CHAPTER 12.6
FREQUENCY CONVERTERS AND DETECTORS

Glenn B. Gawler

GENERAL CONSIDERATIONS OF FREQUENCY CONVERTERS

A frequency converter usually consists of an oscillator (called a *local oscillator* or LO) and a device used as a mixer. The mixing device is either nonlinear or its transfer parameter can be made to vary in synchronism with the local oscillator. A signal voltage with information in a frequency band centered at frequency f_s enters the frequency converter, and the information is reproduced in the intermediate-frequency (i.f.) voltage leaving the converter. If the local-oscillator frequency is designated f_{LO}, the i.f. voltage information is centered about a frequency $f_{if} = f_{LO} \pm f_s$. The situation is shown pictorially in Fig. 12.6.1. Characteristics of interest for design in systems using frequency converters are gain, noise figure, image rejection, spurious responses, intermodulation and cross-modulation capability, desensitization, area local-oscillator to rf, and to i.f. isolation. These characteristics will be discussed at length in the descriptions of different types of frequency-converter mixers and their uses in various systems. First, explanations are in order for the above terms.

Frequency-Converter Gain. The available power gain of a frequency converter is the ratio of power available from the i.f. port to the power available at the signal port. Similar definitions apply for transducer gain and power gain.

Noise Figure of Frequency Converter. The noise factor is the ratio of noise power available at the i.f. port to the noise power available at the i.f. port because of the source alone at the signal port.

Image Rejection. For difference mixing $f_{if} = f_{LO} - f_s$ and the image is $2f_{LO} - f_s$. For sum mixing $f_{if} = f_{LO} + f_s$, and the image is $2f_{LO} + f_s$. An undesired signal at the difference mixing frequency $2f_{LO} - f_s$ results in energy at the i.f. port. This condition is called *image response* and attenuation of the image response is image rejection, measured in decibels.

Spurious Responses. Spurious external signals reach the mixer and result in generation of undesired frequencies that may fall into the intermediate-frequency band. The condition for an interference in the i.f. band is

$$mf_s' \pm nf_1 = \pm f_{if}$$

where m and n are integers and f_s' represents spurious frequencies at the signal port of the mixer.

Example. There is a strong local station in the broadcast band at 810 kHz and a weak distant station at 580 kHz. A receiver is tuned to the distant station, and a whistle, or beat, at 5 kHz is heard on the receiver (refer to Fig. 12.6.2).

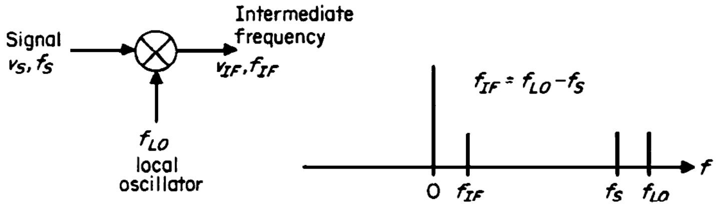

FIGURE 12.6.1 Frequency-converter terminals and spectrum.

An analysis shows that the second harmonic of the local oscillator interacts with the second harmonic of the 810-kHz signal to produce a mixer output at 450 kHz in the i.f. band of the receiver:

$$580 + 455 = 1035 \text{ kHz} = \text{LO frequency}$$
$$2 \times 1035 - 2 \times 810 = 450 \text{ kHz} = \text{i.f. interference frequency}$$

The interference at 450 kHz then mixes with the 455-kHz desired signal in the second detector to produce the 5-kHz whistle. Notice that if the receiver is slightly detuned upward by 5 kHz, the whistle will zero-beat. Further upward detuning will create a whistle of increasing frequency.

INTERMODULATION

Intermodulation is particularly troublesome because a pair of strong signals that pass through a receiver preselector can cause interference in the i.f. passband, even though the strong signals themselves do not enter the passband.

Consider two undesired signals at 97 MHz passing through a superheterodyne receiver tuned to 100 MHz. Suppose, further, that the i.f. is so selective that a perfect mixer allows no response to the signals (see Fig. 12.6.3). Third-order intermodulation in a physically realizable mixer will result in interfering signals at the i.f. frequency and 9 MHz away (corresponding to 100 and 91 MHz rf frequencies, respectively). Fifth-order intermodulation will produce interferences 3 and 12 MHz from the intermediate frequency (103 and 88 MHz rf frequencies).

There is a formula for variation of intermodulation products that is quite useful. Figure 12.6.4 shows typical variations of desired output and intermodulation with input power level. Desired output increases 1 dB for each 1-dB increase of input level, whereas third-order intermodulation increases 3 dB for each 1-dB increase of input level. At some point the mixer saturates and the above behavior no longer obtains. Since the interference of the intermodulation product is primarily of interest near the system sensitivity limit (usually somewhere below –20 dBm), the 1 dB per 1 dB and 3 dB per 1 dB patterns hold. The formula can be written

$$P_{21} = 2P_N + P_F - 2P_{I21}$$

where P_{21} = level of intermodulation product (dBm)
P_N = power level of interfering signal nearest P_{21}
P_F = power interfering signal farthest from P_{21}

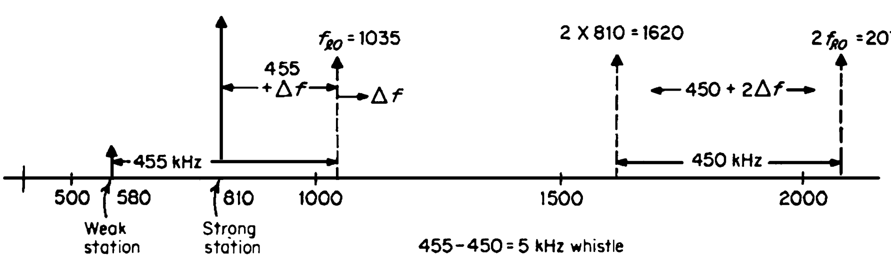

FIGURE 12.6.2 Spurious response in AM receiver.

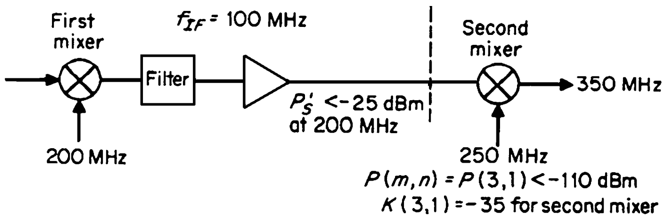

FIGURE 12.6.3 Spurious-response analysis.

P_{I21} is the third-order intercept power. For proper orientation, $f_N = 97$ MHz, $f_F = 94$ MHz, $f_{21} = 100$ MHz in Fig. 12.6.5. The intercept power is a function of frequency. It can be used for comparisons between mixer designs and for determining allowable preselector gain in a receiving system.

FREQUENCY-CONVERTER ISOLATION

There are two paths in a mixer where isolation is important. The so-called *balanced mixers* give some isolation of the local-oscillator energy at the rf port. This keeps the superheterodyne receiver from radiating excessively. The doubly balanced mixers also give rf-to-i.f. isolation. This keeps interference in the receiver rf environment from penetrating the mixer directly at the i.f. frequency. Less important, but still significant, is the LO-to-i.f. isolation. This keeps LO energy from overloading the i.f. amplifier. Also, in multiple-conversion receivers low LO-to-i.f. leakage minimizes spurious responses in subsequent frequency converters.

Desensitization

A strong signal in the rf bandwidth, not directly converted to i.f., drives the operating point of the mixer into a nonlinear region. The mixer gain is then either decreased or increased. In radar, the characteristic of concern is pulse desensitization. In television receivers the characteristic is called cross-modulation. Here the strong

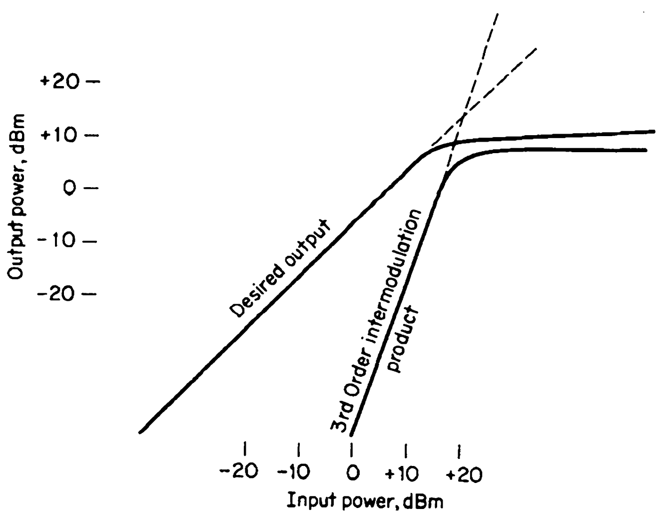

FIGURE 12.6.4 Third-order intermodulation intercept power.

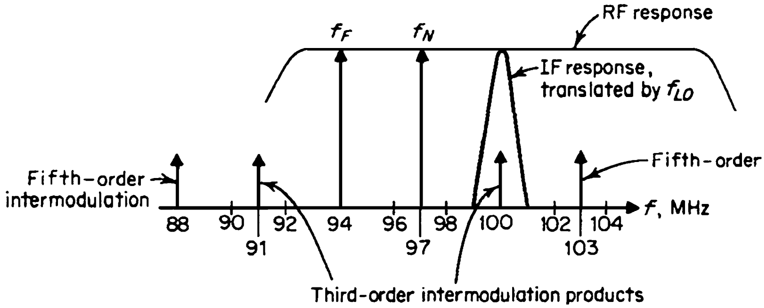

FIGURE 12.6.5 Intermodulation in a superheterodyne receiver.

undesired adjacent TV station modulates the mixer gain, especially during synchronization intervals, where the signal is strongest. The result appears in the desired signal as a contrast modulation of picture with the pattern of the undesired sync periods, corresponding to mixer gain *pumping* by the strong adjacent channel.

SCHOTTKY DIODE MIXERS

The Schottky barrier diode is an improvement over the point-contact diode. The Schottky diode has two features that make it very valuable in high-frequency mixers: (1) it has low series resistance and virtually no charge storage, which results in low conversion loss; (2) it has noise-temperature ratio very close to unity. The noise factor of a mixed-i.f. amplifier cascade is

$$F = L_M(t_D + F_{if} - 1)$$

where L_M = mixer loss
t_D = diode noise-temperature ratio
F_{if} = i.f. noise factor

Since t_D is near unity and L_M is in the range of 2.4 to 6 dB, overall noise factor is quite good, with F_{if} near 1.5 dB in well-designed systems.

The complete conversion matrix involves LO harmonic sums and differences, as well as signal, i.f., and image frequencies. They restrict their treatment of crystal rectifiers to the third-order matrix

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3^* \end{bmatrix} = [Y] \begin{bmatrix} V_1 \\ V_2 \\ V_3^* \end{bmatrix} \quad Y = \begin{bmatrix} y_{11} & y_{12} & y_{13} \\ y_{21} & y_{22} & y_{23} \\ y_{31} & y_{32} & y_{33} \end{bmatrix}$$

where 1 denotes signal port; 2, i.f. port; and 3, image port.

With point-contact diodes, the series resistance is so large that not much improvement is realized by terminating the image frequency, and terminating the other frequencies involved is less significant.

With the advent of Schottky barrier diodes, which have much smaller series resistances, proper termination of pertinent frequencies, other than signal and i.f. frequencies, results in a minimizing of conversion loss. This, in turn, leads to a minimizing of noise figure.

Several different configurations are used with Schottky mixers. Figure 12.6.6 shows an image-rejection mixer, which is used for low i.f. frequency systems where rf filtering of the image is impractical.

There is a general rule of thumb for obtaining good intermodulation, cross-modulation, and desensitizable performance in mixers. It has been found experimentally that pumping a mixer harder extends its range of

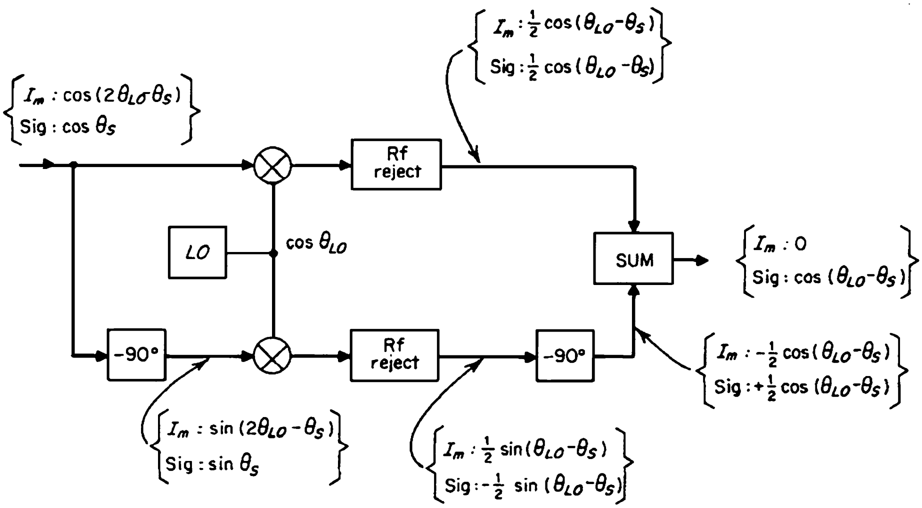

FIGURE 12.6.6 Mixer designed for image rejection.

linear operation. The point-contact diode had a rapidly increasing noise figure with high LO power level and could easily burn out with too much power. The Schottky diode, however, degrades in noise figure relatively slowly with increasing LO power, and it can tolerate quite large amounts of power without burnout. There is a limit to this process of increasing LO power: the Schottky diode series resistance begins to appear nonlinear. This leads to another rule of thumb: pump the diode between two linear regions, and spend as little time as possible in the transition region. Application of these two rules leads to the doubly balanced Schottky mixer. The reason for this is that one pair of diodes conducts hard and holds the other pair off. Hence large LO power is required, and one diode pair is conducting well into its linear region while the other diode pair is held in its linear nonconducting region.

DOUBLY BALANCED MIXERS

The diode doubly balanced mixer, or ring modulator, is shown in Fig. 12.6.7. The doubly balanced mixer is used up to and beyond 1 GHz in this configuration. The noise-figure optimization process previously discussed applies to this type of mixer. It exhibits good LO-to-rf and LO-to-i.f. isolation, as shown in Fig. 12.6.8. Typical published data on mixers quote 30 dB rf-to-i.f. isolation below 50 MHz and 20 dB isolation from 50 to 500 MHz.

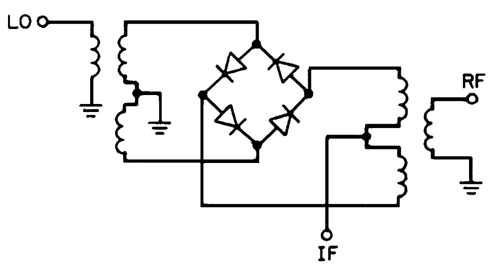

FIGURE 12.6.7 Doubly balanced mixer.

Another feature of balanced mixers is their LO noise-suppression capability. Modern mixers using Schottky diodes in a ring modulator provide somewhat better LO noise suppression. The ring modulator provides suppression only to AM LO noise, not FM noise.

PARAMETRIC CONVERTERS

Parametric converters make use of time-varying energy-storage elements. Their operation is in many ways similar to that of parametric amplifiers (see Section 11). The difference is that output and input frequencies are the same in parametric amplifiers, while the frequencies differ in parametric converters. The device most widely

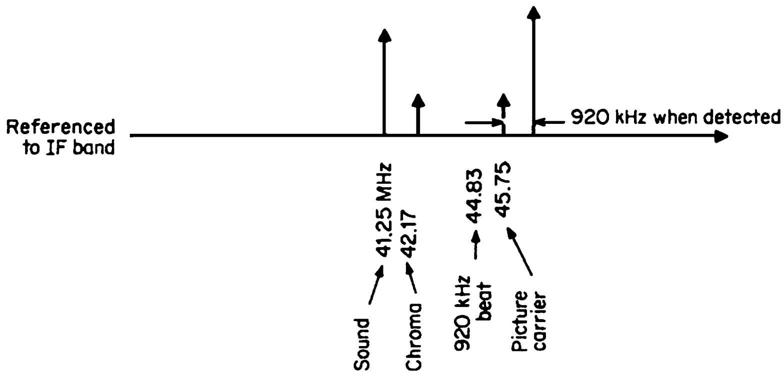

FIGURE 12.6.8 Generation of 920-kHz beat in TV tuners.

used for microwave parametric converters today is the varactor diode, which has a voltage-dependent junction capacitance. The time variation of varactor capacitance is provided by a local oscillator, usually called the *pump*.

Attainable gain of a parametric converter is limited by the ratio of output to input frequencies. Therefore up conversion is generally used to achieve some gain. Because lower-sideband up conversion results in negative resistance, the upper sideband is generally used. This results in simpler circuit elements to achieve stability.

There is a distinct advantage to up conversion; image rejection is easily achievable by a simple low-pass filter.

TRANSISTOR MIXERS

One of the original concerns in transistor mixers was their noise performance. The base spreading resistance r_b is very important in noise performance. The reason is that mixing occurs across the base-emitter junction; then the i.f. signal is amplified by transistor action; however, r_b is a lossy part of the termination at the i.f. signal, image, and all other frequencies present in the mixing process. Hence r_b dissipates some energy at each of the frequencies present, and all these contributions add to appear as a loss in the signal-to-i.f. conversion. This loss, in turn, degrades noise figure.

Manufacturers do not promote transistors used as mixers, probably because of their intermodulation and spurious-response performance. Estimates of intermodulation intercept power go as high as +12 dBm, while one measurement gave +5 dBm at 200 MHz; however, a cascade transistor mixer is used in a commercial VHF television tuner.

MEASUREMENT OF SPURIOUS RESPONSES

Figure 12.6.9 shows an arrangement for measuring mixer spurious responses. The filter following the signal generator implies that generator harmonics are down, say 40 dB. This ensures that frequency-multiplying action is owing only to the mixer under test. The attenuator following the mixer can be used to be sure that a spurious response of the receiver is not being measured. That is, a 6-dB change in attenuator setting should be accompanied by a 6-dB change on the indicator.

Generally the most convenient way of performing the spurious-response test is first to obtain an indication of the indicator. Then tune the signal generator to the desired frequency and record the level required to obtain the original response. This should be repeated at one or two more levels of the undesired signal to ensure that the spur follows the appropriate laws. For example, if the response is fourth-order (four times the signal frequency $\pm n$

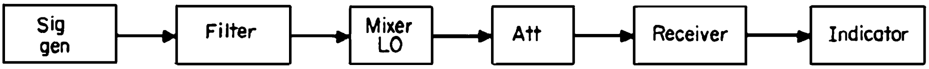

FIGURE 12.6.9 Test equipment for measuring mixer spurious responses.

times the LO frequency), the measured value should change 4 dB for a 1-dB change in undesired frequency level. The order of the spurious response can be determined by either of two methods. The first method is simply by knowing with some accuracy the undesired signal frequency and the LO frequency and then determining the harmonic numbers required to obtain the i.f. frequency. The other technique entails observing the incremental changes of the i.f. frequency with known changes in the undesired signal frequency and the LO signal frequency.

This completes the measurement for one spurious response. The procedure should be repeated for each of the spurious responses to be measured.

The intermodulation test setup is shown in Fig. 12.6.10. In general, a diplexer is preferable to a directional coupler for keeping generator 1 signal out of generator 2. This is necessary so that the measurement is not limited by the test setup. A good idea would be to establish that no third-order intermodulation occurs becuase of the setup alone. To do this, initially remove the mixer-LO circuit. Then tune generator 1 off from center frequency to about 10 or 20 dB down on the skirt of the receiver preselector. Tune generator 2 twice this amount from the receiver center frequency. Set generator levels equal and at some initial value, say –30 dBm. Then vary one generator frequency slightly and look for a response peak on the indicator. If none is noticed, increase the generator level to –20 dBm and repeat the procedure. Usually, except for very good receivers, the third-order intermodulation response is found. Vary the attenuator by 6 dB, and look for a 6-dB variation in the indicator reading. If the latter is not 6 dB but 18 dB, intermodulation is occurring in the receiver. If the indicator variation is between 6 and 18 dB, intermodulation is occurring in the circuitry preceding the attenuator and in the receiver. To obtain trustworthy measurements with a mixer in the test position, the indicator should read at least 20 dB greater than without the mixer, while the generator levels should be lower by mixer gain +10 dB than they were without the mixer. This ensures that the test setup is contributing an insignificant amount to the intermodulation measurement.

With the mixer in test position and the above conditions satisfied, obtain a reading on the indicator and let the power referred to the mixer input be denoted by P (dBm). Turn down both generator levels, and retune generator 1 to center frequency. Adjust generator 1 level to obtain the previous indicator reading. This essentially calibrates the measurement setup. Denote generator 1 level referred to the mixer input by P_{21} dBm. Then the intermodulation intercept power is given by

$$P_{I21}\ \text{dBm} = (3P - P_{21})/2$$

The subscripts on intercept power P_{I21} refer to second order for the near frequency and first order for the far frequency (see Fig. 12.6.4).

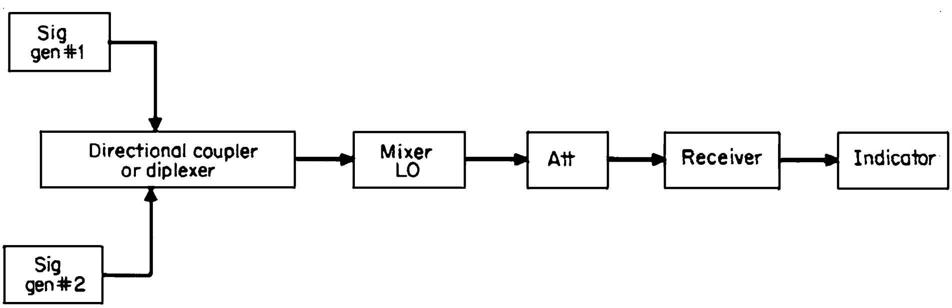

FIGURE 12.6.10 Test equivalent for measurement of mixer intermodulation.

The procedure should be repeated for one or two lower values of P. The corresponding values of $P_{/21}$ should asymptomatically approach a constant value. The constant value of $P_{/21}$ so obtained is then a valid number for predicting behavior of the mixer near its sensitivity limit.

DETECTORS (FREQUENCY DECONVERTERS)

Detectors have become more complex and versatile since the advent of integrated circuits. Up to the mid-1950s most radio receivers used the standard single-diode envelope detector for AM and a Foster-Seeley discriminator or ratio detector for FM. Today, integrated circuits are available with i.f. amplifier, detector, and audio-amplifier functions in a single package.

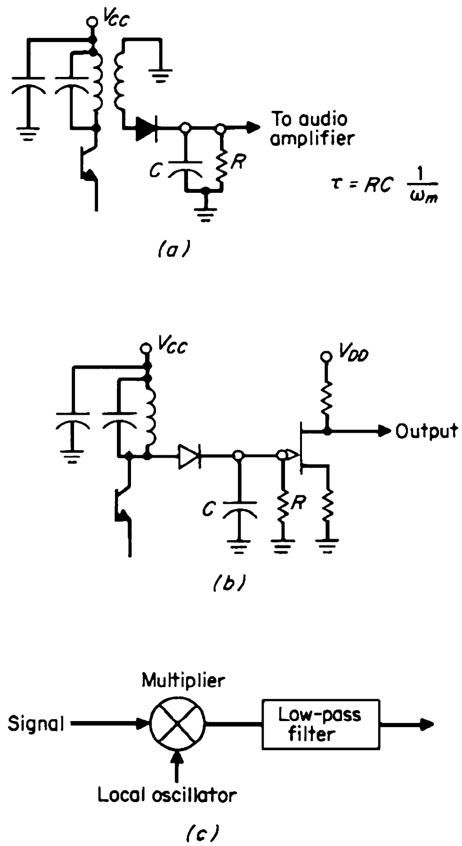

FIGURE 12.6.11 AM detectors: (*a*) AM envelope detector; (*b*) peak detector; (*c*) product detector.

Figure 12.6.11 shows three conventional AM detectors. In Fig. 12.6.11*a* an envelope detector is shown. In order for the detected output to follow the modulation envelope faithfully, the RC time constant must be chosen so that $RC < 1/\omega_m$, where ω_m is the maximum angular modulation frequency in the envelope. Figure 12.6.11*b* shows a peak detector. Here the RC time constant is chosen large, so that C stays charged to the peak voltage. Usually, the time constant depends on the application. In a television field-strength meter, the charge on C should not decay significantly between horizontal sync pulses separated by 62.5 μs. Hence a time constant of 1 to 6 ms should suffice. On the other hand, an AGC detector for single-sideband use should have a time constant of 1 s or longer.

Figure 12.6.11*c* shows a product (synchronous) detector. This type of detector has been used since the advent of single-sideband transmission. The product detector multiplies the signal with the LO, or beat frequency oscillator (BFO), to produce outputs at sum and difference frequencies. Then the low-pass filter passes only the difference frequency. The result is a clean demodulation with a minimum of distortion for single-sideband signals.

The two classical FM detectors widely used up to the present are the Foster-Seeley discriminator and the ratio detector. Figure 12.6.12 shows the Foster-Seeley discriminator and its phasor diagrams. The circuit consists of a double-tuned transformer, with primary and secondary voltages series-connected. The diode connected to point A detects the peak value of $V_1 + V_2/2$, and the diode at B detects the peak value of $V_1 - V_2/2$. The audio output is then the difference between the detected voltages. When the incoming frequency is in the center of the passband, V_2 is in quadrature with V_1, the detected voltages are equal, and audio output is zero. Below the center frequency the detected voltage from B decreases, while that from A increases, and the audio output is positive. By similar reasoning, an incoming frequency above band center produces a negative audio output. Optimum linearity requires that $KQ = 2$, where K is the transformer coupling and Q is the primary and secondary quality factor.

Figure 12.6.12*c* shows a ratio detector, which has an advantage over the Foster-Seeley discriminator in being relatively insensitive to AM. The ratio detector uses a tertiary winding (winding 3) instead of the primary voltage, and one diode is reversed; however, the phasor diagrams also apply to the ratio detector. The AM

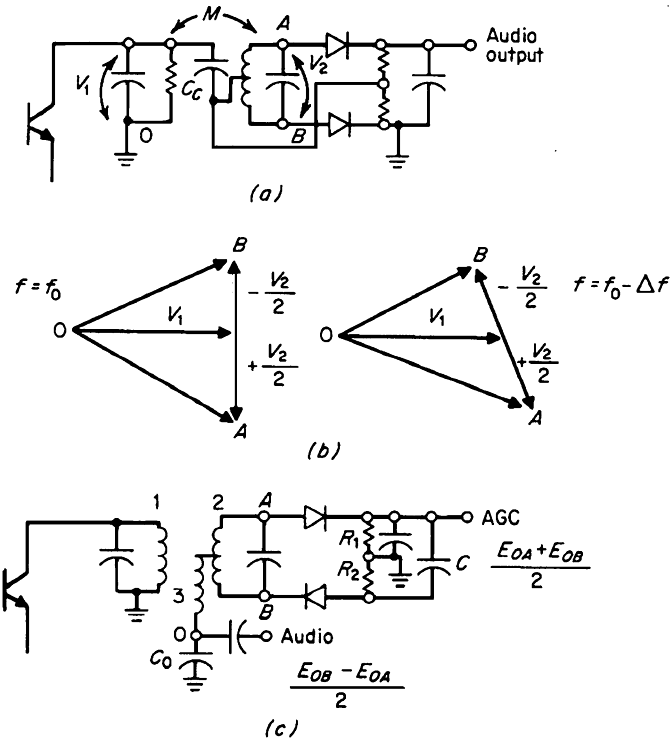

FIGURE 12.6.12 FM detectors: (*a*) Foster-Seeley FM discriminator; (*b*) phasor diagrams; (*c*) ratio detector.

rejection feature results from choosing the $(R_1 + R_2)C$ time constant large compared with the lowest frequency to be faithfully reproduced. The voltages E_{OA} and E_{OB} represent the detected values of rf voltages across *OA* and *OB*, respectively. With the large time constant above, voltage on *C* changes slowly with AM and the conduction angles of the diodes vary, loading the tuned circuit so as to keep the rf amplitudes relatively constant.

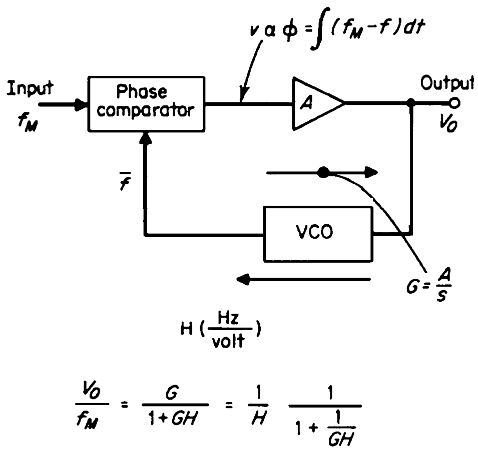

FIGURE 12.6.13 FM detector using phase-locked loop.

Capacitor C_0 is chosen to be an rf short circuit but small enough to follow the required audio variations. In the AM rejection process, AF voltage on C_0 does not follow the AM because the charge put on by one diode is removed by the other diode. With FM variations on the rf, voltage on C_0 changes to reach the condition, again, that charge put on C_0 by one diode is removed by the other diode. The ratio detector is generally used with little or no previous limiting of the rf, while the Foster-Seeley discriminator must be preceded by limiters to provide AM rejection.

With the recent trend toward integrated circuits, there has been increased interest in using phase-locked loops and product detectors. These techniques have been selected because they do not require inductors, which are not readily available in integrated form. Figure 12.6.13 shows a phase-locked loop (PLL) as an

FM detector. The phase comparator merely provides a dc voltage proportional to the difference in phase between signals represented by f_M and f. Initially, f and f_M are unequal, but because of high loop gain, $GH >> 1$, f and f_M quickly become locked and stay locked. Then as f_M varies, f follows exactly. But because of the high loop gain, response is essentially $1/H$, which is the voltage-controlled oscillator (VCO) characteristic. Hence the PLL serves as an FM detector.

AM product detectors also make use of the PLL to provide a carrier locked to the incoming signal carrier. The output of the VCO is used to drive the product detector. Probably one of the most stringent uses of the product detector is in an FM stereo decoder. The *left minus right* (L – R) subcarrier is located at 38 kHz with sidebands from 23 to 53 kHz. There may also be an SCA signal centered about 67 kHz which is used to provide a music service for restaurants and commercial offices. The L – R product detector is driven by a 38-kHz VCO, the output of which also goes to a 2-to-1 counter. The counter output is compared with the 19-kHz pilot signal in a phase comparator, and the phase-comparator output then controls the VCO. Because of the relatively small pilot signal and the presence of L + R, L – R, and SCA information, the requirement for phase locking is stringent.