J. Webster (ed.), *Wiley Encyclopedia of Electrical and Electronics Engineering*

FREQUENCY CONVERTERS AND MIXERS

An essential part of most microwave receivers is the frequency converter, a device that converts the frequency of an incoming signal to another frequency. The output frequency may be *downconverted, upconverted, multiplied,* or *divided*. Important and common is the *mixer*, which downconverts a high-frequency input signal f_s to a much lower and more manageable signal f_{IF}, preserving information concerning the amplitude, frequency, and phase of the input signal. The devices used are nonlinear (i.e., the relationship between current and voltage is not linear). The devices may have either two terminals (diodes) or three terminals (transistors).

Mixers are important for very high frequencies, where amplifiers are not available and direct amplitude or phase detection is difficult. A mixer can be used to downconvert, say, a terahertz frequency signal to a microwave frequency one, where electronic methods are readily available for amplification and any kind of demodulation. In fact, in almost any radio set or mobile telephone receiver or base station, there are several mixers and other types of frequency converters.

Both two-terminal devices (diodes) and three-terminal devices (transistors) are used in mixers. The frequency conversion is accomplished by using the *nonlinear* properties of the device. Virtually all semiconductor devices, such as diodes and transistors, show nonlinear properties in certain bias ranges. Common devices for microwave mixer applications are the Schottky diode, the field-effect transistor (*FET*), and the bipolar transistor. There are many other devices available as well, for example, the superconducting tunneling device (*SIS*) and the superconducting hot electron device for low-noise (high-sensitivity) millimeter and submillimeter wave receivers. For infrared wavelengths, metal–insulator–metal devices and, for optical frequencies, photoconducting devices have been used. Note that ordinary resistors, capacitances, and inductances are linear components.

Figure 1 gives an example of how mixers are used in a receiving system. The antenna is connected directly to the mixer. For example, in a TV–satellite receiver, an amplifier is placed just after the antenna to increase the signal amplitude. In reality there are many other systems (e.g., radar, radio, measurement systems) where mixers are used to frequency downconvert the input signal.

To give a simple illustration of how a diode mixer may work, consider the detector circuit shown in Fig. 2 using an ideal diode (zero resistance in the forward direction and infinite resistance in the backward direction). If a sinusoidal voltage $V(t) = V_{LO}\cos(2\pi f_{LO}t)$ is applied, it will be "rectified" (detected) by the diode. The voltage over the resistance $v_R(t)$ will be a constant dc voltage proportional to V_{LO}. Next, add a *small* signal voltage $\delta v_s \cos(2\pi f_s t)$, that is, $V(t) = V_{LO}\cos(2\pi f_{LO}t) + \delta v_s \cos(2\pi f_s t)$. Assuming that $\delta v_s/V_{LO} \ll 1$, the resulting voltage $V(t)$ will become amplitude modulated, as shown in Fig. 2. The detected voltage $v_R(t)$ over the load resistance R will be proportional to the envelope $V_R(t)$ (assuming $\tau = RC \ll 1/\omega_{IF}$) (see Fig. 3), that is,

$$v_R(t) = V_{LO}\left(1 + \frac{\delta v_s}{V_{LO}}\cos(2\pi f_{IF}t)\right) \qquad (1)$$

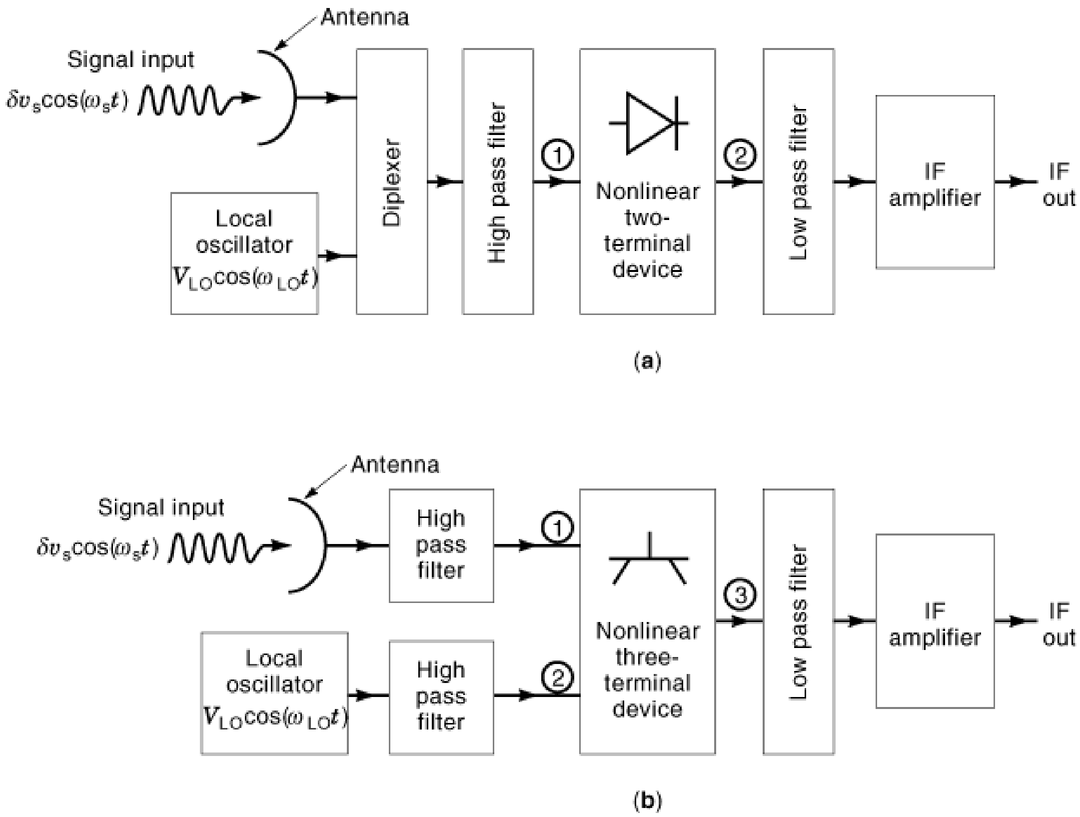

Fig. 1. Typical mixer block diagram. At the input there is a high pass filter that will prevent any low-frequency IF power from escaping to the mixer input (left). The low pass filter will stop any input signal or LO power from going to the IF circuit (right). Part (a) shows a layout for a two-terminal device (diode) and (b) for a three-terminal device (transistor).

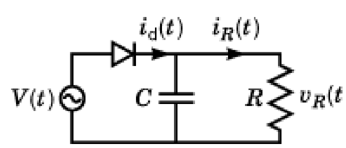

Fig. 2. A rectifying diode circuit. If the capacitance C and the resistance R are large enough, $v_R(t)$ will follow the envelope of $v(t)$ (see Fig. 3).

where $f_{IF} = |f_{LO} - f_s|$ is the intermediate frequency (*IF*). The dc part of Eq. (1) is the "detected local oscillator (*LO*)" and the alternating part is identical to the IF voltage and, in this example, is equal in amplitude to the input signal. This IF signal is fed into an amplifier as described in Fig. 1.

Note that in the simplified example above we have not correctly accounted for a number of parameters, such as the impedances of the LO and signal sources. This means that the IF voltage will not become equal to δv_s. A more detailed description of a more correct calculation is given in the section entitled "Schottky Diode Mixers." For details concerning microwave mixers, see Refs. 1 and 2.

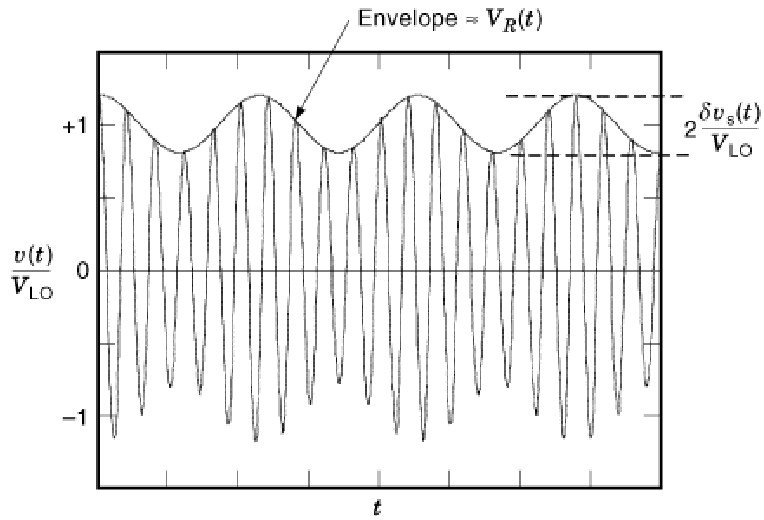

Fig. 3. An example with a drive voltage $v(t)/V_{LO} = \cos(\omega_{LO}\,t) + (\delta v_s/V_{LO})\cos(\omega_s\,t)$, where $\delta v_s/VLO = 0.2$ and $\omega_{LO}/\omega_s = 1.18$. The IF is $0.18\,\omega_{LO}$.

General Properties Of Two-Terminal Non-Linear Devices

Below we describe some general results obtained when a two-terminal device such as the Schottky diode is excited with a sinusoidal signal. The impedance of the Schottky diode is voltage dependent. The speed of this device is indeed high. It shows a nonlinear behavior up to several terahertz.

Frequency Multiplication. Consider a nonlinear diode device exposed to LO (or pump) power yielding a large voltage swing $V(t) = V_{LO}\cos(\omega_{LO}t)$ over the diode. (Note: Below we use $\omega = 2\pi f$). Since the relation between current and voltage is not linear (i.e., $I \neq$ const. $\times V$), the resulting current will *not* have a sinusoidal shape like the input voltage. However, the current in this case is still a *periodic* function versus time with the same periodicity, $\tau = 2\pi/\omega_{LO}$, as the LO frequency and can consequently be expressed as a Fourier series with harmonics of the LO, $n\omega_{LO}$.

$$I(t) = \sum_{n=0}^{\infty} i_n \cos(n\omega_{LO}t + \varphi_n) \tag{2}$$

The component i_0 is the dc component. To obtain power at a particular harmonic, for example, the third harmonic, it is required that the current component at $3\omega_{LO}$ be passing through a resistance R_3 delivering a power of $1/2{i_3}^2R_3$. To avoid any power being delivered at other harmonics it is necessary to ensure that the device is reactively terminated at these harmonics.

In reality, the impedance of most nonlinear devices is complex with both the *real* and the *imaginary* parts voltage (or current) dependent. For a more detailed theory, see the section entitled “Large Signal Analysis by Harmonic Balance.”

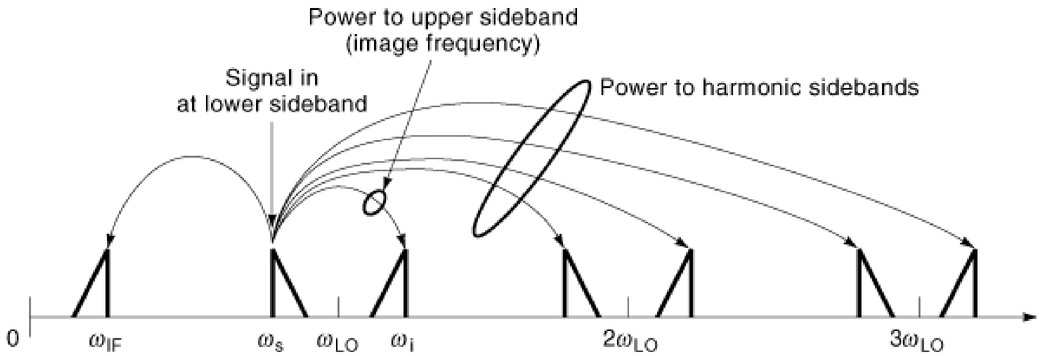

Fig. 4. Power flow in a lower sideband mixer. The input signal power is distributed not only to the IF but also to the harmonic sidebands.

Frequency Conversion. If two signal voltages at ω_{LO} and ω_s are simultaneously interacting with the nonlinear diode impedance, the resulting current can be expressed in a more complex Fourier series,

$$I(t) = \sum_{m=-\infty}^{+\infty} \sum_{n=-\infty}^{+\infty} i_{m,n} \cos((m\omega_{LO} + n\omega_s)t + \varphi_{m,n}) \qquad (3)$$

In the case when both the signal voltage and the local oscillator voltage are large, a large number of frequency conversion products are obtained. Obviously it is possible to generate power at any frequency $m\omega_{LO} \pm n\omega_s$. If for $m = n = 1$ the required output frequency is higher than the signal frequency, one has *frequency upconversion*, and if the output frequency is lower than the signal frequency, one has *frequency downconversion*.

Linear Mixing. As already mentioned, a mixer receiver usually handles "small signals." We have a small signal case if the amplitude of the signal at ω_s is much smaller than the LO amplitude at ω_{LO}. In this case only harmonics of ω_{LO} are important and we are left with

$$I(t) \approx \sum_{m=-\infty}^{+\infty} i_m \cos[(m\omega_{LO} + \omega_s)t + \varphi_m] \qquad (4)$$

In most mixer applications, we are interested in the "intermediate frequency," $\omega_{IF} = |\omega_{LO} - \omega_s|$. An IF load resistance in the circuit will allow power at the IF frequency to be extracted (compare Fig. 2). Equation (4) suggests that power may go not only to the IF but also to *harmonic sidebands*, as illustrated in Fig. 4. The only way to prevent this power loss is to make sure that the harmonic sideband current components are facing impedances that are purely reactive.

Note that an IF signal (Fig. 4) will be created if either $\omega_{su} = \omega_{LO} + \omega_{IF}$ or $\omega_{sl} = \omega_{LO} - \omega_{IF}$. The former frequency ω_{su} is called the upper sideband, and the latter ω_{sl} the lower sideband. For a *lower sideband mixer* the upper sideband is denoted the *image frequency*, and vice versa for an *upper sideband mixer*. For a lower or upper sideband case, some signal power may go to the image frequency. In an *image reject mixer* a filter prevents the image frequency from entering the mixer. In the *image enhanced* mixer, the image terminal is terminated reactively so that the conversion loss is reduced.

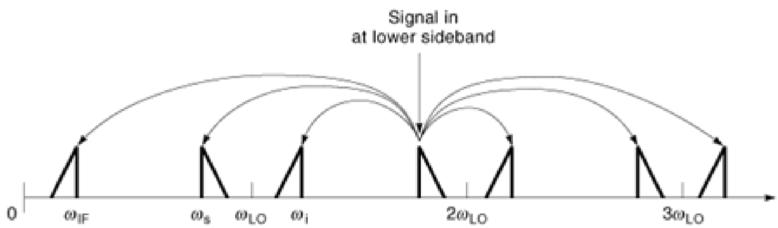

Fig. 5. Power flow in a second harmonic lower sideband mixer.

At high signal powers, there may be confusion because the mixer may produce output signals in the IF band (see intermodulation below) for

$$|m\omega_{\mathrm{LO}} - n\omega_{\mathrm{s}}| = \omega_{\mathrm{IF}} \tag{5}$$

Harmonic Mixer. In a harmonic and small signal linear mixer one has $\omega_{\mathrm{IF}} = |n\omega_{\mathrm{LO}} = \omega_{\mathrm{s}}|$. A spectrum analyzer always uses a harmonic mixer to analyze the signal, and the harmonic number n can be very high (>10). The power flow in a harmonic mixer is shown in Fig. 5.

Intermodulation. All mixer products created by two or more signals are called intermodulation (*IM*) products. Most IM products are unwanted.

For example, if two signals at slightly different frequencies ω_{s1} and ω_{s2} and with a power of the same order of magnitude as the LO are interacting with the nonlinear device, the current will contain frequency products as shown in the following equation:

$$I(t) = \sum_{m=-\infty}^{+\infty} \sum_{n=-\infty}^{\infty} \sum_{k=-\infty}^{\infty} i_{m,n} \cos[(m\omega_{\mathrm{LO}} + n\omega_{\mathrm{s1}} + k\omega_{\mathrm{s2}})t + \varphi_{m,n,k}] \tag{6}$$

All signals at frequencies fulfilling the requirement

$$\omega_{\mathrm{s1}} \approx |m\omega_{\mathrm{LO}} + n\omega_{\mathrm{s2}} \pm \omega_{\mathrm{IF}}| \qquad -\infty < m,n < \infty \tag{7}$$

may create IM products in the IF band. Indeed, intermodulation must be considered a potentially serious problem in all applications where strong signals may occur.

Systems Aspects

In a system, the following properties are important: (1) conversion loss, (2) noise properties, and (3) intermodulation properties.

Conversion Loss. An important property of a mixer is the conversion loss L, defined as

$$L = \frac{P_{\mathrm{s}}}{P_{\mathrm{IF}}} = \frac{\text{Signal power available at the input}}{\text{IF power delivered to the IF load}} \tag{8}$$

In most practical mixers the conversion loss is larger than one. However, it is possible to obtain gain owing to parametric amplification caused by a nonlinear capacitance. In certain configurations using transistors (see Refs. 1 or 4) conversion gain can be obtained. Superconducting mixers, however, due to quantum phenomena, can show stable conversion gain.

There are several loss mechanisms causing the conversion loss:

(1) Losses due to absorption in the nonlinear device
(2) Losses due to power lost to harmonic sidebands
(3) Losses due to reflection at the input port
(4) Losses due to reflection at the IF output port

See also the section entitled “Schottky Diode Mixers.”

Mixer Receiver Noise. The important noise measure in practical applications is always the total receiver noise temperature. The contributions approximately in order of importance are: (1) mixer device noise, (2) the IF amplifier noise, (3) thermal noise from the mixer circuit, and (4) LO noise. There are two different noise measures usually cited in the literature: the *single sideband* (*SSB*) noise temperature and the *double sideband* (*DSB*) noise temperature.

When calculating the noise temperature of a mixer, it is advisable to always start adding up all noise contributions at the IF amplifier input (see Fig. 6). The temperature then becomes

$$T_{\mathrm{in}} = \frac{T_{\mathrm{l}}}{L_{\mathrm{s}}} + \frac{T_{\mathrm{l}}}{L_{\mathrm{i}}} + \sum_{n=2}^{\infty}\left(\frac{T_{\mathrm{l}}}{L_{n+}} + \frac{T_{\mathrm{l}}}{L_{n-}}\right) + T_{\mathrm{M\,out}} + T_{\mathrm{IF}} \quad (9)$$

where L_{s}, L_{i}, $L_{\mathrm{n+}}$, and $L_{\mathrm{n-}}$ are the conversion losses at the signal, image, and upper harmonic sidebands and lower harmonic sidebands, respectively. T_{l} is the noise temperature of the input resistance, $T_{\mathrm{M\,out}}$ is the noise from the mixer diode entering the IF amplifier. As the next step, identify the signal to noise ratio:

$$\frac{P_{\mathrm{s}}}{P_{\mathrm{in}}} = \frac{P_{\mathrm{s}}/L_{\mathrm{s}}}{kT_{\mathrm{in}}\,\Delta f} = \frac{P_{\mathrm{s}}}{kT_{\mathrm{syst}}\,\Delta f} \quad (10)$$

where T_{syst} is by definition the system noise temperature for a *single sideband* receiver. It is assumed that the useful and interesting signal enters only one sideband. Hence

$$T_{\mathrm{syst,SSB}} = T_{\mathrm{in}}\,L_{\mathrm{s}} = T_{\mathrm{l}}\left[1 + \frac{L_{\mathrm{s}}}{L_{\mathrm{i}}} + \sum_{n=2}^{\infty}\left(\frac{L_{\mathrm{s}}}{L_{n+}} + \frac{L_{\mathrm{s}}}{L_{n-}}\right)\right] + T_{\mathrm{MXR,SSB}} + L_{\mathrm{s}}T_{\mathrm{IF}} \quad (11)$$

where we have defined the equivalent noise temperature of the mixer itself, $T_{\mathrm{MXR,SSB}} = T_{\mathrm{M\,out}}\,L_{\mathrm{s}}$.

For the double sideband case, one assumes a useful signal to enter both the upper and the lower sidebands. Hence the signal to noise ratio for this case should be defined as

$$\frac{P_{\mathrm{s}}}{P_{\mathrm{n}}} = \frac{P_{\mathrm{s}}(1/L_{\mathrm{s}} + 1/L_{\mathrm{i}})}{kT_{\mathrm{in}}\,\Delta f} = \frac{P_{\mathrm{s}}}{kT_{\mathrm{syst}}\,\Delta f} \quad (12)$$

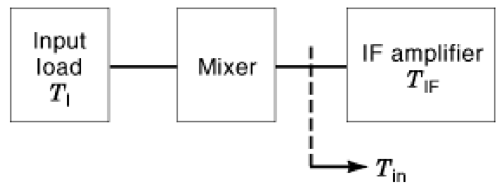

Fig. 6. Receiver configuration for calculating the receiver noise. For the calculation it is wise to determine the noise temperature at the input of the IF amplifier, T_{in} .

that is,

$$T_{\mathrm{syst,DSB}} = \frac{T_{\mathrm{in}}}{1/L_{\mathrm{s}} + 1/L_{\mathrm{i}}} = T_{\mathrm{l}}\left[1 + \frac{L_{\mathrm{s}}L_{\mathrm{i}}}{L_{\mathrm{s}} + L_{\mathrm{i}}}\sum_{n=2}^{\infty}\left(\frac{1}{L_{n+}} + \frac{1}{L_{n-}}\right)\right] + T_{\mathrm{MXR,DSB}} + \frac{L_{\mathrm{s}}L_{\mathrm{i}}}{L_{\mathrm{s}} + L_{\mathrm{i}}}T_{\mathrm{IF}} \tag{13}$$

where the double sideband noise temperature of the mixer itself is defined as

$$T_{\mathrm{MXR,DSB}} = T_{\mathrm{M\ out}}\frac{L_{\mathrm{s}}L_{\mathrm{i}}}{L_{\mathrm{s}} + L_{\mathrm{i}}} \tag{14}$$

Note that if $L_{\mathrm{s}} = L_{\mathrm{i}}$, both the single sideband mixer and system noise temperatures are twice as large as for the double sideband case. The LO noise (if important) can be taken into account by adding a certain amount at the input port. Also note that the noise entering the mixer at the harmonic sidebands may considerably influence the total receiver performance.

Schottky Diode Mixers

The Schottky diode mixer is the most common type of mixer for frequencies from megahertz to terahertz.

Schottky Diode for Mixer Applications. A schematic diagram of a common design of millimeter wave Schottky barrier diode is shown in Fig. 7. Note that the radio frequency (*RF*) current is flowing from the diode contact at the surface of the diode chip to the back contact. Hence the RF series resistance is slightly larger for RF than for dc. The current–voltage (*I*–*V*) characteristic of the junction itself can be calculated from

$$i_{\mathrm{d}} = I_0\left[\exp\left(\frac{qV_{\mathrm{j}}}{\eta kT}\right) - 1\right] \tag{15}$$

where I_0 is the saturation current, V_{j} the junction voltage, q the charge of the electron, k Boltzmann's constant, T the physical temperature, and η the ideality factor, which for good diodes at room temperature is between, say, 1.03 and 1.10.

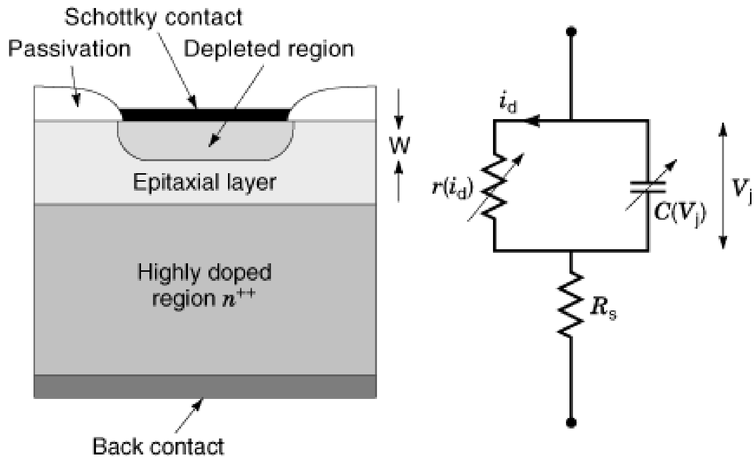

Fig. 7. Simple design of a GaAs Schottky barrier diode and the corresponding equivalent circuit. The higher the frequency the smaller the diodes required.

The small signal RF junction properties of a Schottky diode can be modeled as a nonlinear resistance r_j in parallel with a nonlinear capacitance C_j,

$$r(i_d) = \frac{\partial V_j}{\partial i_d} = \frac{k\eta T}{q i_d} \qquad C(V_j) = \frac{C_0}{\sqrt{1 - \frac{V_j}{V_b}}} \tag{16}$$

where C_0 is the zero bias capacitance and V_b is the built-in voltage of the diode. For more details, see 3.

A common measure of the high-frequency properties is the cutoff frequency, which is defined for the diode at zero bias,

$$f_c = \frac{1}{2\pi R_s C_0} \tag{17}$$

where R_s is measured using the dc I–V characteristic. For good mixer performance this cut-off frequency must be much larger than the signal frequency. The series resistance uses up signal power and should for obvious reasons be made as small as possible. For the conversion efficiency the nonlinear capacitance is of much less importance than the nonlinear resistance. The diode is typically designed as a thin n-doped active layer of 1000 Å (Mott diode), which leaves one with a diode with reduced capacitance variation and a minimum series resistance.

GaAs is preferred for millimeter wave diodes. The main reason is that the high mobility of GaAs yields a low series resistance and consequently a high cutoff frequency. Another advantage (to silicon) is that carriers do not freeze out when the diode is cooled to cryogenic temperatures in order to improve the mixer noise properties [see Eq. (7)].

The noise performance and the conversion efficiency are the prime properties of the diode when used in mixer applications. Noise properties are discussed next.

Shot Noise. The shot noise is due to fluctuations in the electron particle current between cathode and anode of the diode (see Ref. 3). The root-mean-square (*rms*) fluctuations in the current are $\delta i^2 = 2eI\ \Delta f$, where

Δf is a small frequency interval. The noise power of the Schottky diode can then be calculated as

$$P_\mathrm{n} = \frac{\delta i^2}{4\frac{\partial I_\mathrm{d}}{\partial V_\mathrm{j}}} = \frac{1}{2}k\eta T\,\Delta f \qquad (18)$$

Identifying this equation with the ordinary Johnson noise expression, $P_\mathrm{n} = kT\,\Delta f$, it is seen that the equivalent noise temperature of the Schottky diode is

$$T_\mathrm{sh} = \frac{1}{2}\eta T \qquad (19)$$

Note that the noise temperature decreases linearly with decreasing physical temperature. However, the charge transport at room temperature over the Schottky barrier is due to thermionic emission and decreases when the temperature is lowered. Hence for temperatures on the order of 50 K to 100 K and below, temperature-independent tunneling becomes the dominant process for electrons passing the barrier, and the equivalent temperature ηT of Eq. (6) becomes

$$T_\mathrm{tunnel} = \frac{qh}{k}\sqrt{\frac{N_\mathrm{d}}{4\varepsilon m^*}} \qquad (20)$$

where N_d is the doping concentration in the epitaxial layer, ε the dielectric constant of the semiconductor, and $m*$ the effective mass of the electron. In practice, T_tunnel is 50 K for $N_\mathrm{d} = \quad 3 \times 10^{16}$. This doping concentration is recommended for mixers operated at 15 K or 20 K, a typical temperature for commercial cryogenic cooling machines.

1/f Noise. There is excess noise at low frequencies (of the order 100 kHz and lower for a good diode), which is related to surface phenomena at the metal–semiconductor interface. This noise is normally not important in mixers for millimeter wave receivers. However, it is an important limiting factor for certain radar and communications systems.

Thermal Noise. The series resistance is essentially an ordinary resistor and consequently causes ordinary thermal (Johnson) noise. The main noise contribution of this type comes from the substrate (corresponding resistance R_sub) and is denoted as T_sub.

Hot Electron Noise and Intervalley Scattering Noise. Since the diode area is very small in order to make the capacitance reasonably small, a high current density is required to make $r_\mathrm{j}\ R_\mathrm{s}$. This means that electrons may obtain energies larger than the energy related to their thermal movement and *hot electron noise* is obtained. The increase in energy also means that electrons can be transferred from the main Γ valley in the *E–k* diagram (for details see Ref. 3) to the upper L valley, causing fluctuations in the electron velocity and *intervalley scattering noise* is obtained. The hot electron and intervalley scattering noises occur essentially in the undepleted part of the epilayer (resistance R_epi) and are denoted as T_epi.

Large Signal Analysis by Harmonic Balance. When the nonlinear device is pumped by the LO, harmonic currents are created (e.g., see Refs. 4, 5, or 6). The equivalent circuit of a Schottky diode mixer is shown in Fig. 8. Note that the series resistance is assumed to be linear and will be included in the embedding circuit.

The current $I_\mathrm{e}(t)$ contains harmonics of the LO pump frequency, $k\omega_\mathrm{LO}$. For a Schottky diode, this current consists of two parts: one that is associated with the nonlinear resistance $i_\mathrm{d}\ (t)$, and one with the parallel

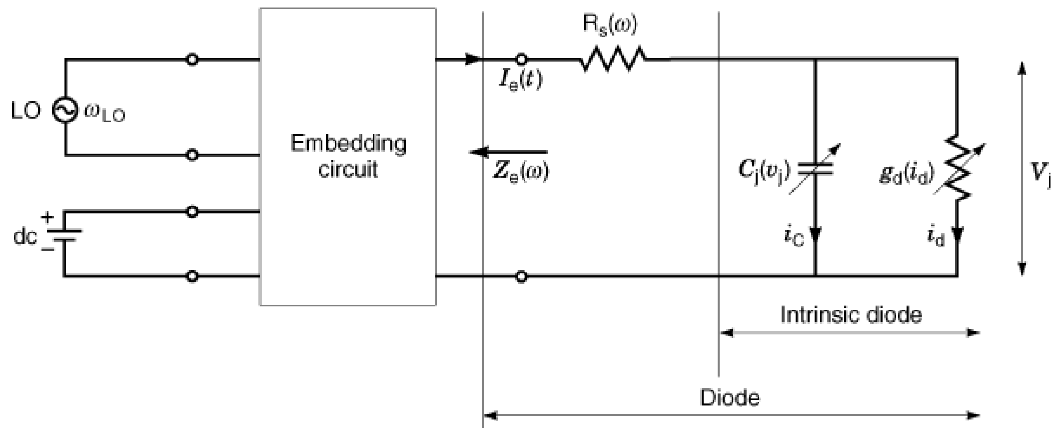

Fig. 8. Equivalent circuit of a mixer. For the intrinsic diode, C_j and g_d are nonlinear and are characterized in the time domain, while the diode series resistance R_s and the embedding impedance Z_e are linear and can be described in the frequency domain.

nonlinear capacitance $i_c(t)$. We have [using now the complex notation; compare Eq. (2)]

$$I_e(t) = i_C(t) + i_d(t) = \sum_{k=-\infty}^{k=\infty} I_{ek} e^{jk\omega_{LO}t} \quad I_{ek} = I^*_{e-k} \qquad (21)$$

This current $I_e(t)$ flows through the embedding circuit, creating voltages at harmonic frequencies at $k\omega_P$p; namely,

$$V_j(t) = \sum_{k=-\infty}^{k=\infty} V_k e^{jk\omega_{LO}t} \qquad V_k = V^*_{-k} \qquad (22)$$

The boundary conditions set by the *embedding circuit* require that

$$V_k = -I_{ek}\left[Z_{ek}(k\omega_{LO}) + R_s(k\omega_{LO})\right] \qquad k = \pm 2,\ \pm 3, \ldots, \pm\infty \qquad (23)$$

$$V_{\pm 1} = V_{LO} - I_{e\pm 1}\left[Z_{e\pm 1}(\pm\omega_{LO}) + R_s(\pm\omega_{LO})\right] \qquad (24)$$

$$V_0 = V_{dc} - I_{e0}[Z_e(0) = R_s(0)] \qquad (25)$$

where V_{LO} and V_{dc} are the LO and dc bias voltages, respectively. The frequency dependence of R_s is due to the skin effect. If $V_j(t)$ is known, $i_d(t)$ and $i_C(t)$ (see Fig. 8) can be calculated from Eqs. (15) and (16). We now have a nonlinear problem to solve in order to determine I_{ek} and V_{ek}. Several iteration type methods have been suggested. See Refs. 4 and 5 for more details.

Having determined the components I_k and V_k, we have $V_{\mathrm{j}}(t)$ and $I_{\mathrm{e}}(t)$, and we can determine $i_{\mathrm{d}}(t)$, $i_{\mathrm{C}}(t)$, $g_{\mathrm{d}}(t)$, and $C_{\mathrm{j}}(t)$.

$$g_{\mathrm{d}}(t) = \sum_{k=-\infty}^{k=\infty} G_k e^{jk\omega_{\mathrm{LO}}t} \qquad G_k = G^*_{-k} \tag{26}$$

$$C_{\mathrm{j}}(t) = \sum_{k=-\infty}^{k=\infty} C_k e^{jk\omega_{\mathrm{LO}}t} \qquad C_k = C^*_{-k} \tag{27}$$

These equations together with the embedding impedance $Z_{\mathrm{e}}(\omega)$ allow us to determine the small signal properties of the mixer.

Small Signal Analyses. The relation between the small signal current and voltage vectors $\delta\boldsymbol{I}$ and $\delta\boldsymbol{V}$ can be expressed in a more general form as

$$\delta\boldsymbol{I} = Y\, d\boldsymbol{V} \tag{28}$$

where

$$Y_{mn} = G_{m-n} + j(\omega_0 + m\omega_{\mathrm{LO}})C_{m-n} \tag{29}$$

where for convenience we use ω_0 for ω_{IF}. It is convenient to form an augmented Y matrix, Y', as indicated in Fig. 9. This augmented network contains the whole mixer, including diode and embedding network, but does not contain signal sources associated with these terminations. Since we define the signal sources as current sources, the augmented network in this case is open-circuited. For the augmented network

$$\delta\boldsymbol{I'} = Y'\, d\boldsymbol{V} \tag{30}$$

and

$$Y' = Y + \mathrm{diag}\left[\frac{1}{Z_{\mathrm{e}m} + R_{\mathrm{s}m}}\right] \tag{31}$$

Inverting Eq. (23) yields

$$\delta\boldsymbol{V} = Z'\, d\boldsymbol{I'} \tag{32}$$

where

$$Z' = \frac{1}{Y'} \equiv \begin{bmatrix} & \vdots & \vdots & \vdots & \\ \cdots & Z'_{11} & Z'_{10} & Z'_{1-1} & \cdots \\ \cdots & Z'_{01} & Z'_{00} & Z'_{-10} & \cdots \\ \cdots & Z'_{-11} & Z'_{-10} & Z'_{-1-1} & \cdots \\ & \vdots & \vdots & \vdots & \end{bmatrix} \tag{33}$$

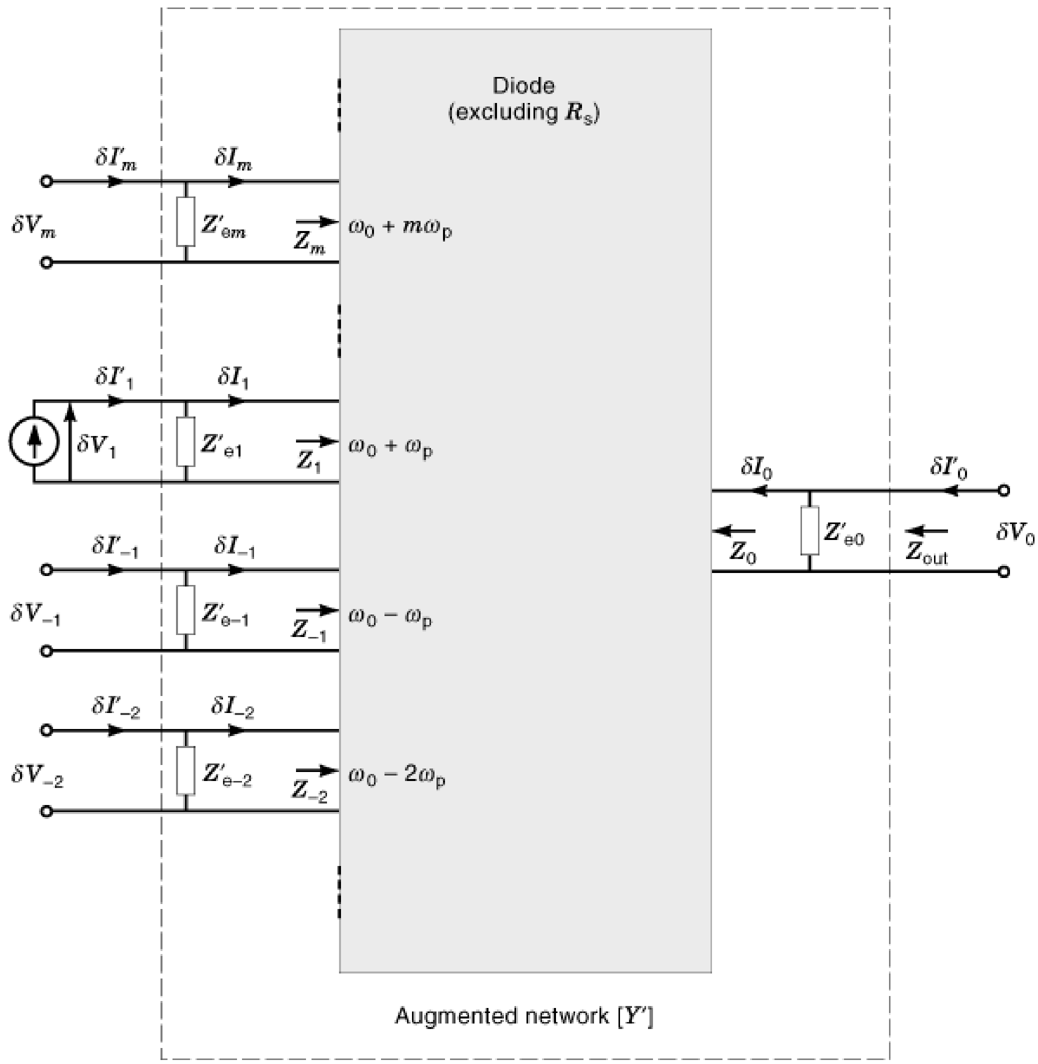

Fig. 9. The *small signal* representation of the mixer as a multifrequency linear multiport network. Notice that R_s is included in $Z'_{ek} = Z_{ek} + R_s$. In a more exact model, R_s is frequency and bias dependent. In this text we use R_{s1} for the series resistance at the input frequency (fundamental mixer) and R_{s0} for the series resistance at the IF. Index 0 indicates parameters at the IF; ω_0 is the IF.

Mixer Port Impedances. The port impedance Z_mm, defined in Fig. 9, can be determined if the corresponding embedding impedance is open-circuited, that is,

$$Z_m = Z'_{mm,\infty} \tag{34}$$

where the subscript ∞ means that Z'mm is evaluated for $Z_{em} = \infty$. The IF output impedance becomes

$$Z_{out} = Z_0 + R_{s0} = Z'_{00,\infty} + R_{s0} \tag{35}$$

Conversion Loss. The conversion loss of a mixer is defined as [compare Eq. (8)]

$$L = \frac{\text{Power available from source } Z_{e1}}{\text{Power delivered to load } Z_e 0} \tag{36}$$

yielding the conversion loss

$$L = \frac{1}{4|Z'_{01}|^2} \frac{|Z_{e0} + R_{s0}|^2}{Re[Z_{e0}]} \frac{|Z_{e1} + R_{s1}|^2}{Re[Z_{e1}]} \tag{37}$$

where Z_{ek} is defined in Fig. 9 (see figure caption) and Z'_{01} is the 01 diagonal element of Eq. (33). A more general expression for the conversion loss from any sideband j to any other sideband i is

$$L_{ij} = \frac{1}{4|Z'_{ij}|^2} \frac{|Z_{ei} + R_{si}|^2}{Re[Z_{ei}]} \frac{|Z_{ej} + R_{sj}|^2}{Re[Z_{ej}]} \tag{38}$$

Equivalent Noise Temperature of the Mixer. The equivalent input noise temperature of the mixer, T_M, is defined as the temperature that the real part of the Z_{el} (lower sideband) must have in order to generate the same noise voltage as the diode itself generates during mixer operation. It is

$$T_M = \frac{\langle \delta V_{N0}^2 \rangle}{4k\,\Delta f} \frac{|Z_{e1} + R_{s1}|^2}{|Z'_{01}|^2\, Re[Z_{e1}]} \tag{39}$$

where $\langle \delta V^2{}_{NO}\text{NO} \rangle$ is

$$\langle \delta V_{N0}^2 \rangle = Z'_0 \langle \delta \boldsymbol{I}'_s\, \delta \boldsymbol{I}_s^{\dagger} \rangle Z_0^{\dagger} + Z'_0 \langle \delta \boldsymbol{I}'_T\, \delta \boldsymbol{I}_T^{\dagger} \rangle Z_0^{\dagger} \tag{40}$$

where

$$\begin{aligned} \langle \delta I'_{Tm}\, \delta I'^{*}_{Tm} \rangle &= \frac{4kT_{eq} R_{sm}\,\Delta f}{|Z_{em} + R_{sm}|^2} \qquad m \neq 0 \\ &= \frac{4kT_{eq} R_{sm}\,\Delta f}{|Z_0|^2} \qquad m = 0 \end{aligned} \tag{41}$$

and

$$\langle \delta I'_{sm}\, \delta I'^{*}_{sn} \rangle = 2eI_{m-n}\,\Delta f \tag{42}$$

I_{m-n} is the current component at the harmonic $(m - n)$ and Z_0 is the zero row of the matrix Z', Eq. (33). Z_{e1} is defined in Fig. 9.

The noise temperature as defined in Eq. (42) is the *single sideband noise temperature* for the lower sideband.

Knowing the diode parameters, the Schottky mixer can be analyzed with high accuracy using commercial software.

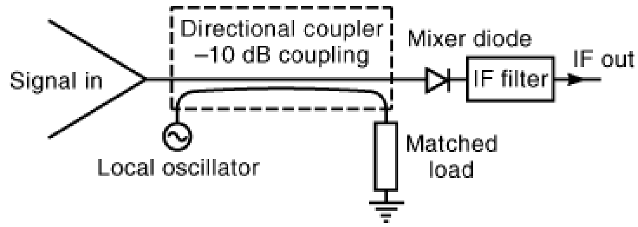

Fig. 10. Example for a single ended mixer configuration. Note that the directional coupler used for injecting the LO couples only 10% of the LO power to the mixer diode, and that the coupler consequently attenuates the signal by a factor of 0.9.

Diode Mixer Topologies and Mixer Design

There are a number of different approaches one may consider in designing mixers. Symmetry properties are one way of defining basic types of mixers, namely,

(1) The single ended mixer with one diode and a common input port for the signal and the LO
(2) The single balanced mixer with two diodes and separate ports for signal and LO
(3) The double balanced mixer with four diodes and separate ports for signal and LO
(4) The double–double balanced mixer with eight diodes and separate ports for signal and LO

For millimeter and submillimeter wave applications, types 1 and 2 have been implemented experimentally and are described in the literature, while all four types are common in microwave frequency applications.

The performance of either mixer depends on the *impedances* seen at the signal, at the IF, and at the harmonic sidebands. The LO should experience a reasonably good match in order to reduce the LO power requirement. In practice, the impedances at the harmonics of the LO and at the harmonic sidebands are very difficult to control. The exception is the impedance at the image frequency, which can often be controlled. The importance is illustrated by referring to properties of a typical broadband mixer (when $L_s = L_i$) used in a single sideband application. In such a mixer a considerable amount of the signal power (≈25%) may end up at the image frequency.

It is also obvious that the *noise* entering the mixer at the image port will be converted with the same efficiency as the signal at the signal port, adding to the system noise.

Single Ended Mixer. A single ended mixer has one input port, used for both the signal and the LO. Hence it is necessary to incorporate a circuit in front of the mixer itself for injecting the LO. This circuit should not significantly attenuate the signal. For example, using a 10 dB directional coupler will attenuate the signal ∼10% and add noise (see Fig. 10). A common way of introducing the LO in microwave and millimeter wave mixers is to use a narrow band diplexer, for example, a filter structure in the input waveguide or a quasi-optical interferometer in the signal path in front of the mixer (e.g., see Ref. 10).

Single Balanced Mixers. In single balanced mixers, the signal and the LO enter the mixer through different ports, isolated from each other. Either 90° or 180° hybrids or baluns are used [Fig. 11(a)] (see Ref. 1 for details). Figure 11(b) shows a low-frequency equivalent circuit of a single balanced mixer. The paths of the signal current i_s and the LO current i_{LO} indicate that they add in one diode and subtract in the other. This causes an imbalance in A, which will slowly cycle at a frequency equal to the IF. Hence the IF power can be subtracted between A and ground. Note that if the LO is noisy, this will not cause any output noise at the IF port.

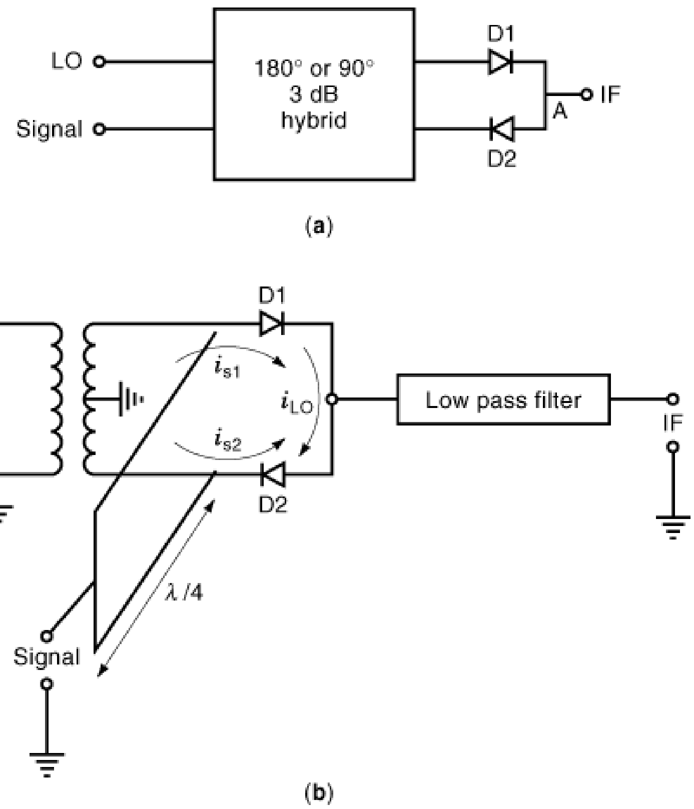

Fig. 11. Basic design of the single balanced mixer (a) and equivalent circuit of the 180° hybrid mixer (b).

The use of two diodes rather than one means that the mixer can handle twice as much power for the same intermodulation as for the single ended mixer. In summary, for the single balanced mixer:

- The signal, LO, and IF ports are isolated from each other.
- The LO noise cancels at the IF port.
- The power handling is superior to the single ended mixer.

Double Balanced Mixers. Essentially, a double balanced mixer is constructed from two single balanced mixers, coupled in parallel and 180° out of phase. The diodes can be arranged in either a star or ring configuration (see Fig. 12). The ring can be arranged very compact as a monolithic circuit.

If the diodes are perfectly identical, the symmetry ensures perfect isolation between the signal and the LO ports. The topology also yields cancellation at the IF port of the even harmonics of both the signal and the LO frequencies. This also means that intermodulation is reduced as compared to the mixers mentioned above. Hence the advantages of the double balanced mixer are:

- Excellent isolation between the signal, LO, and IF ports
- LO noise cancellation at the IF port
- Superior power handling compared to the double balanced mixer
- Superior intermodulation properties compared to the double balanced mixer

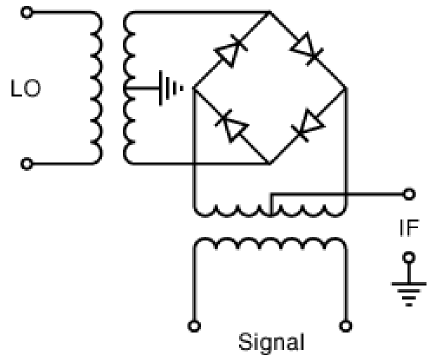

Fig. 12. A double balanced mixer configuration: the ring mixer.

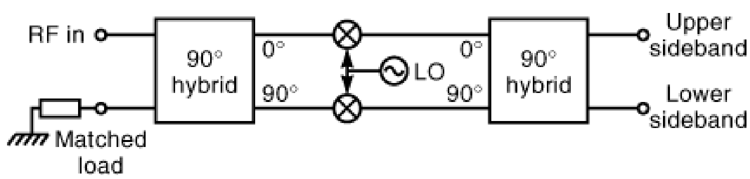

Fig. 13. A single sideband mixer using two balanced mixers and two 90° hybrids. Note that at one output port the upper sideband appears, while at the other output port the lower sideband appears.

Double–Double Balanced Mixers. Double–double balanced mixers are constructed using two double balanced mixers. Eight diodes are used, leading to further power handling capacity and still better intermodulation properties.

Image Rejection and Image Enhancement. The system properties of a single sideband mixer receiver can be improved by introducing a proper circuit at the image frequency. It is of particular importance to reactively terminate the image frequency so that no signal power is lost at the image frequency, and no noise (or any other unwanted signal) at the image frequency can be converted to the IF frequency. Furthermore, if the reactance at the image frequency is chosen properly, the "signal will be reflected back into the mixer," such that the conversion is *enhanced* and/or the noise properties are improved. A stopband filter can be added in the input transmission line to prevent one sideband from reaching the diode. The distance to the diode is chosen to optimize the mixer conversion loss. The conversion loss becomes several decibels lower than the typical 5 dB for a common broadband microwave mixer.

A most elegant method to realize a single sideband mixer is shown in Fig. 13. By using two balanced mixers and two 90° hybrids, it is possible to arrange that the upper sideband and the lower sideband exit the mixer at different ports (see Ref. 1 or Ref. 4 for details).

Harmonic Mixers

In a harmonic mixer a harmonic of the LO frequency, $n\omega_{\mathrm{LO}}$, is used for mixing; that is, the IF is obtained as

$$\omega_{\mathrm{IF}} = |n\omega_{\mathrm{LO}} - \omega_{\mathrm{s}}| \tag{43}$$

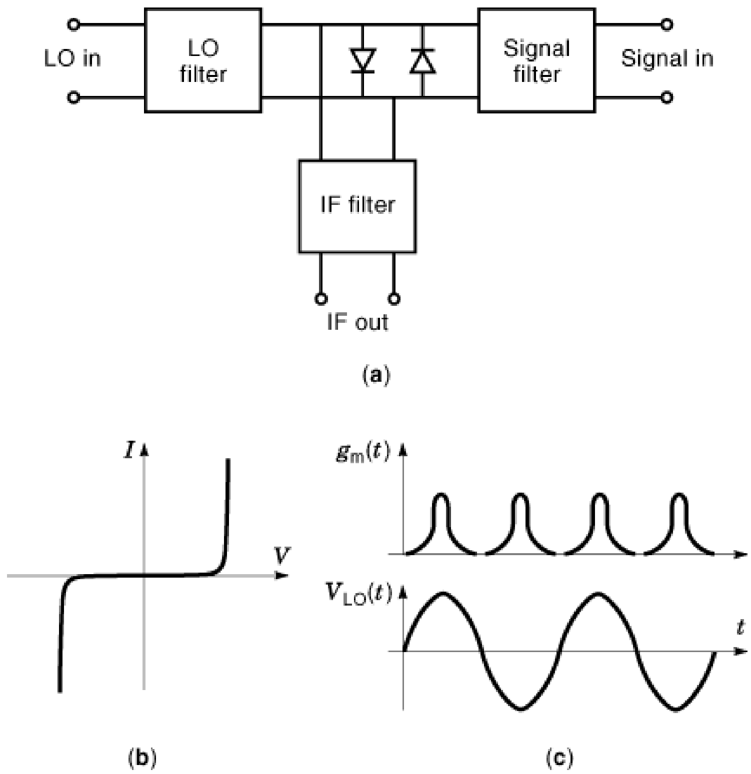

Fig. 14. Subharmonically pumped mixer using antiparallel diodes: (a) the mixer circuit, (b) the dc I–V characteristic, and (c) the resulting waveforms for the LO voltage and the time-dependent small signal conductance.

Harmonic mixers are practical when it is difficult to realize LO power at a frequency near the signal frequency. They are particularly useful at millimeter and submillimeter waves.

Large harmonic numbers n are often used when maximum sensitivity is not required. For example, in spectrum analyzers large harmonic numbers may be used.

Two Diode Subharmonically Pumped Mixers. If two diodes are used in an antiparallel configuration (see Fig. 14), the small signal conductance will vary with twice the LO frequency. Hence the mixer will convert signals located near $2\,f_{\mathrm{LO}}$ and no conversion will occur near f_{LO}. The advantage of the two diode to single diode subharmonically pumped mixers is that no conversion can occur at the fundamental frequency. Moreover, the LO noise will contribute less, since the frequency difference between the signal and the LO is of the order f_{LO}. Another advantage is inherent self-protection against large peak reverse voltage burnout.

Parametric Frequency Conversion

In a parametric frequency converter, a nonlinear reactance, such as a back-biased Schottky diode, is used. Common parametric components are frequency downconverters, frequency upconverters, and frequency multipliers. In a frequency downconverter, a strong pump signal f_{LO} and signal f_{s} (strong or weak depending on application) are applied to the device. The output frequency is

$$f_{\mathrm{out}} = |f_{\mathrm{LO}} - f_{\mathrm{s}}| \qquad \text{frequency downconversion} \qquad (44)$$

For a frequency upconversion, we have

$$f_{\mathrm{out}} = |f_{\mathrm{LO}} + f_{\mathrm{s}}| \qquad \text{frequency upconversion} \qquad (45)$$

The signal frequency in this case may be much lower than the LO (or pump) frequency; that is, the output frequency is not far from the LO frequency.

Note that if the device has no resistive parasitics, no power is lost in the device itself, and 100% efficiency is theoretically possible. However, power may go to harmonics or harmonic sidebands and there are always some parasitic resistances present, for example, the series resistance in a Schottky diode.

If we select a large ratio $f_{\mathrm{LO}}/f_{\mathrm{s}}$, the frequency upconverter may have high gain. This is possible since, by proper choice of the circuit parameters, parametric amplification is achieved (e.g., see Ref. 7). In the case of gain, one has negative resistance in the circuit and one may face stability problems. However, it is very difficult to make a broadband parametric upconverter, since proper impedances have to be realized at f_{s}, f_{LO}, and f_{out}. Parametric converters are used much less today than a few decades ago. The reason is that the quality of mixer diodes and FETs has improved significantly and it is much easier to make diode (or FET) mixers very broadband. This is the reason why resistive mixers are preferred in most applications. Note that FET mixers can be designed for a conversion gain greater than one.

A classical reference concerning varactor circuits is the book by Penfield and Rafuse (7). A parametric downconverter, like the Schottky mixer, can be analyzed using commercial software.

Negative Resistance Diode Mixers

In the current–voltage (I–V) characteristic of, say, the Esaki tunnel diode or the resonant tunneling diode, there is a region that has a differential negative resistance. This means that the mixer can have conversion gain. Tunnel diode mixers have been built and tested. However, a large junction capacitance made the frequency range quite limited, which together with poor power handling, stability problems, and less favorable noise properties means that these mixers have very little practical use today.

Self-Oscillating Mixers

The negative resistance devices can as well promote an oscillation. Hence it is possible to design circuits where the LO is delivered by the same device that is performing the mixing. Besides the devices mentioned already Gunn diodes have also been used in self-oscillating mixers. The sensitivity of such mixers is limited. The advantage may be in applications where the best performance is less important and the lowest price is required, as, for instance, in low price Doppler radar applications.

Bolometer Mixers

Bolometer mixers have been constructed since the 1950s. Since the electromagnetic absorption in bolometer devices can be essentially frequency independent, it should be possible to do mixing to several terahertz. In this type of mixer, one is using the fact that when two signals at slightly different frequencies are superimposed the resulting signal can be described as a signal which is amplitude modulated with the difference frequency. The first useful bolometer mixer was based on InSb devices cooled to temperatures of a few kelvin. When the device absorbs the modulated signal, the electron temperature becomes modulated, leading to a modulation in the

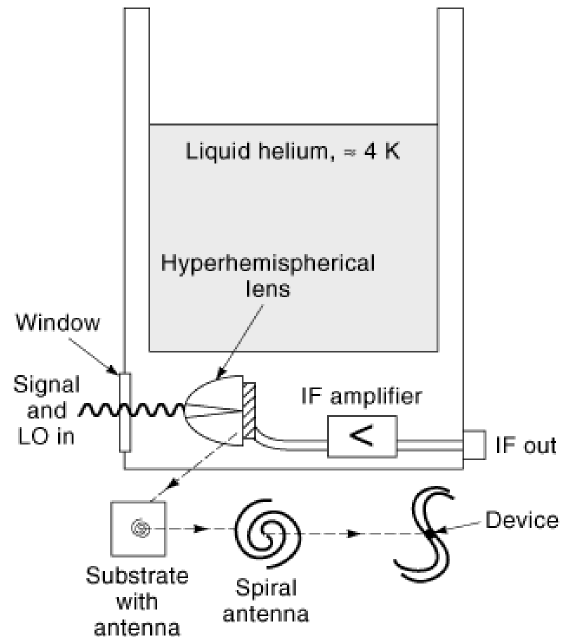

Fig. 15. Schematic description of typical dewar setup for a submillimeter wave SIS or HEB mixer. The LO and the signal are entering together through the dewar window. The radiation is focused on the antenna using a hyperhemispherical lens, downconverted to the IF in the nonlinear device and finally amplified by the IF amplifier. In this figure the antenna is illustrated as a spiral antenna. There are many other possible planar antenna structures available (see Ref. 9 for details).

device resistance. The theory is described in more detail in the section entitled "The Hot Electron Bolometer Mixer" and by Arams et al. (8).

However, the thermal time constant for the InSb device is long, allowing a maximum IF of only about 2 MHz. A more recent bolometer mixer is based on a two-dimensional electron gas in HEMT materials allowing an IF to about 1 GHz (see Ref. 9). However, the most successful hot electron bolometer mixer so far is the superconducting hot electron bolometer mixer.

Mixers Based on Superconducting Devices

Room temperature mixers for frequencies from about 100 GHz to a few THz frequencies use only Schottky diodes. However, if sensitivity is an issue (e.g., as in radio astronomy), there is a better alternative in mixers based on superconducting devices cooled to a few kelvin. Low noise superconductor–insulator–superconductor (*SIS*) mixers have excellent performance up to about 1 THz (see Ref. 10). Superconductor hot electron bolometer (*HEB*) mixers are the best alternative for frequencies above 1 THz.

In Fig. 15 is shown a schematic of a receiver (except for the the input quasi-optics) based on superconducting devices.

The SIS Mixer. The SIS mixer is also called the "quasiparticle mixer." Due to the extremely strong nonlinearity in the *I–V* characteristic of the SIS device, quantum effects are important. Indeed, for certain choices of embedding impedance network, this can result in a conversion gain (see Ref. 11 for details).

In a superconductor and at a temperature below the superconducting transition temperature, electrons form pairs, called Cooper pairs. When they do so, the energy of the electrons near the Fermi energy is lowered

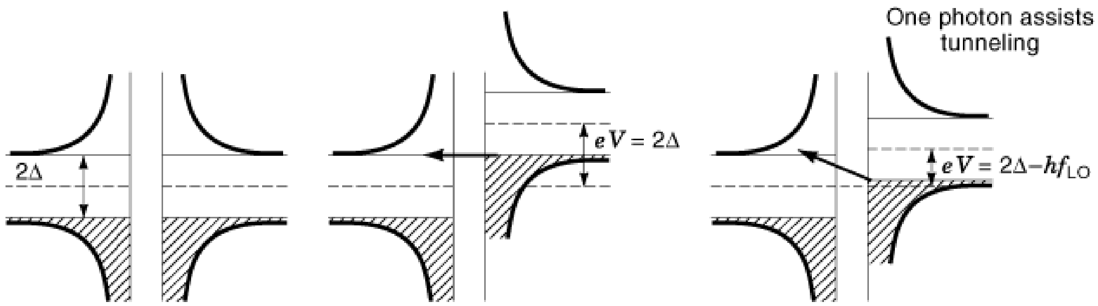

Fig. 16. The SIS device under bias: (a) no bias; (b) for a voltage bias $V > 2\Delta/e$ electrons will tunnel from right to the left; (c) tunneling is assisted by a photon with energy hf_{LO} for the bias voltage $V = (\Delta - hf_{\mathrm{LO}})/e$.

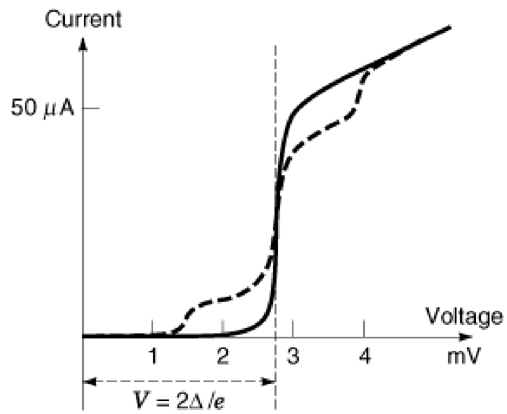

Fig. 17. Typical *I–V* characteristic of a SIS element. The thin line indicates the shape of the pumped *I–V* characteristic, where $f_{\mathrm{LO}}\, h \approx 1.4$ meV ($f_{\mathrm{LO}} = 325$ GHz), while the thick line is for the unpumped device.

by a certain amount Δ. Hence to break up a Cooper pair, an energy of 2Δ is required. This can be described in terms of a bandgap with the energy 2Δ, as shown in Fig. 16.

In the SIS mixer, a tunneling phenomenon is used. In the device, two superconducting films are separated by a thin (~20 Å) layer of insulator. Under bias, Cooper pairs on one side of the isolator break up into two electrons (quasiparticles) that individually tunnel through the isolator and recombine on the other side of the insulator. This is illustrated in Fig. 16. It is interesting to note that the density of states near the band edges becomes “infinite.” This is one important reason why there is such a sharp increase in the current when the device is biased to a voltage $V = 2\ \Delta/e$ (e is the charge of the electron). The *I–V* characteristic is shown in Fig. 17. Note that the voltage scale is in mV, and that 1 meV *corresponds to* 240 GHz. The steps in the *I–V*-curve correspond to dc bias voltages, where exactly $eV = 2\Delta - hf$ ($V \approx 1.35$ mV). Since the *I–V* characteristic is strongly nonlinear within a fraction of a millivolt, the mixer is operating in the quantum regime. Compare also Fig. 18, where the *I–V* character of an SIS device is compared with a Schottky diode. The Schottky diode obviously is not very nonlinear within a voltage interval of 1 mV and is therefore operating fully as a classical mixer for frequencies up to several terahertz. It should also be mentioned that the required LO power is very low, on the order nanowatts. This is of great importance for submillimeter wave mixers, where substantial LO power is difficult to obtain.

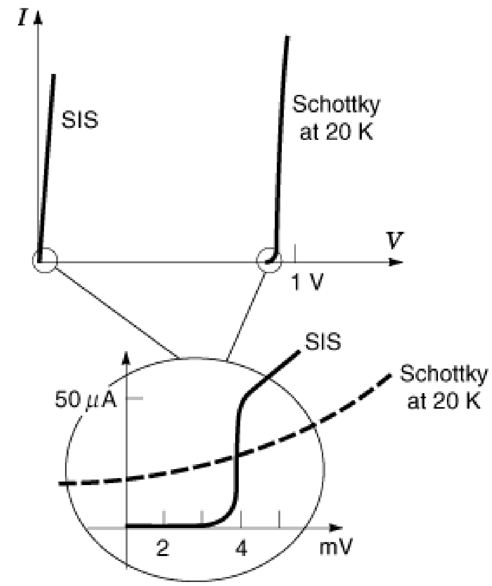

Fig. 18. Comparing the *I*–*V* characteristic of a SIS element with that of a Schottky diode. Note the enormous difference in nonlinearity.

The best SIS devices are realized in so-called Nb trilayer technology (see Ref. 11). The device structure is Nb/Al_2O_3/Nb, where the ≈20 Å thick Al_2O_3 serves as the insulator in the SIS device. For frequencies above about 700 GHz, one is trying to develop devices based on NbN, which has a higher bandgap (≈1.2 THz) than Nb. So far these attempts have not been very successful.

The basic noise in the SIS mixer is shot noise. Comparing the *I*–*V* characteristics of the SIS device and Schottky diode and using the classical theory described in the section entitled "Schottky Diode Mixers," one can see that th the mixer noise and the conversion loss are essentially lower for the SIS mixer. However, the theory must include quantum effects (see Ref. 11 for more details). This leads to a conversion gain that is possibly larger than one, a fact that has been demonstrated in practice.

Hot Electron Bolometer Mixer. A bolometer consists of an absorber that is heated by radiation and a temperature-dependent resistance as a "thermometer." The bolometer has a thermal time constant τ_O limiting the maximum detectable modulation frequency of the absorbed power. This means that the maximum feas- ible IF is $f_{IF} = 1/(2\pi\tau_0)$. When the LO and signal are added together, the instantaneous power variation is described by $[V_{LO}\cos(\omega_{LO}t) + v_s\cos(\omega_s t)]^2\,(1/R_{RF})$ (see Fig. 3). If the bolometer can respond to ω_{IF} but not to $(\omega_{LO} + \omega_s)$, $2\omega_{LO}$, and $2\omega_s$, the bolometer temperature and the resistance will be approximately proportional to $P_{LO} + P_s + 2\sqrt{P_{LO}P_s}\,P_{LO}P_s\cos(\omega_{IF}t)$. Note that the "slow" response of the bolometer device means that there are no harmonics of the LO created and no signal power is transformed to the image frequency, as is the case for both the SIS and Schottky mixers.

A *superconducting hot electron bolometer* (*HEB*) consists of one or several superconducting thin film strips in parallel, deposited on a substrate, for example, silicon, single crystalline quartz, or sapphire. The strips are cooled to the superconducting state and then heated by dc and microwave power to temperatures near the superconducting to normal transition temperature, where the superconductor will gradually become normal (Fig. 19).

The maximum IF is determined by the electron temperature relaxation time τ_O, that is, $f_{IF} < 1/(2\,\pi\tau_O)$, and a major issue is to find ways of making the time constant τ_O short enough. Figure 20 indicates how cooling occurs in the so-called phonon-cooled and the diffusion-cooled bolometer, respectively (compare Refs. 10 and 12).

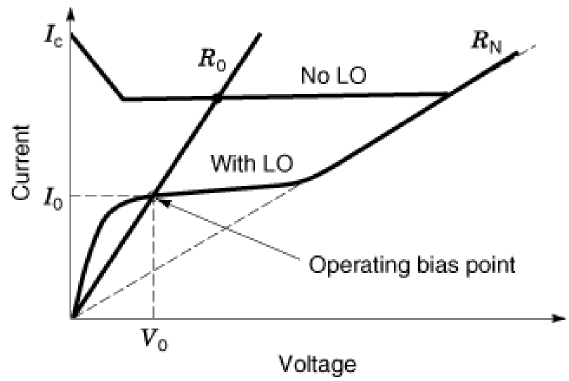

Fig. 19. Current–voltage (I–V) characteristic of a HEB with and without a LO. At the operating point, $V_O/I_O = R_O$. For large bias voltage the whole strip is normal conducting ($R = R_N$). I_c is the critical current, the maximum current in the completely superconducting state.

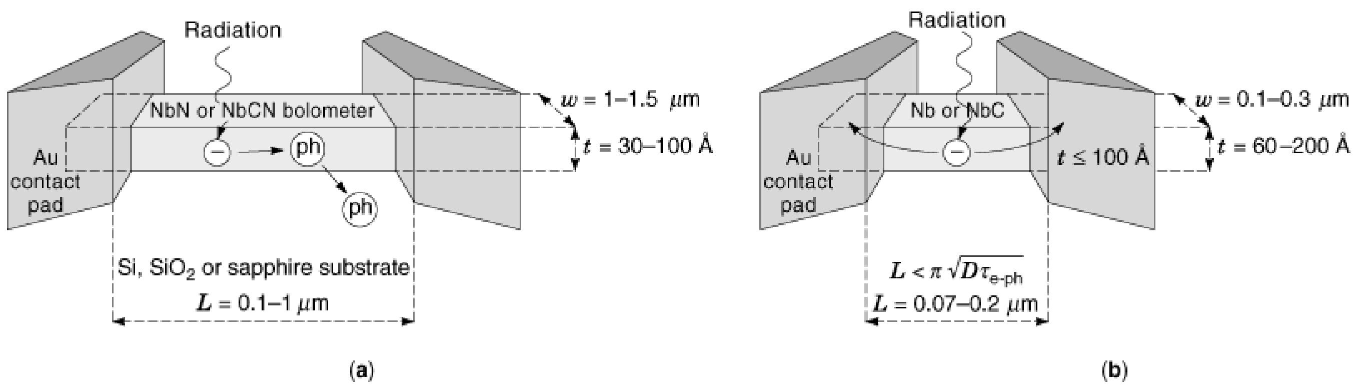

Fig. 20. Two types of bolometer devices: (a) phonon-cooled and (b) diffusion-cooled.

When operating the mixer, the device is absorbing LO power (P_{LO}) and signal power (P_s) as well as power from the dc bias supply ($P_{dc} = V_O I_O$). When the power increases, obviously the electron temperature increases and the resistance of the device increases as $\Delta R = (dR_O/dP)\,\Delta P = C_0\,\Delta P$. Figure 21 shows a simple equivalent circuit of the mixer, where the device is biased by a constant dc current. Consequently, the modulation at the IF of the resistance will cause an IF voltage to appear across the device, causing an IF current through the IF load resistance R_L. The IF current ΔI is superimposed on the dc bias current through the mixer device $R_O(P)$ and will cause a "modulation" of the dc power $\Delta P_{dc}\cos(\omega_{IF}t)$. The total power dissipated in the device is then

$$\begin{aligned} P(t) &= P_0 + \Delta P(t) \\ &= P_{dc} + \Delta P_{dc}\cos(\omega_{IF}t) + P_{LO} + P_s + 2\sqrt{P_{LO}P_s}\cos(\omega_{IF}t) \end{aligned} \tag{46}$$

Assuming that dc and RF power affects the resistance by the same amount, the IF modulation of the device resistance becomes $C_0\,\Delta P(t)$. There is a resulting bias point of the device $V = V_O$ and $I = I_O$. Defining the device dc resistance R_O as the time average of $R_O(P)$, that is, $R_O = V_O/I_O$, one obtains (8) the conversion

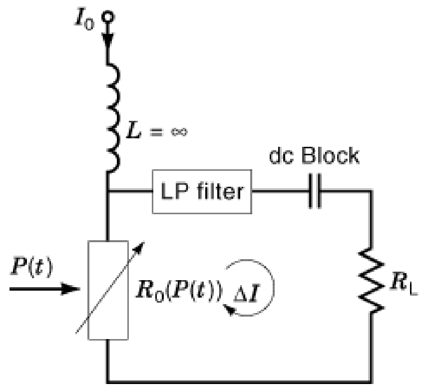

Fig. 21. Equivalent circuit of bolometer with load.

gain,

$$G = \frac{P_{IF}}{P_s} = 2C_O^2 \frac{P_{LO}\, P_{dc}}{(R_L + R_O)^2} \cdot \frac{R_L}{R_O} \left(1 - C_O \frac{P_{dc}}{R_O} \cdot \frac{R_L - R_O}{R_L + R_O}\right)^{-2} \tag{47}$$

where $C_O = dR_O/dP$, R_L is the IF load resistance, P_{IF}, P_S, P_{LO}, and P_{dc}, are the IF, signal, LO, and dc power, respectively, dissipated in the device.

The commonly assumed fundamental limit of −6 dB gain for hot electron mixers is not valid if a negative differential resistance of the unpumped *I–V* curve is available (see Ref. 13 for details). The load resistance for maximum gain is equal to the differential resistance of the *I–V* curve at the bias point of the pumped mixer.

Experimental Results. In experiments on phonon-cooled NbN HEB mixers, noise temperatures of about 400 K (DSB) have been obtained at 600 GHz, and 1000 K (DSB) at 900 GHz. At 2.5 THz a noise temperature of about 1400 K has been obtained. The conversion loss is typically 10 dB, including losses from the optics in front of the mixer.

The noise of these mixers is caused by thermal fluctuations in the bolometer device (causing resistance fluctuations) and by Nyquist noise. For diffusion-cooled mixers a noise temperature of 650 K DSB at 533 GHz was measured by Skalare et al. (14). These experiments indicate that an IF bandwidth of at least 3 GHz is achievable. Later experiments show that it should be possible to obtain at least 50% higher IF bandwidths in practical mixers. A crucial number is the maximum IF bandwidth. For a phonon-cooled HEB mixer, a 3.7 GHz IF bandwidth (−3 dB reduction in conversion gain) has been measured and for diffusion-cooled between 2 and 6 GHz. While this bandwidth is defined for a *conversion loss* increase of 3 dB, the bandwidth defined for when the *noise temperature* has increased by 3 dB is about 1.5 times larger (∼5.5 GHz for the phonon-cooled one).

The LO power needed is less than 100 nW, which is much lower than needed for Schottky diode mixers.

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