APPLICATION NOTE

Mixer and Detector Diodes

Surface Barrier Diodes

Most people who use diodes are more familiar with junction devices than with the surface barrier diodes commonly used in mixer and detector circuits. In a junction diode the rectifying junction is formed between a p-type region and an n-type region of a semiconductor. In a surface barrier diode the rectifying junction is formed between a metal and a semiconductor, which may be either n-type or p-type.

Both devices operate on the same physical principals, the difference being in the construction.

The Schottky barrier diode is made by sputtering or evaporating the barrier metal onto the surface of the semiconductor (silicon or gallium arsenide).

The Schottky barrier type is available with a wider range of electrical properties and package types for more advanced circuits. In this application note we go into the details of the physics, construction, and applications of Schottky diodes.

Types of Construction

Schottky diodes are available from Skyworks on two semiconductor materials—silicon and gallium arsenide. Silicon diodes are available in either n-type or p-type polarity while GaAs diodes are available in n-type only. Skyworks Schottky diodes can be divided into classifications based on packaging and chip construction.

Mounted Beam-Lead Package

In this type, one or more beam-lead Schottky diodes with coplanar leads are bonded onto a ceramic, fiberglass, or plastic substrate. This construction is mechanically rugged, has very low inductance, and is particularly convenient for double-balanced mixers.

Unmounted Chip

These are for those who prefer to use chips; they are available in several different sizes and bonding pad arrays.

Unmounted Beam-Lead Diodes

These are for use in MIC circuits or other special constructions, where minimum inductance or minimum size are important. They are available as single diodes, pairs, quads, and other monolithic arrangements.

Electrical Characteristics and Physics of Schottky Barriers

Schottky barrier diodes differ from junction diodes in that current flow involves only one type of carrier instead of both types. That is, in n-type Schottkys, forward current results from electrons flowing from the n-type semiconductor into the metal; whereas in p-type Schottkys, the forward current consists of holes flowing from the p-type semiconductor into the metal.

Diode action results from a contact potential set up between the metal and the semiconductor, similar to the voltage between the two metals in a thermocouple. When metal is brought into contact with an n-type semiconductor (during fabrication of the chip), electrons diffuse out of the semiconductor into the metal, leaving a region under the contact that has no free electrons ("depletion layer"). This region contains donor atoms that are positively charged (because each lost its excess electron), and this charge makes the semiconductor positive with respect to the metal. Diffusion continues until the semiconductor is so positive with respect to the metal that no more electrons can go into the metal. The internal voltage difference between the metal and the semiconductor is called the contact potential and is usually in the range 0.3–0.8 V for typical Schottky diodes.

When a positive voltage is applied to the metal, the internal voltage is reduced, and electrons can flow into the metal. Only those electrons whose thermal energy happens to be many times the average can escape, and these "hot electrons" account for all the forward current from the semiconductor into the metal.

One important thing to note is that there is no flow of minority carriers from the metal into the semiconductor and thus no neutral plasma of holes and electrons is formed. Therefore, if the forward voltage is removed, current stops "instantly," and reverse voltage can be established in a few picoseconds. There is no delay effect due to charge storage as in junction diodes. This accounts for the exclusive use of surface barrier diodes in microwave mixers, where the diode must switch conductance states at microwave local oscillator rates.

The voltage-current relationship for a barrier diode is described by the Richardson equation (which also applies to thermionic emission from a cathode). The derivation is given in many textbooks (for example, Sze).*

$$I = AA^{**}{}_2 \, exp \left(- \frac{q\phi B}{kT} \right) \left[exp \left(\frac{qv}{NkT} \right) - 1 \right]$$

where

A = area (cm^2)

A** = modified Richardson constant (amp/oK)^2/cm^2)

I = Boltzman's Constant

T = absolute temperature (°K)

ϕB = barrier heights in volts

V = external voltage across the depletion layer (positive for forward voltage) - V - IR_S

R_S = series resistance

I = diode current in amps (positive forward current)

n = ideality factor

The barrier height ϕB is primarily determined by choice of barrier metal and the type (n or p) of semiconductor used. A secondary consideration is the crystal orientation of the substrate. The barrier height is important as it determines the amount of local oscillator power required to drive the diode into its nonlinear region. If there is limited local oscillator available a low barrier diode would be used. If more local oscillator power is available a higher barrier diode could be used to improve intermodulation distortion.

Richardson's equation describes the behavior of the diode but it is hard to use for circuit design. A better equation for circuit designers to use is one in which all parameters are independent of voltage and current. The simplest one that agrees fairly well with Richardson's equation is

$$I = A^*J_0 {}^*(exp(q^*V/nkT) - 1) = I_S \, exp \left(\frac{qv}{nkT} - 1 \right)$$

where

I = current in amps

A = area of Schottky barrier in cm^2*

J_0 = saturation current density in amps/cm^2 *

V = applied voltage in volts

n = ideality factor

T = temperature in Kelvin

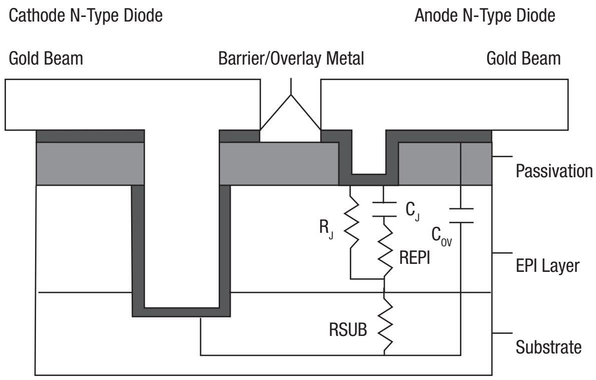
* typical values for these variables and others necessary for computer modelling are included in the following table.

Matrix for mW Spice Diode Model

Frequency	Drive	V_B (V)	C_{J0} Min. (pF)	C_{J0} Max. (pF)	R_S (Ω)	V_F (V)	I_S A
KU	DMF	2	0.05	0.15	12	0.5	3.17E–08
	DME	3	0.05	0.15	12	0.6	6.33E–10
	DMJ	4	0.05	0.15	12	0.8	6.33E–13
X	DMF	2	0.15	0.30	7	0.5	1.27E–07
	DME	3	0.15	0.30	7	0.6	2.53E–09
	DMJ	4	0.15	0.30	7	0.8	2.53E–12
S	DMF	2	0.30	0.50	4	0.5	2.48E–07
	DME	3	0.30	0.50	4	0.6	4.97E–09
	DMJ	4	0.30	0.50	4	0.8	4.97E–12

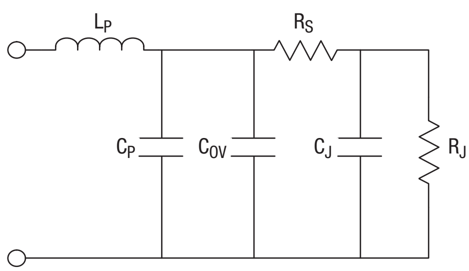
Diode Cross-Section

The following picture shows a cross-section of a typical beam-lead Schottky diode.



where
R_J = junction resistance
C_J = junction capacitance
Repi = resistance of epi layer
Rsub = resistance of substrate (spreading resistance)
Cov = overlay capacitance

The equivalent circuit of these structures is shown below.



where
R_S = Repi + Rsub
C_P = package capacitance (where applicable)
L_P = package inductance (where applicable)

The following table lists C_P and L_P for some standard single diode packages.

Package	CP pF	LP (nH)
130–011	0.10	0.6
207–011	0.13	0.6
247–001	0.15	0.3
325–011	0.14	0.6
404–011	0.09	0.5
464–011	0.03	0.5

Schottky Barrier Diode Capacitance

The total capacitance of a Schottky diode is:

$$C_T = C_J + Cov + C_P$$

where
C_J = junction capacitance
Cov = overlay capacitance
C_P = package capacitance

The junction capacitance is generally measured without bias and is governed by the following equation:

$$C_J(0) = \frac{A*q*E_S*Nd^{1/2}}{(2*(Vi - kT/q))}$$

At an applied voltage C_J(v) can be computed by the following equation:

$$C_J(V) = \frac{Cj(0)}{(1 - (V/Vi - kT/q))^{1/2}}$$

where
A = area of Schottky barrier in cm^2
Nd = doping density of epi layer in cm^3
E_S = dielectric constant of material $*E_0$
V = applied voltage in volts
Vi = built in voltage = ϕ_B -0.15 for n-type silicon with Nd $=10^{17}$

Series Resistance

The series resistance of a Schottky diode is the sum of the resistance due to the epi layer and the resistance due to the substrate. The resistance of the epi is given by the following equation:

$$Repi = \frac{t}{q*Ue*Nd*A}$$

$$= \frac{L}{q\mu_N N_d A}$$

where
L = thickness of epi in cm
μ_n = mobility of electrons for n-type Si (for p-type silicon the mobility of holes would be used)
Nd = doping density of the epi layer in cm^3
A = area of Schottky contact in cm^2

The resistance of the substrate is given by the following equation:

$$Rsub = 2*\rho_S*(A/\pi)^{1/2}$$

Where
A = area of Schottky contact in cm^2
ρ_S = substrate resistivity in Ω-cm

Mixer Diodes Compared To Detector Diodes

Mixer diodes are designed to convert radio frequency (RF) energy to an intermediate frequency (IF) as efficiently as possible. (In practice, the conversion efficiency should be at least 20%.) The reason for doing this is that selective amplifiers at the RF frequency are expensive, so the signal is converted to a lower frequency where high gain and good selectivity can be more easily achieved.

The frequency conversion is obtained by operating a diode with fast response and high cutoff frequency as a switch, turning it on and off at a rate determined by a local oscillator (LO). The output frequency (I_F) is then the difference between the LO frequency and the RF frequency.

A good mixer diode with a high cutoff frequency will be capable of low conversion loss (L_C). This, combined with a low noise figure in the IF amplifier, will result in a low overall noise figure, unless the diode itself generates noise (other than normal thermal noise). Ideally, the mixer diode should accomplish this with a minimum of LO power and no DC bias.

Detector diodes are designed to rectify very low levels of RF power to produce a DC output voltage proportional to the RF power. The diode may be operated at a small DC bias (typically 50 µA) which results in a relatively high RF impedance (typically 600 Ω). As a result, very low capacitance is required to achieve high sensitivity. Since the output is at a very low level, the low frequency, audio frequency excess noise (“1/f noise”) is an important consideration.

Mixer Parameters

The quality of a mixer diode is generally controlled by either low frequency parameters or RF operating parameters.

Low frequency parameters customarily specified are (in order of importance):

Junction Capacitance (C_{J0}) at zero bias

Series Resistance (R_S) or cutoff frequency (f_{CO})

Reverse Voltage (V_B) at 10 mA or 100 mA

Forward Voltage (V_F) at 1 mA

Excess Noise Voltage (1/f noise)

Leakage Current (I_R) at IV

Series resistance is sometimes controlled by specifying dynamic resistance, RT, at some particular forward current. Series resistance can then be calculated by subtracting R_B (R_B =28/I(ma)) from RT. The excessive noise voltage need not be specified unless the IF frequency is less than 1.0 MHz (such as for Doppler radars or autodyne mixers).

Some people prefer to specify RF parameters instead of the above low frequency parameters. In order of importance, the customary parameters are:

Noise Figure (NF in dB)

would be specified in a particular mixer circuit at a particular RF frequency and LO power level.

Conversion Loss (L_C in dB)

would be specified in a particular mixer circuit at a particular RF frequency and LO power level.

RF Impedance (VSWR)

expresses how well the diode and circuit are matched to the LO source at a particular LO power.

IF Impedance (Z_{if})

expresses the low frequency impedance of the driven diode, considered as a source of IF voltage. The IF amplifier should be designed to have its optimum noise figure for this source impedance. This parameter is dependent on LO power, as well as RF and harmonic impedance presented to the diode.

Detector Parameters

As with mixers, a detector diode can be specified by its low frequency parameters, the same ones that apply to the mixer diodes, with the exception that 1/f noise is now second in importance instead of fifth.

Alternatively, a detector diode can be specified by RF parameters, the customary ones being:

Voltage Sensitivity (V/mW)

is the ratio of DC voltage output to RF power input at a particular frequency and power level. Voltage sensitivity depends on bias current and C_{J0}.

Tangential Signal Sensitivity (TSS, in dBm)

is the minimum RF signal level, in dB below 1 mW, that produces a tangential indication on a low frequency oscilloscope. See Figure 1:

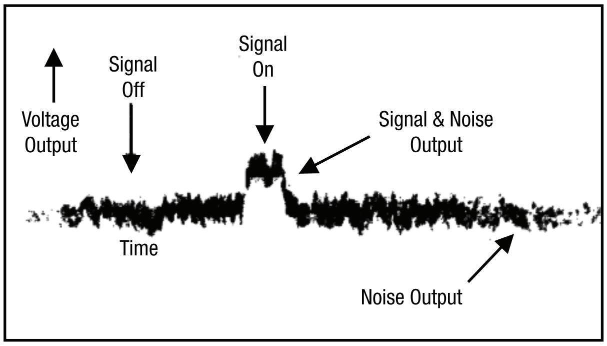


Figure 1. Measurement of Tangential Signal Sensitivity

(Tangential sensitivity depends on voltage sensitivity, diode excess noise voltage, and both RF and video bandwidth).

Video impedance (Z_V, in Ω)

is the low frequency impedance of the diode, considered as a source of video voltage. It is the same as R_T at the bias current used (about 600 W for any diode with 50 µA bias).

Figure of Merit (FM)

This parameter combines voltage output and Z_V to give a convenient bandwidth-independent measure of TSS.

Mixer Diodes

Theory of Mixers

The simplest way to think about the action of a mixer diode is to consider a single-ended mixer consisting of a single diode at the end of a transmission line. The RF signal and the local oscillator drive power are coupled into the same line by filters or hybrids. The local oscillator drives the diode into heavy forward conduction for nearly half a cycle and into reverse bias for the other half cycle. The reflection coefficient of the diode, Γ, then varies periodically as a function of time.

In this model the only effect of the junction capacitance and package parasitics is to transform the source impedance from its actual value to some other number, Z_0, at the semiconductor junction. If the instantaneous junction conductance is G(t), then you have the situation indicated in Figure 2:

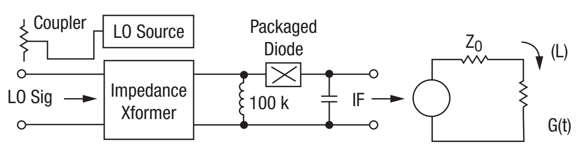


Figure 2. Mixer and Equivalent Circuit

For available LO power, P_L, the generator voltage is:

$$2V_L\,(t) = 2V_L \cos \omega_L t$$

where

$$V_L = (2Z^1{}_0 P_L)^{-0.5}$$

Diode I–V Approximation

The forward diode characteristic is given by the equation

$$L(t) = I_S \exp[(Vt) - IR_S /0.028]$$

This equation can be approximated by a two-piece linear approximation, which has the diode conducting only if the voltage exceeds a forward voltage, V_F:

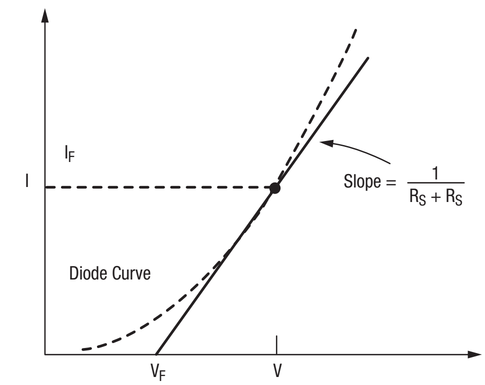


Figure 3. Diode Forward Characteristics

The barrier resistance, R_B, should be evaluated at the peak current using $R_B = 0.028/I_P$. The equation for I_P is

$$I_P = \frac{2V_L - V_F}{Z_Z{}^1{}_0 + R_S = R_B}$$

The approximation can be justified by graphing the equation or by looking at an actual diode on a curve tracer (1 mA/cm). In practice,V_{F1}, the forward voltage at 1 mA, can be used for V_F.

Therefore, the low frequency diode conductance, G is

$$G(t) = \begin{cases} \dfrac{1}{R_S + R_B}, \text{ if } 2VL(t) > V_F \\ \omega^2 C_J{}^2 R_S, \text{ otherwise} \end{cases}$$

If you use this reasoning to compute the time-dependent reflection coefficient, the result is a rectangular waveform (Figure 4).

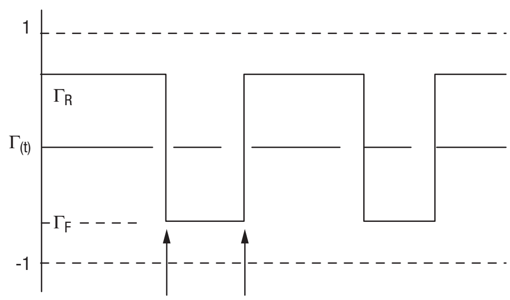


Figure 4. Time Dependent Reflection Coefficient

$$\Gamma_F = \frac{R_S + R_B - Z^1_0}{R_S + R_B + Z^1_0} \approx 1 + \frac{2(R_B + R_S)}{Z^1_0}$$

$$\Gamma_R = \frac{1 - Z^1_0 \omega^2 C_J^2 R_S}{1 + Z^1_0 \omega^2 C_J^2 R_S} \approx 1 - 2Z^1_0 \omega^2 C_J 2R_S$$

The angle, T, is the conduction angle, i.e. the number of electrical degrees of the LO waveform during which the diode is conducting.

$$\Theta = 2 \text{ arc cos } \left(\frac{V_F}{V_T} \right)$$

$$= 2 \text{ arc cos } \left(\frac{V_F}{(8Z^1_0 P_L)^{-0.5}} \right)$$

Typically the conduction angle is between 120° and 170°.

Conversion Loss

In order to handle the mathematics of the mixer, the G waveform must be expressed as a Fourier series

$$\Gamma(t) = \Gamma_0 = \Gamma_1 \cos \omega_L t + \Gamma_2 \cos \omega_L + ...$$

where

$$\Gamma_1 = 2/\pi(\Gamma_F - \Gamma_R) \sin \Theta/2$$
$$- 2/\pi(2 - 2Z^1_0 \omega L^1 C_J{}^2 R_S - 2R_S + R_B/Z^1_0) \sin \Theta/2$$

When there is an incident of RF signal voltage $V_S \cos w_S t$, in addition to the LO voltage, the voltage of the reflected wave is

$$V_R(t) = \Gamma(t) V \cos \omega t$$
$$= \Gamma_0 V_S \cos \omega t + \Gamma_1 \cos \omega_L t \cos \omega_S t + ...$$
$$= \Gamma_0 V_S \cos \omega t + 1/2\, \Gamma_1 V_S [\cos(\omega_L - = \omega_S) t$$
$$\cos(\omega_L - = \omega_S) t]...$$

The important term is the one involving $\omega_L - \omega_S$, because this is the difference frequency (IF). The ratio of reflected power at this frequency to the incident power at ω_S is the conversion efficiency, n.

$$\eta = \frac{P_{IF}}{P_S} = \frac{(0.5\Gamma_1 V_S)^2}{V_S{}^2} = \frac{\Gamma_1{}^2}{4}$$
$$= \frac{4}{\pi^2} \left[1 - Z^1_0 \omega_L{}^2 C_J{}^2 C_J{}^2 R_S \frac{(R_S + R_B)}{Z^1_0} \right]^2 \sin^2 \frac{\Theta}{2}$$

To optimize the conversion efficiency, you clearly want R_S to be zero; however, nature won't allow you to do this. In practice low R_S means large junction diameter and thus high C_J (and vice versa), so diode manufacturers introduce a parameter, the "cutoff frequency," which is essentially independent of junction diameter:

$$f_c = \frac{1}{2\pi R_S C_J}$$

where f_c = cutoff frequency

It is useful to express conversion loss in terms of fc instead of R_S, leaving C_J as the free parameter, since the range of variation of f_c in actual products is limited by material properties, whereas C_J can be designed for almost any value.

$$R_S = \frac{1}{\omega_L{}^2 C_J}$$

$$\eta = \frac{4}{\pi^2} \sin^2 \frac{\Theta}{2} \left[1 - \left(\frac{Z^1_0}{X_C} + \frac{X_C}{Z^1_0} \right) \frac{f}{f_c} - \frac{R_B}{Z^1_0} \right]^2$$

The quantity in parenthesis is close to 2, if the reactance of CJ is between $Z'_0/2$ and $2Z'_0$. So, for a large range of C_J, the conversion efficiency is determined almost entirely by the ratio of LO frequency to the cutoff frequency of the junction, by the peak current which determines R_B, and by the conduction angle.

For this reason, the capacitive reactance should be chosen to be Z'_0 or typically 100 Ω. The exact value is not critical for conversion loss unless very wide bandwidth is desired. Cutoff frequency should clearly be as high as possible. Conduction angle and R_B are determined by LO power and forward voltage. Therefore, LO power should be high and forward voltage should be low.

For high drive levels, Θ is close to 180°, sin t/2 is nearly one and R_S – 0 so the best conversion efficiency is

$$\eta = \frac{4}{\pi^2}\left(1 - 2\frac{f}{f_c} - \frac{R_B}{Z^1_0}\right)^2$$

and the conversion loss, in dB, is

$$L_C \approx 3.9\ dB + 17\frac{f}{f_c} + 9\frac{R_B}{Z^1_0}$$

Actual single-ended mixers, such as the ones used at Skyworks to test Schottky diodes, give results similar to this equation, or slightly better. Theoretically, an actual mixer can be 0.9 dB better than this because of harmonic suppression. That is, instead of the sum frequency and other harmonics being absorbed in the source resistance, they are reflected back into the diode to be remixed with harmonics of the $-\Gamma$ waveform to produce more IF output. In actual diodes this happens automatically if the package is designed to have a low pass characteristic that cuts off frequencies between the operating frequency and the harmonics. In any case, the circuit can be designed to reflect all harmonics back into the diode, and if these reflections are phased properly, you get the full 0.9 dB improvement.

The conversion loss actually measured on production diodes is in general agreement with the previous equations, as indicated in the following figure. The conversion loss points are from a large number of production lots measured at Skyworks over the last several years. As can be seen in Figure 5, the results follow equation (4–15) if 0.9 dB is subtracted for harmonic suppression, and the last term contributes about 0.5 dB.

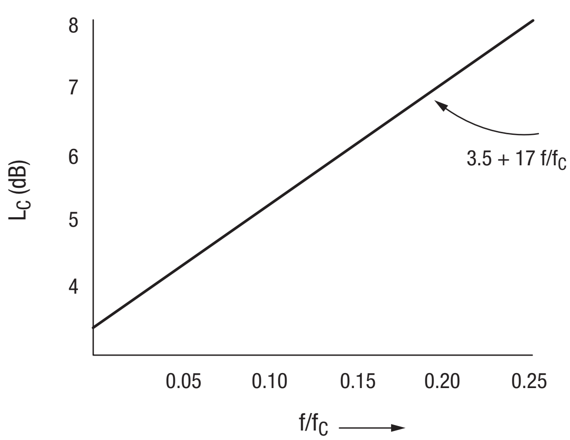


Figure 5. Conversion Loss as a Function of Normalized Frequency

Noise Figure

Definitions and Formulas

In practice, not only the wanted signal comes into the diode to be converted to the IF frequency, but also random signals of various sorts. This noise is also converted to the IF frequency with the same conversion efficiency as the signal. In addition to this, the mixer adds other sources of noise:

1. Image noise—If the signal frequency is $f_L + f_{iF}$, then noise at the frequency $f_L - f_{IF}$ is also converted to the IF frequency with the same efficiency. This doubles the noise at the IF port.
2. Diode thermal noise—The parasitic resistance R_S generates thermal noise. The higher the R_S the more the conversion loss and the higher this contribution is, in direct proportion. This noise source will increase if the diode is run at elevated temperatures.
3. Shot noise—Electron flow across the diode depletion layer generates shot noise. This noise turns out to be half what the thermal noise would be in an ordinary resistor equal to R_B, and will be directly proportional to the absolute temperature of the diode.
4. Excess noise—At low frequencies, the junction noise increases due to trapping of electrons. This noise often has 1/f spectrum and is therefore called 1/f noise. At high current levels there is additional noise due to velocity saturation of the carriers and carrier trapping. This noise has a minor effect on mixers and is discussed in a later section.
5. IF noise—The input stage of the IF amplifier adds some noise of its own. Most mixer specifications assume that the IF amplifier has a noise figure of 1.5 dB.
6. LO noise—The sidebands of the noise from the local oscillator may overlap the signal and image frequencies, thus acting like an excess noise source. (This effect can be eliminated by filtering the LO or by using a balanced mixer.)
7. Harmonic noise—In the wide-open, single-ended mixer design we are talking about, noise at frequencies near harmonics of the LO frequency can also be converted to the IF frequency. This can be eliminated by using a harmonic enhanced design, or by making sure that the package parasitics isolate the junction from the circuit at the harmonic frequencies.

Noise factor is defined as the ratio of the signal-to-noise (S/N) ration at room temperature at the signal input to the mixer to the S/N ration at the output of the IF amplifier. Noise figure is the noise factor expressed in dB. For a moderately heavily driven mixer ($R_B \approx 0$), the noise added from the image and the diode thermal noise (from R_S) exactly makes up for the noise lost in the conversion process, if the diode is at room temperature. Therefore, the noise power going into the IF amplifier is exactly equal to the noise coming in with the signal; but the signal is reduced, so the signal-to-noise ratio is reduced by exactly the amount of the conversion loss.

After adding in the IF noise figure, the result is

NF = *noise figure (dB)*

$= L_C \text{ (dB)} + N_{IF} \text{ (dB)}$

However, the shot noise and the excess junction noise should be considered. The shot noise added by the junction is only half what would be expected from a resistor equal to R_B. For low drive the increase in noise figure is not as great as the increase in conversion loss. If enough LO power is absorbed to heat the diode significantly, one should take into account the temperature of the diode. Also, excess noise (1/f noise) should be taken into account if the IF frequency is low. This is usually accounted for by assigning an effective temperature to the diode, which may be either less or more than room temperature, T_0.

$$NF = L_C \text{ (dB)} + NTR\text{(dB)} + N_{IF} \text{ (dB)}$$

where the NTR, in this model, is

$$NTR = \frac{T_{eff}}{T_0} = 1 - 4 \frac{f}{f_c} \left(\frac{T}{T_0} - 1\right) + \frac{R_B}{Z^1_0} \left(2 - \frac{T}{T_0}\right)$$

NTR = Noise Temperature Ratio

In most specifications, the IF amplifier noise figure is assumed to be 1.5 dB (if the actual amplifier has a different noise figure, the data are corrected to the nominal 1.5 dB). In addition, the diode is assumed to be operated at a junction temperature equal to room temperature.

Therefore, if the IF frequency is not too low the expected noise figure for the single-ended mixer, driven with a quiet local oscillator, is

$$NF \approx 5.4 \text{ dB} + 17 f/f_c + 10 \log_{10} (NTR) + 9 R_B / Z^1_0$$

For IF frequencies below 1.0 MHz the 1/f noise becomes important and the noise figure could be higher than this unless the diodes are selected for low1/f noise. At high local oscillator drive levels, RB decreases, but the high forward current activates additional noise due to traps and velocity saturation, as well as higher temperature. Thus the noise figure increases instead of approaching a constant. In addition, as the reverse swing from the LO approaches diode breakdown, the back resistance, RR, decreases, and conversion loss will be degraded further.

Double Sideband (DSB) Noise Figure

When noise figure is actually measured, a hot source or broadband noise tube (or noise diode) is used as a “signal” source. Unless filtering is used, this kind of source provides “signal” both at the signal frequency and image frequency. Therefore, when the noise source is switched on and off to determine the signal-to-noise ratio at the output of the IF amplifier, twice as much output is obtained with the noise source on than if a single frequency signal were used. The measured noise figure (the so-called “double sideband” noise figure) will be 3 dB lower than the specified (“single sideband”) noise figure. Nevertheless, this kind of measurement is more convenient to do, and usually the measurement consists of measuring the DSB noise figure and adding 3 dB to obtain the SSB noise figure.

There are many other factors, such as line losses, coupler losses, the loss in signal -LO combiner or filter, and the deviation of the IF noise figure from 1.5 dB which must be taken into account as part of the calibration in order to get the correct noise figure for the single diode mixer alone.

Crystal Current

The diode produces DC current as a result of rectifying the local oscillator current. The total current is

$$I(t) = \begin{cases} \dfrac{2V_L \cos \omega_L t - V_T}{Z^1_0} & \text{if } 2V_L(t) > V_T \\ \omega^2 C_J{}^2 R_S V_L \cos \omega_L t, & \text{(otherwise)} \end{cases}$$

The average DC current, or crystal current: (w = t) is

$$\text{crystal current} \quad I_{DC} = \frac{2V_L \left[\sin \frac{\theta}{2} - \frac{\theta}{2} \cos \frac{\theta}{2}\right]}{kT\pi \left(Z^1_0 + R_S + R_B\right)}$$

If you compute the DC voltage by similar reasoning, you find that there is an apparent reverse DC voltage equal to

$$V_{DC} = -Z_0 I_{DC}$$

This is caused by the DC current through the DC circuit assumed to be equal to Z_0. (Actual single ended mixers typically use a 100 Ω resistor.)

VSWR

The VSWR expresses how well the RF diode impedance is matched to the LO source impedance. In terms of the LO current and voltage it is defined as:

$$VSWR = \frac{Z_{LO}}{Z_0} \text{ or } \frac{Z_0}{Z_{LO}}\text{, whichever is larger}$$

The large signal impedance, Z_{LO}, is the ratio of V_{LO} and I_{LO} which are the first order Fourier coefficients of the voltage and current waveforms:

$$V(t) = V_{DC} + V_{LO}\cos\omega_L t + V_2\cos 2\omega_L t + \ldots$$
$$I(t) = I_{DC} + I_{LO}\cos\omega_L t + I_2\cos 2\omega_L t + \ldots$$

$$I_{LO} = \frac{2V_L(\theta - \sin\theta)}{2\pi\,(Z^1{}_0 + R_S + R_B)} + 2\omega^2 C_J{}^2 R_S V_L$$

$$V_{LO} = 2V_L - Z^1{}_0 I_{LO}$$

$$\frac{Z_{LO}}{Z^1{}_0} = \frac{V_L}{Z^1{}_0 I_{LO}} = \frac{2\pi\left(\frac{R_S + R_B}{Z^1{}_0} + 1\right)}{\theta - \sin\theta + \pi\omega^2 C_J{}^2 R_S Z^1{}_0} - 1$$

$$VSWR = \left[\frac{Z_{LO}}{Z^1{}_0}\right]^{\pm 1}$$

In order to reduce radiation of the LO from the antenna, the VSWR should be less than 1.6. This corresponds to a reflection of less than 5% of the LO power.

IF Impedance

When the diode is considered as a source of IF voltage, it is important to know what its low frequency (IF) impedance is. The IF amplifier has to be designed to work optimally when driven from a source of this impedance, or diodes and circuit conditions should be chosen to prove an optimum impedance for the input of the IF amplifier.

If an external DC bias is applied to the diode, the crystal current will change, due to a change in the conduction range. Applying a small reverse DC (or IF frequency) voltage is the same as increasing V_T by the same amount. The IF impedance is the ratio of the applied DC or IF voltage to the change in crystal current.

$$Z_{IF} = \frac{\Delta V_F}{\Delta I_{DC}} = \frac{1}{(dI_{DC}/dV_f)}$$

$$= \frac{2\pi}{\Theta}\,(Z^1{}_0 + R_S + R_B)$$

This is always greater than $2Z_0$ and typically ranges from 200 to 500 Ω.

As an example of the behavior of these parameters as LO power is varied, the following graph shows the noise figure, VSWR, crystal current and IF impedance of an X-band diode. The fixed parameters are V_F = .28 V, R_S = 7w, C_J =.20 pF, and Z_0 = 150 Ω, values appropriate for low barrier diodes in a waveguide test holder, such as those used for testing mixer diodes at Skyworks.

Performance is better at low LO power levels than these formulas indicate because actual diodes have a soft knee in the forward I-V characteristic. Also, the noise figure for actual diodes can be about 1 dB better due to harmonic suppression, but the noise figure goes up at high LO power due to heating and other effects. Nevertheless, these formulas can give you some insight into the meaning of the various RF parameters and their relationship to the capacitance and I-V characteristics of an actual diode.

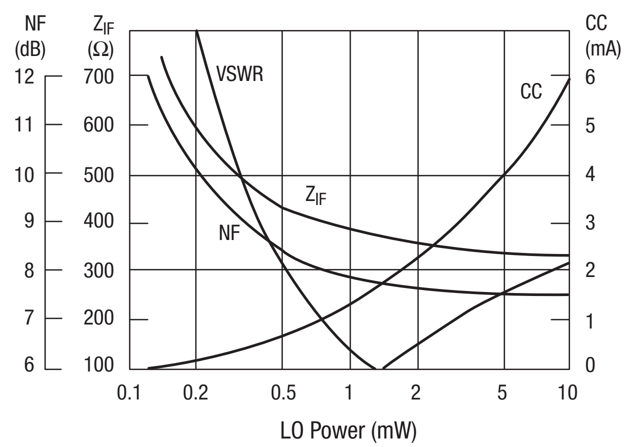


Figure 6. Mixer Parameters as a Function of LO Power

Practical Mixer Configurations

Single-Ended Mixer

The single-ended mixer used in the above analysis has some disadvantages which limit its usefulness.

1. Even with a low VSWR, too much LO power is reflected into the signal port.
2. To couple the LO and signal onto the same line with broad bandwidth requires a coupler which increases the conversion loss, noise figure and multiplies required LO power. (For example, a 6 dB coupler adds 1.2 dB to the conversion loss and noise figure and requires four times the LO power.)

3. If the coupler is unacceptable, a set of filters can be used, but if the IF and LO frequencies are close, the bandwidth will be restricted severely. However, no extra LO power is needed.
4. The mixer is very sensitive to amplitude variations (AM noise) in the LO power, which will increase the noise figure, if the AM noise spectrum overlaps the signal frequency.

Balanced Mixer

For many years, the solution to these problems was to use a balanced mixer containing two diodes driven in opposite phase. In this case, the reflected LO power cancels, but the IF output adds if the diodes are reversed. Conversion loss is the same as for the single-ended mixer.

Twice the LO power is required as for a single diode mixer. The VSWR can be much lower, and the ZIF depends on how the signals are combined (for the transformer circuits it will be half that of a single diode). The noise figure will be reduced dramatically compared to the single-ended mixer because the AM noise from the local oscillator at the signal frequency is cancelled at the IF output, provided the diodes are well enough matched.

Figure 7 shows some of the common balanced mixer configurations, as well as a practical single-ended mixer:

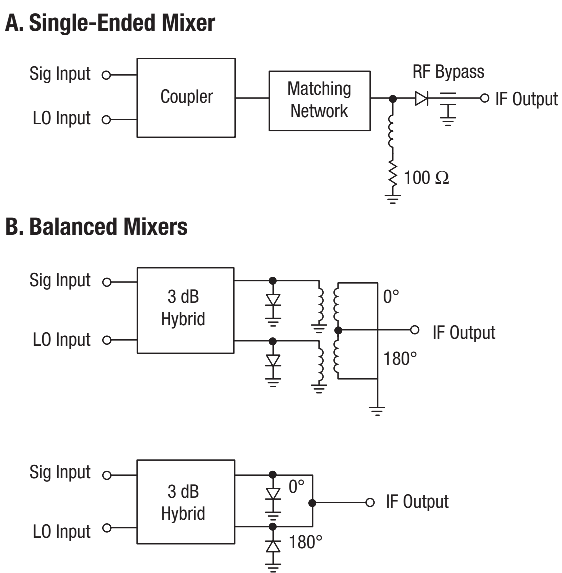


Figure 7. Single-Ended and Balanced Mixers

Double-Balanced Mixers

The use of four diodes in a ring, bridge, or star configuration makes it possible to cancel the LO reflections and noise at both the signal and IF ports, so no filtering is needed at the IF port. This requires the use of very broadband baluns or transformers. In recent years, several manufacturers have developed these double-balanced mixers to the point where bandwidths over 25 GHz are possible. To do this requires that the diodes be physically very close together to avoid inductive parasitics, and exhibit good electrical matching between all four diodes.

The best solution is to make all four diodes simultaneously in a ring configuration using beam-lead technology. (These are available mounted on various carriers, or as unmounted beam-lead quads.) Figure 8 shows one of the most common circuit configurations.

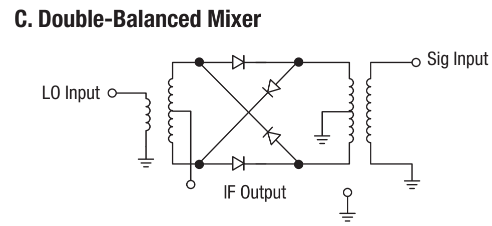


Figure 8. Ring Quad Configuration

If you know how to design broadband baluns or transformers, this kind of mixer circuit is a natural. However, you should remember that in circuits with bandwidth over one octave, harmonic enhancement cannot be used, so there is a penalty in conversion loss.

The easiest way to understand the conversion action is to consider Figure 9:

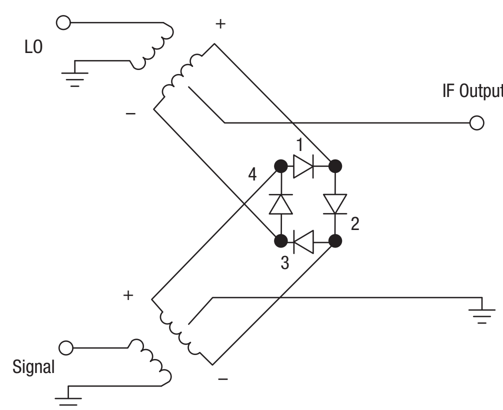


Figure 9. Ring Quad for Analysis

When LO is in "positive" phase, diodes (2) and (3) conduct, and the negative arm of the signal transformer is connected to IF. When the LO is negative, diodes (1) and (4) conduct and connect the positive arm of the signal transformer to the IF output. The two pairs of diodes therefore act like a high-speed SPDT switch. When one goes through the mathematics for the conversion loss (involving the transmission coefficient instead of the reflection coefficient) formulas for conversion loss and noise figure similar to the ones for the single-ended mixer can be derived.

Parameter Tradeoffs

Barrier Height

The barrier height of a Schottky diode is important because it directly determines the forward voltage. In order to get good noise figure the LO drive voltage, V_L, must be large compared to V_T, which is essentially V_{F1}. Normally, it is best to have a low forward voltage (low V_{F1}, or low drive) diode, to reduce the amount of LO power needed. However, if high dynamic range is important, high LO power is needed, and the diode can have a higher V_F and should also have a high V_B (see table below).

Type	Typical V_{F1}	LO Power	Application
Zero Bias	0.10–0.25	<0.1 mW	Mainly for Detectors
Low Barrier	0.25–0.35	0.2–2 mW	Low–Drive Mixers
Medium Barrier	0.35–0.50	0.5–10 mW	General Purpose
High Barrier	0.50–0.80	>10 mW	High Dynamic Range

Noise Figure vs. LO Power

At low LO drive levels, noise figure is poor because of poor conversion loss, due to too low a conduction angle. At high LO drive levels noise figure again increases due to diode heating, excess noise, and reverse conduction.

If high LO drive level is needed, for example, to get higher dynamic range, then V_B should be specified (>5 V). However, nature requires that you pay for this with higher R_S (lower f_c), so the noise figure will be degraded compared to what could be obtained with diodes designed for lower LO drive. Forward voltage and breakdown are basically independent parameters, but high breakdown is not needed or desirable unless high LO power is used.

Such a high breakdown diode will have low reverse current (which is important only if the diode has to run hot).

Silicon vs. GaAs

Typical silicon Schottky diodes have cutoff frequencies in the 80–200 GHz range, which is good for use through Ku-band.

At Ku–band and above or for image enhanced mixers, higher f_c may be needed, which calls for the use of GaAs diodes. These have lower R_S due to higher mobility, which translates to cutoff frequencies in the 500–1000 GHz range.

However, if your IF frequency is low, be careful; GaAs diodes have high 1/f noise. They also have high V_{F1},so more LO power is required.

C vs. Frequency

There is quite a lot of latitude in choosing C_J. However, in general, the capacitive reactance should be a little lower than the transformed line impedance (Z_0). If Z_0 is not known, a good way to start is to use X_C = 100 W. Experience has shown that most practical mixers use an X_C near this value (a little higher in waveguide, and lower in 50 W systems).This translates to the following “rule of thumb” for choosing the junction capacitance of a diode for operation at frequency f (in GHz):

$$C_{J0} \approx \frac{100}{\omega}$$

$$\approx \frac{1.6}{f} \quad \text{(in pF)}$$

Detectors

General

Detectors are typically used to convert low levels of amplitude modulated RF power to modulated DC. The output can be used for retrieval of modulated information, or as a level sensor to determine or regulate the RF level.

Detector diodes act as square law detectors for low-level signals. That is, the output voltage is proportional to the square of the RF voltage at the junction (i.e., proportional to the RF power). At higher signal levels, the detector will become linear, and at still higher levels, the voltage output will saturate, and not increase at all with increasing signal.

Detector Circuits

In general, a diode detector will require a single diode together with an RF impedance transformation circuit and some low-frequency components. The configuration looks like:

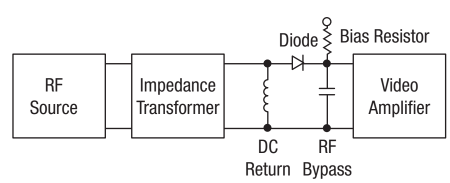


Figure 10. Typical Detector Circuit

The bias resistor generally has a very high impedance compared to the diode constant and bias the diode to a favorable impedance level.

Theory of Detection

Low Level (Square-Law)

Detection occurs because of the nonlinear I-V characteristics of the diode junction. The I-V curve of the junction is the same at microwave frequencies as at DC.

If the junction capacitance is left out of consideration for the moment, the forward I-V curve of the diode (at room temperature) is

$$I = I_S \left[exp \left(\frac{V_J}{0.028} \right) -1 \right]$$

Where $V_J - V - IR_S$ = junction voltage

If the DC current is held constant by a current regulator or a large resistor, then the total junction current, including RF, is

$$I = I_0 = i \cos \omega t$$

and the I–V relationship can be written

$$V_J = 0.028Ln \left(\frac{I_S + I_0 + i cos\, \omega t}{I_S} \right)$$

$$= 0.028Ln \left(\frac{I_0 + I_S}{I_S} \right) + 0.028Ln \left(\frac{i cos\, \omega t}{I_0 + I_S} \right)$$

If the RF current, i, is small enough, the IN-term can be approximated in a Taylor series:

$$V_J \approx 0.028Ln \left(\frac{I_0 + I_S}{I_S} \right) + 0.028 \left[\frac{i \cos \omega t}{I_0 + I_S} - \frac{i^2 cos^2 \omega t}{2(I_0 + I_S)^2} + ... \right]$$

$$= V_{DC} + V_J \cos \omega t + \text{higher frequency terms}$$

If you use the fact that the average value of $\cos^2$ is 0.50, then the RF and DC voltages are given by the following equations:

$$V_J = \frac{0.028}{I_0 + I_S} \quad i = R_S i$$

$$V_{DC} = 0.028/n \left(1 + \frac{I_0}{I_S} \right) - \frac{0.028^2}{4(I_0+I_S)^2} = V_0 - \frac{V_J^2}{0.112}$$

Therefore, the DC voltage decrease from the bias voltage, R_0, depends on the square of the RF junction voltage only. (Note, however, that the number "0.112"is really 4nkT/q and is temperature dependent.)

To get the maximum voltage sensitivity, it is clearly necessary to arrange the circuit to get the maximum possible RF voltage at the junction. That is, the impedance transformer should be designed to have the highest possible impedance at the diode, and the diode should be biased to a high enough impedance (low I_0) so the open circuit RF voltage will not be loaded down too much. In addition, C_J should be low for the same reason.

Voltage Output (Square-Law Region)

The output voltage of a detector will depend on the parasitics and circuit impedances. Suppose the impedance transformer is designed to boost the source impedance to an impedance, o, at the diode. Then the relation between V_J and the available power of the source PRF can be seen in Figure 11.

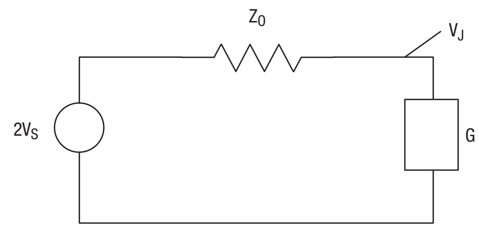


Figure 11.

$$V_S = \sqrt{2Z^1{}_0 P}$$

$$G = \frac{1}{R_s} + \frac{R_s}{X_c^2}$$

As before, the C_J is absorbed into the impedance transformation and the impedance, Z', is assumed real at the junction (i.e., C_J has been "parallel-tuned" to get the highest possible V_J):

$$V_J^2 = \frac{(2V_S)^2}{(1 + Z^i{}_0 G)^2}$$

$$= \frac{8Z^i{}_0 P_{RF}}{(1 + Z^i{}_0 G)^2}$$

The output voltage of the detector will be

$$V_{DC} - V_0 = \frac{-8Z^i{}_0 P_{RF}}{0.112\,(1 + Z^i{}_0 G)^2} = \frac{-71.4Z^i{}_0 P_{RF}}{(1 + Z^i{}_0 G)^2}$$

The impedance $Z'_{0'}$ is usually limited by bandwidth considerations or by the practical design of the impedance transformer. For a fixed $Z'_{0'}$ R_J should be as high as possible (which results in a high VSWR). Most manufacturers specify the output voltage for one microwatt RF input power.

An important special case is $Z'_0 = 50\ \Omega$, because many of the voltage sensitivity specifications are measured by placing the diode in the end of a 50 W line. If the C_J is small enough, the voltage output per unit power input for $Z'_0 - 50\ \Omega$ is

$$E_0 = \frac{V_0 - V_{DC}}{P_{RF}} = \frac{V_J^2}{0.112} = \frac{3570}{1 + \left(\frac{100}{R_B}\right)}\ \mu V/\mu W$$

Remember

$$R_S = \frac{28}{I_0 + I_S}\ \text{, (for } I_0 \text{ in mA)}$$

So for $I_0 = 50\ \mu A$; $RB = 560\ \Omega$, and therefore:

$$E_0 = 3000\ \mu V/\mu W$$

It should be pointed out that the VSWR will be very high for this kind of detector. In this case the VSWR is equal to $R_B/50$, which is over 11 if I_0=50 µA, atypical bias current.

Another important special case is when Z'O is matched to the shunt conductance, Z'O = 1/G. In this case the voltage output is

$$E_0 = \frac{18}{\frac{1}{R_S} + \left(\frac{R_S}{X_C^2}\right)}\ \mu V/\mu W$$

$$= \frac{18\ R_S}{1 + \frac{R_S + R_B}{X_C^2}}\ \mu V/\mu W$$

If the detector diodes are specified at a bias current of 50 mA (R_B – 560 W) and XC is designed to be large, then the matched output voltage is

From the previous equation, the larger X_C, the higher the output voltage, but remember that practical diodes are limited by a finite cutoff frequency so a large X_C automatically means a larger R_S.

$$E_0 = 18\ R_S = 10{,}000 \mu V/\mu W$$

In practice, it is usually sufficient to have $X_C > 20$ W and $R_X < 40$ W which results in no more than 2 dB degradation of the output voltage compared to the above equation.

Sensitivity

Tangential Signal Sensitivity (TSS)

At low power levels, sensitivity is specified by the "tangential signal sensitivity" (TSS). This is the power level that raises the DC voltage by an amount so the noise fluctuations do not drop below the level of the noise peaks with no signal. This is about 4 dB above the minimum detectable signal (MDS). Detection is so inefficient that even for wideband systems, the incoming noise (antenna noise) need not be considered. All the noise is produced in the diode and the video amplifier.

$$V_N^2 = 4kTBR_S + 2kTBR_B\left(1 + \frac{I_S}{I_0 + I_S}\right)$$

To this should be added the noise voltage due to the video amplifier, which can be expressed in terms of fictitious noise resistance, R_a, of the amplifier:

$$V_{NA}^2 = 4kTBR_a$$

The standard value of R_a is 1200 Ω.

The total noise voltage is

$$V_N^2 = 2kTB\left[R_B\left(1 + \frac{I_S}{I_0 + I_S}\right) + 2R_a + 2R_S\right]$$

Since the peak noise voltage is 1.4 times the rms noise voltage, (V_N), the condition for tangential voltage output is:

$$V_{DC} + 1.4\ V_N = V_0 - 1.4\ V_N$$

$$\text{or } V_0 - V_{DC} = 2.8\ V_N$$

For the biased diode measured in a 50 W circuit,

$$\text{Tangential Power} = \frac{2.8\ V_N}{V_{OUT}} = \frac{\left(1 + \frac{50}{R_S}\right)^2 (2.8\ V_N)}{3750}$$

$$= 0.78\left(1 + \frac{50}{R_J}\right)\sqrt{2kTB[R_B + 2R_a + 2R_S]mW}$$

The tangential sensitivity is the tangential power expressed in –dBm. For a diode with 50 µA bias (R_J= 560 W) measured with a video bandwidth of 10 MHz,this is:

$$TSS = 10\ \log_{10}(2828 V_N/V_0)$$

$$= 10\ \log_{10}\left[0.92\sqrt{2kTB[560 + 2R_a\ 2R_S}\ \right]$$

$$= 48.8\ \text{dBm for } R_a \sim 1200\ \Omega$$

Note that if the diode has high 1/f noise, the tangential sensitivity will be reduced considerably.

If the circuit is matched to the diode, the tangential sensitivity will be significantly increased. In this case the TSS is

$$\text{Tangential Power} = \left(\frac{1 + \frac{R_S R_B}{X_C^2}}{18 R_B} \right) 2.8\, V_N$$

$$= 0.157 \left(1 + \frac{R_S R_B}{X_C^2}\right) \sqrt{\frac{2kTB}{R_B} \left[1 + \frac{I_S}{I_0 + I_S} + 2\, \frac{R_S + R_B}{R_B}\right]}$$

For a zero bias detector diode, $I_0 = 0$ and
$R_S = R_0 - R_S = Z_V - R_S$ so the tangential sensitivity is:

$$TSS = 10 \log_{10} \left[0.157 \left(1 + \frac{R_S Z_V}{X_C^2}\right) \sqrt{\frac{4kTB}{Z_V} \left(1 + \frac{R_a}{Z_V}\right)} \right]$$

If you assume typical values as $X_C = 200\ \Omega$, $B = 10$ MHz
$R_a = 1200\ \Omega$, and $R_S = 20\ \Omega$, then the result is:

$$TSS = 10 \log_{10} \left[4.6 \times 10^{-5} (1 + 0.0005) \sqrt{\frac{1}{Z_V} \left(1 + \frac{1200}{Z_V}\right)} \right]$$

= -55 dBm for ZV = 2000 – 5000 Ω

Figure of Merit (FM)

The measurement of TSS is complicated by the fact that the apparent peak noise voltage may not be exactly 1.4 V_N. Depending on the intensity setting of the oscilloscope, the apparent peak noise can be much larger than this, resulting in an error of several dB in the apparent TSS.

To take the operator dependence out of the TSS measurement, FM is introduced, which is defined by

$$FM = \frac{E_0}{\sqrt{Z_V + R_N}}$$

For diodes with zero bias the TSS is calculated from the FM by the formula

$$TSS = 10 \log_{10} \frac{\sqrt{4kTB}}{FM}$$

For biased diodes, the situation is slightly more complicated

$$S = 10 \log_{10} \left(\frac{\sqrt{4kTB}}{FM} \right) + 5 \log_{10} \left(\frac{2Z_V + 2R_a}{Z_V + 2R_a} \right)$$

The relationship is even more complicated if 1/f noise is considered which may be necessary if the diode is biased.

High Voltage Output

At high signal levels, the detector will begin to deviate from square-law behavior. This begins to happen when $V_J = 0.028$ V. For these signal levels, the sensitivity can be calculated from the same formulas as for the crystal current of a mixer if V_T is replaced by $V_{F1} - V_{DC}$. At high signal levels, the diode will develop enough reverse bias to keep the crystal current at the value I_0 and the output voltage will approach twice the signal voltage, V_S. Therefore:

$$V_{DC} - V_F \cong 2V_S = -\sqrt{8Z'_0 P_{RF}}$$

This behavior is called linear detection because of the linear relationship between V_{DC} and V_S.

At higher power levels, the reverse bias behavior of the I-V curve becomes important; as the reverse voltage approaches V_B, the slope of the reverse characteristic becomes comparable to Z'_0, and begins to lead down the circuit. At a little higher power, the diode starts rectifying in the reverse direction as well as in the forward direction, and this results in a limitation of the output voltage.

The whole input/output characteristic of a detector is illustrated in Figure 12.

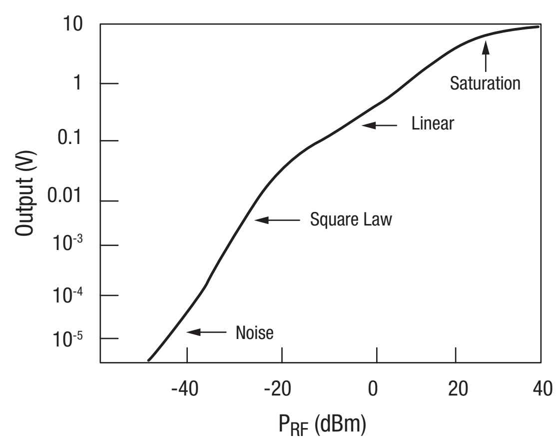


Figure 12. Detector Output Characteristics

1/f Noise

Excess noise due to surface static and traps often has1/f frequency spectrum instead of the uniform spectrum characteristic of thermal noise and shot noise. That is, the noise power per unit bandwidth has a behavior:

$$\Delta\,(V^2{}_{N1}) \sim \frac{A}{f}\;\Delta f$$

To find the total noise voltage, the actual lower frequency limit, f_L, of the video amplifier must be known.

$$V^2{}_{N1} = \int_{f_L}^{f_L + B_A} \frac{A}{f}\,df = A/n\left(\frac{f_L + B}{f_L}\right)$$

Combining this with the thermal and shot noise expressions gives

$$V^2{}_{N1} = A\;Ln\left(1 + \frac{B}{f_L}\right) + 2kTB\left[R_S\left(1 + \frac{I_S}{I_O + I_S}\right) + 2R_a + R_S\right]$$

It is convenient to eliminate the constant A by defining a noise corner frequency f_N, the frequency at which the 1/f noise is equal to the shot noise.

$$f_N = \frac{A}{2KTR_J}$$

In terms of noise corner,

$$V^2{}_{N1} = 2kTB\left\{R_B\left[1 + \frac{I_S}{I_O + I_S} + \frac{f_N}{B}/n\left(1 + \frac{B}{f_L}\right)\right] + 2R_a + R_S\right\}$$

The noise corner can be specified for a diode, but this is complicated by the fact that for typical diodes the excess noise does not have an exact 1/f spectrum, and also because the noise corner can depend on bias conditions. At Skyworks, the 1/f noise output is measured in a bandwidth of 60 kHz (with f_L = 8 Hz) as a measure of 1/f noise. This is sufficient as a qualitative measurement of noise corner frequency, since V_N^2 is proportional to fN. It is interesting to note that for a 50 µA biased diode with a noise corner of less than 3 kHz, the noise output will be less than a 560 W resistor.

Detector Configuration

High Sensitivity

In this type, an impedance transformer is used to raise the impedance to as high a value as practical. Ideally, this should be the zero bias resistance of the diode, but this approach is limited by the R_S and C_J. It is also limited by bandwidth considerations and losses in the impedance transformer. Narrow-band detectors with voltage outputs of 10–30 mV/µW can be achieved this way. Tangential sensitivity approaching -70 dBm (in a << 1 MHz video bandwidth) is achievable with good diodes, high Z_0 (over 10 K), and low noise video amplifiers. Even higher sensitivity can be obtained by reducing the video bandwidth. A schematic is shown in Figure 13.

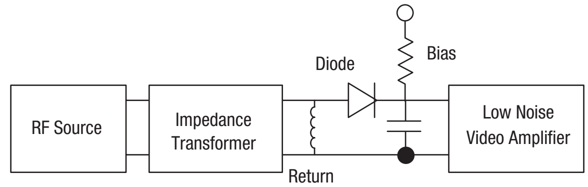


Figure 13. Typical Detector Circuit–High Sensitivity

Wideband

A detector circuit uses a wider band impedance transformer or balun and is limited to a much smaller impedance at the diode, usually 50–200 Ω. For the 50 Ω type the best voltage sensitivity is 3600 uV/uW, (unless the diode package increases the impedance at the chip above 50 Ω), and tangential sensitivities are limited to about -54 dBm (in a 50 MHz band). The configuration is shown in Figure 14.

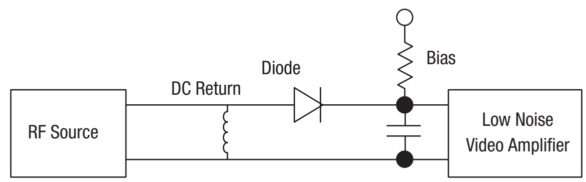


Figure 14. Typical Detector Circuit—Wideband

Flat Detector

The above configuration has a reasonable flat response if the RF source is well matched, but has a high VSWR. Therefore, it is sensitive to any mismatch in the source which will then reflect back some of the reflected signal. To avoid this, a 50 Ω resistor can be included to eliminate the reflections, but this halves the signal voltage available at the diode, and reduces the output to less than 1 mV/µW, and the TSS will not be more than -48 dBm. However, the extremely wide bandwidth and low VSWR of this type of detector make it very useful. The circuit is shown in Figure 15.

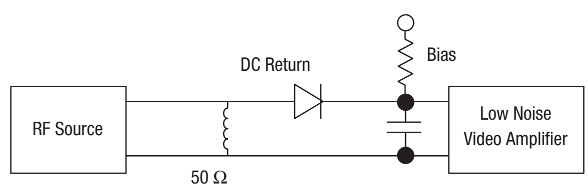


Figure 15. Typical Detector Circuit—Flat Response

Matched Pairs

Detectors that must operate over a temperature range, or must be insensitive to variations of bias supply voltage, must have the reference voltage, Vo, built into the detector. This can be done by using an identical diode as a reference. For this reason, detectors are often sold in matched pairs. A typical circuit is be shown in Figure 16.

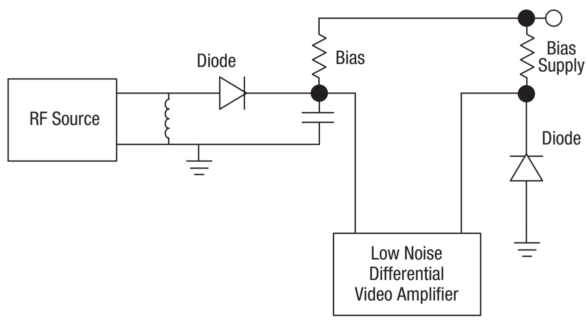


Figure 16. Temperature Compensated Detector

Parameter Tradeoffs

Bias vs. No Bias

Although the zero-bias detector diode looks like a good way to reduce circuit complexity, applying bias to a diode reduces the noise temperature of the resistance R_B at video frequencies. In addition, the bias resistor can be chosen to compensate for the natural temperature variation of R_J (which is proportional to absolute temperature in $°_K$ for constant current). That is, if the resistance is inversely proportional to T, then R_B will be constant over temperature. The video impedance of a zero-bias diode is very temperature dependent. However, a diode operated at zero bias has no 1/f noise. Therefore, this type of diode is the choice for audio frequency output, such as motion detectors. The lack of bias resistor also simplifies the design of impedance matching networks for narrowband, high sensitivity detectors.

Caution should be used in selecting diodes for use in unbiased detector circuits because deviation from square-law behavior can occur at low levels. If a mixer diode or a detector diode not designed for zero-bias operation is used without bias, the small signal resistance, R_B, (video impedance) will be too high. In this case, it will be impossible to get a good match to the diode, even over a narrow bandwidth, and the RF power will be dissipated in lossy circuit elements. Thus the RF voltage at the junction will be much less than it should be, resulting in lower TSS and voltage sensitivity at very low signal levels. When the signal level is increased, the diode self-biases to a lower resistance, R_B, and more of the power reaches the diode. Therefore, the voltage sensitivity increases. The net result is that the detected response is faster than square law at very low signal levels, approaching fourth law or fifth law in many cases. This results in substantial error if a square-law characteristic is assumed, as in many power level measurement applications. This effect does not happen if a zero-bias Schottky diode is used, properly matched, in a low loss detector mount.

C_J vs. Frequency

For most purposes, it is sufficient to have $X_C > 150$ W in a detector diode. This leads to the following “rule of thumb” (for C_{J0} in pF):

$C_{J0} < 1.1/f$ (f = signal frequency in GHz)

which is good for “typical” detectors. However, this is usually too stringent for 50 Ω detectors, especially flat detectors. Conversely, in the case of high output detectors, the C_J may not allow enough bandwidth. In this case, lower C_J should be traded for more R_S, since R_S matters less in detectors than in mixer diodes. Some detector designers use diodes with R_S as high as 100 Ω.

1/f Noise

Detector diodes are usually used in systems whose video bandwidth extends below 10 kHz. In this case 1/f noise voltage becomes much more important than for typical mixer diodes. It can be specified by a noise corner frequency, or by an upper limit or the noise output in a particular audio band. Skyworks diodes are screened using an audio amplifier with a response from 8 Hz to 60 kHz (at 50 µA bias) when low 1/f noise is specified.

Burnout

General

Schottky barrier diodes are more subject to burnout due to incident RF pulses than are typical junction diodes, even the very small junction diodes used in microwave systems. Basically, there are three reasons for this:

1. The barrier diameters are very small (less than .5 mil diameter), resulting in high dissipated power density.
2. The metal semiconductor contact is not as stable chemically as a junction between two regions deep within a semiconductor, and can be damaged by temperatures on the order of 400 °C.
3. Because of lack of charge storage (conductivity modulation) the resistance of the diode at high currents will not be very low (typically around 10 Ω). Therefore, the diode does not protect itself as well as junction diodes, whose dynamic resistance may drop to a few tenths of an Ω at high forward currents or high incident RF power.

Dependence of Burnout Power on Pulse Length

A diode will begin to degrade when some part of the junction reaches a certain temperature. The exact temperature depends on the metallurgy used, and on the degree of perfection of the junction, especially at the edges. All of the metallurgies used in Skyworks Schottky diodes are good for at least 350 °C.

For RF pulses less than 5 ns long, the temperature rise is directly proportional to the total pulse energy dissipated in the epitaxial layer just under the barrier metal. This would appear to lead to the conclusion that the energy content of the RF pulse determines whether the diode will burn out, but the situation is not that simple. For example, if the incoming RF pulse has a peak-to-peak voltage (at the diode) less than the diode breakdown, there will be relatively little dissipation in the junction. At higher pulse voltages, the percentage of the incoming energy that is dissipated will increase. The amount of dissipation in the diode will also depend on the circuit, which determines what happens to the energy reflected by the diode. All that can be said without exact knowledge of both the diode and the circuit is that the susceptibility of the diode to burnout is related to both the power (or voltage) in the incoming RF pulse and pulse duration.

For longer pulse lengths (5 ns to 100 ns) the temperature of the diode junction is dominated by thermal diffusion, and the temperature rise will be proportional to the square root of time for a given power dissipation. Therefore, the burnout is not expected to depend on the total dissipated energy for pulse lengths over 5 ns, but is more related to the incident power (if the peak-to-peak voltage is high enough).

If the pulse length is longer than about 100 ns, the maximum junction temperature is controlled by the thermal resistance of the chip and package. In this case, the burnout rating will depend to some extent on the quality of the heat sink used for the diode.

Burnout vs. Frequency

Because the capacitance of mixer diodes must be smaller at higher frequencies, smaller diameter junctions are used. This, of course, makes higher frequency diodes more susceptible to burnout than low-frequency diodes. For short pulses, the burnout power is approximately inverse with frequency, whereas for long pulses, or CW, the effect is more gradual.

Detector diodes typically have lower capacitance and thus smaller junctions than mixer diodes. This is often not an issue, because detector diodes are not usually exposed to high power RF pulses. However, if the system requires that they be exposed, then the burnout rating should be given serious consideration in selecting the diode.

Transients and Electrostatic Discharges

For the same reasons outlined above, Schottky diodes are subject to burnout due to circuit transients and electrostatic discharges. (The majority of diode burnout problems we encounter are due to these two causes.)

Electrostatic discharge is becoming even more of a problem than it used to be, since most people wear plastic clothes and shoes. A person’s hand can easily acquire a charge of over 5000 V on a dry winter day, and when it touches the diode, it can release as much as 10 amperes of short circuit current in less than a nanosecond. The solution is to always ground your hand, tweezers, pliers, or any other tool before touching the diode. (Also, both terminals of the circuit it goes into should be grounded—someone may have touched one of the conductors and charged it.)

Another way of damaging diodes is to check the front-to-back ratio with a conventional multimeter to see if it is still a diode. (It won’t be.) The ohmmeter batteries in a typical multimeter range from 1.5–9 V, and the leads will be charged to this voltage until they touch the diode. The discharge is usually sufficient to burn out the diode within about 2 nanoseconds (the longer the leads, the worse the effect). This effect can be avoided by using a push-to-test switch across the diode when testing it in this way, or by using a curve tracer instead of a multimeter. Some DVMs are just as bad as multimeters, because they produce digital pulses which hit the diode.

Switching transients in actual circuits can cause the same effect, if there is sufficient inductance between the source and the diode. This can be eliminated by using a small capacitor between the source of the transient and the diodes.