Applications of PIN Diodes

Application Note 922

Introduction

The most important property of the PIN diode is the fact that it can, under certain circumstances, behave as an almost pure resistance at RF frequencies, with a resistance value that can be varied over a range of approximately 1 Ω to 10 KΩ through the use of a DC or low frequency control current.

When the control current is varied continuously, the PIN diode is useful for leveling and amplitude modulating an RF signal. When the control current is switched "on" and "off" or in discrete steps, the device is useful for switching, pulse modulating, attenuating, and phase shifting of an RF signal.

In addition, the PIN's small size, weight, high switching speed, and minimized parasitic elements make it ideally suited for use in miniature, broadband RF signal control components.

This application note describes the important properties of the PIN diode and illustrates how it can be applied in a variety of RF control circuits. Related HP publications of special interest are:

1. AN 929, Fast Switching PIN Diodes.
2. AN 985, Achieve High Isolation in Series Applications with the Low Capacitance HPND-4005 Beam Lead PIN.
3. AN 957-1, Broadbanding the Shunt PIN Diode SPDT Switch.
4. AN 957-2, Reducing the Insertion Loss of a Shunt PIN Diode.
5. AN 957-3 Rectification Effects in PIN Attenuators.
6. AN 979, Handling and Bonding of Beam Lead Devices.

Characteristics of the PIN Diode

A PIN diode is a silicon semiconductor consisting of a layer of intrinsic (high resistivity) material of finite area and thickness which is contained between highly doped p and n type material. When the diode is forward biased, charge is injected into the intrinsic or "I" region. This charge consists of holes and electrons which have a finite lifetime before recombination. The density of charge in the intrinsic region and its geometry determines the conductance of the device, while the lifetime [denoted by τ (tau)] determines the approximate low frequency limit of useful application.

The conductance of the diode is proportional to the stored charge and the charge is in turn related to the diode current by

$$I_d = \frac{dQ_d}{dt} + \frac{Q_d}{\tau} \qquad (1)$$

where I_d = Diode current
Q_d = Charge stored in the diode
τ = Recombination lifetime
d = delta

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

$$Q_d = I_d \tau \qquad (2)$$

If the bias consists of both a constant current and a low frequency RF or time varying signal, then the dc component of stored charge will be "modulated" by the presence of an ac component. The degree of modulation depends on the relative level of the two charge components and the frequency of the RF signal.

This dependence on frequency can be readily seen by solving the Laplace transform of Eq. (1) which yields

$$Q_d(\omega) = \frac{i_d\tau(j\omega)}{1 + j\omega\tau} \qquad (3)$$

where $(\omega) = 2\pi$ freq.

This is plotted in Figure 1(a) and iillustrates that at signal frequencies below $f_c = 1/2\pi\tau$ the RF signal has about the same effect as the dc bias. Above f_c however, the modulation effect decreases by about 6 dB/octave.

The lifetime of PIN diodes is determined by design and is usually based on the desired switching speed. Typically, τ can be in the range of 0.005 µsec to over 3 µsec. For a value of 100 nsec, f_c is ≅ 1.6 MHz. Thus, the diode can be simply visualized as follows: At frequencies well below f_c, the PIN diode behaves as an ordinary PN junction diode. The RF signal incident on the diode will be rectified and considerable distortion of the signal will occur. In the vicinity of f_c, the diode begins to behave as a linear resistor with a small nonlinear component. The signal consequently suffers some degree of distortion. At frequencies well above f_c, the diode appears essentially as a pure linear resistance whose value can be controlled by a dc or a low frequency control signal.

a) Low and High Frequency Equivalent Circuits

Because of this behavior, the equivalent circuit of the PIN diode also depends on the frequency. At frequencies much less than f_c, the equivalent circuit is as shown in Figure 2(a) and is that of a normal PN junction diode.

In this circuit

L_P = Package Inductance
C_p = Package Capacitance
R_s = Series Resistance
R_j = Junction Resistance $\cong \frac{nkT}{qI_{dc}}$

For a typical n = 1.8 and at room temperature

$$R_j \cong \frac{48}{I_{dc(mA)}}$$

I_{dc} = Forward dc Bias Current
$C_{j(v)}$ = Junction Capacitance = a function of applied voltage

At frequencies just below and just above f_c, the equivalent circuit of a PIN diode depends upon the way in which the device was designed. It can reflect behavior which is strongly inductive or strongly capacitive. In addition, operation at moderate bias levels in this frequency range will result in the generation of large amounts of distortion.

At frequencies much higher than f_c, the equivalent circuit is as shown in Figure 2(b). Here the elements L_p, C_p and R_s are the same as in Figure 2(a). The element C_I, represents the I-layer capacitance which is constant and dependent only on the geometry of the I-layer. This capacitance can be measured by conventional bridge techniques and at a low (usually 1 MHz) frequency if the diode is reverse biased past the punch-through voltage to assure that the I-layer is fully depleted. Typical values of C_I for present HP PIN diodes are in the range of 0.02 to 2 pF, depending on the diode design. The element R_I represents the *effective* RF resistance of the I-layer. Although shown as variable, *this resistance is constant with respect to the RF signal*, providing the

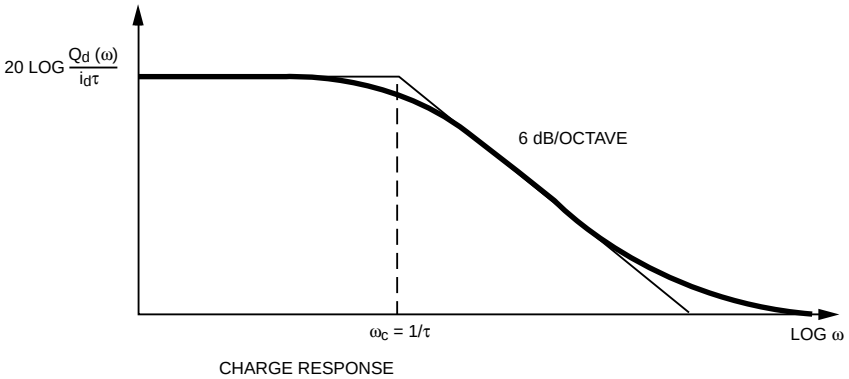


Figure 1. PIN Behavior as a function of Frequency.

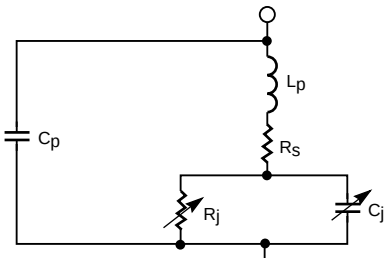


Figure 2(a). PIN Low Frequency Equivalent Circuit.

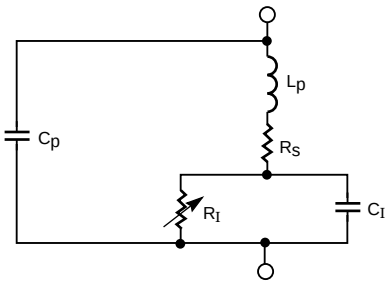


Figure 2(b). PIN High Frequency Equivalent Circuit.

frequency of the signal is much higher than f_c, It is, however, variable by the dc or very low frequency control current.

b) The RF Resistance Characteristic

Although the PIN diode is a two-terminal device, it behaves essentially as a two-port device. With respect to the DC or low frequency control signal, it appears as the circuit shown in Figure 2(a), and with respect to RF signals of frequency > $10f_c$, it appears as the circuit shown in Figure 2(b). The transfer characteristic of this unconventional two-port is governed by the resistance R_I, which can be written as

$$R_I = K/I_{dc}^{x}$$

where R_I is the effective high frequency resistance and I_{dc} is the dc bias current in mA. The dependence of R_I on I_{dc} is similar in form to the dependence of R_j on I_{dc} in the low frequency equivalent circuit; however, the constant K and the exponent x are different. The high frequency resistance function is plotted in Figure 3 for the HPND-4165 diode. Due to a variety of mechanisms that exist in the diode at RF frequencies, both K and x must be determined empirically. For a specific diode design, the exponent x is usually a constant. For the HPND-4165, x is typically 0.92. The constant K and therefore, R_I, however, are highly dependent on the fabrication and process control and its value can vary by as much as 3:1 from diode to diode. For switching or pulse modulating application, the variation of R_I between diodes at a given bias is not significant since the diode is usually switched between a very high and a very low value of resistance by the control current. For analog applications such as attenuating and modulating, and particularly where repeatability and tracking of the attenuation with bias current is desired, this variation of R_I from unit to unit can impose a design and performance limitation. The HPND-4165 is precisely controlled in manufacturing, and resistance values at specific bias points are specified and the slope of resistance vs. bias matched within narrow limits. The specification limits of these parameters are:

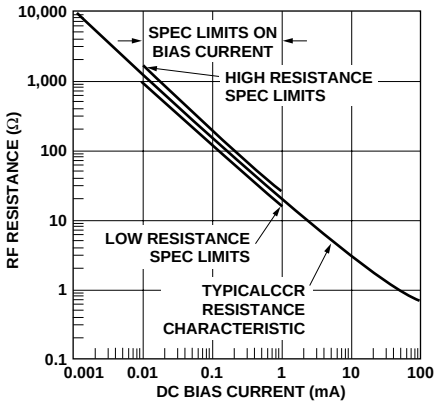


Figure 3. Typical RF Resistance versus Bias Current for HPND-4165.

HPND-4165 PIN Specifications

Parameter	HPND-4165	Test Conditions
High Resistance Limit R_H	1100-1660 Ohms	10 μA
Low Resistance Limit R_L	16-24 Ohms	1 mA
Max Diff. in Resistance vs. Bias Slope Δ x	0.04	10 μA and 1 mA

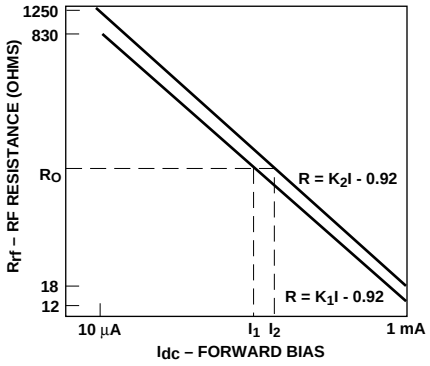


Figure 4. Matching of PIN Diodes by Offset Current.

Both of the resistance limits and the slope are determined by measurement at a test frequency of 100 MHz. An example of these resistance limits is shown graphically in Figure 4. The placement of resistance limits on both extreme values of resistance *and* independently on the slope, assures that the resistance vs. bias curves for individual diodes will be essentially parallel lines lying within the tunnel defined by the

high and low resistance limits. Extremely close tracking between diodes can therefore be achieved by offsetting the bias on one diode with respect to the other. The offset current ratio required to achieve this is given by

$$\frac{I_2}{I_1} \cong \left(\frac{K_2}{K_1}\right)^{\frac{1}{x}}$$

The K values of the individual diodes can be readily determined by measuring the RF resistance of the diodes at a 1 mA dc bias current.

c) Effects of Package Parasitics

Typical HP PIN diode packages are illustrated in Figure 5. With the exception of the Stripline (Style 60) package, all the other packages introduce additional reactive elements in the form of package inductance L_p and package capacitance C_p. These elements are shown in Figure 2 and typical values for some packages are given in Figure 5.

In some applications these parasitic elements can be either "tuned-out" by additional external reactances or actually utilized by forming a resonant circuit around the diode. The bandwidth of such structures is, however, limited. When these elements cannot be tuned out, performance over a broad bandwidth will generally suffer.

As an example, Figure 6 shows the attenuation versus bias characteristics of the same diode in three different packages. In all cases, the device is mounted in shunt across a 50 ohm system. The frequency of operation is 500 MHz and in no case were any package parasitics tuned out. The solid lines are contours of constant attenuation calculated from an assumed equivalent circuit consisting of a parallel resistance R_p and a reactance X_p in shunt across a 50 Ω system. The coordinates of the graph are the resistance R_p and reactance X_p elements. The attenuatlon performance of a particular diode can be easily determined by plotting its equivalent R_p and X_p components at any frequency of interest.

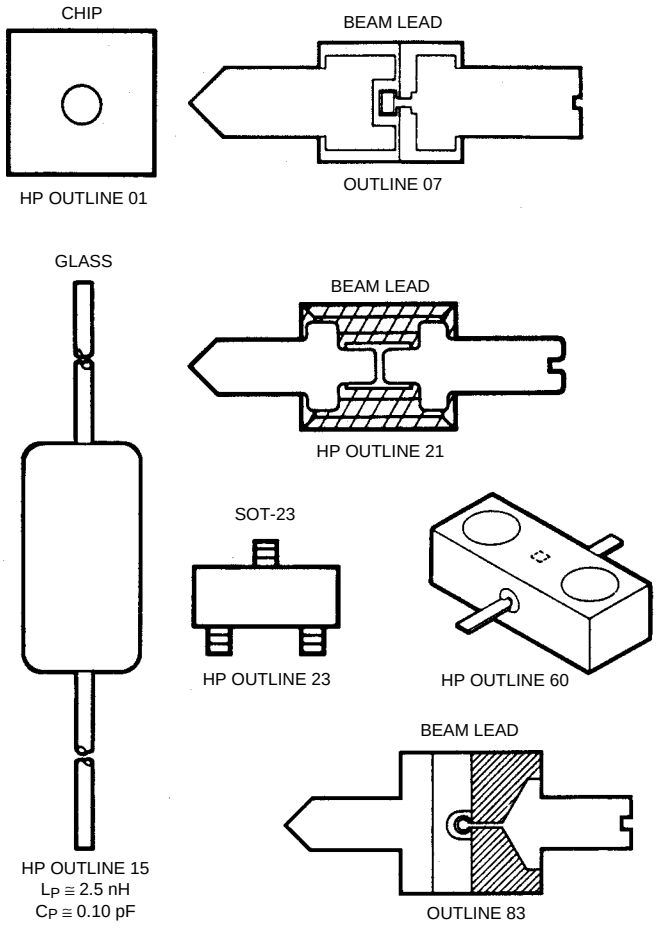


Figure 5. Package Outlines

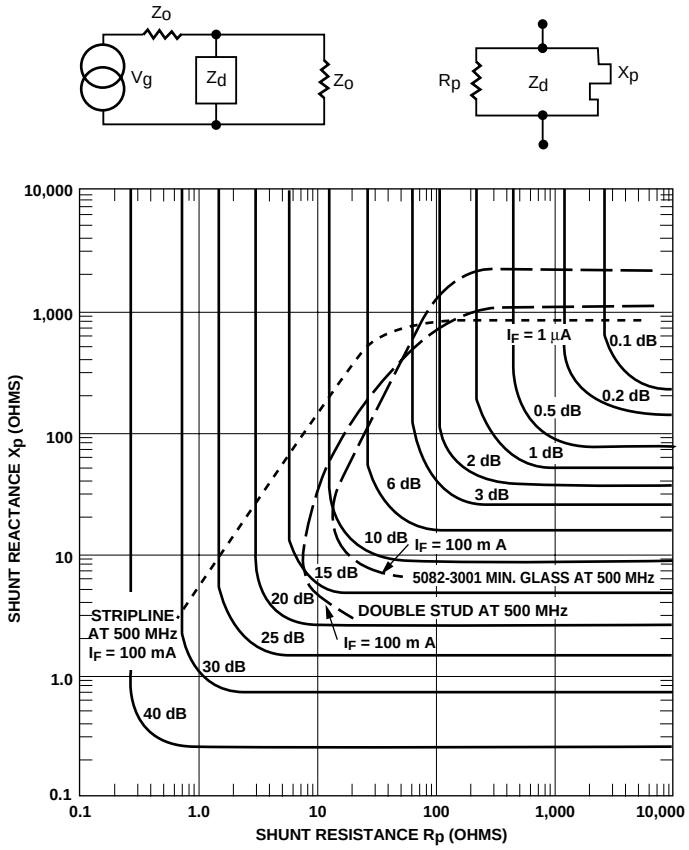


Figure 6. Attenuation Produced by a Shunt PIN Diode as Its Impedance Varies with Changes of Bias Current.

500 MHz is a relatively low frequency, nevertheless differences in the attenuation characteristics of different packaged diodes are evident. At frequencies much higher than 500 MHz, the glass and the double stud diodes will show lower attenuation.

The Stripline package does not exhibit this limitation because the internal reactance elements are proportioned to form a low-pass filter structure with a high (30 GHz) cut-off frequency.

Because of this design, the Stripline PIN diode does not require any external matching elements when used in a 50 ohm Stripline circuit and can be used in the design of extremely wide-band RF control circuits.

d) Distortion in PIN diodes

Two basic methods are used to fabricate PIN diodes. In the case of BULK diodes, a wafer of very pure (intrinsic) silicon is heavily doped on the top and bottom faces to form P and N regions. The result is a diode with a very thick, very pure I region. The epitaxial layer (or EPI) diode starts as a wafer of heavily doped silicon (the P or N layer), onto which a thin I layer is grown. After the epitaxial growth, diffusion is used to add a heavily doped (N or P) layer on the top of the epi, creating a diode with a very thin I layer populated by a relatively large number of imperfections.

These two different methods of design result in two classes of diode with distinctly different characteristics, as shown in the table below:

Parameter	EPI Diode	Bulk Diode
I-layer	Thin	Thick
Lifetime	Short	Long
Distortion	High	Low
Current	Low	High

While manufacturers seldom classify their PIN diodes as bulk or epi devices, one can generally identify them by their lifetime, τ. Lifetime is generally short (35 to 200 nsec.) for epi diodes and relatively long (400 to 3000 nsec.) for bulk diodes.

At frequencies above $10f_c$, the PIN diode acts as if it were a variable resistor. However, it is a non-linear device and, as such, it generates distortion. There are three kinds of distortion; harmonic, intermodulation and crossmodulation. While they are different from each other, they are related. All of them are strongly influenced by the lifetime of a PIN diode, as well as the RF power which is applied to the diode. Long lifetime bulk diodes produce such little distortion that the accurate measurement of it is often very difficult with the best of instruments. Short lifetime epi diodes produce high amounts of distortion; under some bias conditions, such a PIN diode behaves very much like a SRD (Step Recover Diode) comb generator.

PIN diode distortion drops suddenly as bias is set to zero (or reverse bias) or to very high levels of current. Distortion is worst at the intermediate levels of bias where the PIN diode's resistance is some moderate value. Thus, it can be seen that attenuator applications require the use of a long lifetime bulk diode to minimize distortion. In switch applications, where the bias is always at one extreme or the other, epi diodes generally perform satisfactorily and offer the advantage of low current consumption.

In reading the following paragraphs, it should be borne in mind that the bulk diode is always used for attenuator applications and sometimes as a switch, while the epi diode is used as a switching element only.

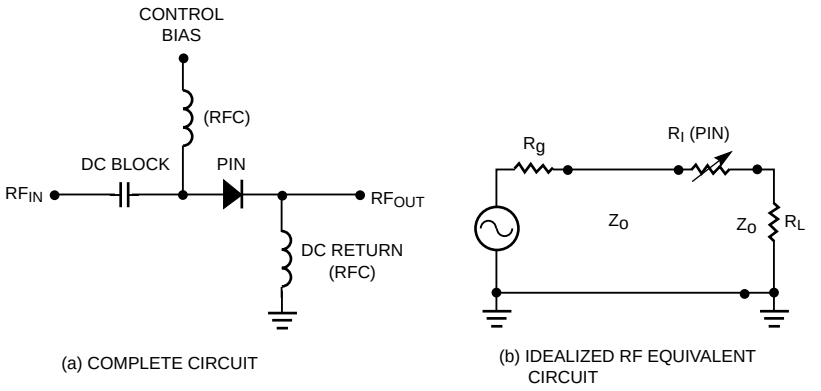


Figure 7. Series PIN RF Attenuator or Switch.

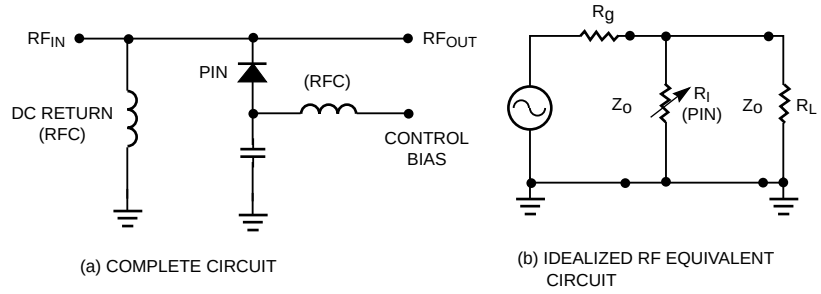


Figure 8. Shunt PIN RF Attenuator or Switch.

PIN Diode Applications

a) Design of Broadband Reflective SPST Switches and Attenuators

The previously described characteristics of the PIN diode make it ideally suited for use in attenuating or switching of RF signals. Two of the simplest circuits that can be used for this are shown in Figures 7 and 8.

The attenuation in the Series PIN circuit is decreased as the RF resistance of the PIN is reduced by increasing the forward current. The opposite occurs for the shunt configuration. If the control bias is switched rapidly between high and low (zero) values, then the circuit acts simply as a switch. When used as a switch, the residual attenuation that exists when the switch is "ON" is usually called Insertion Loss (I.L.). The attenuation provided when the switch is "OFF" is usually called Isolation (I.S.). If the diode is assumed to be a pure resistance at RF, then the attenuation for each circuit is given as:

$$\alpha_{(\text{series})} = 20 \log \left(1 + \frac{R_I}{2Z_o}\right);$$

$$\alpha_{(\text{shunt})} = 20 \log \left(1 + \frac{Z_o}{2R_I}\right);$$

where $Z_o = R_G = R_L$ Circuit, Generator and Load Resistance, respectively.
R_I = PIN diode RF resistance at the specified bias current.

As can be seen, the attenuation is not dependent on frequency and is a function of the ratio of the circuit resistance to the diode resistance. As the bias on the diode is varied, the load resistance as seen by the source also varies; consequently, attenuation is achieved primarily by reflection and partly by dissipation in the PIN diode. Circuits of this type are generally referred to as reflective switches or attenuators.

As was shown previously, a real diode contains several reactance elements due to the diode and the package. Consequently, when a real diode is used, the attenuation characteristic becomes frequency dependent. For an equivalent circuit shown in Figure 9, the attenuations for the two circuits are given by:

$$\alpha'_{(series)} = 10 \log \left[\frac{\left(\frac{R'_S}{Z_o} + 2\right)^2 + \left(\frac{X'_S}{Z_o}\right)^2}{4} \right]$$

$$\alpha'_{(shunt)} = 10 \log \left[\frac{\left(\frac{R'_S Z_o}{R'^2_S + X'^2_S} + 2\right)^2 + \left(\frac{X'_S Z_o}{R'^2_S + X'^2_S}\right)^2}{4} \right]$$

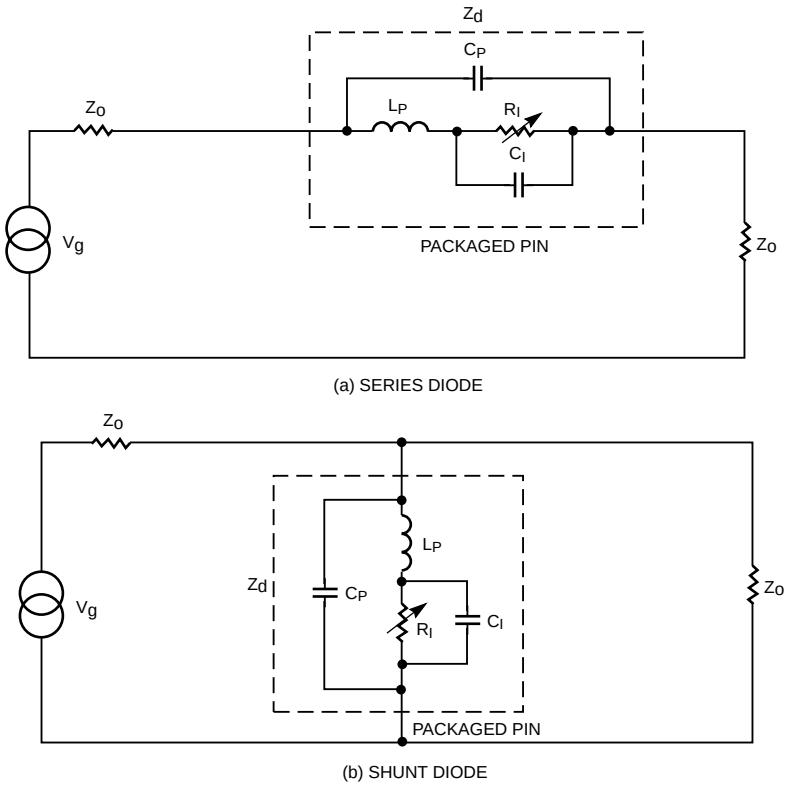


Figure 9. Packaged Diodes Used as Reflective Attenuators.

where R'_s and X'_s are the series equivalent resistance and reactance of the diode impedance, i.e., $Z_d = R'_s + jX'_s$.

These α functions are plotted as a function of frequency in Figures 10(a) and 10(b) which show what values of isolation and insertion loss might be typically obtained for single packaged diodes in either shunt or series reflective attenuators. These curves are based on typical values of diode and package parameters. The upper frequency performance can be improved slightly by tailoring the circuit geometry around the diode package. For example, the high package inductance (~ 3 nH) of the glass packaged diode can be reduced in the series circuit by moving the ground plane or planes of the transmission line closer to

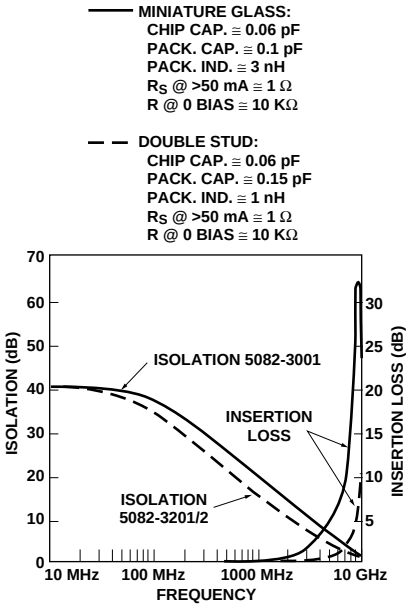


Figure 10(a). Attenuation of Series Diodes Used as a Switch in a 50 Ohm System.

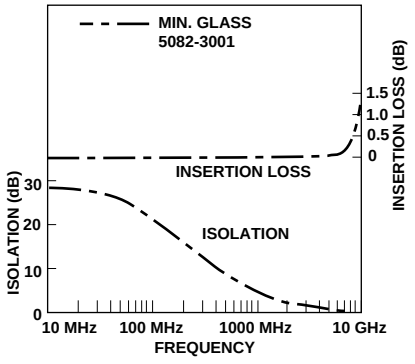


Figure 10(b). Attenuation of Shunt Diodes Used as a Switch in a 50 Ohm System.

the diode package, as shown in Figure 11. This technique reduces the package inductance by making it appear more like a transmission line. It also reduces the series coupling capacitance of the package by cutting the axial electric field across the diode. The former effect decreases the insertion loss and the latter increases the isolation at higher frequencies. The Stripline diode does not begin to exhibit these limitations until approximately 18 GHz because the internal reactances of the package and diode are part of a low pass filter structure. The mini-strip or microstrip package can be made part of a low pass filter by tailoring the lead inductance so the reactance equals the characteristic impedance of the line at the cutoff frequency. This is the frequency where the capacitive reactance of the diode is half the line impedance. The combination of the two leads and the diode forms a three element low pass filter. Other combinations of cutoff frequency and lead inductance may be obtained from tables of filter elements.

b) Design of Resonant SPST Switches

Another way that the performance of a packaged diode can be improved at high frequencies is to "tune-out" the parasitic elements by the addition of external reactances. Such circuits are generally referred to as resonant switches in which the PIN diode is used essentially as a switch which switches the circuit parameters from a parallel resonant condition to a series resonant condition. The high and low impedances produced by the parallel and series resonant circuits, respectively, constitute the "ON" and "OFF" states of the switch. Although the performance can be improved at the design frequency, the bandwidth of resonant switches is necessarily limited, usually to ≤ 10%.

Figure 12 shows two typical resonant switch circuits which are useful at frequencies below 1 GHz, or where the required external inductance L is much greater than the diode parasitic inductance L_p and the external capacitance C is much greater than the package capacitance C_p. Under these conditions, the diode parasitic elements have a negligible effect on performance. The equivalent circuits for the two states of this resonant switch are shown in Figure 12 (c) and (d). When the diode is forward biased, the elements L_1, and C_I are in parallel resonance and the circuit between A and B appears as a high impedance. When the diode is reverse biased, the elements L_2, and C_I are in series resonance and the circuit between A and B appears as a low impedance. In both cases, the impedances between A and B are finite due to the presence of diode resistance and finite element Q's. Both of these effects are represented as loss resistances R_{sp} and R_{ss}, respectively. The expected insertion loss, maximum attenuation, and bandwidth can be obtained from the formulas given in Table 1. For L_1 and L_2 being equal, the resonant frequency in both

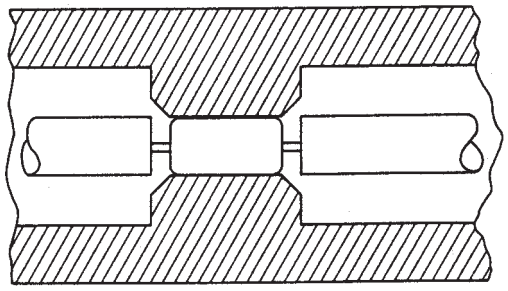


Figure 11. Optimizing a Glass Packaged Diode for Series Reflective Attenuating in a Coaxial Line.

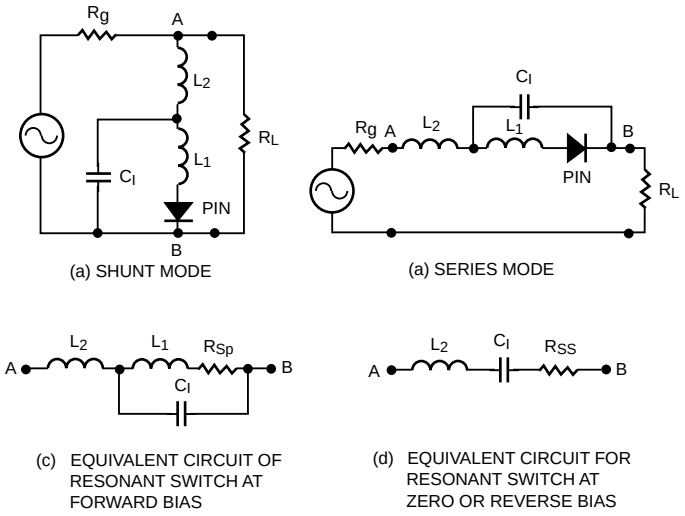


Figure 12. Resonant PIN Switch Circuits.

cases is approximately

$$f_o \cong \frac{1}{2\pi\sqrt{L_1 C_I}}$$

the approximation being due primarily to the fact that diode parasitic elements affect the two resonant states differently. In the parallel resonance state, there will also be a secondary series resonance at approximately $f_2 = 1.4\ f_o$.

At frequencies higher than 1 GHz, the circuits shown in Figure 12 are difficult to design because the values of the external reactances begin to approach those of the diode parasitic reactances and considerable interaction between the various elements occurs. Although this mode is still possible, the synthesis will generally have to be empirical.

Resonant switches can also be built by utilizing the package parasitics and the diode chip capacitance C_I, if these are of the right value, or by tailoring these package parasitics; which is possible with the microstrip package. In this mode of operation, the equivalent circuit of Figure 13 applies. The mode of operation is similar to the one described previously and is evident in Figures 10(a) and 10(b) where it can be observed that as the frequency increases, the insertion and isolation curves approach each other and finally reverse roles as the equivalent circuits approach parallel and series resonance.

The element values are chosen such that L_p and C_I are in series resonance when the diode is reverse or zero biased, and the elements L_p and C_p are in parallel resonance when the diode is forward biased. For maximum isolation and minimum insertion loss, the two resonant frequencies must be equal which dictates that $C_I = C_p$. This, of course, requires a careful choice of the diode and the package for the frequency of interest. In many applications, however, a sufficiently high isolation to insertion loss ratio can be achieved by working on the slopes of two resonant responses of dissimilar frequencies. Consequently, C_I need only be approximately equal to C_p.

In a packaged diode, C_p can be varied over narrow limits by addition of external capacitance if $C_p < C_I$ which is generally true for large diodes, or by reducing the axial electric field around the diode in the manner described previously. The inductance can be controlled by external addition if L_p is less than required. The synthesis of the circuit is therefore very much controlled by the proper choice of the diode chip capacitance C_I, since this parameter cannot be controlled by the user.

The impedance appearing across A-B for the two resonant conditions is then

Series Resonance

$$R_{SR} \text{ where } R_{SR} = \frac{R_I}{1 + \frac{R_I^2}{X_{CI}^2}}$$

Parallel Resonance

$$R_S(1 + Q^2) \quad Q = \frac{X}{R_S}$$

and $R_S = R_I$ @ high forward bias

The isolation, insertion loss, and bandwidth may also be found from Table 1, simply by substituting R_{SR} for R_{SS} and R_S for R_{sp}. X is the reactance of either the resonating inductance or capacitance.

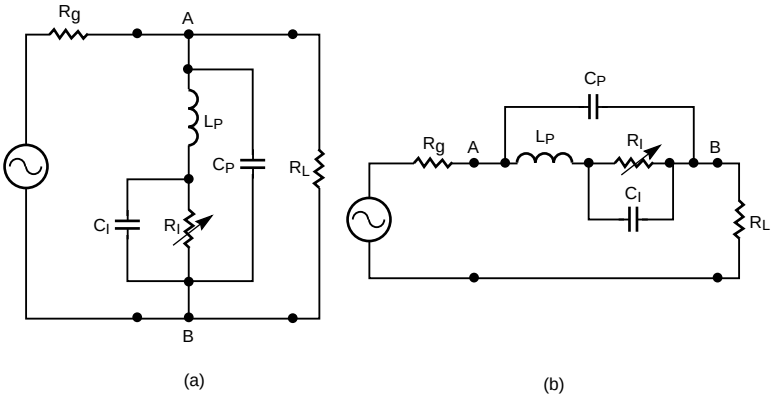


Figure 13. Resonant PIN Switches Realized with Use of Package Parasitics and Chip Capacitance.

Table 1.

	Attenuation (α)	Insertion Loss	Bandwidth
Series Mode Resonant Switch	Forward Bias $20 \log \left(\frac{R_{SP}^2 + X^2}{2Z_o R_{SP}} + 1 \right)$	Reverse Bias $20 \log \left(\frac{R_{SS}}{2Z_o} + 1 \right)$	Forward Bias $[4\pi Z_o C_1 X 10^{\alpha - 20}]^{-1}$
Shunt Mode Resonant Switch	Reverse Bias $20 \log \left(\frac{Z_o}{2R_{SS}} + 1 \right)$	Forward Bias $20 \log \left(\frac{Z_o R_{SP}}{2(R_{SP}^2 + X^2)} + 1 \right)$	Reverse Bias $\frac{Z_o}{4\pi L_2} X 10^{-\alpha/20}$

Applicable when $R_g = R_L = Z_o$

where:

$$R_{SP} = \left(\frac{Q_1}{1 + Q_1^2} \right) X$$

$$R_{SS} = \frac{X}{Q^2}$$

X = reactance of either the resonating inductance or capacitance

c) Design of Multiple Diode and Multi-throw Switches and Attenuators

When the maximum attenuation or isolation requirements are greater than what can be obtained by a single diode, multiple diode circuits using series, parallel or series-parallel arrangements can be used. Examples of such circuits are shown in Figure 14. A simple parallel or series connection of two diodes will only increase the attenuation by a maximum of 6 dB and will also increase the insertion loss.

However, if n diodes are spaced at quarter wavelength intervals, as shown in Figure 14(b), the overall attenuation can be increased by more than n times that of a single diode. The insertion loss, if it is due to parasitic elements, may also decrease because the 1/4 wavelength ($\lambda/4$) spacing produces cancellation of these reactances. Where $\lambda/4$ spacing is not practical, higher isolation can be achieved by using a series-shunt connection as shown in Figure 14(c). In this connection, isolations greater than the sum of that obtained with a single series and a single parallel diode may be obtained.

The discussion so far has been concerned with circuits that are essentially single-pole single-throw switches or two-port attenuators. A variety of multiple throw arrangements are also possible as shown in Figure 15. The design considerations for these circuits are similar to those outlined above except that interaction between diodes must be considered in the design. The SPDT circuit of Figure 15(a) operates as follows: When diode D_1 is forward biased and diode D_2 is zero or reverse biased, the RF power flows from Port 3 to Port 2 and Port 1 is isolated. When the two bias conditions are reversed, RF power flows to Port 1 and Port 2 is isolated. To minimize reactive loading of the open port by the closed port, the diodes are spaced $\lambda/4$ away from the feed point. The RFC provides a dc return for the bias currents and an open circuit for the RF signal. The capacitors C_1 and C_2 provide an RF ground for the diodes and an open circuit for the bias current. For increased isolation, additional diodes can be used which are spaced $\lambda/4$ away from the first set.

When $\lambda/4$ spacing is impractical or its bandwidth restriction is undesirable, a series configuration, as shown in Figure 15(b), can be used.

In this circuit, increased isolation can be provided by placing shunt diodes singly or in $\lambda/4$ spaced pairs at the output ports after the series diodes. This configuration is particularly useful for design of multiple-throw wideband switches. For multiple-throw configurations, additional series or series/parallel sections are added to the common feed point.

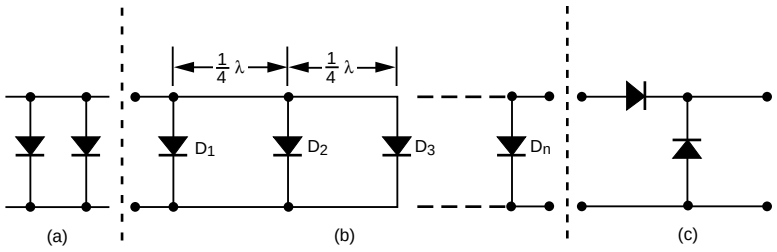


Figure 14. Reflective Mode Attenuators.

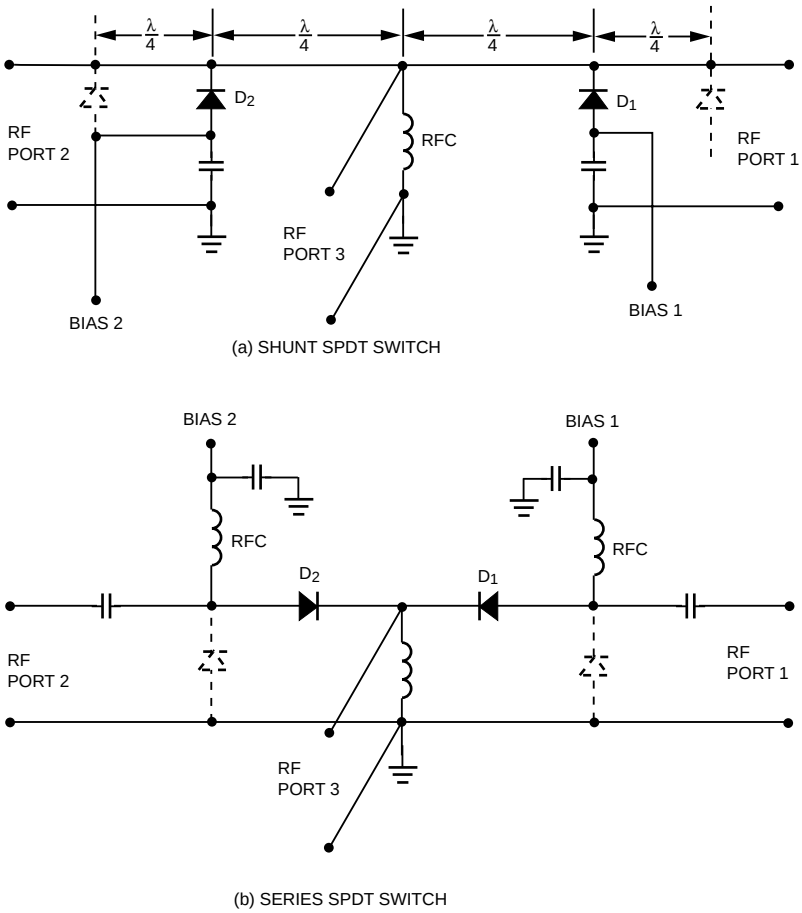


Figure 15. Multiple Throw PIN Switches.

d) Design of Constant Impedance Switches and Attenuators

For some RF systems, the high reflection coefficient at the RF ports of the reflective switches cannot be tolerated. For such applications, a variety of constant impedance switches and attenuators can be built using PIN diodes. Several such circuits are shown in Figure 16. The circuits of Figure 16(a), (b), and (c), operate by absorbing the undesired RF signal power in the PIN diodes. The circuits of 16(d), (e), and (f) operate by reflecting the power to a different RF port. In Circuits a and b, the control current variation through each diode is arranged in such a way that the impedance at both RF ports remain essentially constant at the design value, while the overall attenuation can be varied over a range of less than 1 dB to greater than 20 dB. In Circuit c, the input impedance is kept constant by using a distributed structure with a large number of diodes. The impedance variation of each diode is also shaped so that the diodes in the center of the structure vary more than those near the ports. The resulting tapered impedance structure results in an essentially constant impedance at the ports, while the overall attenuation can be varied up to a range of 40-80 dB, depending on the length of the structure. The π and T circuit structures are generally very compact and are particularly useful at low frequencies in the range of 10 MHz to 500 MHz. In addition to being constant impedance, they exhibit very little change of phase with attenuation and are very useful for providing gain control in RF transistor amplifiers in which bandpass and phase characteristics must be maintained over a large range of gain variation. Performance information on the π attenuator is contained in HP Application Note 1048.

Because of its distributed nature, the circuits of Figure 16(c) are generally too large to be useful at lower frequencies but are very useful above 1 GHz. This circuit is particularly attractive when very large attenuation ranges are required. Additional design information on this type of attenuator is contained in Ref. 1.

The attenuator circuits of Figure 16(d), (e), and (f) achieve constant impedance characteristics by virtue of the properties of the coupling elements which are placed between the RF ports and the PIN diodes. In the circuit of 16(d), the power incident on Port A of a 3 dB, 90° hybrid divides equally between Ports B and C, and Port D is isolated. The mismatch produced by the PIN diode resistance in parallel with the load resistance at Ports B and C reflects part of the power. The reflected powers at B and C combine and exit out of Port D. Port A, in this case, is isolated and therefore appears matched to the input signal. The maximum attenuation that can be achieved with this scheme depends on the directivity of the coupler and the quality of the terminations at Ports B and C when the diodes are unbiased. The VSWR at Port A will depend on the

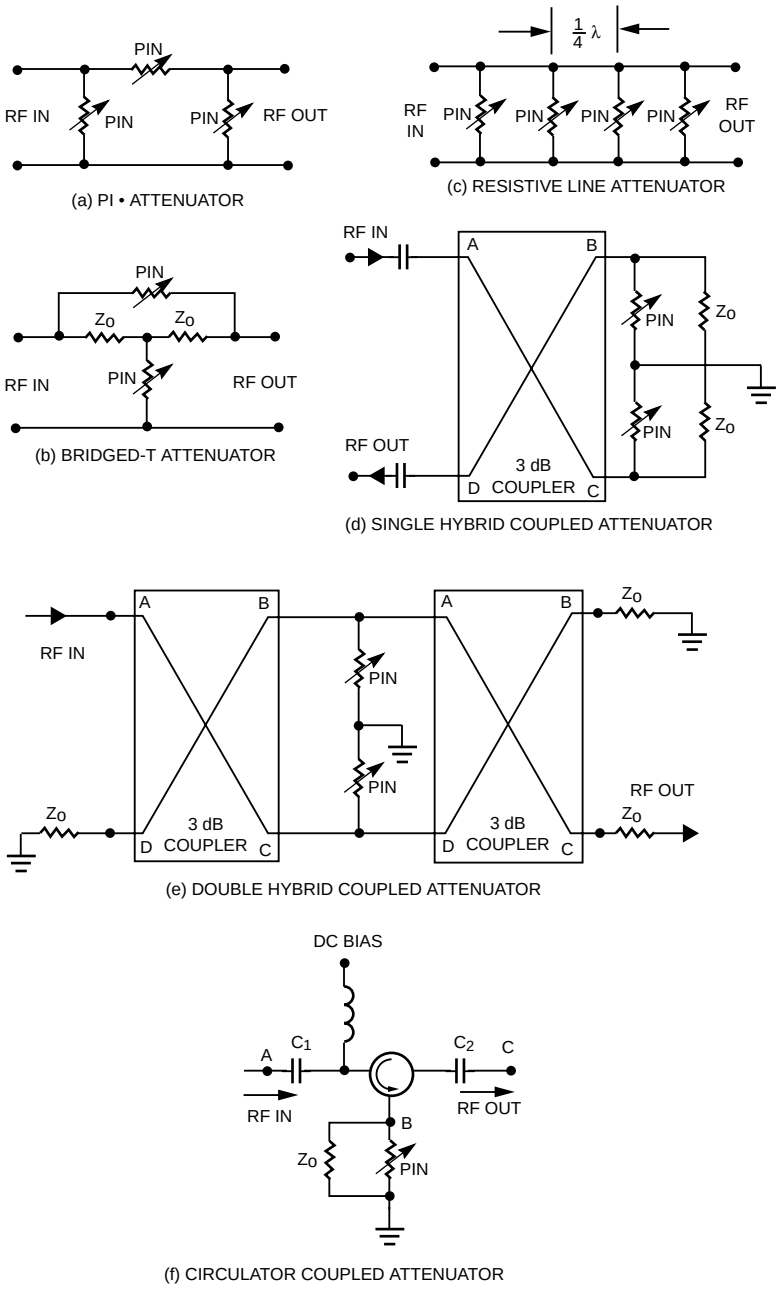


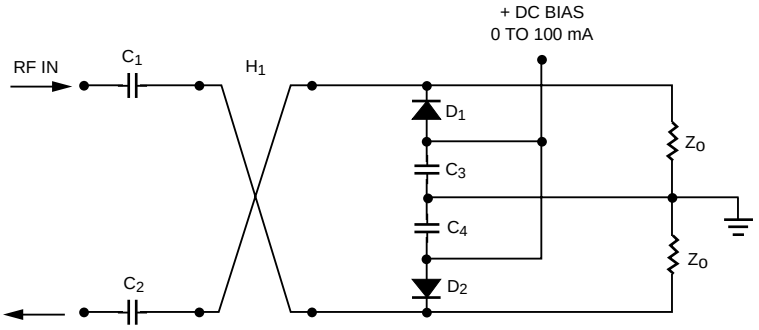
Figure 16. Constant Impedance PIN Diode Attenuators.

equality of the power split in the hybrid and the equality of the reflection coefflcients at Ports A and B. To assure the latter, the diodes must be reasonably well matched and must be spaced equally from Ports B and C. The insertion loss will depend on the losses and the equality of power split in the coupler and the minimum resistance of the PIN diodes when forward biased.

Circuits of this type can be easily and economically constructed in 50 ohm stripline form using the HP Stripline PINs. These diodes require no external matching structure and are very consistent in respect to the position of the effective internal reflection plane. The typical performance that was obtained with this type of circuit is illustrated in Figure 18 for the circuit of Figure 17. As can be seen, this kind of attenuator is useful over a wide frequency range. However, due to the hybrid directivity, there will be a considerable ripple in the high attenuation state. The ease with which the Stripline PIN diodes can be incorporated in such circuits is illustrated in Figure 19. Here broad-band biasing is achieved by applying the bias current to the diode at an "RF ground" point. After passing through the diode, the bias currents are returned to ground through the 50 ohm terminating loads.

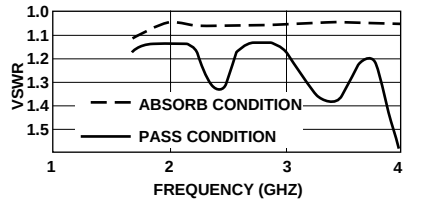
The ripple in the attenuation characteristics of the single hybrid coupled attenuator can be eliminated by using two identical hybrids as shown in Figure 16(e). This circuit eliminates the dependency of attenuation on hybrid directivity and its variation with frequency. The typical performance achieved with the double hybrid circuit using HP diodes is illustrated in Figure 21 for the circuit of Figure 20. The construction of this circuit is also in stripline but differs from the previous one by the biasing scheme. Instead of using ceramic RF bypass capacitors, the diodes make direct contact with the RF ground planes and bias is applied through inductors L_1 and L_2, which appear as a high impedance to the RF signal. As in the previous circuit, the spacing of the diodes from the hybrid ports must be equal and the diodes must be reasonably well matched.

The circuit of Figure 16(f) achieves constant impedance characteristics by virtue of the isolation properties of the circulator. The PIN diode shunts the matched B port impedance Z_o. When the PIN appears as a high resistance, all power is absorbed in Z_o at Port B. As the PIN impedance changes, Port B is progressively mismatched and power is reflected and passed by circulator action to Port C where it is absorbed in a matched load. The maximum attenuation attainable in this circuit depends on the matched port circulator isolation and the quality of the match of the circuit at Port B when the PIN diode is zero or reverse biased. The mechanical and electrical properties of the Stripline PIN diode make it ideally suited for direct integration into the usual stripline ferrite circulator structures.

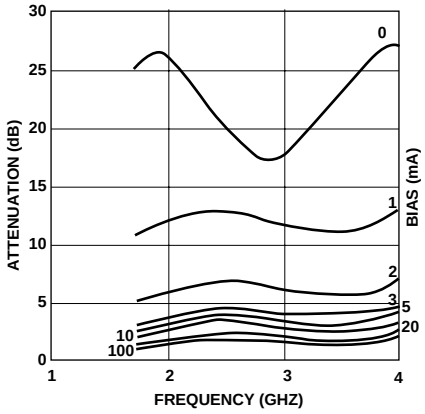


H_1 – 3 dB, 90° 1/8" STRIPLINE COUPLER (15 dB DIRECTIVITY OR BETTER 0.8 TO 4.3 GHz)
D_1, D_2 – HP STRIPLINE PIN DIODE
C_1, C_2 – DC BLOCKS
C_3, C_4 – 480 pF CERAMIC RF BYPASS CAPACITORS
Z_O – 50 OHM STRIPLINE LOADS

Figure 17. S-Band Single Hybrid Coupled PIN Attenuator.



(a) VSWR VERSUS FREQUENCY



(b) ATTENUATION VERSUS FREQUENCY

Figure 18. Performance of an S-Band Single Hybrid Coupled PIN Attenuator.

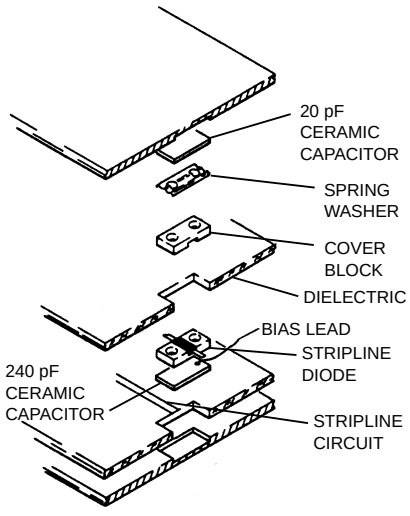
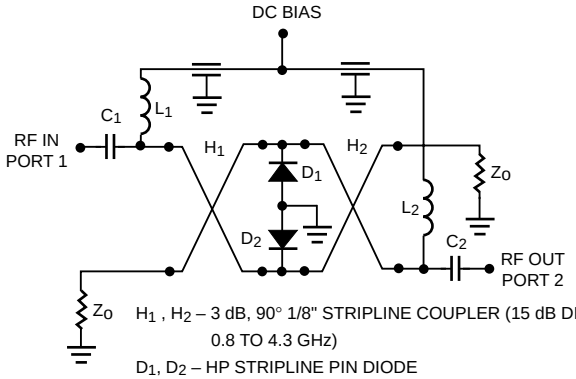


Figure 19. PIN Diode Biasing in Stripline Structures.

PIN Diode Phase Shifters

The high speed switching capabilities and the low "ON" and high "OFF" impedance states of the PIN diode make it also very useful for many types of high speed, current controlled phase shifter applications. Figure 22 shows a number of practical PIN diode phase-shifter circuits. All of these circuits are easily realized in economical stripline form and require no diode matching structures.

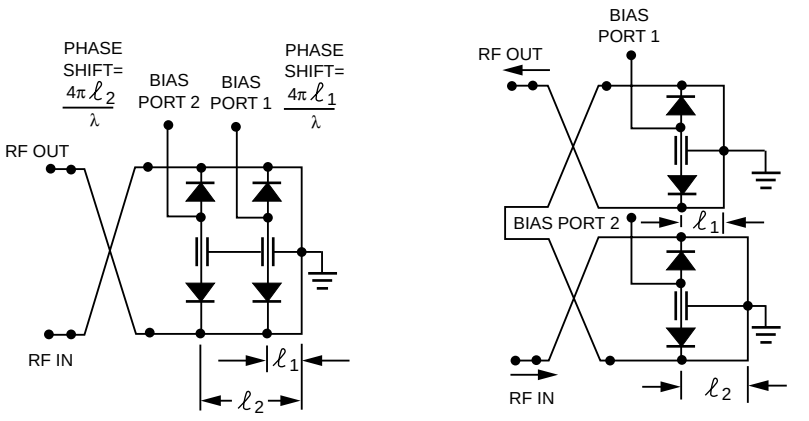
As an illustration, a two-bit, 90°-180°, phase-shifter circuit is shown in Figure 22(a). The circuit uses a conventional stripline 3 dB hybrid and HP stripline PIN diodes. The diode mounting and biasing was as shown in Figure 19. 90° and 180° of phase shift was obtained by biasing Ports 1 and 2, respectively. The insertion loss in any state should be approximately 1 dB and the VSWR < 1.1:1. For smaller phase increments, diodes can be spaced closer, however the effective electrical length of the package and the operating frequency will set the limit on the minimum phase increment in this kind of circuit. At higher frequencies and/or smaller phase increments, electrically shorter low parasitic diode packages, such as the HP beam lead, can be used or the circuit can be modified by cascading several stages of hybrid coupled phase shifters as shown in Figure 22(b). The hybrid coupled phase-shifters are generally economical to build and require a small number of diodes.



H_1 , H_2 – 3 dB, 90° 1/8" STRIPLINE COUPLER (15 dB DIRECTIVITY OR BETTER 0.8 TO 4.3 GHz)
D_1, D_2 – HP STRIPLINE PIN DIODE
C_1, C_2 – DC BLOCKS
L_1, L_2 – BIAS INDUCTORS
Z_O – 50 OHM STRIPLINE LOADS

Figure 20. S-Band Double Hybrid Coupled PIN Attenuator.

Other phase-shifter circuits, as shown in 22(c) and (d), are usually more expensive (circulator cost) or require more diodes. The loaded line circuit which had been popular in early phase-shifter designs requires a very large number of diodes. The design of this kind of circuit is well covered in existing literature, particularly Ref. 2.



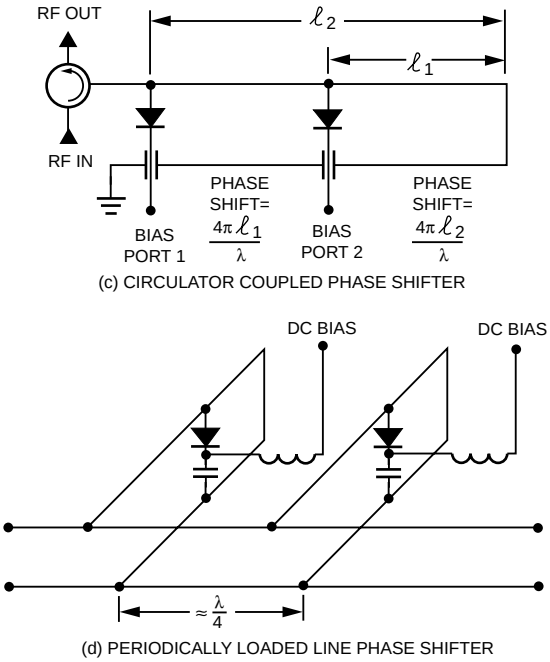


Figure 22. PIN Diode Phase Shifters.

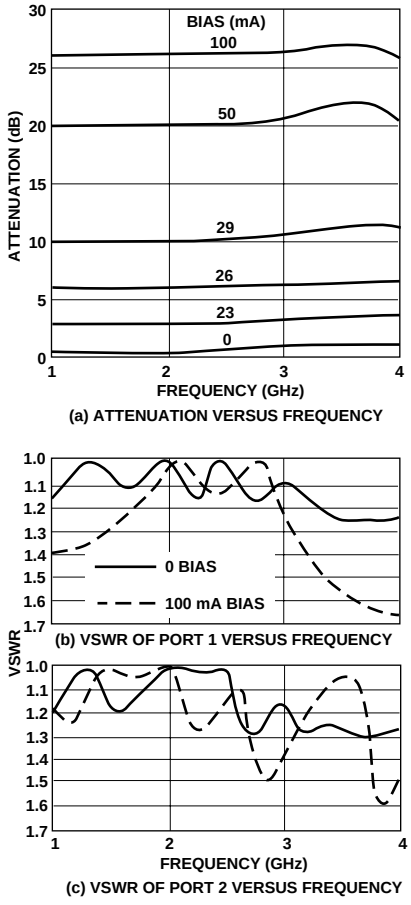


Figure 21. Performance of an S-Band Double Hybrid Constant Impedance PIN Attenuator.

PIN Diode Power Handling

The maximum signal RF power that a PIN diode can handle is limited by either the diode's breakdown voltage or its power dissipation capability. In most cases, power dissipation capability sets the lower limit on signal power handling.

Because PIN diodes are usually operated in a reflective mode, the amount of signal power that can be handled by a PIN in a circuit is usually very much larger than the actual power dissipated in the diode. The ratio of the power dissipated in the diode to the incident power depends on the impedance of the diode relative to the circuit impedance, the number of diodes, and their relative spacing. The following calculation sequence refers to Figures 23 and 24 which relate the CW Power Multiplier to either resistance or attenuation.

1. Read the power dissipation limit from the absolute maximum ratings given in the appropriate data sheet.
2. Determine the CW Power Multiplier from Figure 23 if diode resistance is known. Alternatively, use Figure 24 if circuit attenuation is known.
3. Multiply the power dissipation limit by the CW Power Multiplier determined in (2).

For example, for the series mode and a diode with a low and high resistance state of 1 and 10K ohms, respectively, approximately 2% (CW Power Multiplier = 50) of the input power will be absorbed by the diode in either the "ON" or the "OFF" state. If such a diode can dissipate 3 watts, it can therefore handle 150 watts of signal power. In the shunt mode, the low resistance state corresponds to the highest absorbed power, which in this case is 7.5% (CW Power Multiplier = 13.3) of the incident power. Hence, this same diode can handle 40 watts of signal power.

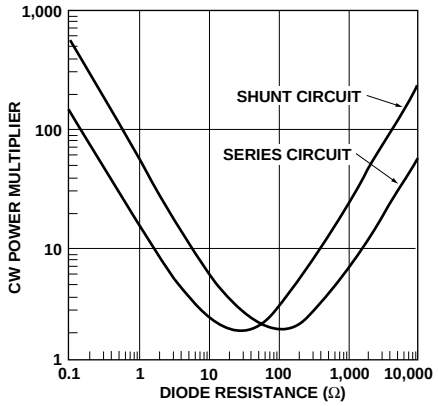


Figure 23. CW Power Multiplier vs. Diode Resistance.

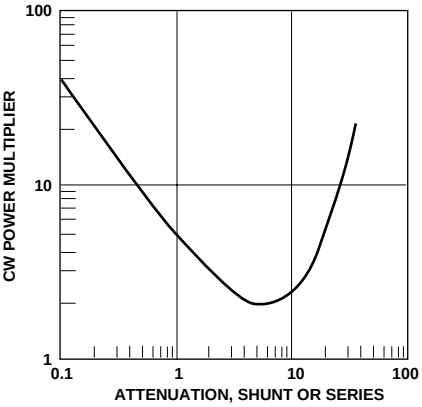


Figure 24. CW Power Multiplier vs. Series or Shunt Attenuation.

If two diodes are used in shunt to gain higher isolation, their spacing determines how much power is dissipated in each. The lead diode in a paralleled diode set will absorb either an equal or a greater amount of power than the second diode, depending upon the electrical spacing of the two diodes. Figure 25 shows what fraction of the input power will be dissipated in the lead diode as a function of total circuit attenuation and diode spacing for a shunt paralleled diode set in a 50 ohm system. Dotted curves indicate the constant resistance contours to which the diodes are biased.

A common switching application uses two shunt diodes in a double throw circuit, as shown in Figure 15. One shunt diode is used in each arm of the switch. The diodes are located a quarter wave from the junction of the two output arms and the input arm.

The power multiplier for either shunt diode in this circuit is much larger than the multiplier for a single shunt diode. For example. the multiplier for a 1 ohm diode is 53, compared to 13.5 for a single diode. For a 0.5 ohm diode, the multiplier is 103, compared to 26 for a single diode. The advantage is approximately a factor of four, so that a shunt diode in this double throw configuration can switch the same power as a series diode.

Since power handling can be many times greater than the power dissipated in the diode. the latter should be minimized and carefully controlled in high power applications. The maximum power that a PIN diode can dissipate is given by:

$$P_{max} = \frac{T_{j(max)} - T_A}{\theta_{jc} + \theta_{jA}}$$

where:

$T_{j(max)}$ = maximum operating diode junction temperature
T_A = Ambient temperature
θ_{jc} = Thermal resistance – junction to case
θ_{jA} = Thermal resistance – case to ambient

The junction-to-case thermal resistance is determined by diode and package design. The case-to ambient thermal resistance is

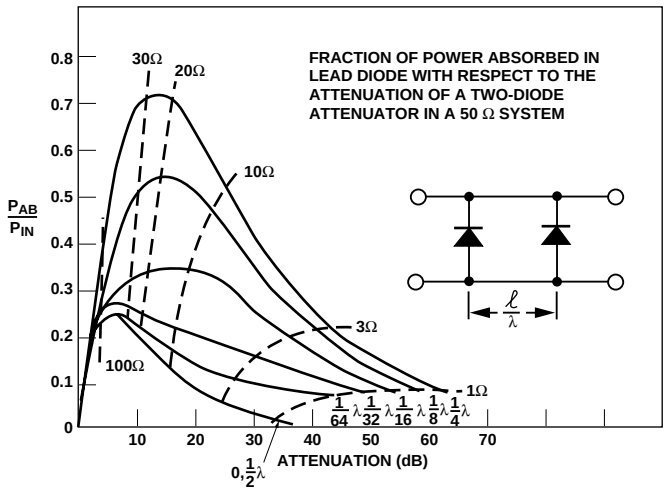


Figure 25. Fraction of Power Absorbed in Lead Diode with Respect to the Attenuation of a Two-Diode Attenuator in a 50 Ω System.

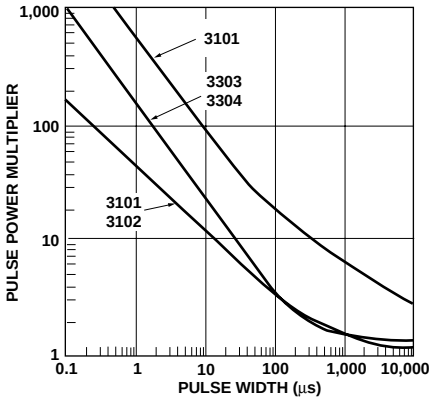


Figure 26. Pulse Power Multiplier vs. Pulse Width.

determined by how well the diode is mounted. For very low θ_{jc} diodes, good thermal mounting is particularly important. The HP Stripline package, with its large, flat, copper body and 2-bolt mounting capability, is particularly well suited for applications requiring a minimum overall thermal resistance.

If the PIN diode is used to control pulsed RF power, the peak power that can be handled increases as the pulse width decreases. This is because the effective thermal resistance decreases with pulse width. For a given pulse width, the maximum peak dissipated power can be determined from Figure 26. For example, at a pulse width of 10 μsec, the peak power handling capacity is increased by ten times.

In all cases, the voltage breakdown limit must be checked. For the series circuit this is expressed as:

$$P_A\,(max) = \frac{(V_{BR} - V_{BIAS})^2}{400}\,W$$

and for the shunt circuit:

$$P_A\,(max) = \frac{(V_{BR} - V_{BIAS})^2}{100}\,W$$

This limit is optimized by using zero reverse bias (V_{BIAS}). However, at low frequencies and/or high power, forward current may flow on the positive swing of the RF voltage, reducing the diode impedance. Reverse bias may be necessary in these cases. Reverse bias may also be needed to reduce switching time. See AN929, Fast Switching PIN Diodes, for details of this application.

References

1. J. K. Hunton and A. G. Ryals, "Microwave Variable Attenuators and Modulators Using PIN Diodes," IRE Transactions on Microwave Theory and Techniques, pp. 262-273, July 1962.
2. Richard W. Burns and Louis Stark,"PlN Diodes Advance High-Power Phase - Shifting," Microwaves, pp. 38-48, November 1965.
3. Robert V. Garver, "Theory of TEM Diode Switching," MTT, Pp. 224-238, May 1961.
4. Jack H. Lepoff. "A New PIN Diode for VHF-UHF Applications" I EEE Transactions on Broadcast & Television Receivers. pp 10-15, Feb. 1971.
5. R. Caverly and G. Hiller, "Distortion in p-i-n Diode Control Circuits" IEEE Trans. Microwave Theory Tech., vol. MTT-35, p. 492, May, 1987.