



Anaren®

Cased/Caseless Components

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Application notes: Cased/caseless

**Mixers &
Modulators**



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General

Mixers are used to convert a signal from one frequency to another. This is done by combining the original RF signal with a local oscillator signal (LO) in a non-linear device such as a Schottky-barrier diode.

The diode output spectrum includes:

- the original inputs, LO and RF
- all higher order harmonics mLO and nRF (where m, n are integers)
- the two primary sidebands, LO ± RF (m, n = 1)
- all higher order products of mLO ± nRF (where m, n are integers)
- a dc output level

The desired output frequency, commonly called the intermediate frequency (IF), can be either the lower (LO-RF) or upper (LO + RF) sideband. When a mixer is used as a down-converter, the lower sideband is the sideband of interest. (Unless specifically noted, we will use the term “mixer” synonymously with down-convertors.)

A microwave *balanced* mixer makes use of the 3 dB hybrid to divide and recombine the RF and LO inputs to two mixing diodes. The 3 dB hybrid can be either the 90° or 180° type. Each has certain advantages that will be covered later. The critical requirement is that the LO and RF signals be distributed uniformly (balanced) to each mixer diode.

Figure 1 is a typical balanced mixer block diagram. The mixer diodes are reversed relative to each other; the desired frequency (IF) components of each diode are then in-phase while the dc outputs are positive and negative, respectively.

The two diode outputs are summed in a tee where the dc terms cancel and only the desired IF component exits at the IF port.

Mixer Types Compared

While Anaren’s 3 dB, 90° and 180° type mixers are in widespread use and quite effective, Anaren also supplies two other basic mixer types. These are the double-balanced mixer and the Ortho-Quad® (quadrature fed dual) mixer. None of the four mixer types is optimum for all parameters. The relative advantages and disadvantages of each type are summarized in Table 1. Careful use of the table will allow the selection of the optimum mixer type for a particular application. A more detailed discussion of the mixer parameters follows the table.

Figure 1 - Simplified Block Diagram of 3 dB Hybrid Balanced Mixer

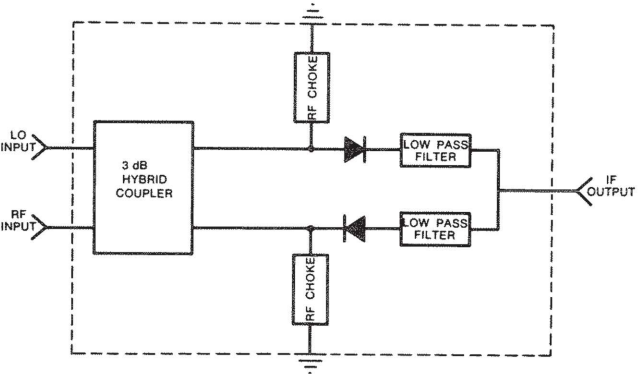


Table 1 - Anaren’s 3 Basic Mixer Types — Performance Comparison

Mixer Type	Typical* Anaren Model No.	No. of Diodes ⁽¹⁾	VSWR ⁽²⁾	Conversion Loss ⁽³⁾	LO/RF Isolation ⁽⁴⁾	Bias	Spurious Rejection ⁽⁵⁾ m-n = 1	Harmonic Suppression ⁽⁶⁾	Intercept Point ⁽⁷⁾ (dBm)	Dynamic Range	IF Bandwidth	Image Focusing Port ⁽¹⁰⁾
90° Hybrid	70116 7H0116	2 2	Good Good	Lowest Lowest	Poor Poor	No Yes	Poor Fair	Poor Fair	+15 +30 ⁽⁸⁾	High Highest	Wide Wide	LO LO
180° Hybrid**	7B0055 70580	2 2	Poor Poor	Low Low	Good Good	Yes Yes	Fair Fair	Good Good	+15 +30 ⁽⁸⁾	High Highest	Wide Wide	RF RF
Double-Balanced	73125	4	Poor	Low	Very Good	No	Good	Very Good	+18	High	Extremely Wide ⁽⁹⁾	RF
Ortho Quad®	74125	4	Poor	Low	Excellent	No	Good	Very Good	+18	High	Extremely Wide ⁽⁹⁾	RF
	7A0126	4	Good	Low	Very Good	No	Good	Fair	+18	High	Wide	Internal Load

NOTES:

- (1) 2-diode types require +7 dBm LO power for best performance; 4-diode types require +10 dBm LO power.
(2) VSWR: Poor: 2.5:1 typ
Good: 1.3:1 typ
(3) Conversion Loss: Lowest: 5 – 7 dB typ
Low 7 – 9 dB typ
(4) LO/RF Isolation: Poor: 10 dB typ
Good: 20 dB typ
Very Good: 25-30 dB typ
Best: 35-40 dB typ
(5) mxn Spurious Rejection: where |m – n| = 1: i.e., 1x2, 2x1, 2x3, 3x2, etc.
Poor: partial rejection of most |m – n| = 1 spurs
Fair: partial rejection of most |m – n| = 1 spurs
(bias adjust will suppress some spurs even further)
Good: potentially rejects all |m – n| = 1 spurs

- (6) Harmonic Suppression: Poor: partial rejection of LO/RF even harmonics
Fair: partial rejection of LO/RF even harmonics (bias adjust will suppress some harmonics even further)
Good: can reject all LO even harmonics
Very Good: can reject all LO and RF even harmonics
(7) Intercept Point: Typical third order intercept point is 6 to 9 dB above the LO power.
(8) This intercept point can only be achieved by using the optimum load line biasing technique (see page 000) and increasing the LO power to approximately +23 dBm.
(9) The IF bandwidth for the 73000 series of balanced mixers overlaps the RF range of the units. Bandwidths from DC to 4 GHz are available.
(10) Image Focusing: defines where image signal energy is focused, LO port, RF port, or to internal termination. Where this image signal is focused can affect mixer performance in phase and amplitude tracking systems. (See page 000 for additional information.)

*Typical models shown are L-band or S-band units.

**Due to improved performance to be found in Anaren’s double-balanced mixer series 73120 and 74120, 180° balanced mixers are no longer available as a separate line. However, they are discussed in this application section as part of the evaluation of the four basic mixer types.

VSWR

The VSWR specifies the degree of impedance match between the mixer LO and RF ports and the 50 ohm system in which it is used.

One of the most desirable characteristics of the 90° hybrid mixer is its excellent VSWR over the full performance range of the hybrid. When a signal is fed into either input, any reflections from the similar mixer diodes will combine at the other input, providing a low VSWR at either port. The input reflection coefficient for a 90° hybrid mixer having diode reflection coefficients, p_1 , p_2 is given by:

$$\rho_{\text{ (mixer, 90°) }} = \frac{p_1 - p_2}{2}$$

If $p_1 = p_2$ (if the diodes are balanced), whatever their value, the input remains matched.

If the mixer uses a 180° hybrid, any reflections from the similar mixer diodes will be focused back to the input port. Unless the diodes look like a very good match, the input VSWR will be poor. The input reflection coefficient for a 180° hybrid mixer having diode reflection coefficients p_1 , p_2 is given by:

$$\rho_{\text{ (mixer, 180°) }} = \frac{p_1 + p_2}{2}$$

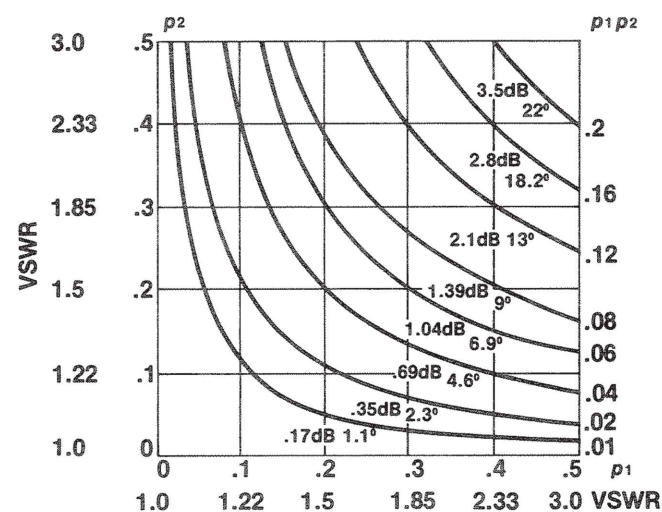
Diode impedance is a function of LO power; therefore VSWR will be affected by LO drive level for this mixer. If the diode impedances are equal, the input VSWR equals the return loss for one diode.

An Ortho-Quad® mixer has the same good VSWR characteristics as the 90° hybrid mixer since it uses 90° hybrids as the coupling mechanism to the diodes.

A double-balanced mixer exhibits poor VSWR characteristics. It uses 180° hybrids as the coupling networks to its diodes and therefore depends on good diode match to provide good input VSWR. In addition, the LO VSWR is frequently degraded even further by the necessity for extracting the IF signal from the LO coupling network. VSWR's of 3:1 or greater are not uncommon for double-balanced mixers.

Low VSWR can be especially important if the mixers are to be used in phase and amplitude tracking receiver systems.⁽¹⁾ The biggest cause of amplitude and phase ripple in a tracking receiver system is the error paths set up by the mismatch of one component beating with the mismatch of another component. Figure 2 shows the phase and amplitude ripple possible from two mismatches on a 50 ohm line. Consider, for example, the errors possible if an antenna and a mixer have a mismatch. If a VSWR of 2.33:1 ($p = 0.4$) is assumed for both components, Figure 2 shows it is possible to have peak-to-peak phase errors of 18.2 degrees and peak-to-peak amplitude errors of 2.8 dB. When it is realized that most microwave *double-balanced* mixers have a VSWR of 3:1 or greater, the typical 1.4:1 VSWR advantage of the Anaren 90° hybrid or Ortho-Quad® mixers becomes apparent.

Figure 2- Amplitude and Phase Ripple Due to Mismatch Errors.



LO/RF Leakage

The LO/RF isolation is a measure of the LO leakage at the RF port of the mixer.

The 90° hybrid mixer has poor isolation characteristics because LO/RF isolation is dependent upon diode match. Any LO energy not absorbed by the diodes is directed to the RF port. The LO/RF isolation for a 90° hybrid mixer having diode reflection coefficients p_1 , p_2 is given by:

$$\text{Isolation (dB)} = 20 \log_{10} \left(\frac{2}{p_1 + p_2} \right)$$

If the diode impedances are equal, the isolation equals the return loss for one diode. Since diode impedance is a function of LO power, the isolation will be sensitive to LO drive.

A 180° hybrid mixer has good isolation characteristics; dependent only on the diode balance regardless of their impedance values. The LO/RF isolation for a 180° hybrid mixer having diode reflection coefficients p_1 , p_2 is given by:

$$\text{Isolation (dB)} = 20 \log_{10} \left(\frac{2}{p_1 - p_2} \right)$$

LO power will not affect isolation if the diodes continue to track each other.

The Ortho-Quad® will have good LO/RF isolation regardless of LO power because a unique, patented phasing network directs any unused LO energy to an internal termination; it will never show up at the RF port as LO/RF leakage. The good LO/RF isolation is independent of LO power level. The Ortho-Quad® is the only one of the four basic type mixers that will provide, simultaneously, good VSWR and good isolation regardless of LO drive level.

The double-balanced mixer has the same good LO/RF isolation characteristics as the 180° hybrid balanced mixer because it uses 180° hybrids as the coupling networks to its diodes. As long as the diodes track each other, the LO/RF isolation will be good and will remain independent of LO power.

For many receiving applications the *amount* of LO power leaking out the RF port is important since it will ultimately be reradiated out of the antennas. A *biased* mixer can be very useful in such applications because LO power and resultant reradiation can be significantly lowered while still maintaining good conversion loss. In fact, using biased 90° balanced mixers may actually result in lower radiated energy from the antenna and better overall performance than when using many double-balanced mixers. Double-balanced mixers at first would appear to offer better results because of their higher LO/RF isolation specs, but the following comparison will serve to illustrate why this may not be the case:

BIASED 90° BALANCED MIXER (Model 7H0118)		TYPICAL DOUBLE-BALANCED MIXER (Same Frequency as Model 7H0118)	
Conversion Loss:	8.0 dB max (0 dBm LO) 8.5 dB max (-4 dBm LO)	Conversion Loss:	9.0 dB max.
LO Power:	0 dBm at 1.8 mA bias -4 dBm at 1 mA bias	LO Power:	+10 dBm (no bias option)
LO/RF Isolation:	10 dB typical (6 dB min)	LO/RF Isolation:	17 dB typical*
LO Leakage at RF Port:	-10 dBm (0 dBm LO) -14 dBm (-4 dBm LO)	LO Leakage at RF Port:	-7 dBm (+10 dBm LO)
LO, RF VSWR:	1.65 typ. 2.0 max.	LO, RF VSWR:	3.0 typ. 4.0 max.*

*Many double-balanced mixers have no published worst case spec.

The Anaren 7H0118 radiated 7 dB *less* LO power at the antenna terminal than the double-balanced mixer. This improved leakage performance is even more impressive when the rest of the specifications are examined. When compared to the typical double-balanced mixer, the Anaren biased, 90° hybrid mixer has:

1. Lower LO/RF leakage levels.
2. Equal or better conversion loss.
3. 10 to 14 dB *less* LO drive required.
4. Much lower VSWR.

Conversion Loss and Noise Figure

Conversion loss and noise figure are two of the most important and closely-related mixer parameters. The primary requirement for a mixer is to provide maximum IF output power with the minimum RF input power while generating the least amount of noise. Minimum losses in the RF, LO coupling networks, high quality Schottky-barrier (hot carrier) coupling networks to the diodes, broadband diode matching and optimum local oscillator drive, all contribute to the good conversion loss and low noise figure of Anaren's mixers.

Conversion loss of the mixer is a measure of the power at the IF frequency relative to the power at the RF input frequency. It is a function of the mixer alone.

Mixer noise figure is commonly stated at a particular IF frequency when measured in a system containing an IF amplifier following the mixer. This IF amplifier contributes to the measured noise figure. The mixer diodes also can contribute a small amount of noise energy to the noise figure. The measured noise figure (NF_0) in decibels is related to the mixer conversion loss [L_C], the IF amplifier noise figure (F_{IF}), and the mixer diode noise-temperature ratio (T) by:

$$NF_0 = L_C + 10 \log_{10} (F_{IF} + T - 1)$$

where L_C is mixer conversion loss in dB
 F_{IF} is the noise factor (a numeric ratio) of the IF amplifier
T is the effective diode noise-temperature ratio.

T can be as low as 0.85 for modern high-quality hot carrier diodes but a more typical value is 1.0. The relationship between measured noise figure and conversion loss then simplifies to

$$NF_0 = L_C + NF_{IF} \quad (2)$$

where NF_{IF} is IF preamplifier noise figure in dB.

When $T = 1$ the mixers' single-sideband noise figure is the same as its conversion loss. Equation (2) then shows that the overall noise figure is obtained by adding the noise figure of the IF preamplifier directly to the mixer conversion loss.

It is instructive to observe what happens to the overall noise figure when $T = 0.85$. By equation (1):

$$NF_0 = L_C + 10 \log_{10} (F_{IF} + 0.85 - 1)$$

When NF_{IF} is equal to 1.5 dB:

$$NF_0 = L_C + (1.413 - 0.15) = L_C + 1.02 \text{ dB}$$

The combined preamplifier/diode noise contribution is only 1.02 dB, 0.48 dB *less* than that produced by the preamplifier alone.

All noise figure measurements relating to Anaren mixers are based on single sideband operation using a 30 or 60 MHz IF preamplifier having a NF_{IF} of 1.5 dB.

Bias Option

In many instances the optimum LO power is not available for best conversion loss. The conversion loss of an unbiased mixer degrades rapidly as the LO power decreases. This is due chiefly to diode mismatch at the RF and IF frequencies. The diode mismatch can be minimized and the conversion loss dramatically improved by biasing the diode with a dc current.

Anaren provides biased 90° hybrid mixers and biased 180° hybrid mixers as standard catalog items. Figure 3 is a block diagram for a 90° biased mixer and Figure 4 shows how conversion loss and optimum bias current typically vary with changing LO power. These biased 90° mixers use a single bias terminal with no internal current limiting resistor.

Figure 3 - Simplified Block Diagram of 90° Hybrid Biased Mixer.

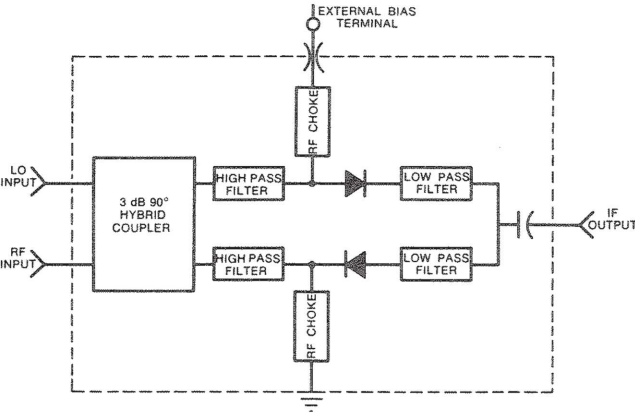
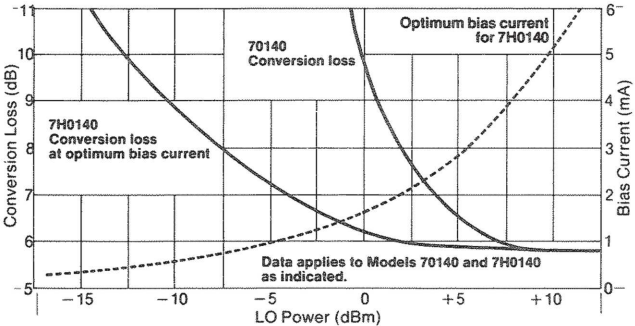


Figure 4 - Conversion Loss and Bias Current versus LO Power for 90° Hybrid Biased Mixer.



The 180° hybrid biased mixers use two bias terminals and most use internal current limiting resistors. Figure 5 is the block diagram for a 180° biased mixer. Figure 6 shows how conversion loss and optimum bias voltage typically vary as the LO power is changed. Note that, unlike the 90° biased mixer, bias voltage must be applied at all times for proper operation of the 180° biased mixer; even when normal LO power (+ 7 dBm) is applied. This is because the only dc return for the diode current is through the large (typically 3.3 K ohm) internal current-limiting resistors.

Figure 5 - Simplified Block Diagram of 180° Hybrid Biased Mixer

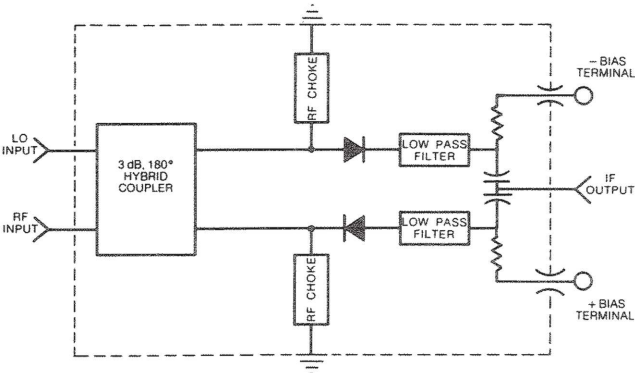
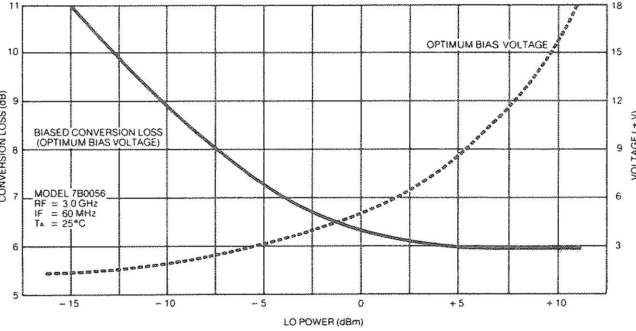


Figure 6 - Conversion Loss and Bias Current vs. LO Power for 180° Hybrid Biased Mixer



It is convenient to have current limiting resistors built into a mixer so that all one has to do is apply the correct bias voltages to the bias terminals. However, it turns out that having direct access to the diodes dc bias point allows a number of bias techniques to be used. One of these techniques, the optimum load line approach, allows the designer to optimize conversion loss and other mixer parameters over a wide LO dynamic range.

These bias techniques were first developed for use with Anaren's 90° hybrid balanced mixers and are available for use with the standard catalog models. Other mixer types can be supplied with similar bias options, on special order, but the discussion which follows will be directed toward the standard catalog line of 90° biased mixers.

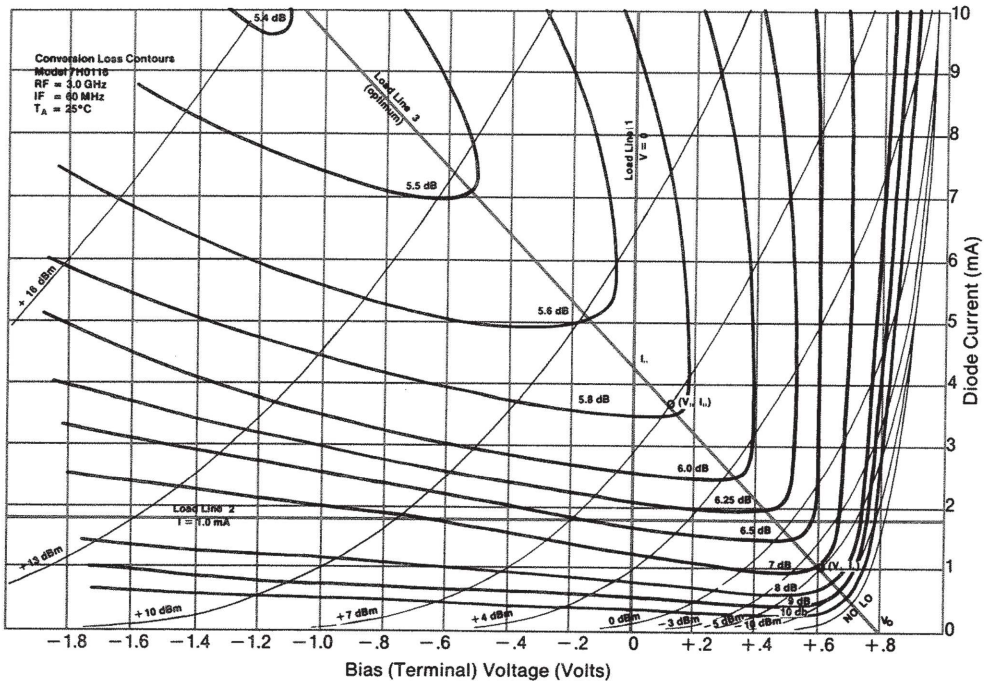
Bias Techniques

Conversion loss, VSWR and spurious response characteristics of a mixer depend upon the impedance presented to the diodes, not only at the RF and IF frequencies, but over the entire spectrum present in a Taylor's expansion of the diode current⁽²⁾. For biased mixer operation the dc impedance presented to the diode is most important. A family of I-V characteristics can be measured and plotted as a function of applied LO power with conversion loss, RF VSWR and spurious response contours then plotted over the I-V curves. Various load lines* can then be plotted revealing mixer performance under various bias and LO conditions.

Figure 7 shows the I-V characteristics and conversion loss contours for a typical mixer and illustrates the three types of load lines most useful for the majority of mixer applications. The first type is the unbiased diode (zero load resistance: voltage = constant = 0) which covers the majority of mixers in use today. The second type is the constant current load line (I = constant) which has been commonly used where insufficient LO power is available. The third case is a new approach⁽¹⁾ which optimizes both conversion loss and RF VSWR over a wide LO dynamic range through the use of an optimum load line.

*Where there is a resistance load in the diode's circuit, the voltage acting on the diode is a function of this external resistance network and any external bias voltage applied. This function can be expressed as an equation and may be represented by what is commonly known as a load line on the diode I-V characteristics. Because the load is a pure resistance it will obey Ohm's Law, and the relationship between current and voltage will be a straight line. Zero load resistance is indicated by a vertical load line, while a horizontal line indicates infinite resistance.

Figure 7 - Conversion Loss Contours



Each Anaren biased mixer is individually characterized by measuring two current and voltage points on the optimum load line. These values are supplied on each mixer test data sheet in the form shown by Figure 9.

Figure 9

LO @ +7 dBm	LO @ -5 dBm
V _H = ____VOLTS	V _L = ____VOLTS
I _H = ____MA	I _L = ____MA

One of the measured points (V_H, I_H) is measured at +7 dBm LO power and the other point current and voltage axes, as LO power. The current intercept (I₀) and voltage intercept (V₀) for the required resistor calculations can then be found graphically by extending the optimum load line through the two measured points to the (V_L, I_L) is measured at -5 dBm illustrated in Figure 7.

Alternately, I₀ and V₀ could be calculated directly by using:

$$V_0 = V_L + \frac{I_L}{m}$$

$$I_0 = mV_0$$

where $m = \frac{I_H - I_L}{V_L - V_H}$ = slope of the optimum load line

Example: Assume a +6V supply and calculate the required values for R₁ and R₂ for optimum load line bias technique of Figure 8b. From Figure 7 or test data information (Figure 9) supplied with mixer:

$$\begin{aligned} V_H &= 0.12V & V_L &= 0.61V \\ I_H &= 3.6\text{ mA} & I_L &= 0.95\text{ mA} \end{aligned}$$

$$\begin{aligned} \text{then } m &= \frac{(3.6 - 0.95) \times 10^{-3}}{(0.61 - 0.12)} \\ &= \frac{2.65 \times 10^{-3}}{0.49} = 5.41 \times 10^{-3} \end{aligned}$$

$$\text{and } V_0 = 0.61 + \frac{0.95 \times 10^{-3}}{5.41 \times 10^{-3}} = 0.61 + .176 = .786V$$

$$I_0 = 5.41 \times 10^{-3} (.786) = 4.25 \times 10^{-3} A$$

Therefore:

$$R_1 = \frac{6.0V}{4.25 \times 10^{-3} A} = 1.412 \times 10^3 \text{ ohms}$$

$$R_2 = \frac{.786(1.412 \times 10^3)}{6.0 - .786} = \frac{1.110 \times 10^3}{5.214} = 213 \text{ ohms}$$

$$\begin{aligned} \text{Use } R_1 &= 1.5 \text{ K ohms} \\ R_2 &= 220 \text{ ohms} \end{aligned}$$

The two-resistor bias network just described minimizes the mixer conversion loss over an extended LO dynamic range. It also turns out that spurious and two-tone intermodulation performance of these mixers can be improved by operating the diodes at high LO powers while the bias network maintains optimum conversion loss.

Both the constant-current and optimum load line bias techniques should be temperature compensated if

performance is to be maintained over wide temperature variations. If temperature compensation is not used, the constant current load line is more tolerant of temperature variations, especially for mixer operation at low LO power.

These mixers having the bias option differ from some types of biased mixers because they can also be used with no external bias. This is accomplished by directly grounding the bias terminal (zero load resistance) and applying sufficient LO power. As shown in Figure 4, an LO drive of +7 dBm is typically required for optimum performance.

Figure 3 shows how the bias circuitry is implemented in the mixer. The bias terminal is an EMI low-pass (pi-network) filter. It does not protect against low frequency (less than 1 MHz) noise or voltage surges. The bias supply should, therefore, be reasonably well regulated and filtered for optimum mixer performance and diode protection.

Intermodulation Performance

Strictly speaking, all mixer outputs other than the desired sideband (LO - RF = IF) are "spurious" responses. However, the spurious responses that will be considered in this section are the two major types of distortion products—single-tone (harmonic) intermodulation and two-tone intermodulation.

A mixer diode, due to its I-V curvature, is an efficient harmonic generator and will produce the entire spectrum of mixing terms present in the Taylor's expansion of diode current. All of these terms are of lower amplitude than the desired IF output but can be reduced even further by various techniques. A *single-ended* mixer filters out many of the higher-order terms with a frequency selective decoupling network behind the diode. *Balanced* mixers depend on the phasing properties of their LO and RF coupling networks, (3 dB, 90° or 180° hybrids) for added spurious suppression. Most *double-balanced* mixers, by virtue of their 4-diode network and symmetry of their LO and RF coupling networks, achieve additional suppression without need for frequency selective filters.

Harmonic Intermodulation

Single-tone harmonic intermodulation products result from the mixer local oscillator and its harmonics beating with the RF input signal and its harmonics. They are of the form mLO ± nRF and are usually classified by their "order". For example, when the third harmonic (m = 3) of the LO beats with the second harmonic of the RF (n = 2) the resulting intermodulation product is a fifth-order term (m + n = 5). This term is also commonly called a "three-by-two" (3x2).

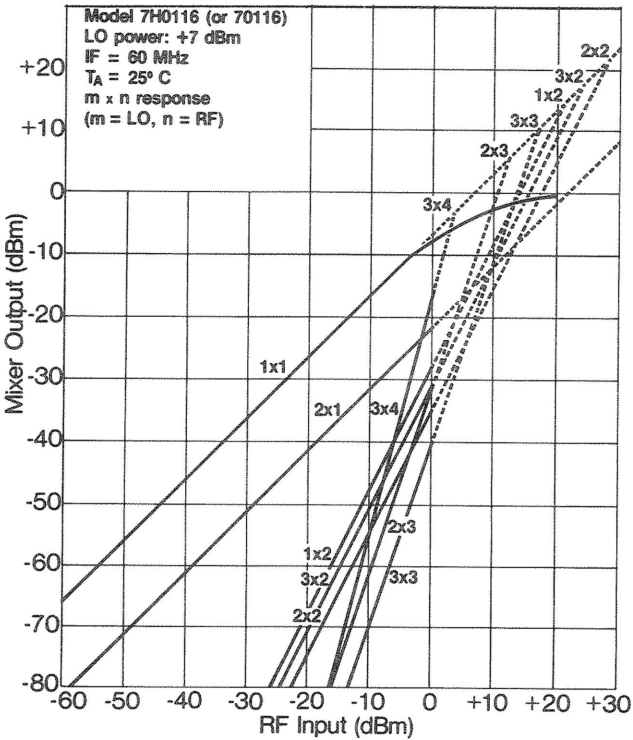
Suppression of single-tone intermodulation products by a balanced mixer* depends primarily upon good mixer balance. The more uniform the distribution of RF and LO power to the diodes, and the more closely matched the diode characteristics, the better the mixer balance.

*Compared to an unfiltered, single-ended mixer.

Suppression of single-tone (harmonic) intermodulation products is dependent upon the RF and LO phase codes at the diode junctions. If the RF and LO harmonics are properly phased, complete suppression of some harmonic intermodulation terms is possible. For example, all four basic mixer types, theoretically, completely suppress the 2x2, 4x4, and other even m x n harmonics. (Conversely, none of the mixers suppress the 3x3, 5x5 or other odd m x n harmonics).

The 90° hybrid balanced mixer has a phase code that provides at least 3 dB additional suppression for all harmonics of the LO and RF inputs, save the 2x4, 4x2, 3x3 and 5x5. It also, theoretically, completely suppresses the 2x2, 4x4, 3x1, 1x3, 3x5 and 5x3 terms. [In this discussion of intermodulation performance we will limit ourselves to 5x5 terms or lower (m, n ≤ 5).] Figure 10 shows typical single-tone intermodulation response for a 90° mixer.

Figure 10 - Single-Tone Intermodulation Response



The Ortho-Quad® mixer provides identical performance to the 90° mixer except there is as much as 15 dB additional suppression for all the harmonics of the RF input.

The 180° hybrid mixer's suppression characteristics depend upon whether the sum (Σ) or difference (Δ) port of the hybrid is selected as the LO port. When the LO signal is applied to the Δ port, all even harmonics of the LO signal are suppressed. (If the LO is applied to the Σ port, second-order intermodulation is suppressed.)

The double-balanced mixer theoretically can provide complete suppression of all even harmonics of the LO and RF signals. Specification sheets for low frequency (<100 MHz) double-balanced mixers typically provide

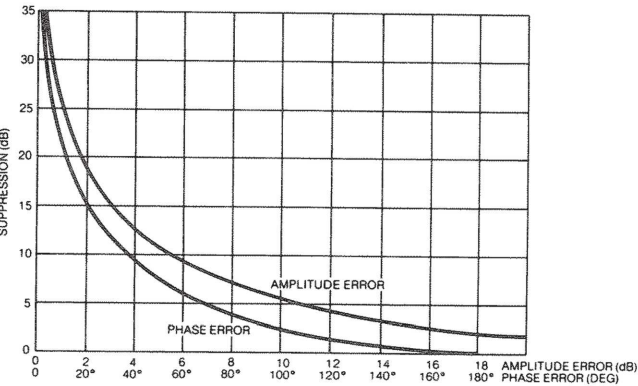
harmonic intermodulation charts for m x n responses which show suppressions as high as 90 dB. Our experience with microwave balanced and double-balanced mixers indicates the microwave system designer should be very cautious about extrapolating this low frequency data to microwave frequencies.

At frequencies of 100 MHz or less it is possible to achieve extremely broadband coupling networks that provide phase and amplitude balances of ±1 degree and ±0.1 dB at the input frequencies and their harmonics. This performance is required to obtain harmonic suppression greater than 50 or 60 dB at -10 dBm input levels. Specifications of ±5° and ±0.3 dB for phase and amplitude balance are typical for microwave LO and RF coupling networks.

Because phase errors are multiplied by the order of the RF and LO harmonic, it is possible to have very large phase errors that will reduce the actual harmonic suppression. As an example: the 2x2 (mxn) suppression for all mixers is theoretically infinite, but a calculation shows the typical microwave balun could possibly have m(5°) + n(5°) = 2(5°) + 2(5°) = 20° phase error.

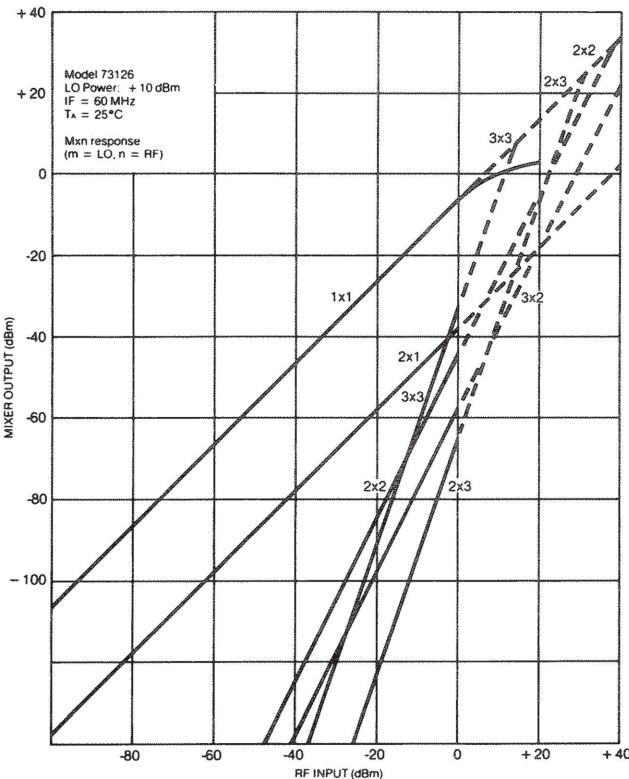
Figure 11, showing the theoretical suppression for various values of phase and amplitude error, gives a maximum suppression value of 15 dB for 20° phase error. At some other point in the band there may be zero phase error, if the balun errors cancel. At that point the suppression would approach infinity (or more realistically, be limited by the balun amplitude balance and/or the diode balance). Our experience has shown the amount of actual suppression to vary by as much as ±15 dB depending on the exact frequency of operation and the particular order of intermodulation.

Figure 11 - Suppression vs. Phase and Amplitude Error



The double-balanced mixer *will* consistently provide the best |m-n| = 1 suppression, however. These terms (4x5, 5x4, 3x4, 4x3, etc.) are important because they, along with the 2x2 and 3x3 terms, lie closest to the desired response. Figure 12 shows typical single-tone intermodulation response for a double-balanced mixer.

Figure 12 - Single-Tone Intermodulation Response

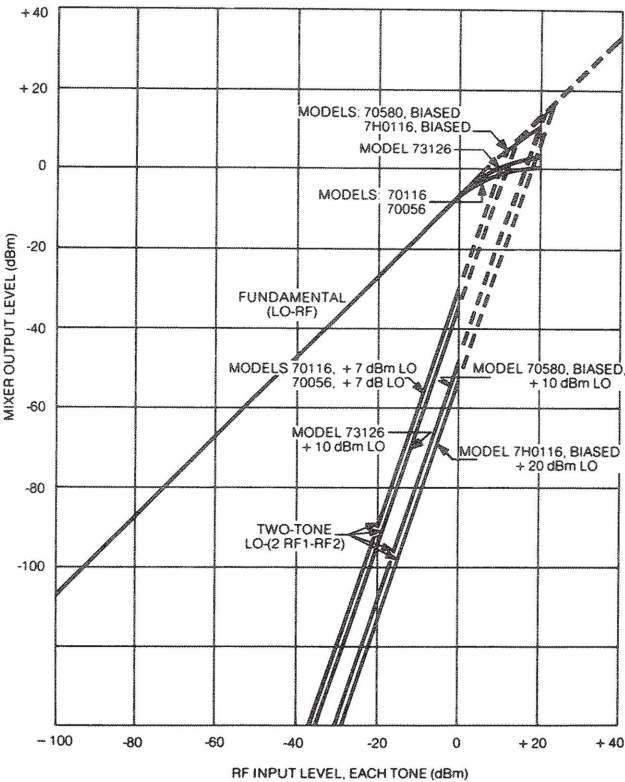


Two-tone Intermodulation/Intercept point

Two-tone intermodulation distortion is the result of two signals being applied to the mixer RF input at the same time. These signals may generate harmonics, mix with each other, then beat with the mixer local oscillator according to the expressions $(2RF_1 \pm RF_2) \pm LO = IF$ or $(RF_1 \pm 2RF_2) \pm LO = IF$. These are the expressions for the *third-order* two-tone intermodulation products. They are usually of primary interest because of their relatively large magnitude and because they are difficult to filter from the desired mixer output.

A convenient method of determining the two-tone third-order performance of a mixer is the “third-order intercept” approach. It makes use of the intersection of the fundamental and third-order responses on a log-log (or dB/dB) scale. A plot of each response will be a straight line with a slope corresponding to the order of the RF response. (The 1x1 fundamental is a first-order response, a 3x2 is a second-order response, a 2x3 is a third-order response, etc.) The third-order, two-tones will have a slope of 3 when plotted on a dB/dB scale. Figure 13 is a plot of the fundamental and third-order responses for some typical mixers. The third-order intercept for the biased Model 7H0116 mixer occurs at an RF *input* level of + 15 dBm. Note this is a theoretical point, graphically obtained by extending the measured third-order and fundamental responses until

Figure 13 - Fundamental and Two-Tone, Third-Order Response



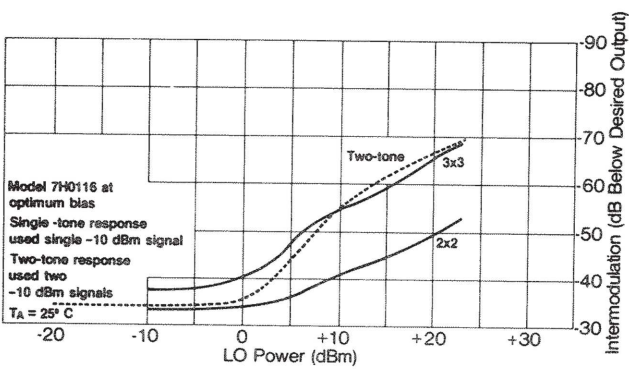
they intercept. Once the intercept point is known the third-order response at any input level can be calculated: with a slope of 3, a 1 dB change in the input level will produce a 3 dB change in the third-order output. Various mixers can be compared for third-order suppression if their intercept points are known—the higher the intercept point, the better the suppression.

Improvement with Bias

Note in Figure 13, the third-order performance of most of the unbiased mixers, regardless of what type they are, is nearly identical. No mixers (balanced or double-balanced, of any type) provide any additional two-tone suppression compared to a single-ended mixer. Two-tone suppression is a function of diode I-V characteristics.

Improved rejection of both single-tone (harmonic) and two-tone intermodulation *can* be improved with the use of high mixer LO power and/or high bias voltage as shown in Figure 14. The intercept point will increase approximately dB for dB, with LO power, if the mixer is properly biased. (Note in Figure 13, the higher intercept point for the biased Model 7H0116 compared to its unbiased version, Model 70116.) Lepoff and Cowley⁽²⁾ have shown this intermodulation improvement to be due to the diode characteristic's departure from the theoretical expression for diode current.

Figure 14 - Intermodulation vs. LO Power



Intermodulation is caused by non-zero values for fourth-order and higher terms in the power series expansion of the diode current-voltage function. For diode voltages near 1 volt they found that the measured diode current drops below the theoretical curve. This meant that the higher-power (intermodulation producing) terms in the power series expansion were smaller at those voltages than they were at lower levels. Therefore, intermodulation rejection could be improved by operating the diode at a higher bias voltage than the *usual zero bias*.

In a balanced mixer in which the two diodes can be operated at different bias voltages (as the Anaren 180° hybrid biased mixers having two bias terminals) some of the higher power terms of one diode's current function can actually be made to cancel the corresponding terms in the other diode's current function. This technique has been used with mixers operating in the 2.2-2.3 GHz telemetry band to obtain greater than 100 dB rejection of spurious components. If the bias voltage is increased still further, the diode current saturates and the second-order (mixing) term decreases in magnitude. There is an optimum bias voltage which minimizes intermodulation yet still provides good conversion loss.

Other high-level mixer designs⁽³⁾ use series diodes or series resistors to achieve the bias effect necessary for good intermodulation performance. These additional elements add complexity, cost, and increased conversion loss and are restricted to optimum performance for high level signals. The Anaren *optimum load line* bias technique allows the lowest conversion loss and the highest intercept point, regardless of signal or LO drive. The LO power can be raised to the level required for best two-tone suppression; without suffering the conversion loss penalty, associated with high LO drive, of mixers not using the load line technique. A special, high-cost, high-level mixer for best two-tone performance is not required. Figure 15 shows the intermodulation performance obtained using the optimum load line bias technique described on page 000. Figure 16 shows performance

Figure 15 - Conversion Loss (at Optimum Bias) & Intermodulation vs. LO Power

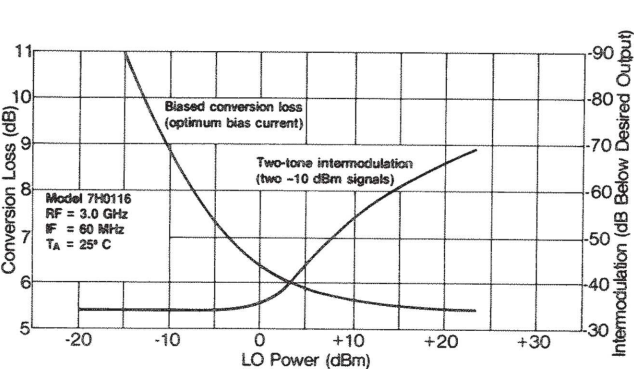
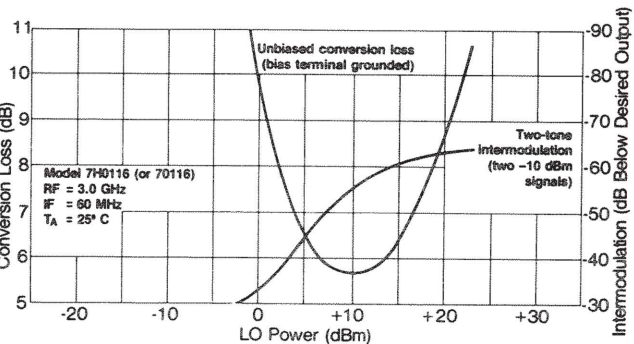


Figure 16 - Conversion Loss (No Bias) & Intermodulation vs. LO Power



when using a non-biased mixer. Comparison of the two figures demonstrates the advantage of using the optimum bias capability. At an LO power of + 23 dBm it provides about 69 dB of intermodulation suppression and a mixer conversion loss of about 5.5 dB. At the same high LO power a normal unbiased mixer provides about 63 dB suppression but the conversion loss has increased to 10.6 dB.

Image Focusing Port

In some mixer requirements, it is important that the internally generated image signal (2LO-RF) be focused where it will not interfere with generation of the desired IF output. One of the sources of phase and amplitude ripple produced by a mixer is the internally generated image signal leaking from the RF port. This signal can be reflected back into the mixer by other component mismatches of the system, mix with the LO and convert to the IF frequency as an error component. The image signal leaking from the LO port does not cause the same problem because it is suppressed by the mixer's common-mode rejection (all four basic mixer types feature good common-mode rejection.)

The 180° mixer and the double-balanced mixer focus the image at the RF ports and the 90° mixer focuses it to the LO port. The Ortho-Quad® mixer does not let the image signal come out either the LO or RF ports. It, and other potentially troublesome harmonic terms, are focused to an internal load. For additional information on other phase and amplitude error considerations see page 116.

IF Bandwidth

The 90° mixer, the 180° mixer and the Ortho-Quad® mixer frequency selective decoupling networks behind the diodes. These networks suppress higher-order spurious components, increase the LO/IF and RF/IF isolation, and minimize the mixer conversion loss at low IF frequencies. A plot of relative conversion loss versus IF frequency for the above 3 mixer types would show the IF response rolling off at approximately 6 dB/octave, and the –3 dB response point at approximately .2 RFL, where RFL is the lower RF band edge. All unbiased units have an IF response that extends down to dc with no conversion loss degradation. Biased units normally contain a dc blocking capacitor(s) in series with the IF output. This limits the low frequency –3 dB response to about 5 MHz.

The double-balanced mixer is available in two versions. One version, in the standard mixer packages, has frequency selective filters at the IF port as described above. The second version, in a miniature package, has an extremely wide IF bandwidth. The IF response extends from dc up into the RF passband because it is extracted with a phasing technique rather than by filtering. These high IF mixers can be extremely useful. Large segments of the RF passband can be downconverted and processed instantaneously using a fixed-frequency (non-scanning) LO. This means a much simpler LO system and elimination of the image and 2x2 and 3x3 spurious response. If the double-balanced mixer is used as a bi-phase (BPSK) modulator, the wide IF band allows very high data rate modulation.

Dynamic Range

Simply defined, dynamic range of a balanced mixer is the power range over which the device can be used. The absolute maximum useful range is limited on the low end by the minimum detectable signal (MDS) and at the high end by mixer saturation level. (The saturation level occurs at an input power level approximately equal to the LO power level less the conversion loss of the mixer).

For most applications however, the dynamic range of interest is usually the “spurious free dynamic range”. The upper limit on this dynamic range is set by the maximum input level that generates third-order intermodulation products just equal to the mixer output noise level.

An expression⁽⁴⁾, using the intercept point (IP) for spurious free dynamic range (R_{sf}) is:

$$R_{sf} \text{ (in dB)} = 2/3 [IP - L_c - MDS]$$

Where L_c is mixer conversion loss (dB). MDS is 3 dB above thermal noise:

$$MDS = -111 \text{ dBm/MHz} + 10 \log_{10} B$$

Where B is the bandwidth in MHz and –111 dBm is 3 dB above thermal noise in a 1 MHz bandwidth.

Example: An optimally biased Model 7H0116 has an intercept point of + 25 dBm. If an IF bandwidth of 10 MHz is chosen, the spurious free dynamic range is:

$$2/3 [+ 25 - 5.5 + 101] = 2/3 [+ 120.5] \text{ or } 80 \text{ dB.}$$

Applications of Balanced and Double-Balanced Mixers

Downconverter

The preceding pages have been concerned with performance of the four basic mixer types when used as downconverters. These basic mixer types can be used singly and/or combined with other components to form Image Rejection mixers. Quadrature IF mixers, etc. These downconversion devices will be discussed later.

Upconverter (Balanced Modulator)

In upconverter or balanced modulator applications the lowest frequency signal is used as an input and applied to the IF port. *Balanced modulation* usually implies a double-sideband, suppressed carrier, output signal with the modulation sidebands located close to the carrier. *Upconversion* usually means that the LO (pump) signal is widely separated from the desired sideband outputs.

For either application, the balanced modulator accepts an RF input and a modulation input (usually called the IF) and produces an output consisting of the upper (RF + IF) and lower (RF – IF) sidebands. The original RF input (carrier) is normally suppressed at the output.

For double-sideband, suppressed carrier applications, the RF input is normally a low-level (–10 dBm typical) signal while the IF input is a low-frequency, modulating signal. The modulating signal is used as the high level switching signal and is usually + 7 to + 10 dBm. The RF output consists of the upper and lower sidebands centered about the suppression RF carrier. The degree of carrier suppression is directly related to the RF input/RF output isolation of the modulator. This carrier suppression may vary from 6 to 40 dB depending on the type of modulator and the frequency involved.

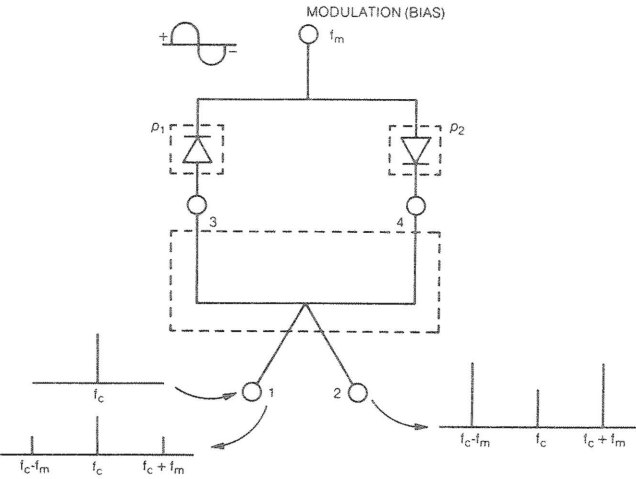
The 90° hybrid mixer and the Ortho-Quad® mixer are *not* used as modulators because their LO and RF coupling networks use 90° hybrids. The 90° hybrid mixer, if used as a balanced modulator, suppresses the sidebands but does not suppress the carrier. The Ortho-Quad® mixer suppresses both the carrier and the sidebands.

The 180° hybrid coupling networks, used in the 180° mixers and the double-balanced mixers, offer the right phasing characteristics so they can be used either as mixers (downconverters) or as modulators (upconverters). The balanced modulator using a single 180° hybrid will be described for simplicity but the double-balanced mixer (used as a modulator) works on the same basic principle. The difference is that separate 180° RF and LO hybrids (baluns) drive a balanced diode quad in a full-wave configuration, for the double-balanced modulator.

Balanced Mixer/Modulator Using Single Hybrids

A balanced mixer/modulator can be constructed by placing two mixer diodes with reverse polarity behind a 180° hybrid and combining the IF in a tee as shown in Figure 17.

Figure 17 - Balanced Modulator Using 180° Hybrid



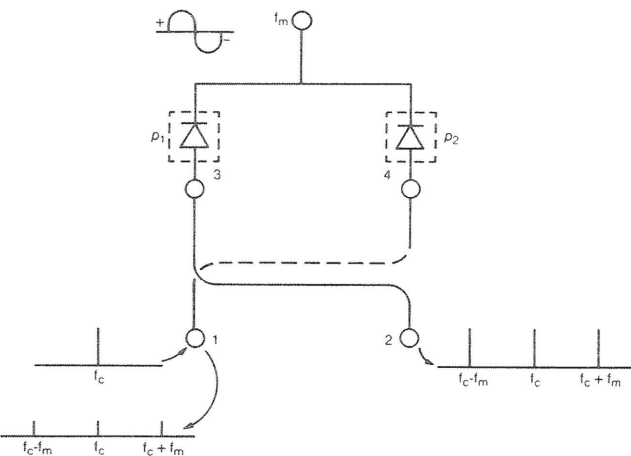
The incident RF carrier sees like reflection coefficients at the diodes and reflects back to the input port. The modulating (biasing) signal in the IF port produces opposite polarity reflection coefficients at the sum and difference frequencies and these modulated signals come out the remaining port.

Assuming perfect isolation for the hybrid, the amount of carrier suppression depends primarily on the similarity of the diodes—if they have equal reflection coefficients all the residual carrier is reflected back to the input port and none shows up at the output. This residual carrier shows up as VSWR at the input port, however. In effect, a balanced modulator using a 180° hybrid trades carrier suppression for VSWR.

Both 180° hybrid, and double-balanced, mixer/modulators provide good carrier suppression and have poor VSWR. The double-balanced mixer provides better carrier suppression because it has higher LO/RF isolation. It also offers more harmonic suppression because of its full-wave, double-balanced, drive. The most serious problems occur because of the poor mixer VSWR. When the mixer ports are not terminated in well-matched loads, reflections can cause severe amplitude and phase distortion due to the VSWR beats. The LO and IF ports are particularly sensitive to reactive terminations, which can cause conversion loss variation of ± 3 dB and spurious response variations of ± 20 dB.⁽⁵⁾

A balanced modulator using a 90° hybrid can be constructed as shown in Figure 18, by placing two mixer diodes with the *same* polarity behind the hybrid. However, this modulator, used alone, provides no carrier suppression. Because both diodes are inserted with the same polarity, the carrier and the modulation sidebands see like reflection coefficients. For a 90° hybrid this means that no energy reflects back to the input (therefore, good VSWR). There is poor carrier suppression because the carrier shows up at port 2 at a comparable amplitude to the sidebands.

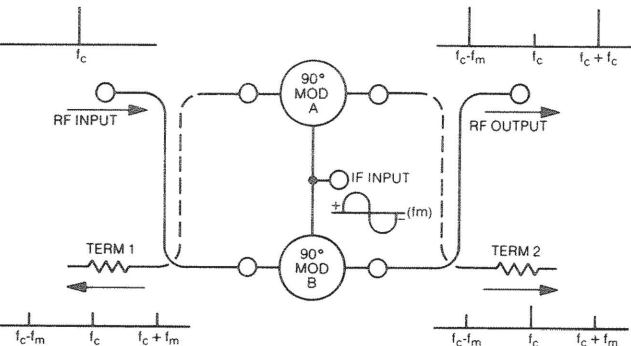
Figure 18 - Balanced Modulator Using 90° Hybrid



A High Performance Balanced Modulator

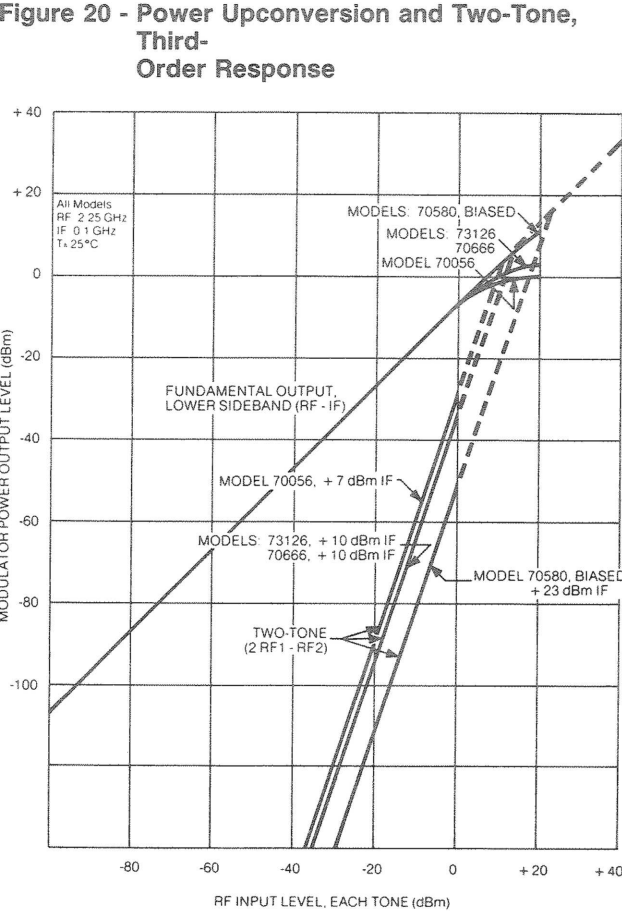
In order to get good carrier suppression and good VSWR simultaneously, Anaren uses the special patented circuit of Figure 19 (which is a balanced modulator version of the Ortho-Quad® balanced mixers).

Figure 19 - 70660 Series Balanced Modulator



As previously noted, the 90° balanced modulators (A, B) of Figure 19 have no carrier suppression of their own, and the unsuppressed carrier shows up at the same port as the modulated signal. However, the carrier and the modulated sidebands have different *phase* relationships. This makes it possible to separate the carrier from the sidebands by using RF phasing networks. The RF phasing properties of the INPUT and OUTPUT 90° hybrid direct the unsuppressed carriers to termination 2 where they are absorbed. The modulated signals, having a different phase code, are focused to the RF output port. Termination 1 absorbs the residual carriers from A and B and provides the good VSWR. The result is the 70660 Series balanced modulator; providing good carrier suppression and good VSWR, simultaneously.

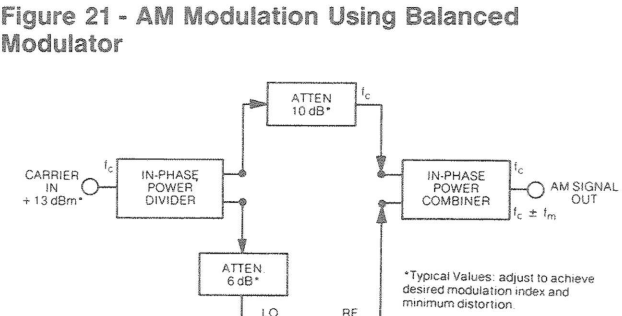
Figure 20 shows typical power up-conversion and two-tone, third-order performance for the 180° balanced, double-balanced, and 70660 series modulators. Note the intercept points for all these mixers are identical, except for the biased Model 70580 that uses the optimum load line technique to achieve good two-tone performance at high low power.



AM Modulator

The balanced modulator can easily be converted to an amplitude modulator by reinserting the required amount of carrier at the output of the mixer. One method of re-establishing the carrier is unbalancing the modulator by applying a dc bias current along with the modulation signal to the IF port. This decreases the RF input/RF output isolation, letting more of the original carrier leak through to the output.

A second approach offers lower distortion and better control of the modulation index by reinserting the carrier after the modulator. This approach is illustrated in Figure 21. The 70660 Series balanced modulators offer the best VSWR at all ports, independent of drive levels. This minimizes the output distortion.



Pulse Modulator

The balanced modulator is used as a pulse modulator by applying the carrier signal to the LO (or RF IN) port and the modulating pulse (either positive or negative) to the IF port. The pulsed output is then taken from the RF (RF OUT) port. The normal pulse requirements are 0 to 20 mA. The IF port will typically switch in less than one nanosecond and will support a pulse of indefinite length. Very clean pulse bursts, with minimal carrier feedthrough are provided for RF levels between 0 and +10 dBm.

Again, the 70660 Series offers the best performance for most applications, because of its very low VSWR and high LO/RF (RF IN/RF OUT) isolation.

Bi-Phase Modulator

Bi-phase modulation is similar to pulse modulation—reversing the IF pulse polarity will switch the RF output phase by 180°. (A bi-polar drive of +10 mA is typically required.)

For those applications where the accuracy of the 0°, 180° phase and insertion loss states is critical, the modulator having the best LO/RF (RF IN/RF OUT) isolation and the best VSWR will provide the best results. The 70660 Series will typically provide phase accuracy of ±2 degrees and insertion loss tracking of ±0.2 dB for bandwidths up to 2 GHz. The absolute insertion loss is typically 3 dB.

For BPSK systems requiring extremely high data rates, the 73000 series of double-balanced mixers offers bi-phase data rated up to 4 gigabits/sec.

Current-Controlled Attenuator/Switch

Balanced modulators can be used as wide-range, current-controlled attenuators or high on-off ratio switches. The isolation between the LO and RF (or RF IN, RF OUT) ports is controlled by the amount of dc current flowing through the IF port. The control current polarity is irrelevant. The maximum attenuation ("OFF" state) is equal to the LO/RF (RF IN/RF OUT) isolation. The minimum attenuation at 10-20 mA of control current is typically 2-3 dB.

The VSWR at the RF ports will vary with the control current for the 180° and double-balanced mixers. The VSWR for the 70660 Series modulator will remain low and independent of control current.

Figure 22 - LO to RF Attenuation vs. IF Current

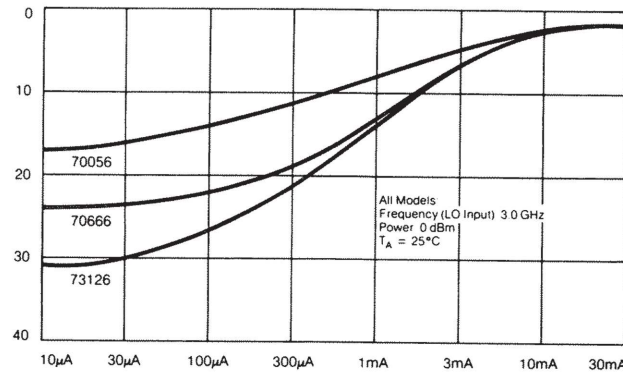


Figure 22 shows typical attenuation vs. control current for the balanced modulators.

NOTE: For all applications requiring bias current, avoid damage to the modulator diodes by *always* driving the IF port with a current-limited source.

Phase Detector

All four basic *mixer* types are useful as microwave phase discriminators. (The 70660 Series of balanced modulators is not designed for use as a phase detector). When identical frequencies are fed to the mixer LO and RF ports, a dc output proportional to the phase difference between the two signals will appear at the IF port.

The 90° hybrid mixer produces an output voltage proportional to the *sine* of the phase difference between the two identical input frequencies:

$$V_{outDC} = K_1 \text{SINE} (\theta_L - \theta_R)$$

Where K_1 depends on the input power and diode characteristics.

$(\theta_L - \theta_R)$ = phase difference of the two inputs.

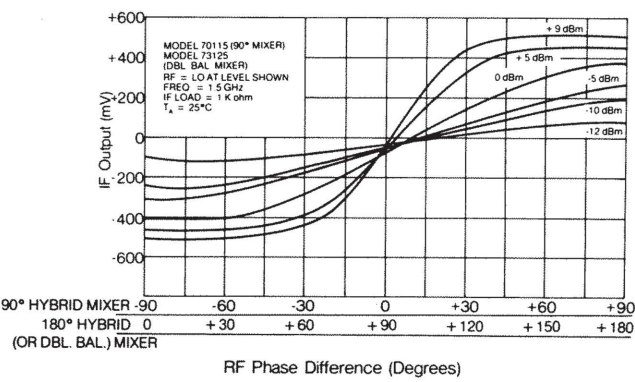
A typical value of K_1 at 0 dBm input power (-3 dBm per diode) is 500 mV/mW. This is the voltage sensitivity when loaded into 1 K ohm.

The 180° hybrid mixer, and the double-balanced mixer, produces a dc output proportional to the *cosine* of the phase difference between the two identical input frequencies:

$$V_{outDC} = K_1 \text{COSINE} (\theta_L - \theta_R)$$

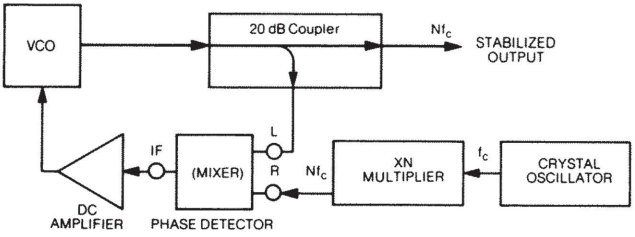
Figure 23 shows IF output voltage as a function of the phase difference of the two inputs. The 90° hybrid mixer produces a dc null when the input signal phases are equal. The 180° hybrid, and double-balanced, mixers produce a maximum negative output when the input signal phases are equal.

Figure 23 - Phase Detector Performance



A typical application for a phase detector is in a phase-locked loop. The block diagram shown in Figure 24 shows a phase detector being used as the error sensor for a phase locked source.

Figure 24 - Block Diagram of Phase-Stabilized Frequency Source



Frequency Discriminator

A phase detector can be converted to a frequency discriminator by adding a power divider and a delay line as shown in Figure 25.

The unknown RF signal is divided to a reference and a delay path. The differential delay (T) creates a phase difference (θ) between the two signals which is a linear function of frequency (f) and is given by

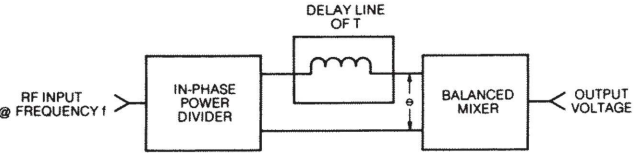
$$\theta = 2\pi fT$$

or

$$\theta = 2\pi f \left(\frac{\ell}{C} \right)$$

where ℓ = length of the delay line
 C = velocity of propagation

Figure 25 - Frequency Discriminator



The 90° hybrid mixer, and the Ortho-Quad® mixer, produce an output voltage proportional to

$$V_{out} = kP \sin \theta$$

The 180° hybrid mixer, and the double-balanced mixer, produce an output voltage proportional to

$$V_{out} = kP \cos \theta$$

k is a constant related primarily to insertion loss and diode sensitivity. P is the RF input signal power.

If the RF input is 0 dBm, K is typically 500 mV/mW when the output is loaded into 1 K ohm.

The output voltage of the 90° hybrid mixer, and Ortho-Quad® mixer, has zero crossings when $\theta = 0, \pi, 2\pi$, etc. For example, when

$$f = \frac{n}{2T} = \frac{nc}{2L}$$

where n = 0, 1, 2, 3, etc.

The output voltage of the 180° hybrid mixer, and double-balanced mixer, has zero crossings when

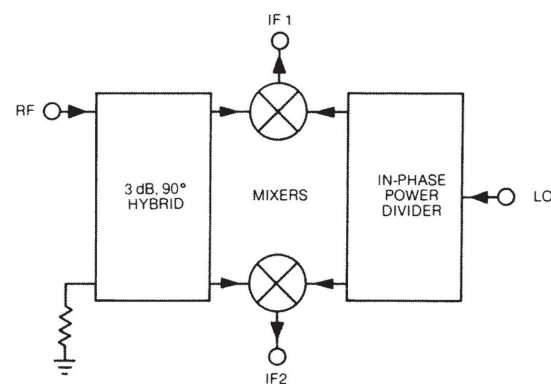
$$\theta = \frac{\pi}{2}, \frac{3\pi}{2}, \frac{5\pi}{2}, \text{ etc.}$$

Near the zero crossing, $\sin \theta$ and $\cos \theta$ are almost linear functions of θ , so the output voltage is nearly linear with frequency (within ± 30 degrees of the zero crossings the deviation from linearity is less than $\pm 5\%$). The *slope* of the output voltage curve is a function of input power and the discriminator must be preceded with a hard limiter to remove the power/frequency ambiguity.

Quadrature IF Mixers

The Quadrature IF Mixer (QIFM) provides two equal amplitude IF outputs that are in phase quadrature. It consists, basically, of two balanced mixers and two hybrids. Figure 26 is a block diagram.

Figure 26 - Block Diagram of a Quadrature IF Mixer



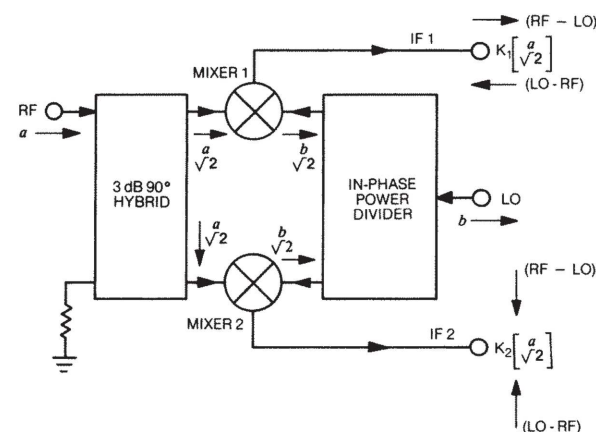
Operation

The LO is introduced to the two mixers in-phase and the RF is introduced through a quadrature hybrid. This introduces a 90° phase difference between the RF and LO signals applied to the mixers, and results in two IF outputs of equal magnitude but in phase quadrature. One IF output leads or lags the other depending on whether the RF frequency is greater or less than the LO frequency. If the LO signal is a fixed frequency, the desired IF output can be produced by an RF signal above or below the LO frequency:

$$RF = LO \pm IF$$

The operation of a QIFM can be understood by following the signal voltage vectors in Figure 27.

Figure 27 - Simplified Vector Operation of Quadrature IF Mixer



With a high level LO signal, b , (typically greater than 0 dBm), an RF input, a , will produce IF outputs equal to: $k_1 (\frac{a}{\sqrt{2}})$ and $k_2 (\frac{a}{\sqrt{2}})$ where k_1 and k_2 are proportional to the conversion losses in mixers 1 and 2. The vectors show the relative phase of the IF outputs for the upper (RF - LO) and lower (LO - RF) sideband responses.

When the RF frequency is below the LO frequency, the IF response (LO - RF) from mixer 1 lags the output of mixer 2 by 90°. When the RF is above the LO, the IF response (RF - LO) leads the output of mixer 2 by 90°.

Another way of looking at the operation of the QIFM is to realize that the two IF outputs are proportional to $a \sin \theta$ and $a \cos \theta$, respectively. θ is the difference phase between signals a and b . Note that θ can be a fixed angle if the two frequencies are identical, or is time varying at the difference frequency ($\Delta \omega t = IF$). Since the IF outputs are proportional to $\sin(a-b)t$ and $\cos(a-b)t$ respectively, and since $\sin(-\theta) = -\sin \theta$ and $\cos(-\theta) = \cos \theta$, one output changes phase by 180° relative to the other when $a > b$ or $a < b$.

The two IF voltages, $a \sin \theta$ and $a \cos \theta$, are the in-phase (I) and quadrature (Q) components of the correlation of the RF and LO signals. The correlation time is the inverse bandwidth of the IF outputs. The QIFM is identical in design and application to the correlator (phase discriminator) provided one of the QIFM inputs, a or b , is strong enough to act as an LO. (The correlator uses dc biased diodes, which enables it to operate with both a and b at very low levels.)

Intermodulation Performance

The mixers shown in Figure 26 could be any of the four mixer types: 90° balanced, 180° balanced, ORTHO-QUAD® or double-balanced. The intermodulation performance of the QIFM is the same as the intermodulation performance of the type of mixer used. (See pages 101-104 for a discussion of intermodulation performance for the four mixer types.) Anaren uses the 90° balanced or the ORTHO-QUAD® in its QIFM designs because they provide better phase and amplitude balance at the IF ports than either the 180° balanced or double-balanced.

Applications

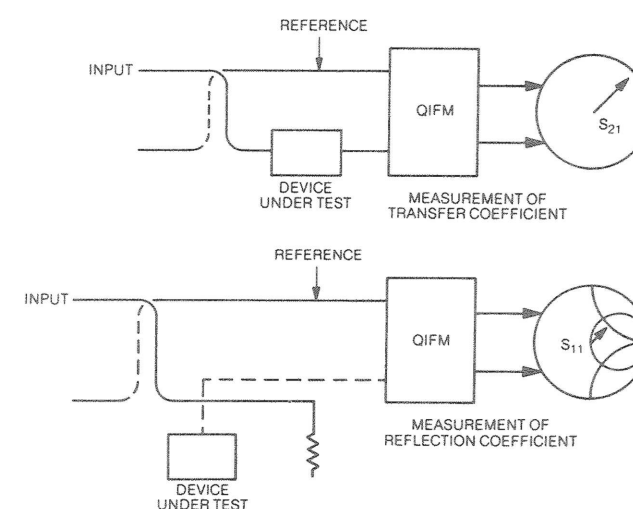
Doppler Systems

This network is used for many applications. The earliest was probably to discriminate between approaching and receding doppler targets. This application is described in detail by Skolnik⁽⁶⁾. For an approaching target, the received frequency is greater than the LO and vice versa for a receding target. The relative phases of the received doppler signals (IF) are used to discriminate between the two situations. This technique is used for motion detector, "burglar" alarm, and radar altimeter applications, to name a few.

Network Analyzer

When the LO and RF signals are at the same frequency, the IF outputs are still proportional to $a \sin \theta$ and $b \cos \theta$ respectively, where θ is the phase difference between the RF and the LO signals. The device can therefore be used as a vector voltmeter, when connected to an X-Y oscilloscope, as shown in Figure 28.

Figure 28 - Quadrature IF Mixer Used as Network Analyzer



With the voltage $a \cos \theta$ applied to the X-axis of the oscilloscope and the voltage $a \sin \theta$ applied to the Y-axis, a pulsed RF signal appears as a strobe whose length is proportional to the RF voltage a and whose angle is θ . This makes the device useful as a test instrument to measure transmission and reflection coefficients — a network analyzer.

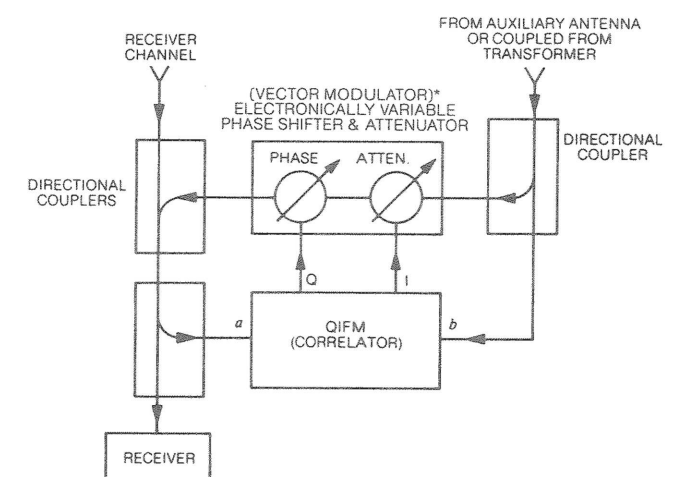
Adaptive Cancellor

The QIFM can be applied as a correlator and used in the adaptive cancellation of two microwave signals.

A typical example of a problem that requires adaptive cancellation is where a transmitter and receiver are operating in close proximity, such as a jammer and a surveillance receiver. Transmitter leakage to the receiver via the antennas usually is sufficiently strong so as to blind the receiver during transmission. Another example is that of a hostile jammer blinding a receiver through its sidelobes.

To combat such situations, short of turning off the receiver, a scheme such as indicated in Figure 29 can result in a significant reduction of the interference at the receiver, while still maintaining look-through at the outside world.

Figure 29 - Adaptive Cancellor



*See page 125 for details.

The essentials of the scheme are that we pick off a strong sample of the interfering signal, (either by means of a coupler in the first case, or by means of an auxiliary omnidirectional antenna in the second case), and add this sample to the signal being received by the receiver. By adjusting the amplitude and phase of the added signal we can arrange that it exactly cancels the interfering signal in the receiver channel. Since the relative amplitude and phase of the interfering signal will vary owing to relative motion of the antennas, vibration, frequency changes, etc., we are required to continuously adjust the phase and amplitude of the cancelling signal to maintain the nulled condition at the receiver. This is achieved by means of a quadrature IF mixer as shown.

A sample of the combined signal finally reaching the receiver is correlated with a sample of the jammer signal. Provided the jammer signal is strong enough to act as an LO, the outputs of the mixer are:

$$a \cos \theta$$

and

$$a \sin \theta$$

where a is the voltage entering the receiver, and θ is the difference phase between the two signals being correlated. Note that θ can be a fixed angle if the two frequencies are identical, or is time varying at the difference frequency ($\Delta \omega t$). Filtering these outputs with a low-pass filter (e.g. 1 KHz cut-off) insures that only signals that are extremely close in frequency give an output. The two error signals are thus the in-phase (I) and quadrature (Q) components of the correlation of the RF signals over a time period equal to the inverse bandwidth of the filter (1 msec. in the case of a 1 KHz cut-off).

If the two RF signals do not correlate, then both $a \sin \theta$ and $a \cos \theta$ are zero. That is, the interfering signal has been cancelled at the receiver. The two error signals I and Q drive the variable attenuator/phase shifter (Vector Modulator) in a null seeking mode to insure a continuing null at the receiver. This adaptive scheme will affect the desired signal at the receiver by less than 1 dB.

Digital (PSK) Demodulator

The QIFM can also be used to demodulate a QPSK digital phase format and recover the two data streams. The QPSK input is applied to the RF port while the regenerated carrier is applied to the LO port. The demodulated data streams are then available at IF ports 1 and 2. Figure 30 shows a basic modulator/demodulator system block diagram. The QPSK modulator is the Quadrature Modulator specified on page 61 and explained on pages 114-115.

Figure 30 - Block Diagram of QPSK Modulator/Demodulator System

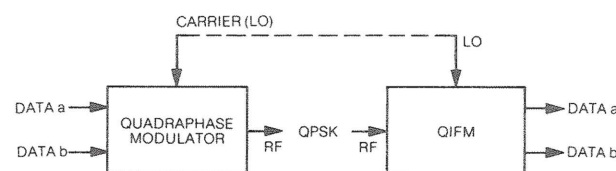


Image Rejection Mixers

From the previous section we know that a quadrature IF mixer can discriminate between the IF outputs produced when the RF frequency is above or below the LO frequency. This fact allows the realization of an Image Rejection Mixer (IRM). The IRM directs each IF output to a specific IF port and suppresses the undesired (image) IF response.

Operation

The desired IF frequency can be produced by an RF signal either higher or lower than the LO frequency by a frequency difference equal to the IF frequency:

$$RF = LO \pm IF$$

Normally, only one of these sidebands is desired and is designated the "real" IF output; the other is termed the "image". The choice of the upper or lower sideband as the "real" response is an arbitrary design decision.

From Figure 27 we know that the IF outputs of the mixers provide voltage vectors $k_1 (\frac{a}{\sqrt{2}})$, with phase dependent on the sideband. If we select the lower sideband ($LO - RF$) as the "real" response and apply the corresponding IF frequency, the vectors will ideally sum to only one IF (real) port and cancel at the other (image) port. See Figure 31.

Figure 31 - Simplified Vector Operation of Image Rejection Mixer

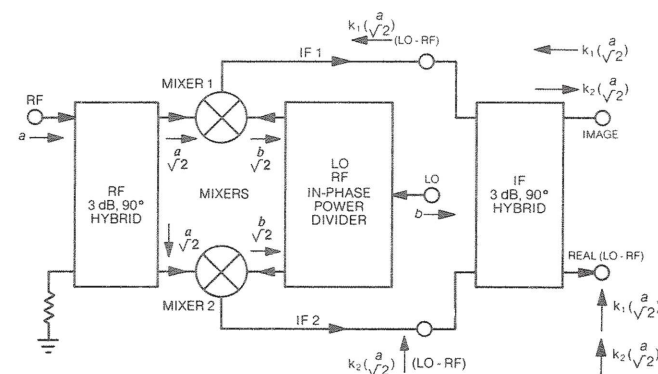
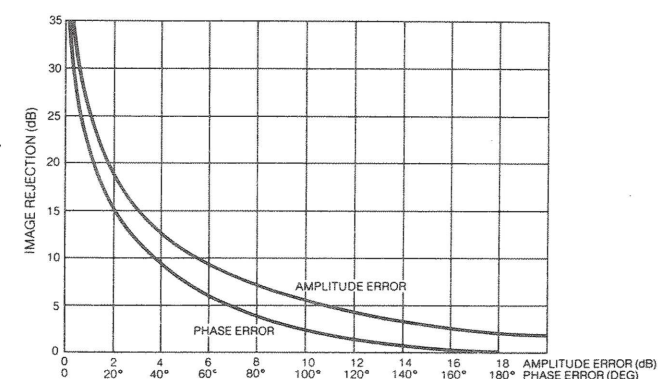


Image Rejection

If $k_1 \neq k_2$, or if the vectors are not in quadrature, they will not completely sum at the real port and cancel at the image port. Image rejection is the ratio of the image signal to the real signal at the real IF port. Good image rejection requires close control of the phase and amplitude of the signal vectors. Phase and amplitude errors occur primarily in the RF portions of the IRM network but can also occur when the mixer outputs are summed by the IF hybrid. Figure 32 is a chart showing how image rejection is affected by phase and amplitude errors. When $k_1 = k_2$, and there is no phase error, the output voltage at the real port is equal to $k_1 a$; the only loss is due to the mixer losses (k_1, k_2).

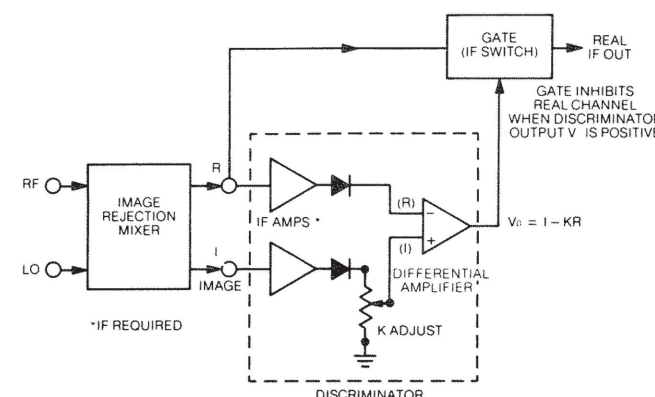
Figure 32 - Image Rejection vs. Phase and Amplitude Error



Complete Image Rejection Using Gating Techniques

Figure 32 shows that it takes only ± 0.5 dB of amplitude imbalance to reduce the image rejection to 25 dB (assuming perfect phase balance.) Typically, 20 dB of image rejection can consistently be obtained over octave bands. This may not be enough rejection for the applications in which image rejection mixers are normally used. Typical applications for these mixers are in wideband, sweeping LO receivers or spectrum analyzers. In both cases, the required dynamic range of the equipment may exceed 80 dB. The 20 dB image rejection is, therefore, not a very practical number for this application. However, once we have discriminated between the real and image responses to this degree, a simple gating scheme, such as illustrated in Figure 33 will give the additional rejection required.

Figure 33 - Gating Technique for Complete Image Suppression



The IF signals from the real (R) and image (I) ports are detected and fed to the inputs of a differential amplifier with an output: $V_0 = I - kR$ where k is an adjustable constant. When the I channel output is greater than the R channel output, the differential amplifier produces a positive output (V_0) which activates the IF switch (gate) and produces no IF output. If the R channel output is greater than the I channel, the amplifier output is negative, the IF gate is not activated, and the real IF is allowed to pass. If the constant k is adjusted to a value of 0.1, then the real channel IF output is inhibited when the real signal does not exceed the image by at least 10 dB. This insures that random signals in the image channel do not inhibit the real channel IF output. This technique does more than suppress the image signal — many intermodulation products can be eliminated by the scheme.

Intermodulation Products

Anaren's image rejection mixers use the 90° balanced, and the ORTHO-QUAD®, mixers because they provide the best amplitude and phase balance and, therefore, the best image rejection. The intermodulation performance of the unaided IRM is therefore the same as these two mixer types. Using a gating scheme similar to the one in Figure 33 used for image suppression, also allows suppression of many other intermodulation products.

Referring to Figure 31, we see that the LO signals appearing at the mixer inputs are in phase with each

other. Thus, *harmonics* of the LO signals are also in phase with each other (the relative phase of the n^{th} harmonic is n times the relative phase of the fundamental). The RF fundamentals, however, are 90° relative to each other. Thus the second, third and fourth harmonics are 180°, 270°, and 360° (0°) respectively.

The second and fourth harmonics of the RF, mixing with any harmonic of the LO (including the fundamental), produce mixer outputs which are not in phase quadrature, but are 180° and 0° respectively. When these signals are summed in the IF quadrature hybrid, they appear equally at the real (R) and image (I) ports. Since there is no difference between I and R, a positive V_0 is obtained, eliminating all these products. (In practice, R and I vary by a few dB, but normally not by the 10 dB threshold required to inhibit the output.) This gating scheme, when properly implemented, can eliminate the 1x2, 2x2, 3x2, 4x2, 1x4, 2x4, 3x4, and 4x4 intermodulation products. A more sophisticated gating scheme is reviewed in detail on page 173.

Bias Options

The catalog's biased IRM's allow performance under "starved" LO conditions. Biased IRM's using the optimum load line technique (see pages 99-103) can be supplied on special order and can provide optimum conversion loss and best intermodulation performance.

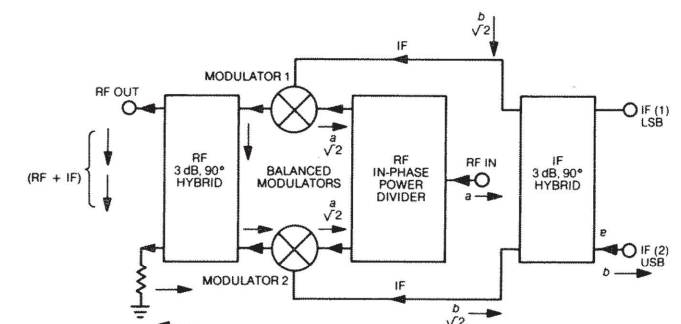
Single Sideband Modulators

A single sideband modulator (SSBM) accepts an RF (carrier) input and an IF input, and produces *either* the upper ($RF + IF$) or lower ($RF - IF$) sideband with the carrier and unwanted sideband suppressed. This is accomplished using the phase cancellation method of single sideband modulation.

Operation

This single sideband modulator consists of two balanced modulators and two hybrids, arranged as shown in Figure 34 (The block diagram is identical to the image rejection mixer shown in Figure 31, except that the balanced mixers must be balanced modulators.)

Figure 34 - Simplified Vector Operation of Single Sideband Modulator



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Although not shown by the vectors in Figure 34, the output of each modulator is a *double sideband* signal, the phase states changing at the IF (modulation) rate. When these double sideband signals are fed to the output quadrature hybrid, one sideband will sum to the RF out port with the other sideband suppressed. By changing the modulating signal from IF1 to IF2, either the lower or upper RF sideband can be selected. Thus it is possible to do sideband selection by IF switching between IF1 and IF2.

The vectors in Figure 34 follow just the upper (RF + IF) sidebands. The upper sideband sums to the RF output port when IF input 2 is driven. If input 1 was driven, the upper sideband would sum to the termination of the RF output hybrid, and the lower sideband (RF - IF) would be available at the RF out port.

Sideband and Carrier Suppression

The amount of unwanted sideband suppression depends on the vectors at the RF hybrid being phase and amplitude balanced, just as the image rejection mixer. (Figure 32 can be used to determine unwanted sideband suppression. As a function of phase and amplitude balance; relabel the vertical axis "sideband suppression.")

The degree of carrier suppression is directly related to the RF input/RF output isolation of the type of balanced modulator used. Anaren uses the balanced modulator versions of the ORTHO-QUAD® mixers as described on pages 106-108. These balanced modulators provide good isolation with the phase and amplitude balance necessary for good unwanted sideband suppression.

Applications

The SSBM is used whenever it is desired to generate a carrier signal having only one sideband. It could be used as the heart of a single sideband transmitter, for example. The more common use, however, is to offset a carrier by a precise frequency increment. The SSBM can offset a carrier either up (RF + IF) or down (RF -

IF) by simply changing IF input connections. The accuracy of the offset will be equal to the frequency stability of the modulation (IF) source.

Low-frequency Modulation

The modulatoin frequency is limited only by the availability of a quadrature drive to the balanced modulators. Aanaren supplies standard IF options from 20 MHz to 320 MHz. These options use an internal octave-band hybrid to provide the quadrature modulator drive.

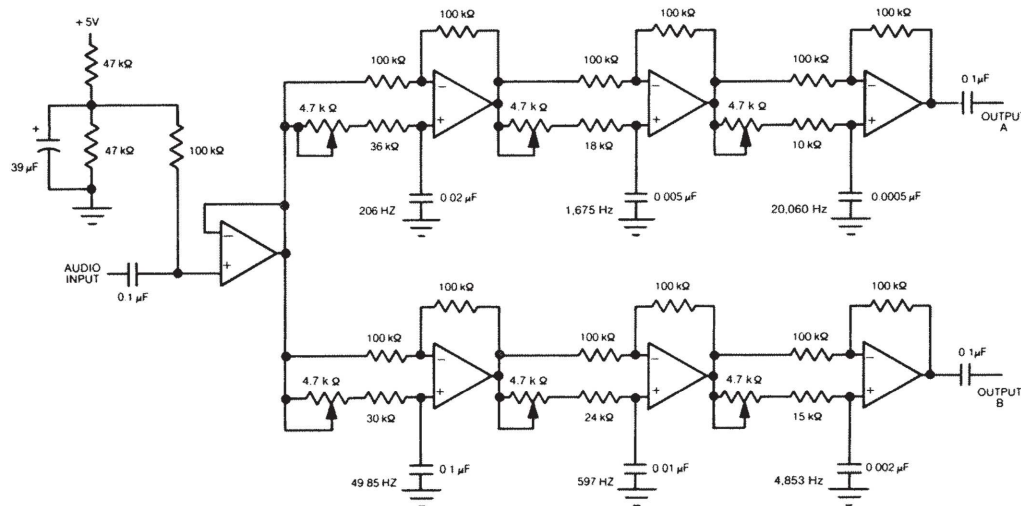
For modulation frequencies below 20 MHz Anaren offers the "—DC" suffixed models that have the modulator inputs brought directly to the IF ports. The customer must supply his own quadrature drive for these models. Lumped component (ferrite toroid) hybrids are available from a number of manufacturers and can be externally connected to the "—DC" IF ports. These devices will extend the IF modulation capability of the "—DC" versions down to about 1 MHz. Below this frequency either digital drive or phase-shift techniques are required.

Figure 36 (page 114) shows a phase-shift method for developing a quadrature drive from 100 Hz to 10 kHz. Any methods used must be able to deliver the equivalent of +10 dBm, or approximately 10 mA across 50 ohms, to the IF ports.

Intermodulation

The IF inputs are normally the high level signals. Their harmonics produce single-tone harmonic intermodulation products that limit the output power dynamic range available from the SSBM. Typically the RF ±3 IF response is only suppressed by 10 to 15 dB. However, the RF ±3 IF response can be improved significantly by making the RF carrier input the higher level signal and the modulation (IF) the lower level signal. This must be considered with care because transposing the power levels will effectively degrade the carrier suppression since this is related to the RF IN/RF OUT isolation. In the case where the modular has very high RF IN/RF OUT isolation the higher level carrier should present no problem.

Figure 35 - Quadrature Drive Outputs from Single Input



OP AMPS ARE LM324 QUADS
+ 5V TO PIN 4 EACH LM324, GND TO PIN 11

Quadrature Modulators

The quadrature (four-phase) modulator is designed for high data-rate digital communication systems. It can be used either in the biphase (BPSK) mode or the quadrature (QPSK) mode.

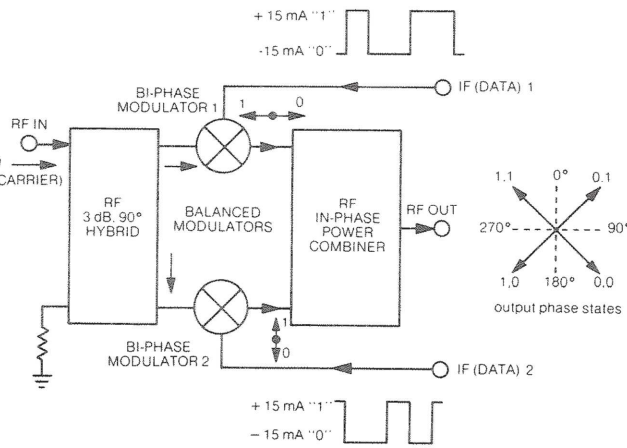
For BPSK operation the input carrier is modulated by identical data inputs at modulation (IF) ports 1 and 2 (the ports can be simply "teed" together). The data input normally consists of a two-level, bi-polar pulse code. The output carrier phase, relative to the input carrier, is phase shifted by 0 or 180° depending on the properties of the data inputs.

For QPSK operation the input carrier is modulated by two independent data channels at IF ports 1 and 2. Each data channel consists of a two-level, bi-polar pulse code. The output carrier phase, relative to the input carrier, is phase shifted 0, 90°, 180° or 270° depending on the properties of the data inputs.

Operation

Figure 36 shows the simplified vector and operation and block diagram of the quadrature modulator. (The block diagram is identical to the single sideband modulator except there is no IF hybrid.)

Figure 36 - Simplified Vector Operation of Quadrature (QPSK) Modulator



The heart of the circuit is the balanced modulator, used as bi-phase modulator (see page 125). The unmodulated carrier signal *a* is split into two carrier components, one 90° out of phase with respect to the other. One data input is used to modulate the 0° to 180° phases of the carrier (modulator 1); the other data input modulates the 90° to 270° phases of the carrier

(modulator 2). The modulator outputs are then combined by an in-phase power divider (combiner) to produce the four phases, 0°, 90°, 180°, and 270°.

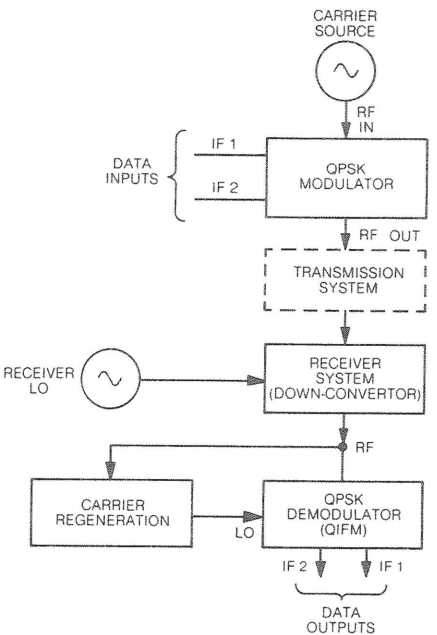
Applications

Digital communication systems are becoming more widely used because of their ability to transmit information over long distances without deterioration, since digital signals, unlike analog signals, can be accurately regenerated. Digital phase modulation (phase shift keying, PSK) allows only discrete phase angles, rather than continuous functions that occur with analog phase modulation. This instantaneous phase difference of the output carrier with respect to the input carrier determines the "digit" being transmitted.

PSK Transmission Systems

Figure 37 shows the elements of a typical QPSK transmission system. PSK systems usually have 2 discrete phrase states with two-phase (biphase, BPSK), and four-phase (quadrature, QPSK) being most common. BPSK generally carries one data stream and QPSK carries two data streams. These phase states are as widely separated as possible to reduce the noise sensitivity.

Figure 37 - Elements of Typical QPSK Digital Transmission System



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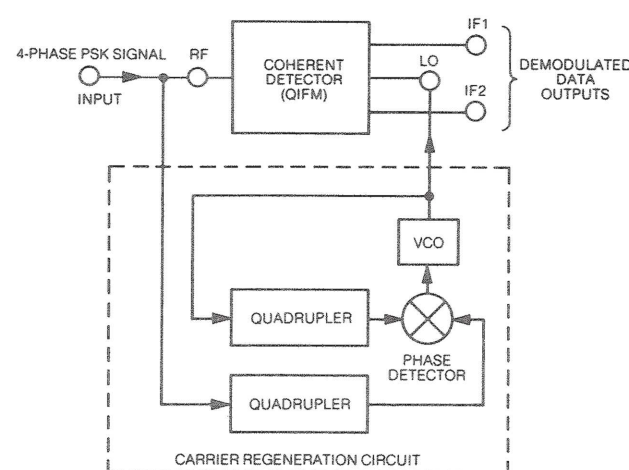
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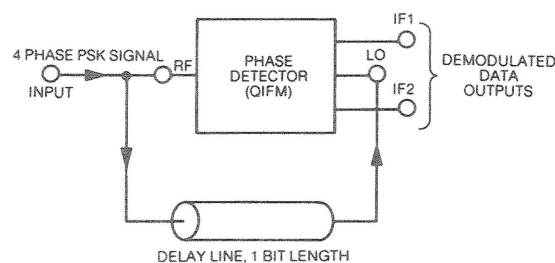
With PSK phase angles of 0° and 180° , 90° and 270° , there is theoretically a perfect carrier null. In order to detect this PSK signal at the demodulator the receiver must therefore regenerate the carrier to use as a coherent phase reference. This can be done by passing the BPSK signal through a frequency doubler and deriving a signal with 0° and 360° phase states. This is now a constant phase signal and it can be filtered and its frequency divided by two, thereby regenerating the original carrier frequency. Or, it can be used as an input to a phase comparator for controlling a phase-locked reference oscillator. The same method is used for regenerating the QPSK carrier except a frequency quadrupler is used. Figure 38 is a block diagram illustrating the coherent detection type of QPSK demodulator.

Figure 38 - Coherent Detection Type QPSK Demodulator



Another, simpler, type of QPSK demodulator is shown in Figure 39. It is called a differentially coherent or, delay, detector. The delay detector is most suitable for low-cost demodulators.

Figure 39 - Delay Detection Type QPSK Demodulator



The delay line is used as a memory for the preceding bit and provides a phase reference for the next symbol. Carrier phase *changes* from bit to bit are, therefore, detected without the need for establishing the fixed phase-lock reference of the carrier regeneration schemes. Either demodulator type uses the QIFM as the demodulating elements.

Quadrature Modulator Performance

Typical performance specifications for the Model 260015 Quadrphase Modulator are as follows:

IF (Data) Signal Levels:

Logic 1: +15 mA

Logic 0: -15 mA

Data Rate: 0 to 100 mB/s (NRZ Format)

Biphase Modulation:

Input Logic Level (IF Ports 1 and 2)	Relative Phase (Degrees)
1	$-(0 + \theta_0) \pm 5^\circ$
0	$-(180^\circ + \theta_0) \pm 5^\circ$

Note: θ_0 is the zero reference phase angle defined as input/output phase shift measured with a logic 1 level present at both IF ports.

Quadrature Modulation:

Input Logic Level		Relative Phase
F Port 1	IF Port 2	(Degrees)
1	1	$-(0 + \theta_0) \pm 5^\circ$
1	0	$-(270 + \theta_0) \pm 5^\circ$
0	0	$-(180 + \theta_0) \pm 5^\circ$
0	1	$-(90 + \theta_0) \pm 5^\circ$

Note: θ_0 is defined as 0° for biphasic modulation.

Input RF power Level: 0 dB (max)

Insertion Loss: 7 dB (typ.)

RF VSWR: 1.5 (typ.)

Isolation: IF to IF Ports: 25 dB (min)

IF to Output Ports: 25 dB (min)

The BPSK signal can be demodulated with any balanced or double balanced mixer covering the modulator RF range.

The QPSK signal can be demodulated with a quadrature IF mixer (QIFM) covering the modulator RF range.

Note that QPSK can give you two independent data channels without multiplexing; biphase only one.

References:

- (1) C.W. Gerst, "New Mixer Designs Boost D/F Performance," *Microwaves*, Vol. 12, No. 10, Oct. 1973. (Reprinted on next page.)
- (2) J.H. Lepoff and A.M. Cowley, "Improved Intermodulation Rejection in Mixers," *IEEE Transactions on Microwave Theory and Techniques* Vol. MTT-14, No. 12, Dec. 1966.
- (3) D. Cheadle, "Selecting Mixers for Best Intermodulation Performance *Microwaves*, Vol. 12, No. 11, Nov. 1973.
- (4) R.I. Disman, "Dynamic Range Performance of Microwave Transistor Amplifiers," *Microwave Journal*, Vol. 16, No. 8, Aug. 1971.
- (5) P. Will, "Reactive Loads - The Big Mixer Menace," *Microwaves*, Vol. 10, No. 4, April 1971.
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Mixer Designs for Phase/Amplitude tracking Subassemblies

The principal requirements being imposed on modern electronic warfare receivers today are wide-frequency coverage, spurious free dynamic range, small size and weight, and for direction-finding functions, extreme accuracies in phase and amplitude characteristics. Previously, broadband DF systems relied upon very accurate RF beam forming networks. As bandwidths have gone to multi-octave coverage, however, it has been necessary to down-convert the RF received signal and to perform beam forming functions at IF frequencies. The translation from RF to an IF frequency must be done, however, while preserving the RF phase codes. This approach has placed considerable stress on the mixer requirements. Specifically, those mixer specifications which are most critical for wide-band ECM receiving systems are:

- Biasing of the mixer—this bears on the LO dynamic range and spurious performance.
- Rejection of spurious signals—this is important when related to system false alarm and dynamic-range requirements.
- LO is RF isolation—important to radiated LO power requirements.
- VSWR—influences phase and amplitude tracking.
- Number of diodes used—this has a direct bearing on the LO power requirements.
- Focusing of image signal energy—critical to phase and amplitude tracking requirements.

The last requirement on the internally generated image signal, (produced by mixing of the LO second harmonic with the RF input (2 LO-RF)), introduces systems phase error. The image power must, therefore, be focused so it doesn't interfere with the generation of the desired IF output.

In order to get mixers to meet these requirements, proper selection of the dc load line is critical.

Most mixers are designed to operate at a constant bias voltage. However, through the use of a relatively simple resistive bias network, it is possible to establish an optimum dc load line, where diode conversion loss and VSWR can be optimized over an extended local oscillator dynamic range. In addition, spurious and intermodulation performance of a diode can be improved through the operation of the diode at high LO power levels while maintaining optimum conversion loss.

Four Mixer Types Analyzed

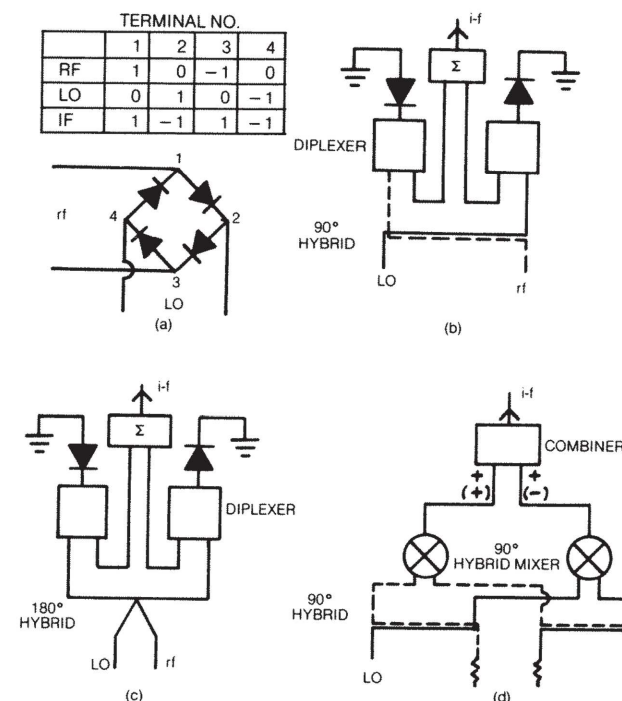
Figure 1 describes four basic mixer types that meet the aforementioned mixer requirements in varying degrees. For the double-balanced mixer in Fig. 1a, a fundamental diode quad is shown. Also shown is a sequence of phase codes or excitation required at each junction.

There are several circuit approaches available for achieving the proper RF and LO excitation and extracting the proper IF signal phase in a double-balanced mixer.

The ORTHO-QUAD^{RM} (quadrature-fed dual mixer), Fig. 1d, is a relatively new approach. In practice, a pair of mixers—either balanced or double-balanced types—are fed in phase quadrature. Only the quadrature-fed dual mixer with a pair of 90° hybrid mixers is considered here.

Table 1 gives a comparison of relative performance of these four mixer types based on the criteria established earlier. Ease of biasing is a direct function of circuit complexity. The 90-degree hybrid mixer and the 180-degree hybrid mixer, Fig. 1b and 1c are the easiest mixers to bias. All the mixers reject 2×2 spurious, none reject 3×3 spurious or two-tone intermodulation but no one mixer is optimum for all parameters.

Figure 1 - Four mixers considered are (a) double-balanced mixer; (b) 90-degree hybrid mixer; (c) 180-degree hybrid mixer; and (d) Ortho-Quad® (quadrature-fed dual mixer). Phase excitations for the double-balanced mixer are shown.



Combining Into a Mixer Subassembly

Let us now see which of these mixer types is most suited for an ECM tracking system. In a phase and amplitude tracking-receiving system, mixers can be arranged such as shown in Fig. 2. With this system, the phase and amplitude code which appears at the various RF ports is translated to the same phase and amplitude code at the corresponding IF ports. Note that the LO and BITE signal are injected through a single port.

The systems engineer can do broadband monopulse direction finding, or build up relatively large synthetic arrays using interferometric techniques with subassemblies of this type. The IF frequency of such systems can range from 60 MHz to 1 GHz or even extend into S-band.

A typical frequency breakdown of a broadband system, taking into account antenna bandwidths, LO power available, and high frequency conversion loss as a function of bandwidth, is a system broken into three bands—the lower band from 0.5-2 GHz, the middle band covering 2-12 GHz and the upper band covering 12-18 GHz. It is possible with today's technology to cover either 2-18 GHz, or even 0.5-1.8 GHz in a single system, but as the overall bandwidth becomes larger and larger, the conversion loss at 18 GHz begins to suffer.

Figure 2 - A phase and amplitude tracking mixer subassembly integrates several mixers, each of which must preserve the RF phase codes applied.

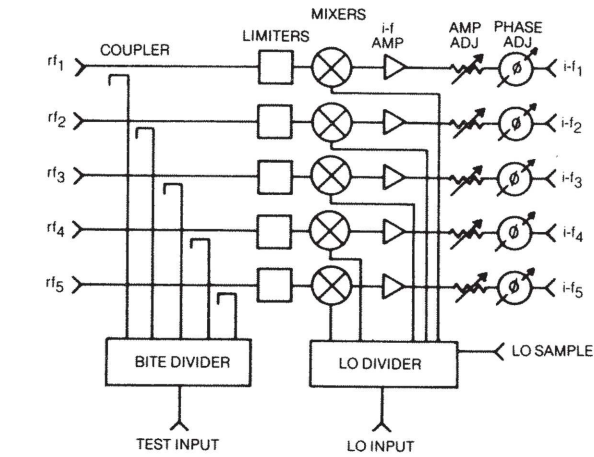


Table 1

	No. of Diodes	VSWR	LO-RF Isolation	Image Focusing Port	Bias ability	Spurious Rejection m-n = 1
90° Hybrid	2	Good	Poor	LO	1	Poor
180° Hybrid	2	Poor	Good	RF	1	Poor
Double-Balanced	4	Poor	Good	RF	3	Good
Ortho-Quad® (Q.F.D.M.)	4	Good	Good	Int. Load	2	Good

- a. VSWR: Poor: 2.5:1 typical
Good: 1.5:1 typical
b. LO-RF Isolation: Poor: 6 dB typical
Good: 20 dB typical
c. Image Focusing: Defines where image signal energy is focused, LO port, RF port, or internally terminated.

It is highly desirable, and almost imperative, that in a wideband multi-channel phase and amplitude tracking system of this type, the LO be operated on the same side of the RF for all real signals. That is, throughout the band, the real signal has to be consistently characterized as LO above the RF frequency or vice-versa. For discussion purposes, let us assume that the LO frequency is always above the RF frequency for the real channel. The phase of the nth IF channel is equal to the phase of the nth LO channel, minus the phase of the nth RF channel to the mixer,

$$\phi^n \text{ IF} = \phi^n \text{ LO} - \phi^n \text{ RF}$$

Ideally, the insertion phase from each RF input port to each mixer should be identical, and likewise, the insertion phase from the single LO terminal to each of the mixers should track. But in the case of a broadband multi-channel subassembly, such as in Fig. 2, the insertion phase from the LO input to a single mixer can be as large as 4 to 5 thousand degrees, and on the RF side, the insertion phase can be as great as two thousand degrees. Errors in the insertion phase produced by either inhomogeneties in the propagation constant (by a very small percentage) or slight constructional line length differences will produce time-delay errors. A time-delay error produces a phase error which varies linearly with frequency.

Let us now consider a 12 to 18 GHz system where this time-delay error causes a phase error of 1 degree per GHz, Fig. 3. For the real signal, the phase error is equal to 12 degrees at 12 GHz and increases linearly to 18 degrees at 18 GHz. At the image frequency, that is where the LO is below the RF, the phase error is simply reversed in sign. This error could be taken out in one of the three places; a time-delay change in the RF channel, a time-delay change in the LO channel or an IF phase shift. Time delay corrections take out the phase error at all LO frequencies and for both real and image RF frequencies. The problem with this approach is that in real life, a variable time delay will be accompanied by a VSWR addition. This VSWR, if it were the only one in the system, would not be a problem, but this mismatch will beat with the already existing VSWR's of the other components and will produce sinusoidal phase and amplitude error patterns as a function of frequency. Such errors are large enough to mask the good done by the time-delay correction.

A phase-shift at the IF port; however, will not affect the performance of the RF and LO paths as far as VSWR beats with frequency are concerned. It thus can

- d. Bias ability: Rated from 1 through 3
1. Easiest to bias
2. More difficult to bias
3. Most difficult to bias
Where |m-n| = 1
ie, 1x2, 2x1, 2x3, 3x2, etc.
3. m x n spurious rejection: Poor: Only partial rejection of |m-n| = 1 spurs
Good: Potentially rejects all |m-n| = 1 spurs

produce a reduction of phase error to zero at one LO frequency (real signal). In the case of the real and image frequency errors in Fig. 3, an IF phase shift of 15 degrees at the IF terminal will produce the phase plots shown in dashed lines. The real channel peak error has been reduced from 18 degrees to 3 degrees. In this case, the image channel peak error has been increased from 18 degrees to 33 degrees. This increase is immaterial if the LO is run consistently on the same side of the RF with only the real channel information being processed.

The biggest cause of amplitude and phase ripple in a subassembly (Fig. 2) is the error paths set up by the mismatch of one component beating the mismatch of another component. The flow diagram of Fig. 4 describes the phase and amplitude ripple of these mismatches:

$$\phi_{pp} = 2 \sin^{-1} \Gamma_1 \Gamma_2 \quad (2)$$

$$A_{pp} = 20 \log \frac{1 + \Gamma_1 \Gamma_2}{1 - \Gamma_1 \Gamma_2} \quad (3)$$

Figure 5 shows the phase and amplitude ripple possible from two mismatches on a 50-ohm line.

There are many possible combinations of error paths due to mismatches of an array of components making up a subassembly. Consider, for example, a hypothetical case where all components in Fig. 2 have a perfect match except the limiter and the RF port of the mixer. If a VSWR of 2.33:1 ($\Gamma = 0.4$) is assumed for both of these mismatches, then from fig. 5, it is possible to have peak-to-peak phase errors of 18.2 degrees and peak-to-peak amplitude errors of 2.8 dB ($\Gamma_1 \Gamma_2 = 0.16$).

Another source of phase and amplitude ripple which can be traced to a mixer is the internally generated image signal leaking from the RF port. This signal can be reflected back into the mixer by the mismatches of other components in the system, mix with the LO and convert to an IF frequency as an error component. Image generation of -12 dB relative to the RF input signal offers the same error path potential as the VSWR of 1.66:1. This image signal leaking from the LO port does not have the same error potential because it is suppressed by common mode rejection.

Figure 3 - Time-delay errors cause a phase error which increases linearly with frequency.

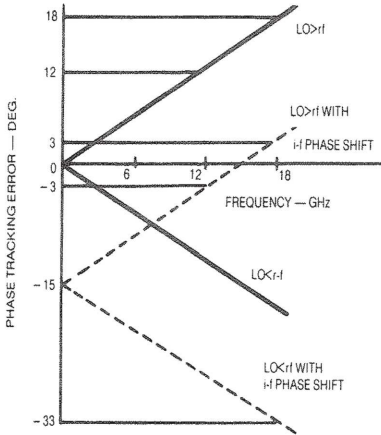


Figure 4 - Flow-graph generation of the transmission coefficient shows cumulative efforts of two mismatches separated by a length of transmission line.

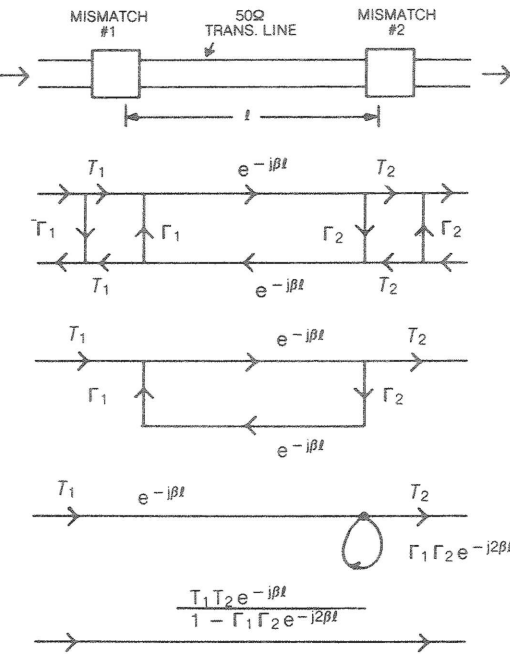
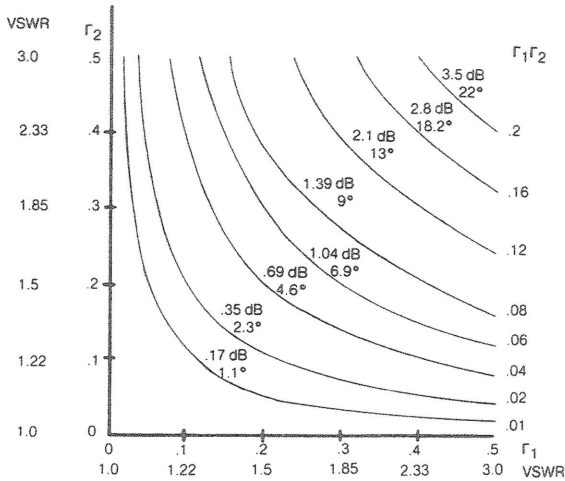


Figure 5 - Two mismatches widely separated by a matched transmission line produce widely varying peak-to-peak amplitude and phase ripple.



Isolation Important

As far as phase and amplitude tracking are concerned, the three parameters that are most important in a mixer are low RF VSWR, low LO VSWR, and image generation at the RF port. These restrictions rule out the double-balanced mixer, Fig. 1a, and the 180 degree hybrid mixer, Fig. 1c, as contenders for a broadband phase and amplitude tracking system. Both the 90 degree hybrid mixer, Fig. 1b and the Ortho-Quad® (quadrature-fed dual mixers), Fig. 1d, have good return loss and do not allow the image signal to come out the RF port. Cross-channel isolation is also very important if significant amplitude imbalance is expected at the RF ports.

It should exceed the greatest amplitude unbalance expected in the system by more than 25 dB to minimize the amplitude-to-phase cross talk. In a phase-only system, where the RF levels are all the same, cross-channel isolation will not produce significant errors.

There are four primary modes that can produce RF to IF cross-channel coupling, Fig. 6. First, is RF leaking through the first mixer across the LO divider and coming in on the LO port to mixer 2. It may then be down converted to the IF frequency at port IF₂. The cross-channel isolation is equal to the RF to LO isolation of the first mixer plus the isolation of the LO splitter at the RF frequency plus the common mode rejection of the second mixer.

The second type of leakage than can occur, Fig. 6b, is IF generated in the first mixer leaking to the LO port through the LO divider and coming in on the LO port of the second mixer. If this mode turns out to be prohibitive, it can be reduced by means of IF filtering on the LO divider.

A third mode of leakage, Fig. 6c, is the sum frequency (LO + RF) generated in the first mixer coming out the LO port, leaking through the LO divider and coming in on the LO port of the second mixer. It then mixes with two times the LO frequency in the second mixer and generates an error signal at the IF. The leakage path is generally not a problem unless the LO divider has no isolation at the sum frequency.

The fourth type of leakage is the image (2 LO - RF) generated in the first mixer leaking out the LO port across the LO divider and coming in on the LO port of the second mixer.

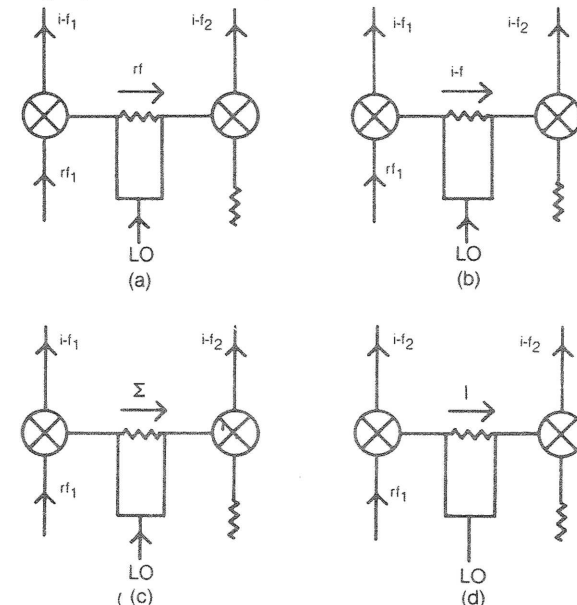
In practice, if the isolation of the LO power divider is maintained in-band and out-of-band, the primary sources of cross channel coupling are RF and image leakage across the LO divider.

The amount of LO power that is re-radiated from the RF terminals is important since it will be ultimately re-radiated out the antennas. The amount of LO power appearing at the RF ports of Fig. 2 is equal to the LO power required per mixer, minus the LO to RF isolation. Under starved LO conditions the typical LO power per mixer is 0 dBm. A 90 degree hybrid mixer affords about 10 dB of LO to RF isolation. The Ortho-Quad® (quadrature-fed dual) mixer provides approximately 25 dB of LO RF isolation. Hence, the LO power appearing at the RF port in Fig. 2 would be either minus 10 dBm or minus 25 dBm, depending upon the type of mixer used. Any further reduction must be made up with phase and amplitude tracking isolators between the RF ports and the antennas.

Mixer Parameters as a Function of LO Power

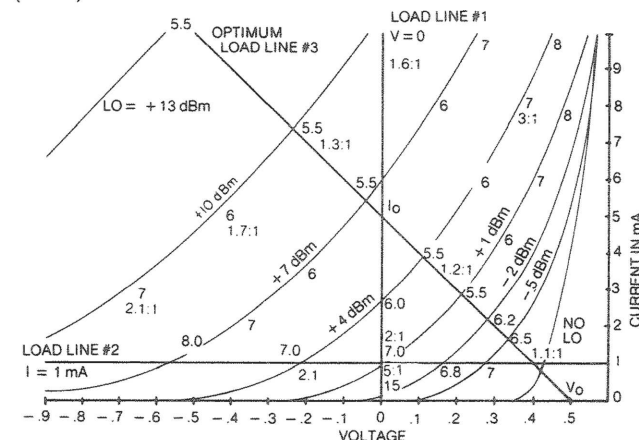
The conversion loss and spurious response of a diode depends on the impedance presented to the diode, not only at the RF and IF frequencies, but over the entire spectrum present in the Taylor's expansion of the diode current. In starved mixer operation, where insufficient local oscillator power is available and a mixer diode must be biased by a dc current the impedance presented to the diode at dc is most important.

Figure 6 - Four Possible Modes of RF to I-F Cross-channel coupling are (a) RF leakage; (b) IF leakage; (c) sum leakage ($\Sigma = LO + RF$) and (d) image leakage ($I = 2 LO - RF$).



The current-voltage characteristics of a hot-carrier diode varies as a function of LO power. This is due to rectification effects in the diode, Fig. 7. The circuit used to obtain these curves is also shown in Fig. 8. With a fixed LO power level applied, the dc voltage is varied, and the dc current through the diode measured. Consider now the response of the diode for three dc load lines, Fig. 7: □ Voltage = constant = 0 □ Current = constant = 1.0 mA. □ A linear load line through V = 0 at 5 mA and I = 0 at 0.5 V.

Figure 7 - Current Voltage Characteristics for a Schottky diode, HP 5082-2506, show diode non-linearities for various LO powers. Conversion loss figures are superimposed on the curves (black)—VSWR (color).



The first load line is for the unbiased diode—the case for the majority of mixers in use today. The second case is when the diode is biased for constant current. Here, the load line is optimized for very low LO power. The third case is a relatively new approach which optimizes both diode conversion loss and RF VSWR

over a wide local oscillator power dynamic range. This requires the use of an optimum dc load line and requires a biasing circuit as shown in Fig. 9.

If the available supply voltage is V_b , then the series resistor (R_1) is chosen such that:

$$R_1 = V_b / I_0 \quad (4)$$

where

I_0 is the dc current value at the intersection of the load line and current axis (Fig. 7).

The shunt resistance R_2 is then chosen to obtain the proper slope of the load line such that:

$$R_2 = V_0 / I \quad (5)$$

where

V_0 is the voltage value at the intersection of the load line and $I = 0$ axis of the diode.

Since there is no current through the diode at this point, the current I through R_2 is:

$$I = \frac{V_b - V_0}{R_1} \quad (6)$$

Combining Eq. (5) and (6), the shunt resistor value is then,

$$R_2 = \frac{V_0 R_1}{V_b - V_0} \quad (7)$$

Figure 8 - Circuit used to derive diode I-V characteristics in Fig. 7 also provides conversion loss.

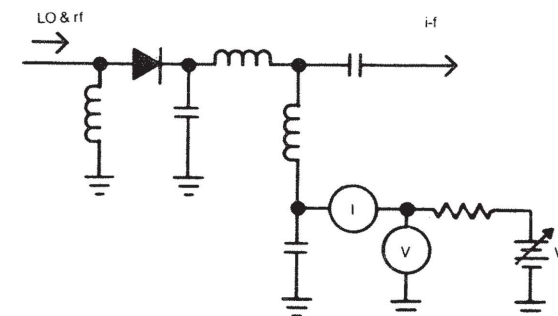
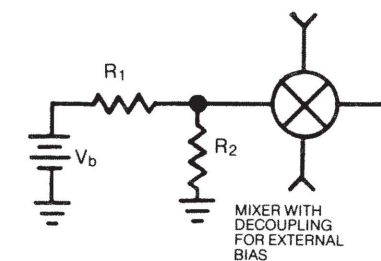


Figure 9 - This bias circuit is required to obtain the optimum dc loading of the mixer diodes.



To obtain the optimum load line of Fig. 7 with a bias supply of 10V; $V_0 = 0.5V$, $I_0 = 5$ mA then,

$$R_1 = \frac{10V}{5 \text{ mA}} = 2k \text{ ohm and}$$

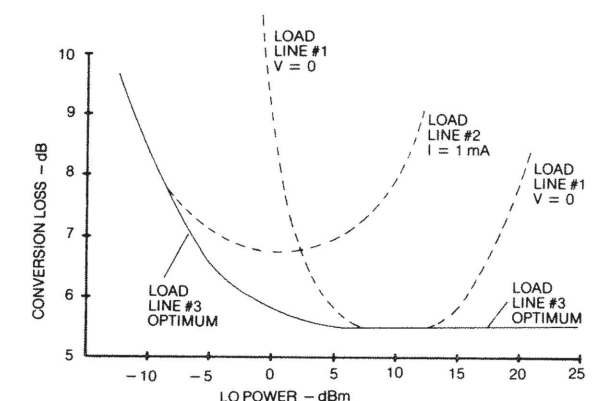
$$R_2 = \frac{(0.5)(2k)}{10-0.5} = 105 \text{ ohms}$$

Figure 10 is a direct plot of conversion loss as a function of LO power for the three dc load line conditions. As can be seen the third load line condition provides the best possible conversion loss over a 35 dB LO dynamic range. The $V = 0$ load line optimizes conversion loss at approximately +10 dBm LO power and the $I = 1.0$ mA load line has an optimum conversion loss only for LO powers below -7 dBm.

Spurious and intermodulation suppression can also be superimposed on the I-V curve as a function of LO power. Fortunately, a load line can be chosen to optimize spurious suppression, conversion loss and VSWR simultaneously. Conversion loss and two-tone suppression $[(2f_1 - f_2); f_1 \text{ and } f_2 \text{ at } -10 \text{ dBm}]$ are plotted in Fig. 11. At high LO powers, (i.e., about +16 dBm) a load line other than $V = \text{constant} = 0$ can be used to improve two-tone suppression with no degradation in conversion loss. For example, the load line in Fig. 11 gives 60 dB two-tone suppression with 5.5 dB conversion loss at +16 dBm. Using the $V = 0$ load, however, the optimum conversion loss point has less than 50 dB two-tone suppression. If the LO power is increased to improve the two-tone suppression, the conversion loss degrades by 2.5 dB.

The $m \times n$ spurious suppression can also be superimposed on the current-voltage curves in Fig. 11 in the same manner as the two-tone suppression. Best spurious performance results with relatively high LO power and by selecting the proper load line.

Figure 10 - By optimizing the load line at which a mixer diode operates, the dynamic range of a mixer can be improved to 35 dB.



A starved LO situation, which is frequently encountered in broadband electronic warfare systems, can produce serious dynamic range problems. The spurious suppression of a mixer is seriously degraded as the LO power is reduced. For example, a choice of mixers will not help the particular situation depicted in Fig. 12, since all of the mixers of Fig. 1 suppress the 2x2 and none of them suppress the 3x3. A gating technique, however, can be used when severe spurious generation limits the dynamic range.

A line graph showing Spurious Suppression (in dB) on the y-axis versus LO Power of Mixer (in dBm) on the x-axis. The y-axis ranges from 30 to 76 dB with major ticks every 10 dB. The x-axis ranges from -10 to 10 dBm with major ticks every 5 dBm. Two lines are plotted: a solid line labeled '2x2' and a dashed line labeled '3x3'. The '2x2' line starts at approximately (-10, 40) and ends at (10, 54). The '3x3' line starts at approximately (-10, 33) and ends at (10, 76). The two lines intersect at approximately (-2, 45).

LO Power of Mixer (dBm)	2x2 Spurious Suppression (dB)	3x3 Spurious Suppression (dB)
-10	40	33
-5	43	40
0	46	47
5	49	54
10	54	76

$$mLO - nRF = +IF \quad (8)$$

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Diagram illustrating the Real Channel IF Receiver architecture. The receiver processes the RF signal through an Image Rejection Mixer, a Diplexer, and various filters and amplifiers to produce the Real IF, Image, and Gate outputs. The diagram includes components like the 2x2 AND 3x3 gate, SUMMER, BANDPASS FILTER, and DETECTOR. The output is labeled as REAL IF, IMAGE, and GATES (WILL INHIBIT REAL CHANNEL IF ACTIVATED).

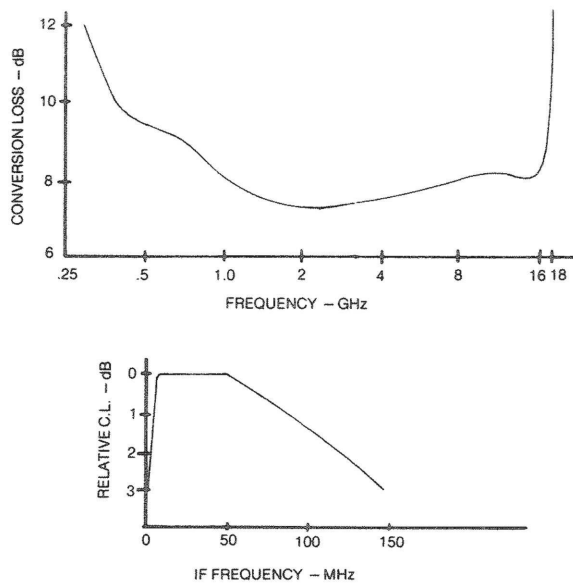
Figure 15 - This spectrum occurs with an image rejection mixer after passing through a narrowband IF filter as the LO is swept over a wide-frequency range.

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The third gate works on the spurious signals with wide LO to RF separation. For a given spurious, such as the 4x3 spur, the range of difference in frequencies between the LO and RF over which this spur can be generated is relatively small. The 3x2 and 2x3 spurs are generated in the vicinity of $LO-RF = 0.5$ of RF min. The 4x3 and 3x4 are generated in the vicinity of $LO-RF = 0.33$ RF min. The 4x3 and 3x4 are generated in the vicinity of $LO-RF = 0.2$ RF min. The high IF ($|m-n| = 1$) port of the mixer can be filtered for one or more of these spurious and followed by a detector. Pre-amplification should not be needed in front of the detector since these signals are higher than the real channel spurious by the amount of the spurious rejection. The output of this detector is then compared to the real channel detector. If the detector output of the $|m-n| = 1$ channel is greater than the real detector, this gate will be activated and the real channel will be inhibited because of the possibility of a spur whose absolute value of $m-n$ is equal to 1.

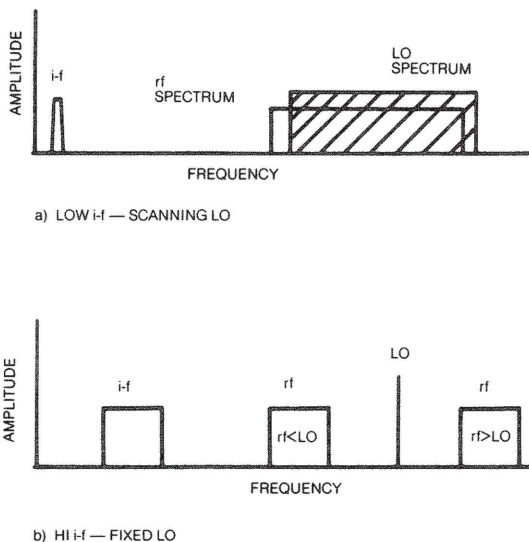
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Figure 16 - A 90-degree hybrid mixer can be designed to cover six octaves or over a 60:1 bandwidth.



The second trend is in high IF mixers. A frequency spectrum comparison between the low IF scanned LO system and the high IF fixed LO system is made in Fig. 17. In a low IF system, the LO is swept through the RF band in a search mode such that IF is always in the center of the IF pass band. When this occurs, the RF frequency is equal to the LO frequency, minus the IF. Circuitry must be provided for rejecting images and 2x2 and 3x3 intermodulation products as well as 2x1 and 3x2 higher-order spurs. In the high IF system, the image and the 2x2 and 3x3 lower-order intermodulation products are not a problem. The LO is a fixed frequency and the IF is typically an octave wide. This means that an RF band—either below or above the LO frequency—can be covered but preselection is required to separate the RF band.

Figure 17 - This spectrum shows the differences between a low IF scanned LO system and a high IF fixed LO system.



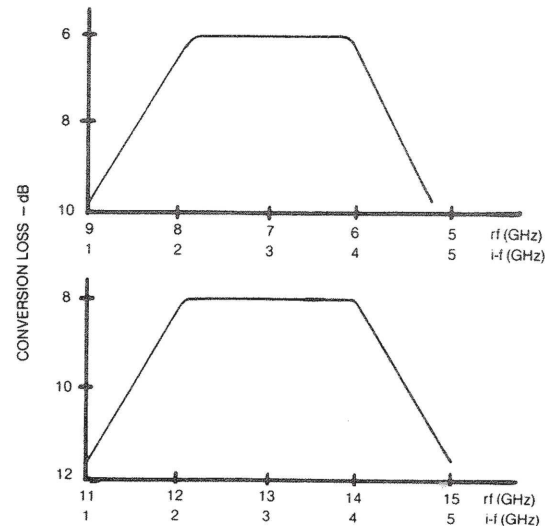
The two major advantages of a high IF system are

- Less complicated LO system
- Elimination of image and 2x2 and 3x3 spurious response.

The major disadvantage is a more complicated IF processor because of the increase of both absolute and percentage bandwidth.

The conversion loss of a mixer with a 10 GHz LO and 2-4 GHz IF is shown in Fig. 18.

Figure 18 - Conversion loss of a high IF mixer with an IF of 2-4 GHz and an LO of 10 GHz.



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