIGNAL MICROWAVE PIN DIODE OPERATION

Under large RF Power control conditions in the Microwave bands (1 GHz and above), the following bias considerations apply:

Forward Bias Condition:

The PIN diode must be forward biased (Low Loss or ON State) so that the stored charge, Qs, is much larger than the RF induced charge that is added or removed from the I-region cyclically by the RF current. This relationship is shown by the inequality: $Q_s \gg I_{rf} / 2\pi f$

Reverse Bias Condition:

High Frequency versus Low Frequency

A PIN diode, designed for high frequency operation is usually fabricated to have low capacitance because the reactance of the diode in the OFF condition must be large compared to the line impedance. The ratio of the PIN’s area to thickness is adjusted to obtain the desired capacitance. The resistivity or doping level of the I-layer is not critical as long as it is greater than 20 to 50 Ohm-cm for operation at 1 GHz. The transit time and the relaxation frequency requirements are easily obtained.

In contrast operation at low frequencies places more constraints on the PIN designer (< 10 MHz or even more so, below 1 MHz). Low relaxation frequency requires very high resistivity levels for the I-layer. Microsemi uses 10,000 Ohm-cm Silicon to obtain the low relaxation frequency. Long transit time requires very thick I-layers. Microsemi manufactures PIN diodes with I-layer thickness of 500 µm. Large values of Q_S are required to control the RF signal at low frequencies and are very critical in attenuator applications where the dc bias current may not be increased without changing the resistance value of the PIN diode. Large values of Q_S (τ > 0.1millisec) are obtained by careful process control and the use of a good passivating surface for the I-layer.

Above 1 GHz, the period of the microwave signal is much smaller than the PIN diode's minority carrier lifetime(τ). In this case, the reverse bias condition (Isolation State) is such that the PIN diode is biased beyond punch through (Appendix A). If large values of RF current are being switched, the reverse bias voltage must be large enough that the RF voltage during its forward excursion does not induce the flow of RF current through the PIN diode. If the PIN diode becomes warm when operating as a high power switch, the reverse bias voltage should be increased to minimize this effect. The PIN diode's reverse breakdown voltage (VBR) must be large enough so that the reverse excursion of the RF voltage does not cause the flow of avalanche current under reverse bias conditions [1,2]. As shown inFigure 1.3.

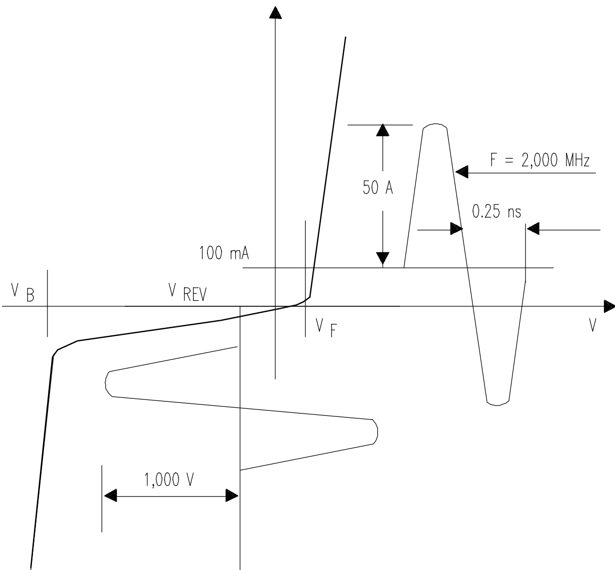


Figure 1.3 RF Voltage and Current Waveforms Superimposed on PIN Diode IV Characteristics

LOW FREQUENCY RF PIN DIODE OPERATION

Below the transit time frequency of the I-region, the PIN diode behaves as a PN junction diode, ie, it rectifies the RF voltage. For frequencies somewhat higher than the transit time frequency but below the Microwave Bands, sufficient reverse bias voltage should be applied to protect the PIN diode from burnout in a high power switch application(Figure 1.3). In this frequency range, lifetime may not be sufficiently large so that the d-c induced stored charge controls the RF power applied. To be completely safe, the reverse bias should be equal to greater than the peak value of the RF Voltage and the VBR should be equal to greater than the peak-to-peak value of the RF Voltage, so that no RF current flows during the positive half of the RF cycle [3,4].

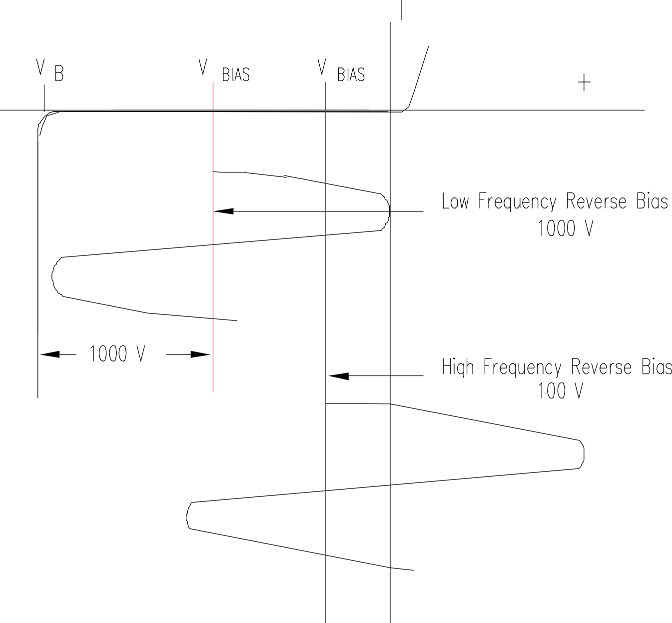


Figure 1.4 L F & H F Voltage Waveforms Superimposed on the I-V Characteristics of a PIN Diode

BIAS-CIRCUIT / RF CIRCUIT ISOLATION

In most applications, it is necessary to provide some degree of isolation between the low-frequency d-c bias circuit and the r-f circuit. Otherwise, RF current can flow into the power supply's output impedance, causing effects that are detrimental to the efficient operation of the power control circuit.

The d-c bias supply is isolated from the RF circuits by inserting a low-pass filter structure between the bias supply and the RF control circuit. For many switch application (Chapter 2), an RF inductor, in series with the bias line, and an RF by-pass capacitor, in shunt with the power supply output impedance, will provide 20 dB or more of d-c / r-f isolation. If higher values of isolation are needed, more complex low-pass filter structures are necessary.

Low-pass filters may significantly increase the switching time of the PIN diode. If a switching time of 100 ns is needed, the low-pass filter must show very little loss to frequencies up to 30 MHz (ie, the filter's cut-off frequency is at least 30 MHz). Shorter switching times require higher filter cut-off frequencies, which may lead to practical construction difficulties. Many commercially available bias tees are not adequate for biasing high power switch prototype circuits because the d-c current rating is too low.

PIN DIODE SWITCHING SPEED CHARACTERISTICS

Switching Speed (T_s) is discussed in detail for specific switch configurations and operating conditions in Chapter 2 and from a diode physics perspective in Appendix A. In switching applications, switching speed is the time required to either fill or remove charge from the I-region. Switching speed depends both on the driver circuit's operating conditions for specific switching states and on the diode's equivalent circuit parameters.

When a PIN diode is forward biased by current, I_F, the current flow results in charge, $Q = I_F \tau$, being stored in the I-region. This stored charge condition causes the PIN diode to be in the low resistance state. If the forward bias current is suddenly removed, the positive and negative charges in the PIN diode will recombine in a time period called τ, the minority carrier lifetime. If a large reverse voltage is applied to

the forward conducting PIN diode, a reverse current, I_R, flows. T_{FR}, or the forward-to-reverse switching time, is expressed in terms of I_F, I_R, and lifetime τ, as

$$T_{FR} = \ln(1 + I_F / I_R)\,\tau \quad \text{(sec.)}$$

The shape of the typical I_F vs time curve, defining T_{FR}, is shown in Figure 1.5.

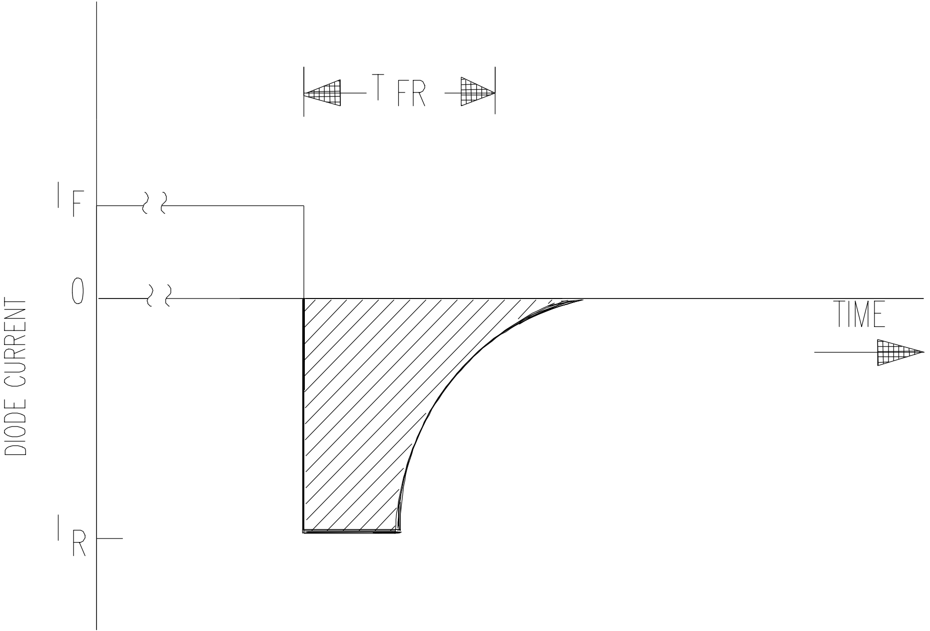


Figure 1.5. PIN Diode Reverse Bias Switching Speed

The speed with which charge is removed from the I-region during turn-off depends on the rise time and amplitude of the switching-voltage pulse applied to the PIN diode. By using spiked waveforms (referred to as overdrive) and by reducing the source impedance of the driver to allow high reverse current to flow, the TFR can reduced substantially.

The time required for the I-region to fill with charge primarily depends on the transit time of the I-region, (ie, the I-region width) and on the reverse voltage and forward bias current that the driver can supply. This reverse-to-forward switching time, T_{RF}, is usually faster than the turn-off time, T_{FR}.

PIN DIODE THERMAL IMPEDANCE

PIN diodes are used to control RF power in circuits such as switches, attenuators, modulators and phase shifters. These PIN diode applications are discussed in detail in the next four chapters. The process of controlling RF power naturally results in some of the RF power being dissipated in the controlling device. The amount of power dissipated is calculated for the various circuit PIN diode circuit configurations in the appropriate chapters.

As a PIN diode dissipates power, its junction temperature begins to rise. The diode's junction temperature depends on the amount of power dissipated, P_d, the ambient temperature T_{amb}, and the thermal impedance, (θ_J), between the diode junction and the diode's ambient temperature. The power rating of a

PIN diode is the power dissipation that will raise the junction temperature from the ambient temperature (usually 25 ^{o}C) to its maximum allowable value, T_{Jmax} (150 ^{o}C).

The maximum power dissipation, Pd, is determined from the relationship:

$$P_d = (T_j - T_a) / \theta_J$$

where T_j is the maximum junction temperature for a Silicon PIN diode (175 ^{o}C) and T_a is the ambient temperature, usually that of the diode's heat sink. P_d is calculated as:

$$P_d = I_{RF}{}^2 R_s + I_{DC} V_{DC}$$

where I_{RF} is the RF current, I_{DC} is the dc current, and R_s is the value of the diode's series resistance at the value of forward bias (d-c) current chosen. Note, that P_d is the maximum power that the PIN diode can dissipate, **NOT** the maximum switched power! The maximum switched power, depends on the PIN diode's bias conditions related to the Characteristic Impedance of the Switch Circuit and the Voltage and Current from the RF Power Source.

WHY YOU SHOULD USE A PIN DIODE

1. Rugged, High Reliability
2. High Voltage Capability — > 2000 Volts
3. High Current Capability — > 25 Amperes continuous
4. High surge Current Capability — > 500 Amperes (1 pulse 8.3 ms ,½ sine)
5. Low Distortion — < -60dBc @ 455 KHz
6. High Power Gain — > 10,000 : 1
7. Fast Switching speed — < 100 ns
8. Small Physical Size
9. Various Thermal Packaging Available
10. RF Relay Replacement - mechanical, mercury, etc.

CHAPTER -2

PIN DIODE RF SWITCHES

NOTES

CHAPTER 2 PIN DIODE SWITCHES

INTRODUCTION

A switch is an electrical component for opening and closing the connection of a circuit or for changing the connection of a circuit device [1]. An "Ideal Switch" exhibits zero resistance to current flow in the "ON" state and infinite resistance to current flow in the "OFF" state. A practical switch design exhibits a certain amount of resistance in the "ON" state and a finite resistance in the "OFF" state.

The use of PIN diodes as the switching element in microwave circuits is based on the difference between the PIN diode reverse and forward bias characteristics [Chapter One]. At lower microwave frequencies, $f < 2$ GHz), the PIN diode (including package parasitics) appears to be a very small impedance under forward bias and a very large impedance under reverse bias. It is the difference in performance between forward and reverse bias states upon which switch operation relies.

Most switch designs to be considered use a difference in reflection, rather than dissipation, to obtain switch performance. Very little power is dissipated by the diode itself, thus permitting small devices to control relatively large amounts of microwave power. Thus, PIN diode switches are reactive networks, where losses are a second order effect. In subsequent sections, we will see that switch circuits resemble filter circuits in many ways.

FUNDAMENTAL PARAMETERS THAT DESCRIBE PIN DIODE SWITCH PERFORMANCE

ISOLATION:

Physically, Isolation is a measure of the microwave power through the switch, that is not transferred to the load, both from Attenuation Loss and Reflection Loss, when the switch is OFF.

As a practical matter, Isolation is a measure of how effectively a PIN Diode Switch is turned OFF. It is determined by calculating the difference between the power measured at the switch output port with the switch biased ON and the power measured at the switch output port with the switch biased OFF.

$$\text{Isolation (dB)} = (P_{out})\text{on (dBm)} - (P_{out})\text{off (dBm)} \qquad \text{Equation 2.1}$$

This equation avoids the problem of accounting for the Transmission Loss through the physical structure of the PIN Diode Switch (all switches have some finite Transmission Loss). Transmission Loss is present whether the switch is ON or OFF.

INSERTION LOSS:

Insertion Loss (I_L) is the Transmission Loss through the physical structure of a PIN diode switch. In the forward biased case (the ON state), large values of bias current plus microwave current may flow through the switch structure, causing significant Ohmic Loss. In the reverse bias case (the OFF or Isolation state), only small values of leakage current flow through the switch, so the reverse bias loss is small.

If the switch is mechanically and thermally designed properly, Ohmic Losses and Thermal Dissipation are minimized and Insertion Loss is relatively low ($I_L < 0.25$ dB).

Insertion Loss is a particularly critical parameter for the Communications System designer. Insertion Loss absorbs signal power, causing the system's Noise Figure to increase by the amount of the Insertion Loss.

PIN DIODE POWER HANDLING LIMITATIONS

The RF System Requirement that usually determines the choice of the particular PIN Diode to be used is the RF power that the switch must handle. The PIN Diode characteristically has relatively wide I-region and can therefore withstand larger RF Voltages than Varactors or microwave Schottky diodes. In Chapter One (Large Signal PIN Diode Operation) the forward and reverse bias conditions, necessary to insure safe high power switch operations were discussed.

In this Chapter, the switch's Power Dissipation is considered as another limiting factor in determining the maximum RF power level that the PIN diode switch can control without overheating. Power Dissipation depends on R_s (which is a function of the forward bias current) relative to Z_o, on the input power to the switch, P_a, as well as on the switch connection chosen. P_d is a very important rating for a PIN switching diode and is given by all manufacturers.

Finally, the maximum RF power that the PIN diode is capable of switching depends on the incident power, P_a, Z_o, the switch connection type, average Dissipated Power (P_d), and on the Reverse Breakdown Voltage (VBR) rating. This parameter is also supplied by most manufacturers, with the stipulation that Z_o = 50 Ohms and that the switch circuit is series-connected.

RF AND MICROWAVE SWITCH DESIGN CONFIGURATIONS

In this and subsequent sections, circuit diagrams of simple and compound switches are given, as well as additional performance information needed to design a switch. We assume in this development, that the individual switch structure is a symmetrical linear two port network and that the characteristic impedance (Z_o) of the input power source, the switch structure, the load impedance, and any transmission lines connecting these components are 50 Ohms. For the more general case, where the input Z_o is not equal to the output Z_o, the reader is referred to reference [2] or any general text on general network theory.

SINGLE POLE SINGLE THROW SWITCHES

SERIES SPST SWITCH

The PIN diode SPST can be used in broadband designs. The maximum isolation (I_{SO})obtainable depends on the diode's Capacitance (C_t). The Insertion Loss (I_L) and Power Dissipation (P_d) depend on the diode's forward biased Series Resistance (R_s). The equations for I_{SO} & I_L and the performance characteristics are given below.

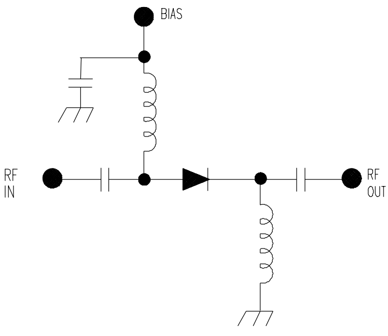


Figure 2.1 Series SPST Switch

For Series SPST Switches:

$$I_L = 20 \log \{ 1 + R_s / 2 Z_o \}$$

$$I_{SO} = 10 \log \{1 + 1 /(4 \pi f C_t Z_o)^2 \}$$

Power Dissipation (P_d) :

$$P_d = \{ 4 R_s Z_o / (2Z_o + R_s)\}^2 P_{av} \quad \text{Watts}$$

where P_{av} is the maximum available power, $V_g^2 / 4 Z_o$ (Watts).

These equations pertain only to matched SPST switches. For VSWR (σ) > 1.0, multiply these equations by the factor [$2\sigma / \sigma + 1$], designated "sigma", to calculate P_d.

Peak RF Current (SPST)

$I_p = \sqrt{2 P_{av} / Z_o}$ Amps

Peak RF Voltage (SPST)

$V_p = \sqrt{8 Z_o P_{av}}$ Volts

If the series SPST switch is not matched, multiply the above equations by the factor "sigma".

SHUNT SPST SWITCH

The Shunt SPST Switch (Figure 2.2) offers high isolation over a broad frequency range (approximately 20 dB for a singled diode switch). Insertion Loss is low because there are no switch elements in series with the transmission line. The diode is electrically and thermally grounded to one side of the transmission line and has higher P_d capability than the SPST circuit. I_{SO} and P_d are functions of R_s. I_L primarily depends on C_t. The design equations are given below.

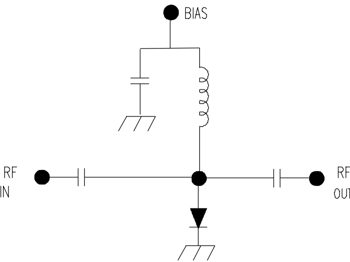


Figure 2.2 Shunt SPST Switch

For Shunt SPST Switches:

$I_L = 10 \log \{1+ (\pi f C_t Z_o)^2\}$ dB

$I_{SO} = 20 \log \{1 + Z_o / 2 R_s\}$ dB

Power Dissipation (Forward Bias):

$P_d = 4 R_s Z_o / (Z_o + 2 R_s)^2 P_{av}$ atts

Power Dissipation (Reverse Bias)

$P_d = \{Z_o / R_p\} P_{av}$ Watts

(where P_{av} is the maximum available power)

Peak RF Current (Shunt Switch)

$I_p = \sqrt{8P_{av} / Z_o}$ Amps

Peak RF Voltage (Shunt Switch)

$V_p = \sqrt{2 Z_o P_{av}}$ Volts

If the shunt switch circuit is not matched, multiply the above equations by the "sigma" factor.

SINGLE POLE DOUBLE THROW SWITCHES

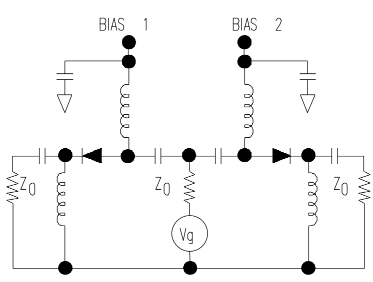


Figure 2.3 Series SPDT Switch

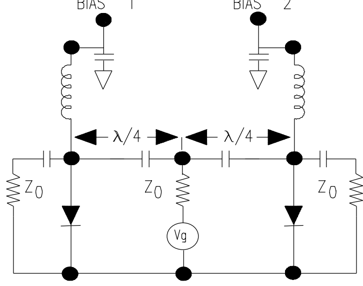


Figure 2.4 Shunt SPDP Switch

The simplest example of the more general Single Pole Multi-throw Switch structure is the Single Pole Double Throw Switch, in which the signal power in a single input transmission line can be connected to either of two output transmission lines.

If the SPDT switch is symmetrical, each switch branch performs like the SPST equivalent; but the Isolation of multi -throw switches is increased by 6 dB. This effect occurs because the OFF branch is shunted by the ON branch and its 50 Ohm termination, causing the RF Voltage across the OFF diode to be 50% less than would be the case for the equivalent SPST switch.

The Shunt SPDT Switch design in Figure 2.4 enhances the electrical performance of this switch by inserting quarter-wavelength transmission lines between the signal power source and the PIN diodes. The isolation of this design is approximately double (ie, 3 dB) that of the Shunt SPST Switch plus 6 dB due to the effect of the multi-throw switch junction. However, the bandwidth is now constrained to less than an octave.

MULTI-THROW SWITCHES

Multi-throw switches are difficult to realize using only shunt diodes. A band-limited shunt multi-throw switch (less than one octave) as shown in Figure 2.5, uses two cascaded quarter-wavelength sections, each terminated by a shunt diode. This configuration gives the OFF branch a high input impedance at the common (signal source) port to prevent impedance “loading” of the ON arm that would otherwise occur.

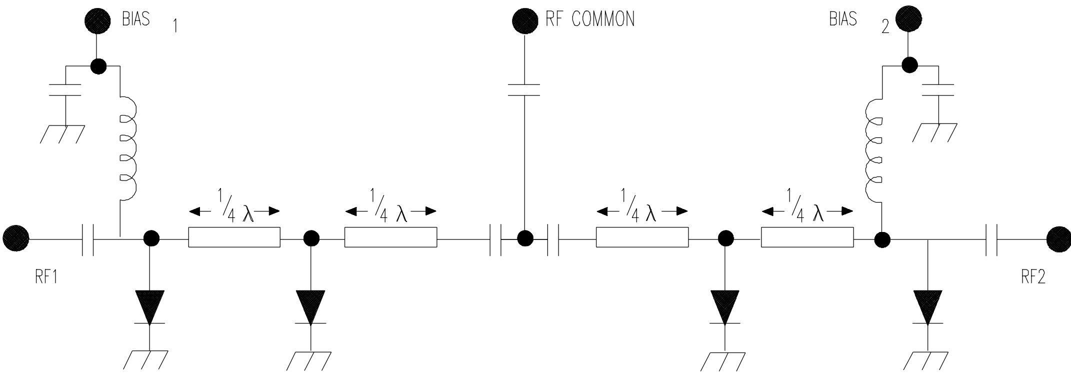


Figure 2.5 Band-Limited Shunt Multi-throw Switch

These configurations can achieve very high isolation (70 to 90 dB) with additional shunt diodes and transmission line sections. These designs would be even more constrained in bandwidth and Insertion Loss increases as sections are added. In the microwave bands, isolation is limited by cross coupling between switch components, causing some direct signal feed-through between input and output ports.

COMPOUND SWITCHES

Compound Switches differ from multi-throw switches in that series-shunt switches are used in combinations to improve overall switch performance. The broad band Insertion Loss of the series switch is combined with the broad band Isolation of the shunt switch in a number of combinations to follow.

SERIES-SHUNT COMPOUND SWITCHES

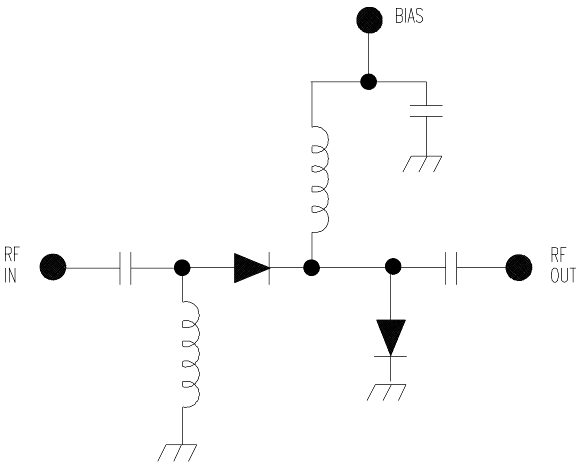


Figure 2.6 Series-Shunt SPST Switch

TEE COMPOUND SWITCHES

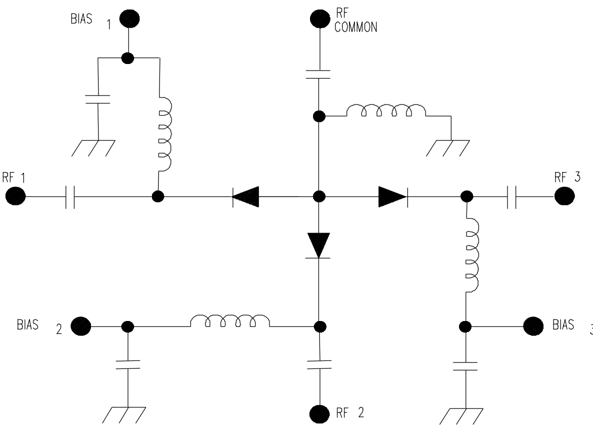


Figure 2.7 TEE SP3T Switch

The simplest compound switches are the Series-Shunt Switch (Figure 2.6) and the TEE Switch (Figure 2.7). These circuits offer improved overall performance but the added circuit complexity degrades the VSWR and the Insertion Loss. Since all diodes are not simultaneously biased in one state or the other, there is an increase in bias circuit complexity. A summary of overall performance parameters for the Series and Shunt SPSTs and for the

Series-Shunt and TEE Compound Switches is shown for comparison in Table I. Performance parameter trade-off is inevitable in any practical switch design.

TABLE I. SUMMARY OF FORMULAS FOR SPST SWITCHES

TYPE	ISOLATION (dB)*	INSERTION LOSS (dB)
SERIES	$10\log\left[1+\frac{1}{\left(4\pi fC_TZ_0\right)^2}\right]$	$20\log\left(1+\frac{R_S}{2Z_0}\right)$
SHUNT	$20\log\left(1+\frac{Z_0}{2R_S}\right)$	$10\log\left[1+\left(\pi fC_TZ_0\right)^2\right]$
SERIES-SHUNT	$10\log\left[\left(1+\frac{Z_0}{2R_S}\right)^2 + \frac{1}{4\pi fC_TZ_0}\left(1+\frac{Z_0}{R_S}\right)^2\right]$	$10\log\left[\left(1+\frac{R_S}{2Z_0}\right) + \left(\pi fC_T\right)^2\left(Z_0+R_S\right)^2\right]$
TEE	$10\log\left[1+\left(\frac{1}{2\pi fC_TZ_0}\right)^2\right]$ $+10\log\left[\left(1+\frac{Z_0}{2R_S}\right)^2+\left(\frac{1}{4\pi fC_TR_S}\right)^2\right]$	$20\log\left(1+\frac{R_S}{2Z_0}\right)$ $+10\log\left[1+\left(\pi fC_T\right)^2\left(Z_0+R_S\right)^2\right]$

* For SPNT Switch, Add 6 dB

TUNED SWITCHES

A simple tuned shunt SPDT switch was shown in Figure 2.4. The presence of quarter-wavelength transmission lines constrain the overall bandwidth but enhance the switch's performance over that bandwidth. Similarly, many RF switch applications operate over a limited frequency band. Distributed lines can be used to improve switch performance as the following examples show.

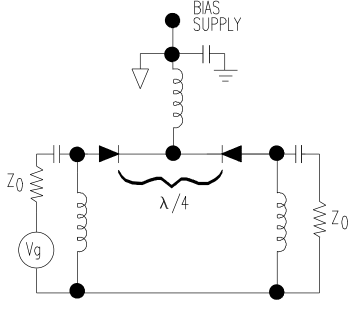


Figure 2.8 Tuned Series SPST Switch

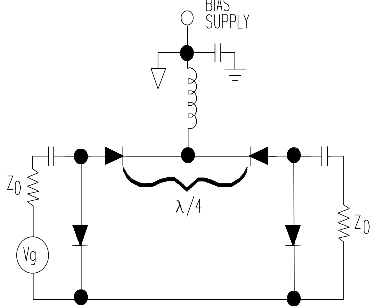


Figure 2.9 Tuned Shunt SPST Switch

The Insertion Loss and Isolation for the circuits in Figures 2.8 & 2.9 can be calculated from the formulas in Table I. The total diode resistance, R_S, used in these calculations is twice that of a single diode SPST switch, unless the bias current is increased to off-set this effect. The maximum Isolation obtainable, using multiple diodes spaced a quarter-wavelength, is twice the dB value obtainable with a single diode switch.

A further increase in Isolation can be obtained by adding more quarter-wavelength sections to these designs. Such tuned switches have band widths less than 10 %, which is quite adequate for wireless radio applications (reference Chapter 6).

TUNED SERIES SPST SWITCHES

Quarter-wavelength spacing reduces the maximum RF voltage across each diode to half of that which would appear across a single diode switch.

Even if the series diode had no quarter-wavelength spacing, the Isolation would increase by 6 dB, because the effective Capacitance is half of that of a single diode.

If this reduction in Capacitance is not primary to the design objectives, diodes with increased Capacitance could be used to increase the power handling capability of the switch

TUNED SHUNT SPST SWITCHES

The maximum isolation obtainable using a Tuned Shunt SPST Switch is twice the dB value obtainable using only a single diode switch. Figure 2.4 shows a Double-throw Tuned Shunt Switch. In this circuit, the Capacitive Reactance of one diode is transformed by the quarter-wavelength line (into an Inductive Reactance) and resonates with the Capacitive Reactance of the second diode. This effect lowers the switch Insertion Loss by about 50%, but narrows the operating bandwidth. As with the Tuned Series SPSTs, quarter-wave spacing can be use higher power diodes with larger values of Capacitance (C_t), but the effective bandwidth of the switch is lowered considerably.

LUMPED CIRCUIT EQUIVALENT OF QUARTER-WAVELENGTH TRANSMISSION LINE

Quarter-wavelength techniques, using distributed line elements, are impractical at frequencies below UHF because of their physical size. Quarter-wavelength lines can be simulated with lumped circuit elements in a network such as that shown in Figure 2.10. The equations for calculating the equivalent L & C values are also shown.

$L = Z_o / 2\pi f_o$ (H)

$C = 1/2\pi f_o Z_o$ (F)

Figure 2.10 Lumped Circuit Equivalent of Quarter Wave Line

TRANSMIT - RECEIVE SWITCHES

Transmit-Receive Switches are a class of Tuned Series-Shunt SPDT Switch, used by designers of Communications Transceivers to alternately connect the transceiver's antenna to either the Transmitter or to the Receiver. Figure 2.11 shows the typical T/R quarter line switch and its lumped circuit equivalent.

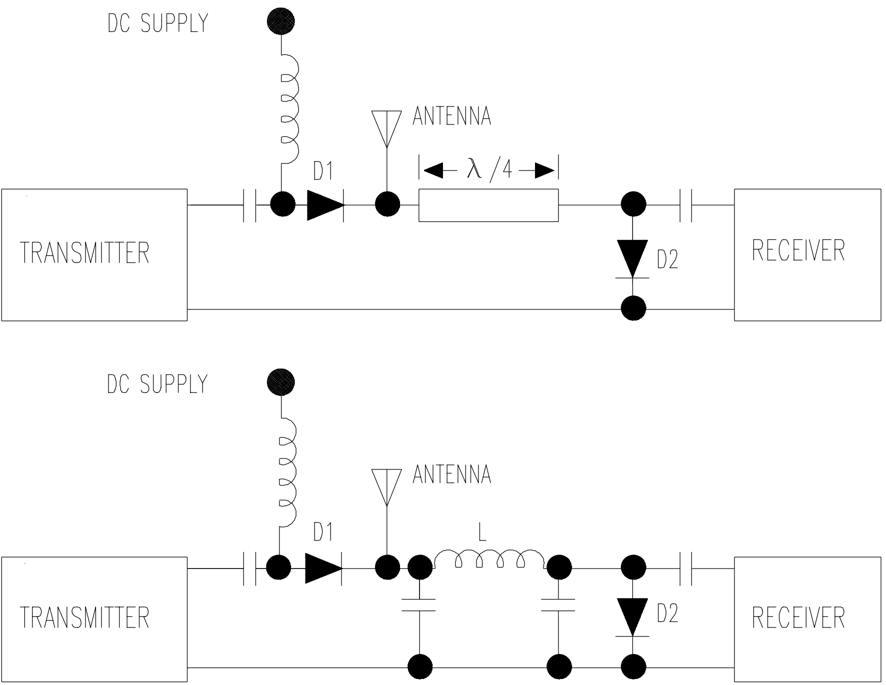


Figure 2.11 Quarter-Wavelength Antenna Switches

The quarter -wavelength line T/R Switch uses the unique property of the quarter-wavelength impedance transformer [3]. Ordinarily, the quarter-wavelength line is used to match two network elements of unequal impedance over a narrow band. If Z_1 and Z_2 are the unequal impedances, then they will be matched if the characteristic impedance of the transformer, Z_o, is related to Z_1 & Z_2 by the equation:

$$Z_o^2 = Z_1 \times Z_2$$

A 25 Ohm signal source can be matched to a 100 Ohm load if they are connected by a quarter wave line of characteristic impedance Z_o = 50 Ohms.

The T/R Switch uses this property to protect the Receiver. Z_o is fixed (usually 50 Ohms) and Z_1 is either the low R_s of a forward biased diode or the isolation state (nearly open circuit) of the reversed biased diode. If Z_1 is nearly a short circuit, the input impedance (Z_2) to the quarter wave line is nearly an open circuit. The transmitter and antenna are disconnected from the receiver. Similarly, when Z_1 is nearly an open circuit (high Impedance), the transmitter is disconnected from the antenna and the receiver is connected to the antenna.

The quarter-wavelength T/R switch is a relatively narrow band SPDT used in many Wireless Telecommunication Transceiver designs. The quarter-wavelength line constrains the bandwidth to 5% to 10%, which is adequate for most communications applications. When both diodes (D_1 & D_2) are forward biased, the transmitter is connected to the antenna and the receiver is protected by the low R_s of D_1 terminating the quarter-wavelength line. When D_1 & D_2 are reverse biased, the transmitter port is isolated by the high reactance of D_1 and the quarter-wavelength line (terminated in an open circuit), and the Receiver port is connected to the Antenna.

The biasing scheme is very simple, requiring only one RF Choke Coil and a few d-c Blocking Capacitors. Greater than 30 dB isolation and less than 0.25 dB insertion loss can be obtained with a UM9401, which has an R_s of 1 Ohm and a C_t of 0.75 pF.

The maximum power, P_{av}, that this T/R switch can handle depends on the power rating of the PIN diode, Pd, and the forward biased diode resistance, R_s. If the antenna has a mismatch (VSWR = σ), P_{av}, is given by the equation:

$$P_{av} = P_d Z_o / R_s \quad \{(\sigma + 1) / 2\sigma\}^2$$

If the antenna is totally mismatched (perhaps the connection is broken), P_{av} is given by:

$$P_{av} = P_d Z_o / 4 R_s$$

We may observe further, that the RF current flowing in both D_1 & D_2 are nearly the same and so, both diodes dissipate about the same amount of RF power.

BROADBAND ANTENNA SWITCHES

If more than 10 % bandwidth is required, more complex switch structures are required. The simplest broad band antenna switch to construct uses two series diodes in a Compound Switch configuration (similar to Figure 2.7) and is shown here as Figure 2.12.

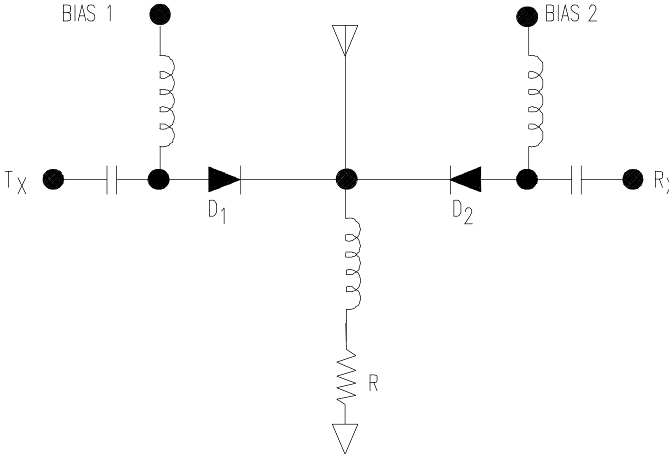


Figure 2.12 Broadband Antenna Switch

Figure 2.12 is a more broad band SPDT switch, but the biasing scheme is more complex, requiring two bias tees and a d-c return coil, because D_1 & D_2 are alternately biased forward or reverse now. When the Transmitter is ON (and the Receiver is OFF), D_1 is forward biased and D_2 is reverse biased. D_1 is reverse biased and D_2 is forward biased when the Receiver is ON and the transmitter is OFF.

The Transmit / Receive isolation state depends solely on the reverse bias Capacitance of D_2, and this becomes the upper frequency limitation of the switch. The Isolation can be increased by using one of the techniques discussed in the "Tuned Switches" section. If D_2 is replaced by two similar PIN diodes is series, the Isolation increases by 6 dB, without reducing the bandwidth significantly. Of course, two diodes will represent an increase in Insertion Loss unless the bias current is increased to off-set the increase in R_S.

Although PIN diode parasitic reactances somewhat limit the bandwidth over which low Insertion Loss and high Isolation can be achieved, the operating bandwidth can also be limited by the bias network, which is a filter network that isolates the d-c bias current from the RF circuit components. The frequency response of this bias network should be measured with the PIN diodes removed from the switch circuit.

D_1 is selected primarily based on its power handling capability. The UM2101 series is recommended for HF Band and the UM4001 or UM4901 for VHF, UHF, and L-Band applications, either in the axial leaded (B package) or insulated stud (D package) because of their excellent thermal properties. For SMT circuit construction, the UPP9401 is recommended for D_1. D_2 is not exposed to high RF currents and therefore should be selected for low Capacitance and low distortion. The 1N5767, the UM7301B, and the UPP1002 (SMT) are recommended for D_2.

As an example, if the UM9401 is used as D_1 and the 1N5767 is used as D_2, the receiver isolation at 50 MHz will be greater than 40 dB, and at 500 MHz, greater than 20 dB.

HIGH POWER BROADBAND ANTENNA SWITCH

An example of a high power broad-band antenna switch, designed to operate over the 10 to 100 MHz band, is shown in Figure 2.13.

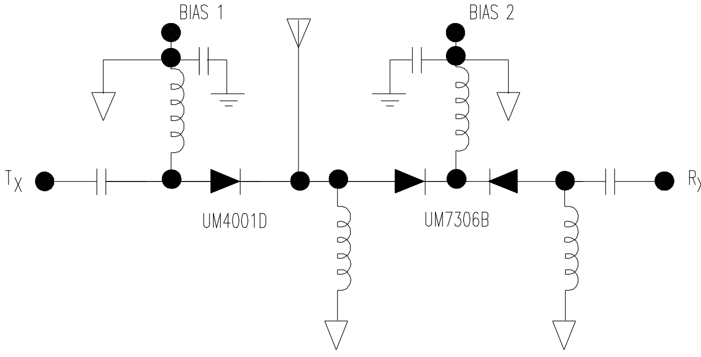
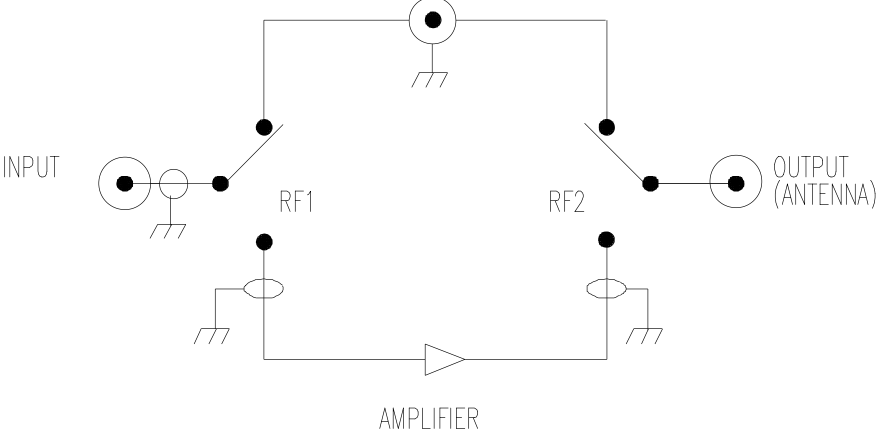


Figure 2.13 High Power Broadband Antenna Switch

This switch can control 1 KW transmitter power with excellent distortion performance (IM3 < -80 dBc).
The forward bias into Bias Terminal 1 is 1 Ampere, for low power dissipation in the transmitter diode and reverse bias of 500 Volts (at Bias Terminal 2) so that excessive RF current does not flow in the OFF state. HF Band (2 to 30 MHz) switches should use the UM2010 series and MF Band (0.3 to 3 MHz) switches should use the UM2310 series of PIN diodes.

MUPTIPLE POLE-MULTIPLE THROW SWITCHES (M x N SWITCHES)

So far, we have only discussed single pole, single or multiple throw switches. A Switch Matrix is a generalization of the concept of the M x N Switch, in which any one of M inputs can be connected to any one of N outputs by means of the network of interconnecting switches. Reference [2] discusses this generalized case.



2.14 Double Pole - Double Throw Switch

The simplest case is the Double Pole-Double Throw Switch or Transfer Switch, which is quite important to RF circuit designers. The DPDT Switch allows a pair of input terminals to be connected to either of two pairs of output terminals as in Figure 2.14. The performance of each pair of connections can be analyzed as a SPST Switch. The DPDT Switch will be discussed in detail in Chapter 7, when it is used as a Transfer Switch for an Amateur Radio Transmitter Antenna. The application is to replace relays in RF Power Amplifiers.

DEVICE	HIGH VOLTAGE >2000 V	HIGH AVERAGE POWER >100 W	HIGH PEAK POWER >10 KW	HIGH POWER CW DUPLEXERS >100 W	ANTENNA SWITCHING >100 W	HIGH FREQUENCY > 1GHz	LOW FREQUENCY <10 MHz	ULTRA LOW FREQUENCY <1 MHz	L C R <
HUM2020	X	X	X	X	X		X		
UM2100		X	X	X	X		X	X	
UM2300		X	X	X	X		X	X	
UM4000		X	X	X	X		X		
HUM4020	X	X	X	X	X		X		
UM4300		X	X	X	X		X		
UM7000		X	X	X					
UM7100		X	X	X					
UM7200									
UM7300		X	X	X	X		X		
UM7500		X	X	X					
UM9401		X	X	X					
UM9415					X				
UMM5050	X		X	X	X				
UPP9401		X		X					
UPP1004		X		X					

CHAPTER - 3

PIN DIODE RF ATTENUATORS

NOTES

PIN DIODE VARIABLE ATTENUATORS

INTRODUCTION

An Attenuator [1] is a network designed to introduce a known amount of loss when functioning between two resistive impedances: $Z_{in} = Z_1$ and $Z_{out} = Z_2$. Z_1 and Z_2 are defined to be terminal impedances to which the attenuator is connected.

MATCHED ATTENUATORS

If the input of the attenuator is matched to Z_1 and the output to Z_2, the circuit is a matched attenuator and the loss is entirely due to Transmission Loss and not to Reflection Loss. The source (input) and the load (output) may be reversed since resistive networks are reciprocal. If $Z_1 = Z_2$, the resulting matched attenuator design is said to be symmetrical, or to exhibit network symmetry. Matched Attenuator Networks may be either balanced or un-balanced (with respect to ground), depending on the exact nature of the source impedance and the load impedance.

Examples of the principle attenuator configurations and their balanced, unbalanced, and symmetrical forms, appear in figures 3.1, 3.2, and 3.3. These will be referred to later in the chapter as PIN diode attenuator designs are obtained.

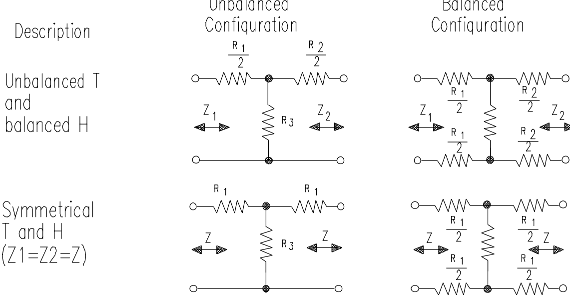


Figure 3.1 Unbalanced T, Balanced H, and Symmetrical T and H

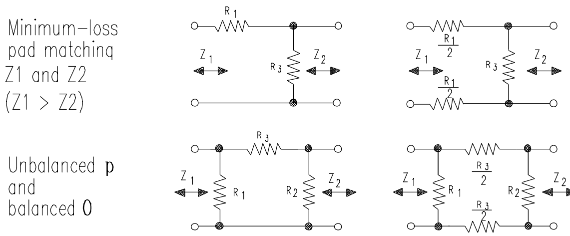


Figure 3.2 Unbalanced ✍, Balanced O, and Symmetrical ✍ and O

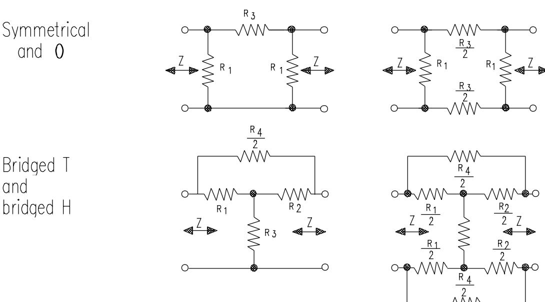


Figure 3.3 Bridged T and Bridged H

Design equations for the unbalanced - symmetrical cases are given below, because of their usefulness in later sections. Symbols used in these design equations have the following meaning:

Z_1 and Z_2 are the terminal Impedances (resistive) to which the attenuator is matched.

$Z = Z_1 = Z_2$ (Symmetrical Case)

N is the ratio of the power absorbed by the attenuator from the source, to the power delivered to the load.

K is the ratio of the attenuator input current, to the output current into the load.

$K = (N)^{1/2}$ for the symmetrical case.

A = attenuation (dB) = 10 log(N) or 20 log (K)

SYMMETRICAL T

$R_1 = Z\ [1 - 2 / (K + 1)]$ $\qquad$ $R_3 = 2Z / [K - 1 / K]$

SYMMETRICAL ✌

$R_1 = Z\ [1 + 2 / (K - 1)]$ $\qquad$ $R_3 = Z\ [K - 1 / K] / 2$

BRIDGED T

$R_1 = R_2 = Z$ $\qquad$ $R_3 = Z / (K - 1)$ $\qquad$ $R_4 = Z\ [K - 1]$

Design equations for the other cases are given in Reference [1].

REFLECTIVE ATTENUATORS:

If the matched condition is not required, simpler networks can be designed as reflective attenuators. These may consist of a simple variable series or a shunt resistive element, that attenuates by exhibiting the necessary mismatch or reflection on the transmission line. In these instances, the attenuation loss is almost entirely due to Reflection Loss although some small amount of Transmissiom Loss may occur. Examples of Reflective Attenuators occur later in this chapter.

PIN ATTENUATOR DIODES

All the basic attenuator configurations can be realized by inserting Current Controlled Resistors (PIN Diodes) in the place of the variable resistances in Figures 3.1, 3.2, and 3.3. In the case of the Symmetrical Microwave Bridged T Attenuator, $R_1 = R_2 = Z_o = 50$ Ohms, and R_3 and R_4 are the variable resistors, replaced by PIN diodes.

Variable attenuators, with PIN diodes as the variable resistance elements, use the forward biased resistance characteristic (Figure 3.4) of the device over nearly its complete forward bias range. The extremely low current range is to be avoided because (see Appendix A) at low current values, the PIN diode's stored charge ($Q_s = I_f$ x ☺) is small and the diode may rectify, causing the attenuator's signal distortion to increase.

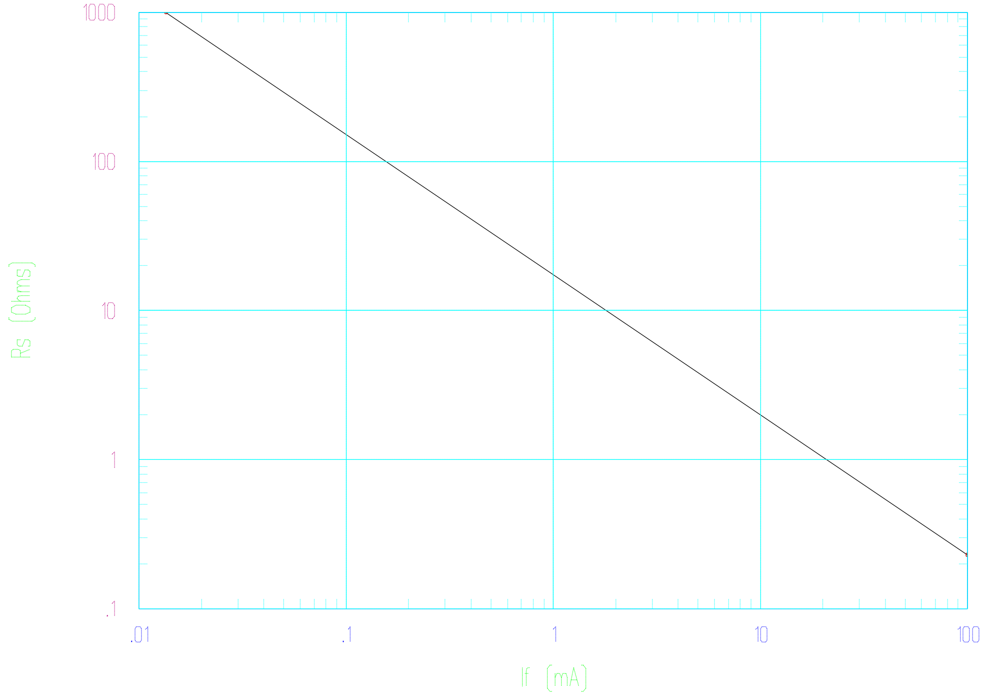


Figure 3.4 Typical Forward Biased Resistance vs Current, UM9552

PIN DIODE ATTENUATOR CIRCUIT APPLICATIONS

PIN diode attenuator circuits are used in automatic gain control (AGC) circuits and power leveling applications. They are also used in high power modulator circuits, which is the subject of Chapter 4. A typical AGC configuration is shown in Figure 3.5.

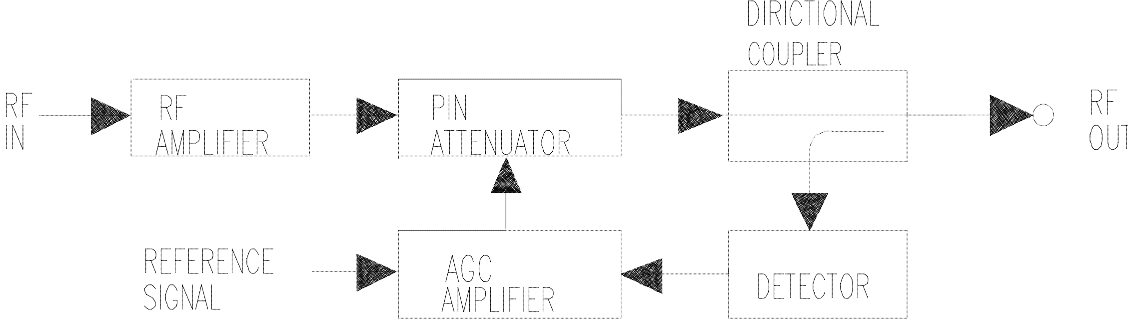


Figure 3.5 RF AGC / Leveler Circuit

The PIN diode attenuator may be a simple reflective attenuator, such as a series or shunt diode mounted across the transmission line. Some AGC attenuators are more complex networks that maintain impedance match to the input power and load as the attenuation is varied across its dynamic range. Other methods are used to implement the

AGC function, such as varying the gain of an RF transistor stage. The PIN diode AGC circuit results in lower frequency pulling and lower signal distortion.

Microsemi Corp. provides a number of PIN diodes designed for attenuator applications, such as the UM2100, UM7301B, UM4301B, UM9552, and the UM9301, which can provide high dynamic range and low signal distortion at frequencies from 100 KHz to 2 GHz. These devices are available in packages designed for standard PC board construction or in packages suitable for Surface Mount Technology.

MICROWAVE MATCHED ATTENUATOR CIRCUITS

The design equations for various matched attenuator circuits configurations have already been given. We now look at the practical implementation of these designs for microwave attenuators.

QUADRATURE HYBRID ATTENUATORS

Quadrature hybrids are commercially available from 10 MHz to 2 GHz, with inherent bandwidths up to a decade. Figures 3.6 and 3.7 are typical quadrature hybrid circuits with series or shunt configured PIN diodes. For 50 Ohm Quadrature Hybrids and branch lines, the attenuation as a function of diode resistance is shown in Figure 3.8.

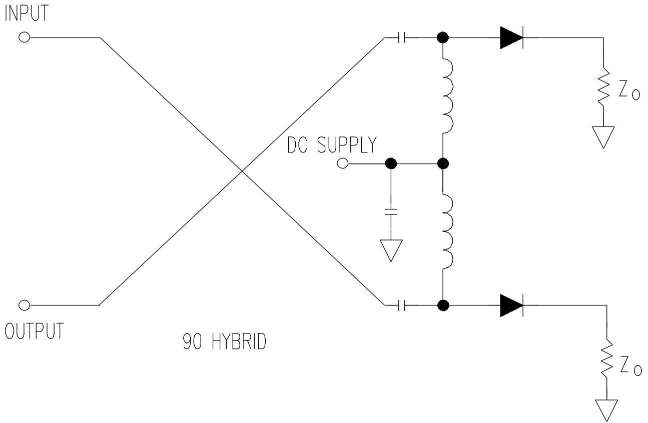


Figure 3.6 Quadrature Hybrid Matched Attenuator (Series Mounted PIN Diodes)

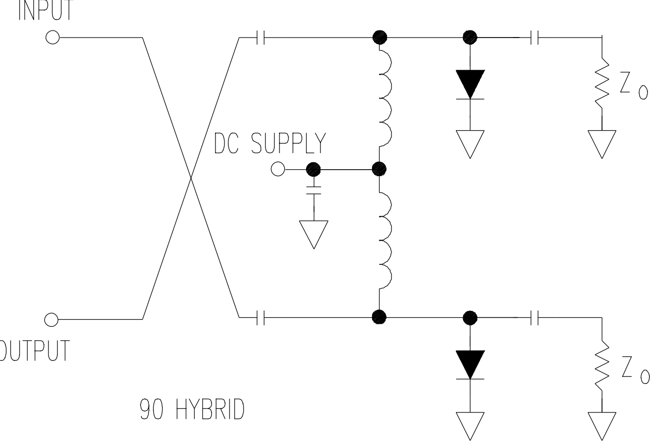


Figure 3.7 Quadrature Hybrid Matched Attenuator (Shunt Mounted PIN Diodes)

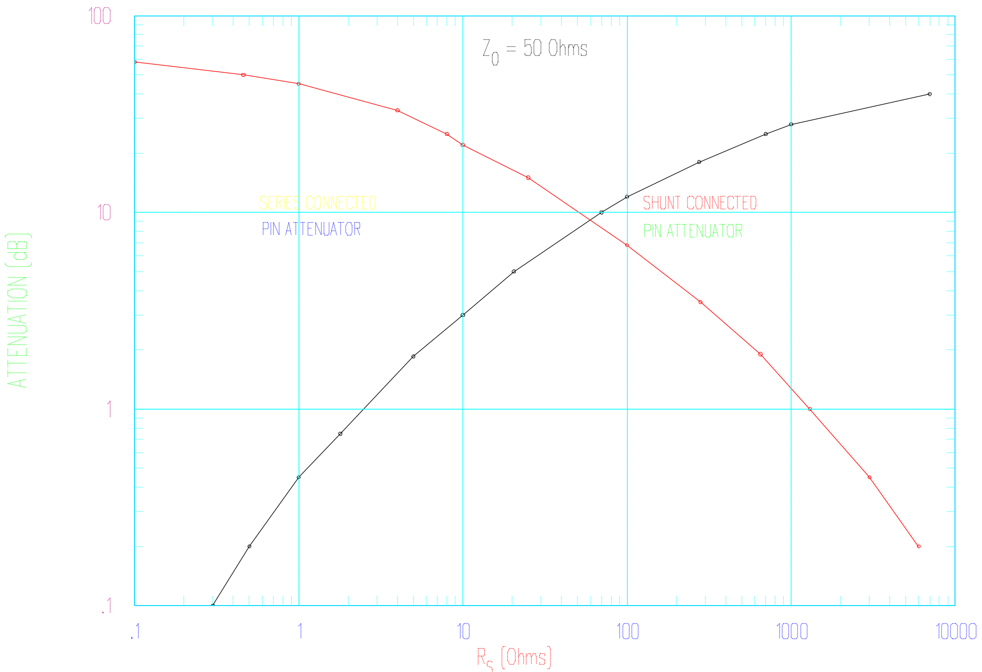


Figure 3.8 Attenuation of Quadrature Hybrid Attenuators

The following equations summarize the performance of these quadrature hybrid attenuators:

Series Connected PIN Diodes	Shunt Connected PIN Diodes
Attenuation = $20 \log \{1 / (1 + 2Z_o / R_s) \}$, dB	Attenuation = $20 \log \{1 / (1 + 2R_s / Z_o)]\}$, dB

The quadrature hybrid configuration can control twice the power of the simple series or shunt diode attenuators because the incident power is divided into paths by the hybrid. Reference [1] shows that the maximum power dissipated in each diode is only 25 % of the total incident power and this occurs at the 6 dB value of attenuation. However, the branch load resistors must be able to dissipate 50% of the total incident power at maximum attenuation. The purpose of the branch load resistors is to make the attenuator less sensitive to differences between individual diodes and to increase theattenuator power handling by 3 dB.

Both types of hybrid attenuators exhibit good dynamic range. The series configured hybrid attenuator is preferable for attenuation levels greater than 6 dB, whereas the shunt configured hybrid attenuator is preferable for attenuation ranges below 6 dB.

QUARTER-WAVE ATTENUATORS

Matched attenuators can also be configured using quarter-wavelength circuit techniques, using either lumped or distributed circuit elements. A quarter-wavelength matched attenuator with series connected diodes is shown in Figure 3.9 and with shunt connected diodes in Figure 3.10. Performance equations are given below the circuit diagrams, and the attenuation vs R_s characteristics are plotted in Figure 3.11 for a transmission system with a characteristic impedance of 50 Ohms.

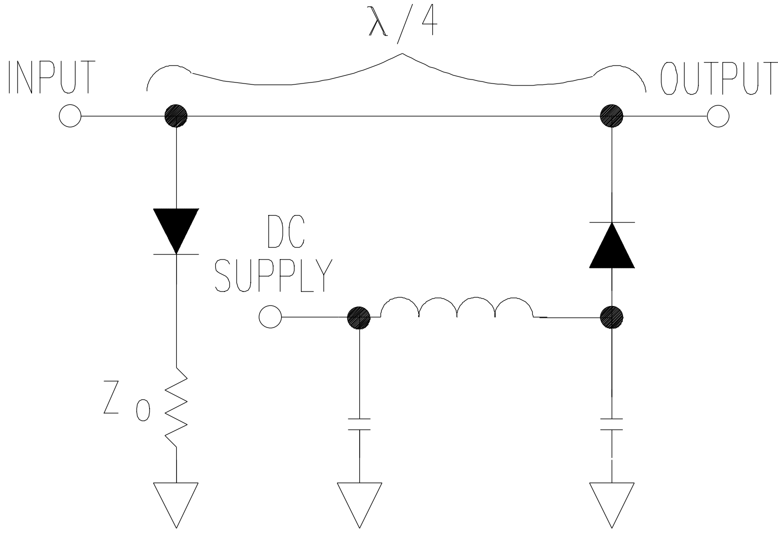


Figure 3.9 Quarter-Wave Matched Attenuator (Series Connected Diodes)

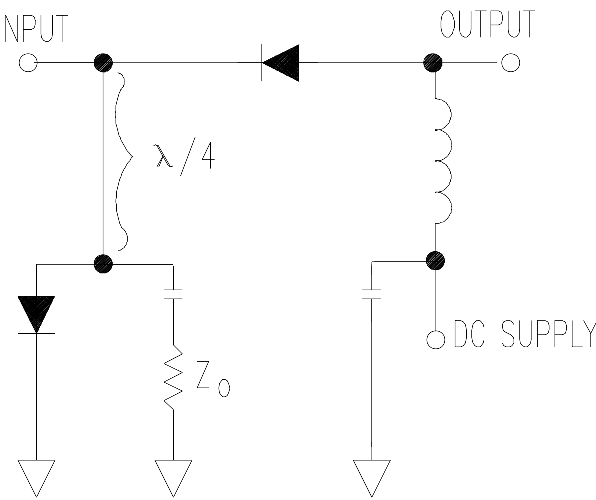


Figure 3.10 Quarter-Wave Matched Attenuator (Shunt Connected Diodes)

The following equations summarize the performance of these Quarter-Wave Attenuators:

Quarter-Wave Attenuator performance equations:

(Series Connected Diodes)	Shunt Connected Diodes
Attenuation = 20 log (1 + Z_o / R_s), dB	Attenuation = 20 log (1 + R_s / Z_o), dB

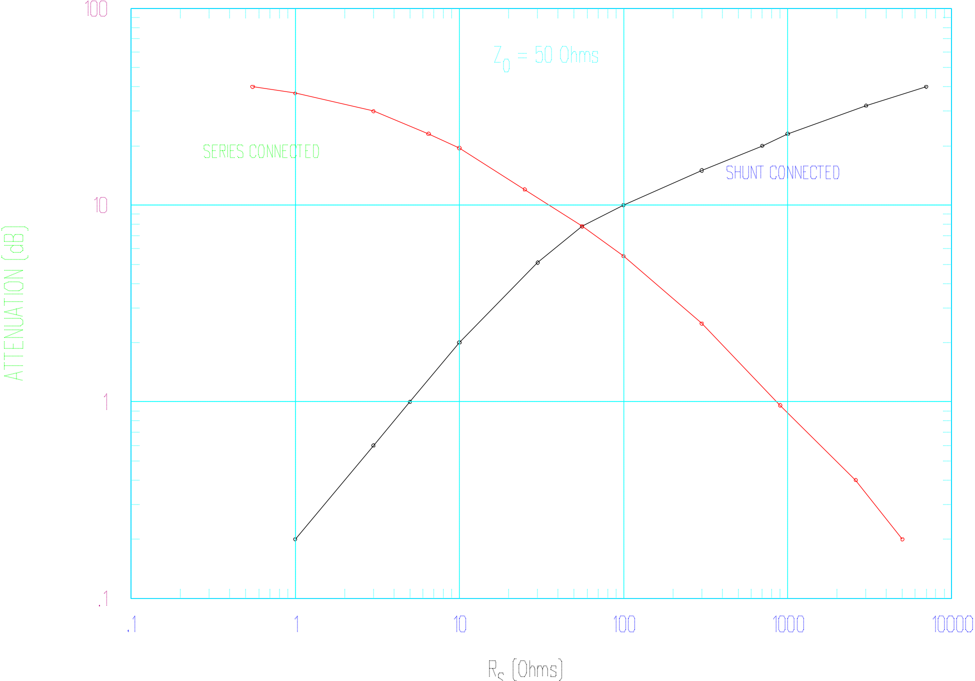


Figure 3.11 Attenuation of Quarter-Wave Attenuators

Quarter-Wavelength Attenuators are matched when both diodes are biased to the same resistance. This usually occurs since both diodes are connected in series to the d-c current supply, and so the same forward bias current flows through both diodes. The series connected configuration is preferable for higher values of attenuation and the shunt connected configuration is preferred for lower attenuation levels.

BRIDGED TEE & π ATTENUATORS

The fundamental attenuator design configurations, together with the design equations, were described in the initial section of this chapter. The most appropriate for matched broadband attenuator applications, especially those in the RF bands from HF Band through UHF Band, are the Bridged TEE & the π circuits. The upper cutoff frequency of these circuits often depends on the bias circuit isolation that can be obtained with practical circuit components. Feed through leakage at higher values of RF may also affect the highest value of attenuation that a particular design can achieve. The Bridged TEE circuit is shown in Figure 3.12 and the π circuit, in Figure 3.14.

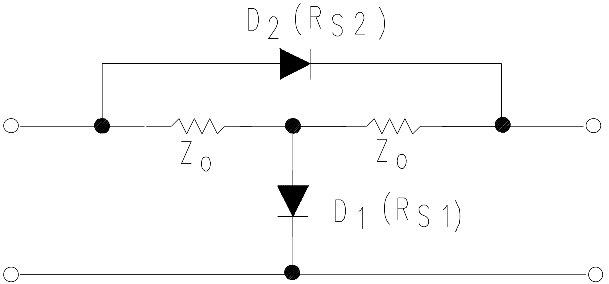


Figure 3.12 Bridged TEE Attenuator Circuit

The attenuation for the Bridged TEE circuit is obtained from the following equations[1,2]:

Attenuation = 20 log (1 + Z_O / R_{S1}), dB, and $Z_0^2 = R_{S1} \times R_{S2}$

These equations can be solved to show that the attenuation depends on the ratio of R_{S2} to R_{S1}, whereas the attenuator match conditions (Z_O) depends on the product of R_{S1} and R_{S2}.

The relationship between the forward biased resistance ($R_{S1,2}$) of the PIN diode and the forward bias current is also needed to determine the sets of values of diode driver currents that are needed to maintain impedance match for each value of attenuation desired. Figure 3.4 shows R_S vs I_f for the UM9552. The design procedure for the Bridged TEE circuit using UM9552's is available [2]. The attenuation curves for the Bridged TEE Attenuator are shown in Figure 3.13.

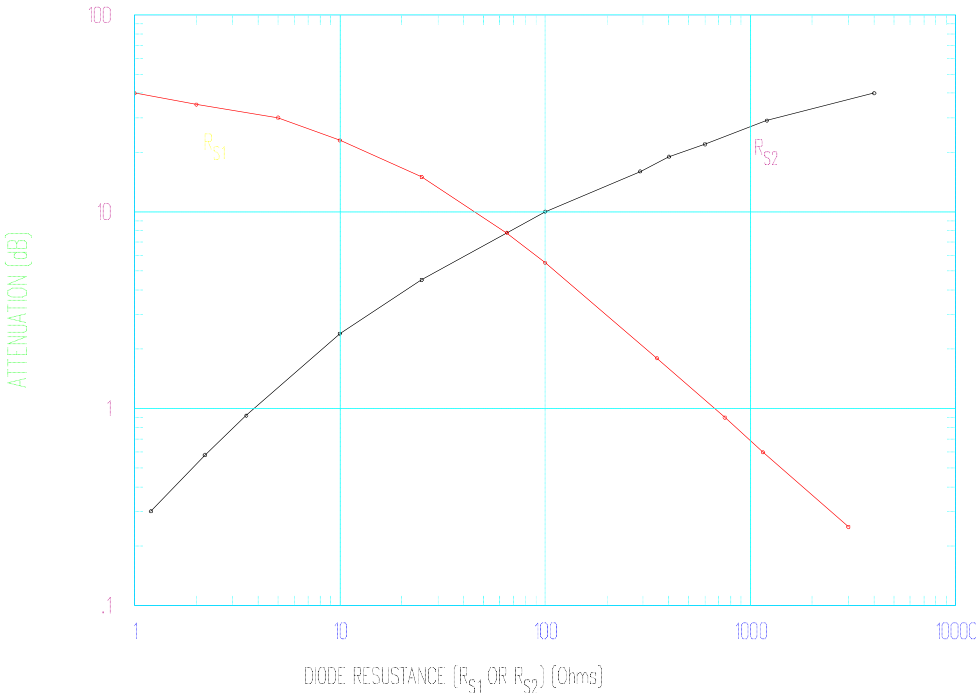


Figure 3.13 Attenuation of Bridged TEE Attenuators

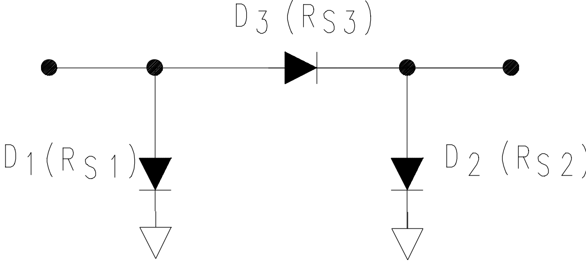


Figure 3.14 Ξ Attenuator Circuit

The 👌 attenuator circuit also has a set of equations that define the dependence of the attenuation state on the values of the three diode resistances[1].

$$\text{Attenuation} = 20 \log \{(R_{S1} + Z_0) / (R_{S1} - Z_0)\} \text{ dB}$$

$$\text{where: } R_{S1} = R_{S2} \text{ (Ohms) and } R_{S3} = 2 R_{S1} Z_0^{\ 2} / (R_{S1}^{\ 2} - Z_0^{\ 2}) \text{ (Ohms)}$$

The 👌 attenuator equations can be solved to obtain the performance curves shown in Figure 3.15. We see that the minimum value of Rs_1 and Rs_2 is 50 Ohms. $Rs_1 = Rs_2$ simply means that the attenuator is symmetrical, ie, the power source and load impedances are the same and equal to 50 Ohms.

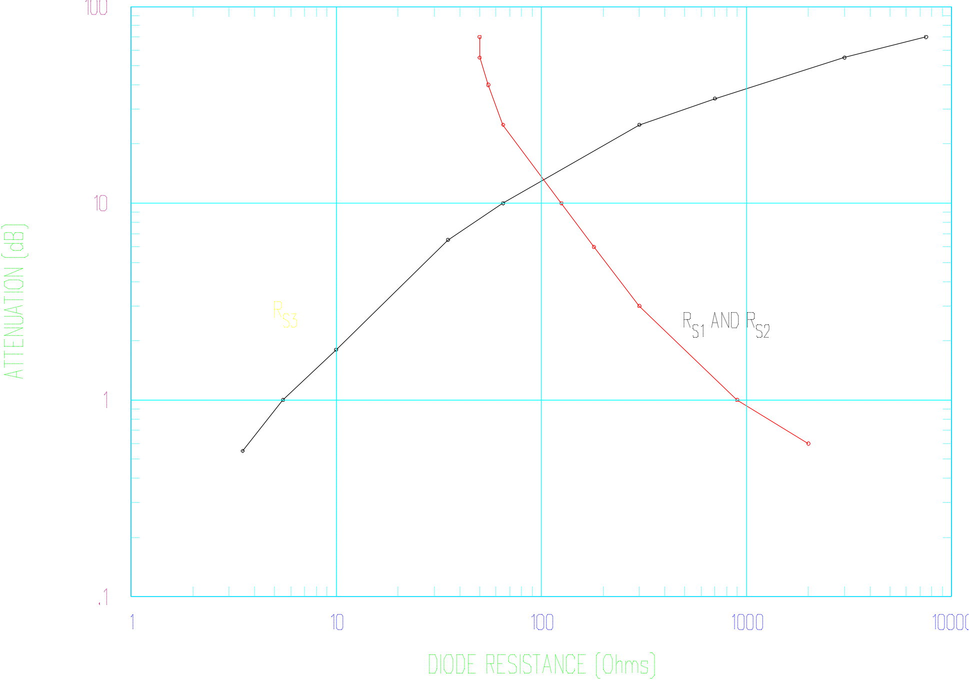


Figure 3.15 Attenuation of 👌 attenuators

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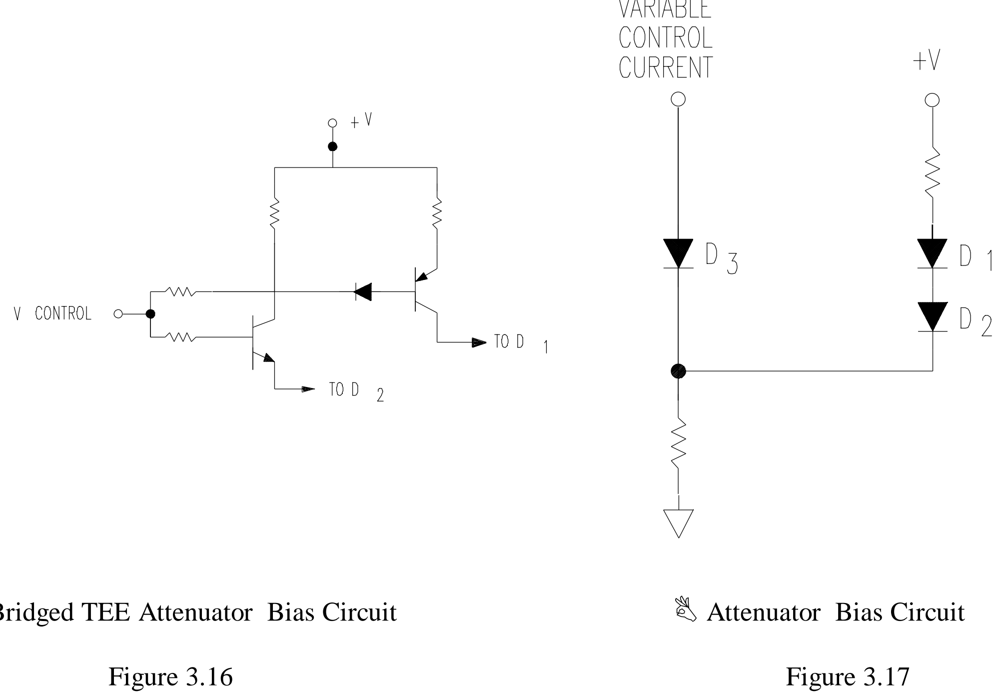
In both the Bridged TEE and the π Attenuator circuits, the PIN diodes are biased at two different values of resistance simultaneously and these must track so that the attenuator remains matched as different values over the dynamic range of the attenuator. Suggested voltage controlled bias circuit are shown in Figure 3.16 for the Bridged TEE attenuator and in Figure 3.17, for the π attenuator.



Bridged TEE Attenuator Bias Circuit

Figure 3.16

π Attenuator Bias Circuit

Figure 3.17

REFLECTIVE ATTENUATORS

In contrast to Matched PIN Diode Attenuator Circuits, Reflective Attenuators can be designed using single series or shunt PIN diode switch configurations (Chapter 2). In this application, the PIN diodes are only biased in the forward direction, utilizing the current control resistance characteristic of the PIN diode. Referring to Figure 3.4, the forward bias current may be continuously varied from high resistance to low resistance values. Attenuation is obtained by introducing impedance mis-match in the transmission line. This causes some of the power to be reflected back toward the power source. This is undesirable in many systems applications because it may cause frequency pulling and power instability. However , Reflective Attenuators are inexpensive to design and build. The attenuation values obtained using these reflective attenuators can be calculated from the following equations:

Series Connected PIN Diode Attenuator: Attenuation = $20 \log(1 + R_S / 2 Z_0)$, dB

Shunt Connected PIN Diode Attenuator: Attenuation = $20 \log(1 + Z_0 / 2 R_S)$, dB

These equations are plotted in Figure 3.18 for series and shunt attenuators with $Z_0 = 50$ Ohms. These equations and curves assume that the PIN Diode Impedance is purely resistive. Above the UHF Band, Capacitive and Inductive Reactances of the packaged PIN diode chip must be taken into account.

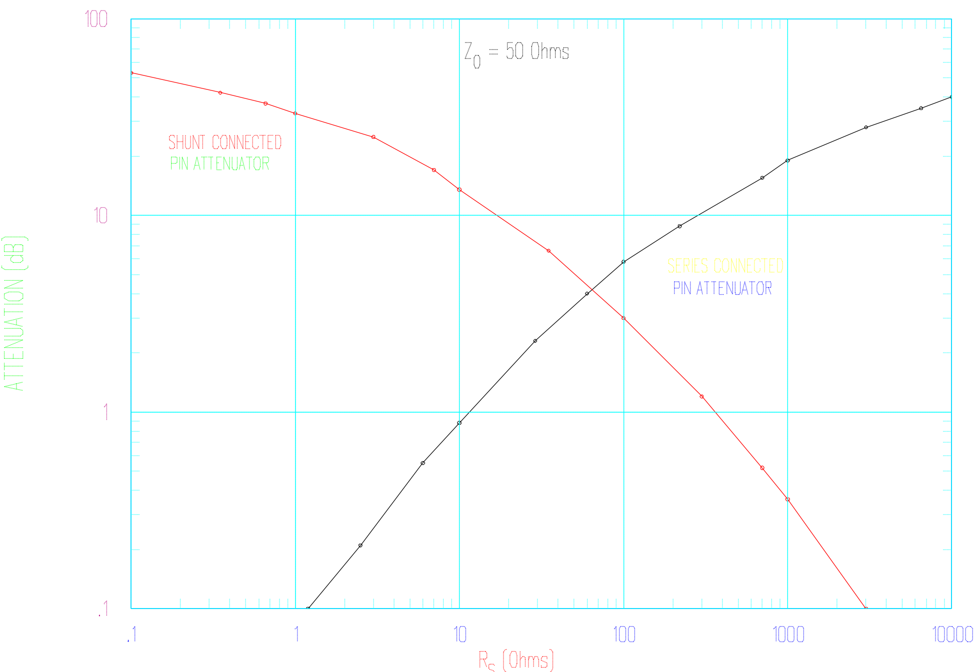


Figure 3.18 Attenuation Of Reflective Attenuators

DISTORTION IN PIN DIODE ATTENUATORS

Distortion is a particularly critical parameter in PIN diode attenuator circuits and is defined, described, and discussed in Appendix E and reference [3].

APPLICATION	**RECOMMENDED PIN DIODE TYPES**
High Power >1 W	UM2100, UM4000, UM4300, UM9552
AGC	UM4000, UM6000, UM7000
Low Frequency	UM2100, UM4000, UM4300, UM9552
Ultra Low Frequency	UM2100, UM9552

CHAPTER - 4

PIN DIODE RF MODULATORS

NOTES

PIN DIODE MODULATORS

INTRODUCTION

In Chapter 1, it was said that a Microwave PIN diode is a semiconductor device that operates as a variable resistor, whose value is defined by the d-c bias state or by a low frequency (compared to the RF Carrier Wave) variable bias.

In Chapter 2, the PIN Diode was described as a Switching Element, whose control current is switched ON & OFF to control the RF signal.

In Chapter 3, the PIN Diode was described as an Attenuating Element, whose control current was varied continuously (but perhaps also in discrete steps) to produce various levels of attenuation to the RF signal.

The main difference between the two applications is the manner in which the bias conditions are defined for the PIN diode circuit. In both applications, only one signal (or one band of signals) was present in the PIN circuit.

In Chapter 4, the PIN diode is described as a Modulator Element. Modulator applications are much more complex to analyze in that two discrete signal frequencies are present in the PIN diode simultaneously. These consist of the RF Carrier Wave (usually a single frequency in the RF or Microwave Bands) and a much slower varying, lower frequency signal (a sub-band of the d-c to 10 MHz range). The lower frequency signal current represents a relatively slowly varying "bias current" that modulates the I-region impedance that the PIN diode exhibits to the RF Carrier Wave current, causing the amplitude of the RF Carrier Wave to change.

A detailed analysis of a specific modulator design depends on the relative maximum amplitudes of the two signals, the location of the two signals in the frequency spectrum, and the waveform of the low frequency modulating signal [1]. The two modulator designs, described in this chapter, are the Continuous Amplitude Modulation and Pulsed Amplitude Modulation. They are readily implemented with PIN diodes. These modulator networks are assumed to be broadband with no restrictions on the impedance termination at various sideband frequencies. The reader is referred to the general literature for other design constraints.

The RF & Microwave modulation techniques to be discussed in Chapter 4 are distinct from the Digital Modulation Techniques that prepare the information signal for transmission through the RF Channel [2].

MODULATION - BASIC CONCEPTS

Modulation [1 , 3] is a process whereby certain characteristics of an RF Carrier Wave are varied or modified in accordance with a message or information signal which may be Analog or Digital in format. Modulation is also called Up-Conversion since the information signal is "up-converted" from the Signal Band (usually some segment of the d-c to 10 MHz band, depending on the waveform of the Signal)) to the RF Carrier Wave Band (usually in the RF or Microwave Bands) for efficient transmission through the RF Channel. Ordinarily, there is at least a 5 : 1 separation in frequency between the Signal Band and the RF Carrier Wave Band for ease in designing the RF Filters needed to provide isolation between the circuit components operating in a multi-band network.

RF & MICROWAVE AMPLITUDE MODULATION

If the RF Carrier Wave is Continuously Amplitude Modulated by an Analog Signal Source, the Modulated RF Wave is always present in the modulated output. This PIN Diode Modulator Circuit is actually a PIN Diode Attenuator circuit in which the PIN diode is "forward biased" by the signal wave while the RF Carrier Wave is also present in the PIN diode. The forward biased Resistance of the PIN diode is (relatively) slowly and continuously varied by the information signal waveform producing a Continuous Amplitude-Modulated RF wave, as shown in Figure 4.1. Note that the RF carrier frequency retains its sinusoidal wave-form while the amplitude envelope varies at the modulation frequency. The RF carrier wave has peak amplitude "A", while the modulation wave has peak amplitude "B". The modulation index "K" is given by: $K = B / A$, and is a measure of the depth of modulation. If $K = 1$, the RF Carrier Wave is said to be 100 % modulated.

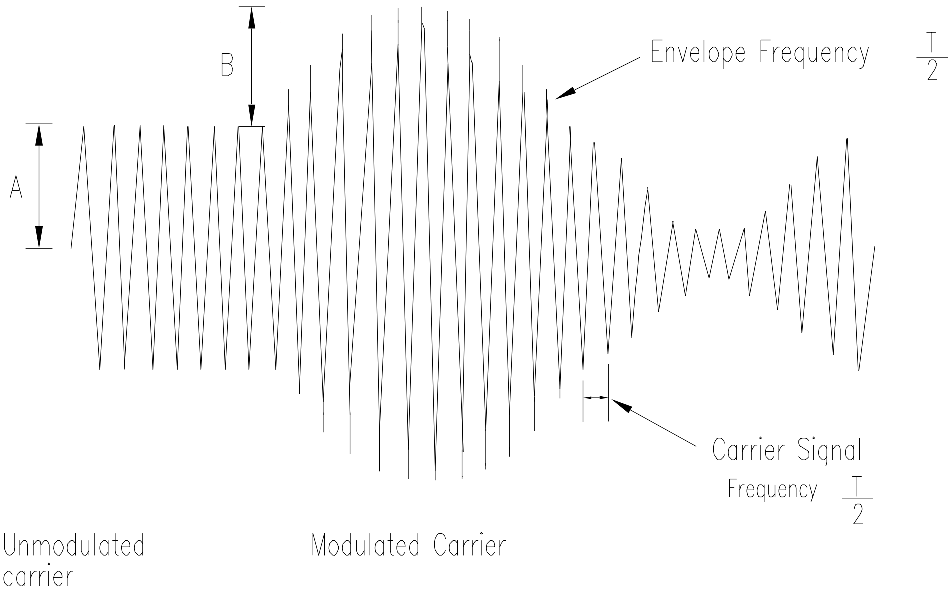


Figure 4.1 Continuous Amplitude - Modulated Wave

The frequency spectrum of the Continuous Amplitude Modulated wave is shown in Figure 4.2, which shows three distinct frequencies: the RF Carrier (F_C), its lower sideband ($F_C - F_S$), and its upper sideband ($F_C = F_S$). The sidebands are separated from the carrier frequency by the magnitude of the frequency of the modulation signal (Fs). Figure 4.2 is the frequency domain representation of the waveform in Figure 4.1 because only the amplitude of each sinusoidal wave and its appropriate location in the frequency spectrum are shown. Both sidebands exist because the modulation network is broadband and they are therefore terminated in the Characteristic Impedance Zo = 50 Ohms.

Balanced Amplitude-Modulation can be used to suppress the Carrier Wave. This can be achieved by using two hybrids, one at each of the Carrier Frequency and the Modulation Signal Frequency, and two PIN diodes, in a balanced network [4]. One of the sidebands can then be filtered to obtain a Single Sideband Output Waveform (SSB-AM), which greatly increases transmitter efficiency.

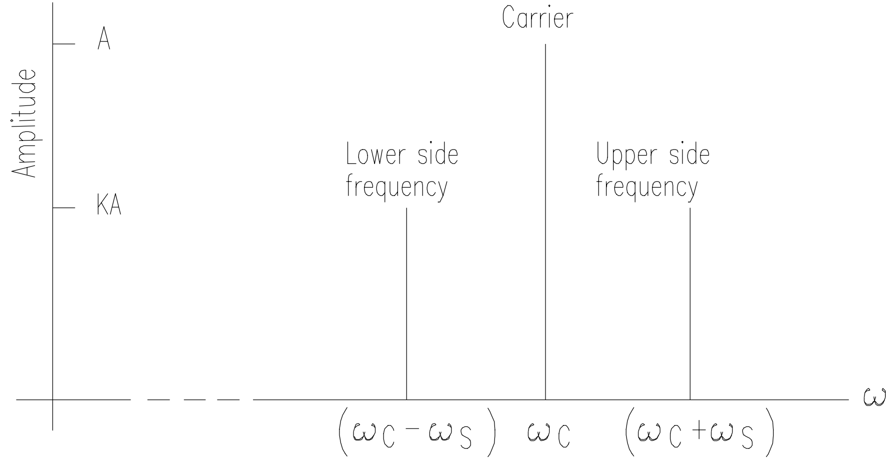


Figure 4.2 Frequency Spectrum of the Continuous Amplitude-Modulated Wave

MICROWAVE POWER MODULATORS

PIN diodes are the preferred active elements for Microwave Power Modulators. The switching speed must be fast enough for the PIN diode to respond to the modulating signal, without introducing non-linear modulation effects. The PIN diode's minority carrier lifetime should be long enough to provide a low level of RF Intermodulation Distortion.

PIN diode Modulator applications use circuit configurations that are similar to PIN diode attenuator circuits. Since the modulation signal is fed into the d-c bias port, the bias circuitry must be sufficiently broadband that the modulation signal is not distorted. Isolation between the modulation insertion port and the RF Carrier input port should be at least 50 to 60 dB. The RF circuitry should be sufficiently broadband to terminate the RF carrier and both sidebands in 50 Ohms.

For pulsed and continuous (linear analog) modulators, the quadrature hybrid circuit shown in Figure 4.3 satisfies the bandwidth and Isolation requirements. Such quadrature hybrids are available from about 10 MHz to 4 GHz in compact form.

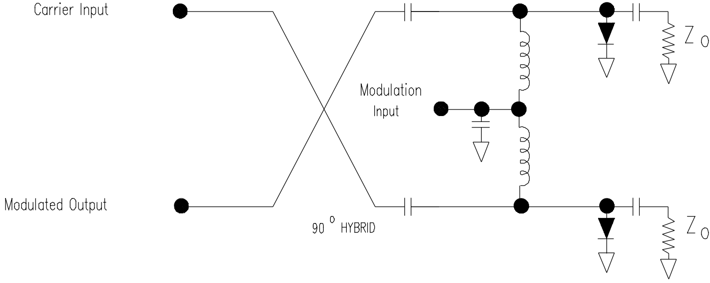


Figure 4.3 Quadrature Hybrid Matched Modulator

Dynamic Ranges of up to 80 dB are achievable in certain Continuous AM designs since the PIN Diode's lifetime characteristic improves the modulation linearity over the AM signal amplitude range. The unique characteristic of large signal PIN Diode Continuous AM modulators is that the PIN Diode device parameters can be adjusted so that modulation efficiency and linearity are optimized.

If the RF Carrier Wave is pulse modulated, no RF signal output is present between pulses. This PIN Diode Pulse Modulator is a PIN Diode Switch circuit that is rapidly biased ON (the low Insertion Loss state) and OFF (the high Isolation State) according to the alternating polarities of the pulsed information signal. In the pulse modulation mode, the RF Carrier Wave is not transmitted during the OFF state. Usually, the output signal of the Pulse Information Source is sufficiently weak that it must be amplified by a modulation driver (Amplifier) circuit so that the PIN Diode can be driven ON and OFF without distortion of the pulsed RF output waveform.

DEMODULATION

Demodulation is described here to complete the view of the Modulator as an integral part of the RF Channel [2]. Baseband Signal Processing prepares the Modulation Signal for Up-Conversion to the RF Channel's Carrier Band. Ultimately, the Modulated RF Carrier is received and Demodulated for additional Baseband Signal Processing. The success with which the original Modulation Waveform is retrieved by this process depends on the linearity (both amplitude & phase) of the modulation process and on the free space characteristics of the RF Channel.

Demodulation or Detection is the inverse process of Modulation. At the Receiver, the Amplitude Modulated Waveform is inputted to the Demodulator and the Modulation Signal is Down-Converted to baseband (d-c to 10 MHz). Ideally, the Demodulated Wave should be a faithful replica of the original Modulation Wave that inputted the Transmitter's Modulation Circuit. A re-labeled version of Figure 4.3 is shown below to indicate that basically, a Demodulator circuit is a Modulator circuit with the inputs and output reversed (Figure 4.4).

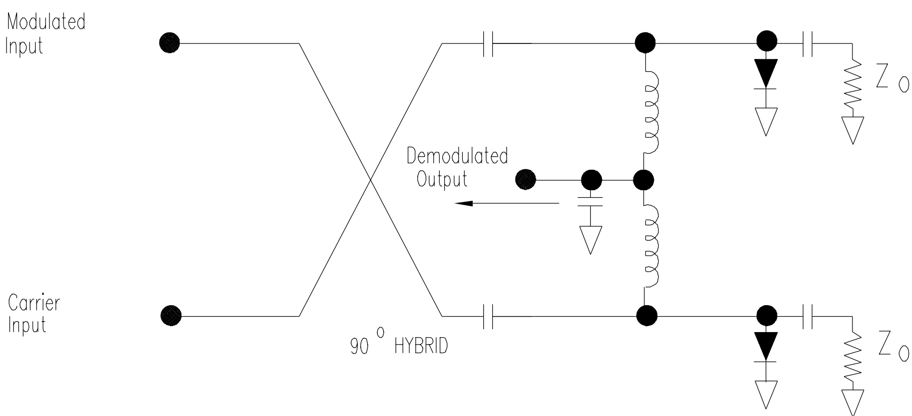


Figure 4.4 Quadrature Hybrid Marched Demodulator

APPLICATION	RECOMMENDED PIN DIODE TYPES
High Power >1 W	UM2100, UM4000, UM4300, UM9552
AGC	UM4000, UM6000, UM7000
Low Frequency	UM2100, UM4000, UM4300, UM9552
Ultra Low Frequency	UM2100, UM9552

CHAPTER - 5

PIN DIODE RF PHASE SHIFTERS

NOTES

MICROWAVE PHASE SHIFTERS

INTRODUCTION

In Chapter Four it was mentioned that “all three characteristics of the RF Carrier Wave: Amplitude, Frequency, and Phase, can be modulated”. Since only Amplitude Modulation of PIN diodes was to be described, a more complete discussion of the RF Carrier Wave form was not required. In the present chapter on Phase Shifters, the phase of the RF Sinusoidal Wave form must be described so that the concept of Phase Shift through an RF or Microwave Circuit has more than an intuitive or heuristic significance to the reader.

TIME VARYING SINUSOIDAL WAVEFORMS

Input signals to RF and Microwave Circuits are usually described physically in terms of single sinusoidal wave forms, or for more complex wave forms, may be described as a composite or summation of a number of sinusoidal wave forms.

In its most frequent form for electronic-circuit applications, the sinusoidal voltage is a function of time of the general form:

$$v(t) = V_o \sin\{2\pi f t + \phi\} \qquad (5.1)$$

where: V_o is the amplitude, f is the frequency, and ϕ is the phase angle with respect to some arbitrary phase angle reference, as shown in Figure 5.1.

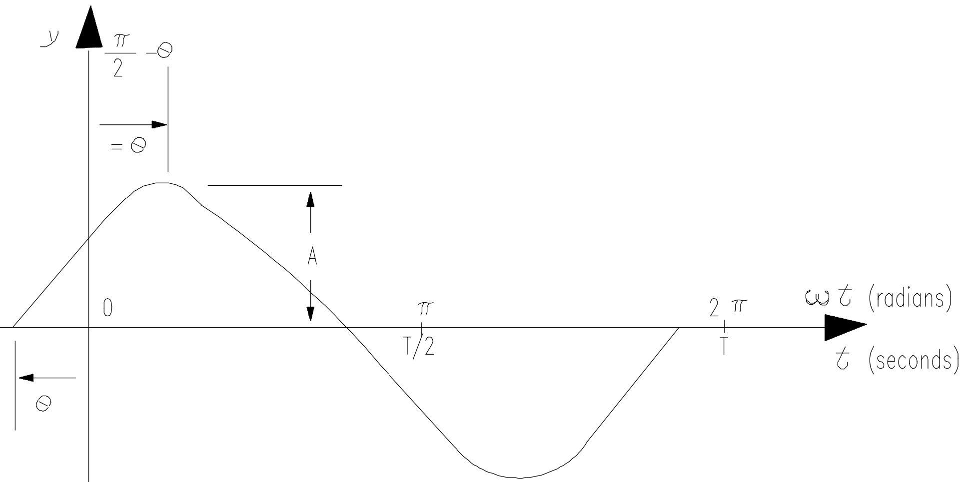


Figure 5.1 The Sinusoidal Function

In Figure 5.1, A = V_o, the frequency (f) is related to the inverse of the period (T) of the wave, and ϕ is the phase angle of the wave with respect to the phase reference. The phase reference can be chosen arbitrarily to define the initial value of the Phase Angle (ϕ) at the input to the circuit.

PHASE ANGLE

Phase Angle [1] is the measure of the progression of a periodic wave in time or space from a chosen instant or position. The Phase of a periodic wave, v(t), the frequency of which corresponds to period T, is the fractional part (t / T) of the period T through which t has advanced relative to an arbitrary origin or phase reference.

Circuit components have some physical size and the RF Voltage wave form passes through these components with a certain finite velocity and time of transit. The total phase angle, ϕ, will increase or "shift" by an amount related to this transit time through these components. This is the residual phase shift associated with a specific component and a specific wave form frequency. Small circuit-components operating at relatively low frequencies (i.e., HF Band) will exhibit very small residual ϕ. In the Microwave Bands, however, these same components will exhibit significantly larger residual phase shifts.

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MICROWAVE PIN DIODE PHASE SHIFTERS

Microwave Phase Shifters utilize these properties of wave propagation through circuit component structures. They are designed to produce the phase shift effects required for the operation of certain classes of Antenna Systems for RADAR and Communications applications. Reference [2] treats the subject of Semiconductor Phase Shifters for Antenna arrays in great detail. The treatment here deals specifically with PIN diode Phase Shifter circuit design.

The PIN diode may be considered as a lumped variable-impedance microwave circuit element. Microwave or RF signals passing through the PIN diodes experience some finite phase shift. A lumped element representation permits the design of compact phase shifter circuits at the RF frequency bands. However, in the higher microwave frequency bands, losses increase and power handling capability decreases. Figure 5.2 shows an example of a PIN diode phase shifter with driver-amplifier connections. Note that there is one driver circuit for each phase shifter bit.

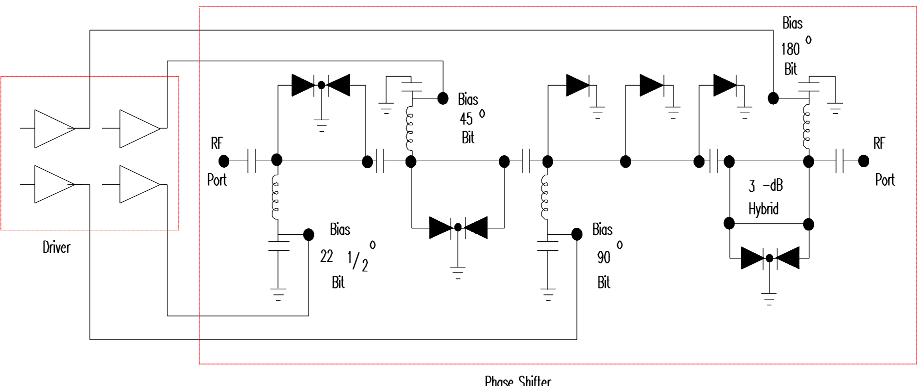

Figure 5.2 A PIN Diode Phase Shifter With Driver-Amplifier Connections

PIN diodes are utilized as series or shunt connected switches in phase shifter designs. The switched elements are either lengths of transmission line or reactive elements. The criteria for choosing PIN diodes for use in Phase Shifters are similar to those used for other switching applications.

In addition, there is the possibility of introducing phase distortion, particularly at high RF power levels or low reverse bias voltages. Microsemi PIN diodes are characterized by thick I-regions and long carrier lifetime and these are the characteristics that result in low phase distortion.

REAL TIME DELAY PHASE SHIFTER

The major application of electronically controllable, rapid-acting microwave phase shifters is in antenna systems for Phased Array Radar. In these systems, a phase shifter is placed in series with each radiator of an array of antennas. The radiating-phase-front direction is controlled by varying the time delay from the source of a common signal to each radiating element of the array. The steering of the array is independent of the radiation frequency.

The circuit used to produce this time delay is the real-time-delay phase shifter or the nondispersive phase shifter. It is essentially a switched section of transmission line. The time delay through the circuit must be independent of the phase-shift state. The phase shifter is nondispersive because its overall phase shift is not directly proportional to frequency.

The Phase-Frequency Characteristics of dispersive and nondispersive phase shifters are shown in Figure 5.3 and are discussed in the next section.

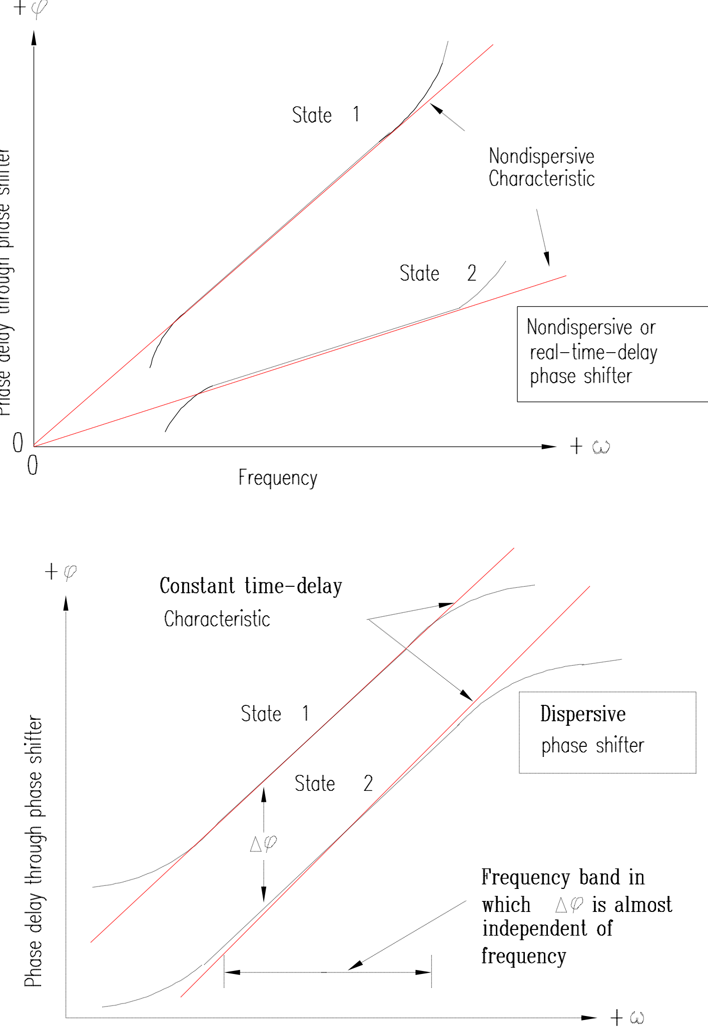

Figure 5.3 Phase-frequency Characteristics of Phase Shifters

TYPES OF REAL TIME - DELAY PHASE SHIFTERS

The simplest design for a real-time-delay phase shifter uses a circulator or a 3 dB coupler to provide matched input and output for the switched elements, as shown in Figure 5.4.

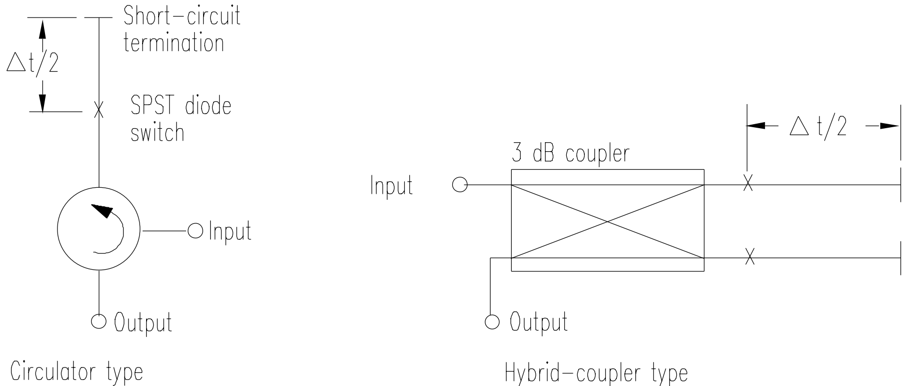

Figure 5.4 Elementary Real-Time-Delay Phase Shifters

In both phase shifters, The SPST diode switches are used as either short circuits or open circuits over a finite bandwidth. Dependence on a section of line for the time delay causes a bandwidth limitation on this class of phase shifters, even if ideal couplers or circulators and identical SPST switches are used. The narrow band performance of the delay lines produces the dispersion that results in nonconstant time delay. For narrow band systems, this design is a viable approach.

POWER LIMITATIONS

The choice of diodes limits the maximum power that can be carried by these phase shifters. PIN diodes have sufficient stored charge to prevent them from producing wave form distortion. Their capability is limited by thermal dissipation of absorbed RF power and RF voltage breakdown. To choose a PIN diode for a specific design, we specify a maximum allowable RF current or voltage applicable to the forward- or reverse-biased diode. Using these two values, we can calculate the maximum power that a phase shifter design may carry. For the hybrid-coupled phase shifter of Figure 5.4, we calculate from the structure of the SPST diode switch that each switch may carry I_{max} Amperes in the forward biased state and V_{max} Volts in the reverse biased state. We then state that

$$P_{max} = [\, I_{max}^2 \;\text{x}\; Z_o \,] \;/\; 8 \qquad (5.2)$$

where Z_o is the characteristic impedance of the transmission line in which the SPST is built (which must be matched to the generator) and P_{max} is the maximum incident power that may be carried by this line when the switch is in the maximum current state.

DISPERSIVE PHASE SHIFTERS

The distinguishing feature of this type of phase shifter is the production of a frequency-independent phase change. Figure 5.5 illustrates a hybrid-coupler type of dispersive phase shifter, which uses PIN diodes as the shunt

Susceptance across the transmission lines. An entire Phase Shifter would contain a number of these “unit Cells” that each contribute a specific Phase Shift Bit, of value d ϕ. For example, one “Cell” could be designed to be a 22.5 degree Bit, another “Cell”, a 45 degree Bit and so on.

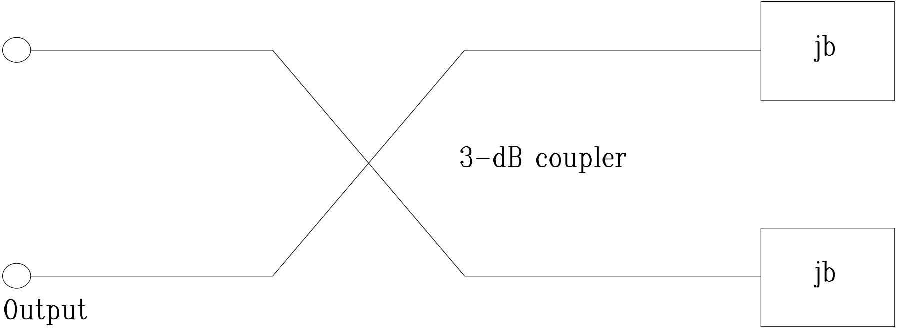

Figure 5.5 Hybrid-coupler Phase Shifter “Cell”

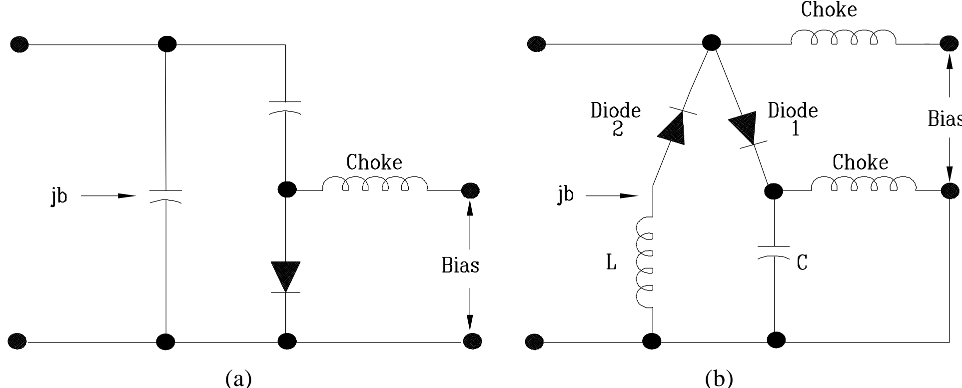

Figure 5.6 Two Possible Loading Elements of the 3 dB Hybrid-coupler in Figure 5.5.

One PIN diode alone is a switchable Capacitance terminating the transmission line. Forward and reverse bias values are chosen such that, in the forward bias case, C+ is $C_o + \Delta C$, and when reverse biased, C- is $C_o - \Delta C$, where C_o and ΔC determine the Phase Shifter Bit value desired [2,3].

The total Bit phase change is the difference between the forward and reverse biased phase changes:

$$\Delta \phi = (\phi+) - (\phi-) \qquad (5.3)$$

which we desire to be frequency - independent. Thus we impose the condition:

$$d(\phi+)/df = d(\phi-)/df \qquad (5.4)$$

which is the condition for equal time delay in the two bias stated. This is the characteristic that distinguishes the dispersive from the nondispersive phase shifter.

The dispersive phase shifter, using one PIN diode per branch, is shown in Figure 5.6 (a). It is suitable for values of $\Delta\phi$ up to approximately 120 degrees. A dispersive phase shifter, using two PIN diodes per branch, is shown in Figure 5.6 (b). This configuration can produce phase changes in excess of 180 degrees.

A GENERAL DISCUSSION OF PIN DIODE PHASE SHIFTER CIRCUITS

A schematic diagram of a 4-bit phase shifter[2] which gives 16 steps in increments of 22.5 degrees, is shown in Figure 5.7.

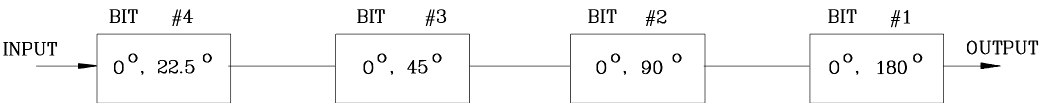

Figure 5.7 Schematic of Transmission 4-bit Phase Shifter

Any of the phase bits may take on three basic circuit topologies.

Switched -line phase bit (Figure 5.8)

Hybrid-coupled phase bit (Figure 5.9)

Periodically loaded-line phase bit (Figure 5.10)

The choice of circuit topology of the phase bit depends on factors such as the number of PIN diodes required, power level, fabrication ease, and cost. Minimum insertion loss condition implies equal loss in each bit state.

SWITCHED-LINE BIT PHASE SHIFTER

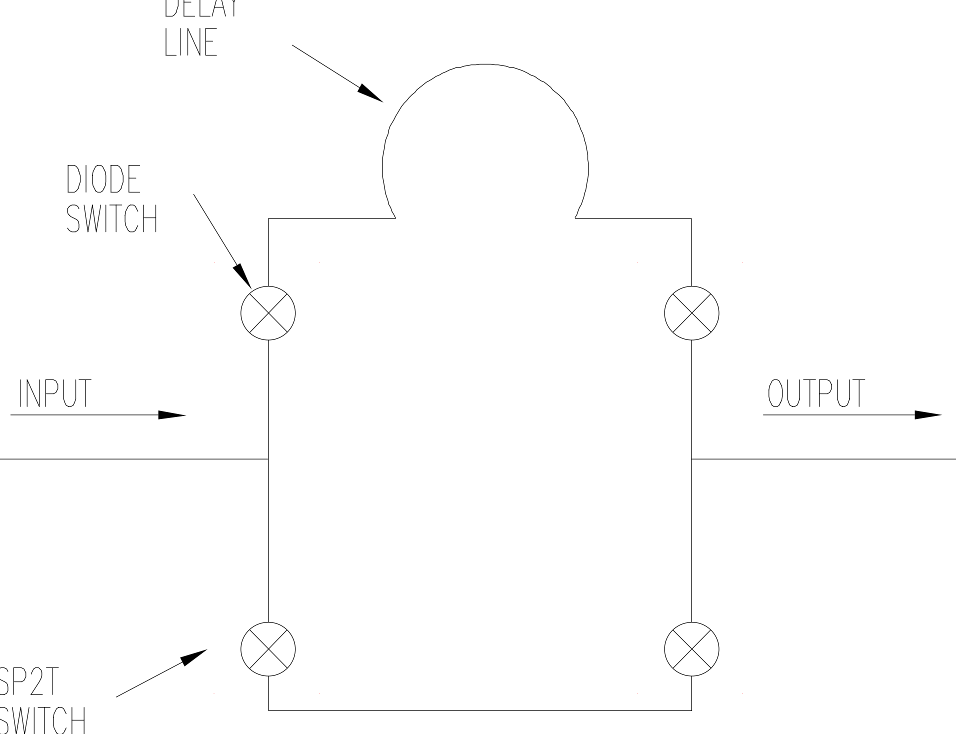

Figure 5.8 Switched-Line Bit Phase Shifter

The switched-line phase shifter is shown in Figure 5.8. This circuit consists of two SPST switches and two lengths of transmission line for each bit. Four PIN diodes are required as a minimum. The transmission line lengths are

arbitrary, so the circuit can be used for bit phase values from 0 to 360 degrees. It can also be used as a time delay network. The Isolation per switch in the OFF branch must be greater than 20 dB to avoid phase errors.

Minimum loss occurs in each SPDT switch when the energy absorbed in the OFF branch is equal to the loss of the ON or pass branch. If the diode resistance is equal in both states, the minimum loss for the bit occurs when the Characteristic Impedance of the T-junction is equal to the average Capacitive Reactance of the PIN diodes. Analysis [2] for this type of bit shows that if the circuit is optimized for minimum loss by properly choosing the Z_o of the T-junction, the peak-power capacity is twice that of any other type of phase bit. If line loss is neglected, the Insertion Loss is the same for all bit sizes.

HYBRID - COUPLED- BIT PHASE SHIFTER

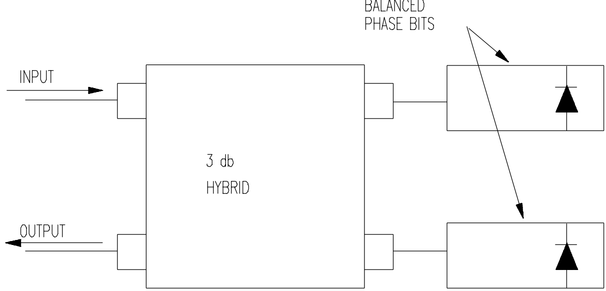

Figure 5.9 Hybrid-Coupled-Bit Phase Shifter

The Hybrid-coupled-bit phase shifter, Figure 5.9, has a 3 dB hybrid junction with balanced phase bits attached to the coupled branches. Analysis of this type [2] shows that the voltage (VBR) required of the PIN diodes depends on the bit size in which the diode is used, if it is assumed that equal power is incident on all cascaded bits. VBR is highest for the 180 degree bit and is reduced by $\sqrt{\sin(☼/2)}$ for the smaller bits. Similarly, the insertion loss is also a function of bit size. If the loss of the 180 degree bit is L_o, then the loss of the smaller bits is $L_o \sin(☼/2)$. The hybrid-coupled-bit phase shifter has the least loss of the three types and uses only two PIN diodes per bit.

An analysis [2] for the hybrid-coupled 180° phase bit shows that the peak-power capability and optimum impedance level for equal loss in both switch states are given by:

$$P = V^2 / 4Z_o \qquad (5.5)$$

where $Z_o = X_c \sqrt{R_F / R_r}$

Equation (5.5) shows that to obtain high-power handling capacity, the PIN diodes must have high VBR, relatively high Capacitance, and operate in transmission lines with low impedance levels.

The breakdown voltage required of the PIN diodes depends on the bit size in which the diode is used. This requirement is the highest for the 180° bit and is reduced by the factor, $\sqrt{\sin(\psi/2)}$, for the smaller bits. If the dissipation loss for the 180^0 bit is L_O, the loss of the smaller bits is $L_O \sin(\psi/2)$. The Hybrid - Coupled Bit Phase Shifter has the least loss of the three types being considered here, and uses the least number of diodes.

Shorter bits can be obtained from the 180 degree bit by using the transformed-switch technique[2], which consists of placing an impedance transformer one-eighth wavelength before the input port of the 180 degree bit. The impedance transformation ratio of the transformer is varied to produce the various phase-bit sizes.

LOADED-LINE PHASE SHIFTER

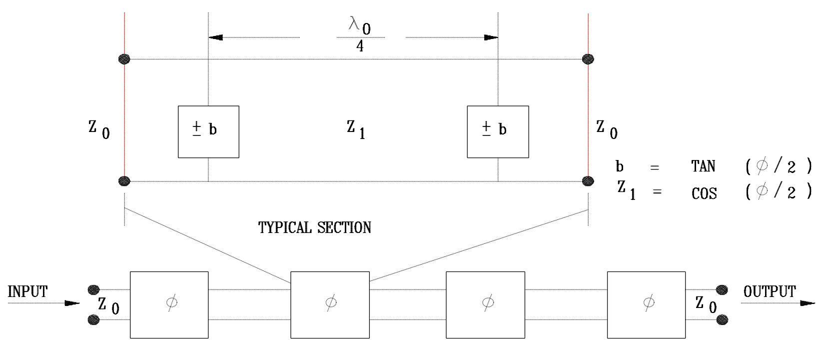

Figure 5.10 Loaded-Line Phase Shifter

The loaded-line circuit uses switched loading Susceptances spaced a quarter-wavelength apart along a transmission line (Figure 5.10). Adjacent loading Susceptances are equal and are switched into either a Capacitive or an Inductive state. Impedance-matched transmission for both states is maintained by choosing the impedance level of the transmission-line section between the PIN diodes

The magnitude of the loading Susceptance, B, and the characteristic Impedance of the connecting transmission line, Z_1, are related to the required phase shift,ψ, per section and the Phase Shifter optimum impedance Z_o by the equations

$$Z_1 = Z_o \cos (\psi / 2) \qquad (5.6)$$

$$B = \tan (\psi / 2) \qquad (5.7)$$

The PIN diodes can be either directly mounted or stub mounted across the transmission line. Average power handling in this configuration is limited by the VBR of the PIN diode and the practical level to which the characteristic impedance can be reduced.

The peak power capacity is a function of the VBR and the phase step size.

For equal insertion loss in each state of the phase step, the peak power capacity for the loaded line phase shifter is one half of the hybrid-coupled-bit design.

The insertion loss of n small phase steps cascaded to achieve 180 degree phase shift is $\pi / 2$ times the loss of the 180 degree hybrid-coupled-bit circuit.

To achieve high power capability, the loaded-line phase shifter uses many PIN diodes and small phase increments.

CONCLUSIONS:

In comparison with the loaded line phase shifter, the hybrid coupled design can handle up to twice the average or peak power when using the same PIN diodes. In both hybrid and loaded line designs, the power handling capability of the maximum bit size is related to the maximum RF current and the peak RF Voltage that the PIN diodes can withstand. The characteristic impedance of the bit circuits is a variable that can be used to adjust the current and voltage stress to be within the device ratings. This means that Z_o is reduced below 50 Ohms to reduce the voltage stress in favor of higher RF currents. The maximum current rating of Microsemi PIN diodes depends on the power dissipation rating while the maximum voltage stress is dependent on I-region thickness.

DEVICE	HIGH VOLTAGE >2000 V	HIGH AVERAGE POWER >100 W	HIGH PEAK POWER >10 KW	HIGH POWER CW DUPLEXERS >100 W	ANTENNA SWITCHING >100 W	HIGH FREQUENCY > 1GHz	LOW FREQUENCY <10 MHz	ULTRA LOW FREQUENCY <1 MHz	L C R <
HUM2020	X	X	X	X	X		X		
UM2100		X	X	X	X		X	X	
UM2300		X	X	X	X		X	X	
UM4000		X	X	X	X		X		
HUM4020	X	X	X	X	X		X		
UM4300		X	X	X	X		X		
UM7000		X	X	X					
UM7100		X	X	X					
UM7200									
UM7300		X	X	X	X		X		
UM7500		X	X	X					
UM9401		X	X	X					
UM9415					X				
UMM5050	X		X	X	X				
UPP9401		X		X					
UPP1004		X		X					

CHAPTER - 6

PIN DIODE CONTROL CIRCUITS FOR WIRELESS COMMUNICATIONS SYSTEMS

NOTES

PIN DIODE CONTROL CIRCUITS FOR WIRELESS COMMUNICATIONS SYSTEMS

INTRODUCTION

Chapter 6 discusses PIN Control Circuits that are appropriate for Wireless Communications Systems, from HF Band (2-30 MHz) through 2.4 GHz. Many of the control circuits discussed in Chapter 2 (PIN Diode RF Switches) & Chapter 3 (PIN Diode RF Attenuators) are suitable for specific wireless system applications. Some of the most important circuit applications, using PIN diodes, are discussed below.

PIN DIODE ANTENNA TRANSMIT/RECEIVE SWITCHES

Transmit / Receive Antenna Switches are commonly used to connect the Transceiver's Antenna to either the Transmitter port or the Receiver port. Physically, this circuit is a Single Pole - Double Throw Switch (SPDT) that in either position, must have very low loss in the "ON" State (less than 0.5 dB) and high Isolation (typically, 30 dB) in the "off" State. Several T / R Switch circuits appeared in Chapter 2 and are repeated here as Figures 6.1 & 6.2.

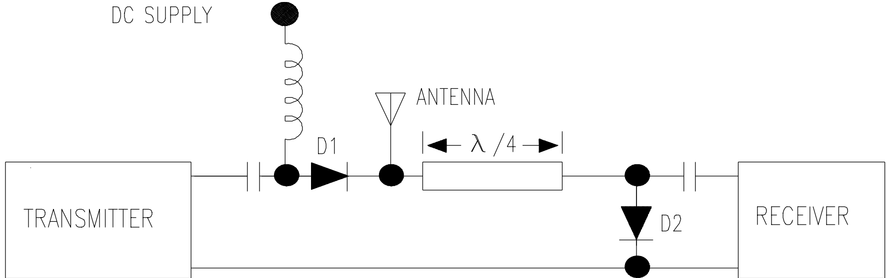

Figure 6.1 QUARTER WAVE ANTENNA SWITCH

Figure 6.1 is a narrow band SPST switch configured to switch the Antenna Port between either the Transmitter or Receiver Ports. The Quarter-wavelength line is the narrow band limitation, but for systems that have been allocated only 5% to 10% signal bandwidth, this is a very practical solution. When both D1 & D2 are forward biased, the Transmitter is connect to the Antenna, and the Receiver is protected by the isolation network of D2 terminating the quarter wavelength line. When D1 & D2 are reversed biased, the Transmitter Port is isolated by D1, and the quarter wavelength line and the Receiver Port are connected to the Antenna. The biasing scheme is very simple, requiring only one RF Choke Coil and a few d-c Blocking Capacitors. The quarter wavelength line can be simulated by a low-pass LC network to conserve board space.

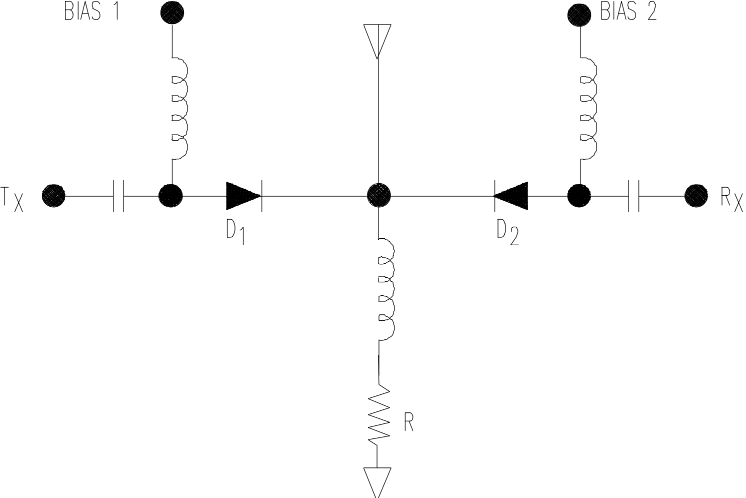

Figure 6.2 BROADBAND ANTENNA SWITCH

Figure 6.2 is a broad band SPDT switch, but the biasing scheme is more complex. It requires two bias tee's and a d-c return coil because D1 & D2 are alternately biased forward or reverse. For example, when the Transmitter is "ON" (and the Receiver "OFF"), D1 is forward biased and D2 is reversed biased. The Transmit / Receive Isolation State now depends solely on the Reverse Bias Capacitance of D2. The two bias conditions are exchanged when the Receiver is "ON". Isolation can be increased by using two PIN diodes in series instead of one for "D2".

The MICROSEMI UPP1001 Powermite® series of low distortion surface mount packaged PIN diodes were designed for low distortion Wireless Communications handset applications where battery power management is an extremely critical issue with circuit designers. They are excellent for handset, mobile, or base station applications where very low Insertion Loss (I_L) is required. The I_L of the UPP1001, mounted in a 50 Ohm network, with 10 mA forward bias current is less than 0.1 dB. The UPP1000's have a nominal Capacitance (Ct) of 1.25 pF, which is adequate for 5% to 10% bandwidth applications (similar to Figure 6.1) without inductive tuning. To use this series in broader applications (similar to Figure 6.2), either two devices can be used in series or a single device can be placed in one branch of a tuned switch design.

PIN DIODE ANTENNA DUPLEXING SWITCH

DUPLEX (communications) pertains to a simultaneous two-way independent transmission in both directions [1]. HALF-DUPLEX (communications) pertains to an alternate, one way at a time, independent transmission. In this sense, the Duplexing Switch, shown both in block diagram form in Figure 6.3 (a) and schematic form in Figure 6.3 (b), is a half-duplexing switch.

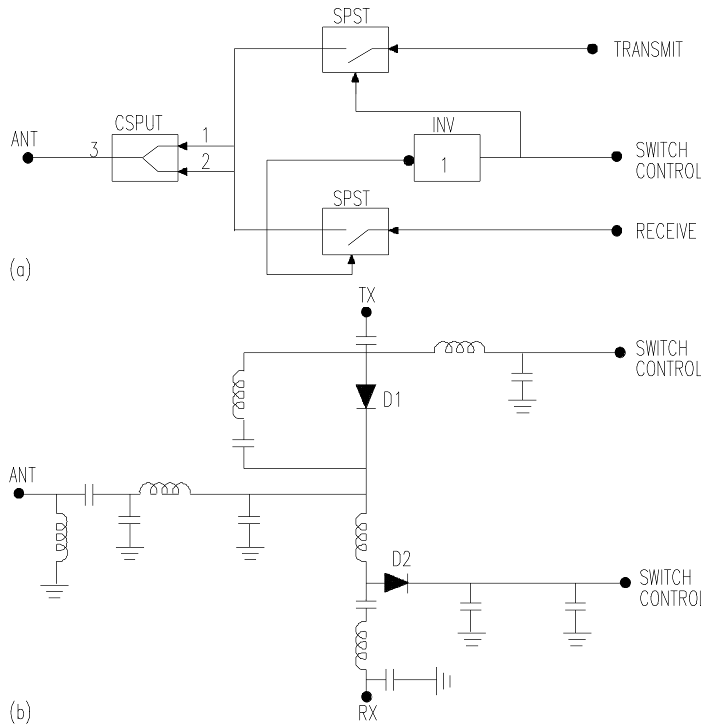

Figure 6.3 Duplexing Switch For Cellular Telephone: (a) Block diagram & (b) Schematic

The block diagram emphasizes the basic function blocks of the duplexing switch. The receive or transmit signals connect to the Antenna port via a power combiner or Y-Junction, which provides a matched, low loss connection to

either signal path. The individual switches are SPST switches, as discussed in Chapter 2. The PIN diode shunts the resonant circuit that determines the pass band of either the receive or the transmit switch. A single dual polarity switch control suffices for both switches, since a polarity inverter provides the opposite polarity for the receive switch.

The receive channel is 40 MHz to 60 MHz below the transmit channel, depending on the specific system application. If Figure 6.3 represents a GSM cellular phone, the receive switching circuit is tuned to 890 to 915 MHz and the transmit switching circuit is tuned to 935 to 960 MHz band. The antenna channel filter network passes both the transmit and receive bands.

PIN DIODE ANTENNA DIVERSITY & T/R SWITCH FOR CELLULAR TELEPHONES

This antenna switch for cellular telephones provides the antenna T/R function as well as the antenna space diversity function for cellular telephones and is installed in vehicles. If provided with a switched d-c- connection port to charge the telephone's battery pack, it is referred to as a "docking switch". The battery pack charging function is not shown because our concern here is with RF switching circuits.

A functional diagram of the antenna diversity and T/R switch is shown in Figure 6.4.

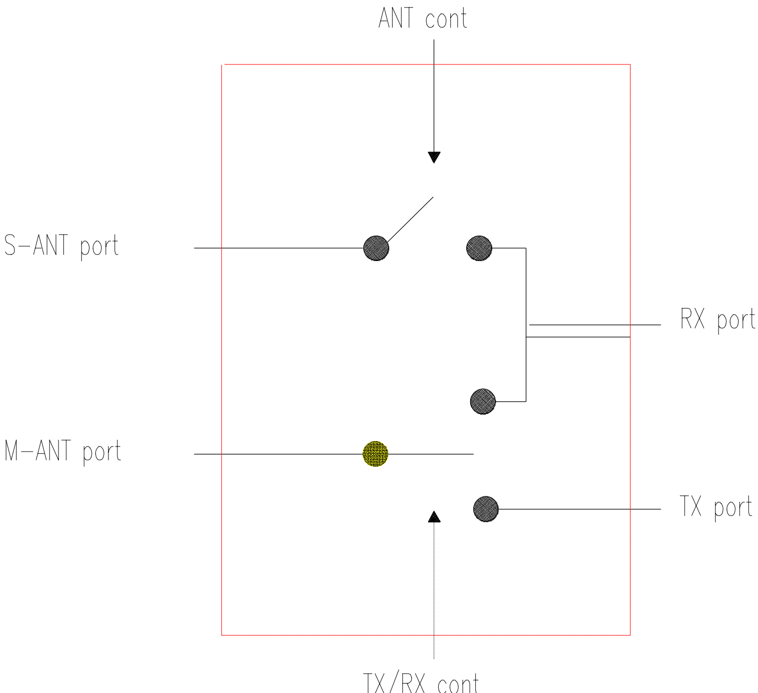

Figure 6.4 PIN Diode Antenna Diversity & T/R Switch for Cellular Telephones

This composite switch contains both a SPDT and a SPST switch (Chapter 2) and would be suitable for either United States or European cellular systems. The T/R switch function is performed by the SPDT switch. The added feature is that the SPST switch is positioned by a signal strength circuit to provide connectivity to the antenna that provides the stronger receive signal to the receive port.

In a typical vehicle mounted application, this composite switch would be designed to switch transmitter power of +38 dBm with a 1 dB compression point of +35 dBm. Isolation between transmit and receive ports would be about

30 dB. Both isolation and 1 dB compression point are specified with the transmitter connected to the main antenna (M-ANT port). The maximum switching speed of the SPST and the SPDT is 1 us. VSWR at all input and output ports is less than 2:1 for all possible positions of the two switches. The SPDT switch could be rendered either as a series PIN diode switch (Figure 2.7) or as a shunt PIN diode switch (Figure 2.8), depending on system design parameters such as bandwidth, power handling, and circuit construction technology. The UPP9401 would be an excellent choice for surface mount SPDT and SPST switches.

PIN DIODE ELLIPTICAL DIRECTIONAL ANTENNA ARRAYS FOR WIRELESS DATA MODEMS

Several firms are adopting an elliptical directional antenna array to provide space diversity for their mobile wireless data modems. The serving base station would have such antenna array as well. An array signal processor, with a sufficient number of antenna elements, can reduce the volatility of the RF link between a mobile unit and the base station. The design concepts were originally described for base stations in [1,2].

A simplified version of the base station array concept can be implemented to lower the BER and to diminish the effects of multi-path fading and co-channel interference for a mobile data modem also. A multi-element array, consisting of either a 6 or 8 antenna elements is used to:

Improve the “front-to-back” ratio (directivity) of the array

Improve the “self-interference” problem of dual arrays

Improve the antenna gain (6 elements yields 6 dB antenna gain)

The available power (P_{av}) from the transmitter output stage depends on the system’s overall linearity and BER specification (Appendix C). PIN diode switches are used to switch the elements of the antenna array, because the power output stage-antenna switch interface can be optimized to achieve the most linear solution [3]. If the effective radiated power from the data modem is 36 dBm @ 2.4 GHz, various combinations of antenna elements and available power gain will result in the following solutions:

Antenna Elements	Antenna Gain	Power Output	Linearity*
8	8 dB	28 dBm	more linear
6	6 dB	30 dBm	less linear

* A quantitative statement of linearity depends on their system parameters than those considered here.

With fewer antenna elements, the transmitter output gain is increased to obtain the required output power, thus decreasing the transmitter-switch overall linearity.

The elliptical antenna array is a specially configured switch matrix. It is a 1 x N switch matrix, where N is either 6 or 8, depending on the particular data modem antenna design. The array is center fed and connectivity between the input feed and the branch antenna elements is provided by a divider/combiner network.

A radial combiner geometry, is given in [3], which has circular symmetry, consists of an array of branch transformers (50 Ohms to 150 Ohms), that connect the antenna elements to the central junction or antenna feed port. An elliptical antenna array would require a more complex network of branch transformers since elliptical symmetry is not so simple as circular symmetry. However, the focus here is on the PIN diode switch design and

bias conditions. The analysis of the shunt mounted PIN diode switches [5] in the branch arms of the radial combiner is generally applicable to other combiner geometries as well.

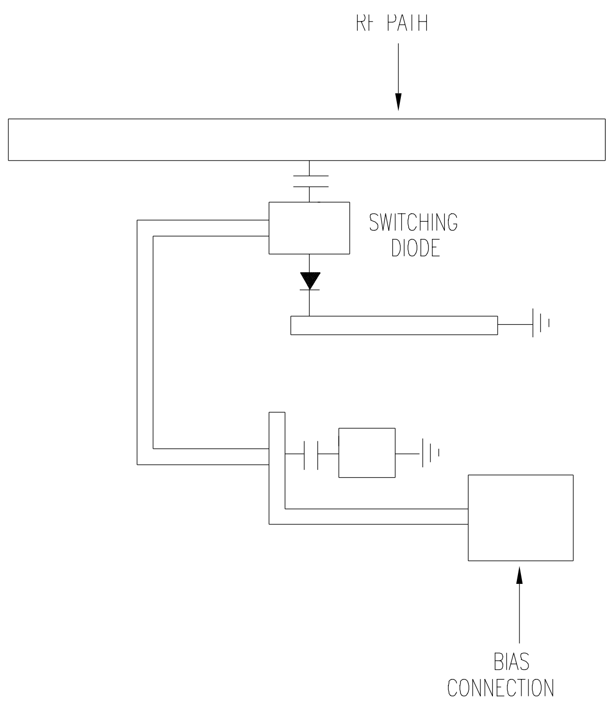

Figure 6.5 Shunt Mounted PIN diode Switching Network

The shunt mounted PIN diode switch is placed about a quarter wavelength from the central junction so that it appears as a virtual short circuit at the point of connection (or an open circuit at the central junction). Each branch antenna switch is connected to the central junction with a transmission line transformer (50 Ohms to 150 Ohms). The exact position of the switch along the transformer is determined by modeling the switch sub-network and transformer combination until a high reflection coefficient at 0 degrees phase angle is observed at the 150 Ohm end of the transformer. This modification of the switch position ensures that an open circuit is seen at the center of the radial switch for antenna ports that are turned off. Figure 6.5 shows the location of the PIN diode and the compensating network.

Shunt switches work by having a PIN diode connected between the transmission and the ground plane. When the diode is in the low impedance state, a short circuit is created and the incident wave traveling along the transmission line is totally reflected. If the PIN diode is in a high impedance state, it appears not to be in the circuit at all, allowing the incident wave to pass unimpeded.

In practice, the PIN diode has finite dimensions and its performance in the “ON” and “OFF” states is affected by the Resistance and Capacitance of the PIN diode junction and the parasitic reactances of the diode package. These effects are compensated by incorporating the PIN diode’s equivalent circuit into the sub-network design model that determines the design of the compensating network.

PIN Diode Automatic Gain Control (AGC) Loops

Automatic Gain Control (AGC) is an important power control function for mobile communications systems. A base station establishes the connection between two mobile units or cellular phones within the cell area that it

controls. It also maintains that link until the transmitted signal of either of the two mobile units is sufficiently weak that the specified Quality of Service (QoS) cannot be maintained. The weaker station is then handed off to an adjacent call's base station controller.

The AGC circuit in the mobile unit is the key element to the establishment and maintenance of a call. A base station monitors the signal strength of each mobile unit under its control. It transmits a control signal to the mobile unit's AGC circuit (via the link maintenance channel) when its transmitter output power must be adjusted to maintain call continuity. Thus, the mobile unit's AGC function is a fundamental part of the base station call link control function.

A block diagram of a generic RF- AGC circuit was shown in Figure 3.5 (Chapter 3). Figure 6.6 a typical block diagram for the AGC function in a cellular radio.

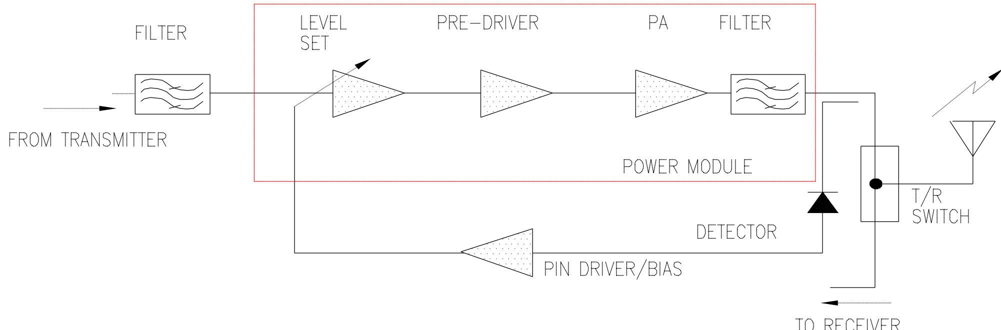

Figure 6.6   Typical AGC Loop For A Cellular Radio

The base station control signal is detected by the detector circuit in the decoupled arm of the AGC circuit. This signal is amplified and fed to the level set attenuator to adjust the input power to the pre-driver. If the mobile unit is approaching the base station, it will signal for a larger amount of attenuation in the mobile unit's level set to provide the smaller amount of transmitter power needed to maintain the call. Attenuation is decreased if more transmitter output power is needed for call link maintenance. Obviously, if the base station senses that the mobile unit cannot increase its transmitter power further, the base station call hand off controller initiates the procedure for call handoff to an adjacent cell's base station.

The level set attenuator circuit can be chosen from the numerous designs presented in Chapter 3. The usual choices are the Bridged TEE PIN Diode Attenuator (Figure 3.12) or the PIN Diode ✍ Attenuator (Figure 3.14). The UM9301 would be an excellent choice for this AGC level set at UHF and above.

OPTIMIZATION OF RF TRANSMITTER LINEARITY

The important parameters for optimizing the overall performance of the transmitter - T/R Switch combination are discussed in reference [4], which appears in Appendix C.

APPLICATION	**RECOMMENDED PIN DIODE TYPES**
WIRELESS	UPP1001-1004, UPP9401

CHAPTER - 7

PIN DIODE CONTROL CIRCUITS FOR HF BAND INDUSTRIAL APPLICATIONS

NOTES

PIN DIODE RF POWER CONTROL CIRCUITS FOR INDUSTRIAL APPLICATIONS

INTRODUCTION

Two new series of thick I-region PIN diodes have been developed by Microsemi - WTN specifically for Industrial RF Power Applications. The UM2100 Series was developed for the 2 - 30 MHz Band (HF Band), and the UM2300 Series was developed for the 200 - 3000 KHz Band (MF Band). The unique characteristics [1] of these diodes will be discussed in terms of their performance in a number of selected industrial applications.

HF BROADCAST BAND ANTENNA SWITCHES

Fundamental work on very thick I-region, long lifetime PIN Diodes was reported in the 1980's by the former RCA David Sarnoff Laboratories [2,3]. These PIN diodes were designed to be capable of replacing the mechanical drive, rotary transmit / receive switches then being used as multi-band antenna switches for exterior communications systems. The bands of interest were the 225 to 400 MHz band (a portion of the UHF band), the 2 to 30 MHz band (HF band), and the 10 to 300 KHz band (VLF band).

Subsequently, Microsemi - WTN has developed two series of thick, long lifetime PIN diodes [1] to provide devices for these and similar high power, lower frequency applications. The UM2100 Series (nominal lifetime of 25 micro-seconds) can switch high RF power (up to 2.5 KW) in HF Band and the newer UM2300 Series (nominal lifetime of 80 micro-seconds) can switch high power (up to 2 KW) through MF Band (200 KHz to 3000 KHz).

A seven-position single pole, useful for HF Band transmitter switching applications, is shown in Figure 7.1. The specifications for the seven- position switch are shown in Table 7.1.

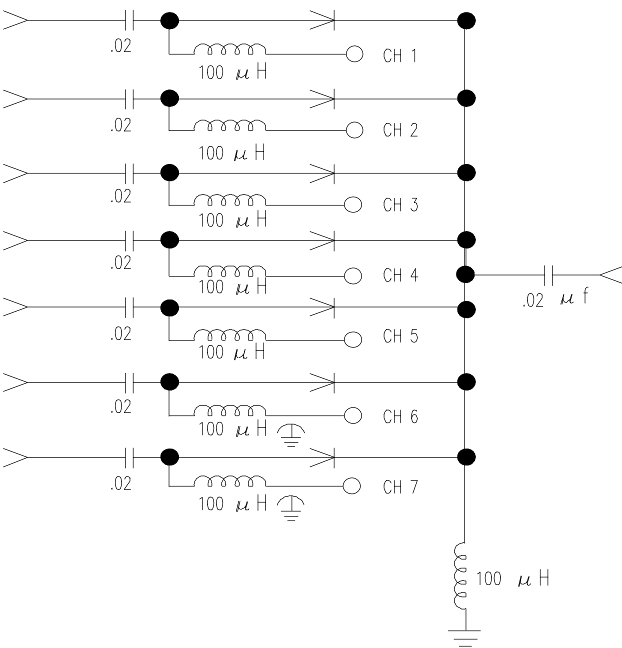

Figure 7.1 HF Band Switch Configuration

Table 7.1 Specifications for 7 Position HF Band Switch

Parameter	Value
Power Handling Capability	1000 W (Max)
RF Voltage	450.0 VRms (Max)
RF Current	8.9 ARms (Max)
Characteristic Impedance	50.0 Ohms
Number of Positions (Channels)	7
OFF State Isolation	25.0 dB (Min)
ON State Insertion Loss	0.5 dB (Max)
Forward Bias Current (ON State)	300.0mA
Reverse Bias Voltage (OFF State)	-400.0 V
IM 3 Distortion Products	-60.0dBc

An equivalent circuit for a switch branch, convenient for calculating the reverse biased switch performance parameters, is shown in Figure 7.2.

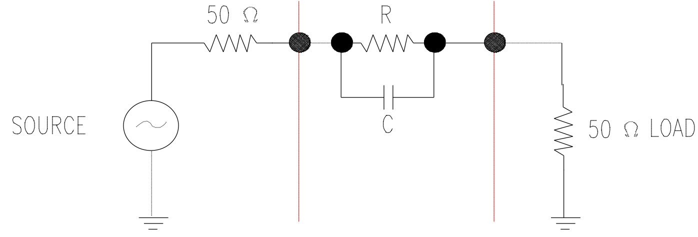

Figure 7.2 PIN Diode Equivalent Circuit

Figure 7.2 is similar to Figure 1 (c), Chapter 1, except that the series Inductance was omitted because it's Reactance is negligible in HF Band. The equations for calculating the Insertion Loss (I_L) and Isolation (I_{SO}) of a Single Pole, Single Throw Switch are given in Chapter 2. In HF Band, the UM2110 PIN diode is recommended for this application. If the PIN diode is forward biased such that Rs < 1 Ohm, then I_L is less than 0.1 dB. The Isolation depends on the PIN diode's Capacitance (Ct) under reverse bias. At 2 MHz, the Isolation is greater than 50 dB, and at 30 MHz, the Isolation is about 30 dB.

We note that the ON state Insertion Loss is constant with frequency but the OFF state Isolation varies with frequency since it is a function of the OFF state Capacitive Reactance of the PIN diode. Referring to Figure 7.1, these Insertion Loss and Isolation calculations are valid from any of the seven input ports (individually) to the common output port. When two ports are in the OFF state, their input port to input port Isolation is increased by 6 dB (above that of a single branch) because the OFF state Capacitance of the two PIN diodes in the dual branch is half that of a single PIN diode. Note that two input ports are not switched on simultaneously.

Appendix E contains a discussion of "Non-linear Effects in Semiconductor Devices" in which the inter-relationship between device non-linearity and rectification of an impressed signal is developed. The main

difference between a pn-junction diode and a PIN diode is that the pn-junction is a non-linear resistance that rectifies the input signal. The PIN diode, over a specific set of operating conditions, presents a linear resistance to the input signal, and thus, does not rectify the signal. These operating conditions are discussed in Chapter 1, sections titled: Large Signal PIN Diode Operation and Low Frequency PIN diode Operation.

It has been experimentally observed [1,2,3] that an incremental increase of signal power while the signal frequency is continually lowered, can cause an on-set of rectification in the PIN diode. This effect is observed when the dc induced stored charge in the I-region is no longer at least five times larger than the rf induced stored charge. The PIN diode rectifies the rf signal because it presents a non-linear impedance to the rf power source and load under these operating conditions.

Under normal operating conditions, PIN diode switches need a relatively small amount of dc bias current to control (switch) large values of rf current. This is the unique feature of a PIN diode. Several hundred milli-amps of d-c current can control several hundred Amperes of rf current.

PN-junction diodes (Varactors) can be used to switch high values of rf power, but the current demands on the switch's power supply are enormous. If a hundred Amps of rf current must be switched by a pn-junction diode, a hundred Amps of dc bias current must be available from the bias current supply. High dc current levels greatly complicate the implementation of rf / dc isolation circuits, especially at low frequencies.

HF BAND TRANSMISSION LINE TUNER

Industrial manufacturing equipment, such as rf sputterers, use a length of RF transmission line to connect the high power (several KW) transmitter or power amplifier to the load, some distance away from the source. Invariably, the load presents an impedance mismatch to the connecting transmission line. The usual solution is to introduce a matching network between the source and the cable. This matching network consists of a number of Capacitors of varying values that can be switched in or out depending on the nature of the mismatch. A typical impedance matching network is shown in Figure 7.3. PIN diode switches are gradually replacing relay switches for this application.

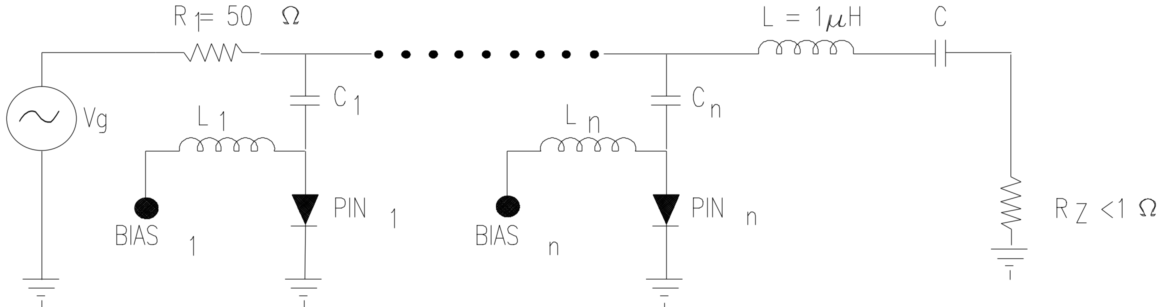

Figure 7.3 HF Band 50 Ohm Transmission Line Matching Network

Sputtering systems delivering 2 KW @ 2 KV line voltage are being produced for the semiconductor industry currently and 5 KW @ 5 KV systems are in the design phase. Microsemi - WTN manufacturers several series of High Voltage PIN diodes for these systems.

HIGH POWER TRANSFER SWITCH FOR HF BAND TRANSMITTERS

A number of manufacturers of HF Band Power Amplifiers for the Amateur Radio market have introduced solid-state switching using PIN diodes[4]. In comparing the various switching devices available to switch the antenna, such as vacuum relays, reed switches, or solid state devices, the ARRL Handbook (reference [5]) states " Perhaps the most modern and elegant approach to switch the antenna between the transmitter and the receiver is the use of PIN diodes. There are no keying-speed constraints when PIN diodes are used, and if the proper devices are

selected, the spectral purity of the output signal will not be affected (by the PIN diode switch). The most important parameter in this regard is the (minority) carrier lifetime".

Reference [5] is referring to the Transmit/Receive Switches as described in Chapter 2. Figure 2.11 shows a narrow band (5 to 10 %) Quarter Wavelength Antenna switch configuration and Figure 2.12 shows a broadband Antenna switch, both adequate for several hundred Watt applications. Figure 2.13 shows High Power Broad band T/R Switch that can control 1 Kilo-Watt transmitter power with excellent distortion characteristics (IM3 less than -80 dBc).

Another antenna switch configuration, shown in Figure 7.4, is the Double Pole - Double Throw (DPDT) Switch or the Transfer Switch, which is used to calibrate the transmitter-antenna interface.

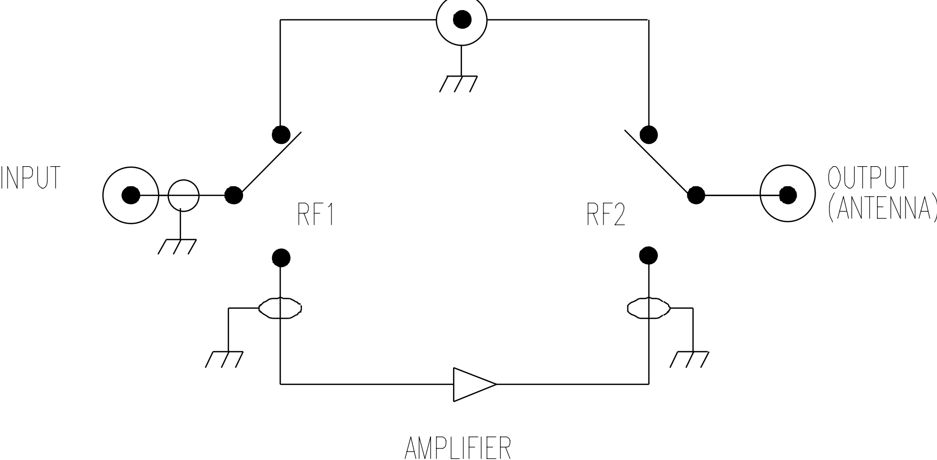

Figure 7.4. Equivalent Circuit of an Antenna Transfer Switch

The transfer switch allows either the "feed through" line or the Power Amplifier to be switched to the Antenna. The connector shown at the center of the "feed through" line may be terminated either in a short circuit (feed through state) or a 50 Ohm load, for transmitter calibration.

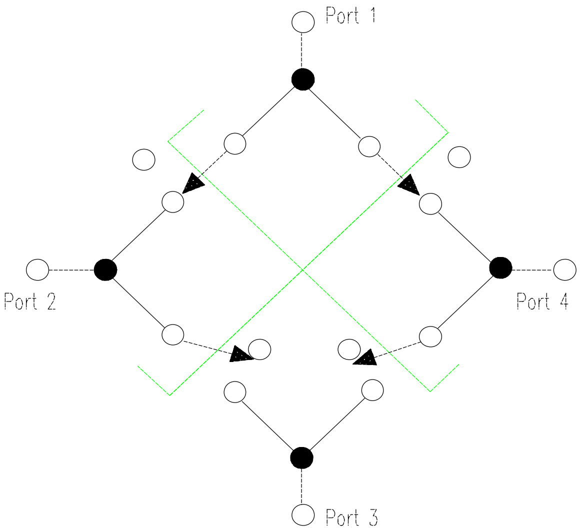

Figure 7.5. Double Pole-Double Throw Transfer Switch

In general, a Transfer Switch (Figure 7.5) is a four port device that has two valid operating states that are defined by the "truth table" (below). SPDT Switches can only connect a single input to either of two loads. SPDT Switches cannot perform the function of inserting sections of transmission line between the source and the load (antenna). Note that Figure 7.5 is not an exact representation of Figure 7.4, since the lower path in Figure 7.4 contains an amplifier and there is no independent Port 3 available.

TRUTH TABLE

Control Input State		RF1 - RF2	RF2 - RF3	RF3 - RF4	RF1 - RF4
1	2				
ON	OFF	ON	OFF	OFF	ON
OFF	ON	OFF	ON	ON	OFF

The Truth Table shows the bias control conditions that set either of the two valid operating states of the Transfer Switch. In control state 1, power flows from Port 2 to Port 4 via branches RF1 - RF2 and RF1 - RF4. In the case of Figure 7.4, this control state would provide the "feed through" line if Port 1 is terminated in a short circuit. Control state 2 "transfers" the inserted path to RF2 - RF3 and RF3 - RF4.

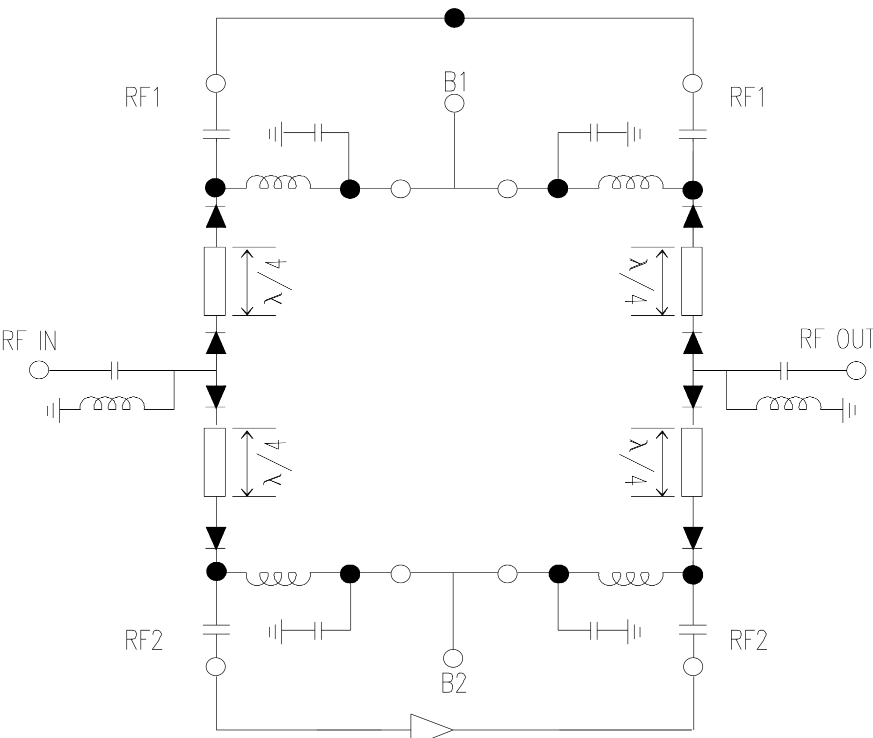

Figure 7.6 DPDT Switch Configured With Two SPDT Switches

The DPDT Transfer Switch in Figure 7.4 can be configured using two SPDT Switches interconnected as shown in Figure 7.6 such that the RF output #1ports of both SPDTs are joined at a common connector terminal and the RF output #2 ports are connected to the amplifier. When bias control #1 is ON (and bias control #2 is OFF), the through line is inserted between the source in the antenna. When bias states are reversed, the amplifier appears between the source and the antenna. The biasing sequence is the same as that indicated in the “truth table” above.

The performance characteristics of this Transfer Switch depend on the choice of the SPDT design. Various SPDT Switch design tradeoffs were discussed in Chapter 2. The choices consist of Series Diode SPDT, Shunt Diode SPDT, and Compound (series and shunt connected diodes) SPDT switches. Each design has its own bandwidth and power handling abilities.

The 4 Port Transfer Switch is a bridge network and hence has extremely wide bandwidth capability. The Transfer Switch shown in Figure 7.7 combines the wide bandwidth characteristic with the high power handling capability of the compound switch. Several hundred Watts and 0.5 - 18 GHz bandwidth switching capability are achievable with this design.

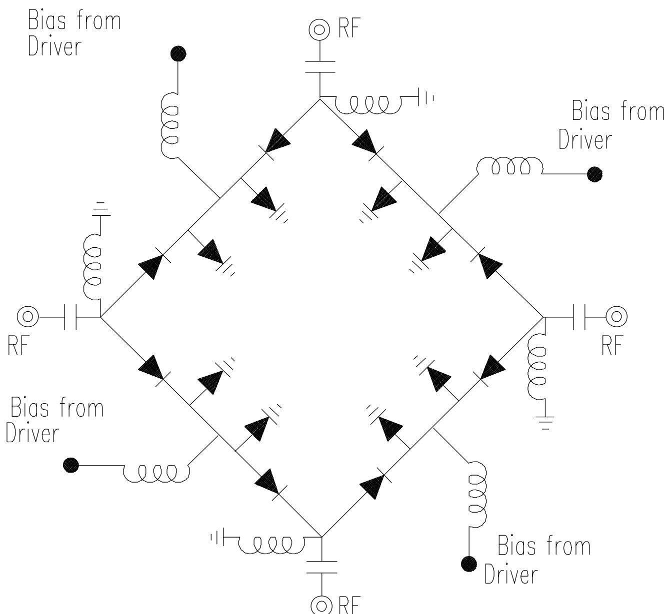

Figure 7.7. Broadband High Power Transfer Switch Configuration

RECOMMENDED PIN DIODE TYPES

DIODES

HUM2020, UM2100, UM2300, UM4000, UM4300

MODULE

UMM5050 5 KW, 5 KV

CHAPTER - 8

PIN DIODE CONTROL CIRCUITS FOR MAGNETIC RESONANCE IMAGING SYSTEMS

NOTES

MAGNETIC RESONANCE IMAGING SYSTEMS

INTRODUCTION

Magnetic Resonance Imaging (MRI) Systems manufacturers use a substantial number of very sophisticated Microwave Components in the RF and Signal Processing parts of their systems.

The basic MRI system used for medical diagnosis consists of a very large, powerful magnet to (0.1 to10 Telsa) surrounding a chamber that is large enough for a patient to lie down inside it.

It also uses a high power, frequency tunable RF source that can be rapidly switched on and off. This produces a large RF field perpendicular to the magnetic field. This RF field is focused by the body coil. The RF source and both coils must be tunable in both frequency and impedance to "match the impedance" of the patient's body.

HIGH - FIELD MRI RF RECEIVER COILS [1]

Special-purpose coils are designed to optimize the signal-to noise-ratio (SNR) from a given region of the body. State-of-the-art coil systems include the use of four or more coils with four separate receivers. This method is often referred to as a phased array system although the signals are not added such that the signal phase information is included [1].

Normally, the RF signal is in the range of 10 to 100 MHz. During a typical set of clinical image measurements, the entire frequency spectrum of interest is of the order 10 KHz, which is an extremely narrow band, considering that the center frequency is about 100 MHz. This allows the use of single-frequency matching techniques for coils because their inherent bandwidth always exceeds the image bandwidth. This is an extremely important consideration when specifying PIN diodes for coil switching elements.

The quality of the MR images depends on the Signal-to-Noise Ratio (SNR) of the acquired signal from the patient. SNR is of the utmost importance in obtaining clear images of the interior of the human body.

MRI RF - COIL DESIGN PRINCIPLES RELATED TO PIN SWITCHING DIODES

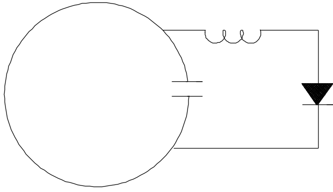

Figure 8.1 Simple Circular Loop With One Switched Gap

Figure 8.1 is a simple model of a circular loop with a single Capacitive gap. The gap is shunted by a series combination of an RF coil (L_S) and a PIN diode. The individual reactances of L_S and C_G are about 50 Ohms at the operating frequency. For simplicity, the bias circuitry and the 50 Ohm RF output line across the PIN diode are not shown. The value of Ls is chosen such that the Inductive Reactance of the coil (L_S) and the Capacitive Reactance of the gap are in parallel (phase) resonance when the PIN diode is forward biased. This parallel resonance causes a large impedance (or zero Conductance) to appear across the gap, causing the RF loop current to decrease to zero (open circuit or OFF state). Multiple PIN Diode Switch Configurations are also used in MRI System designs [1].

A practical MRI coil would have two or more gaps [1]. A second gap is needed to apply an RF synchronization pulse of frequency distribution {[sin x] / x} to time the initial test pulse and the image response pulse. The Capacitive gaps permit the flow of RF current through the MRI Loop. The PIN diode bias network inhibits the flow of RF current through the PIN diode, although the diode must withstand the RF line voltage when it is back biased.

KEY CHARACTERISTICS OF MRI RF - COIL PIN SWITCHING DIODES

Exclusion Of Magnetic Materials

PIN diode must not contain any magnetic materials, either associated with the die, die attach metalization system, or the RF package assembly. The existence of magnetic materials in the PIN diode structure distorts the static magnetic fields associated with the various coils and will interfere with system calibration and accuracy of clinical measurements.

Maximize Signal To Noise Ratio

When the MRI coil switches are OFF (i.e., reverse biased), the receivers are listening to the image return pulse. The receivers' SNR is degraded by the OFF impedance of the RF switch. This effect is specified by the Reverse Bias Leakage Current (I_R) at the PIN diode's Reverse Bias Breakdown Voltage (VBR). I_R @ VBR is alternatively specified by the equivalent parallel resistance (R_P) of the reverse biased PIN diode [Chapter One].

The gradual increase of SNR due to the increase of reverse bias leakage current is the result of poor passivation of the PIN diode's I-region. Microsemi PIN diodes are passivated with a unique proprietary glass passivation process to avoid this problem.

Impedance Matching Of The MRI Coils To The Patient's Body Impedance.

The most usual RF frequencies used for commercial MRI System design are 21 MHz and 64 MHz. The Image search and tune bandwidths are 4 KHz, 8 KHz, and 16 KHz. The absolute values of PIN diode parasitic impedances is less important than their potential variation from lot to lot. For such narrow band applications, these parasitic impedances can be compensated in the initial design of the switch.

PIN DIODE BIAS CONDITIONS FOR MRI COIL SWITCH APPLICATIONS

Since MRI Systems are designed to operate in the HF and VHF Bands, the bias conditions described in Chapter Two for HF Band High Power Switches apply, with the proviso that the PIN diodes are embedded in parallel resonant circuit across a loop gap.

FORWARD BIAS CONDITION

Since the forward bias resistance (R_S) of the PIN diode decreases as the bias current increases (Figure 3.4), Microsemi recommends the maximum bias current level compatible with the bias circuitry available, with 50 mA as a minimum value. The R_S of the forward biased PIN diode in Figure 8.1 adds to the loss of the switch circuit when in parallel resonance, which decreases the open circuit impedance across the gap.

REVERSE BIAS CONDITION

A conservative design would require that the VBR of the PIN diode chosen for a MRI switch design should be at least equal to the peal-to-peak RF Line Voltage and that the reverse bias applied to the PIN diode in the OFF state is at least equal to the peak value of the RF Line Voltage.

RF SWITCH DRIVER CONSIDERATIONS

During the High Power Reverse Bias condition, the RF Voltage should not exceed the PIN diode's VBR. If the RF Voltage swing exceeds the VBR, the driver must have sufficient reverse bias current capability to achieve the desired Switching Speed (Ts), but it must also provide the excess reverse current required during the high power RF pulses. Under this reverse biased leakage condition, the PIN diode may heat appreciably, causing an increase in the leakage current. If the leakage current is large enough, thermal runaway will cause the PIN diode to be destroyed.

RF POWER HANDLING

In a practical SPDT switch for an MRI body coil it would be required to handle 10 KW of RF power.

Power Handling capability is specified by the device manufacturer and depends on:

Maximum Power That Can Be Dissipated By The PIN Diode (P_d Max)

Thermal Impedance From The PIN Die To Thermal Ground (Φ_{JC})

Temperature At Which The Thermal Ground Is Maintained (T_A)

Current MRI designs use PIN diodes mounted on threaded stud packages, either directly on the electrical ground (C-style) or insulated from the thermal ground (D-style).

MICROSEMI PIN DIODES SOLD TO U.S. MRI EQUIPMENT MANUFACTURERS

P / N	VBR
UM 4006	600 V
UM 4010	1000 V
UM 4306	600 V
UM 4010	1000 V
UM 4006	600 V
UM 4010	1000 V
UM 9415	50 V
UM 2106	600 V
UM 2110	1000 V
HUM2010	1000 V
HUM2015	1500 V
HUM2020	2000 V

These devices are available in the glass leaded package (Style B), the 4 - 40 NC Stud package (Style C) and the insulated Stud package (Style D). The UM 2100 series is available in the SMT MELF package. The HUM2020 series is available as an axial leaded or studed versions. The stud thread size is 4-40 NC or 6-32 for the isolated stud version.

APPENDIX - A

PIN DIODE PHYSICS

NOTES

APPENDIX A INTRODUCTION

Appendix A addresses a variety of PIN diode physics topics that may be of interest to circuit designers who wish to probe more thoroughly into such issues as: how a PIN diode controls large values RF current with relatively small values of d-c current (sometimes referred to as "current gain"), what is minority carrier lifetime and how is it related to stored charge (Qs), why might the turn-off time (TRF) of a PIN diode be longer than the "turn-on" time (TFR), and why is the large signal operation of a PIN diode different in the microwave bands than in HF band?

Chapter 1 contains just enough preliminary information about PIN diodes so that Chapters 2 through 5 can be understood. Chapters 2 through 5 are written so that the circuit designer can focus on the specific control circuit function being described. It was felt that the introduction of some of these physics topics in Chapters 2 through 5, would distract many readers from the main focus of a particular chapter.

Appendix A is not meant to be a substitute for a text on semiconductor device physics. The treatment of the above topics varies in thoroughness, but references to the literature are given for those who wish to delve more deeply into a particular issue. Some physics topics are discussed in other appendices also.

Appendix B compares PIN diode circuit characteristics to those of pn-junction devices.

Appendix E discusses the topic of signal distortion in PIN diodes.

I SOME CHARACTERISTICS OF THE PIN DIODE I-REGION

Since the presence of the relatively wide intrinsic layer in a PIN diode is responsible for its unique properties, it is worthwhile discussing the I-layer first and then its characteristics, such as lifetime and stored charge. In fact, the equivalent circuit parameters, the switching times, and the distortion characteristics of the PIN diode are dependent on the properties of the I-layer as well.

Ideally, it is desirable for the I-layer to be intrinsic (the Silicon crystal structure is completely free of chemical impurities or crystal growth imperfections). With an intrinsic I-layer, the RF loss of the PIN diode under reverse bias would be minimized and its Capacitance would vary the least with reverse bias or frequency.

In reality, a truly intrinsic layer is not achievable within a PIN diode structure because it is not feasible to maintain intrinsic resistivity in the I-layer throughout the processing steps of the diode. Thus the concept of p+ and n+ layers separated by an intrinsic layer is somewhat artificial. Microsemi P-I-N diodes have I-layer resistivities of 8,0000 to 12,000 Ohm-cm, either slightly p-type (👌) or slightly n-type (🖫). These are much lower loss devices than competitive devices with much lower values of I-layer resistivity.

The ideal PIN diode has no impurities in its I-layer. Thus, at zero bias, the I-layer is already depleted of carriers. Since there are no mobile carriers in the I-layer to support current flow, the diode is an open circuit to RF signals and its Capacitance does not vary with reverse voltage or frequency.

A practical PIN diode has some space charge in its I-region, due to the presence of impurities. Some reverse bias is required to deplete its I-region of mobile charge. The value of reverse bias at which the space charge in depleted is referred to as the swept-out or "punch through" voltage. It is important to bias a PIN diode beyond the swept-out voltage for most RF applications, since current flow by these mobile carriers constitute an unwanted source of signal loss and noise.

The desired electrical parameters from the use of "intrinsic" silicon are off set by the difficulty of passivating such high resistivity silicon. The impurity level of 10,000 Ohm-cm P-type silicon is 1 impurity atom in approximately 30 billon silicon atoms and it has to be boron (P-type)! This is what makes the passivation process so difficult.

II MINORITY CARRIER LIFETIME

The volume or bulk lifetime in semiconductor is defined [1] as the average time interval between the generation and the recombination of minority carriers in a homogeneous semiconductor. Carriers are constantly being generated thermally in the bulk Silicon [2]. The generation rate is a function of the ambient temperature of the Silicon. Carriers can also be introduced into the bulk Silicon by connecting a Voltage source across the PIN diode junctions. There are two recombination mechanisms that determine the effective lifetime of a semiconductor device.

III RECOMBINATION PROCESSES

BULK RECOMBINATION

The fundamental effect that limits the lifetime in the bulk Silicon is the thermal-equilibrium condition, $pn = n_i^2$, where n_i is the free carrier density in the bulk Silicon at the ambient temperature. The basic recombination process is band-to-band recombination where an electron-hole pair recombines. Some bulk recombination also occurs via recombination centers or traps [2].

SURFACE RECOMBINATION

On the surfaces of the Silicon, nearly the entire recombination process is due to the presence of surface states or traps on the surface. Surface recombination velocity determines the minority carrier lifetime near the Silicon surfaces. Effective device lifetime is determined by the relationship

$$1 / ☺_{effective} = 1 / ☺_{bulk} + 1 / ☺_{surface}$$

IV CHIP GEOMETRY

The PIN diode chip structure is shown in Figure A.1. When the PIN diode wafer is processed, the bulk I-layer may have a minority carrier lifetime [1] somewhere in the range of 1 🖳s to perhaps 10 ms.
This wide range of possible values of lifetime (☺) is related to the particular details of the Silicon wafer processing steps. The lifetime of a PIN diode obtained from the wafer will be much less than that of the wafer.

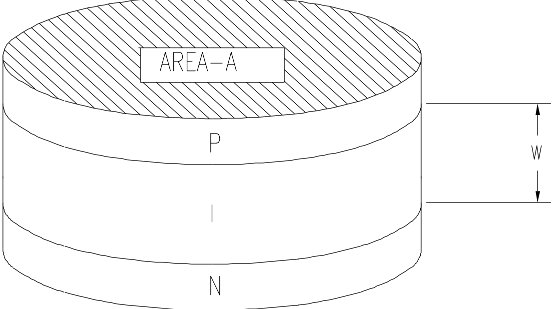

Figure A.1 PIN Diode Chip Structure

The geometry (ie, volume) of the I-region determines the bulk lifetime and is a major factor affecting the effective lifetime of a finished device [3]. The I-region is cylindrical in shape (Figure A.1). The areas of the outer surface also degrade (decrease) minority carrier lifetime. Because the periodicity of the I-region Silicon stops at these surfaces, the rate at which electrons and holes recombine is higher at these surfaces than at the center of the I-region. Figure A.2 shows contours of effective lifetime in Silicon as a function of the mesa (I-region) diameter.

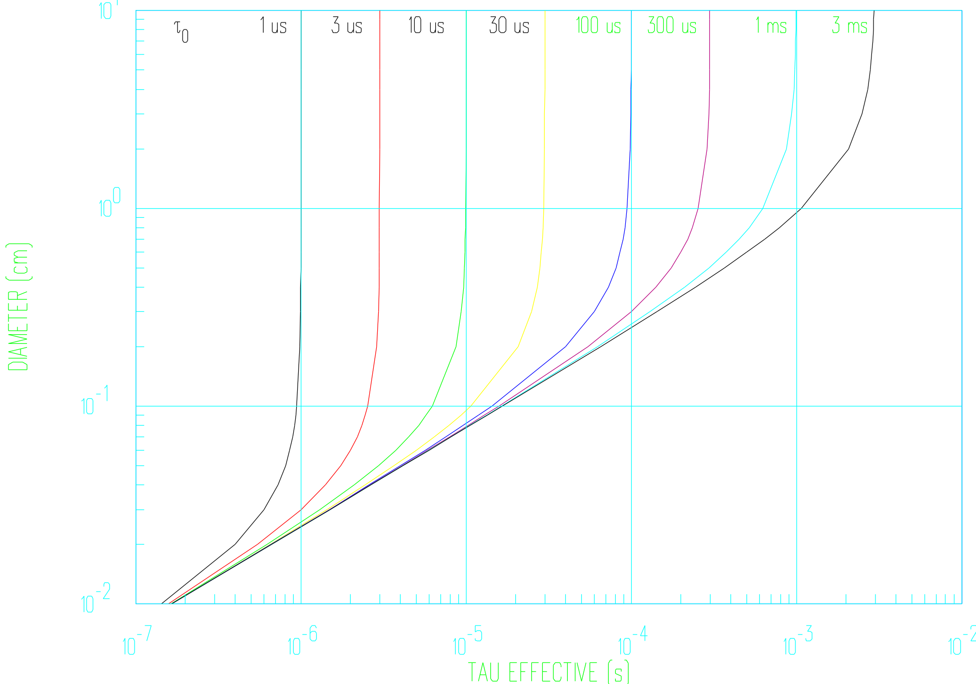

Figure A2. Effective Lifetime vs Mesa Diameter

V STORED CHARGE

When the PIN diode is forward biased, electrons and holes are injected into the I-region, where they have a finite lifetime before they recombine. The charge density in the intrinsic region and its geometry determine the resistance of the device. Lifetime determines the approximate low frequency limit of useful operation.

The diode’s resistance is proportional to the stored charge. The charge is related to the diode current by

I_f = dQ/dt + Q/☺

where I_f = diode forward current, Q = charge stored in the diode, and ☺ = effective recombination lifetime

If the diode is biased with only a constant current, the stored charge is constant and is equal to

Q = I_f ☺

At frequencies below f_O = 1/2✍☺ the RF signal modulates the stored charge the same as the dc bias and the PIN diode behaves as an ordinary PN diode.

VI FORWARD BIAS SERIES RESISTANCE (R_S)

The forward biased PIN diode behaves as a current controlled resistor that presents a linear resistance to the flow of RF current through the diode. This is the property of a PIN diode that enables the device to be used as the RF power control element in linear attenuators and modulators.

The forward bias equivalent circuit of a PIN diode, Figure A.3, consists of the forward bias resistance (Rs) in series with a fixed series inductor (Ls). Rs can be varied over a range of a few tenths of an Ohm to 100,000 Ohms by a direct or low frequency control current.

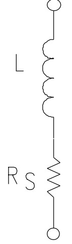

Figure A.3 Forward Bias Equivalent Circuit

The ideal diode equation, with any external voltage due to R_S subtracted from the junction voltage, is

$$I = I_O[e^{-q\,(V-IRs)\,/nkt} - 1]$$ “ideal diode equation”

where n = 1 if the diode’s space charge is diffusion limited and n = 2 if the space charge is recombination limited. Typical rectifiers have n = 1.0 to 1.5 and typical PIN diodes have n = 1.8 to 2.0.

The dc forward voltage characteristics of rectifiers have been extensively treated by Herlet, Benda, and Spenke [4,5]. Both articles do not provide a closed form solution similar to the “ideal diode equation”; however computer programs have been written to utilize their solutions.

At low frequency the resistance of the diode is

$$R_j = nkT/qI_{dc}$$ Ohms

where n = 2 for a well constructed PIN diode.

At room temperature

$$R_j = 52/I_{dc(mA)} \quad \text{Ohms}$$

The total resistance of a PIN diode is made up of R_j and any other series resistance contained in the chip (Ohmic contacts, diffusion layers, or substrate resistance) and resistance of the package and its leads. At frequencies above the dielectric cutoff frequency ($f > f_o$), the PIN diode I-layer's resistance is nearly linear [Appendices B & E]. The resistance of the I-layer is given by [6]

$$R_I = W^2/(2 🖳_{a\,p} ☺\ I_f)$$

where $🖳_{a\,p}$ is the ambipolar diffusion constant, W is the I-layer width, and ☺ is the lifetime in the I-layer

R_I does not contain the I region area explicitly but only the I-layer width. Since ☺ is a parameter that is I-layer geometry dependent, the I-layer resistance implicitly depends on the area as well.

VII CAPACITANCE

The reverse bias equivalent circuit, Figure A.4, consists of the shunt combination of the I-layer capacitance (Ct) and parallel resistance (Rp) in series with the inductance (Ls). The reactance and resistance of the depleted I-layer are represented by Ct and Rp. Ct contains the stray capacitance due to the effects of the diode's package structure as well as the junction capacitance (Cj).

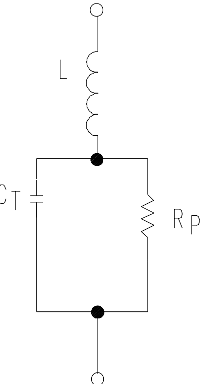

Figure A.4 Reverse Bias Equivalent Circuit

Due to the structure of a N+ and P+ section separated by an I-layer, the N+ and P+ sections accumulate charge. A thin depleted layer with no charge occurs. This layer acts as a parallel plate capacitor (Cj).

$$Cj = 📪_r\ 📪_{🖎} A / W$$

where W varies with the reverse bias voltage.

A well designed low frequency PIN diode will have a flat Capacitance versus Reverse Voltage curve even at low frequency. Figure A.5 shows such a PIN diode with a flat curve to below 4 MHz.

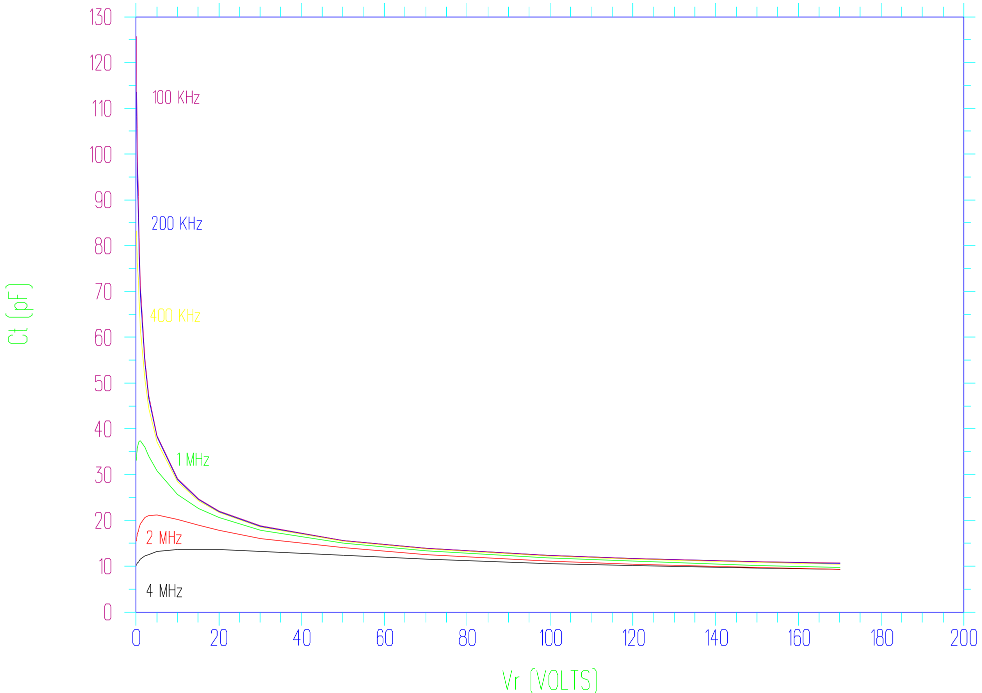

Figure A.5 PIN Diode Capacitance vs Reverse Bias Voltage and Frequency

VIII REVERSE BIAS PARALLEL RESISTANCE (Rp)

The reverse bias equivalent circuit, Figure A.4 shows the shunt loss element (Rp) associated with the PIN diode's total capacitance (Ct). This is the industry's conventional representation of the loss element of a reverse biased capacitor. So zero bias shunt loss is represented by an Rp whose value is infinity (ie, an open-circuit).
A series representation would serve just as well as the shunt representation, and might lend some intuitive insight (zero loss would be represented by a series element whose value is zero!). Reference [7] is an authoritative text on Varactor Applications. Varactors are only useful devices when reverse biased. The main premise of reference [7] is the ease with Varactor circuits can be analyzed if the series representation of the Varactor loss is used.
An ideal PIN structure (Section I, above) would have no shunt loss, because there are no mobile carriers in the I-layer to support conductive current. In practical PIN diode structures, Rp is highly dependent on the wafer processing steps and on passivation of the cylindrical surface of the diode chip prior to packaging. A well passivated PIN diode would have an Rp in excess of 50,000 Ohms when back biased beyond the punch through voltage.

IX REVERSE BIAS EQUIVALENT CIRCUIT

Figure A.6 gives the reader an intuitive feel for the connection between the PIN diode chip model and the reverse bias equivalent circuit. It also shows both the high frequency (microwave bands) and the low frequency (HF band) paths through the model [8].

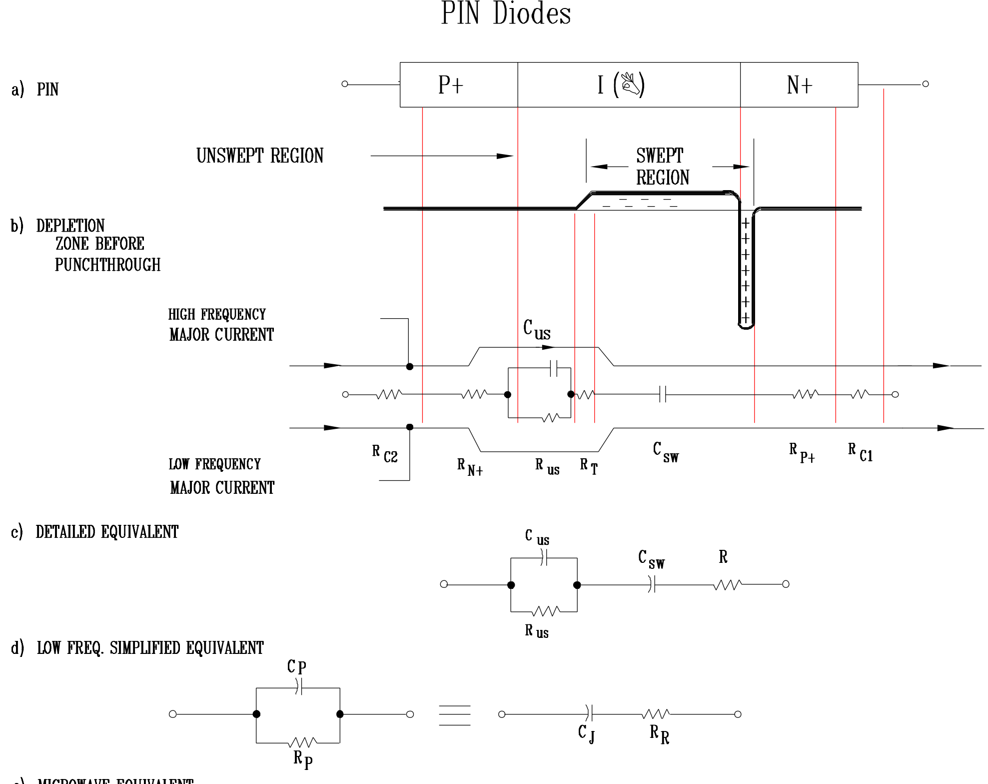

Figure A.6 Reverse Bias Equivalent Circuit

Microsemi's unique passivation process produces very low leakage and very stable reverse characteristics in which the device may be driven hard into reverse breakdown. In a 2 us pulse form currents as large as 100 mA at approximately 3000 volts may be safely handled. Figure A.7 is a typical passivated high voltage PIN.

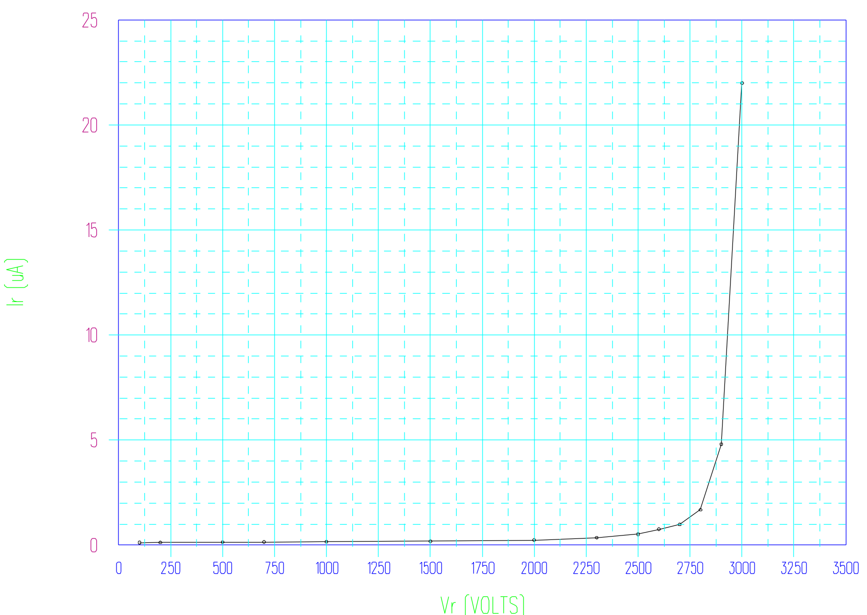

Figure A.7 TYPICAL PASSIVATED HIGH VOLTAGE PIN

X SWITCHING

Up to now the forward and reverse characteristics have been discussed separately. What happens when the PIN diode is switched[9,10]?

REVERSE BIAS TO FORWARD BIAS

The PIN diode, with sufficient reverse voltage to fully deplete the I-layer, will be depleted of mobile carriers. This depletion region will collapse almost instantly and the injection of carriers begins with a finite time to fill the I-layer with charge.

FORWARD BIAS TO REVERSE BIAS

The turn-off time is not a reverse of the turn-on time. When a large reverse bias is applied to a forward bias PIN diode a large current flows limited only by the impedance of the voltage source and the circuitry of the bias feed to the PIN diode. [The RF choke and the low pass filter to supply the bias current should have a very low dc impedance.] The edges of the die closest to the contacts deplete of charges quickly; however, the charges near the center of the die have to diffuse out to the external contacts. As more and more charges are removed, the reverse voltage across the I-layer begins to increase aiding the removal of the charges. By using spiked voltage drivers and by reducing the source impedance, the switching time for a diode can be substantially reduced. Figure A.8 shows the charges in the I-layer and their removal as a function of time.

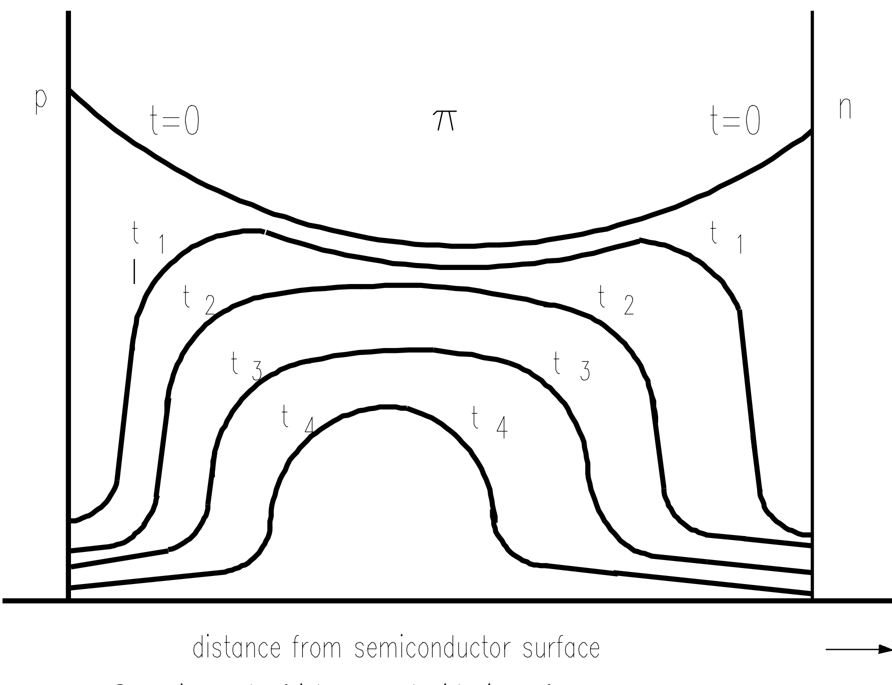

Figure A.8 Charge removal from chip versus time

APPENDIX - B

A COMPARISON OF PIN DIODE & RECTIFIER DIODE CHARACTEISTICS

NOTES

RF Application Note MPD - 101A

A Comparison of PIN Diodes & Rectifier Diodes

W.E. Doherty, Jr,
Microwave Engineering Manager **March 01, 1998**

Introduction

Many RF circuit designers are unclear about how a PIN diode works in an RF circuit and how it differs fundamentally from the circuit performance of a rectifier diode. This Application Note compares the physical properties and electrical behavior of two classes of Silicon semiconductor diodes: those that rectify an a-c signal (Silicon pn-junctions and Silicon Schottky junctions) and those that do not rectify an a-c signal (Silicon PIN diodes). We begin with a discussion of the various electrical properties of Silicon rectifier diodes because the reader is most likely familiar with them already. We then continue with a discussion of unique electrical properties of Silicon PIN diodes and make a comparison of the applications of both classes of diodes.

I Rectifier Diodes

Rectification [1] is generally defined as the process of converting an alternating current to a unidirectional current. A rectifier device conducts current substantially in one direction only. An ideal rectifier diode would be an open circuit in one direction and a short circuit in the other direction. It also would not dissipate power during the rectification process.

Pn junctions and Schottky junctions rectify RF current. The current flow in a pn junction is comprised of minority carriers (holes and electrons), whereas the current flow in a Schottky junction consists only of majority carriers (electrons). The reader not familiar with these types of junctions and the various mechanisms for current flow across semiconductor junctions is referred to reference [2].

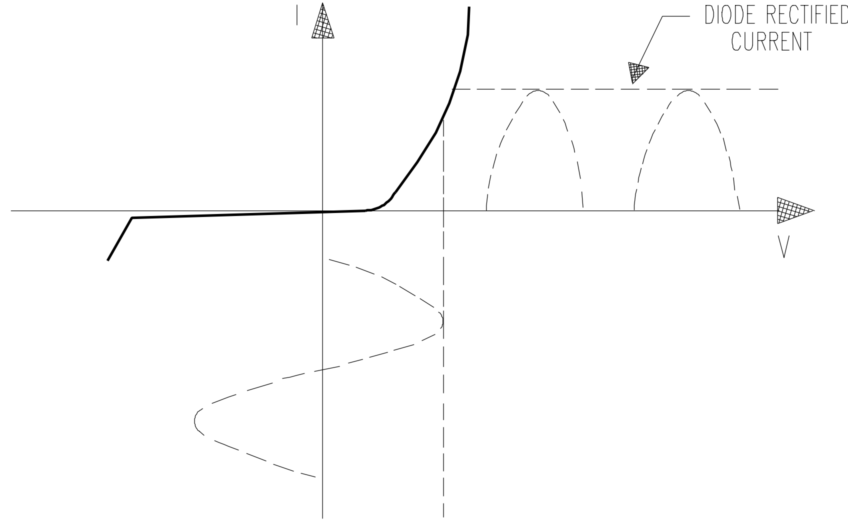

Figure 1. Half-wave Rectification Waveform

The rectifier presents a non-linear resistance (Figure 1) to an a-c current source. The rectifier diode's I V characteristic shows an exponential forward bias curve given by equation 1.

$$I = I_s \left(e^{qv/nkt} - 1 \right) \qquad (1)$$

I_s = Saturation Current
n = Ideality Factor, $1 < n < 2$

The reverse bias "blocking " state, for values of reverse voltage less than the reverse breakdown voltage (VBR), removes the negative half cycles of the input signal. The equation for the reverse bias characteristic is quite complex [2]. Empirically, the reverse bias current increases gradually with reverse voltage until VBR occurs. The breakdown current consists of both avalanche and Zener components and depends mainly on the surface conditions along the periphery of the junction.

The full sine wave signal is shown on the negative vertical axis of Figure 1. The peak voltage of the sine wave is less than VBR. The output current flows in the forward direction only and consists of the forward positive half cycles of the input current wave as shown on the horizontal axis of Figure1. The negative half cycles are blocked from the output by the rectifier's high reverse bias resistance. The current output from the rectifier is unidirectional but it must be filtered to be useful as a d-c bias source.

II Rectification - Non-linear Effects

Equation 1 is the defining relationship for rectifier diodes. The relationship between I and V is exponential and therefore highly non-linear. Heuristically, we observe that a device is linear if the output signal is a faithful replica of the input signal. The input signal's amplitude may increase (gain) or decrease (attenuation) but the shape of the waveform does not change. If the shape of the waveform changes, the device has had a non-linear effect on the input signal. In the case of Figure 1, only the positive half cycles appear in the output of the rectifier. The rectifier has performed a non-linear transformation on the input signal and the output has some added harmonics of the input frequency that the input signal did not have. The subject of non-linearity and signal distortion is discussed in detail in reference [3].

Non-linear devices are highly useful for frequency conversion processes. Frequency up-converters (modulators) and frequency down-converters (mixers) depend on the non-linear devices for their performance characteristics. A rectifier diode is an a-c to d-c converter. However, when non-linear devices are used to switch or attenuate RF power, they will seriously distort the input signal unless biased properly.

III PIN Diode - A Current Controlled Linear Resistor

A PIN diode is a pn-junction with a doping profile tailored in such a way that an intrinsic layer (I-region) is sandwiched between the p-layer and the n-layer. A p-👌 -n diode has a p-👌 junction and a 👌-n junction in series (separated by the length of the I-region).

Microwave PIN diodes are manufactured using an epitaxial process. The 👌 or 🗀 layer is nominally a 100 Ohm-cm layer, grown on a heavily doped p+ or n+ type substrate. These diodes provide adequate switching performance above 2 GHz.

Microsemi Corp's RF PIN diodes are based on high voltage rectifier technology. In these P+✌N+ and P+ 🗀 N+ structures, only the high resistivity layer contributes series resistance (R_s).

Under forward bias conditions, the I-region resistance is given by equation (2).

$$R_s = W^2 / (2\,\mu\,\tau\,I_f) \qquad (2)$$

where: μ = ambipolar mobility of minority carriers

τ = effective lifetime of minority carriers

W = effective width of the I-region

I_f = forward bias current

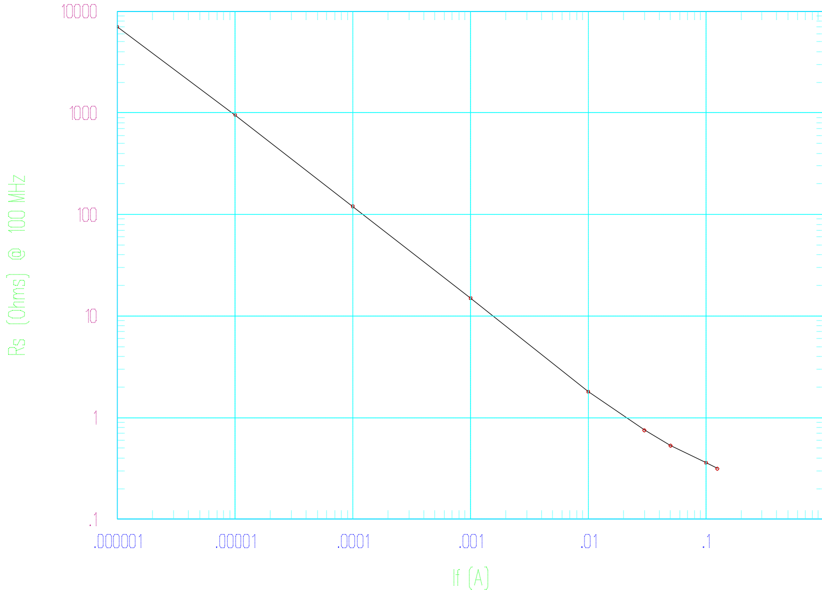

Figure 2. PIN Diode Series Resistance vs Forward Bias Current - UM 9401

This relationship between R_s and I_f is shown in Figure 2 for a UPP 9401. When the PIN diode is forward biased, it is a current controlled resistor that presents a linear resistance to the flow of RF current. Above 50 mA, R_s begins to approach the parasitic resistance of the p+ , the n+ layer, and the Ohmic contact layers. The range of linear operation occurs above a lower cut-off frequency given by equation (3).

$$f_c = 10 / \tau \qquad (3)$$

The RF power that can be controlled depends on the frequency of operation and the peak RF current [3]. PIN diodes can control a relatively large amount of RF current with a small amount of d-c bias current if the frequency of operation satisfies equation (3).

The bias current condition for linear operation is given by equation (4).

$$I_f > I_{rf} / (\pi\, f_o\, \tau) \qquad (4)$$

where I_{rf} is the peak value of the RF current waveform. Optimum bias conditions for high power operation in HF band are discussed in detail in references [3,5]. As a lower bound on equation (4),

$$I_f > I_{rf} / 20 \qquad (5)$$

IV Non-linear Effects In PIN Diodes

PIN diodes are designed and manufactured to enhance the linearity of the forward biased resistance vs forward bias current characteristic of equation (2) and Figure 2. But as the operating frequency approaches the cutoff frequency defined by equation (3) and /or the d-c forward bias current is inadequate to control the RF current, the PIN diode will begin to rectify and non-linear distortion effects will be evident. This effect can be seen experimentally and is discussed in reference [4].

If the RF induced charge (in the I-region) is nearly equal to the d-c induced charge, the forward bias resistance is not only controlled by the d-c bias, but is modulated also by the RF current. Equations (3) & (4) define the conditions for linear operation of PIN diodes.

V Comparison of Rectifier & PIN Diode Applications

So far, the PIN diode discussion has focused on its forward bias characteristic and that implies that the PIN diode is being used in an RF power attenuator circuit. Rectifiers are entirely unsuitable for this application if signal distortion is an issue.

The differences in electrical performance of either the rectifier or PIN diode are most apparent in the RF switch application. Within the constraints of equations (3) & (4), the PIN diode can switch large values of RF current with a small amount of d-c bias current. In VHF band, a d-c bias of 50 mA can control 1 Ampere of RF current if the minority carrier lifetime is adequate (2us). A rectifier diode can be used toswitch 1 Ampere of RF current, but the d-c bias current must be at least 1 Ampere also!

VI Conclusions

RF PIN diodes perform a unique function for RF switch and RF attenuator designers that pn-junction and Schottky junction devices cannot perform. In the switch application, small levels of d-c bias control can control large amounts of RF line current and not distort the input signal waveform. Their unique forward bias R_s vs I_f characteristic provides nearly distortion free RF signal attenuation when biased over the proper current ranges. Longer lifetime PIN diodes are becoming available which will provide the devices needed to switch or attenuate large levels of RF power from MF band through VHF band.

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APPENDIX - C

OPTIMIZATION OF OUTPUT POWER AND IM3'S OF WIRELESS RADIO TRANSMITTERS

NOTES

RF Application Note MPD 102A

The Use of Low Distortion PIN Diode Switches in Digital Communications Links

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Microwave Engineering Manager

March 15, 1998

Introduction

The digital communication link [1] encompasses the entire communication path, from the transmitter's input signal source, through all the encoding and modulating steps, through the transmitter and the RF channel, up to and including the receiver with all its signal processing steps, and terminating at the receiver's output load. Thus the link is more than just the free space RF channel.

In this paper, the overall performance parameters of the digital communications link are described. The non-linear distortion characteristics of Silicon and GaAs switching devices are discussed and related to the the optimization of the RF transmitter's linearity. A comparison of the inter-modulation performance of these switches enables a system engineer to select a switch design for a specific system application.

Digital communications system performance degradation can be related to two major contributors:

The deleterious effects of the RF Communications Link on the propagation of the RF signal

The non-linear noise effects on signal processing in the RF sections of the transmitter and receiver.

The Digital Communication Link

A digital communications link is the free space path between the antennas of two communications units. Propagation through the atmosphere and near the ground results in absorption, reflection, refraction, and diffraction, which modify free space transmission characteristics and cause signal fading.

Signal fading has many forms [2] such as frequency independent and frequency dependent fading, atmospheric multipath fading, and ground reflection multipath fading. RF signal fading effects cause increased Bit Error Rate and Inter-symbol Interference. These effects are offset somewhat by the following techniques:

Slow varying fading can be avoided by handing off a mobile connection to an adjacent cell that offers less fading [3].

Code Division Multiple Access (CDMA) is the usual solution to multipath fading and co-channel interference [3].

Spread-spectrum techniques improve the bandwidth efficiency and interference characteristics of cellular radio systems [4].

Smart antenna techniques overcome many of these various impairments [4].

Non-linear Effects in the RF Sections of the Transmitter & Receiver

Deterioration in communications link performance is also related to nonlinear noise effects in the RF sections of the transmitter and receiver. The effects of these sources of noise or interference are increased by the nonlinear behavior of the transmitter RF power amplifier and the T / R switch.

In evaluating communications system performance, the most important quantity is the Signal-to Noise Ratio (SNR), because the system design depends on the ability of the system to detect a signal in the presence of noise, with an acceptable probability of error. Since the RF output signal is a modulated carrier wave form, it is useful to refer to the average carrier power-to-noise (C/N) and the SNR of interest.

Non-linearity in Silicon PIN Diodes and GaAs MESFET's

The Third Order Inter-modulation Product (IP3) is one measure of device non-linearity and its use implies that the device is a "third -order device" ie, the curve of the third -order inter-modulation power vs input power has a constant slope (Figure 1).

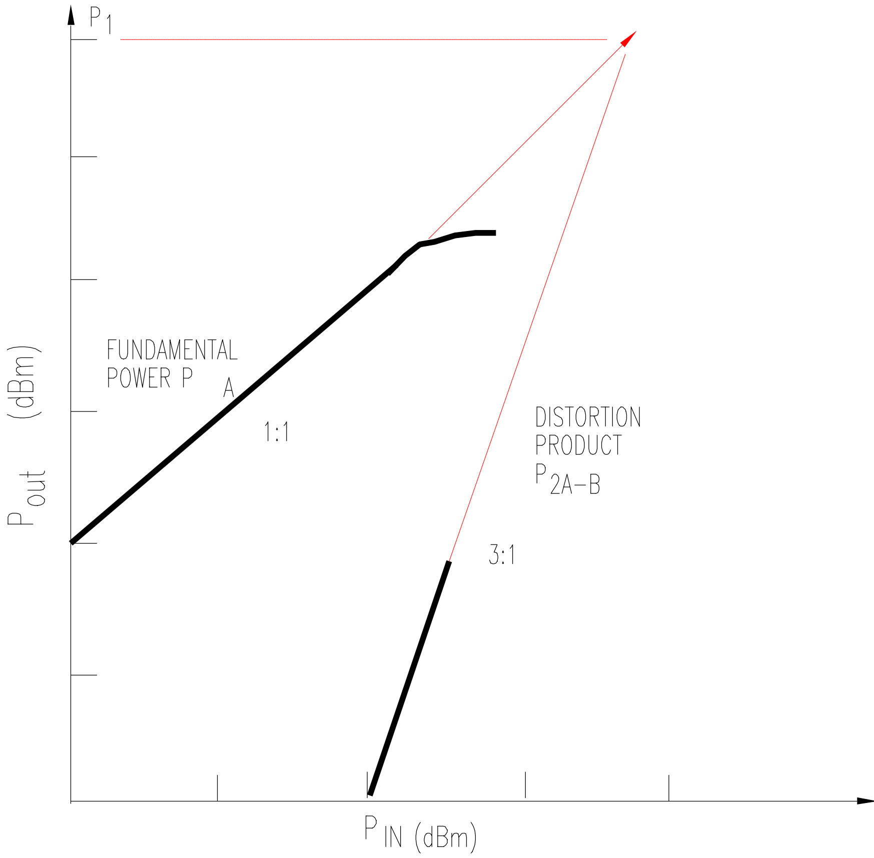

Figure 1. Definition of the Intercept Point

Silicon PIN diodes are third-order devices and their specified IP3 values are valid up to and beyond 10 Watts, depending on the design parameters of the specific PIN diode. Silicon PIN diodes are inherently higher power, lower distortion devices than GaAs MESFETs. In the forward biased "ON" state, the effect of conductivity modulation by the RF signal is minimized by the stored charge, $Q_S = I_F$☺ in the PIN diode's intrinsic region [5,6]. If the transmitter power is increased, the forward bias current can be increased to obtain the same IM3 as specified.

GaAs MESFETS are not third-order devices [7]. IP3 data, measured with input tone power of 5 to 7 dB, is not valid at higher input power levels. GaAs MESFET manufacturers cite the 1 dB compression point, P_{1dB} as a

measure of distortion. But P_{1dB} is not indicative of AM to PM phase distortion. GaAs MESFETs are inherently lower power switching devices. The gate -source junction is a rectifying Schottky junction with higher distortion properties that a PIN diode, at any power level. Table I shows typical T/R switch data for both Silicon and GaAs MESFET switches.

Table I

Typical T/R Switch Device Specifications

	Silicon	**GaAs MESFET**
IL	**0.25 dB**	**1.0 dB**
ISO	**40 dB**	**40 dB**
IP3	**65 dBm**	**45 dBm**
PS	**10 W**	**2 W**
"ON"	**10 mA**	
"ON"		**-10 V**
Test Tones	**1 W**	**10 mW**

Non-linear Distortion Effects

Device Non-linearity

All physical devices have some degree of non-linearity and thus distort the signal transformation process. The elements of a linear equivalent circuit are derived from small variations about the d-c operating point. Thus linear or "small signal" means that the circuit is limited to the first order derivative (of the device's I (v) function. By extension, a non-linear device's equivalent circuit is defined by higher order derivatives.

Non-linear distortion may be specified in terms of gain compression, AM - PM distortion, and inter- modulation products (specified in terms of intercept point, Figure 2). For devices with a well-behaved third order IMD performance (3:1 slope), the intercept point definition enables the inter-modulation performance to be specified independently of incident power level.

Optimization of RF Transmitter Linearity

The important performance parameters for optimizing the Power Amplifier - Transmit / Receive Switch combination [8] are shown in Figure 2. Each RF components makes its own contribution to the overall system's inter-modulation distortion level. In hand-held transceivers, the power amplifier may be a "linearized" class AB amplifier to achieve the desired CNR or IM3 and power-added efficiency.

Often a GaAs MESFET is chosen for the T/R Switch because of convenience and apparent cost saving. Their catalog specifications can be misleading, especially regarding IM3 performance at the power amplifier's output level. Since the GaAs switch's IM3 level is a function of the input ower, the P/A-T/R combination is optimized by backing off the P/A output power until the transmitter distortion specification is met.

The optimum solution is to improve the distortion performance of the T/R switch by using a Silicon PIN diode switch design. C/N and BER improve because the PIN diode's IM3's are much lower (typically 3 to 4 dB. The power amplifier's output power can be increased until the IM3 of the amplifier equals the IM3 of the switch

(Figure 2).This solution can be quite cost effective because the use of a Silicon PIN diode switch dramatically improves the system without resorting to more expensive alternates to raise the Carrier to noise ratio. The IM3 of a Silicon PIN Switch is typically 70 dBc at the 1 Watt input power level . For new designs, a Silicon Pin Switch can be combined with a more highly linearized P/A to achieve a CNR improvement of 8 to 10 dB, without the cost of increasing the system/s carrier power level.

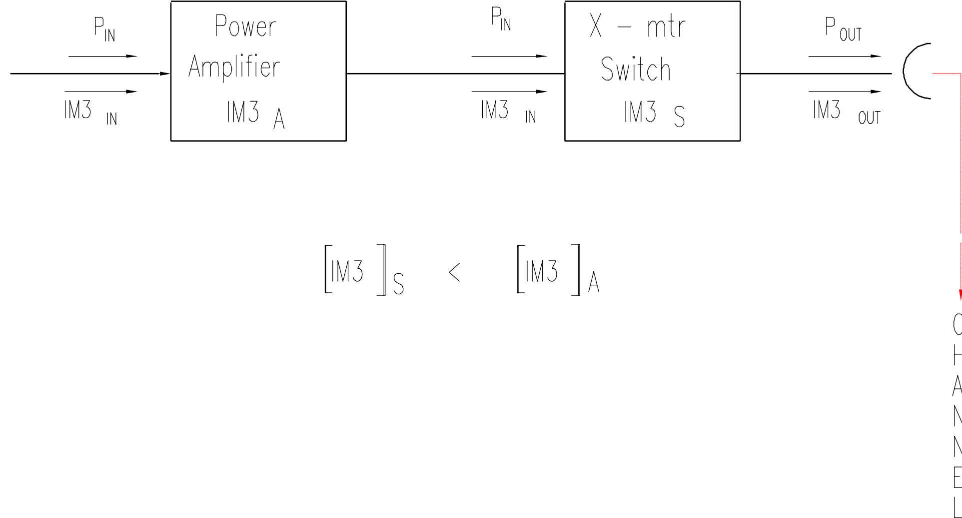

Figure 2. RF Transmitter Optimization

Conclusions

Recent trends in Telecommunications and Wireless Networking place greater linearity demands on RF components. Bit Error Rates (BER) of E -03 are adequate for digital voice transmission. Nomadic Data/ Miltimedia services require a BER of E -10 which relates to an end-to-end IM3 performance of -80 dBc at 36 dBm of effective radiated power. GaAs IC's cannot provide this level of linearity.

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APPENDIX- D

PIN DIODE DRIVER CIRCUITS

NOTES

DRIVER CIRCUITS FOR HIGH POWER PIN DIODE SWITCHES

The information in this chapter is presented as a guide for the RF switch - switch driver interface. References [1,2] provide a fuller treatment of this subject and can be used as a basis for RF switch - switch driver interface designs. Most of the information in this chapter is based on experimental work done on driver circuits used to evaluate thick I-region, large area, long lifetime PIN diodes. These long lifetime PIN diodes present singular difficulties in driver design because of the potentially large amount of stored charge in the I-region.

This chapter treats only two-state driver circuits in which either a forward bias current or a reverse bias voltage is to be applied to a PIN diode. Not covered are analog drivers as used in such circuits as linear attenuators and continuous phase shifters, where a precise bias current or voltage profile might be required.

White[1] lists five performance categories for driver circuits.

1. Input signal power supply buffer and TTL compatibility
2. Switching speed
3. High power, pulse leakage current supply
4. fault detection
5. Complementary drivers

It should also be noted that the relationship between the driver transistors and the PIN diode are as follows:

DRIVERS	PIN
PNP	cathode to ground
NPN	anode to ground

Since high voltage NPN transistors are easier to obtain, high power PIN circuits require anode to ground PIN diodes. Microsemi supplies both polarities of PIN diodes.

Two starter circuits from [1] are shown below. Figure D.1 is a typical circuit and Figure D.2 is the same circuit with speed-up components.

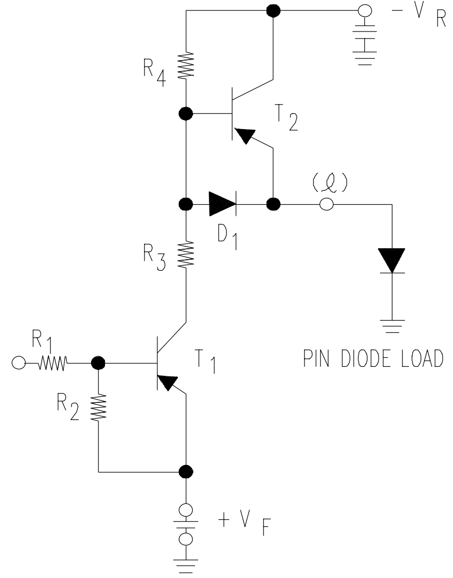


Figure D.1

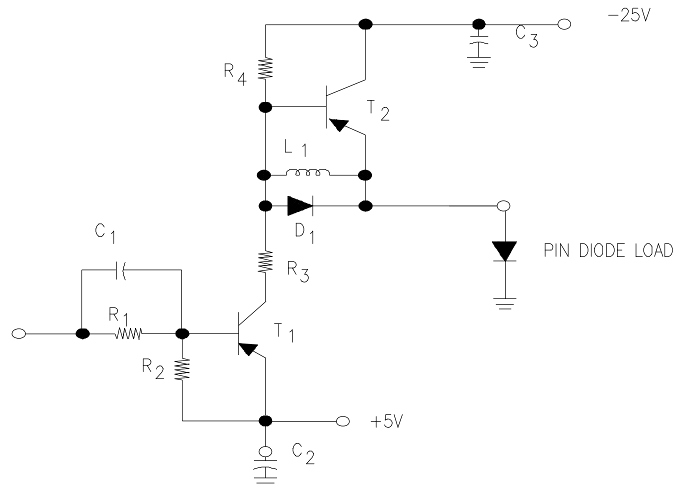


Figure D.2

Typical starting point values for using the circuit in figure D.2:

Q1, Q2 NTE 292 (TO220 PNP)	C1 75 pF
R1 499 Ω	C2 150 µF
R2 1K Ω	C3 0.05 µF
R3 15 Ω 5 W	L1 1 mH
R4 10K Ω	
R5 1K Ω	
D1 1N5811	
PIN diode HUM2020	

An additional good reference for driver circuits is [2].

APPENDIX - E

PIN DIODE DISTORTON

NOTES

PIN DIODE SIGNAL DISTORTION

INTRODUCTION

Signal Distortion in electrical networks is caused by the non-linear characteristics of the circuit elements and components of which the network is comprised. All electrical components exhibit some degree of non-linear behavior under certain operating conditions but the principal contributors are active devices such as semiconductor diodes and transistors. Non-linear distortion [1] may be described in terms of gain compression, AM-PM distortion, and intermodulation products (specified in terms of intercept point).

DISTORTION - BASIC CONCEPTS

(Signal) distortion has been defined [1] as an "undesired change in input signal wave form as the signal passes through an electrical network". The principal sources of waveform distortion are:

A. Non-linear relationship between the input and output ports of the network.

B. Nonuniform transmission at different frequencies in the transmission band.

C. Phase shift not proportional to frequency.

Source A degrades the waveform of a signal traversing a specific electrical network. Sources B & C can degrade signal waveforms propagating through the RF Channel between the two antennas of a wireless radio network (Chapter 6 and Appendix C).

To clarify the effects of component non-linearity on the signal waveform, consider the unmodulated waveform in Figure 5.1 and the amplitude modulated waveform in Figure 4.1. A purely linear transformation of these signals would mean that only the amplitude of these waveforms would change. A linear gain block would increase the amplitude of the wave, but otherwise, the exact shape of the wave would remain intact. Similarly, a linear attenuator section would decrease the amplitude of the wave, but would not change its shape otherwise. A non-linear device changes the input wave form, thereby introducing frequency components that do not exist in the input wave form.

NON-LINEAR EFFECTS IN SEMICONDUCTOR DEVICES

Semiconductor diodes and transistors are composed of pn-junctions or of Schottky junctions. The "I-V" characteristic (or the junction current as a function of the impressed voltage) of either junction is exponential in nature [2]. These I-V characteristics of these junctions are inherently non-linear, in fact, they rectify an impressed signal.

Rectification [1] is generally defined as the process of converting an alternating current to a unidirectional current. A rectifier conducts current substantially in one direction only. The output waveform is definitely not a faithful replica of the input!

PIN diodes behave linearly and exhibit low distortion over the ranges of input power, frequency, and bias conditions specified by their manufacturer. This is due to the presence of stored charge in the I-region, as discussed in Chapter 1 and Appendix A. But at sufficiently low frequencies and /or high RF power, they will begin to rectify because the stored charge is insufficient to control the large RF current present. In general, non-linear effects in PIN diode circuits can be minimized by choosing the appropriate PIN diode and bias conditions for the application.

Distortion Parameters:

All circuits which contain non-linear elements such as diodes and transistors produce certain kinds of distortion to various degrees. The primary source of distortion in PIN diodes is conductivity modulation of the charge within the I-layer of the diode under forward bias and capacitance modulation of the diode under reverse bias. Distortion can be controlled by the proper choice of diode characteristics and by the design of the switch circuit itself. The three types of distortion which are most often of concern are as follows:

HARMONIC DISTORTION:

This is a single-tone distortion product, resulting when a voltage at a single frequency f, applied to a non-linear device, creates spurious voltages at frequencies 2f, 3f, …, Nf. Of most concern, because they are the closest to the desired signal, are the second and third harmonics. The order of the distortion product is given by the frequency multiplier; for example, the second harmonic is a second order product.

INTERMODULATION DISTORTION:

This is a multi-tone distortion product. It results when two or more signals, of equal or unequal amplitude, mix in a non-linear device to produce unwanted signals whose frequencies are related to these of the original input. The number of signals may exceed 10, and analysis becomes very complex. To keep matters as simple as possible, many semiconductor manufacturers make two-tone measurements using two voltages which are equal in amplitude and closely spaced in frequency. Given two such input signals at frequencies f_1 and f_2, one can compute several significant intermodulation distortion products from the equation

$$Mf_1 \pm Nf_2$$

where M, N = 1, 2, 3, ….

The order of the distortion product is given by the sum M + N.

Of the infinite number of distortion products described by this equation, one is of special significance. The third order products:

$$2f_1\text{-}f_2 \text{ and } 2f_2\text{-}f_1$$

These are important because they exist on either side of the original signals f_1 and f_2 and they cannot be removed by filtering in a narrow band system.

CROSS MODULATION DISTORTION:

This is another form of multi-tone distortion. The non-linear device causes modulation on one signal of frequency f_1 to be transferred to a second signal or carrier of frequency f_2. The intercept point for this distortion product is generally 2.5 dB below that for the third order harmonics [3]. Figure E.1 shows the output of a bridge-T attenuator fabricated with UM9552 devices with the following test conditions:

f_1 = 99 KHz f_2 = 101 KHz
f_2 is AM modulated at 50% with a 400 Hz signal
The attenuation is 23.9 dB and the return loss is 23.9 dB

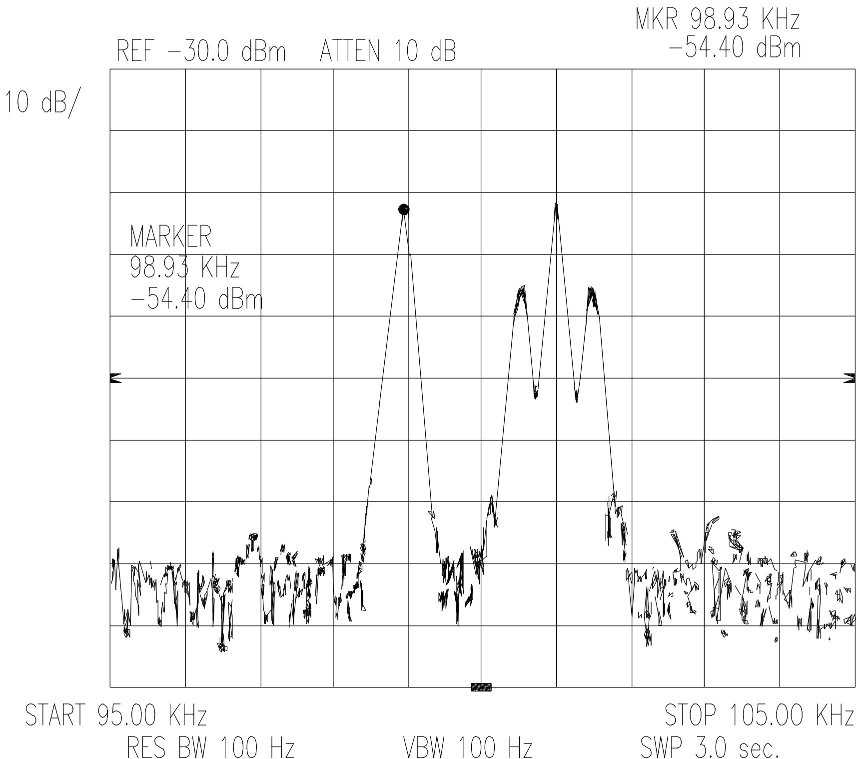


FIGURE E.1 CROSS MODULATION

A theoretical treatment of the effect of non-linearities on multitone signals can be found in reference (4).

Why do we measure third order intermodulation distortion (IMD)?

Two-tone third order intermodulation is a common problem in narrow-band systems. When two (or more) signal are present in a system, strong harmonic components are often generated (See Figure E.1). In cases where two signals are present, the two signals (f_1 and f_2) mix with each other's second harmonic ($2f_1$ and $2f_2$) and create distortion products evenly spaced about the fundamentals ($2f_1$-f_2 and $2f_2$-f_1). Components such as amplifiers, mixers, and filters can generate third order intermodulation products. These distortion products can degrade the performance of many communication systems, such as FM and AM transceivers and high frequency radio teletypes. For example, signals transmitted with excessive third order IMD can interfere with other transmissions. Receivers must also be distortion-free, especially in the preamplifier stages, to preventcrosstalk between adjacent channels.

A good example is using the IS-54 standard for North American Digital Cellular.

Tx = 824 - 849 MHz
Rx = 869 - 894 MHz

For third order IMD, assume two transmitters

f_1 = 824 MHz f_2 = 849 MHz

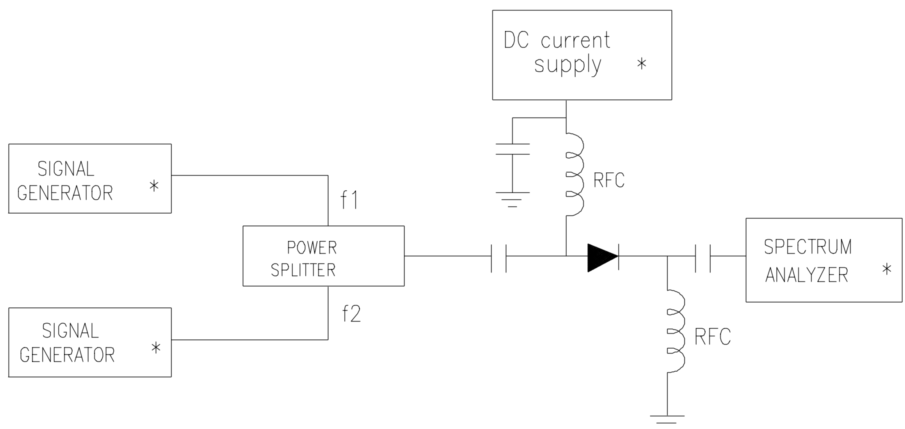
Then $2f_1 - f_2$ = 799 MHz
$2f_2 - f_1$ = 874 MHz this in the middle of the receive band !

This is the reason why component engineers are wanting to specify switching devices with IMD products that are >80 dB below the carrier level.

DISTORTION TEST PROCEEDURE AND SETUP

Distortion is by the two tone Third Order Intermodulation Distortion method. The test setup for the IM3 measurements is show in Figure E.2. This setup was adapted from Hewlett Packard's Product Note 8566B/8568B-1. [Third Order Intermodulation Distortion Measurements. A Downloadable Procedure for HP 8566B and 8568B Spectrum Analyzers]. Our setup is basically the same as HP's without the low pass filters. We use only low power generators but with very low noise and avoid the problem of noisy power amplifiers. The two signal generators are HP 8656B. We have standardized on setting the frequency of the two tones at ✂ 1 KHz of the required frequency. This makes it next to impossible to filter some of the noise the you are trying to measure. It also allows the spectrum analyzer to display the two IM3s and the two tones together.

The two tones, f_1 and f_2, are combined with a HP 11667A power splitter and feed directly to the PIN under test. The PIN is biased [forward or reverse] by inserting two bias tees. If the PIN has any nonlinearity this will generate distortion products. In cases where two signals are present, the two signals (f_1 and f_2) mix with each other's second harmonic ($2f_1$ and $2f_2$) and create distortion products evenly spaced about the fundamental ($2f_1$-f_2 and $2f_2$-f_1). The fundamentals and the IMD products are then displayed on the spectrum analyzer.



- May be controlled with the IEEE-488 BUS CIRCUIT

FIGURE E.2

APPENDIX - F

PIN DIODE RADIATION DETECTORS

NOTES

THE SILICON PIN DIODE RADIATION DETECTORS: UM9441 & UM9442

INTRODUCTION

Silicon PIN diodes have been used as detectors of nuclear radiation for many years [1]. They are more efficient detectors than pn - junctions, and offer good sensitivity to various forms of radiation. They are not restricted to use at low temperature operation but are effective detectors across the entire operating temperature range of -55 ^{o}C to +150 ^{o}C. The PIN diode detects radiation by generating a transient photo current during the time that the radiation pulse is absorbed by the diode's I-layer.

MECHANISM BY WHICH SILICON PIN DIODES DETECT RADIATION

A detailed discussion of the various effects nuclear radiation can have on semiconductor materials is clearly beyond the scope of this brief article. We are concerned here with the mechanism by which a PIN diode detects certain forms of nuclear radiation.

In a heuristic sense, nuclear radiation is the effluence from a nuclear radiation source that consists of energy in the form of particles or waves. The waves are represented as photons or quanta of radiant energy. A radiation detector may be exposed to neutrons (particles), gamma rays, X-rays, or other radiation energy. The radiation can vary in energy density, wavelength, and duration.

When semiconductor material is exposed to nuclear radiation, the energy quanta (Er) is absorbed by the elements of the lattice structure if Er equal to or greater than the band gap energy (Eg) of the material. A hole-electron pair is generated for each quantum absorbed. For Silicon, Eg=1.1eV, which corresponds to a photon of 1.1 micron wavelength [1,2]. Lower energy quanta (or longer wavelengths) are either transmitted through the material or interact with free electrons or lattice site atoms, increasing their steady-state energies.

The predominant reactions of high energy quanta with Silicon are inelastic collisions with electrons and elastic collisions with lattice site atoms. In the former case, the electrons absorb sufficient energy (1.1 eV) to transit from the valence band to the conduction band, creating a pair of carriers (transient effect) that will survive for a time referred to as the recombination time or minority carrier lifetime. Bulk recombination time is typically several microseconds in Silicon PIN diodes. If Er > Eg, secondary ionization effects can occur. If Er is sufficiently large (> 15 eV), lattice site reaction can eject silicon atoms from their equilibrium sites creating permanent damage to the lattice . The presence of these defects after exposure to nuclear radiation adds to the number of carrier recombination sites within the lattice and permanently decreases the minority carrier lifetime of the device. Lifetime can be reduced by as much as a factor of 10, depending on the resistivity of the Silicon and the intensity of the radiation. The device can still function as a radiation detector because lifetime is not a significant parameter when the PIN diode is fully reverse biased.

SILICON PIN DIODES

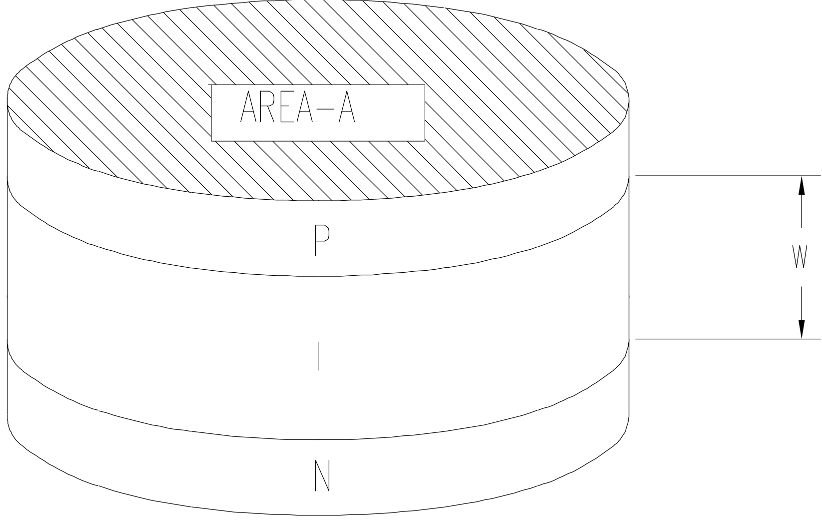


Figure F.1. PIN Diode Chip Structure

Figure F.1 shows the PIN diode chip structure. The high resistivity intrinsic layer is sandwiched between N+ and P+ layers. The PIN diode detects the presence of radiation by the mechanism discussed above. During gamma and/or X-ray irradiation, electron - hole pairs are generated in the silicon.

The PIN diode is reverse biased so that the entire I-layer is swept out (depleted of free carriers). This reverse bias creates an electric force field across the I-layer so that the electrons are swept to the P+ layer and the holes, to the N+ layer. This flow of carriers in response to a pulse of radiation, constitutes a photocurrent that can be measured. To maximize this photocurrent, the I-region must be as large as possible.

In summary, the PIN diode structure possesses the following advantages over pn-junction devices:

Detector Sensitivity (mA of photocurrent per Rad) and photocurrent measurement accuracy are enhanced by the PIN diode structure.

Lot to lot reproducibility of PIN Diode Radiation Detector characteristics is greatly enhanced by Microsemi's PIN diode structure that uses high resistivity Silicon I-layer material.

THE ROLE OF DEVICE PACKAGING

Microsemi PIN Diode Radiation Detectors are available in a glass axial leaded UM9441 and in the TO-39 UM9442. Device packaging also plays a role in the overall performance of a radiation detector. In practice, not all of the radiation pulse is absorbed by the semiconductor material. Some of the radiation is absorbed by the package and other components of the structure. These materials generally emit a spectrum of secondary energies and particles that can be absorbed in the semiconductor material and this gives rise to a photocurrent enhancement effect. The enhancement effect for a particular package can be determined theoretically [1] but is usually determined empirically in practice.

THE UM9441 & UM9442 RADIATION DETECTOR

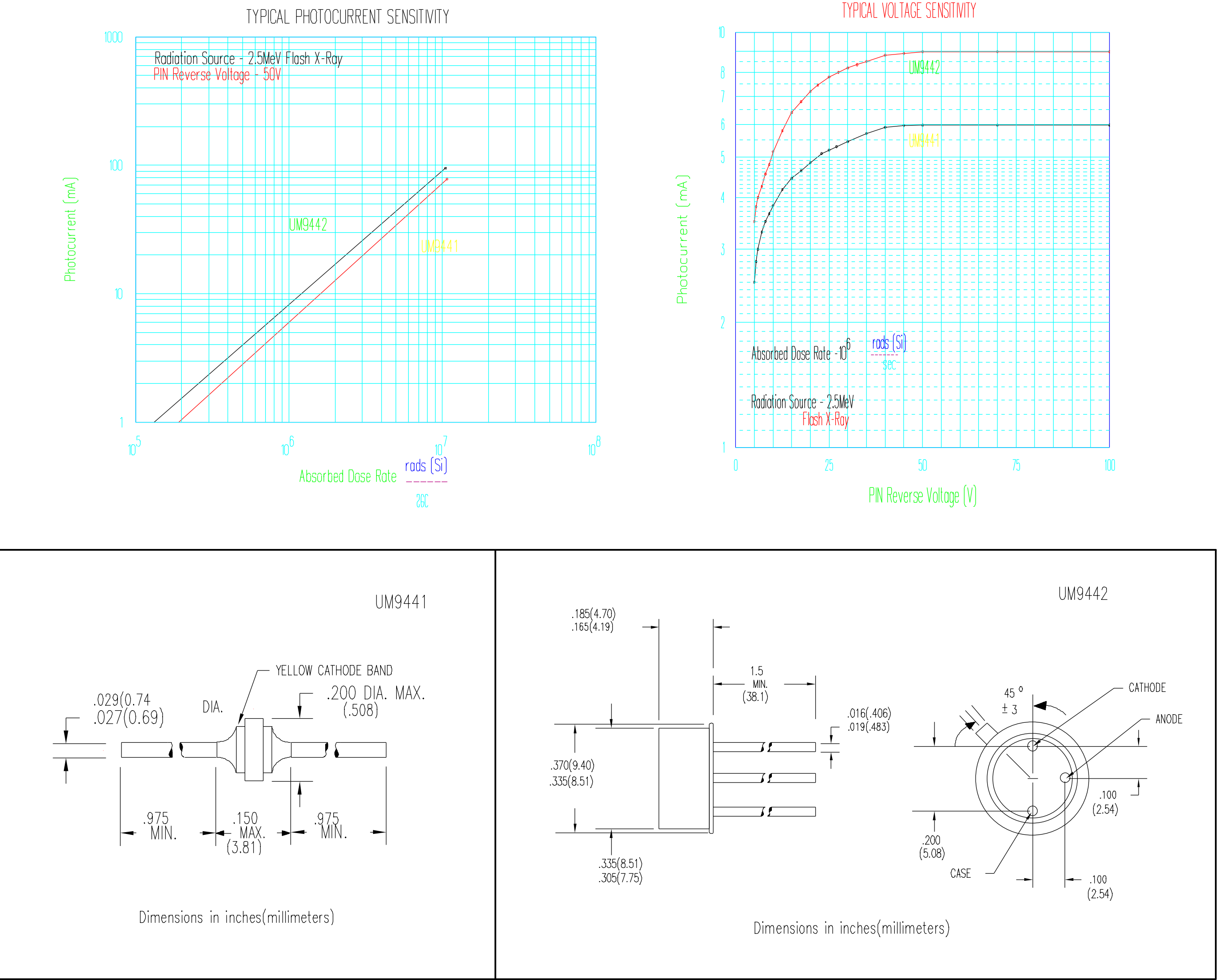
Microsemi Silicon PIN diodes are efficient detectors of nuclear and electromagnetic radiation, including gamma radiation, electrons, and X-rays. The Microsemi UM9441 & UM9442 series utilize high resistivity Silicon and are designed to have uniform area mesa structures to define the active volume of the detectors. Their current sensitivity is proportional to only the I-region volume and is independent of the device temperature as long as the reverse bias voltage exceeds the saturation voltage.

This diode structure minimizes the effects of permanent damage caused by neutrons and other high energy radiation and therefore can withstand high radiation levels. Experiments on UM9441 radiation detectors show that these devices exhibit no degradation in gamma radiation sensitivity when exposed to a total dose of 10 x E 14 neutrons/ cm^2 or a 1 MeV equivalent.

The UM9441 is an axially leaded device constructed by metalurgically bonding the PIN chip between two molybdenum refractory pins that are typically 0.125 inches in diameter and 0.050 inches long. Hyper-pure glass is then fused over this bond to form a voidless seal. Leads are then brazed to the ends of the molybdenum pins. This results in a high reliability HERMETIC package that uses materials that are so well matched thermally, that the UM9441 can withstand temperature shock or cycling from -196 C to + 300 C.
The UM9441 & UM9442 series of radiation detectors can be obtained with compliance to full military high reliability testing. Usually, these devices are procured to a customer's specification control drawing. However, they can be procured as catalog devices with Microsemi's HR2 level screening.

ELECTRICAL SPECIFICATIONS (at 25°C)

	UM9441			UM9442				
Test	Min	Typ	Max	Min	Typ	Max	Units	Test Conditions
Photocurrent	4.0	6.0		6.0	9.0		mA	$V_R = 50V$ $10^6 \frac{rads(Si)}{sec.}$
Photocurrent Rise Time (10%-90%)		10			10		ns	2.5 MeV Flash X-Ray Ion Physics Corp. FX-25
Capacitance		10	15		15	20	pF	F = 1 MHz V = 50V
Reverse Current			1.0			1.0	µA	$V_R = 50V$
Minority Carrier Lifetime	2.0			2.0			µs	$I_f = 10mA$



Figures and features are from UM9441 & UM9442 data sheet

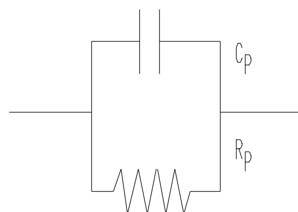
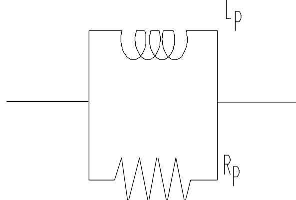
APPLICATION	**RECOMMENDED PIN DIODE TYPE**
RADIATION DETECTORS	UM9441, UM9442 CUSTOM DEVICES

APPENDIX - G

MISCELLANEOUS FORMULAE AND DATA

NOTES

MEASUREMENT PARAMETER FORMULAS

	CONVERSION TO OTHER MODES	Z	D	Q	θ
C_S R_S	$C_P = \frac{1}{1+D^2} C_S$ $R_P = \frac{1+D^2}{D^2} R_S$	$\sqrt{\frac{1}{w^2 C_S^2} + R_S^2}$	❄$C_S\ R_S$	$\frac{1}{w C_S\ R_S}$	-TAN$^{-1}\left(\frac{1}{w C_S\ R_S}\right)$
	$C_S = \left(1+D^2\right) C_P$ $R_S = \frac{D^2}{1+D^2} R_P$	$\sqrt{w^2 C_P^2 + \frac{1}{R_P^2}}$	$\frac{1}{w C_P R_P}$	❄$C_P R_P$	-TAN$^{-1}\left(w C_P R_P\right)$
L_S R_S	$L_P = \left(1+D^2\right) L_S$ $R_P = \frac{1+D^2}{D^2} R_S$	$\sqrt{w^2 L_P^2 + R_S^2}$	$\frac{R_S}{w L_S}$	$\frac{w L_S}{R_S}$	TAN$^{-1}\left(\frac{w L_S}{R_S}\right)$
	$L_S = \frac{1}{1+D^2} L_P$ $R_S = \frac{D^2}{1+D^2} R_P$	$\sqrt{1 + \frac{w^2 L_P^2}{R_P^2}}$	$\frac{w L_S}{R_P}$	$\frac{R_P}{w L_P}$	TAN$^{-1}\left(\frac{R_P}{w L_P}\right)$

$$Z = R + jX = |Z| \left(COS\, q + j\, SIN\, q \right), |Z| = \sqrt{R^2 + X^2}$$

Technical Reference Material

Control Circuits

Important parameters to describe the circuit performance of Control Circuit Elements

Insertion Loss (IL) - Switch is in the "ON" Condition:

Insertion Loss is defined as the ratio

$$IL(dB) = 10\log_{10}\frac{(PL_0)}{(PL_1)}$$

PL = Power delivered to the load with the switch not in the RF circuit.

PL_1 = Power delivered to the load with the switch in the RF current path ("ON Condition").

Isolation (ISO) - Switch is in the "OFF" Condition:

$$ISO(dB) = 10\log_{10}\frac{(PL_0)}{(PL_1)}$$

PL_1 = As above.

PL_2 = Power delivered to the load with the switch in the RF current path ("OFF" Condition).

VSWR (Voltage Standing Wave Ratio)

The SPST switch is a two part network. VSWR at the input port is a measure of the mismatch between the RF Signal Source's internal impedance and the input impedance of the switch. VSWR (input) depends on the switch's load impedance but usually:

$Z_{load} = Z_{10}$
(Characteristic Impedance of the Transmission Medium)

VSWR at the output port of the switch is the measure of impedance mismatch between the switch's load and the switch's output impedance.

The switch as a Reciprocal Circuit Element. If the RF Signal Source's internal impedance and the load impedance equal the transmission line impedance and the switch is a reciprocal network element then:

VSWR (input) = VSWR (output)

VSWR is related to the magnitude of the voltage reflection coefficient.

$$VSWR = \frac{1+|r|}{1-|r|} \qquad |r| = \frac{P_{-1}}{P_R}^{1/2}$$

P_1 = Power incident on the switch

P_R = Power reflected from the switch

Return Loss (RL) is often used to describe the magnitude of transmission line mismatch. Thus RL is defined to be:

$$RL(dB) = 20\log_{10}\frac{1}{|r|}$$

Transmission Loss (TL) describes the transmission properties of the network element and we expect that:

$|r| + |\tau| = 1$

because energy is conserved in the network.

$$TL = 20\log_{10}\frac{1}{|t|}$$

$$|t| = \frac{P_{-1}}{P_T}$$

PT = Power transmitted through the switch to the load.

A conversion chart of VSWR and Return Loss is given. Thus a Return Loss of 20dB represents a VSWR of approximately 1.22.

Switching Equations for Shunt and Series Elements

Insertion Loss (I_L) and Isolation (I_{SO}) for Shunt and Series Switching Impedances are represented by the following equations:

Shunt: $I_L = 10 \log [1 + (R_R/Z_0)]$
$I_{SO} = 10 \log [1 + (Z_0/2R_S]^2$

Series: $I_L = 10 \log [1 + (R_S/2Z_0)]^2$
$I_{SO} = 10 \log [1 + (X_C/2Z_0]^2$

Shunt: $R_R = Z_0$ ([Antilog $(I_L/10)$]-1)
$R_S = Z_0/2$ [{Antilog $(I_{SO}/20)$}-1]$^{-1}$

Series: $R_S = 2Z_0$ [{Antilog $(I_L/20)$ -1 }]
$X_C = 2Z_0$[{Antilog$(I_{SO})/20)$ -1 }]

VSWR, RETURN LOSS, AND TRANSMISSION LOSS VS TRANSMITTED POWER

VSWR	RETURN LOSS (dB)	TRANS. LOSS (dB)	VOLT REFL COEFF.	POWER TRANS. (%)	POWER REFL (%)	VSWR	RETURN LOSS (dB)	TRANS. LOSS (dB)	VOLT REFL COEFF.	POW TRAI (%
1.00		0.000	0.00	100.0	0.0	**1.64**	12.3	0.263	0.24	94.
1.01	46.1	0.000	0.00	100.0	0.0	**1.66**	12.1	0.276	0.25	93.
1.02	40.1	0.000	0.01	100.0	0.0	**1.68**	11.9	0.289	0.25	93.
1.03	36.6	0.001	0.01	100.0	0.0	**1.70**	11.7	0.302	0.26	93.
1.04	34.2	0.002	0.02	100.0	0.0	**1.72**	11.5	0.315	0.26	93.
1.05	32.3	0.003	0.02	99.9	0.1	**1.74**	11.4	0.329	0.27	92.
1.06	30.7	0.004	0.03	99.9	0.1	**1.76**	11.2	0.342	0.28	92.
1.07	29.4	0.005	0.03	99.9	0.1	**1.78**	11.0	0.356	0.28	92.
1.08	28.3	0.006	0.04	99.9	0.1	**1.80**	10.9	0.370	0.29	91.
1.09	27.3	0.008	0.04	99.8	0.2	**1.82**	10.7	0.384	0.29	91.
1.10	26.4	0.010	0.05	99.8	0.2	**1.84**	10.6	0.398	0.30	91.
1.11	25.7	0.012	0.05	99.7	0.3	**1.86**	10.4	0.412	0.30	91.
1.12	24.9	0.014	0.06	99.7	0.3	**1.88**	10.3	0.426	0.31	90.
1.13	24.3	0.016	0.06	99.6	0.4	**1.90**	10.2	0.440	0.31	90.
1.14	23.7	0.019	0.07	99.6	0.4	**1.92**	10.0	0.454	0.32	90.
1.15	23.1	0.021	0.07	99.5	0.5	**1.94**	9.9	0.468	0.32	89.
1.16	22.6	0.024	0.07	99.5	0.5	**1.96**	9.8	0.483	0.32	89.
1.17	22.1	0.027	0.08	99.4	0.6	**1.98**	9.7	0.497	0.33	89.
1.18	21.7	0.030	0.08	99.3	0.7	**2.00**	9.5	0.512	0.33	88.
1.19	21.2	0.033	0.09	99.2	0.8	**2.50**	7.4	0.881	0.43	81.
1.20	20.8	0.036	0.09	99.2	0.8	**3.00**	6.0	1.249	0.50	75.
1.21	20.4	0.039	0.10	99.1	0.9	**3.50**	5.1	1.603	0.56	69.
1.22	20.1	0.043	0.10	99.0	1.0	**4.00**	4.4	1.938	0.60	64.
1.23	19.7	0.046	0.10	98.9	1.1	**4.50**	3.9	2.255	0.64	59.
1.24	19.4	0.050	0.11	98.9	1.1	**5.00**	3.5	2.553	0.67	55.
1.25	19.1	0.054	0.11	98.8	1.2	**5.50**	3.2	2.834	0.69	52.
1.26	18.8	0.058	0.12	98.7	1.3	**6.00**	2.9	3.100	0.71	49.
1.27	18.5	0.062	0.12	98.6	1.4	**6.50**	2.7	3.351	0.73	46.
1.28	18.2	0.066	0.12	98.5	1.5	**7.00**	2.5	3.590	0.75	43.
1.29	17.9	0.070	0.13	98.4	1.6	**7.50**	2.3	3.817	0.76	41.
1.30	17.7	0.075	0.13	98.3	1.7	**8.00**	2.2	4.033	0.78	39.
1.32	17.2	0.083	0.14	98.1	1.9	**8.50**	2.1	4.240	0.79	37.
1.34	15.8	0.093	0.15	97.9	2.1	**9.00**	1.9	4.437	0.80	36.
1.36	16.3	0.102	0.15	97.7	2.3	**9.50**	1.8	4.626	0.81	34.
1.38	15.9	0.112	0.16	97.5	2.5	**10.00**	1.7	4.807	0.82	33.
1.40	15.6	0.122	0.17	97.2	2.8	**11.00**	1.6	5.149	0.83	30.
1.42	15.2	0.133	0.17	97.0	3.0	**12.00**	1.5	5.466	0.85	28.
1.44	14.9	0.144	0.18	96.7	3.3	**13.00**	1.3	5.762	0.86	26.
1.46	14.6	0.155	0.19	96.5	3.5	**14.00**	1.2	6.040	0.87	24.
1.48	14.3	0.166	0.19	96.3	3.7	**15.00**	1.2	6.301	0.88	23.
1.50	14.0	0.177	0.20	96.0	4.0	**16.00**	1.1	6.547	0.88	22.
1.52	13.7	0.189	0.21	95.7	4.3	**17.00**	1.0	6.780	0.89	21.
1.54	13.4	0.201	0.21	95.5	4.5	**18.00**	1.0	7.002	0.89	19.
1.56	13.2	0.213	0.22	95.2	4.8	**19.00**	0.9	7.212	0.90	19.
1.58	13.0	0.225	0.22	94.9	5.1	**20.00**	0.9	7.413	0.90	18.
1.60	12.7	0.238	0.23	94.7	5.3	**25.00**	0.7	8.299	0.92	14.
1.62	12.5	0.250	0.24	94.4	5.6	**30.00**	0.6	9.035	0.94	12.

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POWER CONVERSION dBm TO WATTS

dBm	picoWatts	dBm	nanoWatts	dBm	microWatts	dBm	microWatts	dBm	mil
-80	10.00	-60	1.00	-40	0.100	-20	10.00	0	
-79	12.59	-59	1.26	-39	0.126	-19	12.59	1	
-78	15.85	-58	1.58	-38	0.158	-18	15.85	2	
-77	19.95	-57	2.00	-37	0.200	-17	19.95	3	
-76	25.12	-56	2.51	-36	0.251	-16	25.12	4	
-75	31.62	-55	3.16	-35	0.316	-15	31.62	5	
-74	39.81	-54	3.98	-34	0.398	-14	39.81	6	
-73	50.12	-53	5.01	-33	0.501	-13	50.12	7	
-72	63.10	-52	6.31	-32	0.631	-12	63.10	8	
-71	79.43	-51	7.94	-31	0.794	-11	79.43	9	
-70	100.00	-50	10.00	-30	1.000	-10	100.00	10	1
-69	125.89	-49	12.59	-29	1.259	-9	125.89	11	1
-68	158.49	-48	15.85	-28	1.585	-8	158.49	12	1
-67	199.53	-47	19.95	-27	1.995	-7	199.53	13	1
-66	251.19	-46	25.12	-26	2.512	-6	251.19	14	2
-65	316.23	-45	31.62	-25	3.162	-5	316.23	15	3
-64	398.11	-44	39.81	-24	3.981	-4	398.11	16	3
-63	501.19	-43	50.12	-23	5.012	-3	501.19	17	5
-62	630.96	-42	63.10	-22	6.310	-2	630.96	18	6
-61	794.33	-41	79.43	-21	7.943	-1	794.33	19	7

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APPENDIX - H

SURFACE MOUNT CRITERIA

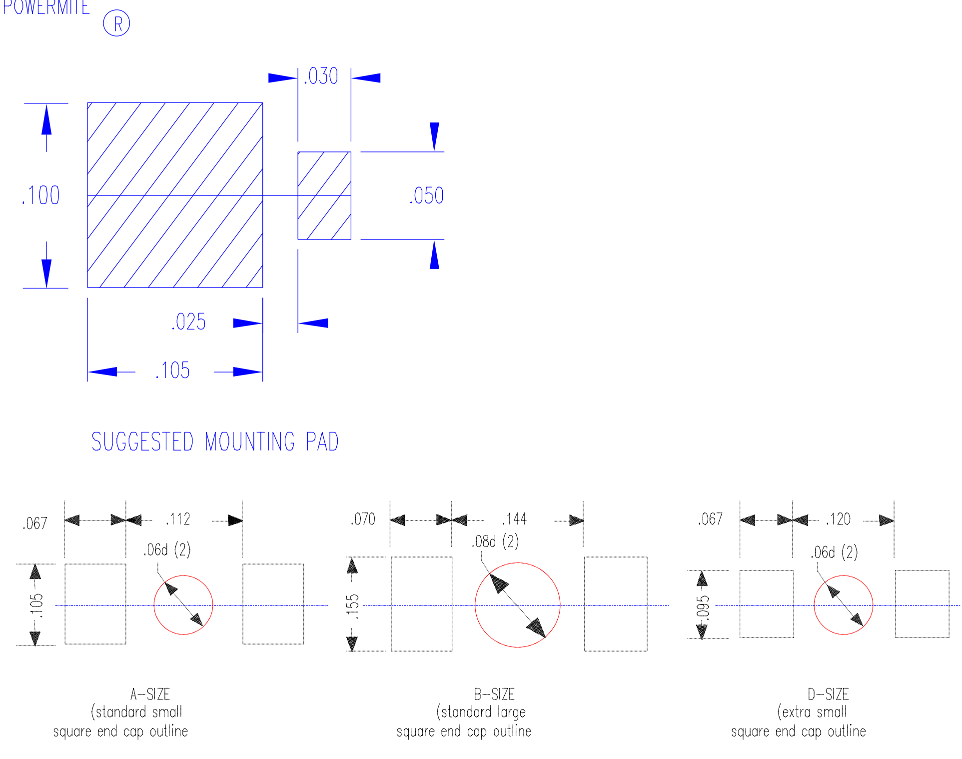
NOTES

SURFACE MOUNT DEVICE CIRCUIT BOARD MOUNTING

Surface mount technology has allowed a great reduction in the size of circuit designs. Many manufacturers are taking advantage of this technology due to the immense availability of SMT devices in the market place today. MSC offers a variety of surface mount packages and semiconductor devices for your designs. The MELF, ceramic F series, and the POWERMITE® packages all allow power diode applications to be surface mounted.

Common surface mounting processes used in the industry are accomplished by using convection or infra red belt furnace, or wave solder equipment. Microsemi surface mount devices may be mounted with any of these methods with one exception. The POWERMITE®, due to its design geometry, may not be mounted with the wave soldering technique, unless the wave is under the board and the POWERMITE® is on top of the board. All other packages may be affixed temporarily to the circuit board with fast curing adhesive systems to accommodate the wave solder technique.

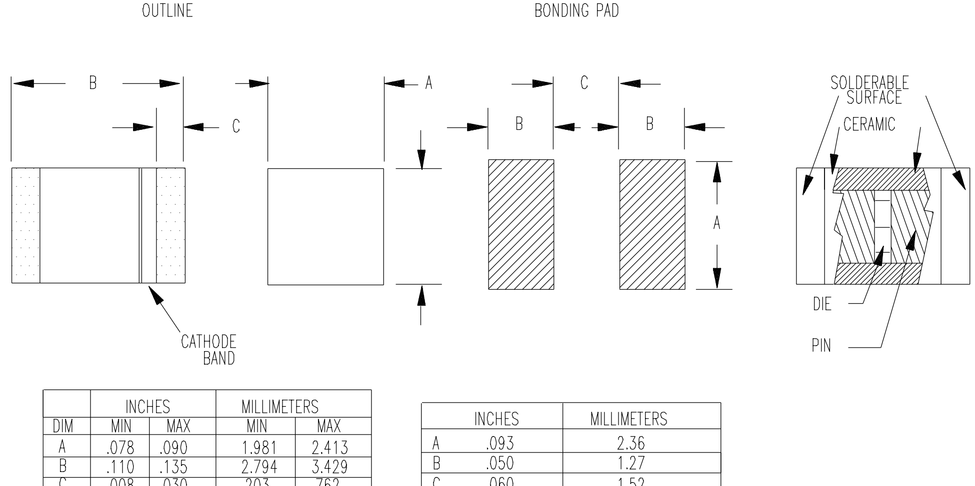
Suggested circuit board trace outlines for MSC surface mount devices are shown below:



NOTES:

1. These dimensions will match the terminals and provide for additional solder fillets at the outboard ends at least as wide as the terminals themselves, assuming accuracy of device placement within .005 inches.
2. If the mounting method chosen requires use of an adhesive separate from the solder compound, a round (or square) spot of cement as shown should be centrally located.
3. Dimensions shown are in inches.

CERAMIC F



DIM	INCHES MIN	INCHES MAX	MILLIMETERS MIN	MILLIMETERS MAX
A	.078	.090	1.981	2.413
B	.110	.135	2.794	3.429
C	.008	.030	.203	.762

	INCHES	MILLIMETERS
A	.093	2.36
B	.050	1.27
C	.060	1.52

The most common solder systems used for attachment to boards are 63/37% Sn/Pb, 60/40% Sn/Pb, and 62/36/2% Sn/Pb/Ag. Flux systems vary, however, RMA flux and water soluble systems are in wide use. The RMA flux requires solvent cleaning processes. Water soluble flux systems meet environmental demands imposed on manufacturers.

Usual techniques used to dispense solder pastes are screen printed through stencils, and commercial solder dispensing systems. The tack which is present in most solder pastes is of great assistance in holding devices to the circuit board prior to reflowing the solder.

The reflow solder process is key to a successful surface mounting operation. The temperature profile parameters suggested to properly solder MSC surface mount devices are shown below. Other profiles may be used to accommodate higher mass boards or when using heat sensitive devices. These must be developed by the user for each case.

Temperature (^{o}C)	>200	>220	>240
Maximum Time (min.)	2.5	1	.33

Temperature Rate of Raise   33 ^{o}C/min. (maximum)

Expected Total Profile Time   12 to 20 minutes

Suggested Atmosphere   Nitrogen, Forming gas, (Air is used by some manufacturers, however, flux charring can result.)

APPENDIX - I

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NOTES

REFERENCES

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APPENDIX C

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