INTRODUCTION

Radiocommunication requires that we shift a baseband information signal to a frequency or frequencies suitable for electromagnetic propagation to the desired destination. At the destination, we reverse this process, shifting the received radiofrequency signal back to baseband to allow the recovery of the information it contains. This frequency-shifting function is traditionally known as *mixing*; the stages that perform it, as *mixers*. Any device that exhibits amplitude-nonlinear behavior can serve as a mixer, as nonlinear distortion results in the production, from the signals present at the input of a device, of signals at new frequencies. Even a rusty screw or bolt on an antenna element can act as a mixer, producing unwanted IMD products that appear at the receiver input.

Although mixers are equally important in wireless transmission and reception, traditional mixer terminology favors the receiving case because mixing was first applied as such in receiving applications. Thus, the signal to be frequency-shifted is applied to the mixer's *RF* port, and the frequency-shifting power or voltage (from a *local oscillator* [*LO*]) is applied to the mixer's LO port, resulting in two outputs at the mixer's *intermediate frequency* (*IF*) port. If the wanted IF is lower in frequency than the RF signal, the mixer is a *downconverter*; if the wanted IF is higher than the RF, the mixer is an *upconverter*. *Converter* may also be used as a term for a single stage that simultaneously acts as mixer and LO.

For a given RF signal, an ideal mixer with a perfect LO (that is, an LO with no harmonics and no noise sidebands) would produce only two IF outputs : one at the frequency sum of the RF and LO, and another at the frequency difference between the RF and LO. Filtering can be used to select the desired IF product and reject the unwanted one, which is sometimes referred to as the *IF image*.

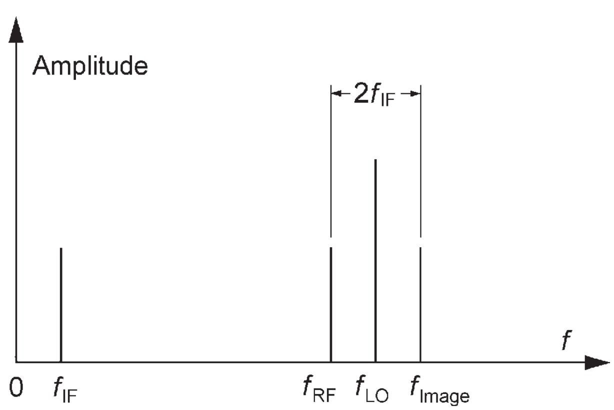


Figure 1 - Relationship between a mixer's image and desired-signal responses. The image is $2f_{IF}$ away from the desired signal.

The simultaneous generation of LO+RF and LO–RF outputs results not from a departure of mixer performance from the ideal, but from the mathematics of mixing itself. Another unavoidable mixing artifact, the *RF image* response, also results from the mathematics of mixing rather than mixer nonideality. Just as a given RF/LO combination produces two IF outputs (LO+RF and LO–RF, the IF and IF image), the mixer will produce output at the desired IF (LO+RF *or* LO–RF) in response to *two* possible RF inputs: one at LO+IF and another at LO–IF (Figure 1). The undesired response, the *RF image* (traditionally referred to as merely as the *image*), is $2f_{IF}$ removed from the desired response. Even if no manmade signals exist at the RF image frequency, reducing a mixer's RF image response can be important because noise at that frequency, including that produced by circuitry between the mixer and antenna, will still be mixed to the desired IF, degrading the signal-to-noise ratio. Filtering and phasing techniques can be used to reduce the RF or IF image responses—filtering if the image is sufficiently removed from the desired response for filtering to provide the necessary rejection, phasing if the desired and image responses are insufficiently spaced for filtering to work, as in the case of a double-conversion receiver in which signals at a high first IF (for example, 50 to 70 MHz), must be converted to a very low first IF, such as 25 kHz.

The output of every real mixer includes a vast number of additional unwanted products, including noise, the fundamentals of the mixer's RF and LO signals and their harmonics, and the sums and differences of the RF and LO and their harmonics. Intermodulation distortion between multiple signals present at the RF port, and IF output resulting from the mixing to IF of LO noise-sideband energy by strong adjacent signals further complicate a mixer's output spectrum and may compromise system performance.

All mixers are *multipliers* in the sense that the various new outputs they produce can be described mathematically as the multiplicative products of their inputs.

Let us now consider the basic theory of mixers. Mixing is achieved by the application of two signals to a nonlinear device. Depending upon the particular device, the nonlinear characteristic may differ. However, it can generally be expressed in the form:

$$I = K\left(V + v_1 + v_2\right)^n \quad \text{(1-1)}$$

The exponent n is not necessarily an integer, V may be a dc offset voltage, and the signal voltages v_1 and v_2 may be expressed as $v_1 = V_1 \sin(\omega_1 t)$ and $v_2 = V_2 \sin(\omega_2 t)$.

When $n = 2$, (1-1) may then be written as:

$$I = K\left[V + V_1 \sin(\omega_1 t) + V_2 \sin(\omega_2 t)\right]^2 \quad \text{(1-2)}$$

This assumes the use of a device with a square-law characteristic. A different exponent will result in the generation of other mixing products, but this is not relevant for a basic understanding of the process. Expanding (1-2),

$$I = K[V^2 + V_1^2 \sin^2(\omega_1 t) + V_2^2 \sin^2(\omega_2 t) + 2\,V\,V_1 \sin(\omega_1 t) + 2\,V\,V_2 \sin(\omega_2 t) + 2\,V_2\,V_1 \sin(\omega_2 t)\sin(\omega_1 t)] \quad \text{(1-3)}$$

The output comprises a direct current and a number of alternating current contributions. We are interested only in that portion of the current that generates the IF; so, if we neglect those terms that do not include both V_1 and V_2, we may write:

$$I_{IF} = 2\,K\,V_1\,V_2 \sin(\omega_1 t)\sin(\omega_2 t)$$
$$I_{IF} = K\,V_2\,V_1 \{\cos[(\omega_2 - \omega_1)t] - \cos[(\omega_2 + \omega_1)t]\} \quad \textbf{(1-4)}$$

This means that at the output, we have the sum and difference signals available, and the one of interest may be selected by the IF filter.

PROPERTIES OF MIXERS

1-1-1 Conversion Gain/Loss

Even though a mixer works by means of amplitude-nonlinear behavior in its device(s), we generally want (and expect) it to act as a linear frequency shifter. The degree to which the frequency-shifted signal is attenuated or amplified is an important mixer property. *Conversion gain* can be positive or negative; by convention, negative conversion gains are often stated as *conversion loss*.

In the case of a diode (passive) mixer, the insertion loss is calculated from the various loss components:

Loss (dB) = Conversion Loss + Transformer Loss
+ Losses due to harmonic generation + Diode Loss **(1-5)**

In the case of a doubly balanced mixer, we must add the transformer losses (on both sides) and the diode losses as well as the mixer sideband conversion, which accounts, by definition, for 3 dB. Ideally, the mixer produces only one upper and one lower sideband, which results in the 3-dB loss compared to the input signal. Also, the input and output transformers add about 0.75 dB on each side, and of course there are the diode losses because of the series resistances of the diodes. This total loss is still defined as conversion loss for all Synergy mixers.

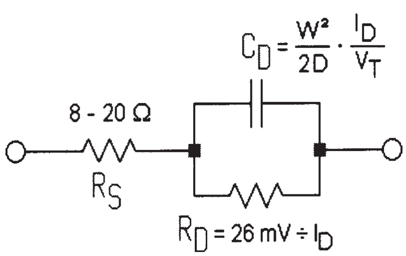


Figure 2—Equivalent circuit of a mixer diode.

Figure 2 shows the equivalent circuit of a diode. It consists of a series (loss) resistor R_S and a time-variable electronic resistor, typically called the *diffusion resistance*, R_D, and a capacitance C_D shunting R_D. C_D can be found from

$$C_D = \frac{W^2}{2D}\frac{I_D}{V_T} \quad \textbf{(1-6)}$$

where D is the diffusion constant, a material-dependent value, and W is the physical width. The average value for R_D is somewhere between the calculated value of 26 mV/I_D and some leakage current, simply because it is generated by a rectification mechanism, which turns the LO power into an RF current and then into a combination of dc and RF currents.

We can calculate the diode loss according to:

$$\text{Diode Loss (dB)} = \log 10\left(\frac{50 + (2 \times R_S)}{50}\right) \quad \textbf{(1-7)}$$

Assuming that R_S = 8 Ω, the diode loss for a diode-ring mixer,

$$\text{Diode Loss (dB)} = \log 10\left(\frac{50 + (2 \times 8)}{50}\right) = 0.5\ \text{dB} \quad \textbf{(1-8)}$$

From (1-5), the insertion loss for this mixer is therefore:

Loss (dB) = 3 dB (conversion loss) +
1.5 dB (transformer loss) +
1 dB (losses from harmonic generation) +
0.5 dB (diode loss) = 6 dB **(1-9)**

This assumes mixing at the fundamental frequency. Sometimes the diode loss resistor R_S is as high as 25 Ω per arm (as in a MOSFET switch), or 50 Ω total. This now results in

Loss (dB) = 3 dB + 1.5 dB +
1 dB + 3 dB = 8.5 dB (Insertion Loss) **(1-10)**

Since the value of R_S is partially determined by the threshold voltage of the diode and the diode diffusion resistance, R_D, a wide range of values can be noticed for different drive levels and mixer topologies. Figure 2 shows a shunt capacitance C_D, the so-called *diode diffusion capacitance*. When the diode is conducting, the influence of this nonlinearity is frequency-dependent, which adds to the insertion loss. In this discussion we have not considered its frequency-dependency. At wireless frequencies, modern Schottky diodes, also frequently called *hot-carrier* diodes, are operated far from their cutoff frequency, resulting in less than 1 dB of additional losses. There are also mixers with special circuitry to terminate the IF image. This is done with a diplexer circuit or equivalent circuitry.

1-1-2 Noise Figure

Like any network, a mixer contributes noise to the signals its frequency-shifts. The degree to which a mixer's noise degrades the signal-to-noise ratio of the signals its frequency-shifts is evaluated in terms of noise factor and noise figure.

For a long time, the literature has stated that the noise figure of a passive mixer, which is pretty much independent of its circuit arrangement, is equal to the mixer's conversion loss. But this neglects the influence of the white-noise contribution of the mixer's diode(s). This is ironic, considering that RF noise generators were long based on thermionic diodes operated in saturation (the 5722 noise diode was a popular type). Such a diode's noise-power output can be readily determined from its saturation current. On the other hand, all Schottky diodes, while conducting, generate white noise that follows the same principle as above. This fact has been practically recognized only by a few companies that make modern noise-measurement equipment. Modern noise-measurement devices measure the noise figure of a system by "hot and cold" technique, an approach based on knowledge of the absolute noise energy emitted under hot conditions (conductance). This method has the advantage that it can be used up to several tens of gigahertz, while the old vacuum-tube-based noise generators ran out of steam at around 1 GHz due to the inability to match the tube to the 50-Ω termination. This was typically accomplished by connecting a 50-Ω resistor between anode and ground (without dc connection), followed by a low-pass filter, which would match the tube capacitance and other parasitics to the required termination of 50 Ω, purely resistive.

In reality, we can take the loss calculation from above and add the Schottky noise generated by the diodes as they are driven by the local oscillator.

If a Schottky diode is used as a noise generator in the conductive mode, it generates a continuous frequency spectrum, possibly up to several gigahertz. There is a mathematical relationship between the noise power spectrum emitted by the diode and the time-averaged current of this diode, which generates the noise. If the noise source impedance is set (typically 50 Ω), the available noise power can be calculated according to:

$$I_R = \sqrt{2e \times I_s \times \Delta f} \tag{1-11}$$

where
$e = 1.6 \times 10^{-19}$ coulombs
I_s = saturation current of the diode
Δf = effective noise bandwidth
For $S_{11} = 0$ or proper termination of this circuit ($R_G = R_{term}$):

$$P_R = \left(\frac{I_R}{2}\right)^2 \times R_i = \frac{e}{2} \times I_s \times \Delta f \times R_i \tag{1-12}$$

Calculated at a bandwidth of 1 Hz:

$$\frac{P_R}{\Delta f} = \frac{e}{2} \times I_s \times R_i \tag{1-13}$$

If

$$\frac{P_R}{\Delta f} = kT_0 \times F \tag{1-14}$$

the noise factor becomes

$$F = \frac{e \times I_s \times R_i}{2kT_0} \tag{1-15}$$

If the values for e and kT_0 are inserted,

$$F = 20 \times I_D \times \frac{26\text{ mV}}{I_D} + \frac{R_S + R_G}{2R_G} \tag{1-16}$$

Example. Assume the passive mixer mentioned above with its 6-dB conversion loss is considered and a dc current of 15 mA results as a function of the LO drive. Since I_D gets cancelled, the noise factor (F) of the diode portion equals

$$F = 20 \times I_D \times \frac{26\text{ mV}}{I_D} + \frac{R_S + R_G}{2R_G} = 0.52 + 0.58 = 1.1 \tag{1-17}$$

The noise figure, NF, is 10 log F, or 0.413 dB. Now this number and the insertion loss must be added. The resulting noise figure would be 6.413 dB. This is consistent with published measurement data.

Exact mathematical nonlinear approach. The exact noise factor of a real mixer is computed by the formula

$$F = \frac{N_0(\omega_{IF}) + kT_0}{K_B T_0 G_{Tc}(\omega_{RF})} \tag{1-18}$$

where

$N_0(\omega_{IF})$ = total noise power (per unit bandwidth) delivered to the IF load at intermediate frequency
K_B = Boltzmann's constant
T_0 = Reference temperature (290 or 300 K is commonly used)
$G_{Tc}(\omega_{RF})$ = Transducer conversion gain from ω_{RF} to ω_{RF}.
Let us now further elaborate on (1-18). We may write

$$N_0(\omega_{IF}) = N_S(\omega_{IF}) + N_{INT}(\omega_{IF}) + N_L(\omega_{IF}) + kT_0 \tag{1-19}$$

where N_S is noise-generated by the RF source resistance and transferred to the IF load through frequency conversion, N_{INT} is noise generated internally to the mixer, and N_L is noise generated by the IF termination. If the source resistance is held at temperature T_0, N_S will basically originate from noise generated at the RF and image frequencies, which are transferred to the IF with approximately the same conversion loss, plus a relatively small contribution transferred from other sidebands with a smaller conversion loss. We may then write synthetically

$$N_S(\omega_{IF}) = 2aK_B T_0 G_{Tc}(\omega_{RF}) \tag{1-20}$$

where a is a coefficient slightly larger than 1. $N_{INT}(\omega_{RF})$ is generated by transformer losses, by the diode Schottky noise and by the diode resistive parasitics, and in principle may take on any value; in particular, it may be zero if both the transformers and the diodes are ideal (that is, if the latter are pure nonlinear resistors). As for $N_L(\omega_{IF})$, by Nyquist's theorem the IF load resistor R_L may be described as a noiseless resistor in series with a noise voltage source whose mean-square voltage (per unit bandwidth) is

$$|V_L|^2 = 4K_B T_L R_L \qquad \textbf{(1-21)}$$

where T_L is the IF termination temperature. If the IF load is driven by a source with an output impedance $Z_{out}(\omega_{IF})$, the noise power actually delivered to the load will obviously be

$$N_{out} = \frac{4K_B T_L R_L^2}{|Z_{out}(\omega_{IF}) + R_L|^2} \qquad \textbf{(1-22)}$$

In addition to N_{out}, the thermal noise originating from the IF termination delivered to the IF load at ω_{IF} will also include contributions from other sidebands that are back-converted by the mixer nonlinearities with a relatively small conversion gain. Thus we may write

$$N_L(\omega_{IF}) = \frac{4bK_B T_L R_L^2}{|Z_{out}(\omega_{IF}) + R_L|^2} \qquad \textbf{(1-23)}$$

where b is slightly larger than 1. If we now introduce the mixer conversion loss, namely

$$L_C = \frac{1}{G_{Tc}(\omega_{RF})} \qquad \textbf{(1-24)}$$

and combine (1-18) with (1-19), (1-20) and (1-23), we finally get the noise factor expression

$$F = 2a + \frac{N_{INT}(\omega_{IF}) + kT}{K_B T_0} L_C + \frac{4bT_L R_L^2}{T_0 |Z_{out}(\omega_{IF}) + R_L|^2} L_C \qquad \textbf{(1-25)}$$

In the normal region of operation of the mixer (sufficient LO drive) we may assume

$$Z_{out}(\omega_{IF}) R_L \qquad \textbf{(1-26)}$$

so that (1-25) becomes

$$F = 2a + \frac{N_{INT}(\omega_{IF}) + kT}{K_B T_0} L_C + \frac{bT_L}{T_0} L_C \qquad \textbf{(1-27)}$$

Multiplying the $\log_{10}$ of F by 10 gives us the exact mixer noise figure in dB.

Table 1-1 shows how the noise figure and conversion loss vary with LO power for a generic diode DBM (Figure 3). “Starving” a diode mixer by decreasing its LO drive rapidly degrades its performance in all respects.

Table 1-1 Noise Figure and Conversion Loss versus LO Power for a Diode DBM

LO Power **(dBm)**	***NF*** **(dB)**	***Conversion Loss*** **(dB)**
–10.0	45.3486	–45.1993
–8.0	32.7714	–32.5264
–6.0	19.8529	–19.2862
–4.0	12.1154	–11.3228
–2.0	8.85188	–8.05585
0.0	7.26969	–6.51561
2.0	6.42344	–5.69211
4.0	5.85357	–5.15404
6.0	5.50914	–4.84439
8.0	5.31796	–4.66871
10.0	5.19081	–4.54960
12.0	5.08660	–4.45887
14.0	4.99530	–4.38806
16.0	4.91716	–4.33322
18.0	4.85920	–4.29407
20.0	4.82031	–4.26763

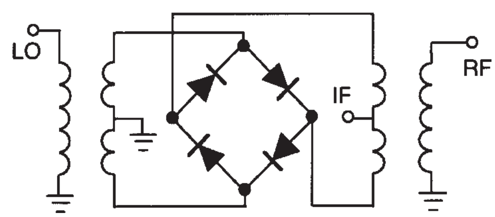


Figure 3 — Generic diode DBM.

1-1-3 Linearity

1-dB compression point. Like other networks, a mixer is amplitude-nonlinear above a certain input level; above this point, the output level fails to track input-level changes proportionally. This figure of merit, P_{-1dB}, identifies the single-tone input-signal level at which the output of the mixer has fallen 1 dB below the expected output level. The 1-dB compression point in a conventional double-balanced diode mixer is approximately 6 dB below the LO power. For Triple Balanced mixers, it is usually 3 dB below the LO power.

1-dB desensitization point. This specification is another figure of merit similar to the 1-dB compression point. However, the 1-dB desensitization point refers to the level of an interfering (undesired) input signal that causes a 1-dB decrease in nominal conversion gain for the desired signal. For a diode-ring Doulbe Balanced Mixers (DBM), the 1-dB desensitization point is usually 2 to 3 dB below the 1-dB compression point.

Dynamic range. The dynamic range of any RF/wireless system can be defined as the difference between the 1-dB compression point and the MDS (minimum discernible signal). These two points are specified in units of power (dBm), giving dynamic range in dB. When the RF input level approaches the 1-dB compression point, harmonic and intermodulation products begin to interfere with the system performance. High dynamic range is obviously desirable, but cost, power consumption, system complexity, and reliability must also be considered.

Harmonic intermodulation products (HIP). These are spurious products that are harmonically related to the f_{LO} and f_{RF} input signals.

$$\text{HIP} = Mf_{LO} + Nf_{RF} \quad \text{(1-28)}$$

Table 1-2 shows relative harmonic intermodulation product levels for a high-level diode DBM.

Intermodulation distortion (IMD). Nonlinearities in the mixer devices give rise to intermodulation distortion products whenever two or more signals are applied to the mixer's RF port. Testing this behavior with two (usually closely spaced) input signals of equal magnitude can return several figures of merit depending on how the results are interpreted. A mixer's third-order output intercept point ($IP_{3,out}$) is defined as the output power level where the spurious signals generated by $(2f_{RF1} \pm f_{RF2}) \pm f_{LO}$ and $(f_{RF1} \pm 2f_{RF2}) \pm f_{LO}$ are equal in amplitude to the desired output signal as shown in Figure 4.

Table 1-2 Typical Spurious Responses of High-Level Double-Balanced Mixer (Decibels Below $f_{LO} \pm f_{RF}$ Response)

RF Input Signal Harmonics		f_{LO}	$2f_{LO}$	$3f_{LO}$	$4f_{LO}$	$5f_{LO}$	$6f_{LO}$	$7f_{LO}$	$8f_{LO}$
$8f_{RF}$	100	100	100	100	100	100	100	100	100
$7f_{RF}$	100	97	102	95	100	100	100	90	100
$6f_{RF}$	100	92	97	95	100	100	95	100	100
$5f_{RF}$	90	84	86	72	92	70	95	70	92
$4f_{RF}$	90	84	97	86	97	90	100	90	92
$3f_{RF}$	75	63	66	72	72	58	86	58	80
$2f_{RF}$	70	72	72	70	82	62	75	75	100
f_{RF}	60	0	35	15	37	37	45	40	50
		60	60	70	72	72	62	70	70

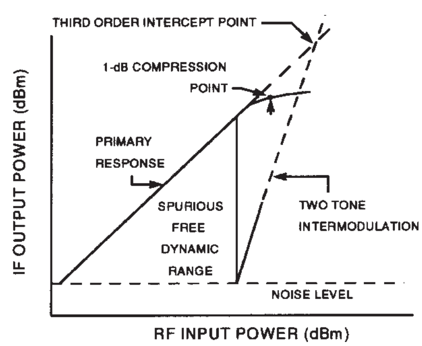


Figure 4 — Mixer linearity evaluation, including compression and two-tone IMD dynamic range. P_{-1dB} for a single-tone cannot be read directly from this graph because the values shown are the result of two-equal-tone drive.

The third order input intercept point, $IP_{3,in}$—IP_3 referred to the input level—is of particularly useful value and is the most commonly used mixer IMD figure of merit. $IP_{3,in}$ can be calculated according to:

$IP_{n,in}$ = IMR ¸ (*n*-1) + input power (dBm) **(1-29)**

where IMR is the intermodulation ratio (the difference in dB between the desired output and the spurious signal, and *n* is the IM order—in this case, 3). In a conventional diode double-balanced mixer, $IP_{3,in}$ is approximately 14 dB above the single-tone 1-dB compression point (P_{-1dB})—approximately 8 dB greater than the local oscillator power.

Although designers are usually more concerned with odd-order IM performance, second-order IM can be important in wideband systems (systems that operate over a 2:1 or greater bandwidth).

1-1-4 LO Drive Level

A mixer's specifications are usually guaranteed at a particular LO drive level, usually specified as a dBm value that may be qualified with a tolerance. Insufficient LO drive degrades mixer performance; excessive LO drive degrades performance and may damage mixer devices. Commercially available diode mixers are often classified by LO drive level; for example, a "Level 17" mixer requires 17 dBm of LO drive.

1-1-5 Interport Isolation

In a mixer, isolation is defined as the attenuation in dB between a signal input at any port and its level as measured at any other port. High isolation numbers are desirable. Figure 5 shows LO-to-IF and LO-to-RF isolation versus frequency for a triple-balanced diode mixer. Isolation is dependent mainly on transformer and physical symmetry, and device balance, but the level of signals applied to the mixer also plays a role, as shown in Figure 6.

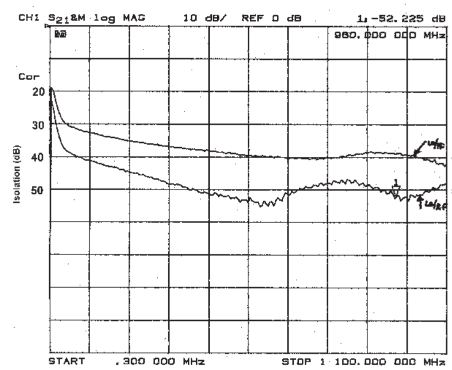


Figure 5 — LO-IF and LO-RF isolation versus frequency for a high-level triple-balanced diode mixer.

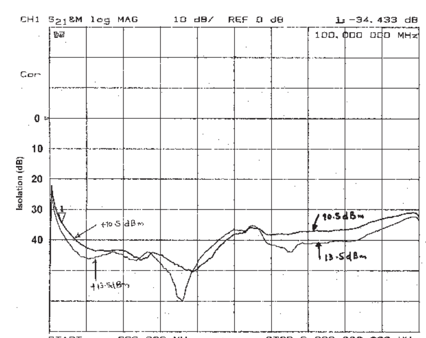


Figure 6 — LO-IF isolation versus frequency and LO drive level for a high-level diode DBM.

1-1-6 Port VSWR

The load presented by a mixer's ports to the outside world can be of critical importance to a designer. For example, high LO-port VSWR may result in inefficient use of available LO power, resulting in LO starvation (underdrive) that degrades the mixer's performance. Figure 7 shows LO-port return loss versus frequency for a high-level diode DBM with two values of LO power. Like interport isolation, port VSWR can vary with the level of the signal applied.

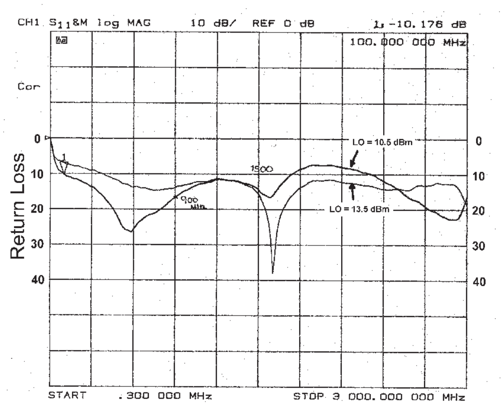


Figure 7 — LO-port return loss versus frequency for a high-level diode DBM.

1-1-7 Dc Offset

Isolation between ports plays a major role in reducing dc offset in a mixer. Like isolation, dc offset is a measure of the unbalance of the mixer. In phase-detector and phase-modulator applications, dc offset is a critical parameter.

1-1-8 Dc Polarity

Unless otherwise specified, mixers with dc output are designed to have negative polarity when RF and LO signals are of equal phase.

1-1-9 Power Consumption

Circuit power consumption is always important, but in battery-powered wireless designs it is *critical*. Mixer choice may be significant in determining a system's power consumption, sometimes in ways that seem paradoxical at first glance. For instance, a passive mixer might seem to be a power-smart choice because it consumes *no* power—until we factor in the power consumption of the circuitry needed to provide the (often considerable) LO power a passive mixer requires. If a mixer requires a broadband resistive termination that will be provided by a post-mixer amplifier operating at a high standing current, the power consumption of the amplifier stage must be considered as well. Evaluating the suitability of a given mixer type to a task therefore requires a grasp of its ecology as well as its specifications.

The simple single-diode-mixer circuit shown in Figure 1-1 is intended only as an illustration of the basic behavior of diode mixer. A practical single-diode mixer would include filtering at its RF, LO and IF ports—RF filtering for image rejection, reduced LO radiation and optimum matching of the RF source to the diode; LO filtering to keep RF out of the LO and optimally match the LO to the diode, and IF filtering to optimally match the diode to its IF (and IF image) load, preferably while providing some rejection of the mixer's unwanted outputs, the strongest (and most potentially troublesome) of which is the LO signal.

1-2-1 Single-Balanced Mixer

Figure 8 shows the schematic of a two-diode, *single-balanced* mixer. It performs multiplicative mixing because its RF and LO signals are applied to different ports. In this more commonly seen two-diode mixer configuration, a balanced transformer drives the diodes out of phase for the LO and in phase for signals present at the RF port. The Figure 9 shows how this mixer's conversion gain and noise figure vary with applied LO power. Figure 10 shows how the mixer's conversion loss and noise figure vary with frequency for a constant LO power.

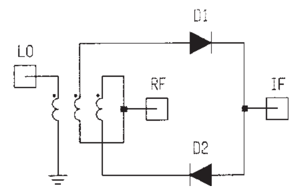


Figure 8 — Schematic of the two-diode (also known as *single-balanced*) mixer.

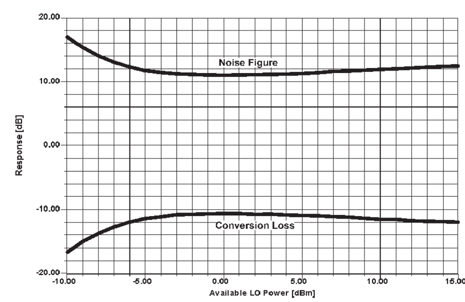


Figure 9 — How the nonideal mixer's conversion loss and noise figure vary with available LO power. In this analysis, LO =500 MHz (13 dBm), RF = 500.455 MHz (–20 dBm) and IF = 455 kHz.

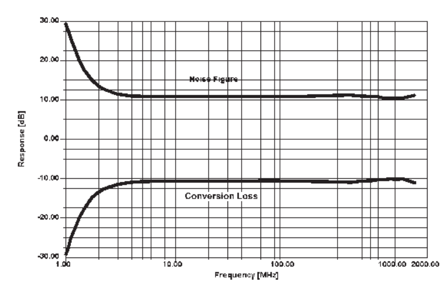


Figure 10 –How the nonideal two-diode mixer's conversion lossand noise figure vary with frequency for a constant LO power. In this analysis, LO = 1 to 1500 MHz (2 dBm), RF = 1.455 to 1500.455 MHz (–40 dBm) and IF = 455 kHz.

The two-diode mixer is used mostly in the frequency range above 1 GHz in a manner akin to a phase discriminator, using step-recovery diodes in the LO feed for enhanced harmonic mixing. Such mixers are mainly used in medium-cost spectrum analyzers or microwave receivers up to several tens of gigahertz, with the necessary transformers and baluns printed on the circuit board.

With perfectly matched diodes and perfect transformer and constructional symmetry, no LO energy arrives at the IF and RF ports, and there is only slight attenuation between the RF and IF ports. Both building and computer modeling such a mixer is impossible: building, because perfectly matched diodes and perfect transformer and constructional symmetry cannot be achieved in practice; computer modeling, because floating-point mathematics runs out of gas in handling the infinite amplitude spread involved in calculating the perfect cancellation of the LO signal as it travels to the RF and IF ports. That said, Figure 11 compares the mixer's *quasi*-ideal port-to-port isolation (matched diodes, a perfect transformer, no stray inductances and capacitances, and a 10-megohm resistors connected from port to port) and nonideal port-to-port isolation (slightly mismatched diodes and 0.5 pF between the upper terminal of the middle winding and ground).

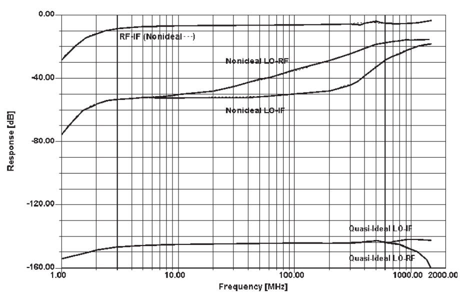


Figure 11 — Port-to-port isolation of the quasi-ideal (identical diodes and no stray capacitance) and nonideal (slightly mismatched diodes and 0.5 pF of stray capacitance between the upper terminal of the middle transformer winding and ground) two-diode mixer. In this analysis, LO = 1 to 1500 MHz (2 dBm), RF = 1.455 to 1500.455 MHz (–40 dBm) and IF = 455 kHz.

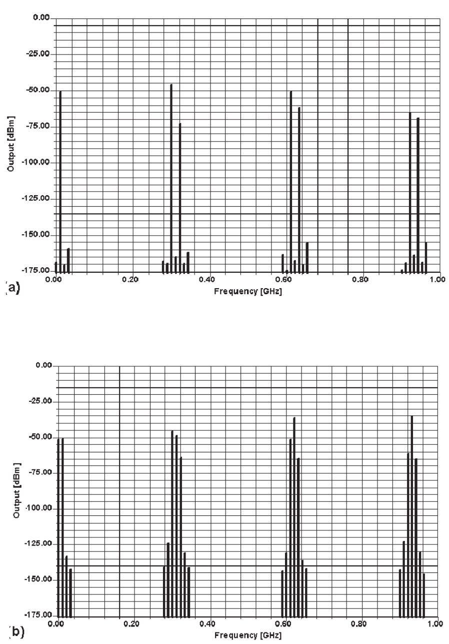


Figure 12 — The nonideal mixer's output spectrum with (a) identical diodes and no stray capacitance and (b) slightly mismatched diodes and 0.5 pF of stray capacitance between the upper terminal of the middle transformer winding and ground. In these analyses, LO = 310.7 MHz (2 dBm), RF = 300.0 MHz (–40 dBm) and IF = 10.7 MHz. Four LO harmonics and 3 LO sidebands were used.

Figures 12a and 12b show the mixer's output spectrum for the quasi-ideal and nonideal cases, respectively.

1-2-1 Subharmonically pumped single-balanced mixer.

Figure 13 shows a single-balanced mixer with a difference: antiparallel diode pairs take the place of single diodes, the RF and IF are buffered from each other only by filtering, and the LO is applied at 1/2 the frequency necessary to provide the desired frequency conversion. The RF-to-IF isolation is limited to that provided by the seriesed input and output filtering, but the LO-to-IF isolation is higher at f_{LO}, and much higher at $2f_{LO}$, than that achievable with a DBM with the LO signal at $2f_{LO}$ (Figure 14). Although the example shown is for an up-converting HF receiver, this technique finds application well into the microwave range as the basis for I/Q modulators, in which carrier leakage must be reduced to a level difficult to achieve with conventional DBMs.

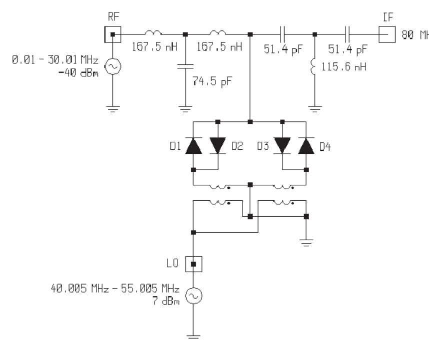


Figure 13 — A subharmonically pumped single-balanced mixer using antiparallel diode pairs. The LO operates from 40.005 - 55.005 MHz to mix 0.01 - 30.01-MHz RF to an IF of 80 MHz.

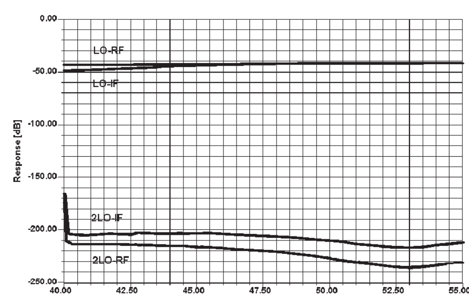


Figure 14 —Simulated interport isolation of the subharmonic SBM. For realism, the diodes and transformers are slightly mismatched.

2-1-1 Diode-Ring Double Balanced Mixer

Adding two more diodes and another transformer to the singly balanced mixer results in a double-balanced mixer (DBM) as shown in Figure 15. A DBM's frequency response is largely determined by the frequency response of its transformers. The low-frequency limit is determined by the inductance of the transformer windings, the reactance of which, at the lowest frequency of interest, should be at least four times the impedance at which the transformer operates. The upper frequency limit is determined mainly by the degradation of the transformers' transmission-line behavior at higher frequencies, although the increasing importance of diode capacitance also plays a role.

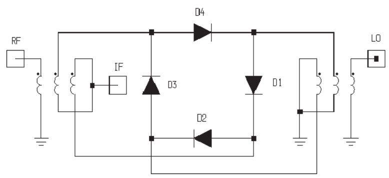


Figure 15 —Schematic of the diode-ring doubly balanced mixer.

A DBM's interport isolation is determined by the symmetry of its transformers, diodes and physical construction. In practice, the effects of diode mismatch can by minimized by using a diode quad ring crossover with closely matched diode.

Figure 16 shows how the DBM's conversion loss and noise figure vary with applied LO power. Figure 17 shows how the DBM's conversion loss and noise figure vary with frequency for an LO power of 7 dBm, with quasi-ideal and nonideal balance. Figure 18 shows how the DBM's port-to-port isolation differs with quasi-ideal and nonideal balance for an LO power of 7 dBm. Figure 19 shows how the DBM's RF- and LO-port return loss varies with frequency; the sharp peak corresponds to a resonance caused by one of the stray capacitances added to simulate less-than-ideal balance in the modeled mixer.

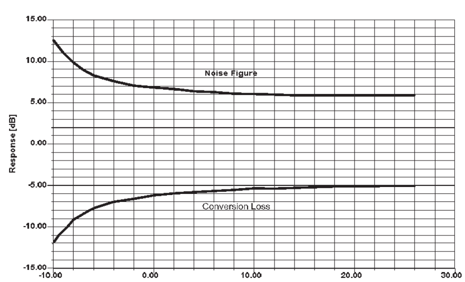


Figure 16 — DBM conversion gain and noise figure versus LO power. In this analysis, LO = 310.7 MHz (–10 to 26 dBm), RF = 300 MHz (–40 dBm) and IF = 10.7 MHz.

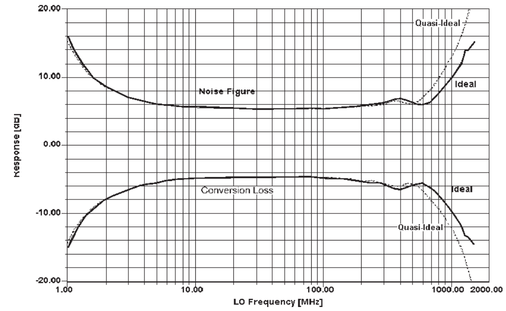


Figure 17 — This plot of conversion gain and noise figure versus frequency for quasi-ideally and nonideally balancedversions of the same DBM reveals that balance plays arelatively minor role in the CG and NF performance achieved.In these analyses, the LO (–7 dBm) sweeps from 1 MHz to 1500 MHz and the RF (–40 dBm) sweeps from 1.455 to 1500.455 MHz to produce an IF of 455

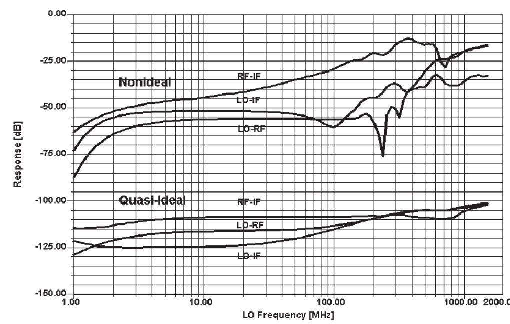


Figure 18 — Interport isolation for DBMs with quasi-ideal and nonideal balance. In these analyses, the LO (–7 dBm) sweepsfrom 1 MHz to 1500 MHz and the RF (–40 dBm) sweeps from 1.455 to 1500.455 MHz to produce an IF of 455 kHz.

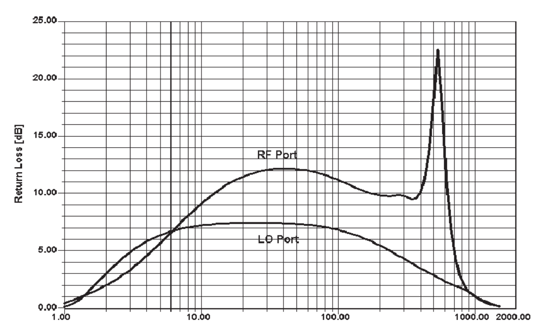


Figure 19 — Return loss versus frequency for the DBM's RF and LO ports. The sharp peak results from a stray resonance.

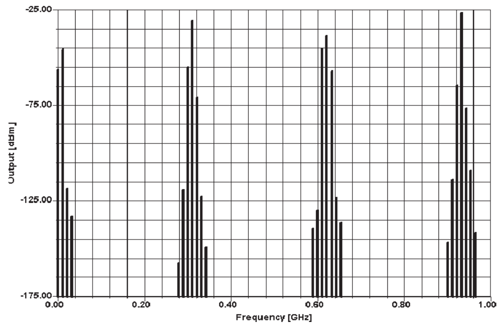


Figure 20 — Output spectrum of a nonideally balanced DBM. In this analysis, LO = 310.7 MHz (–7 dBm) and RF = 300 MHz (–40 dBm) for an IF of 10.7 MHz. Four LO harmonics and 3 LO sidebands were used.

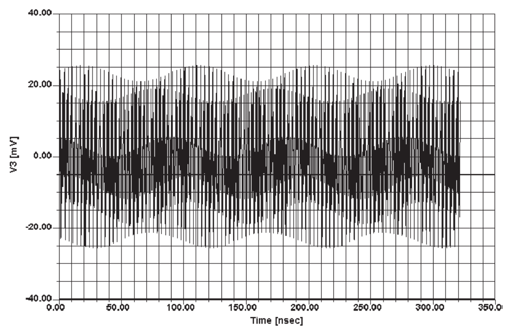


Figure 21 — IF-port voltage waveform of the DBM over 100 cycles of the LO signal. The 310.7-MHz LO and 10.7-MHz IF components are clearly evident.

Figure 20 shows the DBM's output spectrum. Figure 21 shows the DBM's output waveform over 50 cycles of the LO signal.

Two-tone testing the DBM allows us to characterize its IP_3 figures of merit. Figure 22 shows the nonideal DBM's IF and IM_3 responses for LO powers of –5, 1, 7 and 13 dBm. Figure 23 details how the DBM's IP_3 increases with LO drive, and Figure 24 shows the desired IF outputs and close third-order spurs near 10.7 MHz. Figure 25 shows the DBM's output voltage over 100 cycles of the LO signal, and Figure 26 shows the anode-cathode voltage of one of the ring's diodes, also over 100 LO cycles, both under two-tone IMD test conditions.

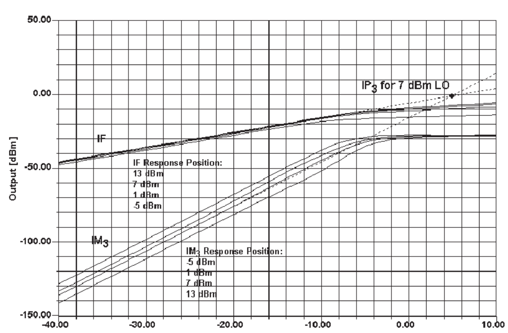


Figure 22 — Diode DBM IF and IM_3 output versus RF power for four LO-drive levels. The responses for LO = 7 dBm have been extrapolated to show IP_3. Within limits, varying a mixer's LO drive affects its linear IF output relatively little while significantly affecting IMD. See Figure 23. For all four analyses, LO = 310.7 MHz, RF1 = 300.0 MHz (–40 to 10 dBm), and RF2 = 300.3 MHz (–40 to 10 dBm); 4 LO harmonics and 3 LO sidebands were used.

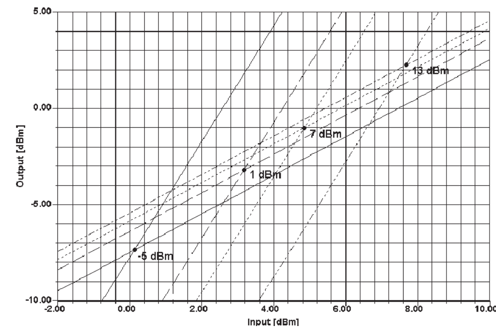


Figure 23 — Extrapolating the responses for all four LO levels represented in Figure 22 shows how varying a diode DBM's LO drive shifts its third order intercept point. Although these curves indicate that the simulated mixer's IP_3 generally increases with LO drive, the improvement in IP_3 is not as great as we might expect. The reason for this is that these four analyses, as well as the other diode-mixer analyses in this chapter, were done using diode models with a threshold voltage (V_J) of 0.23. If high-level diodes with a V_J of about 0.8 V had been used, $IP_{3,out}$ for the 13 dBm LO case shown here would increase to +13 dBm. $IP_{3,in}$ for the 13 dBm LO case would turn out to be 13 dBm + insertion loss = 13 dBm + 7 dB = 20 dBm. The issue of diode damage aside, attempting to increase IP_3 merely by driving a low- or medium-barrier diode harder eventually results in diminishing returns. High-barrier diodes are essential in getting the best IP_3 performance with high LO drive.

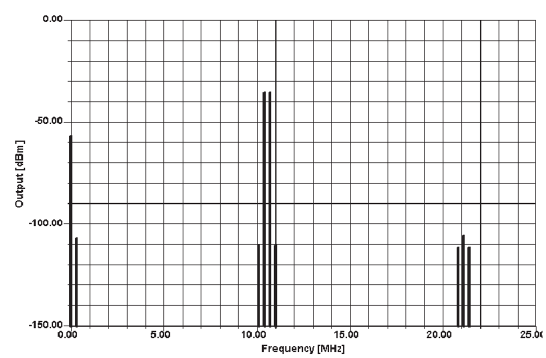


Figure 24 — The DBM's output in the 11-MHz region during two-tone testing. The third-order products are clearly visible above and below the desired output signals. The test conditions for this analysis are those for Figure 22 with LO = 13 dBm.

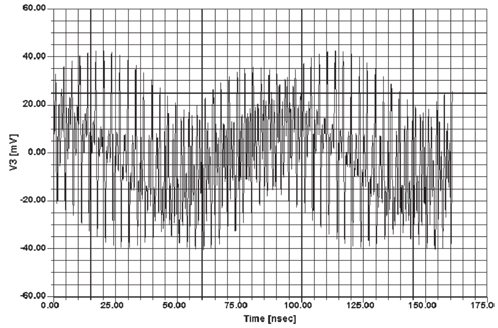


Figure 25 — The DBM's IF-output voltage over 50 cycles of the LO signal. The test conditions for this analysis are those for Figure 28 with LO = 13 dBm.

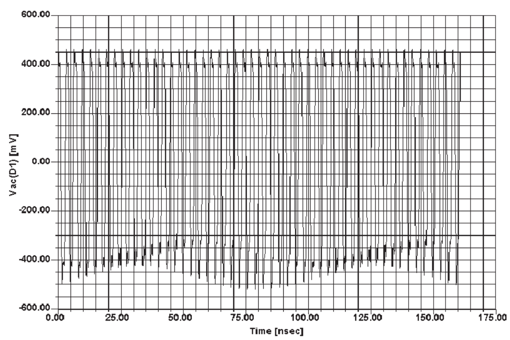


Figure 26 — The anode-cathode voltage of one of the DBM's diodes, also over 50 LO cycles under two-tone IMD test condtions. The test conditions for this analysis are those for Figure 4-28 with LO = 13 dBm.

A double-balanced mixer, unless it is termination insensitive, is extremely sensitive to nonresistive termination. This is because the transmission-line transformers do not operate properly when they are not properly terminated, and the reflected power generates high voltage across the diodes. This effect results in much higher distortion levels than in a properly terminated transformer.

2-2 Applications of mixers

2-2-1 Phase detector. Theoretically, any mixer with a dc-coupled IF port can be used as a phase detector. When two signals of equal frequency are applied simultaneously to the reference and incoming signal ports, the phase detector produces a dc output at the IF port proportional to the cosine of the phase difference (Figure 28).

2-2-2 Binary phase-shift keying (BPSK) modulator. Binary phase modulation occurs when a positive and negative signal current shifts the RF carrier between 0 and 180°. Figure 29 shows a double-balanced mixer operating as a BPSK modulator.

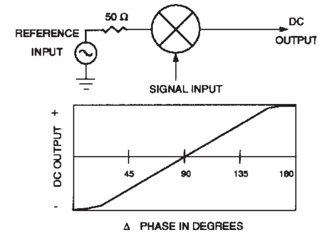


Figure 28 — A mixer with a dc-coupled IF port can be used as a phase detector.

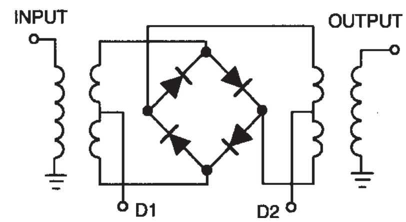


Figure 29 — A diode-ring mixer as phase modulator.

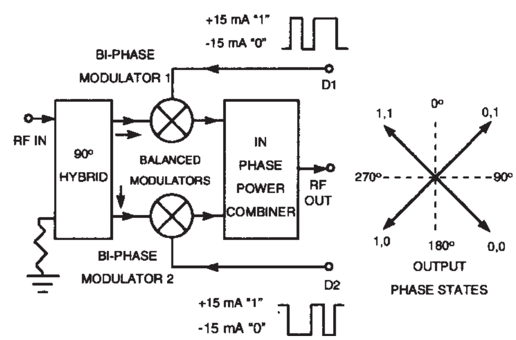


Figure 30 — Two biphase modulators form the basis for a QPSK modulator.

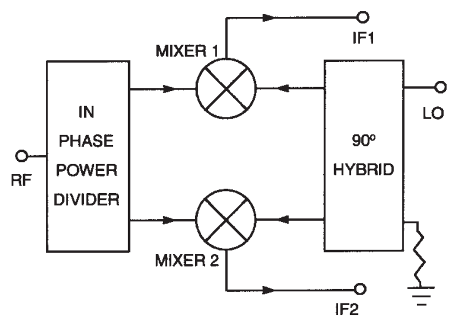


Figure 31 — Quadrature mixer.

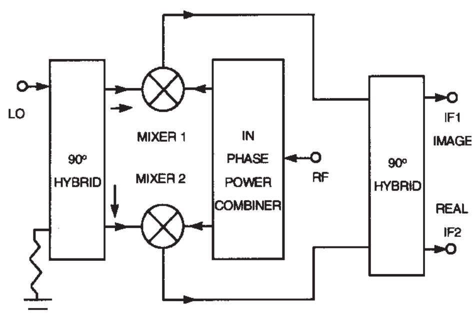


Figure 32 — An image-reject mixer uses phasing to differentiate between its LO+IF and LO–RF IF outputs.

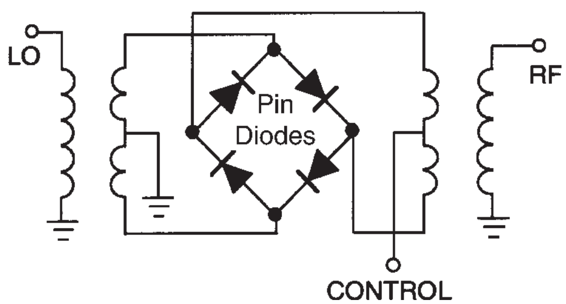


Figure 33 — A diode DBM can be used as a dc-controlled attenuator if PIN diodes are used instead of Schottky devices in its ring.

2-2-3 Quadrature phase-shift keying (QPSK) modulator. A typical QPSK Modulator consists of two biphase modulators, a 90° divider, and a 0° power combiner as shown. Data inputs at the control ports will cause the carrier to shift between 0, 90, 180 and 270° as shown in Figure 30.

2-2-4 Quadrature IF mixer. A quadrature IF Mixer produces two IF outputs in phase quadrature. Its basic structure consists of two double balanced mixers, a 90° splitter and 0° splitter. The basic block diagram is shown in Figure 31.

2-2-5 Image-reject mixer. The image-reject mixer consists of a basic quadrature IF mixer with an additional 90° hybrid at the IF ports as shown in Figure 32. The primary function is to differentiate between the real signal and the image signal. This type of device is especially useful in applications where the desired RF signal and image are so close in frequency that rejecting the image with filtering is not practical.

2-2-6 Diode attenuator/switch. A ring of PIN diodes can be used as electronic attenuators by applying variable forward bias to the diodes (Figure 33). Maximum attenuation is achieved when the current at the control port is zero. The maximum attenuation is the isolation between the input and output port. Minimum attenuation (insertion loss) is achieved when the IF port current is 20 mA.

2-2-7 Single-sideband (SSB) or in-phase/quadrature (I/Q) modulator. SSB or I/Q modulators are useful in discriminating and removing the lower sideband (LSB) or upper sideband (USB) generated during frequency conversion, especially when the sidebands are very close in frequency and attenuation of one of the sidebands cannot be achieved with filtering. This is the case with audio and video modulation, where signals from dc to 10 MHz must be converted to a higher frequency that is appropriate for transmission. In such cases, both sidebands will be very close in frequency to the carrier frequency. With an I/Q modulator, one of the sidebands is easily cancelled or attenuated along with its carrier.

Attenuation of the carrier has been the most troublesome aspect in the design of passive I/Q modulators. Isolation between the local-oscillator (LO) port and the RF port of the mixers, which is the main parameter in determining carrier rejection, is usually insufficient at frequencies above 200 MHz.

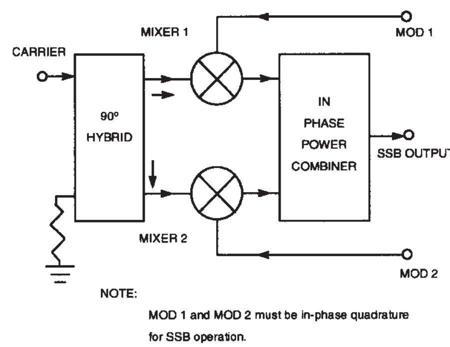


Figure 34 — An SSB modulator matches two high-frequency mixers, a 90° hybrid, and an in-phase power combiner to produce an SSB output signal.

I/Q modulator designs have basically comprised two double-balanced mixers (Figure 34). The mixers are fed at the LO ports by a carrier phase-shifted through a 90°. hybrid. Thus, the carrier signal's relative phase is 0° to one mixer and 90° to the other mixer. Modulation signals are fed externally in phase quadrature to the two mixers' IF ports. The mixers' modulated output signals are combined through a two-way, in-phase power divider/combiner.

The circuit forms a phase-cancellation network to one of the sidebands and a phase-addition network to the other sideband. The carrier is somewhat attenuated and is directly dependent on the inherent LO-to-RF isolation of the mixers and the modulating signal level. In industry-standard I/Q modulators, USB suppression results when the first modulation port (MOD 1) is fed with a signal that is 90° in advance of the signal feeding the second modulation port (MOD 2). Opposite phasing can be arranged by changing the internal phase polarity of the mixers or by interchanging the 90° hybrid output ports to the LO ports of the mixers.

The phase and amplitude imbalances between the various components used in the manufacturing of the I/Q modulators must be tightly maintained for optimum SSB rejection. Matching of the two mixers for conversion loss and insertion phase is extremely critical, since differences in these parameters will add to amplitude- and phase-imbalance errors. The 90° hybrid in the LO port must be in nearly perfect phase quadrature.

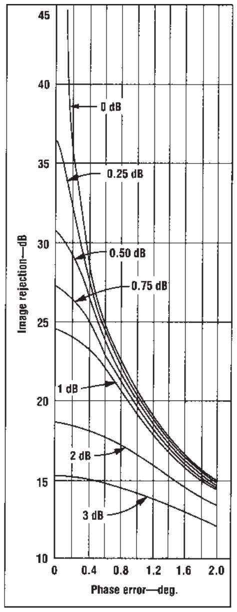


Figure 35 — The level of SSB rejection improves as the phase and amplitude imbalance performance of an SSB modulator improves.

Phase- and amplitude-imbalance errors adversely affect sideband suppression (Figure 35). In most cases, a typical passive I/Q modulator operates with a carrier input level of +10 dBm, which is required to drive the diodes in the mixers to operate in the linear range. The dynamic range of these mixers can be significantly improved by using diodes with a higher barrier height. The LO signal in this case must be increased in order to drive these diodes into conduction in their linear range. Carrier rejection is also a problem when designing an SSB modulator, since only a few decibels of suppression can be achieved in standard high-frequency models. The major contributor to carrier suppression is the inherent LO-to-RF isolation through the mixers. Unfortunately, this isolation is usually poor at cellular frequencies (800 to 1000 MHz), where at least 25 dB of carrier rejection is necessary. In some cases, designers feed a small amount of dc into the IF ports to control the carrier rejection, but this complicates the driver circuitry and calls for temperature compensation when operating at different temperatures. As an example, an SSB modulator is assumed to operate with +10-dBm LO drive with each modulating signal at –10 dBm and in phase quadrature to each other when applied to the modulating ports (MOD 1 and MOD 2). The result will be a modulated signal at –16 dBm, assuming 6-dB conversion loss. For 20-dB carrier rejection with respect to the desired modulated signal, the carrier must be at –36 dBm, which translates to LO-to-RF isolation of 46 dB.

For 20-dB carrier rejection with respect to the desired modulated signal, the carrier leakage must be at –36 dBm, which translates to LO-to-RF isolation of 46 dB.

By employing a subharmonic approach, the performance of SSB modulators can be extended beyond the limits of conventional designs. The approach is based on the use of subharmonic mixers in place of Double Balanced-frequency mixers and is applicable from about 140 to 3000 MHz. Matched antiparallel diode pairs used in single-ended or single-balanced mixer configurations cancel even-order intermodulation products (such as $2f_{LO} \times 2f_{RF}$, $3f_{LO} \times 3f_{RF}$, etc.) at all ports.

Single-ended mixers lack the port-to-port isolation needed for SSB modulator applications. Odd-order products of the RF and LO frequencies (even $f_{LO} \times$ odd f_{RF}) and (odd $f_{LO} \times$ even f_{RF}) appear on all ports, requiring extensive filtering for satisfactory performance. For a single-balanced mixer, even harmonics of the LO combining with odd harmonics of the RF appear at the IF port, whereas odd harmonics of the LO combining with even harmonics of the RF appear at the RF and IF ports. This assumes that a balanced transformer is placed at the LO port, which is a logical choice due to the fact that the highest level signal appears at the LO port. Since the desired odd-order IF products appear at both the RF and IF ports, a need arises for a diplexing network to isolate the RF and IF signals.

The subharmonic modulator design provides a unique way to isolate the RF and IF signals. A single-balanced harmonic mixer offers good LO-to-RF and LO-to-IF isolation but poor RF-to-IF isolation. Fortunately, harmonically related signals are spaced well apart in the frequency spectrum, simplifying filtering of harmonically related signals.

Harmonic mixing also works well with low LO power levels, with somewhat lower 1-dB compression on the RF port than with fundamental-frequency mixing. The ability to operate with LO frequencies that are a fraction of the carrier frequency (1/2, 1/4, 1/6, etc.) significantly reduces the cost of an LO source, especially at higher frequencies. Also, using lower-frequency LO sources helps avoid the signal-leakage problems inherent with higher-frequency LO sources. Minimizing signal leakage, especially at higher frequencies, becomes expensive and bulky. Subharmonic mixing offers several advantages:

- The technique offers the ability to operate at LO frequencies that are 1/2, 1/4, or 1/6 of the carrier frequency. For example, for an IF of 100 MHz at an RF of 2 GHz, the LO can be (2000±100) ÷ 2 = 950 or 1050 MHz.
- The LO's even harmonics are strongly attenuated.
- The filtering requirements for fundamental frequency and odd harmonic signals of the LO are not critical.
- The cost of generating the LO is reduced due to the fact that the LO frequency need only be a fraction of the carrier frequency.

As an example of the performance improvements possible with the subharmonic mixers, units were evaluated at both cellular (935- to 960-MHz) and PCN/PCS (1.8- to 1.9-GHz) bands. For a conventional SSB modulator at 1.9 GHz fed with +10 dBm modulation signals, carrier rejection is barely 10 dB (36).

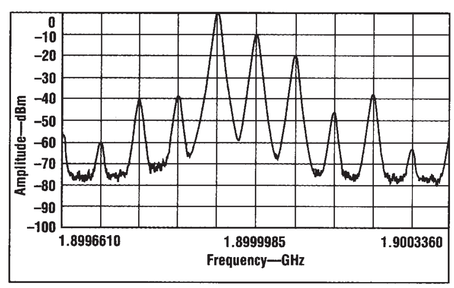
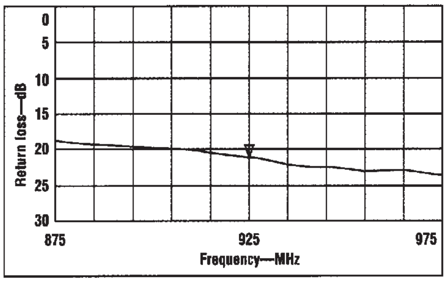


Figure 36 — This plot of carrier and sideband rejection was measured for a conventional SSB modulator operating at 1.9 GHz.



Figures 37 — The SSB modulator's return loss as measured at the local-oscillator (LO) port

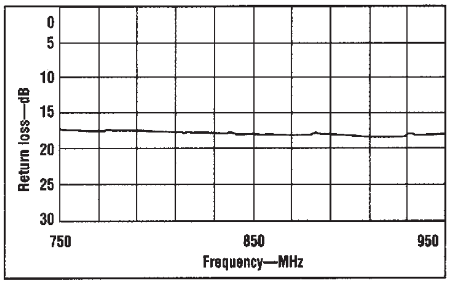


Figure 38 — The novel harmonic SSB modulator's return loss as measured at the RF port.

Sideband rejection can be improved by tuning, but the carrier rejection is controlled by the LO-to-RF isolation of the double-balanced mixers. Conventional double-balanced mixers with high isolation at cellular and PCN bands are very expensive and large when special techniques are used to improve LO to-RF isolation. In contrast, the subharmonic nature of the new approach allows the use of lower-frequency, less-expensive components in the modulators' construction.

The subharmonic modulators offer an improvement of more than 15 dB in carrier suppression compared to the conventional approach.

The measured return loss at the LO and RF ports is better than 18 dB (Figures 37 and 38). Measurements made on a cellular-band SSB modulator reveal carrier rejection on the order of 40 dB. Typical insertion loss is 7 dB while sideband rejection is 30 dB (Figure 39).

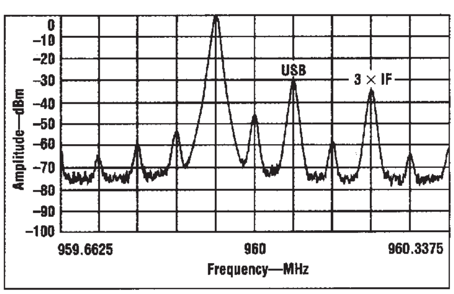


Figure 39 — This plot of carrier and sideband rejection as measured for the novel harmonic SSB modulator operating at cellular frequencies.

By the virtue of harmonic mixing, even-order mixing products are attenuated by about 30 dB with respect to the desired modulated output signal. The fundamental-frequency feedthrough into the output port is approximately 5 dB lower than the desired modulated signal, whereas the fourth harmonic mixing with the modulating signal is approximately 10 dB lower. Typical loss for fourth-harmonic mixing is 17 to 19 dB while maintaining 30 dB of carrier rejection.

Since harmonically-related products are well-spaced in frequency, filtering undesired signals is relatively inexpensive using standard octave-bandwidth filters. Low-cost commercial bandpass filters typically offer better than 40 to 50 dB attenuation of unwanted harmonic signals. Constant-impedance bandpass filters offering good impedance match at desired stopbands can also be used in cases where harmonically related products require impedance termination within a system.

The subharmonic modulator design is easily applied at custom frequencies. Conversion of an SSB modulator with output frequency corresponding to twice the LO frequency to one with output corresponding to four times the LO frequency requires only one component change, in the form of a signal-combining network at the modulator's output. Although the conversion loss of the fourth-harmonic LO component mixing with the modulating signal is in the vicinity of 18 dB, the cost of generating the LO is drastically reduced with the subharmonic modulator. In spite of higher signal loss, the carrier rejection is still at least 30 dB at the fourth harmonic, and harmonically related products can be eliminated with an inexpensive filter.

***2-2-8 Triple-balanced mixer*.** Two diode rings can be combined to form a double double-balanced mixer, or triple-balanced mixer (TBM), as shown in Figure 40. Triple-balanced mixers achieve greater higher dynamic range and interport isolation than double-balanced designs especially above 1 GHz at the expense of LO power and increased complexity and size. The conversion loss of Triple Balanced mixers increases by approximately 6 dB as the IF frequency is lowered below the sampling LO/RF frequency towards DC.

Figure 41 shows the circuit's interport isolation with the circuit configured in a less than ideally balanced form, with small variations in transformer-winding inductance and diode parameters introduced for more realistic modeling. Note that the mixer's interport isolation generally less sloped with frequency, when compared with the DBM (Figure 42).

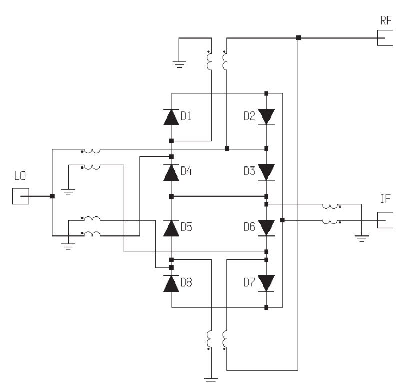


Figure 40—A triple-balanced diode mixer. A limitation of this configuration is that the internal dc common connections associated with its RF and LO transformers disallow usable IF response down to dc.

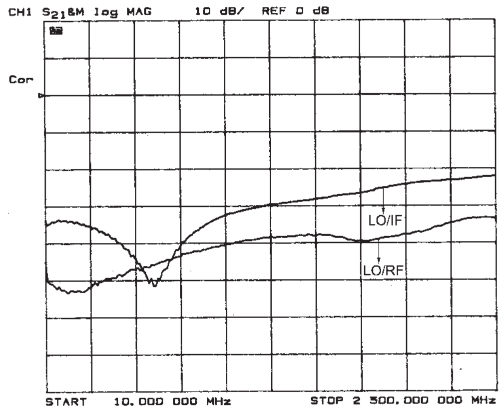


Figure 41—The triple-balanced mixer offers improved high-frequency isolation over a standard DBM.

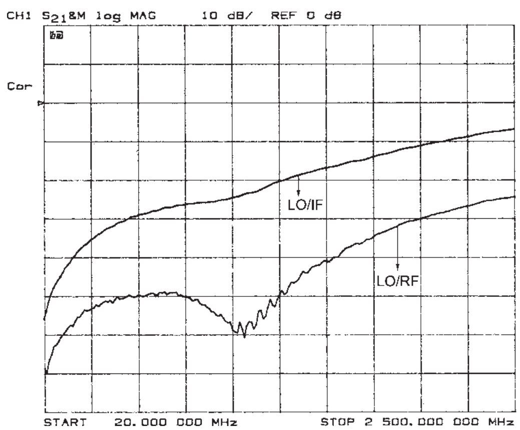


Figure 42—The Double-balanced mixer isolation for a DBM.