# Chapter 6

# Diode Applications

A vast array of diode mixer circuits and technologies have been developed since the early days of radio some 100 years ago. Diodes were the first devices used in "frequency changing" modulators and still are present in the majority of mixer circuits used today. Semiconductor diode technology evolved from the crystal point contact, then the copper oxide rectifier, then the Silicon contact signal diode junction. It is now dominated by Schottky barrier diodes, which are widely available in silicon or gallium arsenide. Diode mixer circuits are implemented using waveguide, lumped element, coaxial, and planar technologies that include the microstrip, slotline, finline, and coplanar families. The circuits presented in this chapter have been selected from a wide range of patents and publications on the subject and are grouped by circuit topologies including single ended, singly balanced, doubly balanced, triply balanced, quadrature, subharmonic, and selected special circuits.

## 6.1 SINGLE ENDED

Diode mixers were first built using point contact crystal diodes, and later Schottky diodes became dominant due to improved performance and reliability. Most single ended diode mixer patents appeared early in the progression of mixer technology. Early mixers were also built using thermionic emission tubes, which suffered from the inherent reliability and variability problems caused by the heat applied to the cathode to excite electron flow. An interesting innovation in 1921 was a point contact rectifier fitted with a socket to become a drop-in replacement for a tube rectifier [1], as depicted in Figure 6.1. The single diode demodulated the amplitude modulated (AM) signal in the usual way, by first rectifying the RF signal and then envelope detecting it to extract the modulated audio signal. This is effectively a single ended mixer used as a direct converter, wherein the AM signal's carrier wave is the LO and the sidebands are the RF. The sidebands mix with the carrier to produce the baseband output.

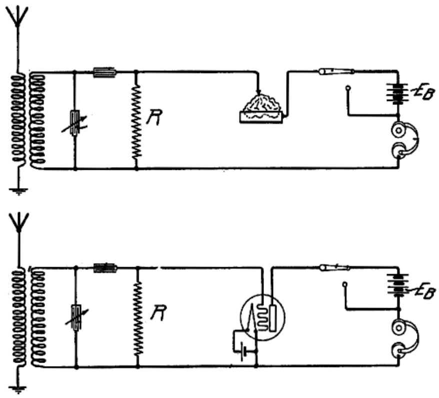



**Figure 6.1** A point contact diode replacement for a tube diode (1921).

### 6.1.1 Coaxial Circuits

An example of an early single ended diode coaxial mixer is found in short wavelength voice communication systems introduced by Ohl in 1939 [2]. It used a point contact rectifier connected between the ends of two parallel semi-circular conducting rods, see 2 in Figure 6.2. The rods together act as the probe in a waveguide aperture antenna, supporting the point contact rectifier at the point of optimum conversion efficiency. A focusing lens is placed in front of the waveguide antenna, and a local oscillator is applied either from a radiating second antenna near the lens, or by directly connecting the LO source to the detector through the two rods. The point contact rectifier was used instead of a tube because of its much lower parasitic capacitance.

The point contact diode comprised a crystal and a very fine tungsten contact wire whose position on the crystal could be moved to optimize performance. The tip of the wire was plated with a highly conductive alloy of gold, silver, and platinum [3]. The authors list the material options for the crystal, including boron, arsenic, tellurium, iron-pyrite, and silicon. Of these, silicon is the best due to its hardness, resistance to oxidation at low temperatures, and electrical resistance. It was noted that the noise produced during the rectification process was dramatically reduced by having a good back contact to ground at the base of the semiconductor. This was accomplished by electroplating the silicon crystal's backside with chromium, or in the case of iron-pyrite plating it with gold. The communication system was dual conversion, with the second mixer also accomplished using a semiconductor diode. Development of crystal mixers at microwave frequencies progressed rapidly during World War II to support the development of radar, which was led by the British.

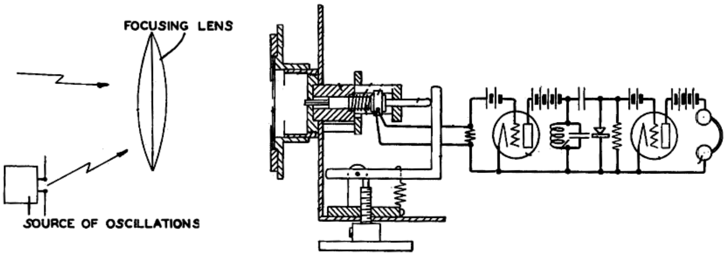



**Figure 6.2** Short wave communication system operating at 1 GHz, 1939.

A crystal rectifier mixer developed during that period in Britain comprises a single ended coaxial design having a circular waveguide RF input [4]. As depicted in Figure 6.3, the RF input to the mixer comprises a transition from waveguide to a telescoping coaxial line that in turn connects to the cathode of the detector. The IF output uses a second coaxial section whose vertical center conductor also connects to the cathode of the detector diode, and utilizes a sliding dielectric block to capacitively short circuit the RF and LO signals to isolate them from the IF output, and to provide the LO ground return. The LO is capacitively coupled to the anode end of the crystal detector via a second telescoping coaxial section. And a fourth coaxial section also connects to the anode end of the detector diode to provide RF and IF ground returns. This ground return section also uses a sliding dielectric block for impedance matching.

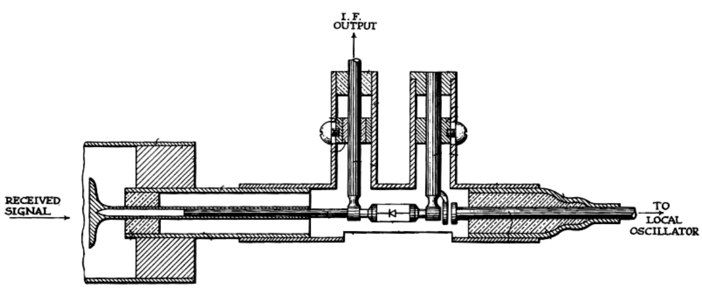



**Figure 6.3** Single ended coaxial mixer, 1945.

Other crystal diode microwave mixers were developed during this period. The invention displayed in Figure 6.4 depicts a mixer having a waveguide RF input port that transitions to a coaxial line using a current loop wire [5]. The coaxial line is connected to the detector diode. The LO is connected to the diode via a second coaxial line. The IF output comprises a third coaxial line connected to the diode via a conical inductor having a higher self resonance due to lower inter-winding capacitance. The IF ground return current flows through the wire loop waveguide probe.

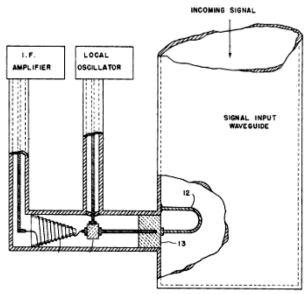



**Figure 6.4** Waveguide mixer using conical inductor, 1951.

Figure 6.5 depicts an improved mixer from this period [6]. It has a rectangular waveguide RF input, and represents a significant advance in bandwidth and simplicity of construction. The waveguide RF input terminates in a backshort, with a "cross conductor" placed at a certain distance in front of the backshort. The rectangular waveguide has the usual cross section whose width is twice the height. The cross conductor runs the full width of the waveguide, and is centered on the waveguide's cross section. The diode is connected perpendicularly to the cross conductor, between it and the waveguide wall, at center of the waveguide width where the E-field is maximum for the $TE_{10}$ mode. Both ends of the cross conductor pass through the waveguide side walls, and continue on to form coaxial lines. The LO is introduced through one coaxial line, and the IF is extracted via the other. The RF energy is coupled to the diode by means of its electric field being parallel to the diode, which is located at the $TE_{10}$ mode maximum. Broad bandwidth is achieved because the LO port has no back short, and LO to RF isolation is achieved through waveguide mode orthogonality rather than by filtering. The RF and LO signals are isolated from the coaxial IF port by means of a back short choke, which allows for a broadband IF port. Also, the diode connection at the waveguide wall provides a broadband ground return for all three signals.

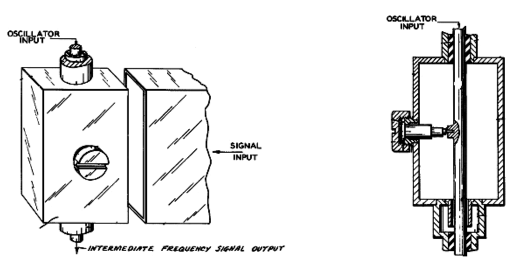
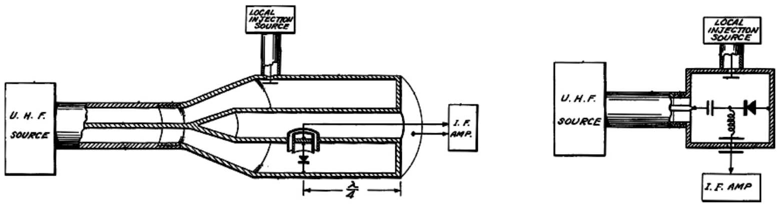



**Figure 6.5** Rectangular waveguide mixer, 1952.

In Figure 6.6 a coaxial mixer is depicted with further improvements, having conversion loss of 5 dB and bandwidth of 80 MHz [7]. The distinguishing characteristics cited are the IF connection to the crystal diode is isolated from the RF and LO signals, and the machining and assembly operations are simplified. The LO is inserted by a capacitive probe that excites a TEM mode in the coaxial structure, coupling it to the crystal diode to generate the desired mixing products. The circuit is claimed to provide LO, RF to IF isolation, with the LO to RF isolation largely dependent on how energy is absorbed by the diode. If the LO and RF are matched to the diode, then isolation is high; otherwise, it will be degraded by the reflections from the mismatched impedances. A second embodiment is Figure 6.6(b), which uses a much simpler construction, and presumably has degraded performance as a tradeoff.




(a) (b)

**Figure 6.6** Coaxial mixer, 1951; (a) diode is mounted on an enlarged coaxial mount; (b) the diode and IF filtering is assembled in a small cavity.

A waveguide subharmonic mixer is depicted in Figure 6.7 [8]. It is designed to operate with an RF frequency at least twice that of the LO. The major goal of the design is to reduce RF signal losses by minimizing RF leakage out the LO port. The mixer comprises a coaxial resonator having a center conductor that is physically less than a quarter wave long at the RF frequency. The center conductor is capacitive loaded at its top by an adjustable screw that allows the capacitance to be tuned by adjusting the screw. The capacitive loading makes the center conductor be electrically a quarter wave long at the RF frequency. The RF is injected into the cavity near the base of the center conductor, where a low impedance exists. The diode anode is connected to the center conductor, also near its base. The diode cathode is connected to a line that runs through a hole in the side wall of the cavity that allows the IF output to connect to a subsequent gain stage, and also provides a capacitive short circuit for the RF and LO ground returns. The LO is injected to the diode anode via a coaxial line that runs up through the hollow center conductor of the coaxial resonator. The LO is connected to the coaxial line via a transformer, with the ground connection of the secondary also providing the IF ground return. An open circuit shunt line that is a quarter wave long at the RF frequency, is also connected to the diode's anode, in parallel with the LO connection. This shunt line provides the anode's RF ground return and isolates RF from leaking out the LO port. But the open circuit shunt stub does not short out the LO, which is at less than half the RF frequency. The high selectivity of the coaxial resonator and open circuit shunt stub cause this design to trade off low conversion loss for narrow band operation.

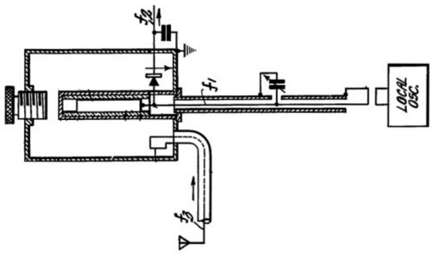



**Figure 6.7** Single ended coaxial mixer, 1952.

Point contact crystal diodes continued to be used at millimeter-wave (mmW) frequencies into the 1950s. Figure 6.8 depicts a single ended diode mixer operating at mmW RF and LO frequencies [9]. The mixer comprises a rectangular waveguide with 0.28 inch width and 0.14 inch height, and having an adjustable back short. A point contact crystal diode is mounted in front of the back short, centered on the waveguide's 0.28 inch width at the TE10 mode's electric field maximum. The crystal element is mounted at the top of the waveguide, with one end of a metal whisker contacting the crystal, and the other end of the metal connecting to the center conductor of a coaxial line that forms the IF output port. The mmW RF and LO are applied together to the diode via the waveguide.

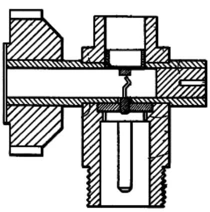


**Figure 6.8** Crystal contact waveguide mixer, 1956.

Figure 6.9 depicts another point contact crystal mixer design [10]. Its goals are to provide broadband operation while minimizing loss of RF energy due to leakage. Another goal is to provide an improved method of packaging and mounting the point contact crystal that facilitates replacement of the diode. The RF and LO inputs to the mixer are incident together at a rectangular waveguide aperture. A fixed back short is placed at the other end of the waveguide. A vertical E-field probe is centered on the waveguide's width (long dimension) and is located approximately a quarter wave at the RF frequency in front of the back short toward the RF and LO input. The tip of the probe is centered on the waveguide's height (short dimension). The probe passes through the top wall of the waveguide and then forms a coaxial line that in turn connects to the whisker of the diode. Thus the RF and LO are connected to the diode. The IF port comprises a coaxial line whose center conductor is a wire connected to the end of the probe. The whisker contacts the crystal inside a removable cylindrical package. The crystal is mounted to a ground connection that provides the broadband ground return for the RF, LO, and IF signals.

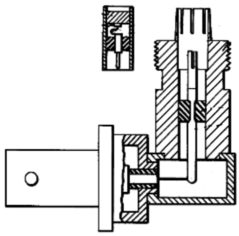


**Figure 6.9** Crystal mixer with replaceable diode, 1960.

Further innovations with single ended diode mixers appear to not be covered by patents until the 1970s and 1980s when Schottky diodes came into usage with stripline and microstrip on printed circuits. The patent 1974 [11] by Otremba, describes a single ended mixer with reactive terminations to the image and LO second harmonic that optimize conversion loss. Two such mixers are combined to form a ratrace singly balanced mixer. The resulting mixer is claimed as an improvement over the 1972 ratrace mixer design by Halford [12].

### 6.1.2 Microstrip Circuits

Stripline and microstrip fabrication became popular for microwave components in the 1960s. Conventional microstrip circuits are built with metallic transmission line traces formed on one side of the substrate while the other side is fully metalized ground plane. Single ended mixers benefit from the simplicity and desirable electrical and mechanical properties of this transmission medium.

Two single ended diode microstrip mixer designs patented in the 1980s are introduced next. The first is in Figure 6.10 illustrating a mixer that is simple to fabricate and assemble, provides sharp filtering, is sufficient to isolate the RF and LO ports, and shorts the image signal to minimize conversion loss [13]. The RF port is on the left at 1, and connects to the diode's cathode by microstrip line 15. The coupled line notch filter 17 presents a high impedance at the LO frequency looking into microstrip line 15 at the diode connection point. Similarly, the LO port is on the right side at 2, and connects to the diode 7 by microstrip line 16. The notch filter 18 causes a high impedance to appear at the RF frequency looking into microstrip line 16 at the diode connection point. Thus the RF and LO signals are connected to the diode and remain isolated from each other's ports. The IF is connected to the diode by the low pass filter 12. The RF and LO ground returns are provided by the capacitive stubs of the low pass filter that are connected to the diode's anode. The IF ground return is provided by the microstrip line 13, which is a quarter wave long at the RF and LO frequencies and thus provides an open circuit to the these while providing a short circuit at the IF. Short circuits are provided at the image frequency to the diode's cathode by the coupled line notch filter 20, and to the anode by the shunt stubs of the low pass filter. Thus a short circuit exists across the diode at the image frequency that provides reduced conversion loss. The frequency components of the current through the diode is given by (5.113) and reproduced in (6.1).

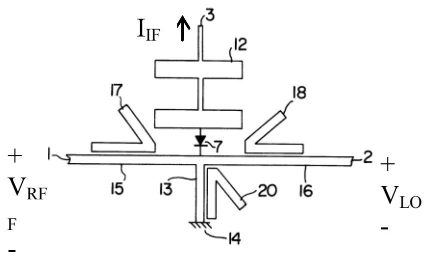



**Figure 6.10** Narrowband single ended mixer.

$$I_d(t) = -\sum_{n=-\infty}^{n=\infty}\sum_{m=-\infty}^{m=\infty} i_{nm}(-1)^{n+m} e^{(n\omega_{LO}+m\omega_{RF})t} \tag{6.1}$$

The second patent is in Figure 6.11 and describes a similar single ended diode microstrip mixer [14]. Its goal is to provide reduced conversion loss over a wider bandwidth. The RF input is at 1, and couples to the diode via microstrip line 2. The LO input is at 4, and couples to the diode first through the edge coupled bandpass filter 5, and then through microstrip line 2.

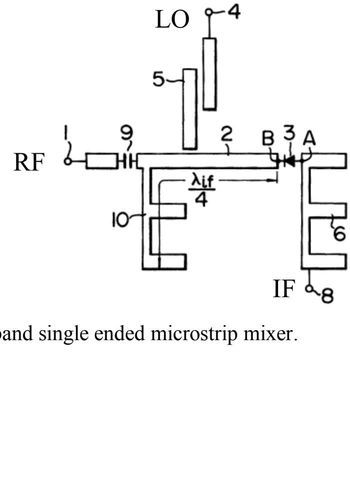



**Figure 6.11** Wideband single ended microstrip mixer.

The RF and LO ground returns are provided by the open circuited stub at A. The IF port is node 8 and connects to the diode's anode via low pass filter 6. The IF ground return comprises the combined lengths of microstrip line 2 and open circuited microstrip line 10, which totals a quarter wave long at the IF frequency. The capacitor 9 blocks the lower IF frequency from exiting through RF port 1, and decouples the RF load from affecting the IF ground return.

## 6.2 SINGLY BALANCED

### 6.2.1 Transformer-Less

A mixer design with the goal of reducing complexity and cost was disclosed by Podell in 1962 [15], which does not need a transformer to provide isolation. Instead, the signal and LO sources, respectively, have low and high impedances, or vice versa, and were applied to the diodes as shown in Figure 6.12. The low impedance source, $F_2$, could comprise a cathode (or emitter) follower stage; and the high impedance source, $F_1$, could comprise an amplifier having a high impedance output. The bias resistor 7 is larger than other resistors in the circuit. The mixer operation is explained by assuming the LO is assigned to $F_1$ and is applied at the point where the two diode cathodes are connected together. Harmonic balance simulation of the circuit with no bias applied shows, for example, with 50 ohm RF source impedance at F2, 50 ohm IF load resistor, and 300 ohm LO source impedance at F1 with LO power at +10 dBm, conversion loss is 13 dB, LO to IF isolation is 17 dB, and RF to IF isolation is 11 dB. Silicon and GaAs diode models were used with very similar results.

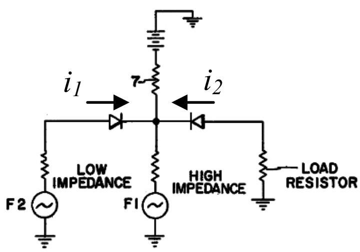



**Figure 6.12** Balanced mixer without a transformer.

The two silicon diodes switch on and off according to the polarity of F1 LO, with LO ground current passing through the IF load resistor on the right and the low impedance of $F_2$ on the left. The voltage-divided signal from $F_2$ is present at the diode cathode connection point when the diodes are switched on, and disconnected with diodes off. The resulting waveform at the IF output load includes the desired IF and other mixing products. The following equations apply to the circuit. Assuming positive polarities from sources F1 and F2, the RF signal is in phase with current $i_1$ and LO is in counter phase. In contrast, LO and RF are in counter phase with current $i_2$. The diode currents are developed in terms of Fourier series as discussed in Chapter 5, according to the polarities of Figure 6.13. Equation (6.4) suggests mixing products with even RF harmonics are suppressed, while those with odd RF harmonics are present. However, given the lack of actual balance in the RF circuit of Figure 6.13(b), suppression of even RF harmonics will be poor. The frequency components of the diode currents are developed in (6.2) and (6.3). To simplify notation, just the phase components of the IF output current are given by (6.4), showing which mixing products cancel.

$$i_1 = \sum_n \sum_m i_{nm} e^{jn(\omega_{LO}t+\pi)} e^{jm\omega_{RF}t} = \sum_n \sum_m i_{nm} (-1)^n e^{jn(\omega_{LO}t)} e^{jm\omega_{RF}t} \tag{6.2}$$

$$i_2 = \sum_n \sum_m i_{nm} e^{jn(\omega_{LO}t+\pi)} e^{jm(\omega_{RF}t+\pi)} = \sum_n \sum_m i_{nm} (-1)^{n+m} e^{jn(\omega_{LO}t)} e^{jm(\omega_{RF}t)} \tag{6.3}$$

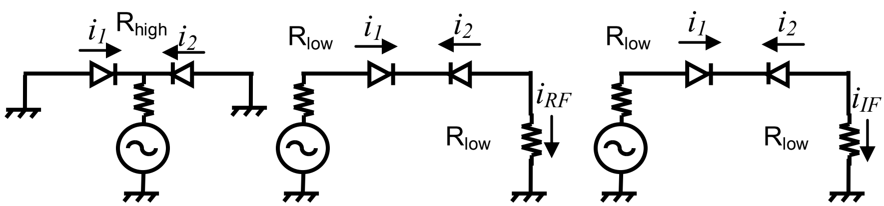



**Figure 6.13** Equivalent circuits at LO, RF and IF frequencies from left to right.

$$i_{IF} = i_1 - i_2 = (-1)^n \left[1 - (-1)^m\right] \neq 0 \text{ for m=odd; =0 for m=even} \tag{6.4}$$

Another example of a singly balanced diode mixer without transformer was disclosed in 1995 and shown in Figure 6.14, using the silicon point contact diode 1N21C. The application is for cell phones where small size and low cost is required, [16]. The main feature is that the goals of the design are realized by removing the transformer.

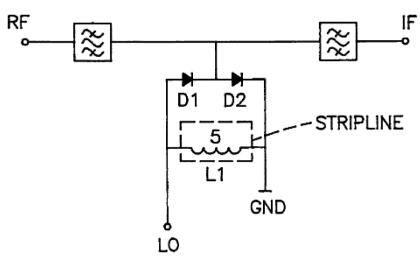



**Figure 6.14** Transformer-less diode mixer.

Instead, an inductor is placed across the diodes to prevent the build up of DC voltage across the two diodes, and at their connection point. The inductance L1 presents a high impedance at the LO frequency, so L1 does not short out the LO. Performance is said to be equal to that of a balanced mixer. Assuming positive RF and LO polarities at their respective ports, the resulting IF current is shown in (6.5) to include odd RF harmonics and suppress even RF harmonics.

$$i_{IF} = i_1 - i_2 = \left[(-1)^m - 1\right] \neq 0 \text{ for m=odd; =0 for m=even} \quad (6.5)$$

### 6.2.2 Lumped Balun

Another mixer designed by Podell, disclosed in 1964, obtained wider bandwidth using transmission line transformers in contrast to designs using conventional transformers having separate primary and secondary windings [17]. The design shown in Figure 6.15 includes a 1:4 impedance ratio transmission line transformer using bifilar wire to obtain RF and LO frequency coverage of 10 - 1000 MHz. The LO source connects to point 10'; the RF source connects to point 44', and the IF output connects to point 62'. The transmission line transformer, A, ideally comprises a binocular (dual hole) ferrite core having two wires passing through each hole. Each pair of wires comprises a transmission line, both of which are connected in parallel on the left at the LO input and connected in series on the right at the diodes. The characteristic impedance of the bifilar transmission lines should be the geometric mean of the input and output impedances. An analysis of the currents and voltages present in the transformer indicates the impedance presented to the diodes on the right side of the core is four times the impedance presented by the transformer to the LO source: the voltage on the right is double that on the left, and the current on the right is half that on the left. The transformer presents the LO voltage to the series connected diodes 36' and 34', making them switch on and off simultaneously.

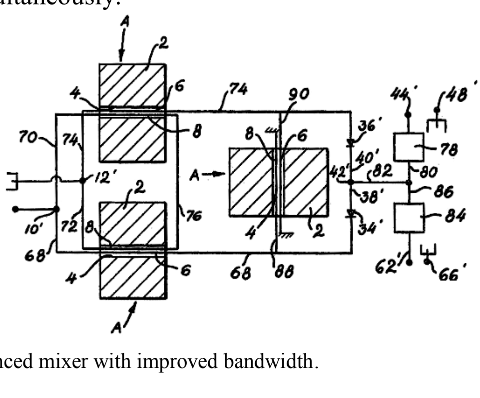



**Figure 6.15** Singly balanced mixer with improved bandwidth.

The RF signal is presented with opposed phase to the diodes at the common point 42'. The IF current is collected at the common point, and then passes through the diplexer element 84 that isolates the RF and IF ports.The RF and IF ground returns are provided by the ferrite transformer, 2, whic uses bifilar wires 6 and 8 operating in the odd mode to provide low impedance paths for the RF and IF currents. In contrast, the mutual inductance of wires 6 and 8 in the core causes a high impedance to the even-mode generated in the wires by the LO signal, keeping the LO from shorting out. To maximize the RF and IF bandwidths, the grounded ends of wires 6 and 8 should connect to ground physically near points 48' and 66'. A self biased version of the mixer of Figure 6.14 is also disclosed.

The mixer of Figure 6.16 disclosed in 1989, comprises two coaxial transmission lines, each having a diode connected in series with the center conductor and located midway along the length of each line [18]. The circuit is similar to that in Figure 6.15, but the implementation is different. Instead of using bifilar wires, this mixer uses coaxial transmission lines surrounded by ferrite material to increase inductance of the outer conductor. The design is reported as having 3.5 to 4.5 dB conversion loss over the 50 – 550 MHz cable band, with nearly a full octave of LO bandwidth. Isolation between LO-RF and LO-IF is 40 dB, input intercept point (IIP3) is at least +20 dBm, with a minimum LO power of +10 dBm. The coaxial lines are much less than a quarter wave long at the LO and RF frequencies. They are connected together at each end: in phase at one end, and opposing phase at the other. The in-phase end has the outer conductors tied together (60) and the inner conductors tied together (38), while at the opposed phase end the inner conductors of both transmission lines respectively tie to the outer conductors of the other transmission line. The LO port is at the opposed phase end, and the RF/IF port is at the in phase end, with the circuit providing isolation between the LO and RF/IF ports. The RF/IF port connects to a diplex circuit (58) that provides isolation between the RF and IF ports. The diodes are in series, with the cathode of one (40) connected to the anode of the other (42). This arrangement causes the RF signal to be applied in opposing phase to the diodes, and the LO signal to be applied in phase. The RF/IF ground return to node 60 is provided by the inside surface of the outer conductor of the coaxial lines. Thus shorter coaxial lines equate to higher RF/IF frequency range. However, shorter transmission lines also tend to short out the LO, raising the upper end of the LO bandwidth. This effect is reduced by increasing the inductance of the outer conductors by surrounding the coaxial lines with ferrite material.

According to the LO phase in Figure 6.16, a negative LO voltage applied to node 72 turns on diode 42, and a positive LO voltage applied to node 70 turns on diode 40; thus, the LO is applied in phase to both diodes and they switch on and off simultaneously. The ferrite material surrounding the transmission lines increases inductance to ground between points 72 - 60 and 70 - 60, thus increasing the low end of the LO bandwidth. And the transmission line action of coaxial lines 44 and 46 enforces equal magnitude currents to flow in opposite directions in the respective inner and outer conductors. This isolates the RF/IF port from the LO port because ideally no net RF/IF current leaves nodes 70 or 72. Similarly, the LO is isolated from the RF/IF port because ideally no net LO current leaves node 38 and the ground point 60 causes a virtual ground to the LO at the RF/IF port .

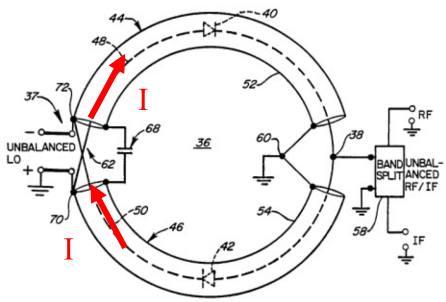



**Figure 6.16** The mixer comprises two diodes in line with the center conductor of the coaxial transmission lines. The mixer was intended for cable TV applications.

### 6.2.3 H-Mixer

The mixer of Figure 6.17, is based on the H Mixer described by Saleh [19] and was disclosed in the invention reported in [20]. Its goals are to minimize conversion loss and maximize $IIP_3$. The improved dynamic range is due to shaping the LO into a square wave to switch the diodes quickly with predominantly odd harmonics. The circuit elements that shape the LO waveform are narrow band, thus the tradeoff is reduced bandwidth.

The square wave LO voltage is shaped by allowing only the odd harmonics of the LO voltage to exist across the diode. The even harmonics are shorted by lines 115 and 115a, which are quarter wave long at the LO frequency and shorted to ground. The lines present an open circuit to the fundamental LO and other odd harmonics, allowing these voltages to exist across the diodes. Further efficiency is obtained by recovering mixing products, including energy from the sum frequency, by open circuiting all mixing products other than the RF, image, LO, and IF. The open circuit to diode currents is provided by the series resonant filter composed of inductor 110 and capacitor 112. Self bias is provided by routing the rectified LO current through resistor $R_{95a}$. Appropriately adjusting the self bias voltage level allows for the optimum tradeoff between LO drive level, conversion loss, and $IIP_3$. Also, circuit balance is improved by forcing the diodes to share the same current.

The hybrid 104 interfaces the RF and LO inputs to the diodes. The hybrid can be either a 180° or 90° type. In the 180° version, the RF (signal) at frequency $\omega_s$, is incident at the sum port of the hybrid, and the LO (pump), at frequency $\omega_p$, is incident at the difference port. Thus the RF is applied in phase, and the LO is applied 180° out of phase, to the two diodes. The RF and LO ground returns are provided by capacitors $C_{122}$. The IF load is represented by resistor $R_{93a}$, which is AC coupled to the diodes. The IF ground return is provided by lines 115 and 115a.

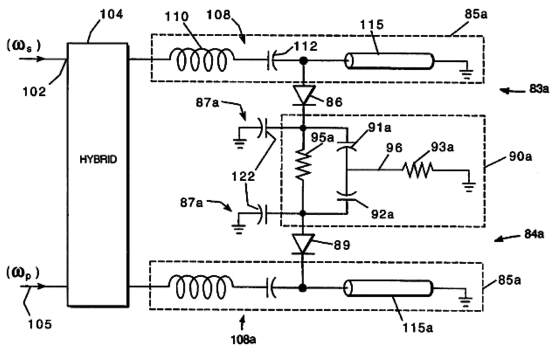



**Figure 6.17** High intercept band limited mixer for operation in the 800 MHz cellular band.

A specific embodiment of the mixer is disclosed that is said to be suitable for cellular bands between 869 – 894 MHz. Component values are: $C_{112}$ = 1 pF; $L_{110}$ = 33 nH; Lines 115 and 115a are 1.42 inches with dielectric constant of 9.6; $C_{122}$ = 25 pF, $R_{95a}$ = 240 ohms; $C_{91}$ = $C_{92a}$ = 20,000 pF; and the load resistor is 50 ohms.

### 6.2.4 Multilayer Ceramic Substrate

The reduced size mixer of Figure 6.18 was disclosed in 2007 [21]. It uses a multilayer ceramic substrate having an integrated Marchand balun for LO, a high pass filter for RF, and a low pass filter for IF. The two diodes are mounted on the top side of the ceramic assembly. The circuit operates with RF at 2 GHz and IF at 50 MHz. The Marchand balun provides the RF and IF ground returns, through the grounded ends of the two balanced lines.

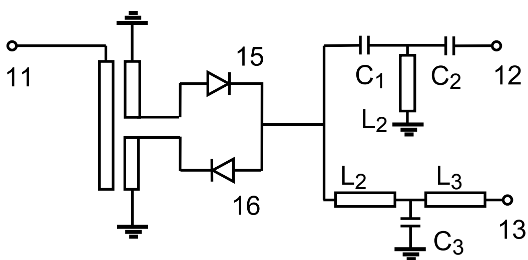



**Figure 6.18** Small sized mixer using a multilayer substrate.

The balun, the high pass filter, and the low pass filter are formed from transmission lines integrated into a stack of multilayer substrates. The mixer diodes are also integrated within the stripline circuit. The Marchand balun is obtained by using two meandered broadside coupled transmission lines.

### 6.2.5 Waveguide and Crossbar

Microwave mixers were first realized using waveguide. Figure 6.19 shows a singly balanced waveguide mixer disclosed in 1946, that uses two rectifying crystals located at opposite ends of a waveguide [22]. The mixer functions as a suppressed carrier modulator for a radio communications system. Compared with earlier designs, its goal is to provide simplified impedance matching, wider bandwidth, and lower distortion. The LO, or carrier wave, having frequency $\omega_c$ is incident at waveguide 13. The RF signal, at frequency $\omega_m$, is applied to the input of transformer 28. The modulated upper and lower sidebands, having frequencies $\omega_c \pm \omega_m$, exit the waveguide with the 90° bend, which is shown on the bottom side of Figure 6.18 but actually extends into the page at point 12.

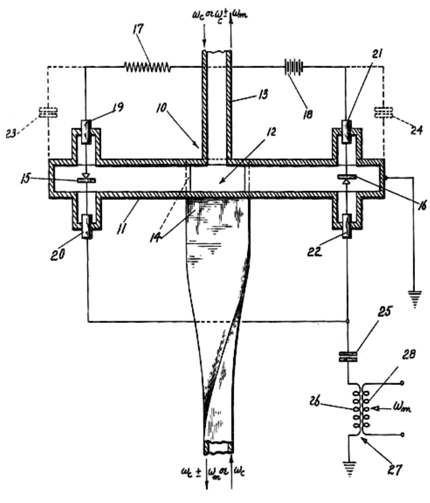



**Figure 6.19** Waveguide suppressed carrier modulator.

The LO electric field (E-field) propagates in waveguide 13 toward the junction at point 12, where it splits into two waves, respectively, traveling toward diodes 15 and 16. The junction at point 12 causes the two E-fields traveling toward the diodes to be 180° out of phase with each other. In contrast, the RF signal is applied in phase to the two diodes. The LO signal switches the diodes on and off to modulate the RF signal, which in turn produces the upper and lower modulated sideband signals in both diodes.

The sideband signals are produced with the same polarity in the diodes, so their resulting E-fields propagate from the diodes toward the junction at point 12 with the same phase. They combine at the junction point 12 and propagate down the waveguide with the 90° bend (into the page). Recalling the LO carrier E-fields traveling toward the diodes are 180° out of phase, they cancel at point 12 and thus are suppressed from exiting along with the sideband E-fields. Bias can be applied to optimize the conversion efficiency of the two crystal rectifiers.

An improved waveguide mixer was disclosed in 1948 [23]. The major innovation of the mixer, shown in Figure 6.20, is locating the two diodes next to each other, thus reducing size, increasing bandwidth, and improving circuit balance. The diodes are in series across the E-plane of the waveguide in which the RF signal propagates. As shown in Figure 6.20, the cathodes of the diodes are AC connected together, separated by two series capacitors.

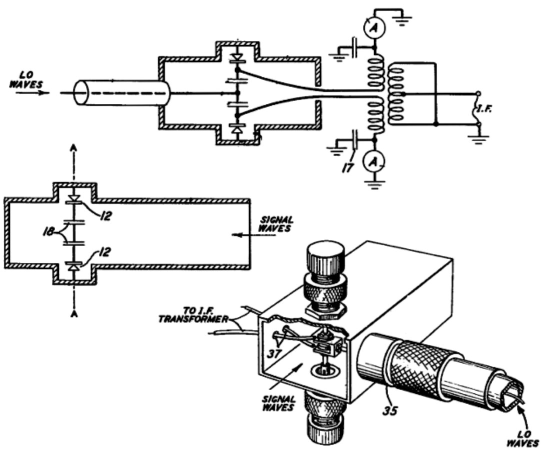
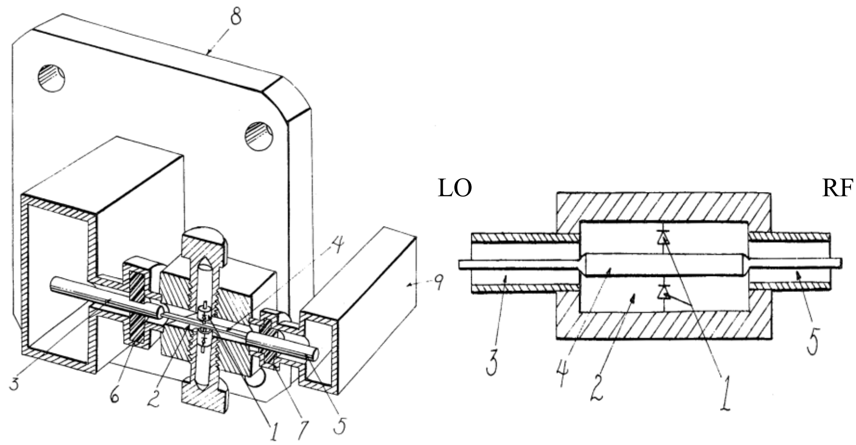



**Figure 6.20** Balanced crystal microwave converter.

The RF voltage signal is in phase with one diode, and 180° out of phase with the other. The RF current passes through the two diodes, and the two series capacitors that also pass the LO but block the IF. The IF is DC connected differentially across the two diode cathodes using wires having inductance that presents a high impedance to the RF signal. The LO is AC coupled in-phase to the diode cathodes through the blocking capacitors. Thus circuit balance isolates the LO from the RF and IF. And the RF and IF are isolated from each other by means of filtering. The magic-T hybrid waveguide patent of Tyrell is referenced [24].

Interest in waveguide as a medium for microwave mixers waned as stripline and microstrip became popular. However, waveguide construction continues to be important for millimeter-wave mixers. A millimeter-wave waveguide mixer, shown in Figures 6.21(a), and (b), was disclosed in 1967 [25]. The goals of the design were wide bandwidth and low conversion loss for an up-converter. The innovation was to place the diodes at the intersection between two coaxial lines and a waveguide that connect the RF, LO, and IF to the diodes.




(a) (b)

**Figure 6.21** Waveguide mixer: (a) LO signal in the waveguide on the left couples to coaxial line 3, and RF signal in the waveguide on the right couples to coaxial line 5. Coaxial filters 6, 7 provide isolation between RF and LO ports. (b) detail showing diode polarity: top and bottom diodes, respectively, have anode and cathode connected to the coaxial line thus are 180° out of phase for LO and RF signals. A simple mathematical analysis demonstrates the IF signals are generated in phase and propagate in the waveguide 2.

This minimizes the signal connection lengths, which in turn reduces the periodicity of mismatch ripples. It also minimizes the lengths of the RF and LO ground returns, which maximizes the upper bound of their bandwidths. The RF signal from the waveguide is coupled to a coaxial line 3 that passes through filter 6 to reject LO leakage. The line continues through the side wall of the waveguide where it transitions to an air-dielectric stripline consisting of cavity 2 with a metallic slab line. The diodes are attached to the slab and grounded to the cavity walls.

The LO is incident at coaxial line 3 and delivered to the diodes, while passing through filter 6 to reject RF signal. Conversely the RF signal is incident at coaxial line 5, and passes through filter 7 that rejects the LO signal. Both RF and LO signals excite the diodes with 180° phase difference since the diodes are in opposed direction with respect to the electric field in cavity 2. The circuit operates as an upconverter so that IF signals generated by the diodes are higher in frequency than the RF input, and are in phase with each other and propagate into the waveguide 2 having been excited by line 4. The diodes are located in the center of the waveguide at the point of maximum E-field, and the up converted upper and lower sideband signals excite a wave that travels in the waveguide 2 toward the viewer of Figure 6.21(b). A waveguide back short also extends into the page, with length equal to a quarter wave at the IF operating band to provide a high impedance across the diodes to maximize conversion efficiency. Figure 6.21(a) depicts waveguides associated with coaxial lines 3 and 5, but a waveguide-to-coaxial transition is not required.

As seen in Figure 6.21(b), the RF and LO connections to the diodes provide immediate ground returns via the inside walls of waveguide that also forms the ground walls of the striplines. The RF and LO signals propagate from the coaxial lines 3 and 5 in TEM mode while the IF propagates out the IF waveguide 2 in TE mode, with the mode orthogonality providing inherent isolation between IF and RF/LO. The cutoff frequency of the IF waveguide is above the operating band of the RF signal, which provides additional RF-IF isolation, and improves conversion efficiency.

An improved mixer design was also disclosed by the same inventor in 1969 [26]. As is evident in Figure 6.22, it has many of the same features as the mixer of Figure 6.21. It places the diodes together at the intersection of the waveguide and coaxial line connections to minimize conversion loss (both absolute and ripple), and maximize bandwidth. But it represents an improvement by its ability to operate over certain frequency combinations not allowed by the mixer of Figure 6.21. The limitation is due to the reflected impedances from the RF and LO filters that short out the diodes at certain RF and LO frequencies. This design solves this problem by changing the means of diplexing the two signals. The innovation is to use one TEM mode transmission line to connect two signals to the diodes, and a waveguide transmission line to connect the third signal to the diodes. The two signals are diplexed onto a coaxial line, with one signal connected via a stepped impedance coaxial low pass filter at point 3. The second signal is connected at point 5 via a waveguide operating as a high pass filter with its cutoff frequency above the band of the first signal. Thus the two signals are isolated from each other, and connected to the diodes, without the presence of impedance reflections that would otherwise limit frequency response. Since the RF and LO signals, as well as the diodes, are coupled to a single bar, this type of structure is known as a crossbar mixer.

The mixer can be used either as an up or down converter. If used as a down converter, the RF signal could be connected to the waveguide 5, which induces the RF signal current onto the coaxial line, which in turn flows to the diodes. The LO is connected to the coaxial line at point 3, and operates lower in frequency than the RF. The coaxial line passes through the RF signal waveguide walls. It then passes through the wall of the second waveguide, and then transitions to stripline before connecting to the diodes. The RF and LO currents on the stripline inside the waveguide cannot excite a waveguide mode since the stripline runs orthogonally to the E-field vector of waveguide 2. The RF and LO ground returns are immediately provided by the diode connections to the waveguide wall, which also forms the ground planes for the stripline.

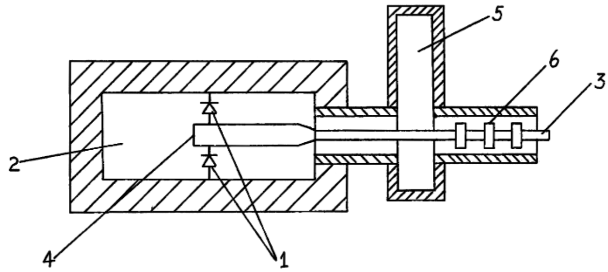



**Figure 6.22** Waveguide millimeter-wave mixer.

The resulting modulated IF diode currents excite the $TE_{10}$ waveguide mode at the point of maximum E-field in waveguide 2, causing the E-field of the IF to propagate into the waveguide toward the observer of Figure 6.22. A waveguide back short also extends into the page, with length equal to a quarter wave at the IF operating band to provide a high impedance across the diodes to maximize conversion efficiency. The mixer can also be used as an upconverter by changing the port arrangement according to the frequencies involved.

### 6.2.6 Finline and Crossbar

Waveguide and stripline/microstrip fabrication offer complementary features. Waveguide mixers can be extremely broadband and offer low conversion loss, but their tightly tolerance machined parts and packaged diodes are expensive. In contrast, stripline and microstrip mixers offer reduced cost and size, but can have higher conversion loss especially at millimeter-wave frequencies. They are lower in cost due to photolithographic fabrication and usage of beam lead or chip diodes. They offer reduced size due to a higher than unity dielectric constant. Combining conventional waveguide and microstrip techniques can provide the benefits of both while minimizing the negatives. In this patent the microstrip lines are of the suspended type, where the circuit traces are on the substrate and the ground planes are the waveguide walls.

The mixer shown in Figure 6.23 was disclosed in 1979 [27]. It is said to have improved performance and reduced cost compared with conventional printed and waveguide mixers. Schematically, it is similar to the mixer of Figures 6.21(a) and (b), but its construction is different. It comprises two machined metal blocks with a printed circuit board sandwiched between them. Two beam lead diodes connect in series, anode to cathode, to the crossbar. The other ends of the diodes are grounded. The diodes are centered on waveguide 3 at the E-field maximum.

Any diode mixer can be used as an up or downconverter, with the RF, LO, and IF connections made at any port according to their frequencies. The following describes a down converter application. The LO is incident at waveguide 2, where it's E-field induces a current in line 14, the suspended microstrip, that is a quarter wave long at the center of the LO band. The resulting current is AC connected, via printed capacitor 17, to the crossbar that in turn connects to the diodes.

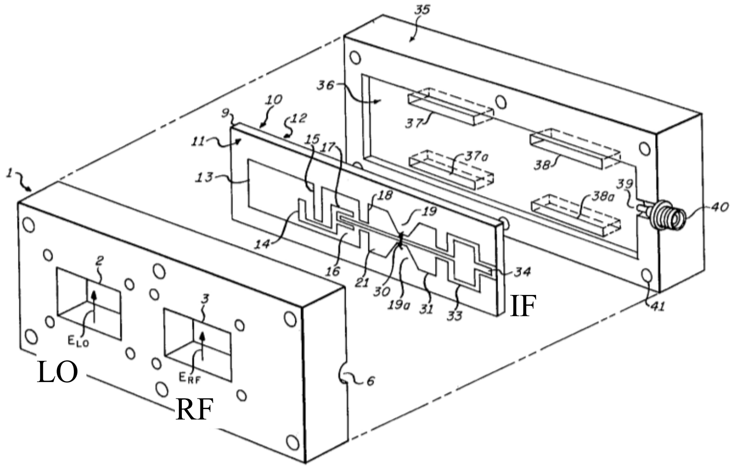


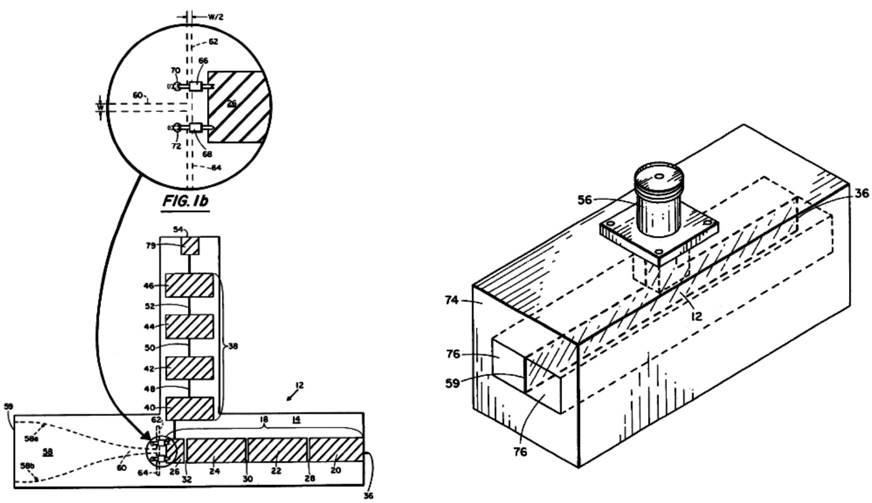

**Figure 6.23** Waveguide mixer with blocking capacitor, LO probe, IF filter and diode connections all to the crossbar printed on a suspended substrate and mechanically sandwiched between two blocks that together form a waveguide.

The LO field is from the center conductor to ground so the LO is 180° opposed in phase with one diode, and in phase with the other. The RF signal is incident at waveguide 3, with electric field applied to both diodes in series, inducing currents in the diodes that are equal in phase with both diode polarities. Back shorts are provided for the RF and LO waveguide transitions to the suspended substrate line. The large signal LO causes the diodes to modulate the RF signal input, resulting in IF currents that add in-phase at the diode connection point, and pass to the line on the right of the diodes. Isolation is provided between the LO and IF ports by the printed capacitor 17 that passes the LO but blocks the IF; and, the printed lowpass filter 33 that blocks the LO but passes the IF. The RF and LO ground returns are provided immediately via the top and bottom diode connections to the ground conductors at 19 and 19a. The mixer of Figure 6.23 was built using HP 5082-2299 beam lead diodes. With RF frequency at 11.8 GHz, DSB noise figure measured 6.5 dB, including 2.6 dB of downstream IF noise figure contribution.

Finline is also used to transition between waveguide and printed microstrip or stripline transmission lines. Finline comprises fin shaped metalized regions that gradually transition without the need for a quarter wave long back short and thus can be very wideband. The mixer of Figure 6.24, disclosed in

1981, is similar schematically to that of Figure 6.15, in that the RF and LO do not coexist on the same transmission line, but different in terms of fabrication [28]. It has one waveguide port, and two microstrip ports. The RF signal is incident at waveguide port 76, where the E-field excites the finline 58 with field vectors beginning at metallization 58b and ending at 58a, creating an odd-mode that transitions into slotline 60. The finline traces are metalized onto the back side of the soft dielectric substrate. Slotline 60 terminates in slotlines 62 and 64, that are connected in series across slotline 60, and ideally have widths approximately half that of slotline 60. The lengths of slotlines 62 and 64 are both equal to a quarter wave at the center of the RF operating band.

Two GaAs beamlead diodes are connected in series anode-to-cathode across the end of slotline 60. They are located on the top side of the substrate and connect across slotline 60 via holes in the substrate. The anode-to-cathode connection point of the diodes also connects to a section of microstrip, 26, with length that is preferably less than one tenth of the LO's wavelength. This microstrip line is diplex connected to the IF and RF, with the IF connection via a low pass filter, and the RF connection via a band pass filter. The low pass filter is composed of series sections of high and low impedance microstrip lines, with a high impedance line connected to microstrip 26 to reject the RF. The bandpass filter comprises microstrip line resonators that are capacitively end coupled. Microstrip 26 is kept relatively short so the high impedance provided to the IF by the first end-coupled capacitor does not appear as a short in the IF band.




(a) PCB circuitry. (b) Waveguide mounted circuitry.

**Figure 6.24** Finline and microstrip millimeter-wave mixer.

The LO currents run through the series diodes, with a virtual ground at the anode-to-cathode connection, which provides LO-IF and LO-RF isolation. As with singly balanced mixers in general, this circuit balance also provides rejection to AM noise on the LO, and rejection of second order intermodulation products at the RF port. Slotlines 62 and 64 keep slotline 60 from shorting out at the diodes, and essentially are the equivalent of the waveguide back short in conventional waveguide. The RF and IF ground returns are provided by the diode connection at slotline 60, which also connect to the backside metallization under the RF and IF microstrip lines. These currents have to run around the ends of slot lines 62 and 64, which limits the upper end of the RF bandwidth, but should not degrade bandwidth of the lower frequency IF.

Another finline mixer is shown in Figure 6.25 that was disclosed in 1983 [29]. Like the mixer of Figure 6.24, it has two metal blocks, between which is sandwiched a printed substrate. All mixing elements are contained within the crossbar printed line. In contrast, it has two waveguide ports and one microstrip port. It is similar schematically to the mixer of Figures 6.21(a) and (b). The LO is incident at the finline waveguide port, which transitions first to a balanced microstrip line, and then to a single ended microstrip line. The length of the finline transition and balanced microstrip is at least three wavelengths at the LO center band. Ground metallization begins on the back side of the substrate at the demarcation between balanced and single ended microstrip. The microstrip line then connects to the two diodes at their anode-to-cathode connection point, so the LO excites the diodes 180 degrees out of phase.

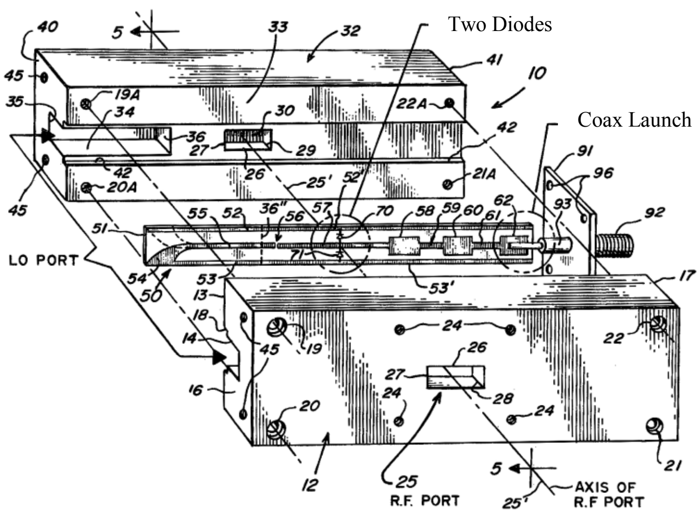
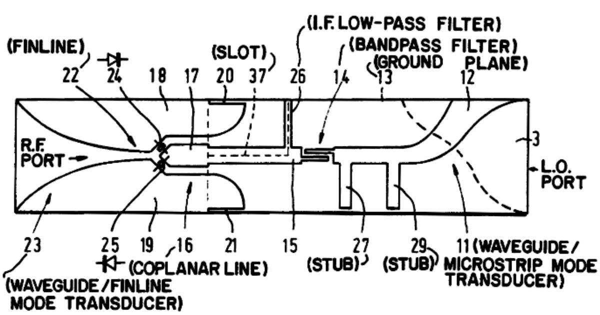



**Figure 6.25** Complete finline millimeter-wave mixer.

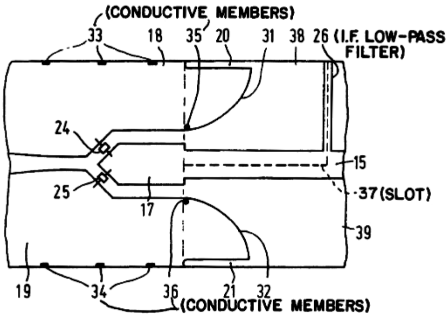
The diodes are connected in series, and located at the center of the RF waveguide at the point of maximum E-field. The RF signal voltage is incident across the diodes, exciting them in equal phase. A quarter wave waveguide back short is located behind the diodes, to provide a voltage maximum across the diodes at the center of the RF frequency range. A rectangular opening is made in the substrate's ground plane to facilitate the back short. The isolation between RF-LO and RF-IF is provided by the orthogonality between LO and IF microstrip lines and the RF waveguide. LO-IF isolation is provided by the diplexer, which is composed of the blocking capacitor 56 in the LO microstrip, and the printed low pass filter in the IF line. Typical operating frequency ranges are given as: RF & LO at 26 to 100 GHz, and IF at DC to 14 GHz.

Another finline mixer disclosed in 1983 is shown in Figure 6.26 [30]. This example makes use of finlines, slotlines, microstrip and coplanar lines printed in both sides of substrate. Like the mixer of Figure 6.24, the diodes are series connected across the slotline port, with the diode anode-to-cathode connection point diplex connected to the other two ports. In contrast, the back side slotline arrangement is different, and the high frequency diplexed port uses finline instead of microstrip. Like the mixer of Figure 6.25, this mixer has two waveguide ports, but both are finline and there is no waveguide back short.

The RF signal is incident at the waveguide port with the finline to slotline transition that ends with the two diodes series connected across the slotline. The anode-to-cathode connection point of the diodes connects to the center conductor of a coplanar waveguide (CPW). The CPW line is a quarter wave long at the RF center frequency. The end of the CPW opposite the diodes transitions to microstrip. At the demarcation between CPW and microstrip, backside metallization abruptly begins on the substrate to provide the microstrip ground plane. At this same point, the grounded sides of the CPW form capacitive stubs that couple the CPW ground current to the microstrip ground, to facilitate the LO and IF ground returns. These capacitive stubs, 31, 32, acting together comprise a short circuit to the slotline mode (odd-mode) of the CPW. The quarter wave length of the CPW causes this short circuit to reflect back as an open circuit across the diodes, to maximize RF voltage.




(a) Full mixer circuit




(b) Diode detail

**Figure 6.26** Layout of finline millimeter-wave mixer on microwave flexible substrate.

The LO signal is incident at the other waveguide port, which has a finline to microstrip transition. In contrast to the mixer of Figure 6.15, this finline transitions to single ended microstrip without passing through an intermediary balanced microstrip line. The microstrip line connects to a bandpass filter comprising two shorted stubs, 27-29, each a quarter wave long at the LO frequency and spaced at a quarter wave, which are present to short out the IF to improve LO-IF isolation. The IF low pass filter comprises five transmission line sections, alternating between high and low impedance, that are all a quarter wave long at the RF and LO frequencies. The first high impedance section 26 is shown in Figure 6.25 and the other sections are contained in another substrate not shown in the figure.

### 6.2.7 Stripline and Microstrip (Single Sided)

A major goal of mixer designs is to obtain a single sided circuit (i.e., one having only top-side metal patterning, and with unpatterned ground plane on the bottom side). This approach is desirable not only for MMIC realizations but also for thick film, thin film, and low cost microwave soft susbtrates.

The stripline mixer shown in Figure 6.27, which uses a ratrace magic-T hybrid, was disclosed in 1963 [31]. The application of the mixer was for an airborne FM altimeter using two antennas, one for transmit and the other for receive. The antenna connectors are orthogonal to the plane of the stripline, and extend into the page in Figure 6.27. The operating frequency was centered at 1.6 GHz, and swept over 50 MHz. The ratrace hybrid has four ports that are arranged so the "sum" port connects to the two diode ports in phase, and the "difference" port connects to the two diode ports 180° out of phase. The RF and LO, respectively, can be connected to either the sum or difference ports. RF and LO power is delivered to the diodes, and the RF and LO ports are isolated from each other due to cancellation.

In the mixer of Figure 6.27, the LO is connected to port 34a, which in turn is connected via a microstrip line to the sum port at point 40. The sum port connects to each diode port via a quarter-wave long microstrip line that maintains an equal phase shift to both diode ports. The RF signal connects to port 42a, and then to the difference port at point 44. The difference port connects to diode port 50 via a quarter-wave line, and to diode port 52 via a three-quarter-wave long line. Thus the difference port connects to the two diode ports 180° out of phase. Adding the lengths of the sum and difference port lines, the circumference of the ratrace hybrid is seen to be 1.5 wavelengths long at band center, which is the standard dimension for a ratrace. The diodes 54 and 56 are separated from the diode ports of the ratrace hybrid via a microstrip line that is about a half wavelength long at the operating frequency, 1.6 GHz. Matching stubs 80 and 82 are shorted to ground to provide impedance matching and the ground return path for the IF. The IF is coupled from the diodes at points 64 and 66 via a lowpass filter, which in turn connect to an audio transformer (not shown) at points 72 and 74. The open circuit shunt stubs 58a and 58b provide the RF and LO ground returns. They are short enough to provide a high impedance to the IF output, so RF ground return is accomplished without loading down the IF.

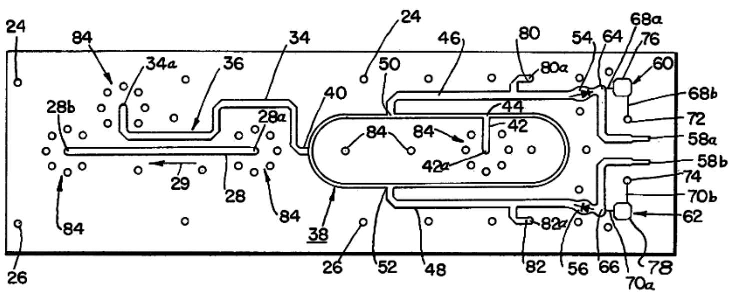



**Figure 6.27** Stripline balanced mixer. Port connections: LO at 34a, Rx antenna at 42a, IF across 72-74, TX antenna at 28b, and TX at 28a.

The mixer of Figure 6.28 disclosed in 1970, is an improved ratrace mixer said to have reduced noise figure and delay distortion, [32]. The trade-off for the improved conversion efficiency is the small operating bandwidth of about fifteen percent. The RF is connected to the sum port at point 14. The LO is connected at the difference port via microstrip 19. The RF and LO ground returns are accomplished with open-circuited quarter-wave stubs 34a, b. The open-circuited stubs 35a, b provide a low impedance ground return for the sum and second harmonic signals, enhancing efficiency without loading down the IF signal. The IF output is taken by summing in-phase the outputs of the two diodes through a lowpass filter comprising high impedance lines 37a, b that provide a high impedance to the RF and LO, followed by microstrip line 38 that gives low impedance to RF and LO. The IF ground return is obtained through the short-circuit quarter-wave lines 33a, b, which provide a low impedance to the IF, and high impedance to the RF and LO. The one-eighth wave long stubs 31a, b and 32a, b are harmonic suppression filters that further improve conversion efficiency. The in-phase port capacitive coupled at point 21 is an auxiliary port that can be used for circuit monitoring.

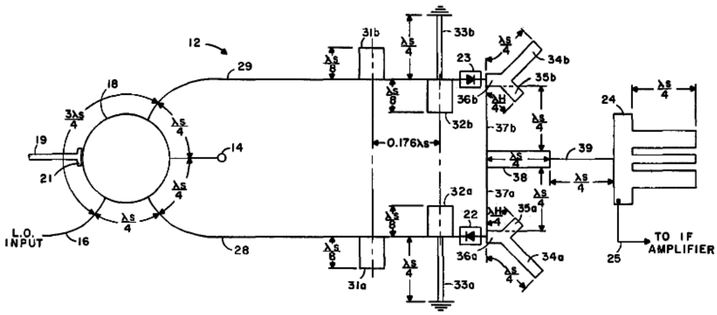



**Figure 6.28** 1970 Ratrace on microstrip or stripline.

One difficulty with the ratrace hybrid is the sum or difference port can be surrounded (blocked) by the two IF lines, requiring a microstrip crossover for the RF or LO port. The mixer of Figure 6.28 overcomes this problem by using an orthogonal connector directed into the page to obviate the need for the crossover. The practical realization of this mixer is in Figure 6.29 where the RF is connected to the sum port via a microstrip to waveguide transition at point 14. The LO is applied via microstrip16.

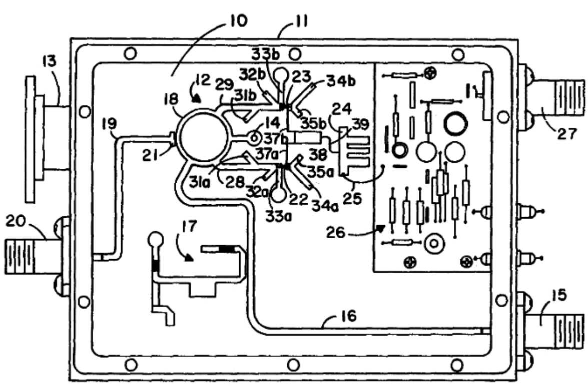



**Figure 6.29** Layout of microstrip ratrace mixer at bottom side of block converter.

Another difficulty is the relatively large size of the ratrace hybrid. The mixer of Figure 6.30, disclosed in 1979, overcomes both difficulties by using only half the ratrace ring [33]. The mixer is disclosed as having the RF connected to line 16, the LO connected to line 15B, and the IF connected to line 18. The RF is incident at the anode-to-cathode connection point of the two diodes. The opposite ends of the diodes are connected to the ends of a microstrip line that is one half wavelength long at band center. The LO is capacitive coupled to one end of the half wavelength line, and the IF is connected via a lowpass filter to the center point of the half wavelength line, thus forming a diplex filter.

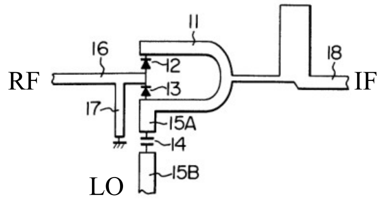



**Figure 6.30** Reduced ratrace mixer has a simple circuit layout but operating bandwidth is only about 5 % of the LO frequency.

The half wavelength line provides a 180° phase shift at center frequency between the anode of diode 13 and the cathode of diode 12. This boundary condition provides the LO ground return, and the virtual ground at the diode connection point that isolates the RF and LO ports from each other. The high impedance line of the IF lowpass filter and its subsequent low impedance open-circuit stub are both sized to present a high impedance to the RF and LO signals, and a low impedance to the IF. The IF ground return is provided by the short circuited stub 17 that is a quarter wave long at the RF and LO. The line is a low impedance to the low frequency IF, and a high impedance to the RF and LO. The capacitor at 14 is a high impedance to the IF, thus it keeps the IF current from leaking out the LO input.

The next circuit in Figure 6.31, disclosed in 1986, can be considered another variant of the ratrace mixer [34]. The goal of this design is further size reduction by combining the RF preselection filter and the mixer balun. Another goal is to provide a mixer with a high frequency IF output. The RF is connected at point 1, the LO is at point 13, and the IF is at point 16.

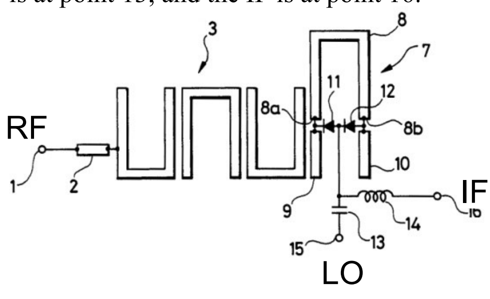



**Figure 6.31** Reduced ratrace mixer with differential mode for RF signals and common mode for the LO. The mixer is bandwidth limited by the frequency dependency of line 8.

The RF signal passes through the bandpass filter 3, and is coupled into the line 8. Line 8 is approximately a half wave long at center frequency, thus it provides the ground return path for the RF current. The LO is capacitive coupled to the mixer at the anode-to-cathode diode connection point. The LO ground returns are provided by open-circuit stubs 9 and 10, which are a quarter-wave long at the LO frequency. The LO and IF are isolated by means of diplex filtering from capacitor 13 and inductor 14. The RF is isolated from LO and IF by circuit balance. The IF return is not shown and can be accomplished by an inductor to ground.

A single sided mixer, shown in Figure 6.32, using a Marchand balun instead of a ratrace hybrid was disclosed in 1990 [35]. It is built using low cost epoxy glass substrate material that has substantial insertion loss at higher microwave frequencies. The mixer is able to operate up to about 10 GHz with low conversion loss by using a section of low loss coaxial line, soldered to the substrate to function as a Marchand balun. The low loss coaxial cable comprises two quarter wave sections, $L_1$ and $L_2$. The center conductors of $L_1$ and $L_2$ are connected in series, but their outer conductors are separated from each other at the middle of their combined length. The RF is connected to the center conductor at the left end of $L_1$, and the center conductor at the right end of $L_2$ is open with the outside conductor shorted to ground. Since $L_2$ is a quarter wave long, the short circuit at the right end of $L_2$ causes a voltage maximum at the outer conductor at the left end of $L_2$. And since the current entering $L_2$ equals the current exiting $L_1$, the voltage at the right end of $L_1$ is substantially equal and opposite to that at the left end of $L_1$. The two voltages would be exactly equal and opposite if the impedance at the RF input were zero ohms, but the impedance is presumably 50 ohms, which is still low relative to that at the right end of $L_2$. The voltage difference across the disconnected outer conductors excites equal and opposite currents running in the microstrip lines to diodes $D_1$ and $D_2$. The field lines from the resulting odd mode in the coupled microstrip lines are not totally contained within the lossy epoxy glass substrate, thus reducing loss. Thus the RF signal is divided and directed to the two diodes 180° out of phase. The RF ground returns are composed of the stubs $S_1$ and $S_2$, and the low impedance LO connection point at capacitors $C_1$ and $C_2$. The LO is connected in-phase to the diodes, with current return provided by the open circuit stubs at the cathode ends of the diodes. The IF is connected differentially at the stubs $S_1$ and $S_2$, via inductors $CH_1$ and $CH_2$ that are high impedances to the RF and LO. The IF is blocked from leaking out the LO by capacitors $C_1$ and $C_2$. The IF ground return is provided by the ground vias at the end of the microstrip lines running under the coaxial lines. The length of the coaxial line is on the order of quarter wave length and the diameter is suggested by the author to be in the order of one-tenth of the cable length. A mixer was built with 11 mm of cable length and 1.2 mm diameter on top of a 0.8 mm thick eploxy glass substrate, operating over 4.2-4.4 GHz with LO-RF isolation of 26 dB, and NF at 10 dB.

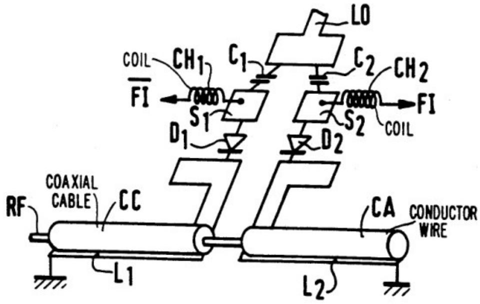



**Figure 6.32** Microstrip mixer with coaxial line Marchand balun.

### 6.2.8 Stripline and Microstrip (Double Sided)

Conventional microstrip circuits are patterned on only one side of a substrate or PCB, with the other side fully metalized and grounded. In double sided circuits, both sides of the substrate are patterned, generally to obtain balanced transmission lines. Double sided balanced mixer circuits are generally more broadband than single sided ones. The extra degree of freedom can be crucial in allowing the three circuits (RF, LO, and IF) to occupy the same physical metal while remaining isolated from each other and completing their own circuit paths without impeding the others. Double sided construction allows magic-T hybrids to be built using balanced microstrip, slotline, CPW, and combinations thereof, in contrast to single sided microstrip that relies almost exclusively on microstrip line length that is inherently narrow band. Double sided construction improves the ability to simultaneously obtain broadband balun structures and current return paths to the RF, LO, and IF.

An early circuit built with double sided construction, shown in Figure 6.33, was disclosed in 1967 [36]. It uses a Marchand balun composed of a microstrip line 34, on the top side of the substrate, crossing over a slotline located on the bottom side, to form the difference port of a magic-T. In the following discussion the LO is incident at the difference port, and the RF is incident at the sum port. The microstrip line crosses perpendicularly to the slotline, and continues to form an open circuited quarter wave shunt stub 33. The quarter wave stub provides a broadband virtual ground that effectively connects the microstrip line to one side of the slotline. The slotline opens up the ground plane under the microstrip and excites a wave in the slot. The slotline 38 divides the ground planes under the microstrip and behaves as a waveguide so that it provides a high impedance at the crossover point since it is a quarter wave length long from the top back short. Thus the frequency response of the balun is dependent on the dimensions of the microstrip open circuit stub and the slotline dimensions. Bandwidth is maximized with a low impedance microstrip stub and high impedance slotline backshort. The sum port 42 is connected across the slot line balun by microstrip lines 40 and 45, which at low frequencies short out the slot line. However, when the combined length of lines 40 and 45 are a certain length, they present a high impedance in parallel with the slotline. Under this condition the LO signal transfers from the slotline to the microstrip lines, and couples to ports 39 and 41. It is at those ports that currents from the sum (RF) port and difference (LO) ports coexist, and travel toward diodes 54 and 56 via microstrip lines 43, 44, 47, and 49, which provide impedance matching. The other side of the diodes from lines 44 and 48 are connected to low impedance open circuit shunt stubs that provide the broadband ground returns to the RF and LO currents. The IF currents generated by the diodes are summed in phase by the high impedance microstrip lines connecting to line 53 that in turn connects to the IF port. The disclosure provides equations governing the impedance relationships within the balun, based on the topology being similar to the Marchand balun. The impedance of the microstrip line 37 is $Z_a$; the impedance of the open circuit stub 33 is $Z_b$; and the impedance of slot line 38 is $Z_{ab}$. Defining R as the balanced impedance across the slotline gap, $R^2$ is selected to be approximately equal $Z_{ab}Z_b$. The LO impedance is then given by the familiar Marchand balun equation, (4.12).

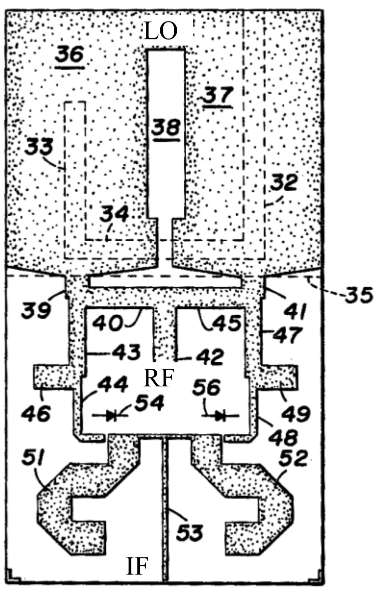



**Figure 6.33** Double-sided balanced mixer built with microstrip and slotline structure.

$$Z = r + jx = \frac{jRZ_{ab}\tan\left(\frac{\pi f}{2f_0}\right)}{R + jZ_{ab}\tan\left(\frac{\pi f}{2f_0}\right)} - jZ_b\cot\left(\frac{\pi f}{2f_0}\right) \quad (6.6)$$

Assuming the impedance of line 32 equals that of the LO generator, a matched condition is obtained when the following condition is true:

$$Z_a = \frac{R}{2}\left(1 + \frac{Z_b Z_{ab}}{R^2}\right) \quad (6.7)$$

The mixer of Figure 6.33 was built using Teflon fiberglass board and gave the following performance: Conversion loss was 8 to 10 dB over about 1.5 to 3.5 GHz, with +13 to +16 dBm LO power. LO to RF isolation measured 18 to 20 dB. Given (4.1) for input impedance to the balun as a function of frequency, and given the diodes in series to the LO and in parallel to the RF, mismatch loss was calculated at 1.89 dB for LO, and 0.76 dB for RF.

### 6.2.9 Coplanar Waveguide and Slotline

The goal of the mixer in Figure 6.34 is to minimize conversion loss, with the tradeoff being reduced bandwidth. Disclosed in 1974, this mixer uses image and sum enhancement to obtain a very low measured conversion loss of 2.6 to 3.15 dB [37]. The mixer uses slotline and CPW to form the hybrid junction where the two GaAs Schottky diodes are located, as shown in the enlarged view of Figure 6.35.

The dotted lines in Figure 6.34 represent metalized traces on the top side of the substrate, and the solid lines represent metal patterning on the bottom ground plane side. The LO input is coupled to microstrip line 26 via the directional filter ring 52. The IF output is taken directly from microstrip line 26, with isolation between LO and IF provided by the directional filter. Microstrip line 26 transitions from the top to bottom side through via 44, to connect to the left end of CPW line 22. Both diodes are located at the opposite (right) end of the CPW line, connected between the end of the CPW line and the two ground planes separated by slot line 18. The RF signal input is incident at 50 ohm microstrip line 31 and impedance transforms via microstrip line 32 to the 100 ohm slot line transition. The RF signal energy then travels along slot line 18 to appear across the diodes. Thus the diodes are connected in parallel relative to the CPW line that conveys the LO and IF signals, and they are connected in series relative to the slotline conveying the RF signal. The diodes are in parallel for both the LO and IF signals so the diode impedance is halved at these frequencies. The diodes are in series for the RF signals so the diode impedance is doubled and provide a close match to the 100 ohm slot line impedance.

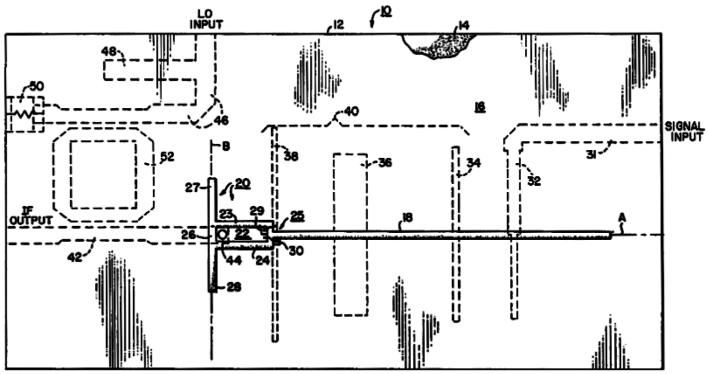



**Figure 6.34** Image and sum enhanced mixer. The solid lines indicate top side slots engraved in the metalized ground plane. The dashed lines indicate far side microstrip line circuitry.

Image and sum enhancement is accomplished by providing low impedances approaching short circuits across the diodes at these frequencies. While the low impedance allows little or no voltage to exist across the diodes at the image and sum frequencies, the associated currents can flow. Referring to Figure 6.34, the electrical lengths of $L_1$ and $L_2$ equal a quarter wave at the sum frequency, and the combined electrical length of $L_1$ and $L_2$ equals a quarter wave at the signal frequency. Thus the short circuits at the ends of the slotlines $L_1$ transform to short circuits at the diodes at the sum frequency, and they transform to open circuits at the diodes at the signal frequency.

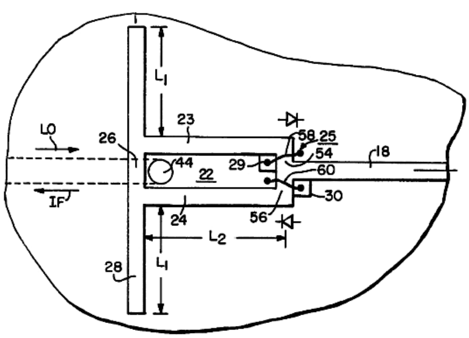



**Figure 6.35** Detail of the hybrid junction including the diodes.

A low impedance is also provided across the diodes at the image frequency by the three-element image reject filter 40, formed by the microstrip elements above the slotline 18. The measured performance is described in Table

6.1. An equivalent electrical circuit representing the mixer is found in Figure 6.36 and it is easily related to the fields located within the coplanar line 22, 23, 24.

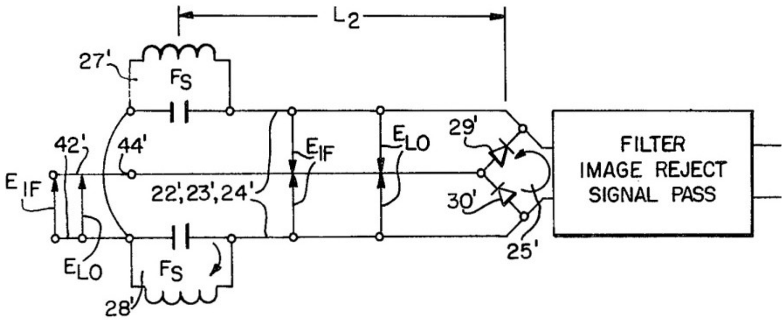



**Figure 6.36** Equivalent electrical circuit indicating the electric fields and the open circuits imposed by the short quarterwave length slotline stubs.

**Table 6.1**
Measured Data for the Mixer

| | |
|---|---|
| Signal frequency | 9.4 GHz |
| LO frequency | 7.8 GHz |
| Signal Bandwidth | 1.0 GHz |
| Conversion Loss | |
| 1.0 GHz band | 3.15 dB |
| 0.5 GHz band | 2.6 dB |
| LO Power | 40 mW |
| Input P1dB | +13 dBm |
| Image band isolation | >25 dB |
| VSWR (across signal band) | <1.4:1 |

The singly balanced mixer of Figure 6.37 was disclosed in 1985 [38] along with a doubly balanced version. It uses microstrip and CPW lines to form a hybrid junction. The mixers disclosed followed similar work published in 1980 by Ogawa et al [39, 40]. The RF and LO inputs are connected to CPW lines, with the LO CPW center conductor connecting to the common point of the diodes in the CPW mode, so the diodes are in parallel to the LO. The E-fields established in the CPW by the LO have the polarity depicted in Figure 6.37 by the solid arrows, and the fields established by the RF are depicted by the dashed arrows. The RF CPW line connects across the LO CPW line at the point where the LO CPW line necks down to have narrowed gaps and line width in the CPW going toward the diodes.

The LO CPW line going toward the LO input has much wider line and gap widths that present a high impedance back short to the two slotline modes set up by the RF CPW connection. This high impedance seen by the RF signal looking toward the LO port causes the RF energy to propagate toward the diodes, with the resulting electric fields set up so the diodes are in series to the RF and LO electric fields. LO to RF isolation is achieved because the center conductor and ground planes of the RF CPW are both connected to the grounds of the LO CPW. This circuit orthogonality allows very little of the LO voltage to become present across the RF CPW line. The RF and LO ground returns are provided by the CPW immediately where the diodes are connected, thus the upper end of the RF and LO frequency ranges are not limited by parasitics from their ground returns. The gaps in the CPW line prior to the diodes widen out to provide a higher impedance to the RF and LO modes for impedance matching. The RF and LO bandwidths are limited by the CPW line lengths and impedances, and the electrical length of the top side ground metal running from the diodes to the substrate edge where the top side ground metal connects to the bottom side metal. The electrical distance from diodes to substrate edge is ideally a quarter wave at the center operating frequency. As frequency decreases below this point, approaching DC, the impedance in parallel with the diodes approaches a short circuit. The same is true as frequency increases toward the point of having a half-wave electrical distance that also presents a short circuit across the diodes.

LO to IF isolation is provided by diplex filtering. The LO is high pass filtered to the diodes by the blocking capacitor on the LO CPW line at the gap in the center conductor metal trace. The IF is low pass filtered to the diodes by the alternating high and low impedance microstrip lines between the diodes and the IF port.

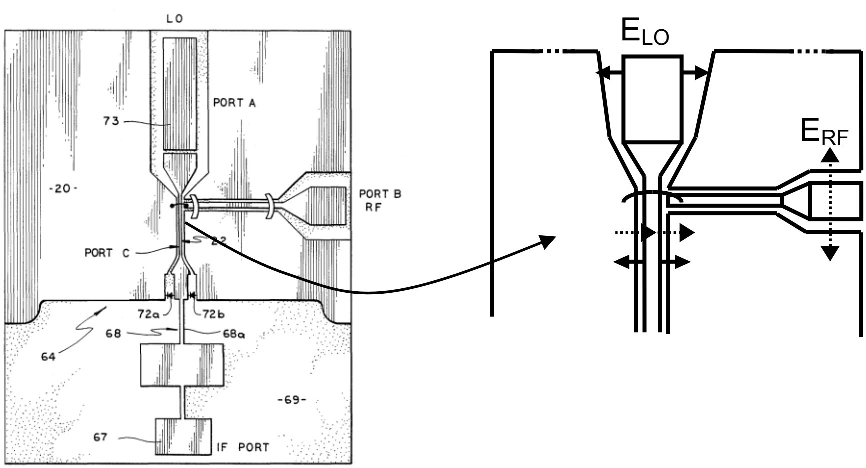



**Figure 6.37** Mixer built using CPW transmission lines. The detail illustrates how the fields from RF and LO are injected into the CPW transmission line.

The low pass filter leads off with a high impedance connection to the diodes to keep the LO from shorting to ground in the low impedance open circuit stubs. Bottom side metallization exists on the lower half of the substrate between the IF port and the diodes, to provide the microstrip ground for the IF low-pass filter. The IF ground return currents run along the top side metal, from the diode connection points to the substrate edges where the topside ground metal connects to the bottom side ground metal. Thus some considerable parasitic phase shift exists in the IF ground return path, which limits the IF bandwidth. The lower end of the RF and LO bandwidths generally coincide with the upper end of the IF bandwidth. The previous mixer circuits of Figures 6.34 and 6.37 were built using ceramic substrate materials. Double sided mixers are also commonly built using "soft" dielectric materials composed of Teflon based materials. The mixer of Figure 6.38, disclosed in 1976, uses soft dielectric, which allows a balanced microstrip line to be cut out and twisted 180 degrees to form the Magic-T hybrid junction [41]. This junction provides the phase relationships to the diodes to achieve singly balanced operation, and obtain isolation between the RF and LO ports. The RF port is at point 14, with the RF signal applied to the two diodes in phase.

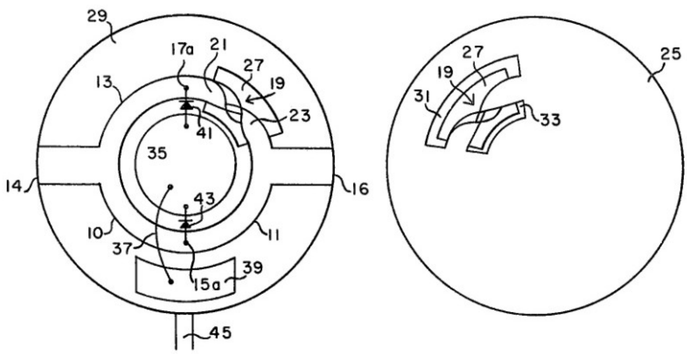



**Figure 6.38** Singly balanced mixer formed by cutting the soft dielectric material and twisting the balanced microstrip line 180° to form the hybrid junction. This is similar to the ratrace with the half wave long line replaced by a 180° phase shifter.

The LO port is at point 16, with the LO signal applied 180 degrees out of phase to the diodes. The RF and LO ports are thus isolated from each other, and the correct phase relationships are presented to the diodes. IF currents are summed at metalized pad 35 that provides capacitive low impedance ground returns for the RF and LO signals, but does not short out the low frequency IF. Bondwire 37 functions as an inductor that forms a low pass filter with capacitance from metalized pads 35 and 39. Operation over 26 to 40 GHz with VSWR less than 2:1 is claimed. The mixer of Figure 6.38 is very similar to the rat race mixer by providing the same RF and LO phase relationships to the diodes; however, it does so without using quarter wave lengths and thus is more broad band than the rat race hybrid.

Figure 6.39 shows another variation of the rat race hybrid that has improved bandwidth, but by routing the IF on the opposite side of the circuit board from the RF and LO, thus obviating the IF crossover that limits the IF upper frequency [42]. The left of Figure 6.39 shows the IF circuitry on substrate 31, and the right side shows the RF/LO circuitry on substrate 21 seen from the back side. The mixer is constructed by flipping substrate 21 such that the letters D and G, respectively, align with each other. A metal ground plane is placed between substrates 21 and 31. The circle AM is three half-wavelengths long, so that the LO signals are 180 out of phase at the diodes and the RF signals are in phase to the diodes. The diodes are placed in holes at 22 and 23 so they connect, respectively, to the IF circuitry at points 32 and 33. Microstrip lines T22/T23 are 44 ohms, and T12/T13 are 72 ohms; both provide impedance matching between the diodes and the IF port. The DC ground return for currents caused by rectification of the LO is provided by the inductive microstrip line of length L, which ends in the ground via 35. The frequency of operation is 9.3 to 9.9 GHz, with an IF of 1860 MHz. The LO ground return takes place by open stubs filters P2, P3. They also give a low impedance to RF signals if IF frequency is sufficiently low.

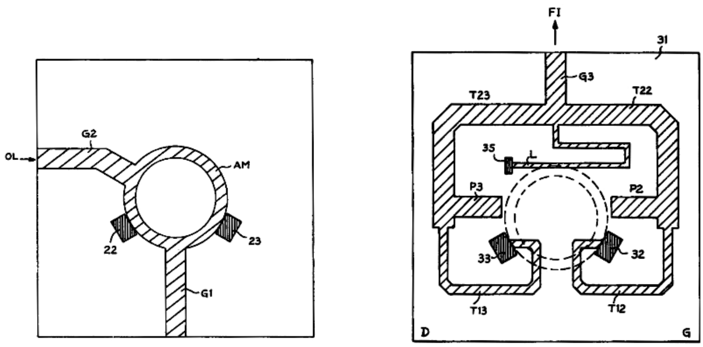



**Figure 6.39** Ratrace mixer with IF (left) on opposite side of board from RF and LO (right).

A major challenge with singly balanced mixers is obtaining the LO ground return without shorting out the RF signal. The ground return is readily obtained in doubly balanced mixers by the second pair of diodes, which unfortunately are not available in singly balanced mixers.

The mixer of Figure 6.40, disclosed in 1985, solves this problem of shorting the RF, and also reduces the physical size, by capacitively coupling the negative RF line to ground [43]. The LO and IF signals use a grounded CPW line whose center conductor ties to the joinning point of the two diodes. Thus the diodes are in parallel to the LO and IF. The opposite ends of the diodes are connected to the ground planes adjacent to the GCPW center conductor, which provide the LO and IF ground returns. The LO and IF are isolated from each other by a diplex filter comprising a high pass series capacitor for the LO, and a series inductor and shunt capacitor lowpass filter for the IF.

The RF is connected across the diodes in series in such a way that the RF is not shorted out by the LO/IF ground lines. This is accomplished by connecting the positive RF balanced microstrip line to a diode at one GCPW ground plane, using an inductive wire. The negative RF balanced microstrip line on the bottom side of the substrate connects to the opposite diode lead on the top side of the substrate, at the other GCPW ground plane, through a via and then using an inductive wire 29. The negative RF balanced microstrip on the bottom side of the substrate is capacitively coupled to the ground plane with a gap 34 cut in the metal. The capacitive loading is claimed to reduce the size of the RF balun, as well as keep it from shorting across the ground plane. The diodes are in series to the RF, and they provide the RF current return. It appears that the RF and LO frequencies may not overlap each other.

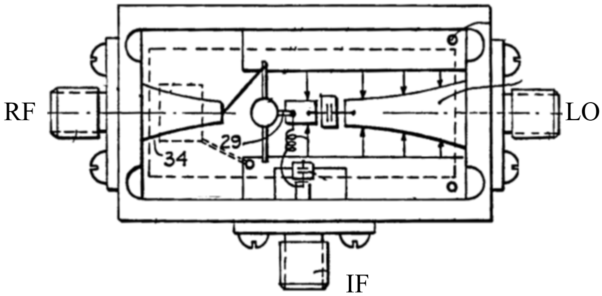



**Figure 6.40** Singly balanced mixer using RF balun with capacitively coupled ground return.

### 6.2.10 MMIC Singly Balanced Diode Mixers

An early MMIC mixer, shown in Figure 6.41, was disclosed in 1968 that uses a quadrature coupler to deliver 9.5 GHz RF and 8.5 GHz LO signals to the mixer diodes for use in a radar [44]. A silicon substrate was employed, with diffused PN junctions. An off chip IF transformer was used to collect the differential 500 MHz IF currents from the two surface oriented diodes at 22 and 23. The mixer is claimed to have 5 dB conversion loss.

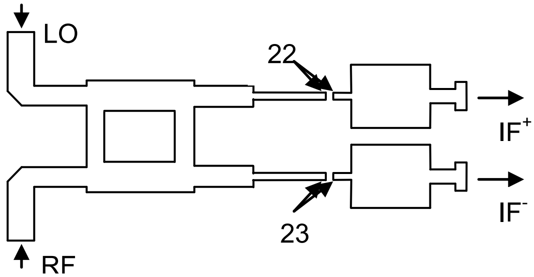



**Figure 6.41** Singly balanced quadrature MMIC mixer.

Another MMIC circuit, operating at 94 GHz and measuring 0.035 by 0.080 inches is shown in Figure 6.42 [45]. It was disclosed in 1970 and includes an integrated 31.3 GHz Gunn oscillator, diode frequency tripler, and mixer integrated onto a 0.004 inch thick GaAs substrate. The Gunn diode oscillator is located at the left, with the diode consisting of area 26, which is a heavily doped n-type region. A varactor diode frequency tripler is located in the middle, which has conversion loss as low as 4 dB.

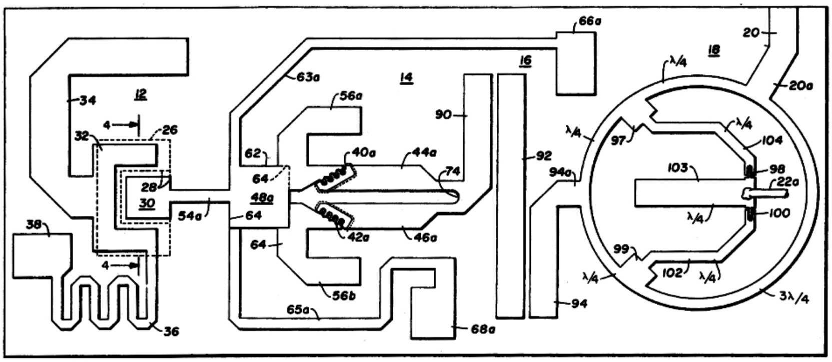
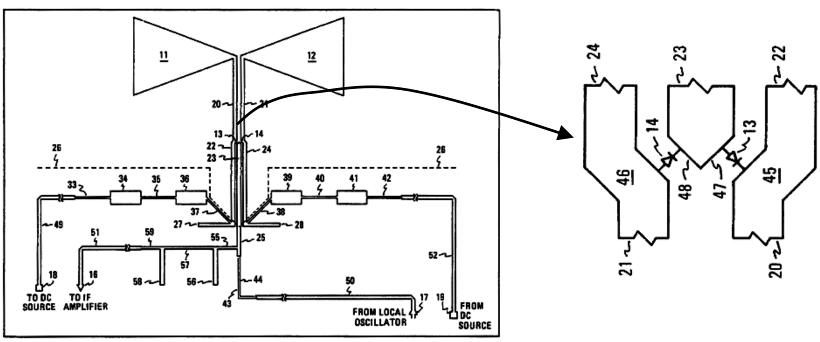



**Figure 6.42** Integrated Gunn oscillator, frequency tripler, and mixer.

The rat race mixer is at the right side, with the LO injected on the left side of the ring at point 94a. The RF signal port connects to microstrip line 20, and the 100 MHz IF is extracted through the bondwire 22a. The GaAs mixer diodes are at 100 and 98. The microstrip line 103 forms the RF and LO ground returns, as well as part of the lowpass filter for the IF.

The MMIC mixer of Figure 6.43, disclosed in 1985, takes advantage of the two balanced coplanar strips 20 and 21, feeding a bow-tie antenna, to form a hybrid junction by connecting the two strips to the two outer lines of a three-strip coplanar transmission line 22, 23, and 24 [46]. The three coplanar strips run parallel to each other, with the center strip connected to the diode common connection point. Thus the two diodes 13 and 14 are connected in series across the two balanced strips, and in parallel with the three-strip coplanar line that runs to the LO and IF ports. The electrical length of the coplanar strips between the hybrid junction and the bow-tie antenna is a quarter wave at the signal frequency. The dotted line 26 indicates the demarcation between where back side ground metallization is applied below the line, and not applied above it. The circuit was designed for operation at 30 GHz with an IF of 2 GHz.

Beginning at the diodes, the three-strip coplanar line runs a half wave length to a point where DC is injected onto the two ground strips, and the center strip continues on to the IF and LO. Positive DC is injected on one ground strip, and negative on the other, according to the diode polarity to improve diodes sensitivity. DC is injected onto both ground strips via low pass filters composed of alternating 90 ohm and 20 ohm microstrip lines, which reject the RF and LO by 25 dB. Open circuit, 30 ohm, microstrip shunt stubs that are a quarter wave long at the LO frequency are also connected to the points where DC is injected to the ground strips, to provide the RF and LO ground returns. The IF and LO are diplex filtered from the center strip to provide LO to IF isolation. The IF low pass filter comprises 90 ohm series microstrip lines with 30 ohm shunt stubs. The IF low pass filter rejects RF and LO by 30 dB, and allows the IF to go to DC for doppler radar and phase detector applications. The LO high pass filter consists of two coupled microstrip lines that form a capacitor to block the low frequency IF. The RF and LO are isolated through mode orthogonality at the hybrid junction. Broadside coupled hair pin filter sections can be added to the back side of the balanced coplanar strips, between the diodes and the bow-tie antenna, to short across the lines at the sum and image frequencies, in turn to minimize conversion loss and noise figure. A doubly balanced version of the mixer is also disclosed. An equivalent electrical circuit to represent the transmission line functions is in the next figure, facilitating understanding of mixer operation.




(a) Full circuit (b) Diode detail

**Figure 6.43** MMIC mixer connected directly to a bow tie antenna using coplanar strips to build a hybrid junction.

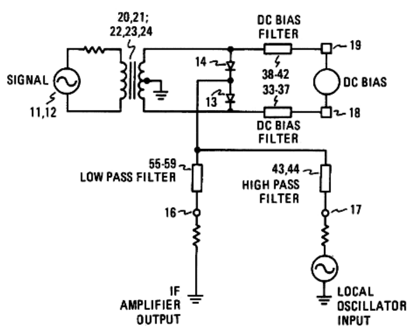



**Figure 6.44** Equivalent circuit for the singly balanced planar mixer MMIC with bow-tie antenna.

## 6.3 DOUBLY BALANCED

The doubly balanced topology continues to be the work horse of diode mixers. Compared with single-ended and singly balanced circuits, the doubly balanced topology offers significantly better IM rejection and port-to-port isolation, with two of the three frequency bands able to overlap without filtering depending on type of mixer. And all this with substantially the same implementation cost in terms of materials, labor, and space as the singly balanced. The doubly balanced mixer originated in the late 1920s, with the first patent granted in 1932 [47]. The goals of this mixer were to obtain isolation and IM rejection using circuit balance with copper oxide diode rectifiers. The mixer comprises two diode pairs, each connected cathode-to-cathode as shown in Figure 6.45. The balanced LO connects to terminals 5,6, with the RF and IF at terminals 7 – 8 and 9 – 10, respectively; however, the RF, LO, and IF can be applied at any of the three specified ports.

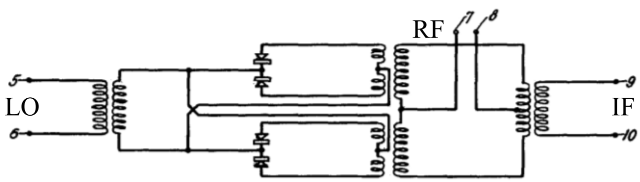



**Figure 6.45** Doubly balanced mixer using two diode pairs, 1932.

This mixer was simplified to the modern doubly balanced circuit comprising one diode ring quad and two center tapped transformers in a patent granted in 1935 [48], shown in Figure 6.46. This circuit dramatically simplified the implementation by cutting the number of transformers in half while providing the same functionality. These early doubly balanced mixers were realized using conventional transformers with magnetically coupled wire coils. Observe that generator 5 can switch the diodes on/off but it is not impressed either at the output load or at the input generator. Let's assume the signal at 11 is the modulating signal. In a given cycle of generator 5 it will switch on diodes 1 – 3 which allows signal 11 to be impressed on load 7. in the next cycle generator 5 will switch on diodes 2 – 4, causing the signal from 11 to appear at load 7 in reversed phase.

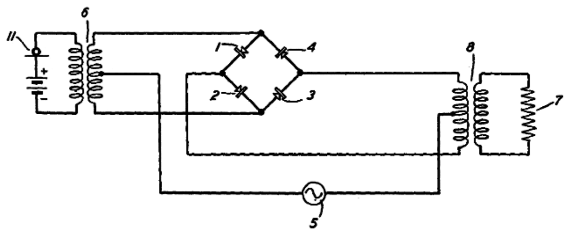



**Figure 6.46** First patented diode ring quad doubly balanced mixer, 1934.

This action represents a modulated carrier wave with carrier suppressed. For proper operation it is desirable the generator or LO signal 5 is of much higher amplitude than the modulating signal 11.

### 6.3.1 Lumped Element

The DB mixer of Figure 6.47 was disclosed in 1979 for use in a television receiver [49]. Such a high volume commercial application requires low cost parts and construction techniques. The goal of this design is to simplify assembly and provide low conversion loss and low IM distortion over a relatively broad bandwidth. This DB mixer design is an attempt to minimize the parasitic reactances normally associated with wire and core baluns. The mixer comprises a metal ground plane over which two dual hole (binocular) ferrite cores are placed. Both cores have a glass packaged axial leaded Schottky barrier diode located inside their two holes. The diode leads are connected either to the ground plane or to a capacitive plate separated from the ground plane by a dielectric layer. Each core hole also has a metal ground return wire passing through it, 130, 136 and 230, 236. Both edges of return wire are shorted to ground carrying a current which is reverse the current through the diode. This is important to neutralize the core magnetization, reducing inductive reactance to these currents, thus increasing bandwidth of operation. The RF is applied in opposing phase to capacitors $C_4$ and $C_6$, which induce currents in the four diodes as they are being switched on and off by the LO. The signal current in diodes 220 and 120, respectively, return to ground through capacitors $C_5$ and $C_3$. The signal currents in diodes 222 and 122, respectively, return to ground directly through ground connections 224 and 124. The IF currents are taken differentially through the windings of the RF transformer 52 that resonate with capacitors $C_4$ and $C_6$.

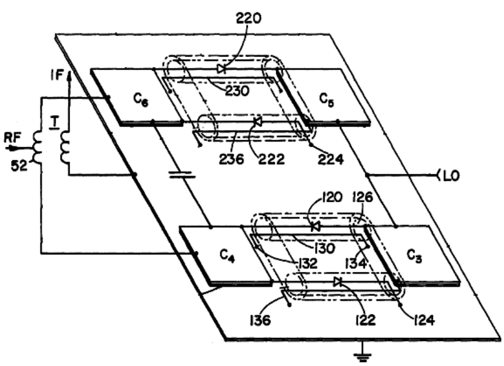



**Figure 6.47** DB mixer with LO balun having no windings on the ferrite cores.

The DB mixer of Figure 6.48 includes an inductor in parallel with the LO balun connections to the diode ring [50]. The inductor is said to resonate out the averaged diode junction capacitance. This nonlinear capacitance contributes to degrade intermodulation and its effect is neutralized by a resonating inductance placed in parallel with the diodes at the LO frequency. The inductance is orthogonal to the RF so it does not degrade conversion performance. Notice the IF is extracted from the center tap of RF transformer.

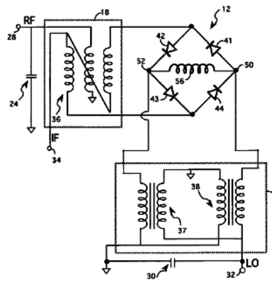



**Figure 6.48** DB mixer using inductor to resonate out parasitic capacitance.

The DB mixer of Figure 6.49 is said to minimize the time for the diodes to switch between conductive and non-conductive states [51]. The goal is improved IM distortion. The reduced time is due to the inductor 14, which acts as a current source. The RF is connected across nodes 1 and 2; the LO is connected across nodes 5 and 6, and the IF output is across nodes 7 and 8.

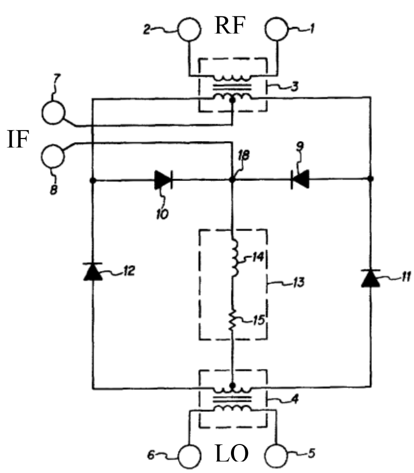



**Figure 6.49** Doubly balanced mixer that minimizes time for diodes to alternate between on and off states.

Each of the four diodes has one of the four possible RF/LO phase combinations, and the IF is connected to pass only mixing products with odd RF and odd LO harmonics; thus, this circuit qualifies as a DB mixer even though it deviates from the normal ring and star configurations. The similarity of inductor usage depicted in Figures 6.48 and 6.49 is striking.

### 6.3.2 Coaxial Transmission Line

The doubly balanced mixer of Figure 6.50 uses the star mixer topology [52] that was introduced in 1937 with vacuum tube technology [53]. The invention uses a Magic-T composed of transmission lines to provide the four diodes with the appropriate relative phase angles for the RF and LO signals. The transmission lines can be realized in any medium, including waveguide, twisted pair wires, coaxial lines, stripline, or microstrip. In contrast to more narrow band mixer circuits, like the ratrace, that depend on line lengths to provide phase cancellation, the wide bandwidth of this design is achieved by arranging the transmission lines so the RF and LO baluns connect to the diodes while presenting to each other broadband high impedance back-shorts.

In the schematic of Figure 6.50 port (1) has connections 1,1' and port (2) has connections 2,2'. This mixer was introduced by Mouw, and a magic-T that is discussed in chapter 4. Port (1) can be the RF input and port (2) the LO input, or vice versa. The IF port is connected to node 39, at the junction of the four resistors R31, R32, R41 and R42 representing the diodes at the LO input drive level. Port (1) has two transmission lines connected in parallel to it, as does port (2). The four transmission lines, 21 – 24, are arranged to isolate port (1), port (2), and node 39 from each other, while delivering the RF and LO energy in the correct relative phase angles to the four diodes. Transmission lines 21 and 22 are connected in parallel to port (1), while transmission lines 23 and 24 are connected in parallel to port (2). The other ends of transmission lines 21 – 24 connect to the diodes and to each other at nodes 3, 3', 4, and 4'. Transmission line 23 connects across resistors R31 and R41, while line 24 connects across R32 and R42.

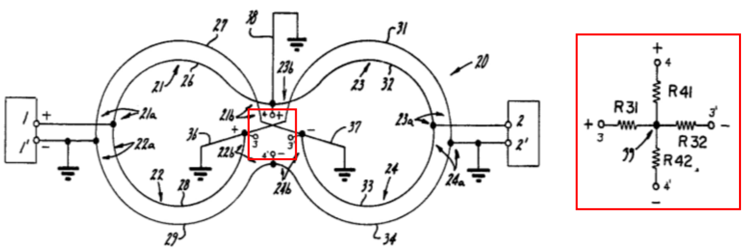



**Figure 6.50** Schematic of a doubly balanced star mixer. The diodes, represented as resistors in the figure on the right, insert into the mixer at 3, 3', 4, and 4'.

The point where R31 and R41 connect together, and the point where R32 and R42 connect together, both present virtual grounds to transmission lines 23 and 24, and so are isolated from port (2). Thus, relative to port (2), these two points can be connected together to form node 39. Similarly, transmission line 21 connects across R41 and R32, while line 22 connects across R31 and R42, with virtual grounds at the two resistor connections relative to port (1). Since port (1) presents equal positive voltages at nodes 3 and 4, and equal negative voltages at nodes 3' and 4', both port (2) and node 39 are isolated from port (1). Similarly, port (1) and node 39 are isolated from port (2). Thus ports (1) and (2) and node 39 are isolated from each other, and connected to the four diodes. This hybrid junction is limited in frequency as follows: At the diode-end of transmission line 23, its two wires connect to the two positive wires 26 and 28 of transmission lines 21 and 22; And at the diode-end of transmission line 24, its two wires connect to the two negative wires 27 and 29 of lines 21 and 22. Thus the positive wires of lines 21 and 22 form their own high impedance transmission line, as do the negative wires of lines 21 and 22, comprising two high impedance lines presenting back-shorts, respectively, to lines 23 and 24. The reverse is also true, as the positive and negative wires of transmission lines 23 and 24 present high impedance back-shorts to transmission lines 21 and 22. As frequency is swept, the impedance of these back-shorts go through peaks and nulls that establish the mixer's bandwidth. With careful physical arrangement of transmission lines and the use of ferrite absorber, many octaves of bandwidth can be achieved for the RF and LO circuits. In contrast, the IF bandwidth is limited by the inductance of its ground return. The IF current must flow through the ground (negative) wires of transmission lines 21 – 24 to the RF and LO connectors, and then back through the mixer housing walls to get to the ground side of the IF connector. Figure 5.50 shows additional IF ground returns 36, 37, 38 that if used must not short out the RF and LO, constraining the IF to be below the RF and LO frequency ranges.

### 6.3.3 Waveguide Mixers

The DB star mixer of Figure 6.51 has wider bandwidth than previous waveguide mixer designs, allowing for an RF to LO frequency ratio greater than 2:1 [54]. This is achieved by locating the diodes immediately adjacent to each other, rather than at multiples of a quarter wavelength as in previous waveguide mixers, which generally were all singly balanced. The LO is presented to the four diodes via the coaxial transmission line on the left of figure. The RF is presented to the diodes via the waveguide section running perpendicular to the page and toward the reader. And the IF is connected to the diodes via the coaxial line low pass filter on the right of figure. Thus the IF and LO ports are isolated from the RF through mode orthogonality, and the LO and IF ports are isolated from each other through a diplex filter.

The diodes are located at the middle of the waveguide's long dimension (horizontal), where the electric field is at its maximum for the $TE_{10}$ mode. The RF port waveguide back-short extends into the page, and terminates in a short circuit. The distance between the diodes and the short circuit is a quarter wave length at the RF frequency, which maximizes the RF voltage across the four diodes. The IF connection to the diodes is via a rectangular metal bar that forms a stripline within the waveguide cavity, then protrudes through the waveguide wall to form the coaxial line low pass filter that the blocks RF and LO leakage out the IF port. Thus the RF signal is impressed across the series-parallel connected four diodes, and isolated from the IF port via mode orthogonality and filtering from the coaxial low pass filter.

The LO port begins as a coaxial line at the left of figure. Two other coaxial lines connect in parallel to the first coaxial line, with the opposite ends of their center conductors connected across the four diodes at the middle of the waveguide. The polarity of the LO is such that LO current alternately passes through diodes 1 – 4 then diodes 3 – 2 as LO polarity changes. The result of the RF and LO connections to the diodes is that each of the four diodes has one of the four possible RF/LO relative phase combinations. The IF is connected to the diodes to reinforce mixing products containing odd RF and odd LO harmonics. The IF ground return is via quarter wave stubs 11, which offer a high impedance to the high frequency RF and LO signals, while offering a low impedance to the lower frequency IF signal.

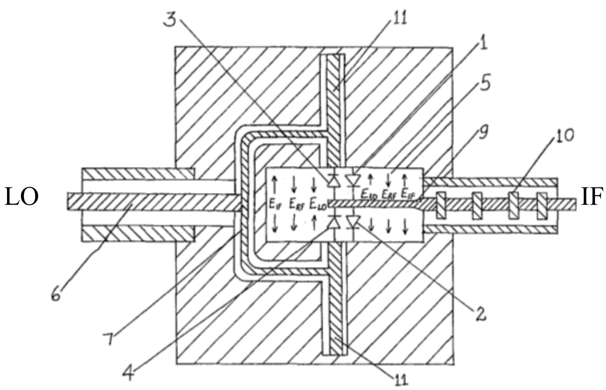



**Figure 6.51** Waveguide doubly balanced diode mixer with coaxial ports for the LO and IF signals. The RF port is the waveguide perpendicular to the paper, and the IF return for diodes 3,4 is by means of shorted quarter wavelength stubs 11.

The doubly balanced mixer of Figure 6.52 uses finline construction, [55]. Typical conversion loss of this mixer is 6 dB at Ka band. The mixer allows for either SB or DB operation depending on how the diodes are connected. The figure depicts the diodes connected to form a DB mixer. The mixer is constructed using waveguide split into two halves (blocks) with the break located where the current flow is at a minimum. The assembly comprises the two metal halves, the PCB dielectric circuit card, the IF SMA connector, and the four beam lead Schottky barrier diodes. Assembly simply involves attaching the beam lead diodes and SMA connector to the suspended substrate card, and screwing the two blocks together. The mixer provides low conversion loss and has superior spurious suppression compared with SB mixers. It also can be biased for low LO power operation.

A waveguide hybrid junction, disclosed in [56], combines energy from the LO in the H-plane waveguide 20 and RF in the E-plane waveguide 28. The waveguide ports are perpendicular to each other, thus they are isolated electrically from each other, and they set up modes inside the assembly that are also orthogonal to each other. Two finline transitions 70 and 72, couple to the LO and RF electric fields, with the LO coupled in equal phase to both finlines, and the RF coupled in opposing phase. The RF and LO electric fields transition down the finline traces, going from wide gap to increasingly narrow gap, where at the most narrow gap, the resulting slotline splits into two series connected slotlines, each of which is terminated by a diode. Thus both finlines are terminated in two series connected diodes that each comprise a diode pair.

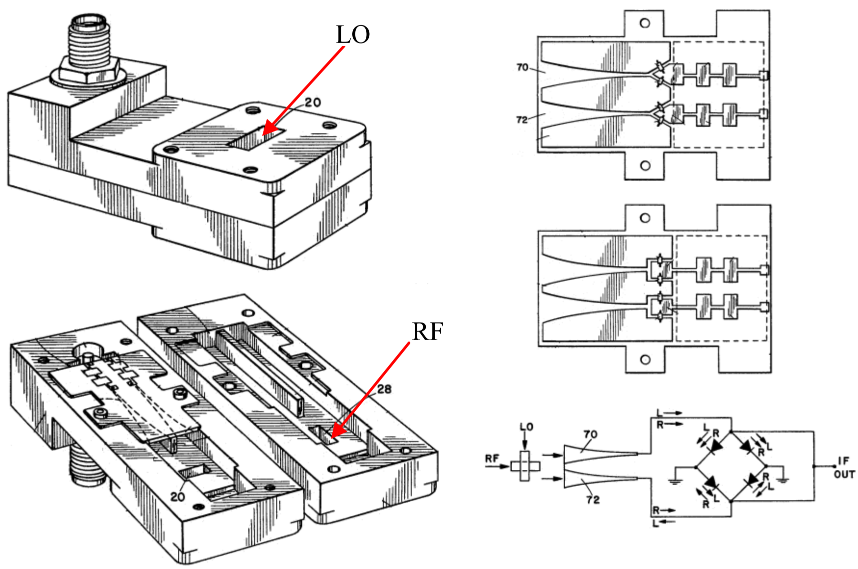



**Figure 6.52** Doubly balanced finline mixer. Mixer construction is shown on the left. The top and middle right show the finline transitioning into slots where the diodes are mounted, with two options for diode connection. Bottom right indicates the electrical equivalent circuit.

The common point of each diode pair connects to a low impedance section at the end of a microstrip low pass filter, which is at a virtual ground relative to the finline. The opposite, high impedance ends of both low pass filters are connected in parallel, with the resulting connection point being the IF port. Thus the IF is isolated from the two finlines, and so also the RF and LO ports, by means of the virtual ground connections and high frequency rejection from the low pass filters.

### 6.3.4 Microstrip

Producing a broadband fully planar (single-sided) microstrip DB mixer remains an important challenge in the field of mixer design. This is certainly important for MMIC realizations, but also for reduced cost thick-film, thin-film, and SMT construction. The current state of the art involves using various combinations of Lange couplers to connect to the nonlinear mixing devices either in quadrature or in 0°/180° relative phase [57]. The current state of the art for MMICs also involves using broadside coupled balanced spiral transformers, which are the smallest realizations.

For many years the ratrace hybrid has been the most popular circuit for planar microstrip balanced mixers. The mixer of Figure 6.53 uses a ratrace hybrid for the LO and IF, and a bifilar wire balun, $T_3$, to realize a DB mixer for a television receiver application, [58]. The DB mixer uses four Schottky diodes arranged in a ring configuration with a paralleled resistor-capacitor in series with each diode for self bias. Rectified LO current circulates around the ring, with equal current in all four diodes, so the voltage developed across the four resistors is also equal.

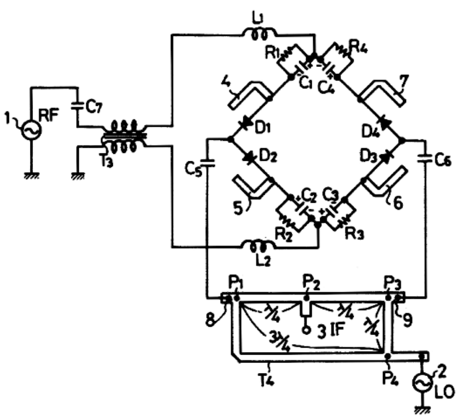



**Figure 6.53** Ratrace balun applied to ring mixer for LO insertion and IF extraction. The RF signal is inserted by a bifilar transformer.

The result is a higher RF input level can be applied before the 1-dB input compression point is reached. The capacitors pass the RF and LO current to minimize resistive losses to them. In addition, four shunt open circuit shunt stubs short the image frequency to minimize conversion loss.

An interesting application shown in Figure 6.54 was disclosed by [59], in which a split ring microstrip resonator is employed. It provides image rejection, and balun function since input is single ended and output is balanced. The ring resonators are tuned with capacitors, which can be implemented as voltage tuning varactor diodes. A semi-lumped resonator can be used to resonate at the required signal frequency, while providing a short circuit at the image or other mixing frequency to optimize performance. High volume commercial mixer applications require minimal cost, small size, and good performance. One such mixer design is depicted in the schematic diagram of Figure 6.55 [60], with the microstip circuitry shown in Figure 6.56. This minimalist doubly balanced mixer topology employs single sided microstrip construction with a monolithic diode cross over quad. No vias or tightly controlled tolerances are required. The mixer is designed for an RF and LO frequency range of 4 to 5 GHz, with IF at 30 MHz. The stated performance is conversion loss of 4.5 to 6.5 dB, with +13 dBm LO drive. The LO-RF and LO-IF isolations are 25 to 50 dB; RF input power at 1 dB compression is +8 dBm; input third-order-intercept-point is +20 dBm; and R-I isolation is 22 dB minimum. The IF band starts at DC and goes up to 500 MHz. Conversion loss and isolation are given in Figure 6.57.

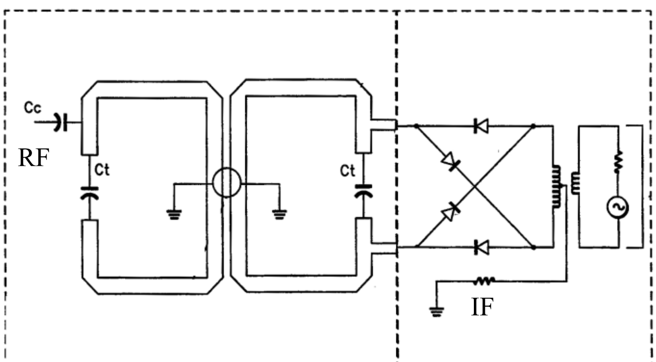



**Figure 6.54** Split ring bandpass filter/balun combination.

The mixer operates as follows. The LO connects to the crossover diode ring quad at points 34 and 35 through microstrip line 28 that is a half wavelength long at the LO center frequency. The 180° phase shift of this line ensures the maximum LO voltage exists across the diodes. A second microstrip line, which is a quarter wavelength long at the LO center frequency, connects between the point 35 and ground. This line is a quarter wave shorted stub that presents a high impedance to the LO signal, and provides the IF ground return. The halfwave line imposes a 180° phase shift between points 34 and 35.

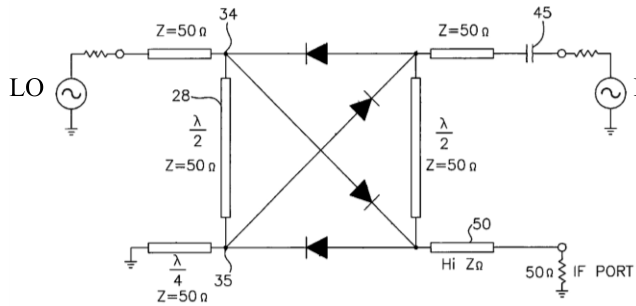



**Figure 6.55** Schematic of a low cost mixer.

The RF balun has the same structure as the LO, with the addition of a simple diplex filter to isolate the RF and IF ports. The high pass portion of the diplexer is a series capacitor 45, and the low pass portion is a high impedance line 50. The nominal value of the series capacitor is 8 pF, which has 7 ohms impedance at 30 MHz, and 1000 ohms at 30 MHz.

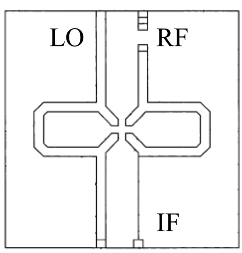
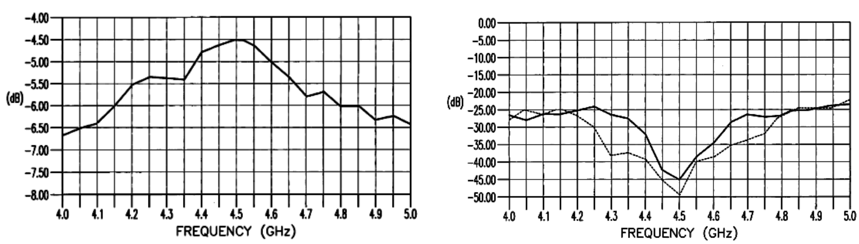



**Figure 6.56** Microstrip implementation of a low cost mixer.




a) Conversion Loss b) Isolation: (——) L - R and (···) L - I

**Figure 6.57** LO is 3.7 – 4.7 GHz @ + 13 dBm; RF is 4 – 5 GHz @ - 10 dBm.

### 6.3.5 Planar Star Mixers

Microstrip mixers using a diode star configuration similar to that of Figure 6.50 have also been realized in microstrip using hybrid and MMIC forms [61]. They require a modification on the conventional Marchand balun for its application, in the manner illustrated in Figure 6.58.

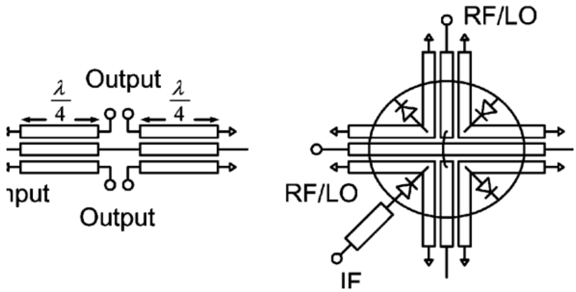
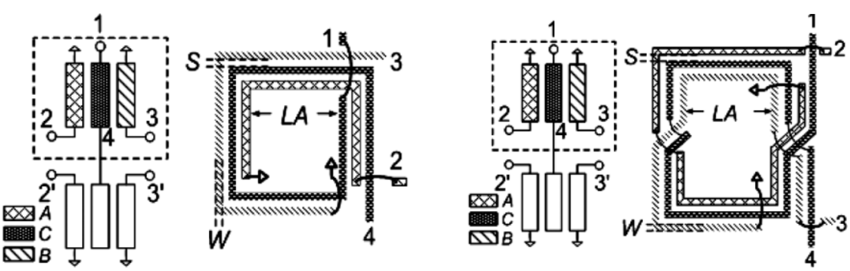



**Figure 6.58** Marchand balun modified for star mixer application. From [61].

These are useful for high microwave frequency and millimeter wave MMIC applications. To obtain reduced size suitable for MMIC circuits, spiral-transformer Marchand baluns have been used in MMIC mixers [62]. The referenced paper refers to these as transformers, but strictly speaking they are Marchand baluns. Two transformer versions were proposed: the first uses a double-spiral transformer on GaAs; the second uses two cascaded trifilar transformers on GaAs. For both versions the even mode impedance is maximized, and the odd mode impedance, $Z_{0d}$, is set approximately equal to (6.8), where $Z_d$ is the diode impedance and $Z_0$ is the source impedance. The layout and schematic of each are in Figure 6.59, detailing the connections and number of layers. The length of the coupled line section is a quarter wave long at the frequency of operation.

$$Z_{0d} = 0.5\sqrt{Z_d Z_0} \tag{6.8}$$




(a) Double spiral (b) Trifilar transformer

**Figure 6.59** Balun transmission line schematic and layout. From [62].

The mixer chips and their performance are shown in Figures 6.60 and 6.61. The double spiral circuit provides the lowest conversion loss, but with less bandwidth due to higher capacitance from the air bridges used to form the transformer. The trifilar baluns are wider in bandwidth and still have good conversion loss within the 26 – 45 GHz frequency band.

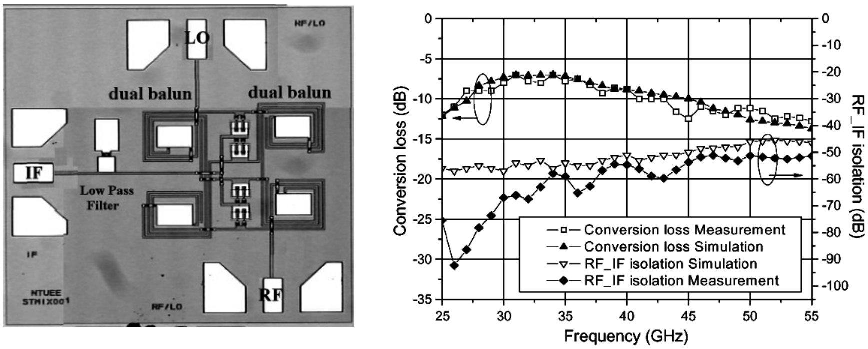



**Figure 6.60** Star mixer and performance using double-spiral baluns. From [62].

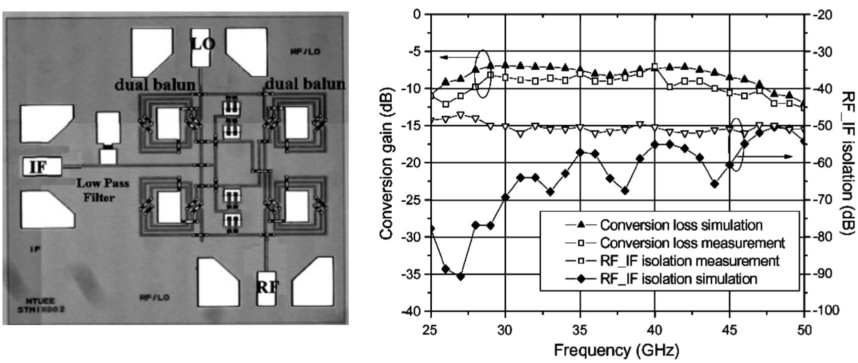



**Figure 6.61** Star mixer and performance using tri-filar transformer. From [62].

### 6.3.6 Balanced Microstrip

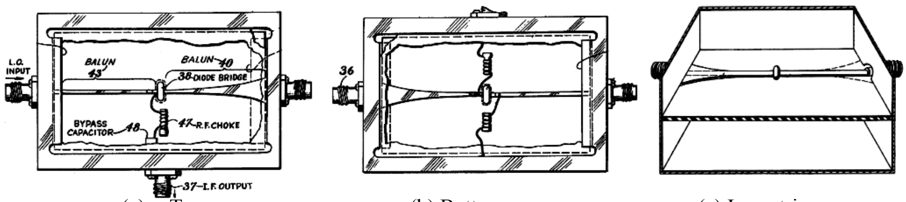
Conventional microstrip is single ended with signal path on the top side and ground on the bottom. Balanced microstrip refers to usage of both top and bottom sides to run signal lines that are 180 degrees apart in phase. Its objectives are to provide a microstrip construction of a microwave mixer having broad frequency coverage with low conversion loss and high isolation.

### 6.3.6.1 Ring Configuration

An important balanced microstrip mixer design was disclosed in February 1970, and is shown in Figure 6.62 [63]. It comprises a single substrate with balanced microstrip lines connected to a Schottky diode ring quad, with the whole assembly being suspended in air inside a metal enclosure. The ring quad comprises four diodes connected anode to cathode. Coaxial connectors are placed on the outside walls of the enclosure for the RF, LO, and IF connections. The RF and LO connections are made through the respective walls to the single-ended microstrip lines. The two microstrip lines begin in a single ended configuration at the housing walls, and transition to balanced microstrip at the diode quad in the middle of the housing cavity.

The width of the top and ground plane sides of both microstrip lines taper equally, forming a balanced microstrip line at the point of connection to the diode ring quad. The taper provides a transition from unbalanced to balanced operation. For an ideal tapered balun the voltage on the top trace will be equal and opposite to the voltage on the bottom trace, and the taper will produce the proper impedance transformation to match the RF and LO signals to the diodes. The equal and opposite voltages divide the power equally across the diodes, and provide the virtual ground needed for isolation to the opposite balanced microstrip line. To analyze and optimize the amplitude and phase balance of the baluns, the top and bottom traces at the balanced end of the tapered line can each be connected to a port using an EM simulator. The grounded ends of the ports connect to a (imaginary) conducting wall placed in the middle of the cavity, perpendicular to the balanced lines and running through the diodes. With the RF input designated as port 1, and the two new ports designated as ports 2 and 3, the quality of amplitude and phase

In the schematic of Figure 6.63 the single ended RF input on the left is transformed up in impedance to a balanced line that connects to two DC blocking capacitors, which in turn connect to the diodes. Two inductors are series connected across the capacitors, with their common point being the IF output port. The impedance of the two blocking capacitors is high enough at the IF frequencies to block the IF current from exiting the RF port, so it exits to the IF port through the two series connected inductors. The capacitor impedance is also low enough at the RF frequencies to pass the RF signal into the diodes. The impedance of the two inductors is low enough to pass the IF current, but high enough to keep the RF signal from shorting out. The LO circuitry is the same, with the exception that the series inductors connected to the LO balanced line provide the IF ground return. The IF port couples inductively to the diodes, which restricts the IF frequency to the MHz range. Also cavity resonances and unwanted propagation modes are suppressed by introduction of microwave absorber material.




(a) Top (b) Bottom (c) Isometric

**Figure 6.62** Balanced microstrip doubly balanced mixer.

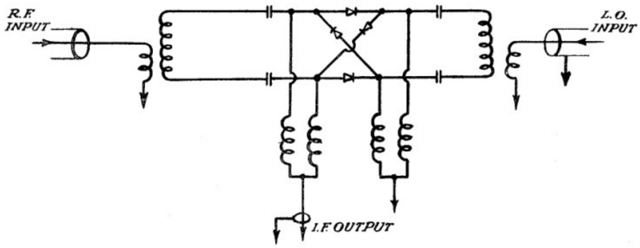



**Figure 6.63** Electrical circuit for the balanced microstrip.

The RF and LO connections in general are interchangeable, but differences in performance can be expected depending on their respective tapers. The balanced microstrip lines are suspended in air, away from the metal covers, to maximize the even mode impedance. The use of soft substrate with low dielectric constant contributes positively since the even and odd mode phase velocities are closer in value, in contrast to high dielectric materials.

### 6.3.6.2 Star Configuration

The star mixer using Mouw's broadband magic-T [52] is discussed in the realization of three mixers [64]. The three use suspended substrate coplanar strips (SSCPS), conductor-backed coplanar strips (CBCPS), and absorber-backed coplanar strips (ABCPS). The circuit configuration is given in Figure 6.64, and the schematic is the same as in Mouw's coaxial line version. The RF and LO lines comprise a magic-T that connect to the four corners of the diode star, and the IF connection is at the center point of the star. It is interesting to note that the grounded ends of the IF CPS are floating, and thus contribute to limiting the IF bandwidth. The planar circuit is used to realize Mouw's hybrid junction using CPW to CPS T-junctions. The even mode impedance is increased by elevating the substrate above ground as with the SSCPS. In contrast, the even mode impedance is dampened by the absorber in the ABCPS case. The odd mode impedance exists across adjacent coupled strips, and the even mode impedance exists between the strips and ground.

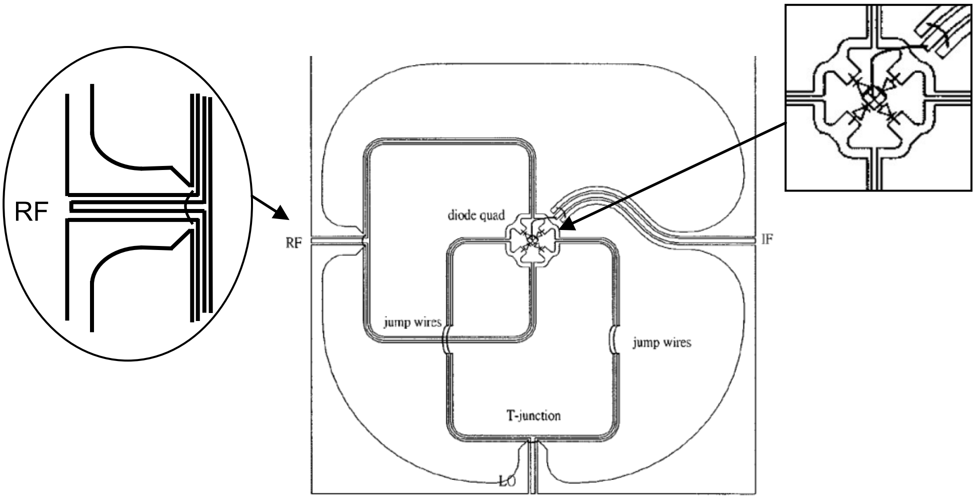



**Figure 6.64** Very wideband star mixer using absorber, with RF launcher and diode connections detailed. From [64].

Using ABCPS with a 0.025 inch thick alumina substrate, measured data for a few mixer designs shows conversion loss is below about 11 dB over about 1 to 34 GHz. It provided the widest bandwidth, with degradation in conversion loss of about 1 dB. The limitation to operational bandwidth in doubly balanced planar mixers is the crossover parasitic inductance and capacitance. Some variations of this circuit were disclosed in earlier years [65, 66], one of which is given in Figure 6.65, utilizing double sided construction with CPW and slotline transmission lines on the substrate top side to 34 GHz. The importance of this mixer is the fact that the RF and LO transmission lines overlap each other on the left side of the diodes to form a magic-T, leaving the IF output free of parasitic crossover lines, allowing for extremely wideband IF.

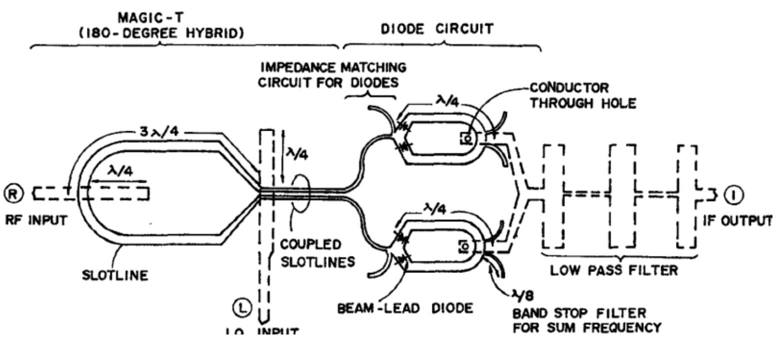



**Figure 6.65** Double-sided DB mixer using CPW and slotline. From [65].

The star nature of the diode connection is better explained in Figure 6.66 by the equivalent electrical circuit. The circuit uses two pairs of diodes, one with common anodes and the other with common cathodes. The ground connection is common for all diodes and IF is extracted the common diode connection point. The LO and RF currents are applied as indicated in the figure.

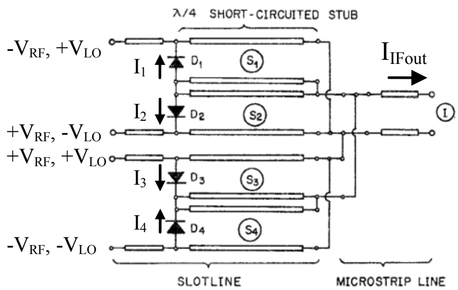



**Figure 6.66** Transmission line equivalent circuit for the mixer in figure 6.65, detailing the star nature of the diode connections. From [65].

The frequency components of currents through diodes $D_1$, $D_2$ are given by (6.9) and (6.10). The diode currents are defined as flowing from anode to cathode. The currents on diodes $D_3$, $D_4$ are similar, and the total IF output current is given by (6.11).

$$I_1 = \sum_n \sum_m i_{nm} e^{jn(\omega_{LO}t+\pi)} e^{jm(\omega_{RF}t)} \tag{6.9}$$

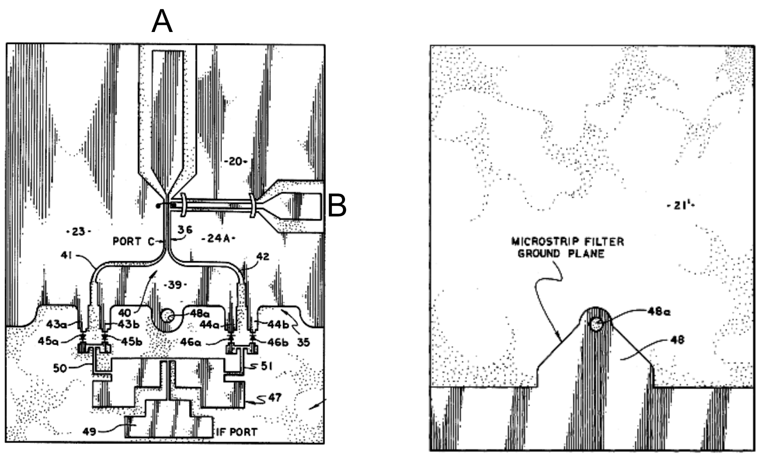
$$I_2 = \sum_n \sum_m i_{nm} e^{jn\omega_{LO}t} e^{jm(\omega_{RF}t+\pi)} \tag{6.10}$$

$$I_{IFout} = -I_1 - I_2 + I_3 + I_4 \tag{6.11}$$

$$I_{IFout} = \sum_n \sum_m i_{nm}\left[1-(-1)^n\right]\left[1-(-1)^m\right]$$

$I_{IFout} \neq 0$ for n = odd and m = odd

Another realization of this approach is in Figure 6.67, [67]. The signal applied at port A propagates on the coplanar line with electric field from center conductor to ground. The coplanar line splits into two slotlines, ideally making the electric fields across the left and right diode pairs be equal in magnitude but 180° opposed in phase. In contrast the signal applied at port B couples into the coplanar line ideally making the electric fields across the left and right diode pairs be equal in both magnitude and phase. The circuit is clearly the paralleling of two singly balanced mixers.




( a) Top view of the mixer. (b) Bottom view showing the ground side

**Figure 6.67** DB mixer using CPW and slotline for the RF/LO hybrid (ports A and B). The IF output circuit comprises a microstrip lowpass filtered power divider to combine outputs from the two diode pairs. The via provides an IF ground return.

## 6.4 TRIPLY BALANCED

Triply balanced mixers remain the most widely used circuit for achieving wide dynamic range with multi-octave frequency coverage and overlapping LO, RF, and IF frequency bands. As explained more fully in Chapter 5, they combine two doubly balanced mixers such that two of the three ports (RF, LO, and IF) are connected together differentially and the third is connected in phase. Thus they are equally referred to as double-doubly balanced mixers. Figure 6.68 shows a triply balanced mixer block diagram. The RF input is at point G, the IF output at point O with both connected differentially to the two doubly balanced mixers. The LO is connected in phase to the mixers [68]. All three ports are interchangeable depending on the respective operating bandwidths. A similar block diagram was granted in 1993 in another patent [69]. Triply balanced mixers have been built using various technologies including: conventional transformers with magnetically coupled wire coils, transmission line transformers using bifilar wire and ferrite cores, printed balanced microstrip lines on double sided substrates and in multilayer LTCC, and using spiral transformers in MMIC form. Numerous triply balanced mixers have been published. A MMIC mixer with RF and LO ports covering 6 to 10 GHz and IF centered at 5 GHz was reported having less than 10 dB conversion loss and better than 25 dB isolation [70]. A triply balanced

MMIC mixer using CPW and slotline was reported with frequency coverage as follows: RF at 6 – 20 GHz, LO at 8 – 18 GHz, and IF at 2 – 7 GHz. Conversion loss ranges from 6.2 to 9.8 dB, with isolation between ports greater than 20 dB [71]. More recently a triply balanced MMIC mixer using lumped element baluns was reported with frequency coverage as follows: RF at 2.0 – 5.5 GHz, LO at 5.0 GHz, and IF at 2.0 GHz [72].

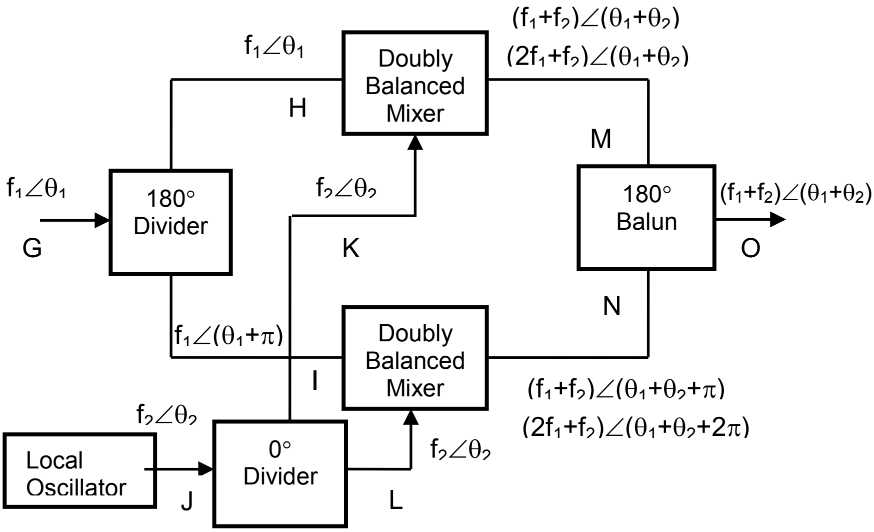



**Figure 6.68** Triply balanced mixer block diagram with input signal at frequency $f_1$ applied to G and LO signal applied at J. At the IF output, 0, the $f_1+f_2$ products add constructively and $2f_1+f_2$ cancels.

### 6.4.1 Triply Balanced Circuits

Numerous triply balanced mixers have been disclosed in the patent literature. The first patented triply balanced mixer uses two diode ring quads connected together with balanced microstrip lines [73]. Figure 6.69 shows the circuit schematic, and Figure 6.70 shows a top view of the implementation. The circuit has two diode ring quads with opposing polarities, connected to each other at opposite corners where the IF balun also connects differentially. The RF and LO ports comprise points 13 and 16 that each connect to two balanced microstrip lines in parallel. The other ends of the balanced microstrip lines connect to each other and to the diodes, forming a magic-T hybrid as discussed in Chapter 3. Point 13 is the sum port, point 16 is the difference port, points 71 and 73 comprise the first diode port, and points 72 and 74 comprise the second diode port. Power incident at ports 13 and 16, respectively, is applied to the two ring quads in phase and 180° out of phase. The top and bottom diode ring quads each comprise a doubly balanced mixer. Thus this circuit is in accordance with the triply balanced block diagram of Figure 6.68.

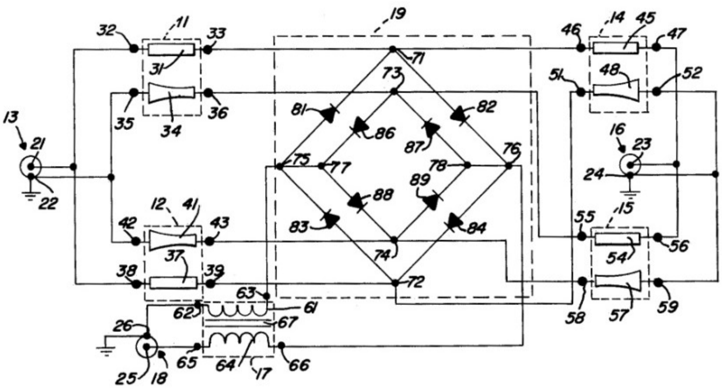


**Figure 6.69** Triply balanced mixer schematic. The RF and LO ports, respectively, are at 13 and 16, or vice versa. The IF port is at 18.

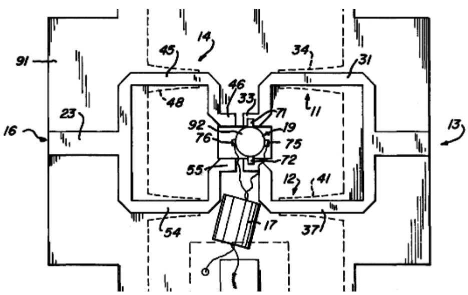
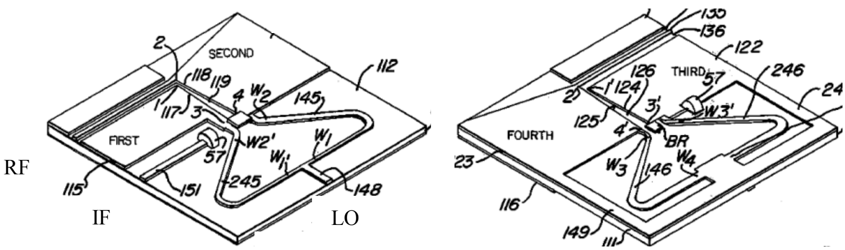


**Figure 6.70** Triply balanced mixer implementation showing top view. Bottom view is similar.

The mixer is said to have an RF and LO frequency range of 0.5 to 12 GHz, and IF range covering DC to 0.5 GHz. While the IF does have a DC connection to the diodes, the lowest usable IF frequency range does not extend to DC. Given the limited inductance offered by the ferrite core to the grounded wire of the two-wire IF balun, points 75 and 77 short to ground at very low frequencies.

Subsequent improvements were made resulting in the triply balanced mixer whose implementation is shown in Figure 6.71 [74]. The circuit is the same except the difference portion of the magic-T is implemented using CPW and slotline, and the in-phase portion is implemented using constant impedance unbalanced-to-balanced microstrip transitions with cosine curve taper. The construction is two sided, with one diode quad attached to the top and the other to the bottom of the substrate. The LO input is at port 148 that connects to a broadband coupled four port balun, built with balanced microstrip lines 245, 145 on top and lines 246, 146 on the bottom. The RF access is by means of a coplanar line which transforms into a double slot balun comprising slot 119 at the top and 126 at the bottom. The IF is extracted using a bifilar transmission line wounded on a ferrite core, 57.

The mixer is said to have RF and LO frequency range of 2 to 20 GHz, and IF range covering 0.05 to 6 GHz. The preferred length of the constant impedance lines is a quarter wavelength at the lowest operating frequency, but said to be usable to 1/40th of a wavelength. Isolation between ports is typically 20 to 30 dB. The constant impedance and CPW-slotline baluns, respectively, are said to have usable bandwidths of about 0.1 to 24 GHz and 1 to 20 GHz.




(a) Top view of triply balanced mixer (b) Bottom view of triply balanced mixer

**Figure 6.71** Triply balanced mixer using a double slot balun and coplanar and balanced microstrip.

The triply balanced mixers discussed so far have at least one interconnection involving a microstrip line crossing over another or an ungrounded (inductive) line. These introduce parasitic capacitance or inductance that limits performance. A triply balanced mixer was disclosed that is without such interconnects [75]. Figure 6.72 shows the schematic and Figure 6.73 shows the top side of the planar mixer.

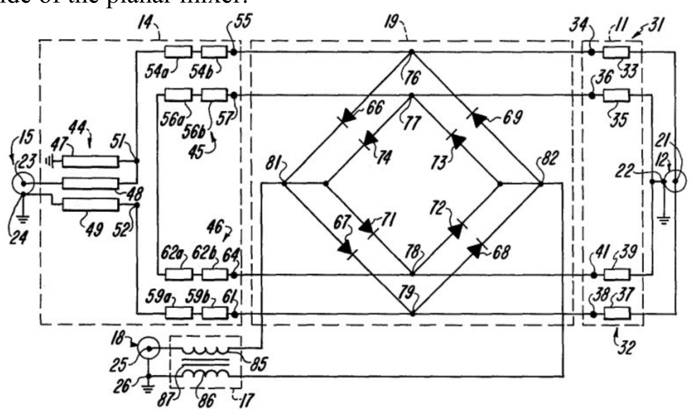


**Figure 6.72** Schematic of an alternative triply balanced mixer.

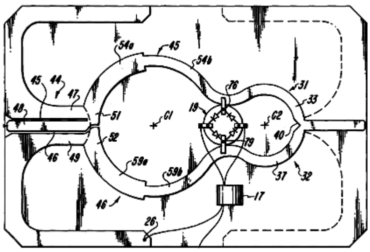



**Figure 6.73** Top side of triply balanced mixer substrate, illustrating the edge coupled balun, 44, followed by balanced lines 45, 46. The balun on the right is simply composed of balanced lines 31, 37.

The bottom side is shown in Figure 6.74, further illustrating the balun details. The mixer uses the two diode ring quads having the same polarity, and the difference portion of the hybrid magic-T comprises three edge coupled microstrip lines.

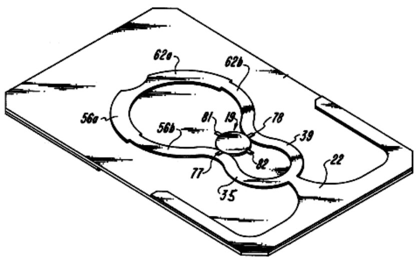



**Figure 6.74** Bottom side of triply balanced mixer substrate.

Another method to avoid crossover connections is to use orthogonal balanced microstrip construction, having RF and LO baluns that are physically perpendicular to each other. Although challenging in terms of assembly, this construction approach is very common and utilizes the leads from the two diode ring quad packages to make connections to the two boards. The leads exit the diode package and first solder to the four microstrip traces at the end of the horizontal balun, then two leads bend upwards and two leads bend downwards and solder to the four microstrip traces of the end of the vertical balun.

A hybrid alternative was invented by Marki that combines the planar substrate with the orthogonal interconnections. A single planar PTFE (teflon) substrate has both RF and LO baluns printed onto it, such that one can be cut out and bent ninety degrees to form the orthogonal balun. Removing the crossing and ungrounded interconnections in the RF and LO circuits dramatically improves their performance. However, in the triply balanced mixers discussed so far the IF balun still connects to the diodes using bifilar wires having significant inductance that limits the upper end of the IF bandwidth to about 8 GHz. An improvement was made in 1982 to replace the wire and core IF balun with balanced microstrip lines that allows for IF coverage up to 15 GHz and higher, [76]. Triply balanced mixers can also be realized using two star diode quads as discussed in Chapter 5, with an example design shown in Figure 6.75 [77]. The star quad has four diodes all connected at a common point, with the other end of each diode comprising four more connection points for a total of five, in contrast to the diode ring quad that has only four connection points. The IF is connected differentially across the common points of the two star quads, and the RF and LO signals couple into the other diode connection points through the magic-T circuitry comprising the U-shaped balanced microstrip lines. The mixer shown in Figure 6.75 has one star quad on top-side, and the other one (not shown) on bottom-side of the double sided alumina substrate.

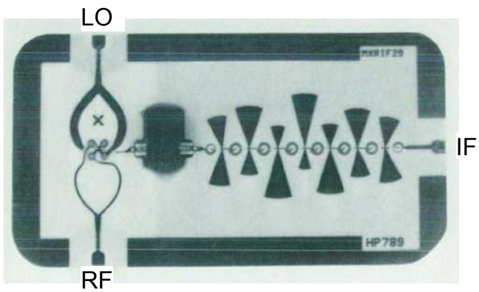



**Figure 6.75** Triply balanced mixer using single substrate and two star diode quads. From [77].

The triply balanced mixers discussed so far all utilize double sided balanced microstrip. In contrast, a single sided microstrip design is described that uses two diode cross over ring quads, and five baluns that each comprise a multi-section Wilkinson power divider and two open circuited couplers [78].

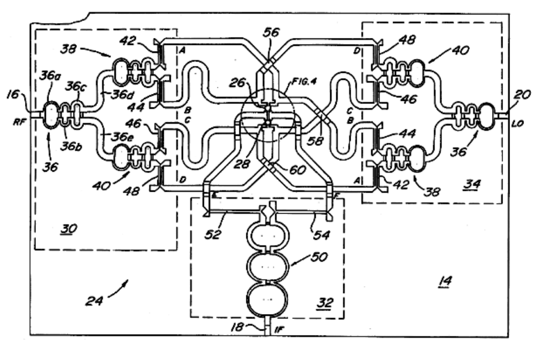



**Figure 6.76** Single-sided triply balanced mixer.

Figure 6.76 shows the cost for avoiding double sided construction is the necessity for numerous crossing microstrip lines that can limit isolation and bandwidth. Much work has been done to implement triply balanced mixers in multilayer low temperature co-fired ceramic (LTCC) to reduce size and cost. One such mixer has LO frequency coverage over 0.5 to 3.5 GHz with conversion loss in the range of about 5.7 to 7.5 dB, and isolation typically 25 to 30 dB [79]. It employs transformers with 1:4 impedance ratio to connect RF and LO, and 1:1 impedance ratio for the IF connection.

Another LTCC triply balanced mixer achieves high LO isolation by swapping the LO and IF ports, and extracting the IF at the center tap of the RF balun [80]. Figure 6.77 shows the schematic for the mixer. In the conventional triply balanced circuit, the LO is isolated from the RF port by means of the hybrid magic-T alone. In this mixer, it is claimed that "double isolation" is achieved by adding to this the isolation afforded by the diode ring quads, since the LO and RF have orthogonal diode connections Some interesting circuit variations of the triply balanced mixer use diode bridge quads in place of diode ring quads.

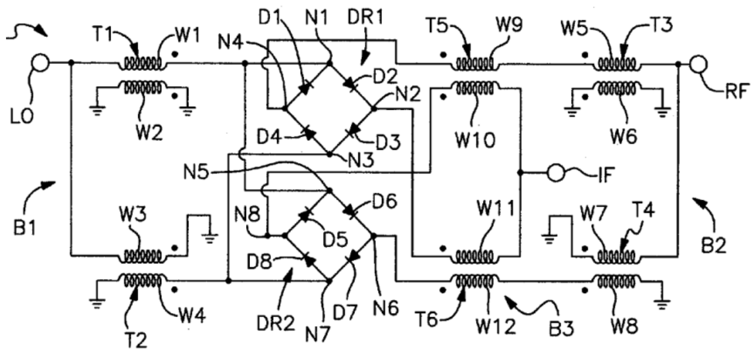



**Figure 6.77** LTCC triply balanced mixer with high isolation.

### 6.4.2 Bridge Quad Mixers

An ancestor to the modern triply balanced topology was patented in 1940 [81]. Figure 6.78 shows it comprises four diode bridge quads, and has balanced connections to the RF, LO, and IF ports. While it is not obvious at first glance, the circuit conforms to the triply balanced block diagram of Figure 6.68. And it achieves triply balanced performance in terms of overlapping RF, LO, and IF frequency coverage, and improved IM suppression. Two of the four diode bridge quads comprise one doubly balanced mixer, and the remaining two bridge quads comprise the other. For purposes of analysis, the first doubly balanced mixer, DBM1, shall comprise bridge quads 1 and 4, and second doubly balanced mixer, DBM2, shall comprise bridge quads 2 and 3. The signal source balun 5 connects to DBM1 and DBM2 in phase at point B and again at point D, while the IF output load circuit connects to them differentially across points A and C. The LO 6 connects differentially to DBM1 and DBM2. Using the method in Chapter 5, with n and m, respectively, referring to LO and RF harmonics, it can be shown that the IF diode currents of DBM1 combining at point A comprise odd RF and odd LO harmonics as shown below. This shows the two bridge quads 1 and 4 together comprise a doubly balanced mixer. The IF diode currents of DBM2, which combine at point C, give the same result.

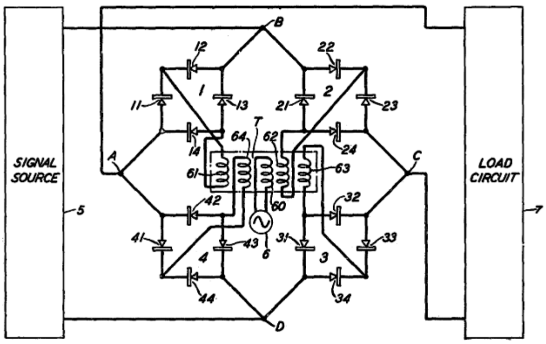



**Figure 6.78** Predecessor to the modern triply balanced mixer, 1940.

The frequency components of the current emerging from node A is given by (6.12). Diode currents are assumed to flow from anode to cathode. The RF signal voltage at B is positive and at D it is negative. Notice that $i_A \neq 0$ when n and m are both odd.

$$i_A = -i_{11} + i_{14} - i_{41} + i_{42} \tag{6.12}$$

$$i_A = \sum_n \sum_m i_{nm}\left[-(-1)^{n+m} + (-1)^n - 1 + (-1)^m\right] e^{jn(\omega_{LO}t)} e^{jm(\omega_{RF}t)}$$

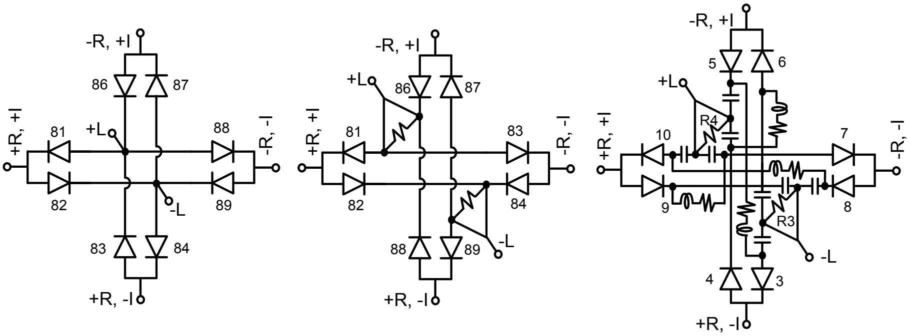
$$i_A = \sum_n \sum_m i_{nm}\left[(-1)^n - 1\right]\left[1 - (-1)^m\right] e^{jn(\omega_{LO}t)} e^{jm(\omega_{RF}t)}$$

Two important mixer circuits are known as the "class IV" or the "Termination Insensitive Mixer" (TIM), and the Paramixer, both of which are arguably variations of the triply balanced mixer. The comparison between these mixer types is depicted in Figure 6.79. The triply balanced mixer has two diode ring quads comprising diodes 81, 82, 83, 84, and diodes 86, 87, 88, 89, with numbering corresponding to the triply balanced mixer in Figure 6.69.

The TIM was patented in 1980, [82]. The two ring quads of the triply balanced mixer are replaced by two bridge quads that are connected by 100 ohm resistors at the LO connection points. The TIM diode bridge quads comprise diodes 81, 82, 83, 84, and diodes 86, 87, 88, 89, with numbering corresponding to the TIM in Figure 6.80. In this circuit configuration, the IM products comprising even RF and even LO harmonics are forced to pass through the two 100 ohm resistors minimizing overall distortion. In contrast, in the triply balanced mixer these currents circulate around the two diode ring quads and cannot be resistively terminated. Attenuating these products is desirable because they include the second harmonic of the RF signal that is a constituent part of the third-order single-tone and two-tone IM products. In comparison to the triply balanced mixer, the TIM has reduced variation in IM suppression versus frequency when reactive loads such as filters are placed adjacent to the RF and IF ports. One viewpoint is that energy from mixing products other than the desired IF reflect off the reactive load and re-mix to add constructively or destructively causing variation in the IM product. The internal termination resistors dampen some of these IM products, without attenuating the desired IF. The cost is increased circuit complexity and reduced bandwidth due to parasitic inductance from longer interconnects.

Another reason that mixers built using bridge quads instead of ring quads may have superior IM suppression is that the peak LO voltage may not be clamped by the diodes as much. This is important because IM suppression improves as the ratio of RF to LO voltage decreases. In a modern doubly balanced mixer schematic such as the one on Figure 5.41(a), the LO forward biases diodes $D_1$ and $D_2$ on, and reverse biases diodes $D_3$ and $D_4$ off, with the reverse bias limited by the turn on voltage across $D_1$ and $D_2$. The RF current is applied differentially across the diode ring quad. As the RF voltage rises, given the polarity shown, diodes $D_2$ and $D_3$ are slightly forward biased, and diodes $D_1$ and $D_4$ are slightly reverse biased, which perturbs circuit balance at the RF frequency and generates IM products containing RF harmonics. In contrast, if the polarity of diodes $D_3$ and $D_4$ were reversed, changing the ring to a bridge quad, the LO voltage is no longer limited by the diodes. However, a second bridge quad must be introduced to utilize the polarities of both LO half cycles, and the circuitry introduced to separate the two bridge quads can help reduce the clamping effect between them.

The Paramixer [83, 84] also uses diode bridge quads in place of ring quads to further its goal of maximizing dynamic range. It was patented in 1978, two years prior to the TIM patent. The TIM and Paramixer have identical RF circuits, but the Paramixer includes DC bias decoupling to ensure DC current is equal throughout each bridge quad. DC current is routed through the diodes of each bridge quad as if it were a ring quad. The Paramixer also includes filters in the LO and IF circuits that terminate out-of-band signals, and an impedance transformer to optimize the RF port impedance match. The goals are improved termination insensitivity to IM products, and multioctave operation. The schematic of a termination insensitive mixer is shown in Figure 6.80, [82], illustrating the two ring mixers and the 100 ohm bridge resistors. The RF, LO, and IF baluns comprise balanced transmission lines that can be realized using balanced microstrip or bifilar wire wrapped around a ferrite core.




Triply Balanced Class IV TIM Paramixer

**Figure 6.79** Comparison between triply balanced, TIM, and Paramixer.

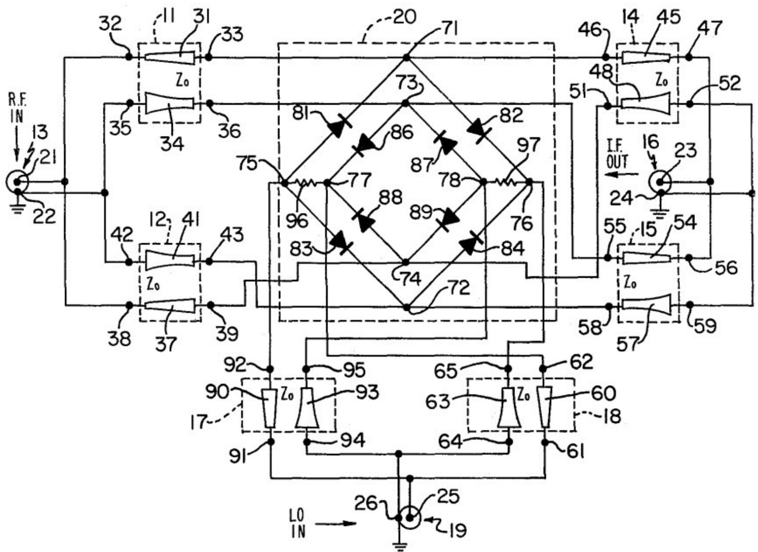



**Figure 6.80** Class IV termination insensitive mixer (TIM).

The Paramixer circuit is depicted in Figure 6.81. It comprises the same magic-T hybrid as already discussed for the TIM circuit. It should be noted that diode bridge quads 56 and 58 alternately switch on and off by the LO, causing a short/open circuit across points a and c, and open/short across points b and d. It can be seen that these conditions for points a, b, c, d cause the IF output to equal the RF signal envelope with its phase reversed at the LO frequency. The resistors $R_3$ and $R_4$ are shown to provide a constant impedance to the LO currents.

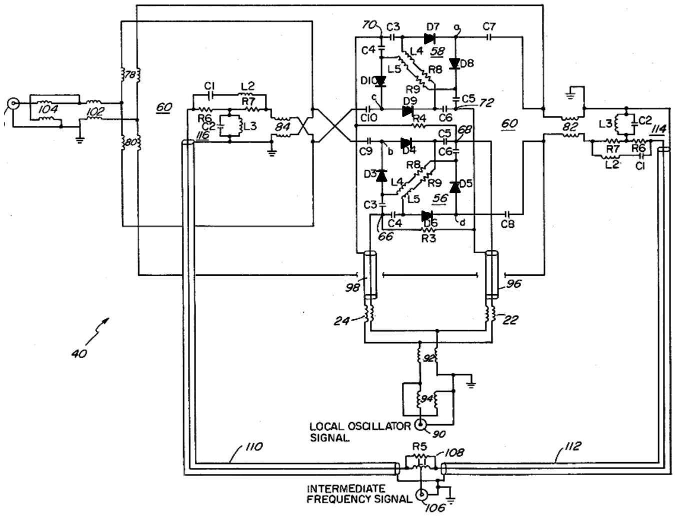



**Figure 6.81** Paramixer circuit schematic.

Details of the RF and IF circuits are given in Figure 6.82. The transmission line transformers 24 and 22 enforce odd mode currents, while suppressing even mode ones. The current, $I_S$, from the "first RF source" splits into two halves that each equal $I_S/2$, and pass through transformers 22 and 24. The positive voltage from transformer 22 turns off diodes $D_4$, $D_5$, the current is then routed through resistor $R_4$, flowing to node 70. At the same time the positive voltage from transformer 24 turns on diodes $D_7$, $D_9$. The negative voltage from transformer 24 turns off $D_3$, $D_6$ forcing current to pass through resistor R3 and then to node 72 that turns on $D_8$, $D_{10}$.

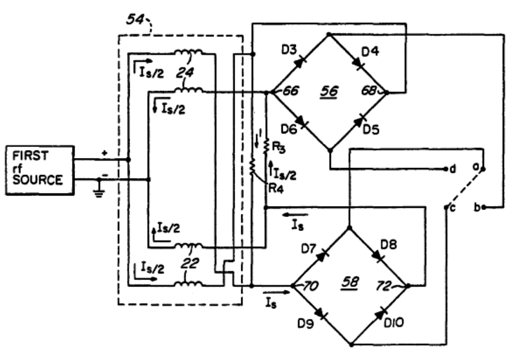



**Figure 6.82** Paramixer circuit: schematic for the LO section of the circuit.

The overall effect is that during the positive cycle the top quad is off and bottom quad is on, and vice versa. Constant loads to the RF and LO sources are maintained as the diode bridge quads alternately turn on and off.

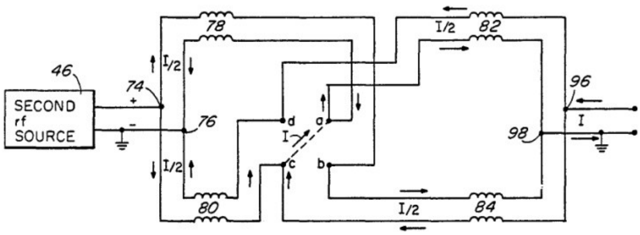



**Figure 6.83** Paramixer circuit: schematic for the RF and IF sections of the circuit.

Resistors $R_3$ and $R_4$ have values equal to the impedance of the transmission line transformers 22 and 24, and should equal the impedance offered by a single diode to the LO power incident at it. This applies equally to the TIM. It is further pointed out in the patent out that to obtain the maximum dynamic range, the operating point for the diode forward current, in quadrant 1 (upper right) of the diode *I, V* curve, should be at half the peak RF signal current. And the operating point for the diode reverse voltage in quadrant 3 (lower left) should be at half the peak RF signal voltage. These two conditions are satisfied simultaneously if the characteristic impedance of the RF signal generator and resistors $R_3$ and $R_4$ equal the maximum reverse voltage of the RF signal divided by the maximum forward current of the RF signal. Furthermore, the bias decoupling resistors $R_8$ and $R_9$ should equal half the value of $R_3$ and $R_4$ for optimum impedance match in Figure 6.81.

### 6.4.3 Multiple Mixer Circuits

The approach used by triply balanced mixers to combine two doubly balanced mixers using in-phase and 180° power combiners has been extended to combining four doubly balanced mixers to further increase dynamic range [85]. Figure 6.84 shows the triply balanced mixer, designated by the author as a "second order" cancellation circuit. Figure 6.85 shows two second order circuits combined to form a "third order" cancellation circuit. The resulting performance for both is shown in Table 6.2 for three cases: individual mixers, second order circuit, and third order circuit.

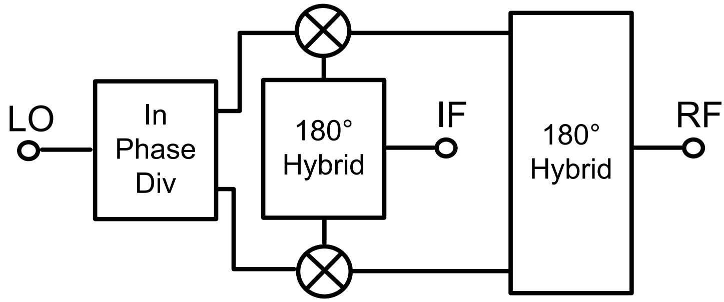



**Figure 6.84** Second order cancellation block diagram. After [85].

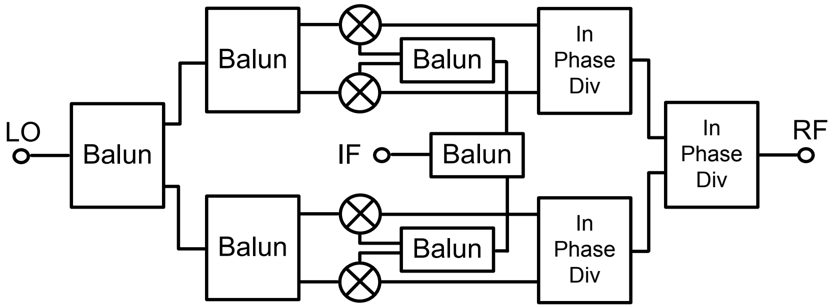



**Figure 6.85** Third order cancellation block diagram. After [85].

LO power increases 3 dB going from individual mixers to second order, and another 3 dB going from second to third order. Conversion loss increases 1 dB going to second order, and another 1 dB going to third order. $IIP_3$ increases 4.2 dB going to second order, and another 4.3 dB going to third order. $P_{1dB}$ increases 4 dB going to second order, and another 4 dB going to third order. L - R isolation increases about 27 dB going to second order, and another 18 dB going to third order. L - I isolation increases about 22 dB going to second order, and another 4 – 9 dB going to third order. Results are also presented for two and four mixers combined using quadrature couplers.

**Table 6.2**
Performance of Second and Third Order Circuits

| | Mixer 1 | Mixer 2 | 2nd Order Cancel dB | Mixer 3 | Mixer 4 | 2nd Order Cancel dB | 3rd Order Cancel dB |
|---|---|---|---|---|---|---|---|
| Conv Loss, dB | 5.7 | 5.7 | 6.7 | 5.7 | 5.7 | 6.8 | 7.7 |
| IIP3, dBm | 9.5 | 9.5 | 13.7 | 9.5 | 9.5 | 13.5 | 18 |
| P1dB , dBm | 1 | 1 | 5 | 1 | 1 | 5 | 9 |
| L - R Isolation, dB | 28 | 29 | 56 | 28 | 29 | 57 | 75 |
| L - I Isolation, dB | 37 | 38 | 59 | 36 | 36 | 64 | 68 |

After [85]

## 6.5 QUADRATURE MIXERS

In down conversion applications rejection of the image frequency can be achieved by filtering or by phase cancellation. This is also true for up conversion applications where the upper or lower sideband is selected and the other is rejected. The phase cancellation approach requires usage of quadrature hybrids that introduces a 90° phase difference between the inputs. An early diode mixer utilizing quadrature hybrids to form a doubly balanced mixer is depicted in Figure 6.86 [86]. It supports the difference frequencies but cancels the sum frequency. Given RF and LO frequencies $f_{RF}$ and $f_{LO}$, the difference frequencies $f_{RF} - f_{LO}$ and $f_{LO} - f_{RF}$ appear at the IF output, but the sum frequency $f_{RF} + f_{LO}$ is suppressed. The four inductors provide the IF ground returns. The circuit of Figure 6.86 can be analyzed using the method presented in Chapter 5 to show which mixing products appear at the IF output.

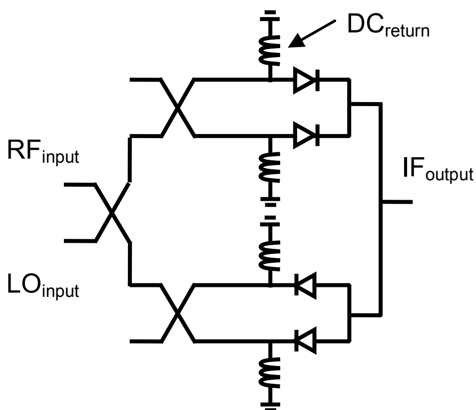



**Figure 6.86** Doubly balanced mixer using quadrature couplers, suppresses the sum frequency.

Assigning integer m to represent RF signal harmonics, and integer n to represent LO harmonics, the IF output current is given by (6.13). The sum frequency is given by m=n=1, while the difference frequencies correspond to m = -1, n=1 and m=1, n= -1. For the sum frequency, (6.12) equals zero.

$$i_{IF} = \sum_{n} \sum_{m} i_{nm}\left(j^{-m}(-1)^n + j^{-n} - j^m(-1)^n - j^n\right)e^{jn(\omega_{LO}t)}e^{jm(\omega_{RF}t)} \quad (6.13)$$

An improved quadrature mixer circuit is shown in Figure 6.87 that has separate quadrature couplers for the RF and LO inputs [87]. The result is higher isolation between RF and LO ports. The top 6.87(a) circuit has IF output $f_0$ with difference $f_S - f_L$ and $f_L - f_S$ signals, but the sum product $f_S + f_L$ cancels.

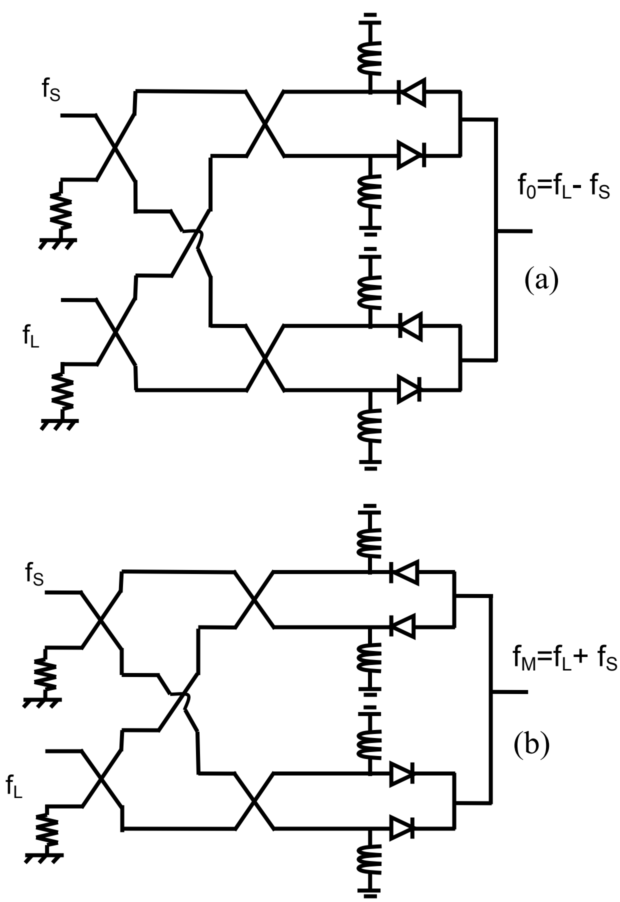



**Figure 6.87** Doubly balanced mixer using quadrature couplers with IF equal to the (a) difference frequency, (b) sum frequency.

The bottom circuit is the opposite of Figure 6.87(b), with IF output $f_M$ only at the sum $f_S$+$f_L$, with signals at the difference frequencies cancelled. The analysis of Chapter 5 is again applied with m and n, respectively, representing RF and LO harmonics. Equations (6.14), and (6.15) give the phase components of the output current for $f_0$ and $f_M$:

$$I_{f0} = \left(-j^n(-1)^m + j^{-m} - j^m + (-1)^m j^{-n}\right) \tag{6.14}$$

$$I_{fM} = \left(-j^n(-1)^m - j^m(-1)^n + j^{-m}(-1)^n + (-1)^m j^{-n}\right) \tag{6.15}$$

A QPSK demodulator was patented [88] where the two input RF signals are phase shifted and then mixed with two singly balanced mixers delivering separate IF outputs as depicted in Figure 6.88. For both IF outputs the sum product is suppressed. The top IF output, $IF_1$, provides the two difference products, $f_{RF1} - f_{RF2}$, and $f_{RF2} - f_{RF1}$ with the same phase angle of 180°. In contrast, the bottom IF output, $IF_2$, provides the $f_{RF1} - f_{RF2}$ with 90° phase shift, and the $f_{RF2} - f_{RF1}$ product with -90° phase shift. Using the same analysis method for the frequency components of current, one obtain (6.16) and (6.17) that give the phase components of the currents for outputs $IF_1$ and $IF_2$. The integers m and n, respectively, refer to the $RF_1$ and $RF_2$ ports.

$$IF_1 = I_{nm}\left\{j^{-m+n} - 1\right\} \tag{6.16}$$

$$IF_2 = I_{nm}\left\{-j^{n} + (-1)^{n} j^{-m}\right\} \tag{6.17}$$

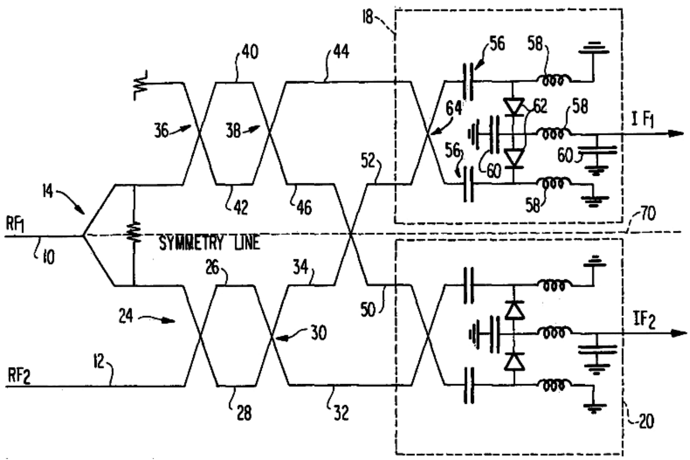
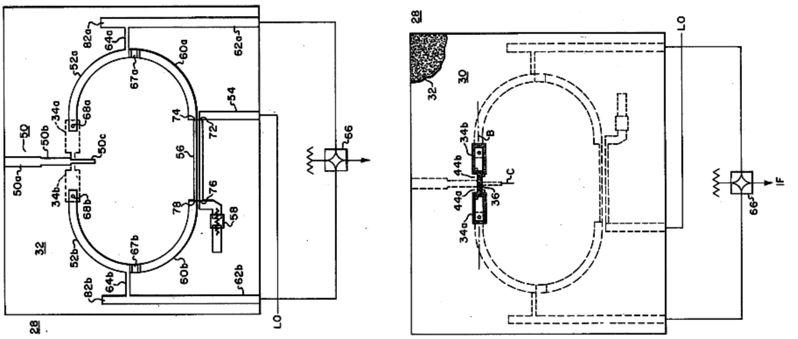



**Figure 6.88** Doubly balanced mixer using in-phase and quadrature couplers, with sum frequency at $IF_1$, and difference frequencies at $IF_2$.

A quadrature mixer using microstrip on the substrate top side, and CPW-slotline on the bottom side was disclosed in 1978 [89]. The top side is shown in Figure 6.89(a), and the bottom side is in Figure 6.89(b). The mixer combines two singly balanced mixers on the bottom side, each comprising two Schottky diodes, connected in-phase to the RF signal, and in quadrature for the LO input and IF output. The RF signal for each SB mixer is driven differentially using a top-side microstrip line that couples energy into the bottom side slot line that in turn drives each diode pair differentially. At this point, the slotline transitions to a CPW line that connects the LO and IF signals to the diodes, which appear in parallel to ground to the CPW lines.




(a) (b)

**Figure 6.89** Quadrature mixer using double sided substrate with microstrip on top side, and CPW-slotline on bottom side. (a) top side, (b) bottom side.

The other end of the CPW lines connect using vias to the top side microstrip lines, which in turn connect to the LO and IF quadrature couplers using a diplex filter to isolate LO from IF. An image rejection mixer product line that was popular in 1980s used this circuit arrangement but reversed the RF and LO ports to improve RF port VSWR and conversion loss.

## 6.6 SUBHARMONIC MIXERS

The basic theory of antiparallel-diode-pair (APDP) subharmonic mixers (SHM) is presented in Chapter 5, where it is shown that SHMs generate an IF output frequency equal to $\pm nf_{LO} \pm mf_{RF}$; generally n=2 and m=1 unless otherwise noted. The switching element of a n=2 SHM goes through two conduction cycles during each cycle of the fundamental LO, generating a conductance waveform rich in even harmonics and suppressed odd harmonics. The switching element capable of delivering this feature comprises two diodes, connected in parallel with the cathode of one connected to the anode of the other, and vice versa. This is known as the APDP. The relative amplitudes of the harmonics are optimized by adjusting the level of the fundamental LO that in turn varies the duty cycle of the switching waveform. With conventional balanced mixers, suppression of mixing products with even LO or even RF harmonics is achieved using baluns. In contrast, SHMs built using a single APDP achieve suppression without baluns, but because of the inherent suppression they are sometimes referred to as balanced. To add to the confusion multiple APDP can be connected using a combination of 0 – 180 degree baluns and/or quadrature couplers to achieve image rejection and improved suppression to undesired mixing products. Some desirable properties of SHMs are the very low level of LO second harmonic leakage power, low LO AM noise, and minimal circuit complexity.

A balanced subharmonic mixer dating to the 1930s uses two diode bridge quads connected together with the LO pumping them, and the RF and IF isolated from each other using balanced low-pass and high-pass diplex filters [90]. The circuit is shown in Figure 6.90 configured as an up-converter. The low frequency RF input is at port 24, the high frequency IF output passes through the load resistor 25, and the LO is applied at 17. The LO is isolated from RF and IF by means of circuit balance, and alternately switches the diode bridge quads on and off. The diode currents are analyzed as done in Chapter 5, with n and m, respectively, referring to LO and RF harmonics. The phase components of the diode currents from the left and right diode bridge quads passing to the IF output load 25 are:

$$
\begin{aligned}
i_{Lfet} &= i_{11} - i_{13} - i_{12} + i_{14} = (-1)^n\left[(-1)^m - 1\right] \\
i_{Right} &= -i_{41} + i_{43} + i_{42} - i_{44} = \left[(-1)^m - 1\right] \\
i_{Lfet} + i_{Right} &= \left[(-1)^n + 1\right]\left[(-1)^m - 1\right] \qquad (6.18) \\
&\neq 0 \text{ for n=even \& m=odd}
\end{aligned}
$$

By adding the left and right current components it is shown that the total current passing to the high frequency IF load contains only even LO and odd RF harmonics, as expected for the doubly balanced subharmonic mixer. Furthermore, since the current in the left and right diode quads each contain only odd RF harmonics and any LO harmonic, both support the RF current, which is $f_{RF} = 1f_{RF} \pm 0f_{LO}$. The LO current can also be shown to be supported in each diode quad.

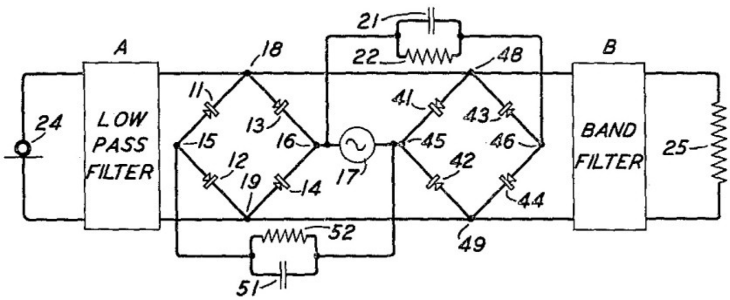
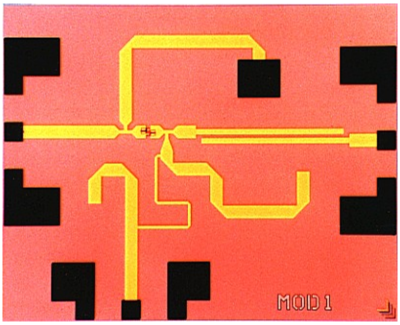



**Figure 6.90** Doubly balanced SHM using two diode bridge quads with balanced diplex filter for RF to IF isolation, 1930.

Diode SHM circuits have evolved so the majority of modern implementations use a single APDP in either a series or shunt configuration. The GaAs MMIC SHM of Figure 6.91 uses the series APDP with open and short circuit stubs to provide narrow-band filtering between the RF, LO, and IF. The mixer operates over the frequency range of 36-40 GHz, with typically 9.5 to 10.6 dB conversion loss, -69 dBm second LO harmonic leakage power, with fundamental LO drive at +6 dBm, [91]. The short circuit shunt stub left of the APDP provides the RF and IF ground returns while presenting an open circuit at the fundamental LO. On the right side of the APDP the open circuit shunt stub provides the LO ground return, the edge coupled capacitor blocks the IF and passes RF, while the series-L shunt-C distributed lowpass filter passes IF and blocks RF. The intended application for this SHM requires low second harmonic LO leakage power and circuit simplicity that equates to small size, making the circuit a good fit.

The SHM is analyzed by assigning the polarities to the diode currents, $i_1$ and $i_2$, and then finding the IF output current, $i_{IFout}$. Again the integers n and m, respectively, are LO and RF harmonics. The result is that mixing products where n ± m is an odd integer exit at the IF output, while products where n ± m is an even integer circulate around the APDP. Thus the desired $2f_{LO}$-$f_{RF}$ product appears at the IF output port, while the $f_{LO} \pm f_{RF}$ products cancel at the IF port. The phase components of the diode currents comprising the IF output current are given by (6.19).

$$i_1 = (-1)^m$$
$$i_2 = (-1)^n$$
$$i_{IFout} = i_1 - i_2 = (-1)^m - (-1)^n \tag{6.19}$$



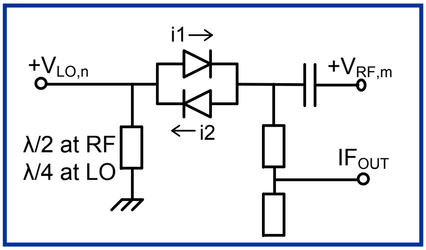




**Figure 6.91** A mmWave SHM with APDP with L-R isolation provided by the proper terminations and the series connection. From [91].

The SHM of Figure 6.92 uses the shunt APDP configuration, and operates at the astounding frequency range of 840 to 900 GHz. The SHM is used for this extremely high frequency application to improve feasibility of LO signal generation, [92]. The RF input on the left side transitions from waveguide to a beam-lead GaAs membrane that contains the APDP diodes connected shunt to ground immediately at the RF input to minimize loss. The fundamental LO is injected through the waveguide at the center of the assembly where it couples to the wide printed traced that forms a probe. LO energy propagates left towards the diodes due to the open circuit presented to the right caused by the high-impedance line. In contrast, the IF exits to the right, being conveyed from the diodes through the series-L shunt-C low pass filter.

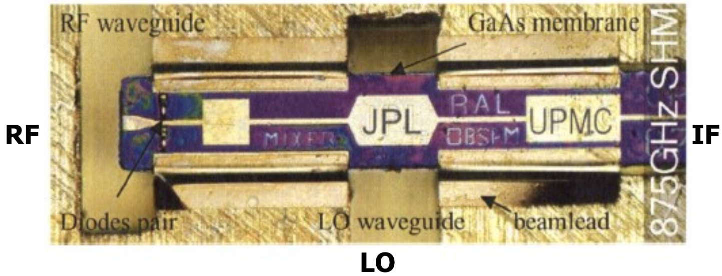



**Figure 6.92** An APDP connected in parallel to the signal line operating as a SHM at 896 GHz. From [92].

The SHM of Figure 6.93, [93], uses the series APDP but in a novel configuration that provides wider bandwidth while preserving circuit simplicity. The MMIC SHM operates over the frequency range of 23 – 37 GHz, with typically 9.4 – 12.0 dB conversion loss. The RF and LO are isolated by means of an edge coupled directional coupler that also injects both into the left side of the APDP. The shunt capacitor to ground at the right side of the APDP provides the RF and LO ground return. The IF is coupled from the APDP through the shunt-C series-L lowpass filter, and the IF ground return is provided on the left side of the APDP by the shorted quarter wave stub.

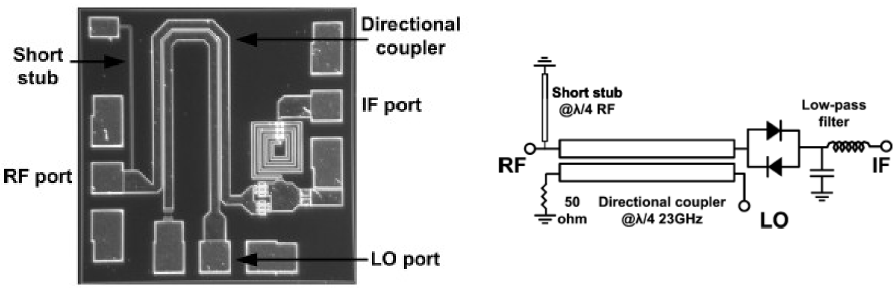



**Figure 6.93** Wideband series APDP SHM with L - R isolation provided by the coupler. From [93].

The SHM of Figure 6.94 uses two GaAs beam lead diodes in a ratrace configuration printed over a 0.015 inch thick alumina substrate, [94]. The SHM operates over the multi-octave frequency range of 10 – 110 GHz in both fundamental and subharmonic modes. Conversion loss in the SHM mode ranges from about 12 to 20 dB. DC bias is injected to allow for LO drive power down to 0 dBm.

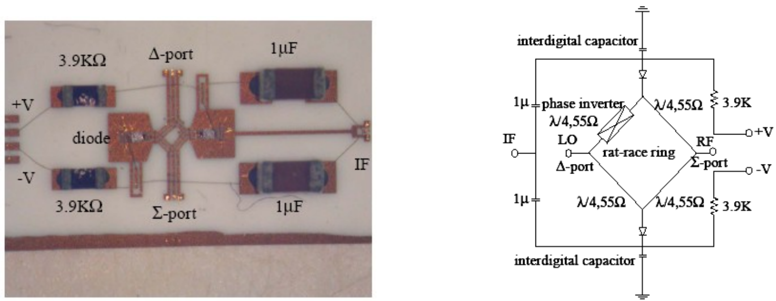



**Figure 6.94** Multioctave ratrace SHM. From [94].

The SHM of Figure 6.95 uses four APDP arranged in a ring, with balanced RF and LO connections to provide isolation between RF and LO, [95]. This is an improvement over a single APDP that provides isolation only by filtering. Thus the SHM of Figure 6.95 allows RF and LO frequencies to overlap, and operate over multioctave range. The SHM operates over the RF frequency range of 4 – 40 GHz, given a 2 – 20 GHz fundamental LO. The IF operates up to 1 GHz, so it can be diplex connected with the RF. Conversion loss is typically 8 to 12 dB over the 4 – 40 GHz range, with fundamental LO to RF isolation of typically 24 to 26 dB, and isolation of the second LO harmonic at typically 55 to 60 dB.

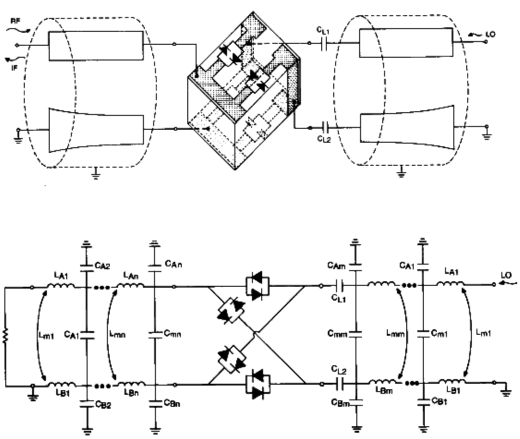



**Figure 6.95** Multioctave triply balanced SHM. From [95].

An alternative subharmonic mixer makes use of coplanar and slot lines as disclosed in patent [96]. The circuit in Figure 6.96 shows the transmission line schematic for a core cell of the mixer where LO signal is applied at node 6 and propagates in the CPW to modulate diodes 81, 82 with polarity depicted by the solid arrows. A high impedance slot line extends to the right of the diodes, which presents a high impedance to the LO, causing the majority of LO power to be delivered to the diodes. That slotline, however, delivers the RF signal power to the diodes with the polarity depicted by the dashed arrows. The diodes are in parallel to the LO signal, and in series to the RF signal. The RF and LO signals are isolated from each other by the mode orthogonality. The wire bond 7 equalizes the ground potential on the CPW; the distances d1 = d2 = $\lambda_{LO}/4$.

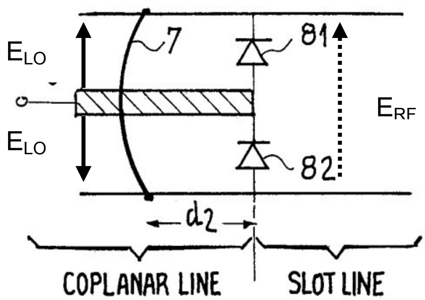



**Figure 6.96** SHM core cell, composed of coplanar line on left of diodes and slotline on the right.

The complete mixer is given in Figure 6.97 where two core cells are combined. The LO is combined on the left side by means of a coplanar Wilkinson power splitter. The RF signal is coupled on the right using a CPW line, with each half of the CPW line behaving as a slotline for each cell. Two diodes are in series in each slotline, and two of these are in parallel by combining the slotlines to form the CPW, so the RF input impedance averages to that of a single diode. The IF current is collected at the CPW center conductor by adding the currents from diodes $i_{15} + i_{16}$ and subtracting the currents from diodes $i_{17} + i_{18}$. The current in the diode, due to application of $V_{RF}$ and $V_{LO}$ voltages is approximately given by:

$$i_{15} = I_S e^{(V_{RF}+V_{LO})/V_T}$$
$$i_{16} = I_S e^{(V_{RF}-V_{LO})/V_T}$$
$$i_{17} = I_S e^{(-V_{RF}+V_{LO})/V_T}$$
$$i_{18} = I_S e^{(-V_{RF}-V_{LO})/V_T}$$

Adding the branch currents gives the IF current defined as:

$$i_{18} = 4I_S \cos[2\pi(2f_{LO} \pm f_{RF})t] \tag{6.20}$$

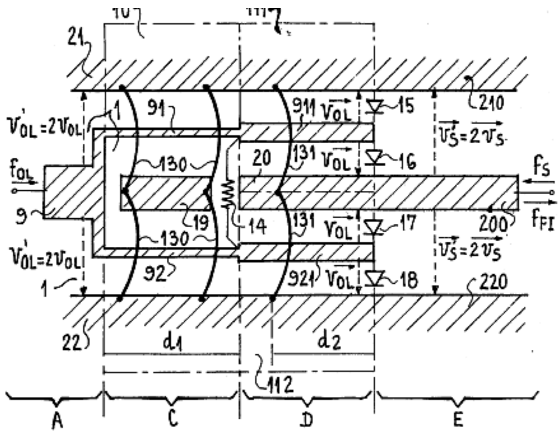
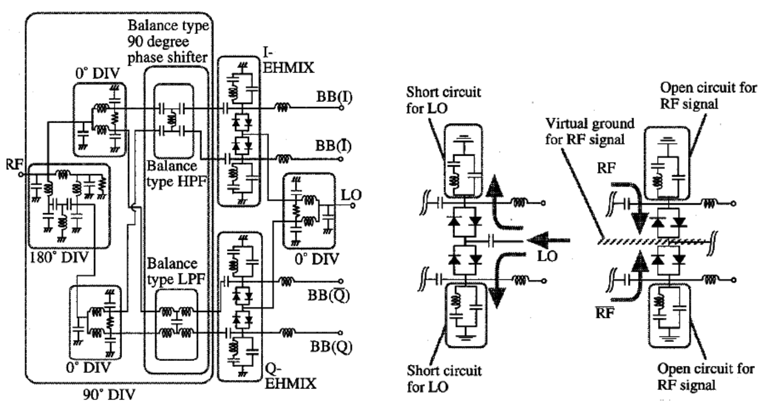


**Figure 6.97** Complete CPW/slotline SHM using four diodes. IF signal decoupling is by additional circuit not shown.

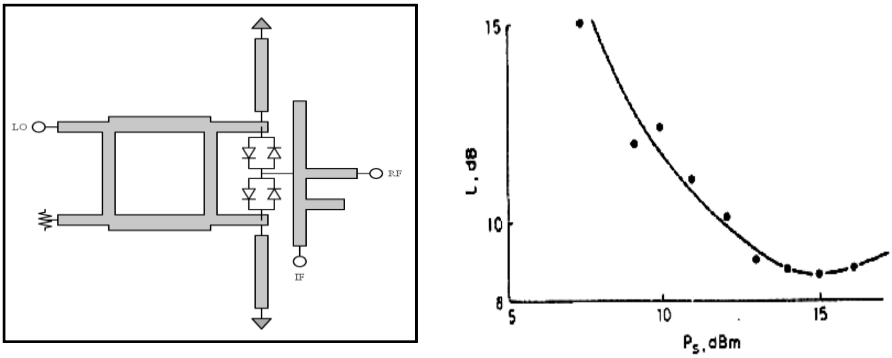
The SHM of Figure 6.98 uses two singly balanced APDP with the RF driven in quadrature, the LO driven in phase, and having two differential IF outputs that are in quadrature, [97]. A detailed view of current polarities in the diodes is shown in part (b) of Figure 6.98. The LO current splits into two halves that are routed through the diodes and return to ground. The RF current is applied in opposed phase to the diodes, sinking into a virtual ground at the joining point. The quadrature baseband down converter operates over the frequency range of 1.4 to 1.7 GHz, with typically 6 dB conversion loss.




(a) Mixer configuration ( b) Detail of currents on diodes

**Figure 6.98** Quadrature SHM with differential baseband output. From [97].

Many SHM applications require higher than the second LO harmonic. The SHM shown in Figure 6.99(a) operates with the fourth LO harmonic, [98]. The LO power must be divided in half with phase angles offset by $\pi/2$, which can easily be provided by 90° hybrid. The open stub on the right is at $\lambda_{LO}/4$ shorting the LO fundamental and is an open circuit to the RF signal.




(a) (b)

**Figure 6.99** Quadrature SHM using fourth harmonic LO. From [98].

The short stub is an open at the LO frequency and a ground return to the RF frequency. The conversion loss presented by the author is in Figure 6.99(b), showing nearly 8.5 dB conversion loss at a pump power of +15 dBm at 2.5 GHz. The signal frequency is at 10 GHz.

The SHM of Figure 6.100 operates with $12^{th}$, $14^{th}$, and $16^{th}$ LO harmonics using APDP configuration, [99]. This mixer was designed to convert an RF signal band from 35 to 40 GHz down to DC to 1.5 GHz using diodes built on a 0.15 μm HEMT process. The RF is filtered by means of a band-pass edge-coupled transmission line filter and a quarter wave length open stub filters RF on the IF, LO side. The IF/LO diplexer uses lumped LC elements. The circuit layout is in Figure 6.101 depicting a large area dedicated to the IF/LO diplexer and a smaller area to the RF side.

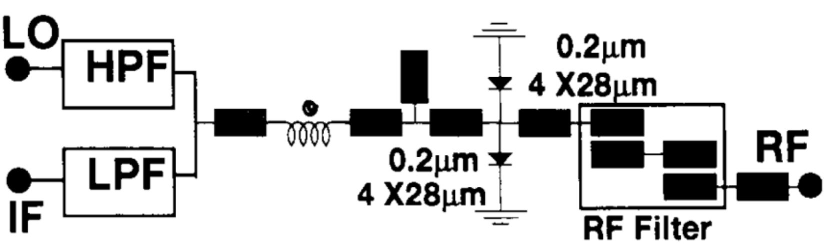



**Figure 6.100** A MMIC high harmonic order downconverter mixer. From [99].

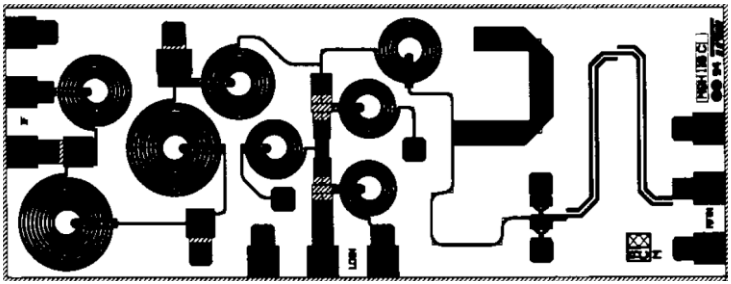


**Figure 6.101** Layout for the mixer occupying a space of 1x2.5 $mm^2$. From [99].

The conversion loss for the circuit operating at the 16th LO harmonic of a S band source (2 – 3 GHz) is on Figure 6.99, obtained from pumping + 17 dBm into the diodes. The LO to RF isolation is in the order of 65 dB.

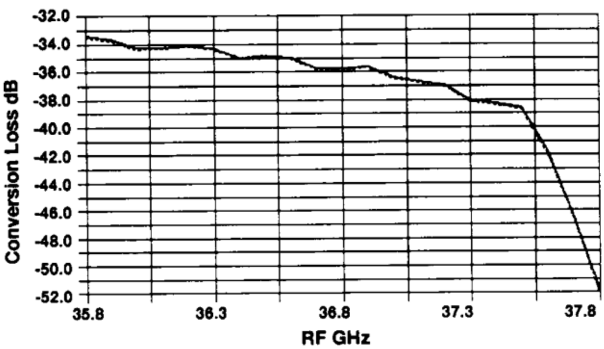



**Figure 6.102** Conversion loss versus RF frequency, with $F_{LO}$ = 2.23 GHz, $P_{LO}$ = +17 dBm. From [99].

## 6.7 SUMMARY

This chapter summarized the important diode mixer circuits and some lesser known ones that have interesting features. A discussion of their operating principles and performance levels was given. The various topologies presented include single ended, singly balanced, doubly balanced, and triply balance with TIM and Paramixer variations. Quadrature and image rejection mixers were also covered. This material is expected to be helpful to engineers by providing a starting point for new designs, and a view of what performance levels have already been achieved.

**REFERENCES**

[1] F.W. Young, "Carrier Wave Receiving System," US Patent 1,526,408, issued February 17, 1925.

[2] R.S. Ohl, "Detector System for Very Short Electric Waves," US Patent 2,378,944, issued June 26, 1945.

[3] Edward B. Craft and Jonathan W. Harris, "Alloy," US Patent 937,284, issued October 19, 1909.

[4] E.C. Cork et al, "Circuit Arrangements For Mixing Oscillations," US Patent 2,455,657, issued December 7, 1948.

[5] W.W. Salisbury, "High Frequency Mixer," US Patent 2,547,412, issued April 3, 1951.

[6] R.V. Pound, "Ultrahigh Frequency Mixer," US Patent 2,605,399, issued July 29, 1952.

[7] Alexander Horvath, "Crystal Mixer for Multiplex Broadcasting," US Patent 2,567,208, issued September 11, 1951.

[8] B.F. Wheeler et al, "High-Frequency Mixer Circuit," US Patent 2,616,037, issued October 28, 1952.

[9] E.G. James et al, "Crystal Contact Devices," US Patent 2,731,561, issued January 17, 1956.

[10] S.E. Howe, "Microwave Crystal Mixer," US Patent 2,961,531, issued November 22, 1960.

[11] Klaus Otremba, "Frequency Converting Device for RF Signals," US Patent 3,943,450, issued May 9, 1976.

[12] Ben R. Hallford, "Microwave Balanced Mixer Circuit," US Patent 3,659,206, issued April 25, 1972.

[13] Shinkawa et al, "Mixer Circuit Arrangement," US Patent 4,211,977, issued July 8, 1980.

[14] Saka et al, "Microwave Integrated Circuit Mixer," US Patent 4,450,584, issued May 22, 1984.

[15] Allen F. Podell, "Mixer Employing Dual Diode Circuit Biased in Their Exponential Operating Regions," US Patent 3,189,836, issued June 15, 1965.

[16] Risto Vaisanen, "Transformless Diode Mixer," US Patent 5,379,458, issued January 3, 1995.

[17] Allen F. Podell, "Balanced Mixer Circuit and Inductive Device Usable Therein," US Patent 3,327,220, issued June 20, 1967.

[18] Pierre Dobrovlny, "Radio Frequency Mixer," US Patent 4,823,400, issued April 18, 1989.

[19] A. A. M. Saleh, *Theory of Resistive Mixers*, PhD Thesis, MIT, Massachusetts, January 1970, pp. 24 - 35, 94-97, and 116-137.

[20] Dean F. Peterson, "High-Intercept Band Limited Mixer," US Patent 5,493,720, issued February 20, 1996.

[21] Ryo Tatsumi, Mitsuhiro Matsumoto, and Tomoya Bando, "Mixer and Converter Using Same," US Patent 7,292,837B2, issued November 6, 2007.

[22] Paul J. Pontecorvo, “Modulator,” US Patent 2,462,893, issued March 1, 1949.

[23] George Rodwin, “Balanced Crystal Microwave Converter,” US Patent 2,576,481, issued November 7, 1951.

[24] Warren A. Tyrrell, “Coupling Arrangement for Use in Wave Transmission Systems,” US Patent 2,445,895, issued July 27, 1948.

[25] George Ctirad Spacek, “High Frequency Converter,” US Patent 3,584,306, issued June 8, 1971.

[26] George Ctirad Spacek, “High Frequency Converter,” US Patent 3,638,126, issued January 25, 1972.

[27] Erich H. Kraemer and John C. Rolfs, “Integrated Circuit Planar High Frequency Mixer,” US Patent 4,276,655, issued June 30, 1982.

[28] John Reindel, “Millimeter Wave Printed Circuit Mixer,” US Patent 4,406,020, issued September 20, 1983.

[29] Chi P. Hu, “Millimeter Wave Stripline Planar Mixer,” US Patent 4,412,354, issued October 25, 1983.

[30] Robert N. Bates and Philip M. Ballard, “R. F. Balanced Mixer,” US Patent 4,542,535, issued September 17, 1985.

[31] Howard Putnam, “Strip Line Hybrid Ring and Balanced Mixer Assembly,” US Patent 3,310,748, issued March 21, 1967.

[32] Ben R. Hallford, “Microwave Balanced Mixer Circuit,” US Patent 3,659,206, issued April 25, 1972.

[33] Keiro Shinkawa and Chuichi Sodeyama, “Mixer Circuit,” US Patent 4,249,263, issued February 3, 1981.

[34] Akira Takayama, “Frequency Conversion Circuit,” US Patent 4,677,693, issued June 30, 1987.

[35] Dominique Podvin, “Balanced High-Frequency Mixer Having a Balun Circuit,” US Patent 5,142,697, issued August 25, 1992.

[36] George D. Vendelin, “Stripline Compensated Balun and Circuits Formed Therewith,” US Patent 3,571,722, issued March 23, 1971.

[37] Lawrence E. Dickens and Douglas W. Maki, “Integrated Circuit, Image and Sum Enhanced Balanced Mixer,” US Patent 3,939,430, issued February 17, 1976.

[38] Eugene C. Findley, “Coplanar Microwave Balun, Multiplexer and Mixer Assemblies,” US Patent 4,739,519, issued April 19, 1988.

[39] H. Ogawa, M. Aikawa, and K. Morita, “K-Band Integrated Doubly balanced Mixer,” *MTT-28*, March 1980, pp. 180-185.

[40] M. Aikawa, H. Ogawa, “A New MIC Magic-T Using Coupled Slot Lines,” *IEEE Transactions on Microwave Theory and Techniques*, Volume MTT 28, June 1980, pp. 679-683.

[41] John Reindel, “Microcircuit Reverse-Phased Hybrid Ring Mixer,” US Patent 3,950,703, issued April, 13, 1976.

[42] Maurice Marchand and Yan Haentjens, “Ultra-High Frequency Hybrid Mixer,” US Patent 4,377,005, issued March 15, 1983.

[43] Clement F. Lacour and Yannick Latouche, “Microwave Mixer,” US Patent 454,112 issued September 10, 1985.
[44] Philip R. Thomas and Tom M. Hyltin, “Microwave Integrated Circuit Mixer,” US Patent 3,416,042, issued December 10, 1968.
[45] Tom M Hyltin, “Integrated 94 GHz Local Oscillator and Mixer,” US Patent 3,534,267, issued October 13, 1970.
[46] Stephen J. Nightingale, “Singly balanced Planar Mixer,” US Patent 4,607,394, issued August 19, 1986.
[47] C. K. Keith, "Frequency Translating System," US Patent 1,855,576, issued April 26, 1932.
[48] F. A. Cowan, "Modulating System," US Patent 2,025,158, issued June 7, 1934.
[49] Pierre Dobrovolny, “Integrated Mixer Arrangement,” US Patent 4,152,653, issued May 1, 1979.
[50] Xiaohui Li, “Compensated Ring Mixer,” US Patent 5,854,974, issued December 29, 1998.
[51] Martin L. Grogan, “Modulation Device for Substituting for a Doubly balanced Mixer,” US Patent 4,313,094, issued January 26, 1982.
[52] R. B. Mouw, “Hybrid Junctions and Mixers and Modulators Constructed There from,” US Patent 3,512,090, issued October 18, 1967.
[53] W. Hahnle, “Arrangement for Modulation,” US Patent 2,144,655, issued January 24, 1939.
[54] George Ctirad Spacek, "High Frequency Converter," US Patent 3,638,126, issued January 25, 1972.
[55] Joseph S. Wong, Kuo-Ing Chung, Kenneth A. Rudenauer, “Orthogonal Hybrid Fin Line Mixer,” US Patent 4,480,336, issued October 30, 1984.
[56] R. M. Walker and Nicholas P. Kernweis, “Orthogonal Hybrid T for Microwave Waveguides,” US Patent 3,192,489, issued September 15, 1961.
[57] Y. Lai, C. Lin, C Lin, Y. Wang, "A New Ka-Band Doubly Balanced Mixer Based on Lange Couplers," *IEEE Microwave and Wireless Components Letters*, Volume 18, No. 7, July 2008, pp. 458-460.
[58] Sadao Igarashi, “Doubly balanced Mixer Circuit,” US Patent 4,306,311, issued December 15, 1981.
[59] Stephen B. Einbinder, “Balanced Split Ring Resonator,” US Patent 5,361,050, issued November 1, 1994.
[60] Rod Kirkhart, "Low Cost Planar Double Balanced Mixer," US Patent 7,305,217B2, issued December 4, 2007.
[61] S. Basu, S. Maas, “Design and performance of a Planar Star Mixer,” *IEEE Transactions on Microwave Theory and Techniques*, Volume MTT 41, No. 11, November 1993, pp. 2028-2030.
[62] C. Kuo, C. Kuo, C. Kuo, S. Maas, Huei Wang, “Novel Miniature and Broadband Millimeter-wave Monolithic Star Mixers,” *IEEE Transactions on Microwave Theory and Techniques*, Volume MTT 56, No. 4, April 2008, pp. 793-802.

[63] Donald Neuf, “Doubly Balanced Microwave Mixer Using Balanced Microstrip Baluns,” US Patent 3,652,941, issued March 28, 1972.
[64] C.Y. Chang, et al, “Ultra-Broad-Band Doubly Balanced Star Mixers Using Planar Mouw’s Hybrid Junction,” *IEEE Transactions on Microwave Theory and Techniques*, Volume 49, No. 6, June 2001, pp. 1077-1085.
[65] H. Ogawa, M. Aikawa, and K. Morita, “K-Band Integrated Doubly Balanced Mixer,” *IEEE Transactions on Microwave Theory and Techniques*, Volume MTT 28, March 1980, pp 180-185.
[66] M. Aikawa, H. Ogawa, “A New MIC Magic-T Using Coupled Slot Lines,” *IEEE Transactions on Microwave Theory and Techniques,* Volume MTT 28, June 1980, pp. 679-683.
[67] Eugene C. Findley, “Coplanar Microwave Balun, Multiplexer and Mixer Assemblies,” US Patent 4,739,519, issued April 19, 1988.
[68] R. H. Merriam, "Quadruple-Balance Mixer," US Patent 3,515,993, issued October 24, 1967.
[69] Robert A. Sowadski, "Spurious Product and Local Oscillator Radiation Suppression System for Use in Super-Heterodyne Radio Receivers," US Patent 5,179,728, issued January 12, 1993.
[70] T. N. Ton, G. S. Dow, T. H. Chen, M. Lacon, T. S. Lin, S. Bui, and D. Yang, "An X-Band Monolithic Double Double-Balanced Mixer for High Dynamic Receiver Application," *IEEE MTT 1990 International Microwave Symposium Digest*, pp. 197-200.
[71] J. Eisenberg, J. Panelli and W. Ou, "A New Planar Double Doubly-Balanced MMIC Mixer Structure," *IEEE Microwave and Millimeter Wave Monolithic Circuits Symposium,* 1991, pp. 69-72.
[72] Hwann-Kaeo Chiou and Hao-Hsiung Lin, "A Miniature MMIC Double Doubly Balanced Mixer Using Lumped Dual Balun for High Dynamic Receiver Application," *IEEE Microwave and Guided Wave Letters*, Volume 7, No. 8, August 1997, pp. 227-229.
[73] Thomas A. Milligan and Peter Will, "Broadband High Frequency Mixer," US Patent 4,063,176, issued December 13, 1977.
[74] Anthony M. Pavio, "Broadband High Baluns and Mixer," US Patent 4,125,810, issued November 14, 1978.
[75] Warren L. Seely, "Broadband Mixer with Coplanar Balun," US Patent 4,355,421, issued Octpber 19, 1982.
[76] B. C. Henderson, "Full Range Orthogonal Circuit Mixers Reach 2 to 26 GHz," *Microwave System News*, January 1982, Volume 12, No. 1, pp. 122-126.
[77] R. S. Bischoff, R. C. Blanc and P. B. Harper, "New Technology in Synthesized Sweeper Microcircuits," *Hewlett-Packard Journal*, April 1991, pp. 36-45.
[78] William J. Drennen, "Planar Triply-Balanced Microstrip Mixer," US Patent 5,740,528, issued April 14, 1998.
[79] Radha Setty and Daxiong Ji, "Triple Balanced Mixer," US Patent 6,917,796 B2, issued June 12, 2005.

[80] Doron Gamliel, "Ultra High Isolation Mixer," US Patent 7,013,122 B2, issued March 14, 2006.
[81] R. S. Caruthers, "Double Balanced Modulating System," US Patent 2,205,843, issued June 25, 1940.
[82] Peter Will, "Broadband Doubly Balanced Mixer Having Improved Termination Insensitivity Characteristics," US Patent 4,224,572, issued September 23, 1980.
[83] Donald H. Steinbrecher, "Doubly Balanced Mixer with Optimized Dynamic Range," US Patent 4,112,374, issued September 5, 1978.
[84] Donald H. Steinbrecher, "Doubly Balanced Mixer with Optimized Dynamic Range," US Patent 4,230,956, issued October 28, 1980.
[85] F. X. Li, G. Corsetto, "Multi-Order Cancellation Technology," *Microwave Journal*, October, 1997, pp. 84-91.
[86] Theodore S. Saad, "High Frequency Signal Combining," US Patent 3,532,989, issued January 24, 1966.
[87] Robert Tenenholtz, "Balanced Mixer," US Patent 3,706,042, issued December 12, 1972.
[88] William H. Childs and Christopher E. Mahle, "Planar QPSK Demodulator," US Patent 4,281,293, issued June 28, 1981.
[89] Lawrence E. Dickens, "Image Phased and Idler Frequency Controlled Mixer Formed on an Integrated Circuit Dielectric Substrate," US Patent 4,118,670, issued October 3, 1978.
[90] R. S. Caruthers, "Modulating System," US Patent 2,086,601, issued June 13, 1937.
[91] Kenji Itoh, Akio Iida, Yoshinobu Sasaki, Shuji Urasaki, "A 40 GHz Band Monolithic Even Harmonic Mixer with an Antiparallel Diode Pair," *1991 IEEE MTT International Microwave Symposium Digest,* pp. 879-882.
[92] B. Thomas, A. Maestrini, D. Matheson, I. Mehdi, P. de Maagt, "Design of an 874 GHz Biasable Sub-Harmonic Mixer Based on MMIC Membrane Planar Schottky Diodes," *33$^{rd}$ International Conference on Infrared, Millimeter and Terahertz Waves*, IRMMW-THz, September 2008, pp. 1-2.
[93] Che-Hung Lin, Yu-Ann Lai, Jui-Chieh Chiu, and Yeong-Her Wang, "A 23–37 GHz Miniature MMIC Subharmonic Mixer,, *IEEE Microwave and Wireless Components Letters,* Volume 17, No. 9, September 2007, pp. 679-681.
[94] Chun-Hsiang Chi, Chung-Hung Wu, Wei-Ting Wang, Chi-Hau Lai, Dow-Chih Niu, and Chi-Yang Chang, "A 10-110GHz Fundamental/Harmonic Rat-Race Mixer," *Proceedings of the 37th European Microwave Conference*, Munich Germany, October 2007, pp. 652-655.
[95] J. L. Merenda, D. Neuf, P. Piro, "4 to 40 GHz Even Harmonic Schottky Mixer," *IEEE 1988 MTT International Microwave Symposium Digest*, pp. 695-698.
[96] Michel Houdart, "Microwave Subharmonic Mixer Device," US Patent 4,485,488, issued November 27, 1984.

[97] Mitsuhiro Shimozawa, Kenji Kawakami, Hiroshi Ikematsu, Kenji Itoh, Nobuyuki Kasai, Yoji Isota, and Osami Ishida, "A Monolithic Even Harmonic Quadrature Mixer Using a Balance Type 90 Degree Phase Shifter for Direct Conversion Receiver," *IEEE 1998 MTT International Microwave Symposium Digest*, pp. 175-178.
[98] J. D. Buchs, Gunther Begemann, "Frequency Conversion Using Harmonic Mixers with Resistive Diodes," *IEE Journal of Microwave Optics and Acoustic*, Volume 2, May 1978, pp. 71-76.
[99] R. Katz, S. Maas, A. Sharma, D. Smith, "A Novel Monolithic HEMT Harmonic Mixer At Q-Band," *IEEE 1995 Microwave and Millimeter-Wave Monolithic Circuits Symposium,* pp. 39-42.