

## INTRODUCTION

Many microwave and RF systems require the frequency of a signal to be translated to a higher or lower frequency. Also, there are applications for the generation of a relatively low frequency voltage or current that is proportional to the amplitude of a higher frequency signal. The properties of a Schottky diode can be exploited to perform these tasks.

This application note is a survey of the physical and electrical characteristics of Schottky mixer and detector diodes. It reviews the semiconductor and electrical properties of these diodes and illustrates how they are used in a number of receiving circuits. It also presents a number of tables and criteria to select an appropriate Schottky diode depending on the requirement of the mixer or receiving system.

This application note is divided into eight sections:

- I. A discussion of the fundamentals of Schottky diodes including the physics of Schottky junctions and their characteristics such as resistance, capacitance and barrier heights. These properties ultimately determine the performance of all mixer and detector diodes.
- II. A discussion of the principles of variable resistance mixer diodes and the diodes' RF properties such as noise figure, conversion loss and impedance.
- III. A discussion of the principles of detector diodes and their RF properties such as sensitivity and video resistance.
- IV. A comparison of the differences in mixers and detectors when used in receivers.
- V. A discussion of common mixer, modulator and multiplier circuits which use Schottky diodes. Some of the advantages and disadvantages of different circuits are discussed.
- VI. A glossary containing definitions of the major terms used in discussing mixer and detector circuits and mixer and detector diodes.
- VII. Tables and graphs to aid in the selection of an appropriate mixer circuit or diode for a circuit based on the system's receiver requirements.
- VIII. A Selection Guide to help select the most appropriate microwave diode.

This application note has been the standard Schottky diode reference since it was written in the mid-1980's. Since then, the fundamental principles of mixer technology and Schottky junction physics have not changed, but many of the implementations of these technologies have evolved and improved. This note has been extensively revised to reflect these advances in diode and circuit design.

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## I. Schottky Diode Fundamentals

A Schottky barrier diode uses a rectifying metal-semiconductor junction formed by plating, evaporating or sputtering one of a variety of metals onto n-type or p-type semiconductor material. Generally, n-type silicon and n-type GaAs are used in commercially available Schottky diodes.

The properties of a forward biased Schottky barrier diode are determined by majority carrier phenomena. A pn junction diode's properties are determined by minority carriers. Schottky diodes are majority carrier devices that can be switched rapidly from forward to reverse bias without minority carrier storage effects. Because of this characteristic they make superior microwave mixer and detector diodes.

The normal current/voltage (I/V) curve of a Schottky barrier diode resembles that of a pn junction diode with the following exceptions:

1. The reverse breakdown voltage of a Schottky barrier diode is lower and the reverse leakage current higher than those of a pn junction diode made using the same resistivity semiconductor material.
2. The forward voltage at a specific forward current is also lower for a Schottky barrier diode than for a pn junction diode. For example, at 2 mA forward bias current a low barrier silicon Schottky diode will have a forward voltage of ~0.3 volts while a silicon pn junction diode will have a voltage of ~0.7 volts.

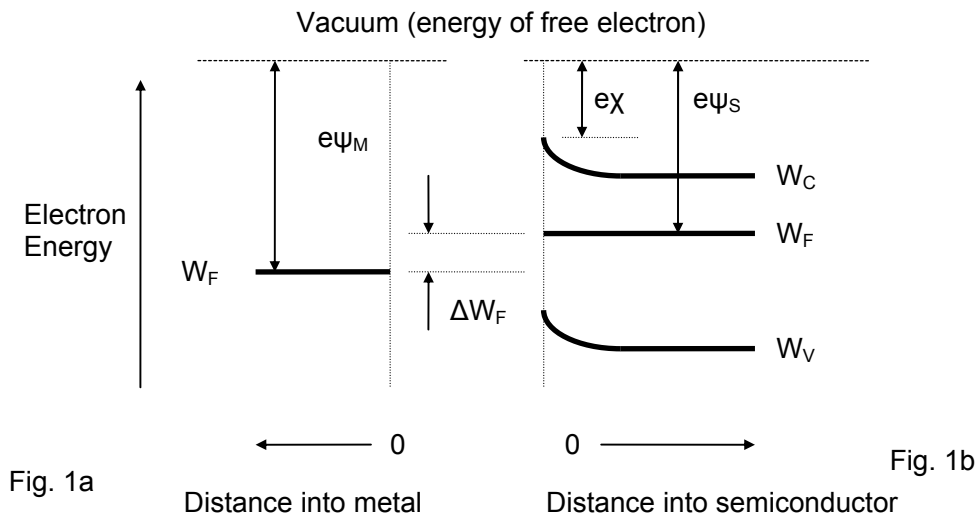
In order to understand the major electrical properties of a Schottky barrier diode, the physics of the barrier and the current across the barrier must be understood.

Figure 1 shows the electron energy levels in a metal as a function of distance from the surface of an isolated metal and on an isolated neutral n-type semiconductor with a net negative surface charge, which explains the curvature of the conduction and valence band energy plots.

In Figure 1a,  $e\psi_M$  is the vacuum work function or the potential required to remove an electron from the Fermi level,  $W_F$ , to a position outside of the metal. Typical values of  $e\psi_M$  are a few volts.  $e\psi_M$  is a constant value for a given, atomically pure metal, but varies with surface contamination.

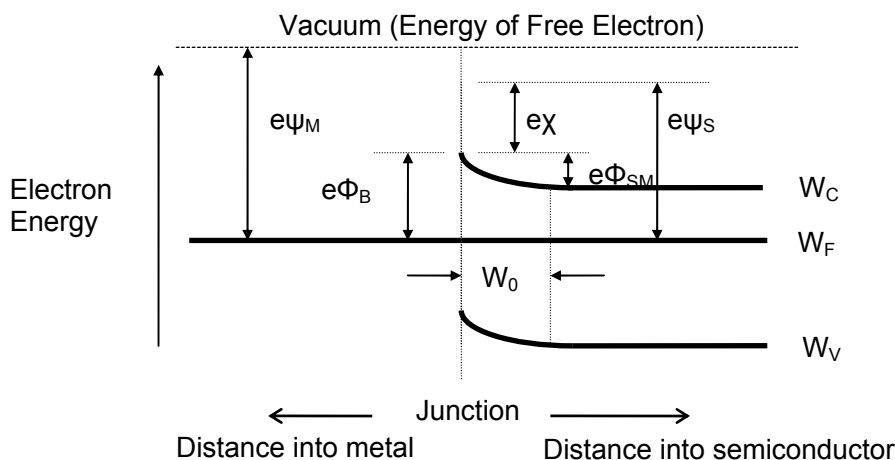
In Figure 1b,  $W_V$  and  $W_C$  are the energy levels of the semiconductor's valence and conduction bands, respectively. As in the metal,  $W_F$  is the Fermi level of the semiconductor and is a function of its doping. Note that the Fermi level of the semiconductor is not equal to the Fermi level of the metal. The energies  $e\chi$  and  $e\psi_S$  are the energies required to remove an electron from the conduction band and Fermi level respectively to a free position outside the semiconductor.

Figure 1b shows the effect of a net negative surface charge on the semiconductor. The Fermi level remains a straight, horizontal line as required by equilibrium. The effect of the surface charge is to bend the energy level of the valence and conduction bands near the semiconductor surface. Thus, the effect of surface charge is to alter the energy levels at the semiconductor surface.



**Figure 1. Energy-Level Diagrams vs. Distance for Metal and Semiconductor Surfaces in Isolated Equilibrium**

A Schottky barrier is formed when materials such as Figures 1a and 1b are brought into direct contact. The Fermi level or chemical potential of the materials must remain constant across the junction or interface at zero bias. Initially when the metal and semiconductors are brought into contact, their Fermi levels are not equal. There will be a net current transport from one material to the other. Then a potential barrier will form between the materials to make the carrier flow in each direction equal so that the net current is zero. In this condition the two materials are in thermal and charge equilibrium and the Fermi level is continuous across the junction. The result of this effect is a Schottky barrier junction.



**Figure 2. Energy Level Diagram of Metal and Semiconductor After Contact (Idealized)**

Figure 2 shows the energy level diagram for the Schottky barrier junction. Note that a potential barrier has formed in the semiconductor to adjust the electron flow from metal to semiconductor and semiconductor to metal. Then an electron in the metal at the Fermi level will encounter a potential barrier of  $\Phi_B$ . An electron in the conduction band of the semiconductor will see a potential barrier  $\Phi_{SM}$ . If the metal is atomically pure and the semiconductor does not have a surface charge, the value of  $f_B$  will be  $(\psi_m - \chi)$ . The presence of any impurities or surface charge on the semiconductor will alter the value  $\Phi_B$  somewhat. The quantity  $\Phi_B$  is often called the **barrier potential** or **contact potential** of a Schottky barrier.

If a voltage is applied to the metal-semiconductor junction in either direction, the Fermi level will no longer be continuous across the junction. Then the equal and opposite carrier flows which existed at zero bias will be changed so that a net current will flow in one direction (or the other, depending on the polarity of the applied voltage). In forward bias, the metal is positive with respect to the semiconductor; the bias will reduce the barrier  $f_{SM}$  for electron flow from the semiconductor to the metal, but the barrier for electron flow from the metal to the semiconductor will remain approximately the same. Thus a net positive current will flow due to the increased flow of electrons from the semiconductor to the metal.

For reverse bias, with the metal more negative with respect to the n-type semiconductor, the barrier for electrons flowing from the semiconductor to the metal increases. This almost eliminates this current component. To the first order, in reverse bias, the barrier for electron flow from the metal to the semiconductor remains constant and represents a net negative current.

The Schottky junction current model described above is called the thermionic emission model. It depends on energetic electrons crossing a potential barrier. For a complete treatment of current characteristics in Schottky diodes, the transport mechanism of electrons that quantum tunnel through a thin barrier must be added to the model.

## A. CURRENT VS VOLTAGE RELATION

The current/voltage (I/V) relationship for a Schottky barrier diode is given by the following equation known as the Richardson equation.

(1)

$$I = I_S \left( e^{\frac{qV}{nkT}} - 1 \right)$$

$$I_S = AA^*T^2 e^{-\frac{q\phi_B}{kT}}$$

Saturation Current,

where:

- A = junction area
- A\* = modified Richardson constant (value varies by material and dopant) = 110 A/(°K<sup>2</sup>-cm<sup>2</sup>) for n-type Si
- T = absolute temperature in K
- q = electronic charge = 1.6 \* 10<sup>-19</sup> C
- f<sub>B</sub> = barrier height in volts
- k = Boltzman's constant = 1.37 \* 10<sup>-23</sup> J/K
- n = ideality factor (forward slope factor, determined by metal-semiconductor interface)

The barrier height of a Schottky diode can be determined experimentally by fitting the forward I/V characteristic to the Richardson equation. Notice that f<sub>B</sub>, the potential barrier for electrons in the metal moving towards the semiconductor, influences the forward current.

The barrier height is important because it determines the local oscillator power necessary to bias the diode into its non-linear region. See Figure 48 for this relationship. In many high frequency receiver systems the available local oscillator power is limited so low barrier Schottky diodes must be used. Schottky diodes have been fabricated with several metals and alloys using p- and n-type silicon and n-type gallium arsenide, with barriers ranging from 0.27 eV to 0.90 eV. (See Table I for barrier heights of common metals, compounds and metal mixtures used for silicon & GaAs Schottky diodes).

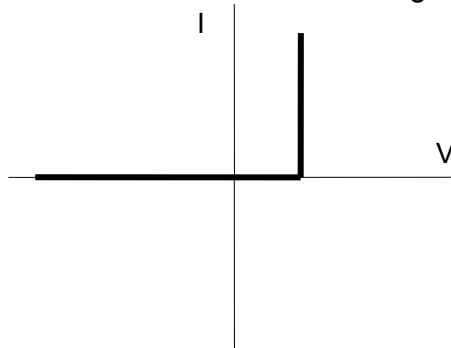
Semiconductors				
Crystal	Silicon (n type)	Silicon (p type)	Gallium Arsenide (n type)	
Orientation	<111>	<111>	<100>	<111>
Layer Doping	Nd $\sim 1 - 10 \cdot 10^{16}$	Nd $\sim 1 - 10 \cdot 10^{16}$	Nd $\sim 10^{17}$	Nd $\sim 10^{17}$
Metals	Barrier Height (eV)			
Au	0.81	0.34	0.90	0.99
Cr	0.55	0.52	-	-
Mo	0.60	0.42	-	-
Ni	0.55	0.51	0.76	0.81
Pd	0.72	0.35	-	-
Pt	0.85	0.28	0.86	0.93
Ti	0.48	0.61	0.76	0.82
W	0.69	0.45	0.80	0.86
Metal Silicides	Barrier Height (eV)			
Ni-Si	0.65 - 0.75	-	-	-
Pd-Si	0.72 - 0.76	0.30 - 0.36	-	-
Pt-Si	0.80 - 0.85	0.26 - 0.29	-	-
Ti-Si	0.56 - 0.63	-	-	-
W-Si	0.60 - 0.65	-	-	-
Alloys or Metal Mixtures	Barrier Height (eV)			
Ni-Cr	~ 0.55	-	-	-
Ti-W*	~ 0.50 - 0.70	0.25 - 0.45	-	-

\* Depending on Mixture

**Table 1. Experimental Values of the More Common Metal-Semiconductors and Metal Silicide Barrier Heights in eV on Silicon and Gallium Arsenide<sup>7,9</sup>**

### B. SCHOTTKY DIODE EQUIVALENT CIRCUIT

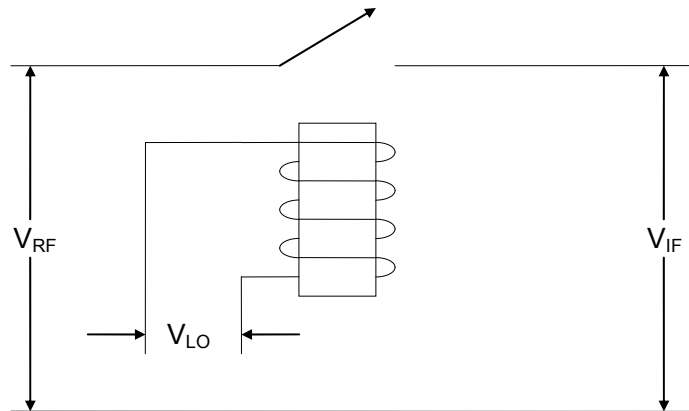
The ideal Schottky barrier mixer diode would have the following I/V characteristic:



**Figure 3. Ideal Schottky Diode I/V Characteristic**

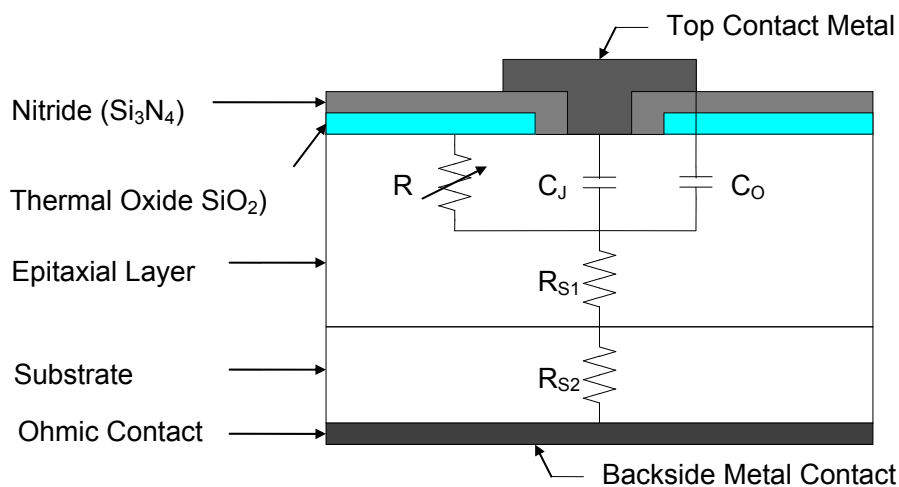


The ideal mixer diode can be considered as a series switch controlled by  $V_{LO}$ .  $V_{IF}$  is produced by gating  $V_{RF}$ . An ideal mixer diode would have no series resistance under forward bias and no capacitance under reverse bias. However, in practice this is not possible.



**Figure 4. Ideal Mixer Diode**

Figure 5a shows the cross section of a typical Schottky diode die. This die has two layers of passivation on its top surface, surrounding the metal-semiconductor Schottky junction. The  $\text{SiO}_2$  ( $\epsilon_r \approx 4$ ) is formed by oxidizing the top surface of the Si epitaxial layer with very pure, de-ionized water vapor. This type of passivation is frequently called “thermal oxide”, since the operation takes place in a very clean, tightly controlled furnace at approximately  $900^\circ\text{C}$ . The resulting passivation is very efficient but vulnerable to contamination from metals and other materials. Another layer of passivation,  $\text{Si}_3\text{N}_4$  ( $6.7 \leq \epsilon_r \leq 7$ ), is deposited on top of the  $\text{SiO}_2$  to substantially reduce this vulnerability. This type of die is compatible with die attach and semi-automatic wire bonding methods normally used with diode packages and with hybrid circuits.

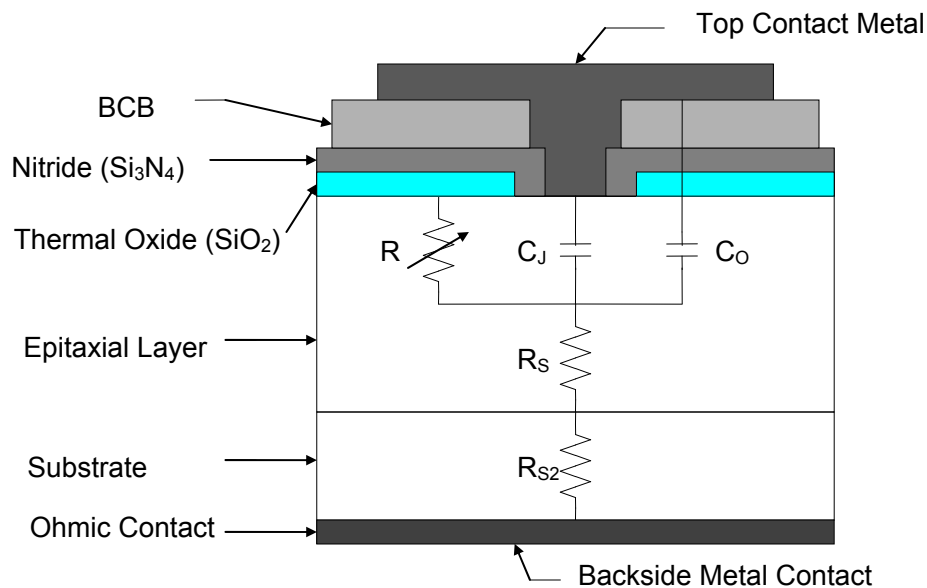


**Figure 5a. Schottky Diode Die with Equivalent Circuit**

Market forces require continually improved, higher frequency electrical performance from semiconductors with lower prices. These conditions require that diodes be packaged in plastic packages, such as the SOT-23, SOT-323, SOD-323, etc., using automated assembly techniques. However, the implications of these requirements are in mutual opposition. Better performance at higher frequencies requires lower junction capacitance, which is achieved by reducing the area of the metal-semiconductor junction. The optical recognition systems used with automated wire bonding assembly equipment have minimum feature sizes smaller than which they cannot recognize properly. This minimum feature size is much larger than the metal-semiconductor junction area that is required for an RF or microwave Schottky diode.

At first glance this problem appears easy to solve by simply increasing the diameter of the metal that is deposited on top of the passivation layers, as shown in Figure 5a, to produce a feature large enough to be optically detected and recognized. This approach can substantially increase the diode's overlay capacitance ( $C_0$ ) to the point that the total diode capacitance becomes too large for high frequency operation.

Since the minimum top metal size is determined by the capability of the optical recognition system used, the only alternatives that the diode designer has is to either make the dielectric layers of the overlay capacitance (the passivation layers) thicker or to use materials with lower relative dielectric constant.



**Figure 5b. Schottky Diode Die with BCB and Equivalent Circuit**

Recent advances in material science have produced many polymers, one of which, benzo-cyclobutene (BCB) is particularly well suited for use with microwave semiconductors. Its low relative dielectric constant ( $\epsilon_r = 2.7$ ) and dissipation factor along with its superior mechanical strength make BCB a good material to use as a third, topmost layer of dielectric in small

capacitance diodes that must have very large diameter top contacts in order to be compatible with automated assembly. Such a die is shown in Figure 5b.

Note that in Figures 5a and 5b the thicknesses of the substrate and epitaxial layers are not drawn to scale. In actual practice, the substrate is typically many times thicker than the epitaxial layer.

In actual mixer operation the Schottky junction can be modeled as a nonlinear resistance ( $R_j$ ) and a shunt capacitance ( $C_j$ ). The nonlinear resistance is the element used for mixer and detector action and will be discussed in detail later. The nonlinear resistance can be obtained from the basic I/V relation for the Schottky barrier (see equation 1). The elements  $R_{S1}$  and  $R_{S2}$  represent resistive losses in the epitaxial layer and substrate layer respectively. These constant resistive losses are generally included in the term  $R_S$ , the total series resistance. The remaining circuit model element is the overlay capacitance ( $C_O$ ), which is the parasitic capacitance that results from the contact metal extending beyond the active region, over the passivation. Figure 6 shows an equivalent circuit for a beam lead Schottky device.

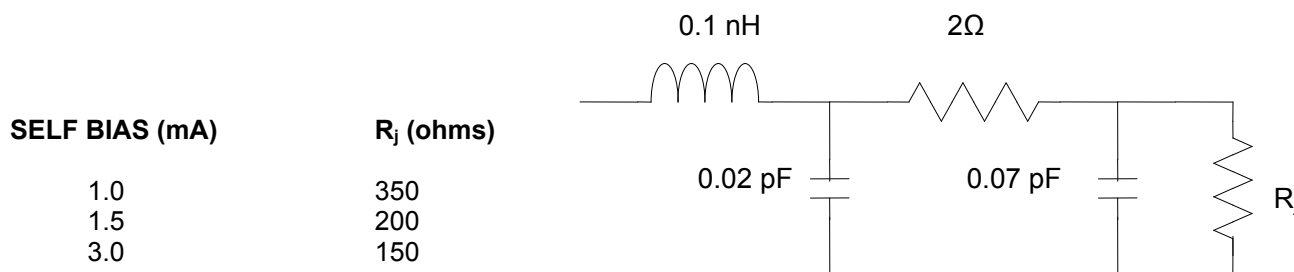


Figure 6. Equivalent Circuit for a MA40415 Beam Lead Device

### C. TOTAL CAPACITANCE OF A SCHOTTKY DIODE

The total capacitance of a packaged Schottky barrier diode is given by:

(2)

$$C_t = C_j + C_o + C_p$$

where:

- $C_j$  = metal - semiconductor junction capacitance
- $C_o$  = overlay capacitance across the oxide layer
- $C_p$  = package capacitance

The overlay and package capacitances can be either substantially reduced or eliminated by using SURMOUNT or beam lead diodes.

## 1. Junction Capacitance

The junction capacitance of a Schottky barrier diode is given by:

$$(3) \quad C_j(V) := \left[ \frac{q \cdot \epsilon_s \cdot S \cdot N_D}{2 \cdot \left( \phi_{sm} - \frac{kT}{q} - V \right)} \right]^{\frac{1}{2}}$$

or

$$C_j(V) := \frac{C_j(0)}{\left( 1 - \frac{V}{\phi_{sm} - \frac{kT}{q}} \right)^{\frac{1}{2}}}$$

where:

- $\epsilon_s$  = electric permittivity of the semiconductor
- $N_D$  = donor density in n-layer
- $\Phi_{SM}$  = barrier voltage seen by electrons in the semiconductor for traversal into the metal
- $V$  = applied voltage
- $C_j(0)$  = junction capacitance at zero volts

A convenient method for determining the barrier voltage  $\Phi_{SM}$  for a specific metal semiconductor combination is to plot  $(1/C_j)^2$  versus voltage. The intercept on the voltage axis is given by  $\Phi_{SM} - kT/q$ . Note: The capacitance versus voltage relation is governed by the barrier seen in the semiconductor while the current voltage relationship is governed by  $\Phi_B$ , the barrier seen by electrons in the metal. These barriers differ in potential by the separation of the Fermi level in the semiconductor from the conduction band divided by the electronic charge or  $(e_C - e_f)/q$ .

## 2. Overlay Capacitance

As seen in Figure 5, the overlay capacitance  $C_O$  is the parasitic capacitance of the contact metallization extending beyond the active junction area and over the passivating oxide. If the effects of surface charges on the semiconductor or depletion of the semiconductor-SiO<sub>2</sub> interface by the applied voltage are neglected, the overlay capacitance can be modeled as a parallel plate capacitor with the SiO<sub>2</sub> layer as a dielectric. Then  $C_O$  becomes:

$$(4) \quad C_O := \frac{\epsilon_1 \cdot A_1}{W_O}$$

where:

- $\epsilon_1$  = electric permittivity of SiO<sub>2</sub>
- $A_1$  = area of overlay region (annular ring)
- $W_0$  = thickness of oxide passivation

The overlay capacitance is a parasitic element which should be minimized for optimum diode performance. Reducing  $C_O$  to a minimum value becomes especially important for frequencies above X band, but there is a trade-off with the contact size. It is normally very difficult to attach wire bonds to contact sizes smaller than 1-2 mils. When junction capacitances for Schottky diodes are specified they normally include this overlay capacitance. Usually  $C_O$  is no more than ~0.02 pF for 1-2 mil diameter contact sizes.

#### D. SERIES RESISTANCE

The total series resistance shown in Figure 5 consists of the resistance of the undepleted epitaxial layer ( $R_{S1}$ ) plus the resistance of the substrate ( $R_{S2}$ ). A low frequency model, which neglects skin effect, will be discussed.

The contribution of the undepleted epitaxial layer to the diode resistance is given by:

(5)

$$R_{S1} := \frac{\rho l}{A}$$

$$R_{S1} := \frac{1}{q \cdot \mu_e \cdot N_D \cdot A}$$

where:

- $\rho$  = resistivity of undepleted epitaxial layer
- $l$  = thickness of undepleted epitaxial layer
- $A$  = area of Schottky junction
- $\mu_e$  = electron mobility in undepleted epitaxial layer (assumes layer is n-type)
- $N_D$  = donor density in undepleted epitaxial active layer

The resistance contributed by the substrate may be modeled by using the resistance of a contact dot and the size of the junction on a semi-infinite semiconductor substrate. This model is normally valid because the active diode diameter is usually much less than the thickness of the substrate. Using this model,  $R_{S2}$  becomes:

(6)

$$R_{S2} := \frac{\rho S}{2 \cdot d}$$

$$R_{S2} := \left( \frac{\rho S}{4} \right) \cdot \sqrt{\frac{\pi}{A}}$$

where:

$\rho$  = substrate resistivity

$d^s$  = active junction diameter

Using equations 5 and 6, the total resistance  $R_S$  becomes:

(7)

$$R_S := \frac{1}{q\mu_e \cdot N_D \cdot A} + \left( \frac{\rho S}{4} \right) \cdot \sqrt{\frac{\pi}{A}}$$

The above analysis totally neglects skin effect, which may increase the substrate contribution to  $R_S$ . For a high frequency model,  $R_{S1}$  will be given by the same expression as above, but in order to model  $R_{S2}$  one must consider that current will flow in a surface layer only one skin depth thick in the substrate. The first component of  $R_{S2}$  to consider will be the spreading resistance of the current into the area directly under the active region one skin depth thick into the substrate. The second will be the resistance of the top surface of the chip. This component may be approximated as the resistance of an annular ring of inner diameter  $d$ , outer diameter  $D$ , the total chip width, and the thickness  $d$  which is the skin depth.

The third component of  $R_S$  is the resistance of the chip side walls, modeled with a thickness  $d$ . The total  $R_S$  at millimeter wave frequencies is the sum of these three components plus the resistance of the active epitaxial area. It is normally not necessary to consider skin effects below approximately 50 to 60 GHz for most diodes.

## E. FIGURE OF MERIT

The cutoff frequency (Figure of Merit) of a Schottky barrier diode is maximized by minimizing the  $R_S C_j$  product. Furthermore, mixer conversion loss ( $L_C$ ) can be shown to be directly proportional to the product of diode series resistance ( $R_S$ ) and junction capacitance ( $C_j$ ). By converting these parameters to semiconductor properties of the active junction, the following figure of merit for a Schottky barrier diode can be obtained:

14

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(8)

where:

$$\text{Figure of Merit} := R_S \cdot C_j \cdot \alpha \cdot L_C \cdot \alpha \cdot \frac{W \cdot \sqrt{\epsilon}}{\mu \cdot \sqrt{N_D}}$$

$\epsilon$  = electric permittivity of the semiconductor

$W$  = undepleted epitaxial layer thickness

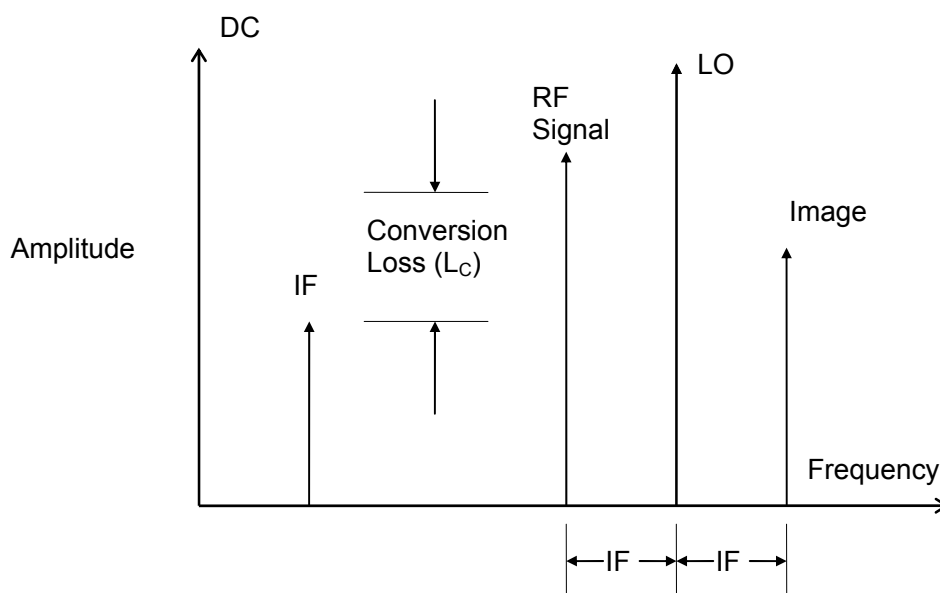
$N_D$  = carrier concentration in active region

$\mu$  = carrier mobility in active region

## II. Principles Of Mixer Diodes

Frequency mixing is the conversion of a low power level signal, (commonly called the RF signal) from one frequency to another by combining it with a higher power (local oscillator) signal in a device with nonlinear impedance. Mixing produces a large number of new frequencies which are the sums and differences of the RF and local oscillator signals and their respective harmonics. In a down converter mixer the intermediate frequency (IF) is the desired output signal. In most applications this signal is the difference of the RF and local oscillator frequencies.

The relationship of these signals to the mixing function is shown in Figures 7 and 8



**Figure 7. Frequency Relationships in a Mixer**

Signals at two different frequencies can produce an output signal at the IF.

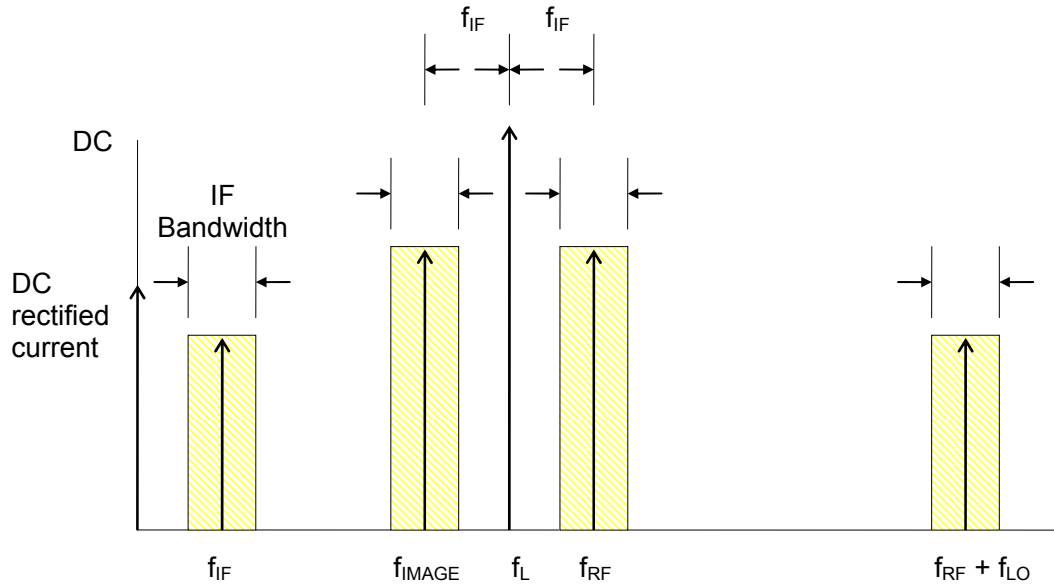
The first of these signals is called the signal frequency or the RF signal. This signal typically has been modulated by another circuit or system. In a down converter mixer, the RF signal frequency is either  $f_{LO} + f_{IF}$  or  $f_{LO} - f_{IF}$ .

The second signal frequency that can produce an output signal at the IF is called the image frequency ( $f_{IM}$ ). The image frequency is offset from the LO frequency by the IF frequency. Energy at the image frequency can degrade noise figure, produce interference or increase distortion of the receiver system. A properly designed mixer will terminate the image frequency signal. In a down converter mixer, if the desired RF signal frequency is  $f_{LO} + f_{IF}$  then the image frequency is  $f_{IM} = f_{LO} - f_{IF}$  and vice versa.



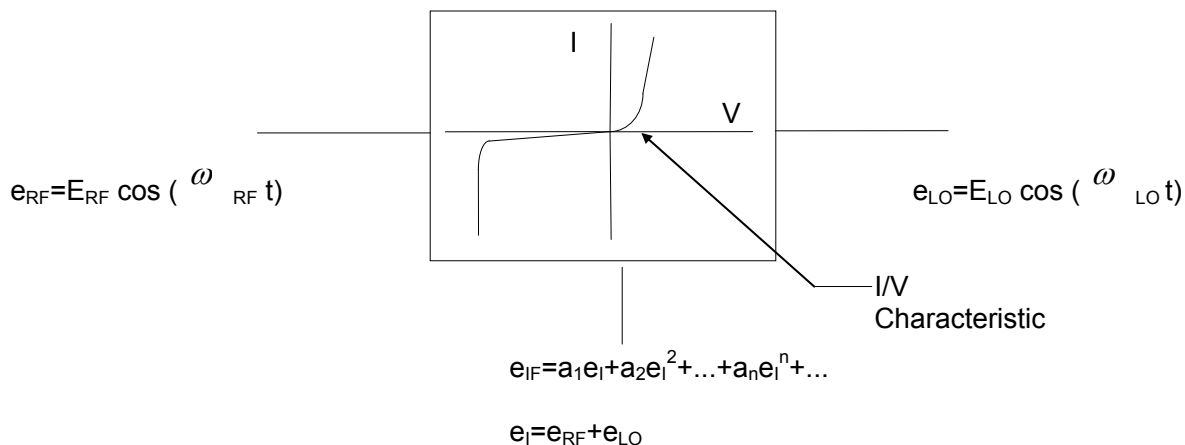
A major consideration of all mixers is the conversion loss which is the reduction of signal power when it is converted from the RF to the IF frequency. The conversion loss ( $L_C$ ) is illustrated graphically.

Figure 8 shows in more detail some of the many frequency components which are generated in a mixer. Note that the RF and image signals can appear on either side of the LO frequency.



**Figure 8. Frequency Relationships in a Mixer**

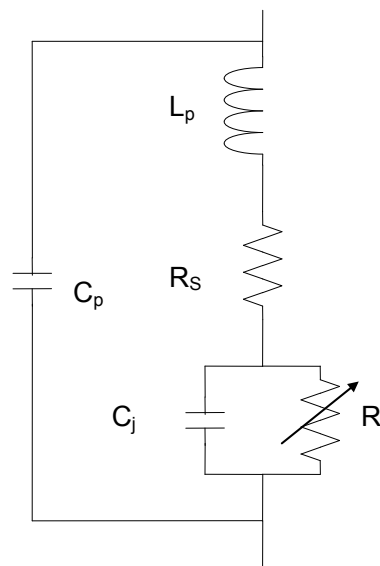
When the RF and local oscillator signals are combined in a variable resistance diode, the frequency components are given by a series expansion. This phenomenon has also been described as multiplication of the RF and local oscillator signals in the time domain. Some of the frequency components are shown below<sup>1,8</sup>:



$$e_{IF} = \begin{matrix} a_1 E_{LO} \cos(\omega_{LO} t) & + & \dots & \text{LO signal} \\ a_1 E_{RF} \cos(\omega_{RF} t) & + & \dots & \text{RF signal} \\ a_2/2 (E_{LO}^2 + E_{RF}^2) & + & \dots & \text{DC component} \\ a_2 E_{LO} E_{RF} \cos((\omega_{LO} - \omega_{RF}) t) & + & \dots & \text{Lower sideband signal} \\ a_2 E_{LO} E_{RF} \cos((\omega_{LO} + \omega_{RF}) t) & + & \dots & \text{Upper sideband signal} \\ a_2/2 E_{LO}^2 \cos(2\omega_{LO} t) & + & \dots & \text{2nd harmonic of local} \\ & & & \text{oscillator signal} \\ a_2/2 E_{RF}^2 \cos(2\omega_{RF} t) & + & \dots & \text{2nd harmonic of RF signal} \\ & + & \dots & \text{etc.} \end{matrix}$$

### A. EQUIVALENT CIRCUIT OF A MIXER DIODE

The Schottky mixer diode may be shown as a nonlinear resistance,  $R_j$ , shunted by a capacitance,  $C_j$ , in series with a resistance,  $R_s$ . This equivalent circuit is shown in Figure 9. The resistance is the nonlinear barrier resistance at the rectifying contact. The capacitance is the barrier and overlay capacitance. At low frequencies the barrier capacitance does not affect rectification but at microwave frequencies its shunting action will reduce the RF voltage across the barrier. Since it is impossible to tune out  $C_j$  with an external inductance at microwave frequencies because of the presence of  $R_s$ ,  $C_j$  must be kept small to minimize reduction in rectification efficiency. The diode package parasitics are represented by the series package inductance ( $L_p$ ) and the shunt package capacitance ( $C_p$ ). The effects of both  $L_p$  and  $C_p$  must be considered when packaged diodes are used.



**Figure 9. Equivalent Circuit of Packaged Mixer Diode**

## B. BASIC MIXER DIODE RF PARAMETERS

A fundamental limitation on the sensitivity of a microwave receiver employing a diode mixer arises from the fact that in the frequency conversion process only a fraction of the available RF signal power is converted into power at the intermediate frequency. Some RF signal is also converted to the usually unwanted image frequency and other harmonics, too. This overall loss is dependent primarily on the diode junction properties, and secondarily on the diode's package parasitics (i.e., mismatch of signal power by  $R_S$ ,  $C_j$ ) and on the match at the input and output ports of the mixer. An additional limitation on performance arises from the fact that the mixer diode itself generates noise (noise temperature ratio) when it is driven by the local oscillator. The conversion loss and the noise temperature ratio are the parameters of most interest in the microwave mixer diode. The mixer diode is completely characterized by the following parameters: conversion loss, noise temperature ratio, receiver noise figure, RF impedance and IF impedance.

### 1) Conversion Loss Theory

The conversion loss of a mixer diode is dependent on several factors, including both the package and the Schottky diode die. Conversion loss,  $L_C$ , can be considered to be the sum of several losses.

The first component of total diode conversion loss can be called the matching loss which is dependent on the degree of impedance match obtained at both the RF signal and IF ports. Less than optimum match at either of these ports will result in a reduction in the available RF signal at the diode and the inefficient transfer of the IF signal. The matching loss can be expressed as:

(9)

$$L_1(\text{dB}) := 10 \cdot \left[ \log \left[ \frac{(S_{RF} + 1)^2}{4 \cdot S_{RF}} \right] + \log \left[ \frac{(S_{IF} + 1)^2}{4 \cdot S_{IF}} \right] \right]$$

where  $S_{RF}$ , and  $S_{IF}$  are RF and IF SWRs respectively.

The second component is the loss of signal power due to the diode's parasitic elements and, is called the diode's parasitic loss. The parasitic elements causing this loss are the junction capacitance ( $C_j$ ) and the series resistance ( $R_S$ ). The diode parasitic loss is the ratio of the input RF signal power to the power delivered to the junction variable resistance,  $R_j$ :

(10)

$$L_2(\text{dB}) := 10 \cdot \log \left( \frac{P_{in}}{P_{out}} \right)$$

Expressing this loss in terms of diode parameters:

(11)

$$L_2(\text{dB}) := 10 \cdot \log \left[ 1 + \frac{R_S}{R_j} + (\omega \cdot C_j)^2 \cdot R_S \cdot R_j \right]$$

where  $R_j$  is the time average value of junction resistance as established by the local oscillator drive level. The minimum value of  $L_2$  occurs when  $R_j$  is equal to  $1/(\omega \cdot C_j)$ :

(12)

$$L_{2\min}(\text{dB}) := 10 \cdot \log(1 + 2 \cdot \omega \cdot C_j \cdot R_S)$$

Since the value of  $R_j$  is strongly dependent on the local oscillator drive level, the value of  $L_2$  is a function of LO drive.  $R_S$  is also a weak function of drive level. If the LO drive is increased above the optimum value,  $L_2$  will increase due to power dissipation in  $R_S$ , while decreasing LO drive also gives insertion loss increase due to the shunting effect of the junction capacitance. In general, for many mixers  $L_{2\min}$  occurs when  $R_j$  is in the range of 250 ohms. This normally occurs at a diode rectified current of approximately 1 to 1.5 mA.

The third component is the actual conversion loss at the diode junction. This loss depends mainly on the voltage versus current characteristics of the diode and the circuit conditions at the RF and IF ports. The nonlinear behavior of the diode is represented by a time-varying conductance,  $G$ , which is dependent on the DC characteristics of the diode and local oscillator voltage waveform across the diode. Conversion loss and impedance values can then be calculated for the various image terminations by means of linear network theory. The minimum conversion loss ( $L_3$ ) at the diode junction for a broadband mixer (image properly terminated) in terms of incremental conductances is given by:

(13)

$$L_{3\min} := 2 \cdot \frac{\left[ 1 + \left[ 1 - 2 \cdot \frac{g_1^2}{g_0} \cdot (g_0 + g_2)^{\frac{1}{2}} \right] \right]}{\left[ 1 - \left[ 1 - 2 \cdot \frac{g_1^2}{g_0} \cdot (g_0 + g_2)^{\frac{1}{2}} \right] \right]}$$

where  $g_0$ ,  $g_1$  and  $g_2$  are incremental conductances, which are derived from a series expansion of the diode conductance obtained from the diode I/V equation:

(14)

$$I := I_S \left( e^{\frac{qV}{nkT}} - 1 \right)$$

It can be shown that  $L_{3 \text{ min}}$  approaches, as a limit, a value of  $\sim 3$  dB. Thus, for an ideal mixer diode, the theoretical minimum conversion loss is 3 dB under broadband conditions because a maximum of half the incident RF power is delivered to the IF port and the remaining RF power is dissipated at the image termination.

Under narrow band conditions, the image frequency can be reactively terminated such that RF power at the image frequency recombines with the local oscillator signal to improve the conversion loss of the diode. Under ideal conditions, theory predicts that a conversion loss of 0 dB for open or short circuited image terminations can be obtained. Values as low as 1 to 1.5 dB have been obtained in laboratory image recovery mixers.

The overall conversion loss,  $L_C$ , of a mixer diode is the sum of the three loss components,  $L_1$ ,  $L_2$  and  $L_3$ .

(15)

$$\begin{bmatrix} \text{Conversion} \\ \text{Loss} \end{bmatrix} = \begin{bmatrix} \text{Matching} \\ \text{Loss} \end{bmatrix} + \begin{bmatrix} \text{Parasitic} \\ \text{Loss} \end{bmatrix} + \begin{bmatrix} \text{Junction} \\ \text{Loss} \end{bmatrix}$$

or

$$L_C = L_1 + L_2 + L_3 \text{ (dB)}$$

For most production mixers a conversion loss of 4.5 to 6 dB is a reasonable value that can be obtained without extensive fine tuning.

## C. NOISE IN MIXER DIODES

### 1) Noise Temperature Ratio

In variable resistors or varactor mixers, there are three main sources of increased noise. The first is the thermal noise, which is present in all conductors at thermodynamic equilibrium. The second is shot noise, which is generated by moving charge carriers under the influence of an electric field. The third component, which increases with decreasing frequency, is usually referred to as 1/f or flicker noise. The noise temperature ratio includes the effects of all three of these contributors.

### 1C) Thermal Noise

The thermal noise for a Schottky barrier is given by the expression:

$$(16) \quad \overline{i^2} = 4 k T G B$$

where:  $k$  = Boltzmann's constant  
 $G$  = diode conductance  
 $B$  = bandwidth under consideration  
 $i^2$  = mean square noise current

### 2C) Shot Noise

The sources of shot noise in a Schottky barrier are similar to that of pn junctions. In a Schottky diode under forward bias there is a net flow of electrons from the semiconductor to the metal, giving rise to DC current,  $I$ . Equal and opposite components of saturation current,  $I_S$ , also flow across the barrier. These currents do not produce a net current in the external circuit, but do produce shot noise. Total shot noise is attributed to the three components. The resulting shot noise current is given by:

$$(17) \quad \overline{i_n^2} = 2q(I + 2I_S)B$$

In terms of diode AC conductance ( $G$ ), the noise temperature ratio ( $t_B$ ) of the barrier is defined as:

$$(18) \quad t_B = \frac{\overline{i_n^2}}{4kTGB}$$

As shown,  $t_B$  is the ratio of the diode mean square noise current to the mean square thermal noise current of a passive conductance. Using the I/V equation for a Schottky barrier diode,  $t_B$  can be reduced to

$$(19) \quad t_B := \frac{1}{2} \cdot \left( 1 + \frac{I_S}{I + I_S} \right)$$

The noise temperature ratio,  $t$ , of the composite diode, consisting of the Schottky barrier with noise temperature,  $t_B$ , and series resistance,  $R_S$ , with its thermal noise is given by the expression

(20)

$$t = \frac{R_j * t_B + R_S}{R_S + R_j}$$

where  $R_j$  is the dynamic resistance of the barrier (reciprocal of  $G$ ). Values of  $t$  and  $t_B$  less than one have been measured experimentally for Schottky barrier diodes. When the silicon Schottky barrier diode noise is due entirely to shot noise:

(21)

$$t_B := \frac{1}{2} \cdot \left( 1 + \frac{I_S}{I + I_S} \right)$$

The saturation current is usually much smaller than 1. The saturation current,  $I_S$ , for a platinum-silicon (n-type) Schottky barrier diode is  $\sim 2 \times 10^{-14}$  amps and the rectified current,  $I$ , is usually 0.1 to 1 mA under local oscillator bias conditions. Thus, for ordinary DC forward biases

(22)

$$t_B \approx 1/2$$

Under optimum local oscillator excitation, symmetry effects reduce the shot noise to much smaller values. At the same time, however, conversion of the source and image thermal noise, together with the series resistance's thermal noise, results in a noise temperature,  $t$ , close to 1.0. Normally, Schottky diodes have  $t < 1.0$ .

### 3C) Flicker Noise (1/f)

Flicker noise is a type of noise whose magnitude is inversely proportional to the frequency at which it is measured. It occurs in thin metal films, carbon resistors, copper oxide rectifiers, crystal varistors and all other semiconductor devices. The causes of flicker noise are not fully understood, although it is probably a surface effect due to large dependence of the noise magnitude upon the condition of the conducting material's surface and the environment surrounding it. Schottky diodes generally have lower "1/f" noise when compared to point contact diodes and are very suitable for applications involving a low IF frequency, e.g., Doppler radars. In general the lowest 1/f noise is obtained with back diodes. Unpassivated Schottky diodes tend to have less 1/f noise than those with an oxide passivation. However, unpassivated diodes are more susceptible to environmental stresses.

## D. OVERALL RECEIVER'S NOISE FIGURE

The most important criterion of mixer performance is its contribution to the overall receiver's noise figure. The noise at the output of a receiver is the sum of the noise arising from the input

termination (source) and the noise contributed by the receiver itself (i.e., due to the IF amplifier and mixer diodes). The noise factor is the ratio of the actual output noise power of a device to the noise power which would be available if the device were perfect and merely amplified the thermal noise of the input termination without contributing any noise of its own. Noise factor is given by the relation:

(23)

$$F := \frac{\left( \frac{S_i}{N_i} \right)}{\left( \frac{S_o}{N_o} \right)}$$

where:

- $S_i$  = available signal power at the input of receiver
- $N_i$  = available noise power at the input of receiver
- $S_o$  = available signal power at the output of receiver
- $N_o$  = available noise power at the output of receiver

The noise figure is the noise factor in decibels (i.e.):

(24)

$$NF(\text{dB}) := 10 \cdot \log \left[ \frac{\left( \frac{S_i}{N_i} \right)}{\left( \frac{S_o}{N_o} \right)} \right]$$

The overall noise figure of a receiver depends on the conversion loss ( $L_C$ ) of the mixer, the noise temperature ratio ( $t$ ) of the mixer diode and on the noise figure of the IF amplifier ( $F_{IF}$ ). It is given by the relation:

(25)

$$NF = L(t + F_{IF} - 1)$$



Mixer diodes are usually specified using  $F_{IF}$  of 1.5 dB. This allows comparison of different diodes under similar test conditions.

The mixer noise can also be expressed in terms of mixer input noise temperature,  $T_M$ :

(26)

$$T_M = T_o * t_o$$

where:

$T_o$  = measurement temperature

$t_o$  = Noise temperature of the diode

### E. MIXER DIODE RF IMPEDANCE

The RF impedance of the variable resistance mixer diode is a property of prime importance in the design of mixers.

Any impedance mismatch at the signal and LO frequencies not only results in signal loss due to reflection but also affects the IF impedance at the IF terminals of the mixer. This effect becomes more serious for mixer diodes with low conversion loss. The RF impedance of a mixer diode can be measured by a SWR method or directly with a network analyzer. The RF impedance is affected by local oscillator power. Normally this power is part of its specification.

The RF impedance is a complicated function depending on package geometry, size and shape of package parts and composition of the semiconductor and its junction parameters. To establish a good match between a semiconductor chip and RF transmission line, an impedance matching transformer is generally required.

### F. MIXER DIODE IF IMPEDANCE

The IF impedance is the impedance seen looking into the IF port of a mixer. It is important to match this impedance to the IF amplifier input impedance. The pertinent mixer diode IF impedance ( $Z_{IF}$ ) is that impedance at the output terminals of the mixer when the mixer diode is driven by a local oscillator. The IF impedance is a function of the local oscillator power level and also depends on the RF properties of the mixer and circuits connected to the RF terminals of the mixer. The IF impedance of a mixer diode driven by a LO is given in terms of its incremental conductances. For the broadband case it is:

(27)

$$Z_{IF} = \frac{1}{g_O} \cdot \left[ 1 - 2 \frac{g_1}{g_O} \cdot \left( \frac{g_O}{g} \right)^2 \right]^{\frac{1}{2}}$$

where  $g_o$ ,  $g$ , and  $g_1$  are incremental conductances.

An accurate measurement of  $Z_{IF}$  is essential for measuring noise temperature ratio ( $t$ ) and conversion loss ( $L_C$ ) of a mixer diode. It is normally done with an admittance bridge. Almost all mixer diodes have their  $Z_{IF}$  specified at a moderate RF frequency, i.e. 30-50 MHz, and at a fixed LO drive power level.

### III. Principles Of Detector Diodes

#### A. BASIC DETECTOR DIODE CHARACTERISTICS

RF and microwave signals can be detected by direct rectification using a nonlinear semiconductor such as a Schottky barrier diode. The sensitivity, however, is often mediocre in comparison to that of a good superheterodyne receiver.

Figures 13 and 14 illustrate the detecting function. The input signal is an RF signal whose amplitude, as a function of time, is the desired output. Optional DC bias to the detector diode may represent an additional input. The output of a detector is a low frequency signal called the video signal. Its amplitude is proportional to the square of the voltage amplitude of the RF signal. The frequency relationships in a detector are illustrated in Figure 13b. The RF input is shown as a carrier with amplitude modulation sidebands. The video signal will be a low frequency signal related to the amplitude modulation of the RF input as shown.

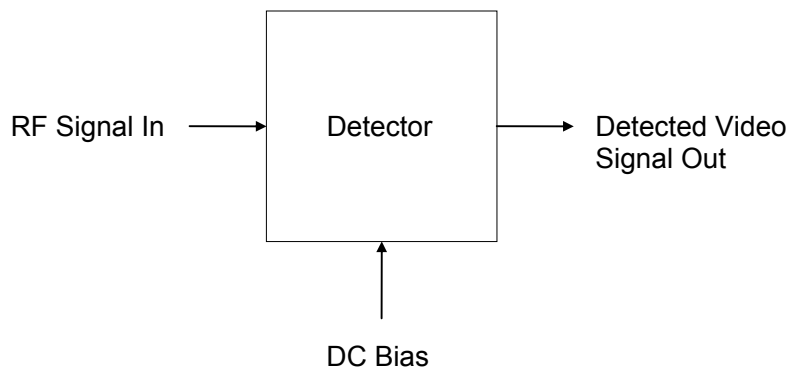


Figure 13a

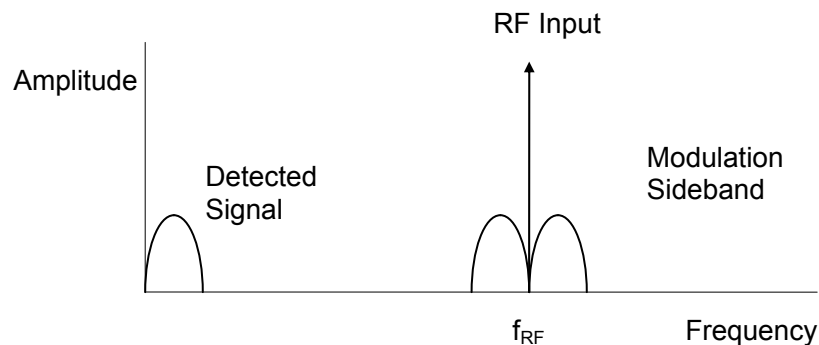
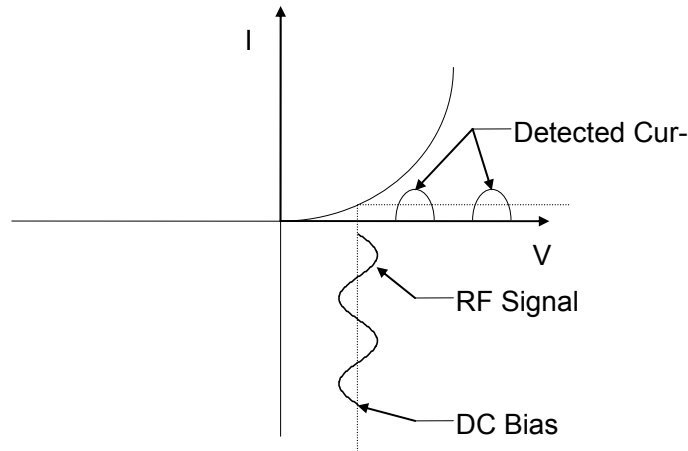


Figure 13b

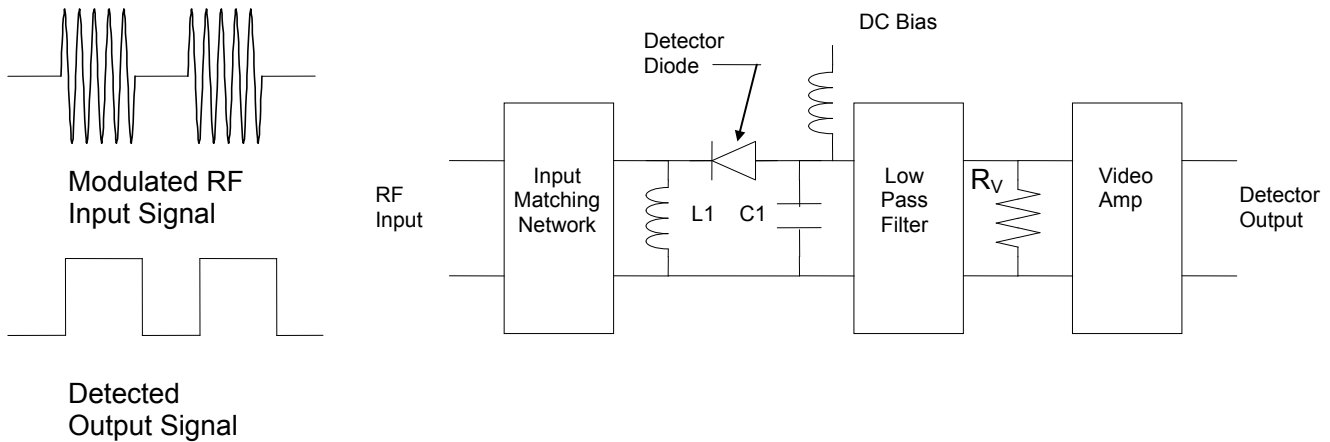
At small RF power the output current is proportional to the square of the RF input voltage. A "real" detector diode has the approximate characteristics shown below.



**Figure 14. Application of Signal Voltage to Schottky Diode**

### B. THE VIDEO DETECTOR

The block diagram of a typical video detector circuit is shown below:

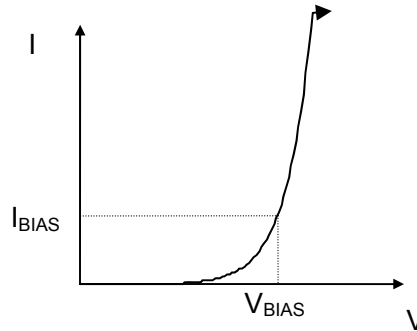


**Figure 15. Video Detector Circuit and Waveforms**

where:

- L1 = the return for the DC and demodulated signal
- C1 = the bypass capacitance for all RF signal components
- $R_v$  = input impedance of video amplifier

The I/V characteristic for an ideal detector diode is:



**Figure 16. Ideal Detector Diode I/V Characteristic**

(28)

$$I = I_S \left( e^{\frac{qV}{nkT}} - 1 \right)$$

where:

$I_S$  = reverse saturation current

$n$  = ideality factor (which equals one for an ideal diode)

$q/kT$  = 38.6 volts<sup>-1</sup> at  $T = 300K$

Assume that the voltage across the diode consists of a bias voltage  $V_{bias}$  and a small RF voltage ( $\delta v$ ). If a Taylor series expansion is performed about the bias point  $V_{bias}$ :

(29)

$$I = I(V_{BIAS} + dv)$$

$$I = I_{BIAS} + \left. \frac{\partial I}{\partial V} \right|_{V_{BIAS}} \delta v + \frac{1}{2} \left. \frac{\partial^2 I}{\partial V^2} \right|_{V_{BIAS}} (\delta v)^2 + \dots$$

If the RF signal voltage  $\delta v = V_{RF} \cos(\omega_S t)$ , then the RF signal current is

(30)

$$G_{RF} = \left. \frac{\partial I}{\partial V} \right|_{V_{BIAS}} V_{RF} \cos(\omega_{RF} t)$$

$$G_{RF} = \left. \frac{\partial I}{\partial V} \right|_{V_{BIAS}} = \frac{q * (I_{BIAS} + I_S)}{nkT}$$

Next, assume that the input signal is amplitude modulated:

$$(31) \quad v = V_{RF}(1 + m \sin(\omega_M t)) \cos(\omega_{RF} t)$$

where:

$\omega_{RF}$  = RF frequency

$\omega_M$  = modulation frequency

Substitution into the Taylor series yields:

$$(32) \quad i = I_{BIAS} + G_{RF} V_{RF} (1 + m \sin(\omega_M t)) \cos(\omega_{RF} t) + \frac{1}{2} \frac{\partial^2 I}{\partial V^2} \Big|_{V_{BIAS}} V_{RF}^2 (1 + m \sin(\omega_M t))^2 \cos^2(\omega_{RF} t) + \dots$$

We are interested in the demodulated components of the above current since the RF currents are bypassed by C1. Therefore,

$$(33) \quad I_m = \frac{1}{2} \frac{\partial^2 I}{\partial V^2} \Big|_{V_{BIAS}} V_{RF}^2 (m \sin(\omega_M t) - \frac{m^2}{4} \cos(2\omega_M t)) \approx \frac{1}{2} \frac{\partial^2 I}{\partial V^2} \Big|_{V_{BIAS}} m V_{RF}^2 \sin(\omega_M t)$$

This result shows that  $I_m$  is proportional to the modulation signal,  $m \sin \omega_M t$ . It also shows that the video output is proportional to RF power,  $V_{RF}^2$ . This is why it is called a square law detector. Finally, note that the conversion efficiency is related to the second derivative of the I/V curve, i.e. the change in slope.

$$(34) \quad \frac{1}{2} \frac{\partial^2 I}{\partial V^2} \Big|_{V_{BIAS}} = \frac{q^2(I_{BIAS} + I_S)}{2(nkT)^2} = \frac{q^2 I_{BIAS}}{2(nkT)^2}$$

since  $I_{BIAS} \gg I_S$ .

Since  $\partial^2 I / \partial V^2$  increases with forward bias, it is evident that the output current at the modulation frequency can be increased by the application of forward bias.

The magnitude of the demodulated current:

$$(35) \quad I_m = \frac{m V_{RF}^2}{2} \left( \frac{q^2 I_{BIAS}}{(nkT)^2} \right) = \frac{qm V_{RF}^2}{2nkT} G_{RF} = \frac{qm P_{RF}}{nkT}$$

since

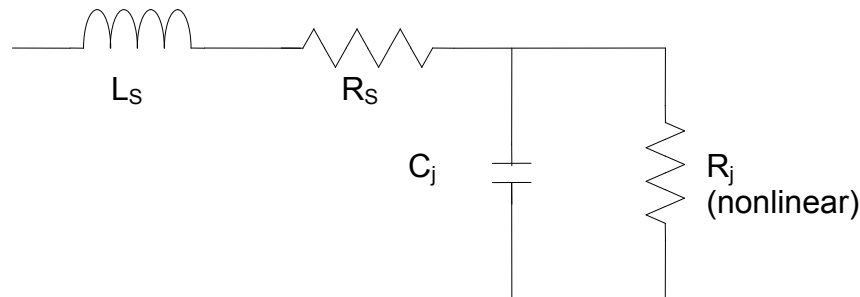
$$P_{RF} = V_{RF}^2 \frac{G_{RF}}{2}$$

For example, for square wave modulation where  $m = 1$ :

(36)

$$\frac{I_m}{P_{RF}} = \frac{qm}{nkT} = 38.6 \frac{\mu A}{\mu W}$$

In actual practice,  $I_m/P_{RF}$  is usually several microamperes per microwatt. The reason the theoretical value is not obtained is that an actual diode has other losses. The approximate equivalent circuit is shown below.



**Figure 17. Detector Diode Equivalent Circuit**

The non-linear resistor represents the I/V curve of the diode. At RF frequencies, it is represented by  $R_{RF} = 1/G_{RF}$ . The loss due to the parasitic circuit elements is given by:

(37) 
$$L \text{ (dB)} = 10 \log [1 + R_s/R_j + \omega^2 C_j^2 R_s R_j]$$

Note that the loss increases with frequency, so diodes with small  $C_j$  are required for good microwave detectors. The parasitic reactances ( $L_s$ ,  $C_j$ ) are often helpful in matching  $R_{RF}$  to  $Z_0$ . It is common for  $R_{RF} > Z_0$ , so  $L_s$  and  $C_j$  serve as a step down transformer.

If the video impedance ( $R_V$ ) is chosen to match the output impedance of the detector, then:

(38)

$$P_M := \frac{1}{2} \cdot I^2 \cdot m_{RMS} \cdot R_V$$

or

$$P_M = \frac{1}{4} * R_V * \frac{q^2 m^2}{(n * k * T)^2} * P_{RF}^2$$

The conversion efficiency is  $P_M/P_{RF} = K P_{RF}$  where K is a constant whose value is determined by the detector diode and detector circuit design.

This relation states that conversion efficiency decreases as  $P_{RF}$  decreases. This is a fundamental limitation of video detection.

Small signal detection is also limited by noise. In the video detector,  $1/f$  noise dominates. The detection capability of a video detector is characterized by its tangential signal sensitivity (TSS) which is expressed in dBm. Its relation to video bandwidth is:

(39)

$$TSS \propto \sqrt{B}$$

where B is video bandwidth.

A useful relationship is:

(40)

$$TSS|_{BW1} = TSS|_{BW2} - 10 \log \sqrt{\frac{BW1}{BW2}}$$

The sensitivity of a low level video detector depends primarily on the following three factors:

the RF matching structure determines the amount of total incident energy that is imposed on the active junction for rectification

the rectification efficiency, output impedance and noise properties of the diode determine the response of the diode junction to incident microwave radiation and

the input impedance, bandwidth and noise properties of the video amplifier at the detector output will affect the overall detector sensitivity.

### C. DETECTOR DIODE ELECTRICAL CHARACTERISTICS

The following section discusses the most important parameters for detector diodes as they are normally used in diode specifications.



### 1) Video Resistance ( $R_v$ )

$R_v$  is the real part of the diode's small signal impedance. This parameter has been shown to be dependent on the DC bias current and the diode's series resistance.

$$R_v = R_j + R_s$$

where:

$R_j$  = small signal junction resistance

$R_s$  = diode series resistance

$R_j$  can be determined by taking the first derivative of the diode I/V relationship.

(41)

$$I := I_S \left( e^{\frac{qV}{nkT}} - 1 \right)$$

(42)

$$R_j = \left( \frac{dI}{dV} \right)^{-1}$$

or

$$R_j := \frac{nkT}{q} \cdot \left( \frac{1}{I + I_S} \right)$$

where:

$I_S$  = saturation current

$q$  = electronic charge

$n$  = ideality factor

$T$  = Temperature (K)

Normally  $I_S \ll I$ , then

(43)

$$R_j := \frac{nkT}{q \cdot I}$$

or

$$R_j := \frac{0.026}{I}$$

for the case of  $n = 1$ ,  $T = 300$  K, and  $I$  is expressed in mA. Most common video detectors will have video impedances in the range of 500 to 10K ohms in normal usage.

## 2) Voltage Sensitivity ( $\gamma$ )

The voltage sensitivity of a detector or a detector diode is the ratio of open circuit video signal voltage to the RF input power.

(44)

$$\gamma = \frac{V_{OCV}}{P_{IN}}$$

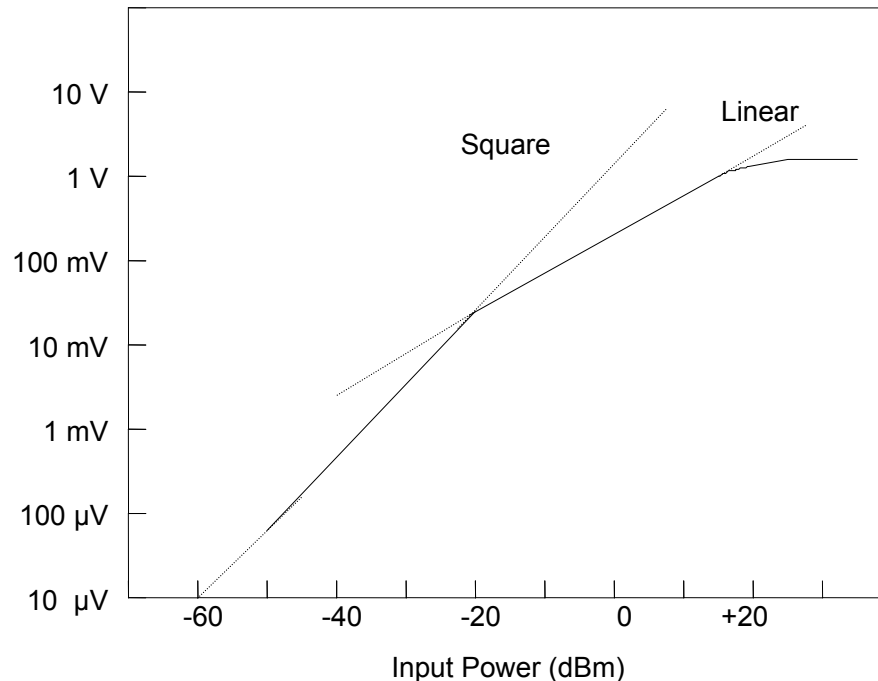
where:

$V_{OC}$  = open circuit video voltage

$P_{IN}$  = RF power incident on the detector

Voltage sensitivity is usually expressed in units of millivolts per milliwatt. To assure that the detector diode is in the square law range,  $\gamma$  is usually measured at -20 to -30 dBm input power levels.

Figure 18 shows a typical detector voltage sensitivity characteristic and the normal square law relationship of a Schottky diode detector.



**Figure 18. Voltage/Sensitivity Characteristics of a Detector Diode**

### 3) Current Sensitivity (b)

The current sensitivity,  $b$ , for a detector diode is the ratio of short circuit video current to the RF input power.

(45)

$$\beta := \frac{I_{SVC}}{P_{IN}}$$

where:  $I_{scv}$  = short circuit video current

The units of  $b$  are milliamps per milliwatt.  $g$  and  $b$  are related as follows:

(46)

$$\gamma := \beta \cdot R_V$$

In terms of diode parameters and physical constants,  $b$  can be expressed as:

(47)

$$\beta := \frac{q}{2 \cdot n \cdot k \cdot T} \cdot \left( \frac{1}{1 + \frac{R_S}{R_j} + \omega^2 \cdot C_j^2 \cdot R_S \cdot R_j} \right)$$

where:

- $q$  = electronic charge
- $n$  = ideality factor
- $k$  = Boltzman's constant
- $T$  = absolute temperature
- $C_j$  = junction capacitance
- $R_S$  = series resistance
- $R_j$  = junction resistance

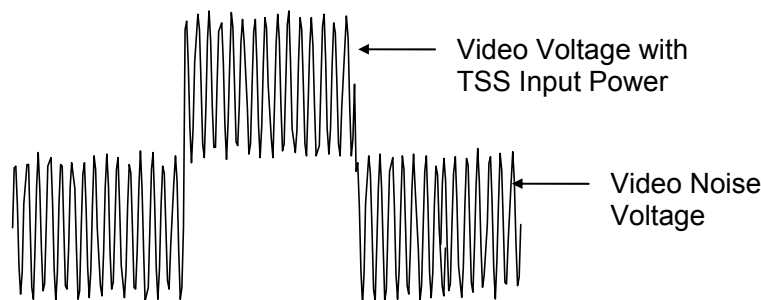
### D. NOMINAL DETECTABLE SIGNAL (NDS)

The nominal detectable signal (NDS) is the RF power level that must be applied to the detector diode so that the video power out of the detector is 3 dB higher than the video output noise level. NDS is a measure of the maximum useable sensitivity of a video detector.

### E. TANGENTIAL SIGNAL SENSITIVITY (TSS)

TSS is the most common sensitivity rating for detector diodes. The measurement is performed as follows.

An observer sets the detector's pulsed input power level to a value where, *in his opinion*, the video noise voltage peaks as observed on an oscilloscope with no signal present are the same level as the lowest noise peaks in the video signal when an RF input signal pulse is incident on the detector. Obviously, this TSS measurement technique is inherently subjective. Figure 19 is a representation of the TSS level measurement.



**Figure 19. Representation of TSS Measurement**

In order to eliminate the subjectivity of the TSS measurement, diode manufacturers define the TSS signal level to be when the video output signal is 8 dB greater than the video noise signal.

### F. FIGURE OF MERIT (FM)

Some old point contact diodes use a figure of merit (FM) to characterize their sensitivity. The FM is as follows:

(48)

$$FM := \frac{\gamma}{\sqrt{R_V}}$$

This figure of merit does not consider shot and 1/f noise introduced by the bias current and therefore is of limited value for describing Schottky barrier detectors.

Using FM, TSS for a given detector-video amplifier combination can be expressed as:

(49)

$$P_{TSS} := \frac{3.22 \cdot \sqrt{B \cdot F_V}}{FM} \cdot 10^{-7}$$

where:

B = bandwidth of video amplifier expressed in Hz

F<sub>V</sub> = noise figure of video amplifier expressed as a ratio

P<sub>TSS</sub> = power level at TSS expressed in mW

Figure 20 shows P<sub>TSS</sub> versus video bandwidth for two values of FM, 220 and 130, for a video amplifier with a noise figure of 3.5 dB.

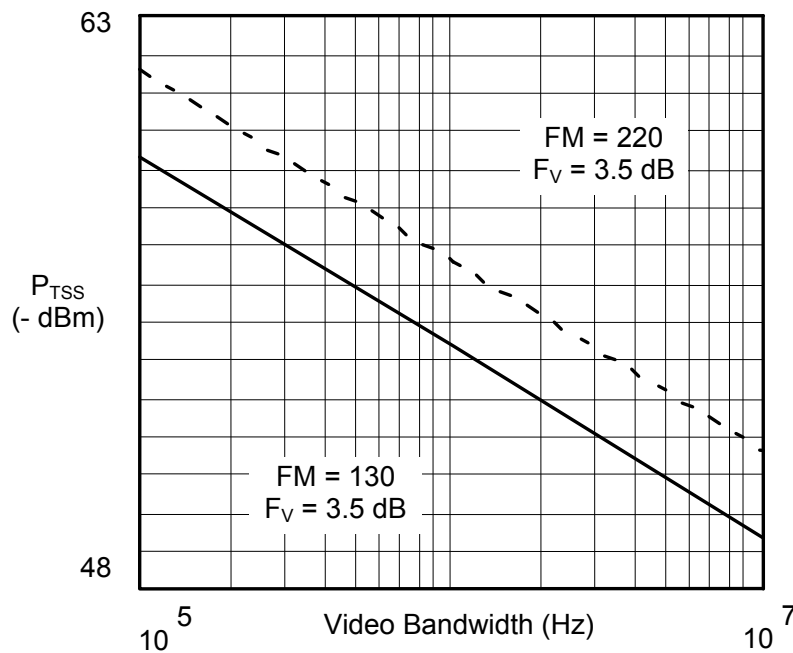


Figure 20. **P<sub>TSS</sub> vs. Video Bandwidth for Two Values of Detector Diode Figure of Merit, FM.**

## G. VIDEO BANDWIDTH

Although the detector diode itself may have a wide bandwidth capability, the circuit in which the detector diode is used will determine the video bandwidth of the overall detector. The typical detector circuit, shown in Figure 15, has its low frequency video response limited by the Inductance of the RF choke and the series coupling capacitor to the video amplifier. The high frequency video response is limited by the amplifier input impedance and the RF bypass capacitance. The upper frequency 3 dB roll off point is given by:

(50)

$$f_{3dB} = \frac{R_V + R_A}{2 \cdot \pi \cdot R_V \cdot R_A \cdot C_T}$$

where:

$R_V$  = detector diode video resistance

$R_A$  = amplifier input resistance

$C_T$  = sum of amplifier input capacitance and capacitance of RF bypass capacitor

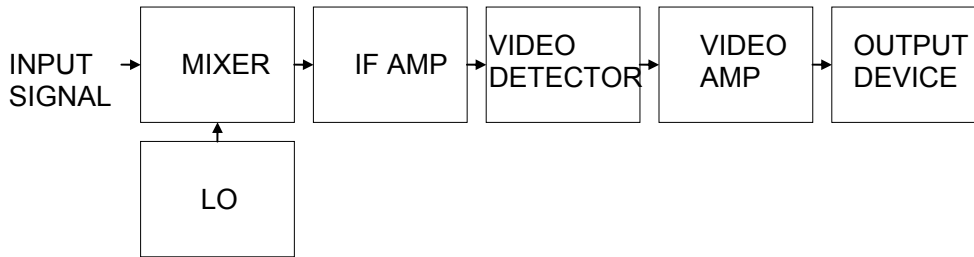
## IV. Comparison Of Mixers And Detectors For Receiving Systems

### A. CHOICE OF MIXERS VS. DETECTORS

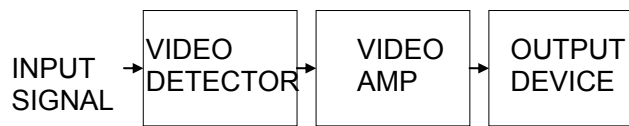
Mixers and detectors both downconvert microwave signals so that they may be displayed or processed further. Low noise amplification (up to 100 dB) is more readily achieved at VHF and below than at microwave frequencies.

Most mixer (superheterodyne) systems use IF amplification at an intermediate frequency (30 - 200 MHz) and then use a second down converter such as a video detector to recover the modulating signal that was superimposed on the microwave carrier. Such a superheterodyne detection system is shown in Figure 21. A microwave receiver with 10 dB noise figure and 1 MHz IF bandwidth would have a maximum sensitivity of - 104 dBm.

A single detection system is shown in Figure 22. Such a system, using only video amplification, can achieve a tangential signal sensitivity (TSS) of perhaps - 60 dBm for a 1 MHz video bandwidth compared with the - 104 dBm for the super heterodyne system. However, the single detection system has the advantage of simplicity, low cost and potentially wide bandwidth.



**Figure 21. Superheterodyne Detection System**



**Figure 22. Single Detection System**

## V. Mixer Circuits

There are a large number of circuits that use Schottky mixer diodes. These circuits were developed to enhance system performance and to optimize a specific design characteristic such as the receiver’s third order intercept point or its operating bandwidth.

The following sections describe a number of these circuits briefly and give advantages and disadvantages of each circuit.

These circuits include:

- |                            |                               |
|----------------------------|-------------------------------|
| Single Ended Mixers        | Balanced Mixers               |
| Double Balanced Mixers     | Double-Double Balanced Mixers |
| Image Reject Mixers        | Subharmonic Mixers            |
| Image Recovery Mixers      | Phase Detectors               |
| Bridge Quad Mixers         | Sampling Circuits             |
| Frequency Multipliers      | Quadrature Phase Modulators   |
| Single Sideband Modulators |                               |

## A. SINGLE-ENDED MIXERS

The simplest and least expensive form of diode mixer is a single ended design which employs a single Schottky barrier diode (see Figure 23). LO power is applied through a directional coupler. The coupler increases the conversion loss and the required LO power. The match between the diode and the input transmission line determines the SWR for the signal input, as well as the amount of LO power appearing at the signal input port. The amount of LO power seen at the signal port is determined by the directivity of the coupler and the diode's impedance match. It is measured in terms of LO-to-RF isolation. High isolation is important in many receiver applications since the LO can be radiated by the receiver antenna.

The IF output signal is extracted from the diode by a low pass filter, eliminating the RF and LO signals. A high pass filter is needed on the RF side of the diode to prevent loss of IF energy in this direction. This filtering requires that the IF frequency always be less than the lowest RF or LO frequency. If the IF response is extended down to DC, the DC component of the diode's voltage will be present at the IF port. All other harmonics and intermodulation products will also be present and must also be suppressed by filtering, as required by the receiver system design. Another drawback to a single ended mixer is that amplitude variations (AM noise) present in the LO source are not suppressed. This noise can increase the receiver noise figure if any AM noise signals are present at the signal and image frequencies.

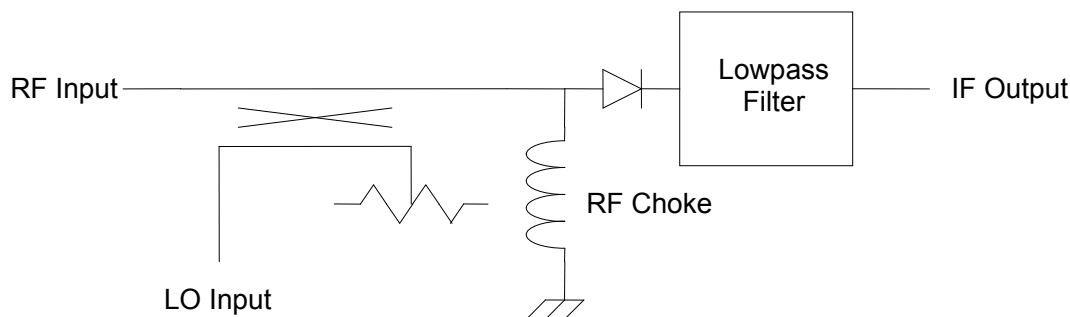


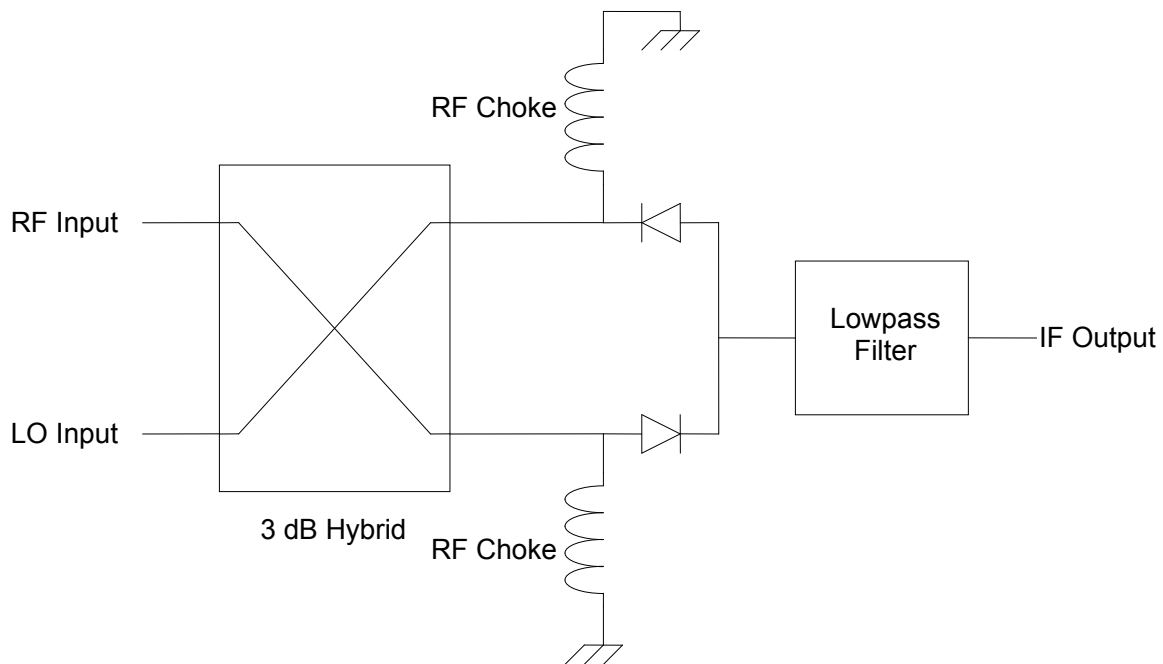
Figure 23. Single Ended Mixer Schematic

## B. SINGLE BALANCED MIXERS

Improvement upon the single ended mixer can be obtained by combining two Schottky mixer circuits in a balanced configuration where the two diodes are driven in opposite phase (see Figure 24). A 3 dB hybrid is used to supply the RF and LO power to the two mixer diodes. This hybrid serves as a balun transformer for the diode pair



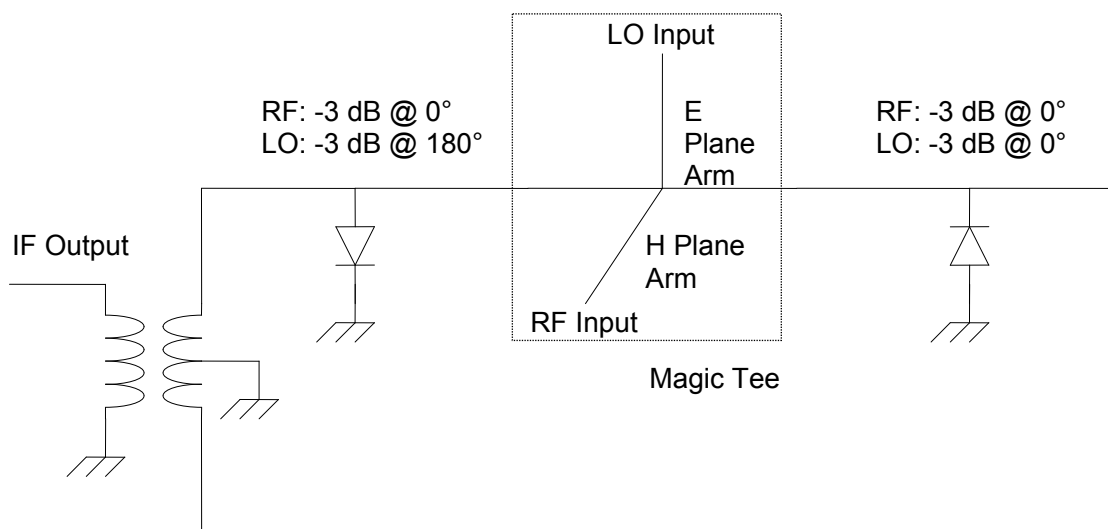
The IF signal is taken from the virtual ground point. A single balanced mixer can use either a  $90^\circ$  or  $180^\circ$  hybrid. Each type of hybrid has certain advantages and disadvantages, but both balanced designs offer improved performance over single ended mixers since spurious responses are reduced, the DC component at the IF output is canceled and there is convenient separation of the LO and RF inputs.



**Figure 24. Balanced Mixer**

### 1) Balanced Mixers Using $180^\circ$ Hybrids

The first approach combines two single ended mixers with the diodes in parallel and  $180^\circ$  out of phase. The hybrid suppresses the even harmonics of one of the input signals (LO or RF). It is usually designed to suppress the harmonics of the LO signal because normally the LO signal is at a much higher power level than the received RF signal. The degree to which the harmonics and intermodulation products are suppressed depends upon the balance of the hybrid and how closely the mixer diodes are matched to each other

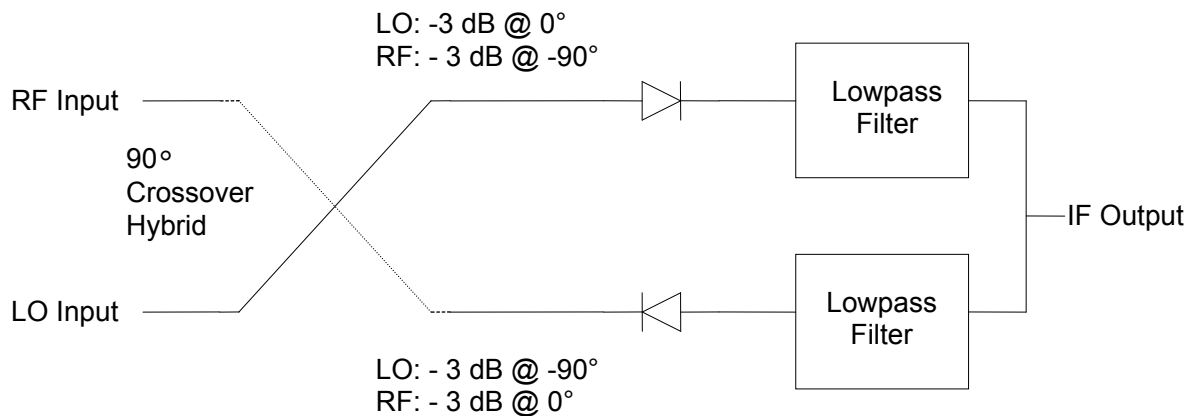


**Figure 25. 180° Balanced Mixer Utilizing a Magic Tee as the Balanced Coupling Mechanism**

The properties of the 180° hybrid determine the characteristics of this class of mixer. If the two output arms of a 180° hybrid are terminated in identical impedances, all reflected power is directed back to the input port. With good diode balance the reflected LO power appears at the LO port and not at the signal port. As a result, LO-to-RF isolation of the 180° balanced mixer is normally 20 dB or greater. This same property of the hybrid, however, causes the SWR for both LO and signal ports to depend on the diode match to the transmission line, (as in the case of a single ended mixer). SWR is typically 2.0:1 unless care is taken to carefully match the impedance of the diodes to that of the transmission line. As discussed with the single ended mixer, appropriate filtering is needed before and after the diodes to separate RF and IF frequencies in order to obtain optimum conversion loss.

## 2) Balanced Mixers Using 90° Hybrids

Balanced mixers designed with 90° hybrids exhibit significantly different RF properties than those using 180° hybrids. The achievement of harmonic and intermodulation suppression for this mixer type is more complicated and depends on the particular intermodulation product of interest. In general, there is suppression of harmonics and intermodulation products of both RF and LO signals.



**Figure 26. 90° Hybrid Mixer**

Equal matched terminations at the 90° output ports of the hybrid result in all the reflected power being directed to the fourth, normally isolated port. Good diode balance, therefore, leads to a low SWR (typically less than 1.5:1) at either the signal or LO ports. The LO or RF isolation depends on the match between the diode and transmission line impedances. This match is normally poor (typically 7 dB return loss) unless careful impedance matching is undertaken at the diode ports.

Appropriate filtering is needed at both the RF and IF ports to separate the RF and IF frequencies for optimum conversion loss.

The 90° hybrid-based mixer has been widely used for broadband (octave bandwidth) single balanced mixer designs because the 90° hybrid is relatively easy to fabricate in coaxial, stripline or microstrip transmission media.

In summary, for the 180° hybrid mixer the RF-LO isolation is less dependent on the match of the coupled arms at the expense of SWR and vice versa for the 90° hybrid mixer.

### 3) Balanced Mixer LO Requirements

Since two mixer diodes are employed in a single balanced mixer, twice as much LO power is required as for a single ended mixer. The noise figure of a balanced mixer will be reduced when compared to the single ended mixer if AM noise from the local oscillator is present at the signal frequency. LO AM noise is canceled at the IF port of a balanced mixer if the diodes are well matched.

The matching criteria for diodes are similar. The IF impedance and conversion loss at the diodes' bias current are determined by the local oscillator power. These diode characteristics are normally specified at approximately 1 mA rectified current.

### C. DOUBLE BALANCED MIXERS

A double balanced mixer consists of two single balanced mixers connected in parallel and 180° out of phase with each other at the RF frequency, as shown in Figure 27. At low frequencies (less than 3 GHz), conventional center tapped toroidal transformers are used for balun transformers to apply the LO and RF signals and for extraction of the IF signal. The transformers convert the single ended LO and RF input ports to a balanced configuration. The two baluns also isolate the LO and RF ports from each other and from the IF port. The IF signal is taken from the virtual ground point of the four diode ring. This results in high isolation between the LO and the IF ports. At higher frequencies, (above ~3 GHz), the baluns are generally implemented as distributed circuits. The IF signal then must be selected by suitable filtering. The symmetry provided by this topology ensures better isolation between the RF and the LO ports and the suppression of even harmonics of both the RF and the LO input signals if all the diodes as well as the circuit are well balanced.

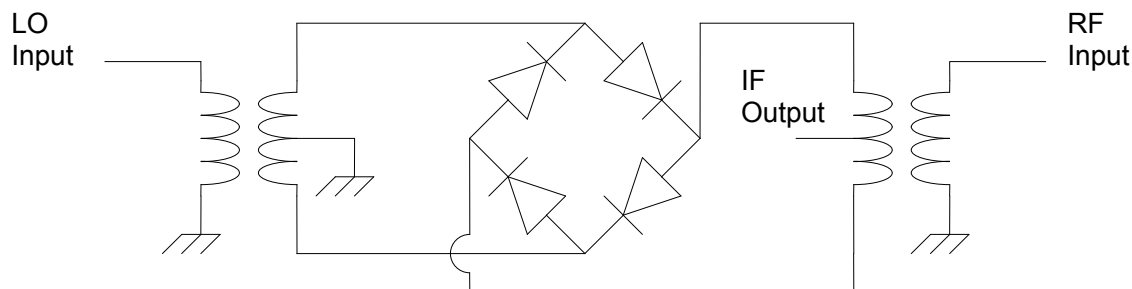


Figure 27. Double Balanced Mixer

In a double balanced mixer the diodes in the ring quad must have very similar junction capacitances and forward voltage drops at the current set up by the LO signal, which is usually in the 1 to 10 mA range.

In operation, a double balanced mixer can be visualized as a simple SPDT switch which switches the RF signal on and off at the rate of the LO frequency. This switching is controlled by the local oscillator signal voltage. When it is applied to the diode ring it will drive pairs of diodes alternately into low impedance (conduction) or high impedance (reverse bias) depending upon the instantaneous polarity of the LO signal. Since the center of the LO transformer's secondary is grounded, the action of the LO signal voltage is to alternately ground the midpoint of the two conducting diodes. Thus the effect of the local oscillator signal on the diode quad is to alternately ground each end of the secondary of the RF input. This switching action "mixes" the two signals (RF & LO) to produce both desired and undesired mixing products of the RF and LO signals and their harmonics. Unlike the case of the balanced mixer where the RF signal is switched off during half of the LO cycle, in a double balanced mixer the RF signal voltage is connected to the output during both halves of the LO cycle. The result is a considerable difference in the output waveform of the single and double balanced mixers. At the IF port of a double balanced mixer even order harmonics of the LO or RF signals are either eliminated or appear with their amplitudes substantially reduced. Double balanced mixers also reduce the

power level of the two tone intermodulation products because less RF voltage appears across each diode for a given RF input power level.

### 1) Distortion Products

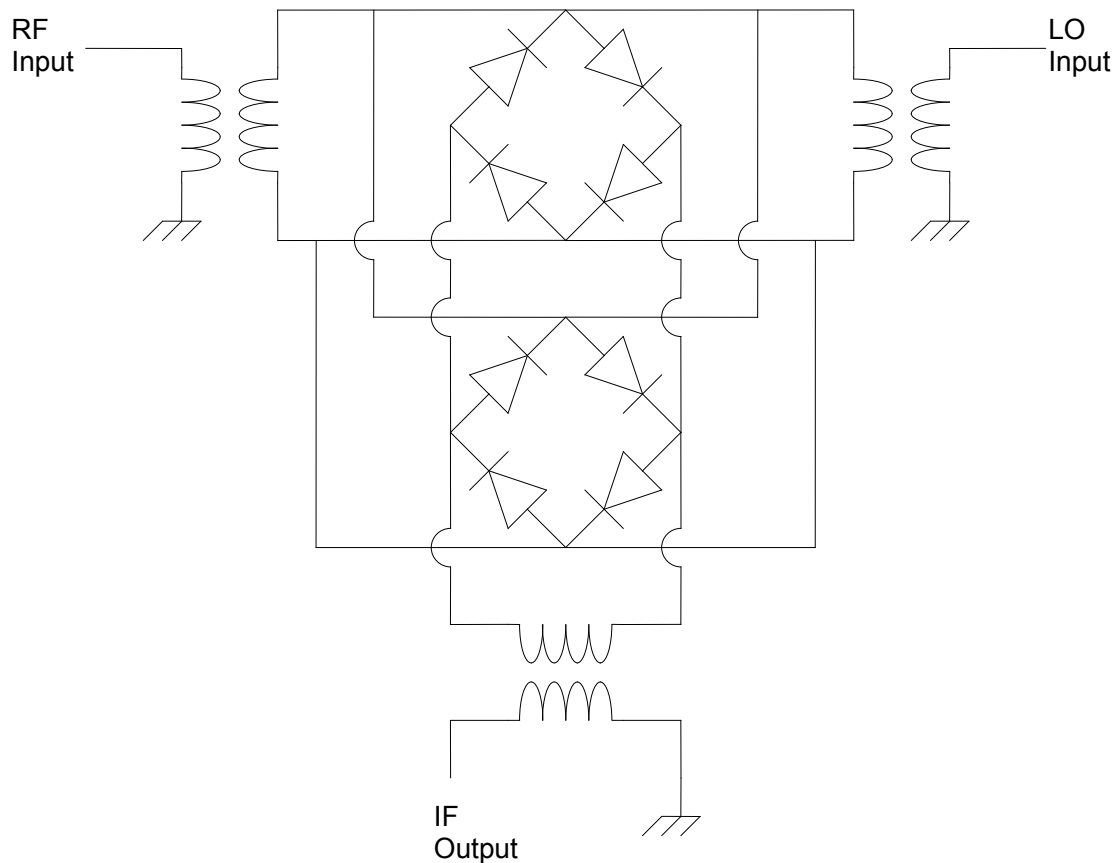
There are several forms of distortion products which occur in all types of mixers. With a single frequency input to the mixer, single tone distortion or harmonic intermodulation distortion produces signals which distort the desired IF output. For systems with a narrow band IF the single tone distortion products will occur primarily out of the IF band. However, for wide band IF mixers, some of these intermodulation products can be troublesome. As the power of the single RF input increases, conversion loss compression may occur. This produces a second form of distortion. If the power level of the RF signal to the mixer is kept well below the power level of the LO, (i.e. ~15 dB down) the mixer's RF-to-IF conversion loss is constant and independent of the RF signal drive power. However, as the RF signal power increases and approaches the same power level as the LO, conversion loss compression occurs, because the RF signal will also forward bias the diodes in the ring. This distorts the normal phase relationships of the mixer which were established by the LO signal. This distortion occurs when the RF voltage is large enough to bias the nonconducting diodes into conduction or the conducting diodes into nonconduction. Increasing the LO drive power incrementally does not significantly improve the conversion loss 1 dB compression point since a great deal more LO power is required to sufficiently increase the reverse bias voltage across the nonconducting diodes. Use of a dual ring quad which uses two diodes connected in series in each leg of the ring increases the mixer's conversion loss compression level because the effective barrier voltage of each leg of the ring quad has been doubled. Each leg's apparent reverse breakdown voltage is also increased to the sum of the individual diode's breakdown voltages. However, this type of double balanced mixer will usually have a higher conversion loss due to the increase in series resistance of each leg of the quad.

When two RF input signals are present simultaneously, two tone intermodulation occurs. This form of distortion is particularly troublesome because some of the intermodulation products are within the desired IF bandwidth. These distortion components are not easily removed by symmetry or filtering.

The double balanced mixer will provide cancellation of AM noise from the local oscillator and LO-to-RF isolation equal to that of a single balanced mixer without the bandwidth restrictions imposed by the 3 dB coupler. Higher dynamic range is also obtained with a double balanced mixer due to the higher levels of LO power which can be used with four diodes. Lower IF impedance which is the result of four diodes in parallel is another benefit. However, impedance matching of the RF and LO ports is more difficult because each port sees the impedance of two diodes in series. This impedance can be quite high.

#### D. DOUBLE-DOUBLE BALANCED MIXERS

When high IF frequency mixers are required, the best mixer choice is a double-double balanced configuration (see Figure 28). The inclusion of the IF balun circuit limits IF current flow through the mixer and decreases the conversion loss compared to other mixer types. The double-double balanced mixer divides the signal power between twice as many diodes as a double balanced mixer and thereby achieves an improvement of 3 dB in dynamic range over the double balanced mixer. Because this circuit requires the extra balun and diode ring, circuit size and parts count are both increased. The double-double balanced mixer also requires 3 dB greater LO power than the double balanced mixer. Because the two diode rings are connected in parallel to the IF and LO baluns, the SWR of a double-double balanced mixer is lower than that of a double balanced mixer.



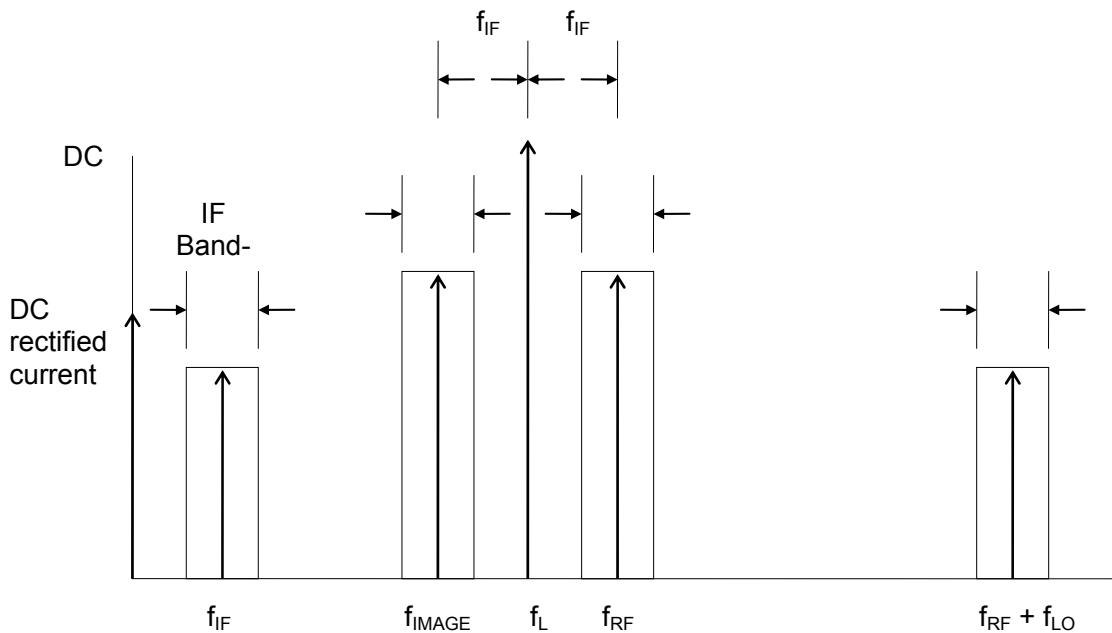
**Figure 28. Double-Double Balanced Mixer**

Another advantage of the double-double balanced mixer is that the IF output frequency can be as high or higher than the LO frequency because of the separation of the LO and IF balun transformer.

### E. IMAGE REJECT MIXERS

For a given IF frequency,  $f_{IF}$ , a signal either above or below the LO frequency,  $f_{LO} \pm f_{IF}$  produces an IF output. If one of these high frequency signals is considered to be the desired signal, then the other is commonly termed the image signal. All mixers normally have four output frequency components:

1. the signal frequency ( $f_{RF}$ )
2. the LO frequency ( $f_{LO}$ )
3. the desired lower frequency ( $f_{RF}-f_{LO}$ )
4. and the LO frequency plus IF ( $f_{LO}+f_{IF}$ ), which is called the image frequency,  $f_{IM}$  (see Figure 29).

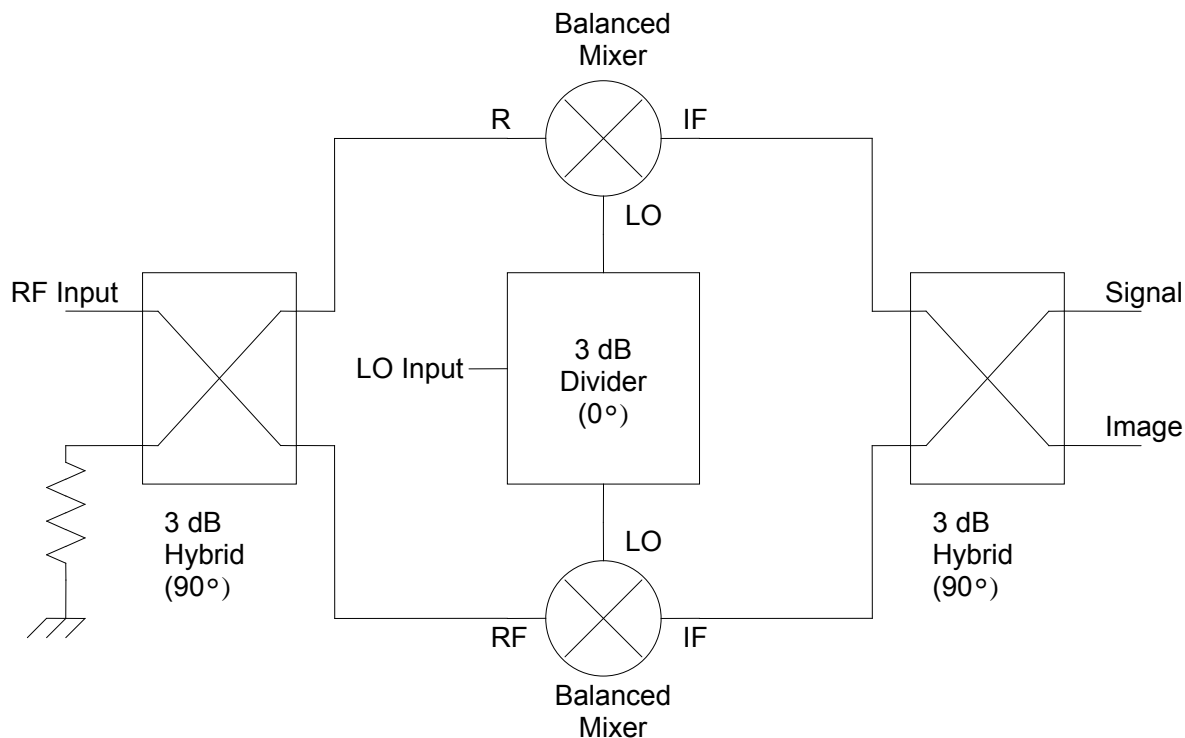


**Figure 29. Frequency Components of a Mixer**

In many applications, it is desirable to either eliminate or distinguish the image response from the desired signal response. If the IF frequency is sufficiently high and the RF bandwidth narrow enough so that the signal and image frequency bandwidths do not overlap, the image response can be eliminated by appropriate input filtering. This type of design is suitable for narrow band systems where a high degree of image rejection is desirable. For broadband applications, especially octave bandwidth mixers, filtering cannot be used for image rejection. In this case the  $f_{IM}$  can be rejected by phasing techniques.

Two single balanced mixers are frequently combined to form an image rejection mixer. A  $90^\circ$ , 3 dB power divider or hybrid is used to divide the RF signal into two quadrature signals which are input to the two mixer RF ports. The LO signal to the mixer is applied through an in phase power divider (see Figure 30). The IF outputs of each mixer are then combined through a  $90^\circ$  hybrid. With this circuit, the signal frequency response appears at one output of the IF hybrid and the image at the other. Either the signal or the image (upper or lower sideband) response can then be suppressed by terminating the appropriate IF output port. The degree of image rejection depends on the amplitude and phase balance between the two mixers.

In addition to image rejection, this type of mixer provides other attractive characteristics including good RF input SWR and a 3 dB improvement in RF signal power handling over other designs because the 3 dB hybrid splits the incoming RF signal without much degradation of noise figure. The LO power required is 3 dB higher than that for a balanced mixer also due to the power division. The conversion loss is usually a little higher due to the additional losses of the RF and IF hybrids. This type of mixer is very useful if an RF amplifier is used in front of the mixer, since the mixer can eliminate image noise amplified by the RF amplifier.



**Figure 30. Image Reject Mixer**

This same circuit can be used as a double sideband modulator by applying an LO signal to the LO port and the baseband modulation signal to the IF port. This results in a balanced modulated RF signal at the RF port. This RF signal will contain both modulation sidebands. The carrier (LO signal) will be suppressed and the conversion loss relationship and isolation characteristics of a double balanced mixer will still apply.



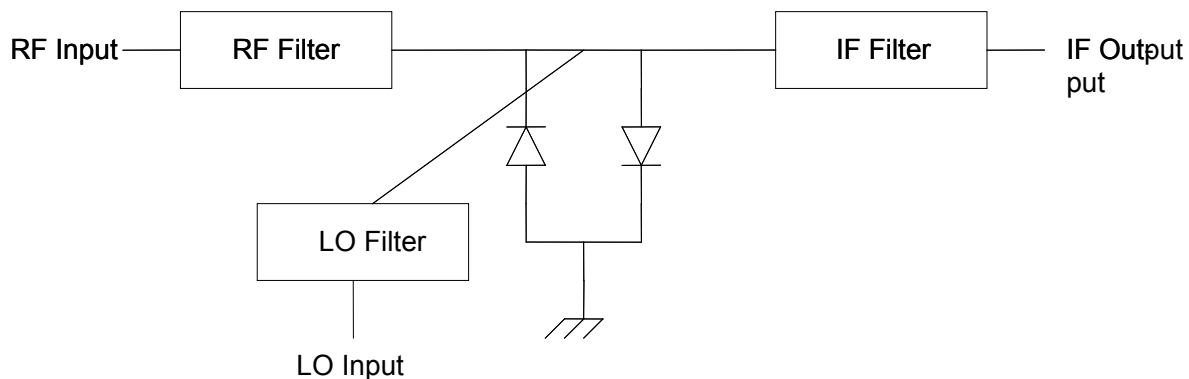
## F. SUBHARMONIC MIXERS<sup>2,3</sup>

Subharmonic mixers are advantageous to use at millimeter frequencies because the LO frequency can be one quarter to one half of the RF frequency. This allows the designer to use a more cost effective (lower frequency) local oscillator design.

In a standard single ended or balanced non-subharmonic mixer, the diode is pumped into its forward conduction by the LO once for every input cycle of the local oscillator waveform. The variation in the resistance of the mixer diode during forward conduction results in frequency mixing.

In a subharmonic mixer, two diodes in an anti-parallel pair (in parallel but connected anode to cathode) are used. The local oscillator sweeps each diode to a forward conduction once each LO cycle. This process requires that the local oscillator frequency be only half that which is needed for a comparable single ended mixer (see Figure 31). This process can occur at any even fraction of the LO frequency but as the fraction becomes smaller (i.e., 1/4, 1/8, ...) the effective LO power decreases.

Single ended subharmonic mixers use one antiparallel pair. Many of the mixing products are canceled by symmetry. All even LO  $\pm$  even RF products and all odd LO  $\pm$  odd RF products are canceled at all three ports. Only odd LO  $\pm$  even RF and even LO  $\pm$  odd RF are generated. Unfortunately the LO, IF and RF ports are not isolated. All have the same frequency outputs. Filters are needed for signal selection and isolation.



**Figure 31. Unbalanced Subharmonically Pumped Mixer**

A balanced subharmonic mixer which uses two antiparallel diode pairs helps alleviate the filtering problem (see Figure 32). The antiparallel diodes must be matched at the LO port. In the balanced design all the even LO  $\pm$  odd RF products are canceled at the LO port and odd LO  $\pm$  even RF products are canceled at the RF and IF ports. Therefore, LO-to-RF and LO-to-IF

isolation is inherent in this circuit. The desired even LO  $\pm$  odd RF products appear at both the RF and IF ports, necessitating filtering. Additional advantages of the balanced subharmonic mixer are that it has twice the conversion efficiency and half the IF intermodulation products as that of an unbalanced subharmonic mixer.

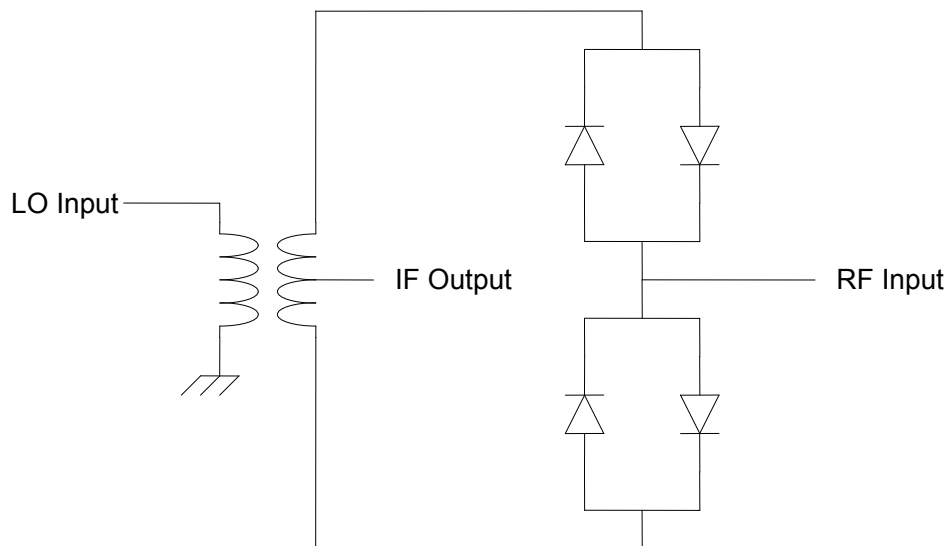


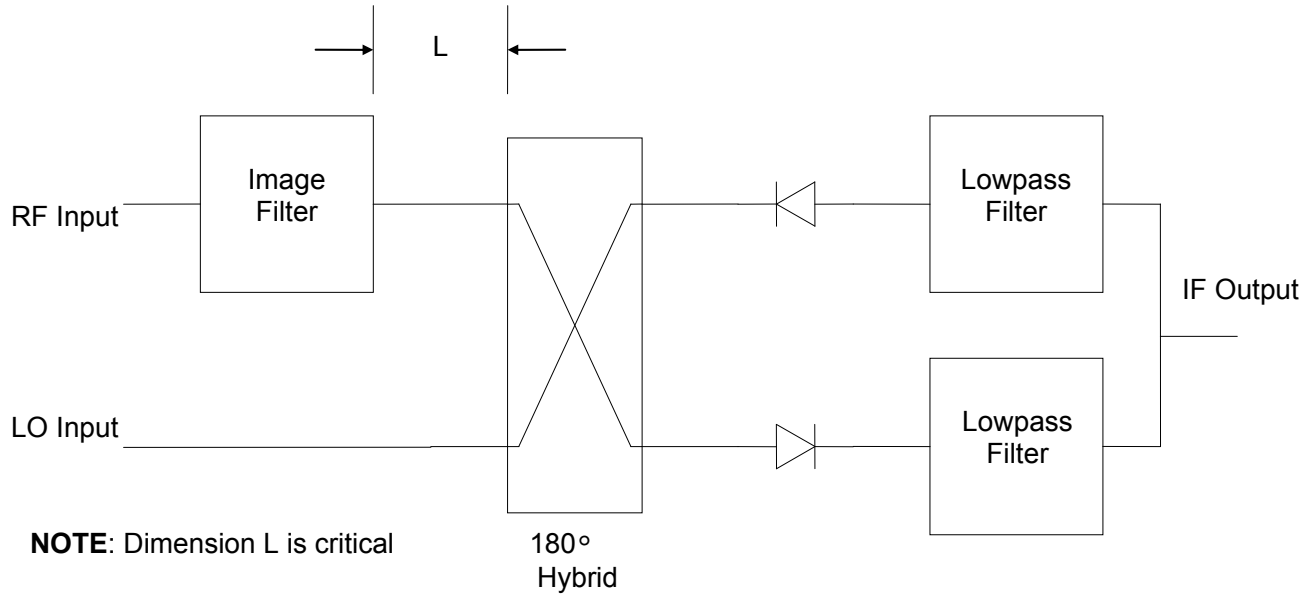
Figure 32. Balanced Subharmonically Pumped Mixer

### G. IMAGE RECOVERY MIXERS

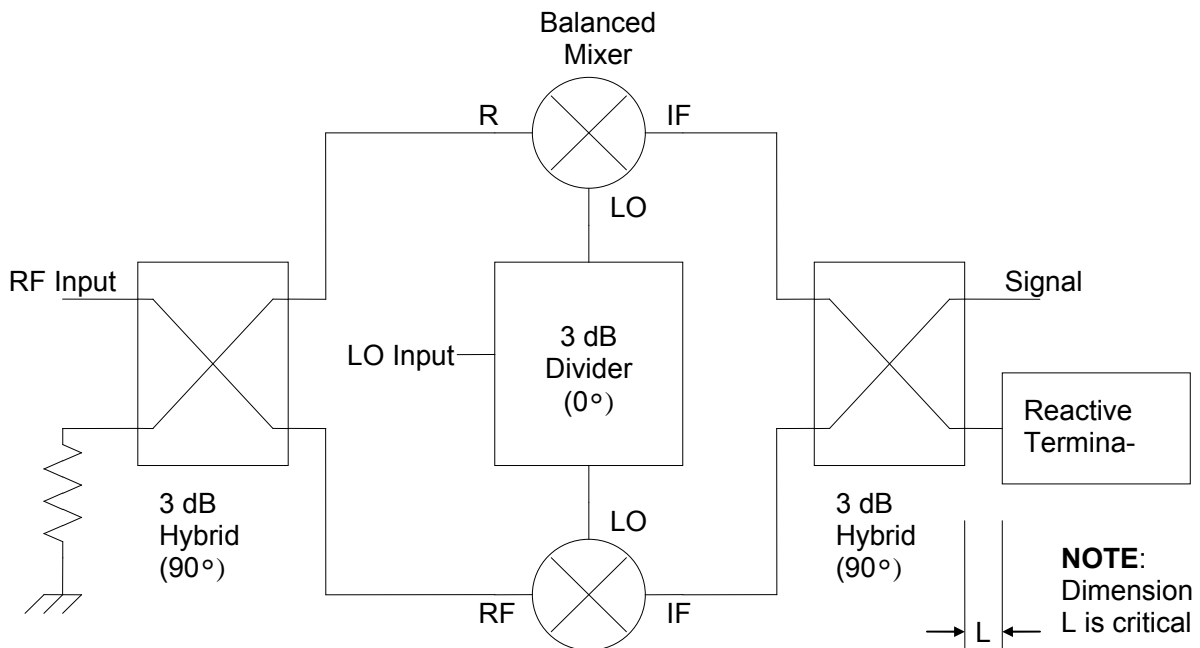
A mixer not only responds to a signal at the signal frequency, it also responds to energy at the image frequency ( $f_{IM}$ ). This is a major element in conversion loss and noise figure. This energy is produced by two mechanisms. If a signal is received at the upper sideband (i.e.  $f_{RF} = f_{LO} + f_{IF}$ ) it can mix with the second harmonic of the local oscillator signal to produce energy at the image frequency ( $f_{IM} = 2f_{LO} - f_{RF}$ ). A mismatch at the IF port can reflect part of the IF frequency output back into the mixer. This reflected signal can be combined with the LO signal to produce energy at the upper and lower sidebands ( $f_{LO} \pm f_{IF}$ ). One of these sidebands is the image frequency. In both cases, the unwanted energy at  $f_{IM}$  can be reconverted to produce additional power at the IF frequency. This will decrease conversion loss. This is the basic principle behind image recovery mixers. They are usually used for narrow bandwidth systems when the lowest conversion loss and noise figure are required.

The same methods used to reject  $f_{IM}$  of an incoming signal can be used for image recovery of an internally produced signal. When using a balanced mixer with a  $180^\circ$  hybrid, the generated image frequency appears at the RF input port. For narrowband applications, i.e., 1-2% bandwidth maximum, a filter at the RF input port can be used to reflect  $f_{IM}$  back into the mixer. By adjusting the electrical length between the filter and mixer, the image can be reflected back to produce IF energy in the proper phase. This will minimize conversion loss. This phasing of the

image is critical, since degradation in the conversion loss can also result from improper tuning. Normally the filter should appear as an open circuit to  $f_{IM}$  (see Figure 33).



**Figure 33. Image Enhancement Mixer**



**Figure 34. Image Enhancement Mixer with Reactive Termination**

## H. PHASE DETECTORS

The ring quad double balanced mixer can also be used as a phase detector. If RF signals of equal amplitude and frequency are applied to the LO and RF ports, a DC signal that is a function of the phase difference between the signals will appear at the IF port. The DC voltage is:

(51)

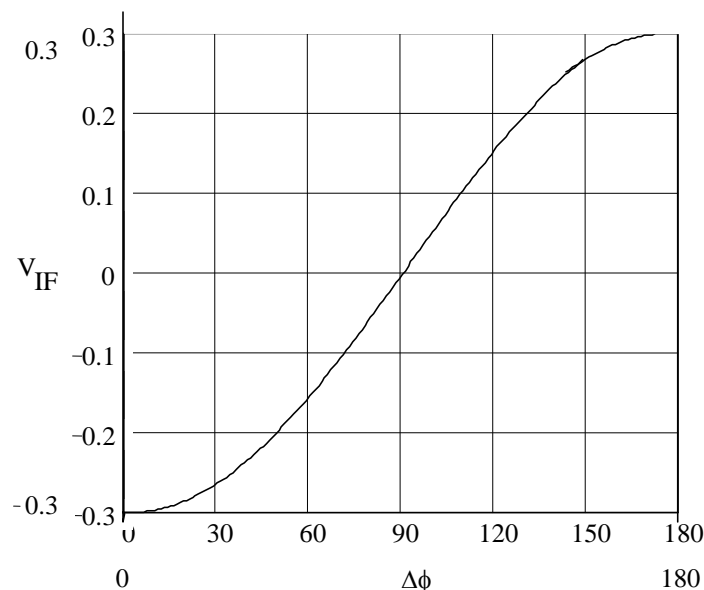
$$V_{IF} := V_{OFFSET} + V_1 \cdot \cos(\Delta\phi + \pi + \phi_{OFFSET})$$

The offset voltage,  $V_{OFFSET}$ , is an additive constant caused by unbalance in the diode ring or in the transformer structures,  $\phi_{offset}$  is the phase difference introduced by the mixer and  $\Delta\phi$  is the phase difference between the signals. With proper design,  $V_{offSet}$  and  $\phi_{offset}$  can be minimized so

(51A)

$$V_{IF} := V_1 \cdot \cos(\Delta\phi + \pi)$$

the constant  $V_1$ , will be a function of the RF signal amplitude and the IF load resistance. By operating the phase detector close to the zero crossing of the cosine function, the output voltage can be made approximately linear with the phase angle for small variations of the angle.



**Figure 35. Phase Detector Performance Curve Illustrating the Linearity of the Output Voltage**

## I. OTHER RING QUAD APPLICATIONS

Other applications of the double balanced mixer include attenuators, switches and biphas modulators. When an RF signal is applied to the LO port and no signal applied to the IF port, the RF output is isolated from the LO input by 20 to 30 dB. If a positive DC voltage is applied to the IF port, the RF applied signal will be transmitted to the RF port with a specific phase. If a negative voltage is applied at the IF port, the RF input signal will be transmitted to the RF port with the same amplitude but in the opposite phase. It is therefore possible to modulate the RF signal through 180° using alternate polarity DC voltage on the IF port. This modulating technique is very useful for digital communications where the IF signal (modulation) can be a stream of binary data.

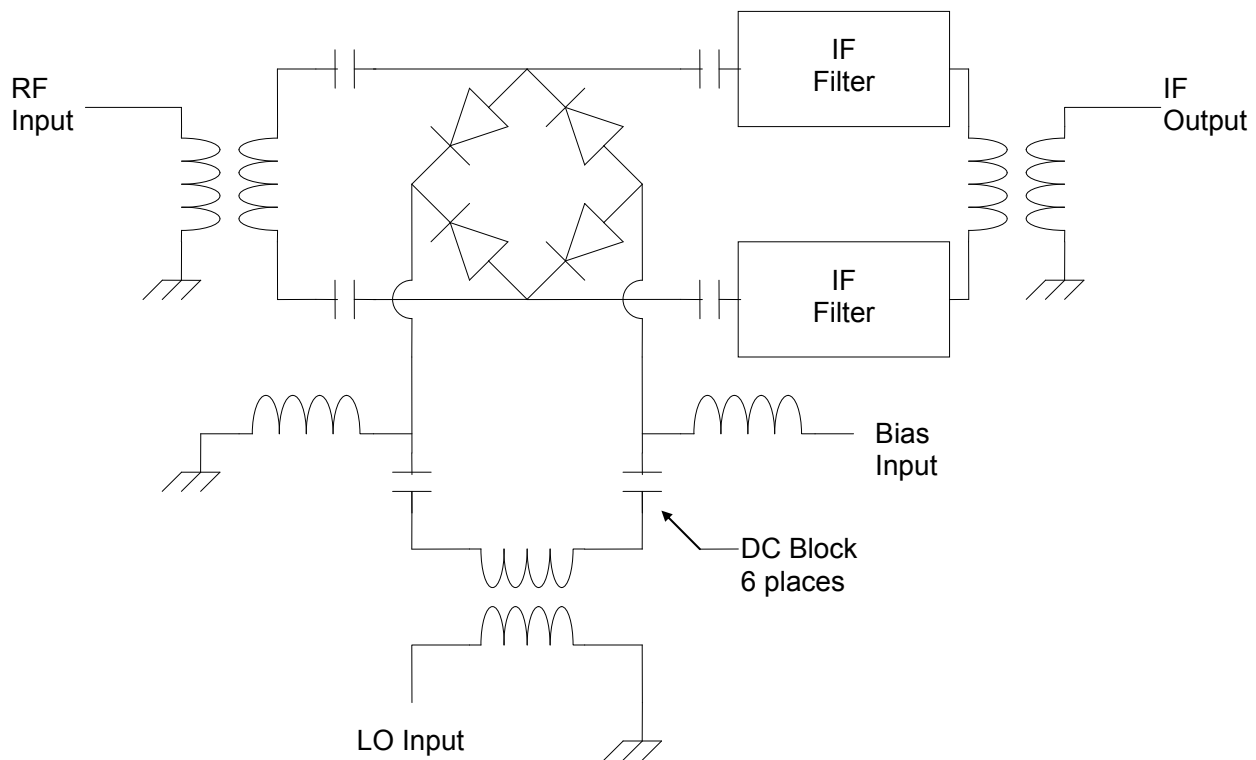
A double balanced mixer can be used as an attenuator by varying the level of the DC current applied to the IF port. This can control the level of the LO signal transmitted to the RF port. By changing the current level from 0 to 2 mA per diode a current controlled attenuator can be produced. The maximum attenuation will depend on the mixer's isolation. Values of 20 - 30 dB are normally obtainable.

## J. BRIDGE QUAD APPLICATIONS

### 1) Biasable Bridge Quad Mixers

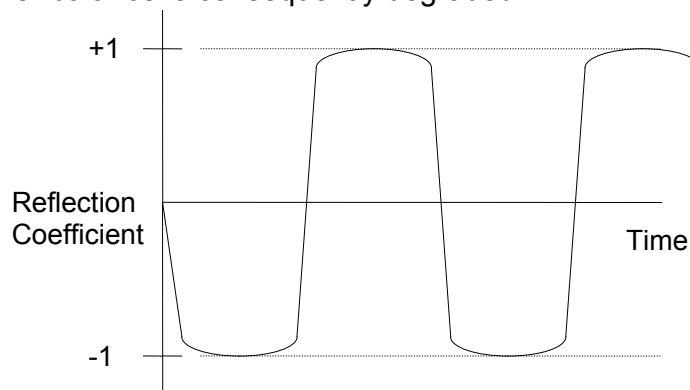
Bridge quads can be employed in biasable bridge quad Schottky diode mixers (see Figure 37). A biasable mixer is very attractive in applications where little LO power is available. In a bridge quad, the diodes are arranged in a rectifier ring (see Figure 37). The bridge quad circuit still provides good LO-to-RF isolation because, like the double balanced mixer, the RF and LO signals are applied at each other's virtual ground points. The bridge quad does not provide RF-to-IF isolation, because the IF must be coupled from the RF signal connection points. An IF filter is necessary to separate the RF and IF. Unlike a ring mixer, the biasable bridge quad mixer does not have the proper symmetry to suppress even harmonics of both the RF and LO. It will have spurious signal levels that are dependent upon which ports are chosen to receive the LO and RF signals. There is less distortion if the IF is taken from the RF signal connection ports. In the normal mode of operation, mixing products having even harmonics of the RF will be suppressed.

A biasable bridge quad mixer operates as follows. The LO pumps all four diodes in phase. The RF is short-circuited when the diodes are on (forward biased) and open-circuited when the diodes are off. In these two states the reflection coefficients are -1 and + 1, respectively. The output IF signal for a bridge quad mixer is the product of the RF incident signal and the periodic square wave reflection coefficient. Since the alternating reflection coefficient is "square like", (i.e. on or off) it produces reflection coefficients in the frequency domain of odd harmonics only as shown in Figure 37.



**Figure 36. Biasable Bridge Quad Mixer**

A potential problem with this circuit is that small differences in the I/V characteristics of the diodes can severely unbalance the mixer when it is operated under DC bias. The bridge is, in effect, two pairs of diodes, with each series-connected pair biased to exactly the same voltage. Because of the exponential dependence of the diode current upon voltage, it is possible for one pair of diodes to conduct much more heavily than the other, if the diodes are not identically matched. When the LO signal is applied, the diode pairs then have very different conductance waveforms and the mixer balance is consequently degraded.



**Figure 37. Biasable Bridge Quad Reflection Coefficient**

## 2) Bridge Quad Sampling Circuit

The bridge quad can also be used in sampling circuits. The bridge quad barrier height must be chosen so that the expected input signal will not cause the diodes to conduct in the forward direction under normal RF drive conditions. If the input signal does become too large, a reverse bias can be applied to the diodes to prevent them from conducting. Sampling of the signal is accomplished through the selective application of a short duration, high level DC pulse that will switch the diodes into conduction. Since symmetry is essential for both AC and DC balance, the diodes must be well matched. Symmetry of the diodes tends to cancel even harmonics of the RF input at the output. This results in a "clean" sampled signal. Because of their higher barrier heights and low series resistance, GaAs bridge quads are very suitable for use in sampling circuits.

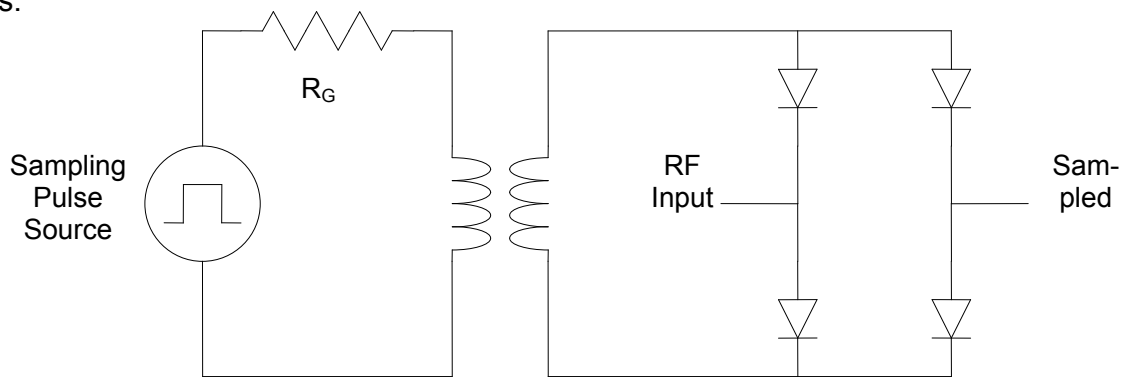


Figure 38. Bridge Quad Sampling Circuit

## K. FREQUENCY MULTIPLIERS

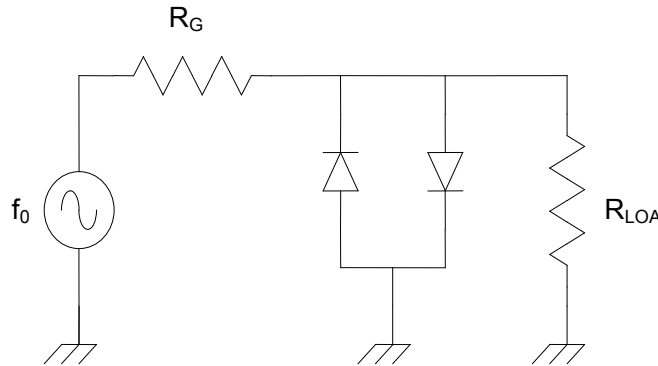
Schottky barrier diodes employed in a balanced configuration can be used as frequency multipliers (see Figure 39). The undesired frequencies (odd or even harmonics) are terminated by symmetry. Therefore, no additional circuit losses are incurred in the filtering out of unwanted harmonics. This helps minimize the conversion loss and provides a clean broadband signal with a minimum of spurious signals.

### 1) Even and Odd Harmonic Multipliers

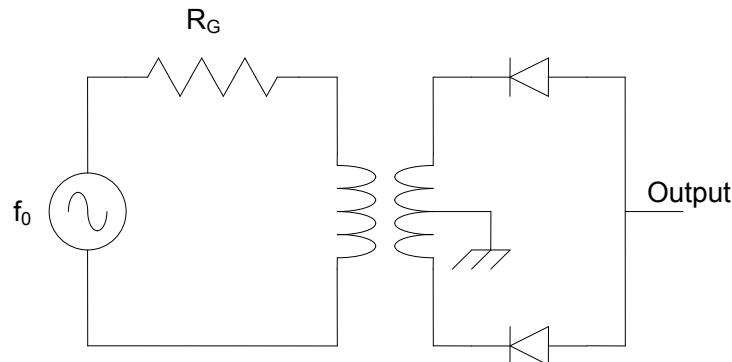
Both odd and even harmonic frequency multipliers can be built. Each design employs a pair of balanced, antiparallel, Schottky barrier diodes. The difference between the odd and the even harmonic multipliers lies in the way in which the diodes are mounted. To achieve the odd harmonic multiplier two diodes are set in anti-parallel to ground and to both the input and the output signals. When building an even harmonic multiplier, the diodes are set in anti-parallel to the input signal, but they are in series to the output signal (see Figures 39a and 39b).

When driven by the incoming RF signal the odd harmonic multiplier produces only odd harmonics of the input signal frequency. There is no DC voltage generated across the diode pair.

Because of this there is no diode DC power dissipation loss and even harmonics are canceled by symmetry. In a similar manner the even harmonic multiplier produces only even harmonics with intrinsic odd harmonic cancellation.



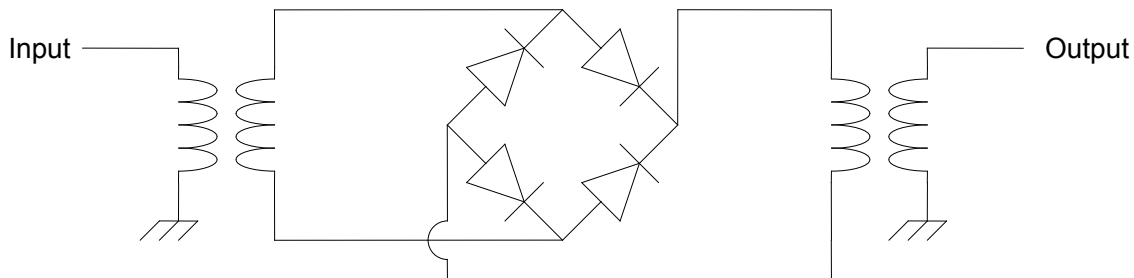
**Figure 39a. Odd Harmonic Multiplier**



**Figure 39b. Even Harmonic Multiplier**

### 2) Bridge Doubler

These design principles can also be applied to more than two diodes in a balanced configuration. For example, bridge quads are commonly employed in bridge doubler circuits. i.e.



**Figure 40. Bridge Doubler**



Analysis of the harmonic content of the circuit reveals that only even harmonics of the input signal are present at the output. The suppression of the odd harmonics of the input signal is dependent upon the balance of the diodes in the bridge circuit (see figure 40).

Typical doubler conversion losses are in the range of 6 to 10 dB and dependent on drive levels. Normally a bridge quad doubler can be used up to input powers of 15 to 20 dBm. Gallium arsenide Schottky diodes are very well suited for bridge doublers because they usually have higher reverse breakdown voltage and lower series resistance than similar silicon diodes.

## L. QUADRATURE PHASE MODULATORS

Quadrature phase modulators are used in many high capacity data links and radio relay equipment. A quadrature phase modulator can be built as shown in Figure 41. It uses two bi-phase modulators (as described in Section I), a  $90^\circ$  hybrid and an in-phase combiner.

The carrier input is split by the hybrid and one port is delayed  $90^\circ$ . The modulation inputs are DC voltage pulses through the IF port. These pulses are the  $0^\circ$  and  $180^\circ$  modulation inputs. The lower modulation signal is delayed  $90^\circ$  (see figure 41). When combined, the signals produce the  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  states. The output of this circuit is a single sideband, suppressed carrier signal. All even and half of the possible odd harmonics are also suppressed.

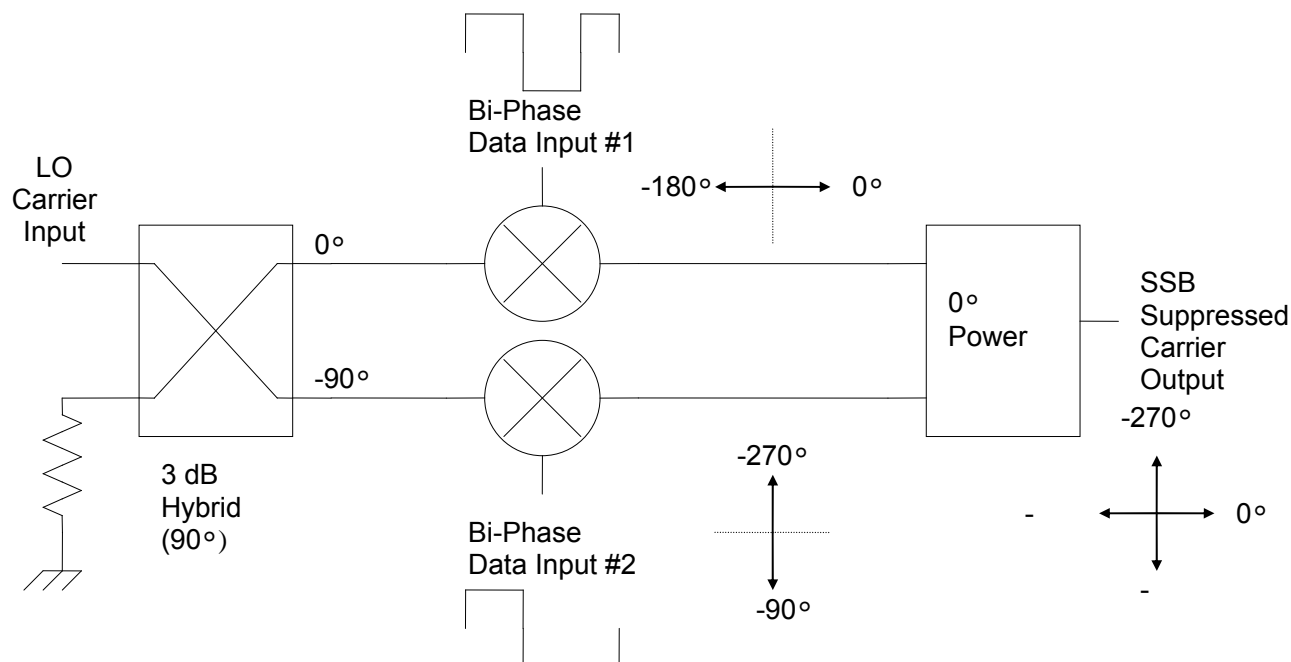


Figure 41. Quadrature Phase Modulator Block Diagram

### M. FREQUENCY DETERMINATION--A QUADRATURE IF MIXER

A quadrature IF mixer (QIFM) can be used to determine the relative frequency of two RF signals or it can determine if the frequency of an RF signal is above or below that of a known LO signal.

A quadrature IF mixer can also be used as a phase detector if the two input signals have the same frequency. Figure 42 shows the circuit of a typical QIFM.

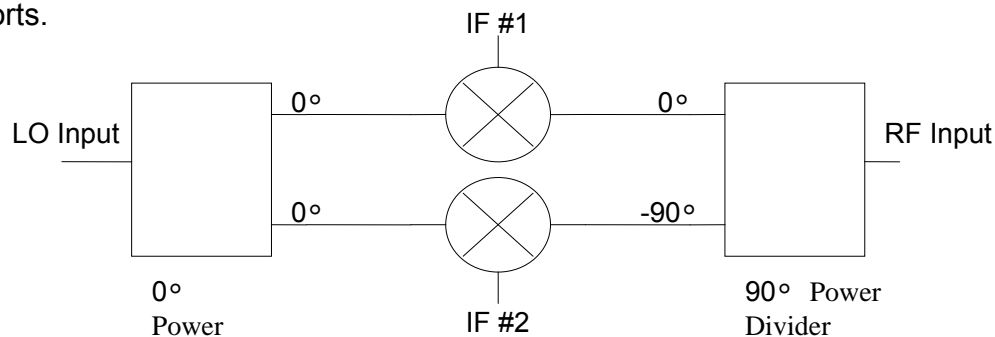
The LO or known frequency is fed through an in-phase power divider to two quadrature mixers. The unknown frequency enters through a 90° power divider. Because the unknown signal enters the two mixers 90° out of phase, the signals available at the IF ports will also vary by 90°. If the unknown frequency is above the LO frequency, the signal from IF port 2 will lead the signal from IF port 1. If the unknown frequency is below the LO frequency, the signal from IF port 2 will lag the signal from IF port 1.

When the two input frequencies are equal the IF ports will only produce a DC voltage with the output voltage proportional to the relative phase angle between the signals (see Figure 35).

There are many uses for QIFMs. The major applications are in phase detection, Doppler radar systems, and quadrature phase shift keying (QPSK) demodulators. When the QIFM is used as a phase detector, the IF outputs are proportional to the sine and cosine of the phase difference of the two input signals. If these two signals are applied to the horizontal and vertical inputs of an oscilloscope, the result will be a polar display of the magnitude and phase difference of the two RF input signals.

When used in a Doppler radar system, the IF output of the QIFM will indicate the velocity of a target, since with an approaching target, the return signal is higher in frequency than the LO and lower in frequency when the target is receding from the radar system.

Another use of the QIFM is as a demodulator for digital QPSK signals. A sample of the carrier signal is injected into the LO port as a reference. The QPSK modulated signal is applied to the RF port. The two original bi-phase digital signals, 0°, 180° and 90°, 270° are extracted from the IF output ports.



**Figure 42. Block Diagram of the Quadrature IF Mixer<sup>5</sup>**

## N. SINGLE SIDEBAND MODULATORS<sup>12</sup>

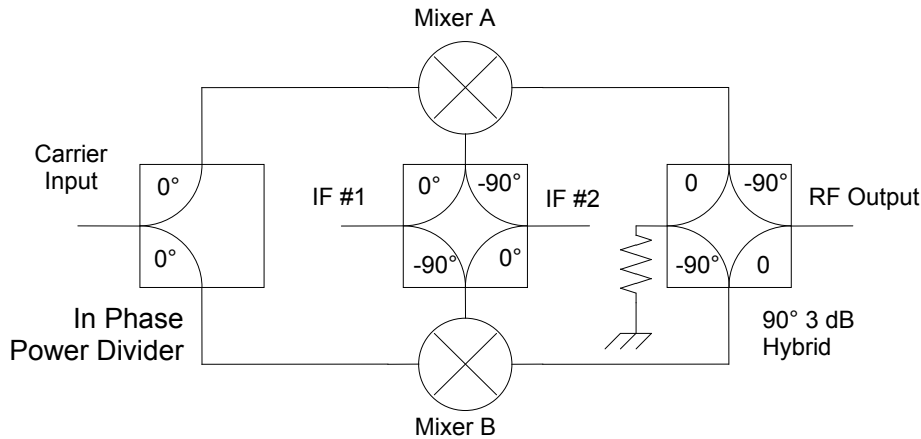
Single sideband (SSB) modulators are used to modulate a carrier signal with a low frequency baseband signal and transmit only one sideband.

Single sideband modulators are often used in digital cellular telephone transmitters. They provide the interface between the digitally encoded voice and the RF transmission channel by converting the I and Q signals from the channel codec into the RF signal that the telephone transmits. Essentially, they modulate a high frequency carrier with the lower frequency voice data to create all forms of analog and digital modulation. Where practical, a direct modulation technique is preferred because it eliminates much of the filtering and LO requirements necessary for multiple up conversion transmitter topologies.

The output signals of the symbol modulator, which are two orthogonal analog signals: an in-phase (I) signal and a quadrature (Q) signal to accommodate phase information, are input to the I and Q ports of the single sideband modulator. The amplitudes of these signals are analogous to the X and Y values in a Cartesian plane. The modulation process is analogous to a rectangular-to-polar coordinate conversion with the vector-summed I and Q transitions producing data-bit-specific RF phase transitions. With proper scaling of the I and Q input ratios, the vector sum of the I and Q channels forms a vector of any phase or amplitude.

In cellular radio systems, the channel carrier spacing is very narrow compared to the modulating frequency. Therefore, the third harmonic of the modulating frequency must be suppressed to reduce interchannel interference. Unlike the QPSK modulator (as described in section L), the SSB modulator must handle the data as a linear signal with the RF carrier acting as the higher power level (LO) signal. For example, the DCS1800 and GSM standards use Gaussian minimum-shift-keying (GMSK) modulation. GMSK makes use of band-limited modulating data to minimize the requirement for output filtering at RF. Due to this band-limiting, the RF output is also band-limited when a linear modulator is used. This enables the system to meet stringent channel bandwidth requirements, thereby compressing many RF channels together to maximize user capacity.

A SSB modulator can be built as shown in Figure 43. The input carrier signal is applied to an in-phase, 3 dB power divider selected for the frequency band of interest. The outputs of this power divider feed two double balanced mixers with DC-coupled IF ports for audio modulation. The modulating signal is applied to the IF ports of the mixers through an IF 90° 3 dB hybrid. The outputs of the two mixers are combined by a 90° 3 dB hybrid covering the carrier signal band.



**Figure 43. Block Diagram of the Single Sideband Modulator<sup>5</sup>**

The output of the upconverter is  $f_{\text{CARRIER}} \pm f_{\text{IF}}$  (the carrier is suppressed). When the signal goes through the  $90^\circ$  hybrid combiner, one of the two frequencies,  $f_{\text{CARRIER}} + f_{\text{IF}}$  or  $f_{\text{CARRIER}} - f_{\text{IF}}$  appears out of phase and is canceled.

If the modulating signal is applied to:

- IF1, the upper side band is produced at the RF output.
- IF2, the lower side band is produced at the RF output.

The sidebands can be easily switched by selecting the IF port to which the modulating signal is applied.

When using a single sideband modulator the balance of the phase and amplitude relationships between the mixers is critical to obtain high suppression of the unwanted sideband. It may be necessary to trim the signal path lengths and losses to get good suppression. A value of 20 to 25 dB is obtainable over narrow bands, i.e. 5 to 10% bandwidth. Normally the IF hybrid design is not critical and 1 to 1000 MHz hybrids work well.

## A. FREQUENCY TERMS

### 1) Signal Frequency (RF)

Signal Frequency is the frequency of the desired RF or microwave signal onto which information has been modulated. This is the frequency that is to be converted to a different (normally lower) frequency.

### 2) Local Oscillator Frequency (LO)

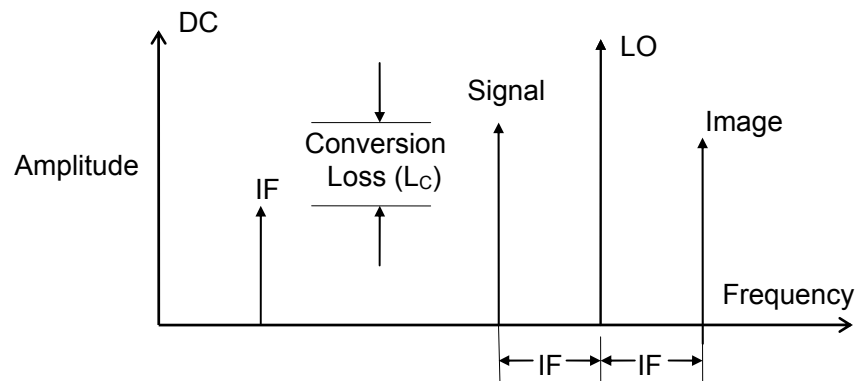
The local oscillator frequency is the frequency of the signal which provides the RF bias to the mixer diode(s). The LO signal is normally at a higher power level than the signal frequency power. An optimum local oscillator power is required to obtain low conversion loss and good RF match to the mixer diode(s). Optimum local oscillator powers are in the range of -3 to +10 dBm for most common diodes.

### 3) Intermediate Frequency (IF)

The IF frequency is the desired output frequency from the mixing process and is normally the difference between the LO and RF signal frequencies,  $LO - RF$  or  $RF - LO$ . When an upconverter is used the IF is the baseband modulating signal.

### 4) Image Frequency

The image frequency ( $f_{IM}$ ) is the frequency at which a (typically) unwanted signal is produced by the interaction of the RF signal and the second harmonic of the LO signal,  $2f_{LO} - f_{RF}$  or  $f_{LO} + f_{IF}$  (see Figure 44). In most mixers, the image frequency signal is terminated in a broadband termination. Since most mixers have reciprocity, termination of the image causes loss of half the signal power. This is the reason that the minimum diode conversion loss is  $\geq 3.0$  dB unless an image enhancement circuit is used. The four frequencies of a normal diode mixer are illustrated below.



**Figure 44. Mixer Frequency Relationships**

## B. TYPES OF MIXERS BY FREQUENCY OUTPUT

### 1) Down Converter (Mixer)

In a down converter (usually simply called a mixer) the desired IF output signal is obtained from the difference of the LO and RF signal frequency. Normally it is at only a small fraction of the signal frequency.

### 2) Up Converter (Modulator)

In an up converter the input signals will be the LO and the IF signals. The RF output signal is the LO + IF. Normally the IF signal is modulated prior to up conversion.

## C. MIXER DIODE TERMS (CHARACTERISTICS)

### 1) Conversion Loss

Conversion loss is the loss of signal power that results from the conversion from the RF signal frequency to the IF frequency in a down converter or conversion from the IF signal frequency to the RF frequency in an up converter. It is defined as a power ratio:

(52a)

$$L_C = \frac{\text{IF Output Power}}{\text{RF Signal Input Power}}$$

Conversion loss may also be expressed in dB:

(52b)

$$L_C = 10 \log \frac{\text{IF Output Power}}{\text{RF Signal Input Power}}$$

When referred to a mixer diode, it is the loss in an optimum single ended mixer carefully designed to minimize losses in the RF and LO coupling networks. Conversion loss normally includes power transferred to the image frequency which is resistively terminated.

### 2) Noise Figure

The noise factor and noise figure (NF) of a mixer diode are closely related to its conversion loss. Noise factor and noise figure are usually measured single sideband in a single ended mixer. Noise factor is the ratio of the signal to noise ratio at the mixer input to the signal to noise ratio at the mixer output.

(53)

$$\text{Noise\_factor} := \frac{\left( \frac{\text{Signal}_{\text{in}}}{\text{Noise}_{\text{in}}} \right)}{\left( \frac{\text{Signal}_{\text{out}}}{\text{Noise}_{\text{out}}} \right)}$$

(54)

$$\text{NF} := 10 \cdot \log(L_C (T_M + f_{\text{IF}} - 1))$$

where:

- $L_C$  = diode conversion loss (expressed as a power ratio)
- $T_M$  = diode effective temperature ratio (normally < 1.0)
- $f_{\text{IF}}$  = IF amplifier noise figure (expressed as a ratio)

Normally this simplifies to  $\text{NF} \sim 10 \log L_C (f_{\text{IF}} - 1)$

For a perfect mixer diode, NF (in dB) = 3.0 dB plus the IF amplifier noise figure and any circuit losses.

### 3) SWR or Match

SWR or match refer to the input standing wave ratio (SWR) of a single diode in a fixed tuned holder at the LO frequency. This is normally stated at a fixed LO drive, normally enough to produce ~1 mA of rectified current.

### 4) IF Impedance

IF impedance is the average of the time varying impedance of a mixer diode at a nominal IF frequency (usually 30 MHz). It is measured with an admittance bridge at a fixed rectified current (normally 1 mA) set up by the LO drive. Most Schottky diodes will have IF impedances in the range of 150-400 ohms at 1 mA.

### 5) Burn Out

Mixer diodes can be destroyed by static discharge or excessive incident RF power. Most Schottky diodes fail by becoming a short circuit. Burn out is defined as the maximum RF power which the diode can withstand without damage. It normally is in the range of 50 to 500 mW CW and up to 1 to 5 watts for pulses less than 2 to 5 nanoseconds long.

### 6) 1/f or Flicker Noise

The effective noise temperature ratio ( $T_M$ ) of most Schottky diodes is normally close to thermal, i.e. equal to that of a resistor of the same impedance at intermediate frequencies above 100 kHz to 1 MHz. However, as the IF frequency is reduced below 100 kHz to 1 MHz, the noise temperature ratio begins to rise rapidly. It normally increases as the inverse of the IF frequency.

This rise increases the noise figure of the diode at low frequency because  $NF = 10 \log (L_C T_m + L_C (f_{IF} - 1))$ . Diodes designed for low audio frequencies normally will have 1/f noise specified at a single frequency, i.e. 100 Hz or 1 kHz as required by the system.

### 7) Barrier Height

Barrier height is the difference between the work function of the metal and the electron affinity in the semiconductor<sup>9</sup>. Barrier height is expressed in term of volts or electron volts.

The barrier height of a Schottky junction determines the I/V characteristics of that diode. This can be important because it determines the local oscillator power necessary to bias the junction to its optimum nonlinear operating point. As an approximation, the optimum local oscillator power will increase as the square of the barrier height, provided the same mixer circuit characteristics and junction capacitance values are used. Table I shows a list of the barrier heights of many metals and metal-silicide systems used for microwave mixer diodes. Figures 48 & 49 show the approximate local oscillator power requirement vs. barrier height.

## D. DETECTOR DIODE CHARACTERISTICS

### 1) Tangential Signal Sensitivity (TSS)

The tangential signal sensitivity is a direct measure of the signal to noise ratio of a detector and it defines the maximum sensitivity of a detector. It is defined as the input power at which a signal to noise ratio of 2.5:1 is produced. TSS usually is measured on an oscilloscope (see below). Because it is a noise measurement the amplifier bandwidth must be defined. Usually this is 1 MHz. It is expressed in dBm (i. e., -55 dBm).

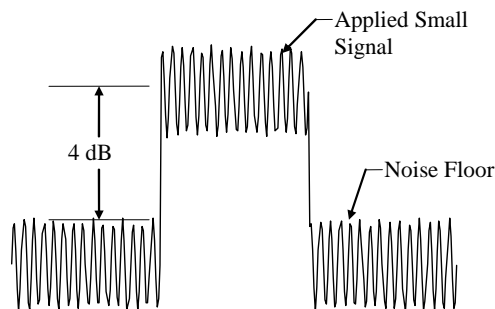


Figure 45.

TSS



## 2) Voltage Sensitivity ( $\gamma$ )

Voltage sensitivity ( $\gamma$ ) is a measure of the output voltage available at a standard input power using a defined load resistance (usually 1 megohm). It is normally measured in the square law range of the diode, typically with  $P_{IN} = -30$  dBm. Voltage sensitivity is given in mV/mW.

## 3) Video Impedance ( $R_V$ )

Video impedance of a detector diode is the Norton equivalent impedance of the diode when it is modeled as an RF-input-voltage-controlled current source. The video impedance of a diode is affected by the DC current flowing in the diode. It is normally specified with a given small DC current, i.e., from 1 to 100 microamperes.

Some ZBD (Zero Bias Detector) diodes are used without an externally applied bias current. Normal video impedances for these diodes can range from 1-2 kilohms to megohms. The video impedance can affect the pulse fidelity of a video detector as the RC time constant of the amplifier/detector depends on  $R_V$  and the bypass capacitor.

The video resistance of a diode is the slope (AC) resistance of a detector diode. It will determine the voltage sensitivity of a detector diode.

The video impedance of a diode is the RF impedance looking into the diode from the video amplifier. It is used to match the detector to the video amplifier.

## E. RECEIVER SYSTEM CHARACTERISTICS

### 1) Receiver Sensitivity

The following equation for the sensitivity of a receiver shows the parameters which affect a receiving system's sensitivity:

$$(55) \quad S = -114 + NF_o + 10 \log_{10} B + 10 \log_{10} (S/N)$$

where:

$S$  = receiver sensitivity in dBm

$B$  = receiver bandwidth in MHz

$NF_o$  = receiver overall noise figure in dB

$S/N$  = minimum acceptable receiver output signal-to-noise ratio in dB

## 2) Doppler Shift

Doppler radars utilize the fact that RF energy reflected by a moving target is apparently shifted in frequency as a result of the relative motion between the source and reflecting target. The amount of this frequency shift is directly proportional to the target's velocity relative to the radar's transceiver. The same effect occurs with sound waves when an automobile sounding its horn is moving with respect to an observer. The sound pitch is higher when the horn is moving toward the observer and decreases as it moves away. The Doppler shift frequency  $f_d$  is given by:

(56)

$$f_d := 2 \cdot v \cdot \frac{f_0}{c} \cdot \cos\phi$$

where:

- $f_0$  = transmitter frequency in Hz
- $c$  = velocity of light ( $3 \times 10^8$  meters per second)
- $v$  = velocity of the target (meters per second)
- $\phi$  = angle between microwave beam and target's path

**Note:**  $\cos\phi = 1$  when the target moves directly toward or away from the signal transceiver. Velocity  $v$  is expressed as a vector so it determines the sign of the Doppler shift frequency.

## 3) Typical Doppler Radar System

A typical Doppler radar system consists of an RF (i.e., microwave) section, a signal processing section and a bias supply.

In order to design a Doppler radar system, one must first know:

1. The maximum range at which the target is to be detected. This determines the overall sensitivity required of the transceiver.
2. The maximum and minimum target speeds that the system is to measure. This determines the required frequency characteristics of the IF amplifier.

The commercial Doppler systems such as police radars and intrusion alarms usually operate with a "zero IF" because the transmitter source (often a Gunn oscillator) is also used as the local oscillator for the mixer. Using this technique, the signal amplification is most easily applied to the IF signal at the Doppler shift frequency. For example, if the transmitter frequency is 10.525 GHz, a vehicle traveling 50 mph (80.5 km/hour) will cause a Doppler shift of 1568 Hz. A police radar's IF amplifier bandpass frequency should be approximately 50 Hz to 5000 Hz.

The maximum range of a radar system can be determined by the following equation:

(57)

$$R_{\max} := \left( \frac{P_t \cdot G_a \cdot K}{F} \right)^{\frac{1}{4}}$$

where:

$P_t$  = transmitted power

$G$  = antenna gain expressed as a ratio

$F$  = receiver noise figure expressed as a ratio

$K$  = a constant

This expression shows that the effective range of a radar system is inversely proportional to the fourth root of the overall receiver noise figure.

Mixer Type	VSWR			Port-to-Port Isolation			LO Power Requirement	P <sub>1dB</sub>
	RF	LO	IF	RF/IF	LO/RF	LO/IF		
Single Ended	Depends on Matching Circuits			Depends on Filters			Low	Low
Single Balanced (180°)	Depends on Matching Circuits			Depends on Filters	Poor	Depends on Filters	Moderate	Moderate
Single Balanced (90°)	Good	Good	Depends on Matching Circuits	Depends on Filters	Poor	Depends on Filters	Moderate	Moderate
Double Balanced	Good			Good			High	High
Double-Double Balanced	Good			Good			Highest	Highest
Sub-Harmonic	Depends on Matching Circuits			Good	Depends on Filters	Depends on Filters	Moderate	Low
Image Reject	Good			Good			Very High	High

Mixer Type	LO AM Noise Rejection	LO Spurious Signal Rejection	3rd Order IM Intercept	Low order Spurious Response Rejection
Single Ended	None	Depends on Matching Circuits	Low	None
Single Balanced (180°)	Good	Depends on Matching Circuits	Moderate	2f <sub>LO</sub> ± 2f <sub>RF</sub> : Good 2f <sub>LO</sub> ± f <sub>RF</sub> : Good if Designed to Suppress LO Harmonics f <sub>LO</sub> ± 2f <sub>RF</sub> : Good if Designed to Suppress LO Harmonics
Single Balanced (90°)	Good	Good	High	2f <sub>LO</sub> ± 2f <sub>RF</sub> : Good 2f <sub>LO</sub> ± f <sub>RF</sub> : None f <sub>LO</sub> ± 2f <sub>RF</sub> : None
Double Balanced	Good	Good	High	2f <sub>LO</sub> ± 2f <sub>RF</sub> : Good 2f <sub>LO</sub> ± f <sub>RF</sub> : Good f <sub>LO</sub> ± 2f <sub>RF</sub> : Good
Double-Double Balanced	Good	Good	Highest	2f <sub>LO</sub> ± 2f <sub>RF</sub> : Good 2f <sub>LO</sub> ± f <sub>RF</sub> : Good f <sub>LO</sub> ± 2f <sub>RF</sub> : Good
Sub-Harmonic	Good	Depends on Matching Circuits	Low	Rejects all mixing with odd harmonics
Image Reject	Good	Good	High	2f <sub>LO</sub> ± 2f <sub>RF</sub> : Good 2f <sub>LO</sub> ± f <sub>RF</sub> : Good f <sub>LO</sub> ± 2f <sub>RF</sub> : Good

**Table 1<sup>4</sup> Comparison of the Characteristics of Common Mixer Circuits**

Type of Mixer or System	Frequency Range	Suggested Diode Types	Comments
Millimeter Receiver Mixers	20 - 100 GHz	GaAs Beam Lead GaAs Flip Chip	Best noise figure
Millimeter Up Converter	20 - 100 GHz	GaAs Beam Lead GaAs Flip Chip	Lower conversion loss allows higher output power
Subharmonic Mixer	20 - 100 GHz	GaAs Beam Lead GaAs Flip Chip	Better conversion loss, noise figure and dynamic range
High Dynamic Range Mixers	100 MHz - 12 GHz	Si Dual Barrier Packaged or Beam Lead	Highest dynamic range. However, noise figure is a little worse above 10 GHz due to extra Rs of 2nd junction
	10 - 20 GHz	GaAs Ring Quad Beam Lead or Flip Chip	Good IM3, better conversion loss than Si. Larger LO drive than required by Si.
Modulator or Sampling Switch	10 MHz - 5 GHz	Si or GaAs Bridge Quad Packaged or Beam Lead	GaAs can have higher $V_B$ than Si but also has higher $V_F$ than Si
Balanced Mixer	1 - 18 GHz	Si - Packaged, Beam Lead or SurMount	Si less expensive than GaAs, operates with lower LO power
Double Balanced Mixer	1 - 18 GHz	Si - Packaged, Beam Lead or SurMount	Si less expensive than GaAs, operates with lower LO power
Starved LO Mixer	1 - 18 GHz	Si - ZBD or Low Barrier Packaged, Beam Lead or SurMount	Lowest LO drive requirement
Phase Detector or Bi Phase Modulator	100 MHz - 18 GHz	Si - Packaged, Beam Lead or SurMount	Match and symmetry of a monolithic quad are better than that of discrete quad
Image Reject Mixer	1 - 18 GHz	Si Ring Quad - Packaged, Beam Lead or SurMount	Match and symmetry of a monolithic quad are better than that of discrete quad

**Table II Selection Guide for Diode Configuration and Material by Type of System**

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