

easy-to-build ssb transceiver

for 1296 MHz

Complete description
of a simple
sideband system
for the amateur
23-cm band —
the same technique
can be used
for the other
uhf bands

The simple microwave ssb system presented here was used to achieve Northern California's first recent two-way ssb communications on 1296 MHz, between WA6UAM and K6UQH, on April 14, 1974.* Aside from any precedent which may have been established, the method used to transmit and receive microwave ssb represents a significant breakthrough in that it is simple, straightforward, in-

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expensive and readily reproducible by any uhf enthusiast. Neither specialized tools nor elaborate test equipment is required to build this equipment — equipment that provides the capability for line-of-sight ssb contacts on the amateur 23-cm band.

Fig. 1 shows the scheme generally used for the transmission and reception of uhf ssb signals. The received signals are down-converted in the conventional manner into the high-frequency spectrum where they are detected by the station receiver. Similarly, a high-frequency signal generated by the station exciter is heterodyned up to uhf, then amplified and transmitted. Note the high degree of redundancy present in this system. Both the transmit and receive converters use a mixer and local-oscillator chain, the function of each being essentially identical to its counterpart.

Assuming the availability of a bilatera

*Some years ago, K6HCP and W6GDO ran successful 1296-MHz ssb schedules, each using a 2C39 as a simultaneous mixer and LO doubler. The resulting ssb output was easily copyable, if not exactly spectrally pure. Bandpass filters were used to attenuate the undesired products at the antenna. The equipment currently used by WA6UAM and K6UQH, described in this article, uses diode balanced mixers with injection at the ultimate LO frequency. This method of heterodyning produces clean 1296-MHz ssb without excessive intermodulation products

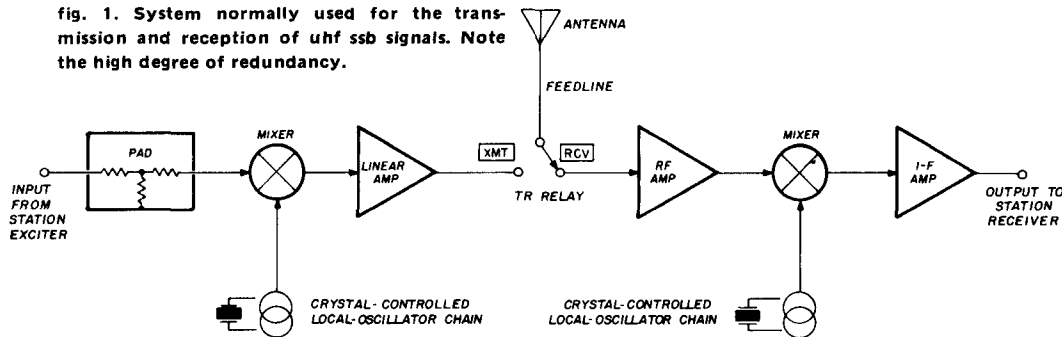
Editor

mixer (one which operates equally well in both the *forward* and *backward* directions), the system can be simplified as indicated in fig. 2. Obviously, a passive mixer must be used in this application. Any active device designed to provide conversion gain in, say, the *up* direction,

of numerous stages of linear amplification *after* the transmit conversion.

It is evident from fig. 2 that by eliminating redundant circuitry, the TR switching complexity has increased three-fold. Assuming that means could be found for eliminating the requirements

fig. 1. System normally used for the transmission and reception of uhf ssb signals. Note the high degree of redundancy.



cannot function as a *down* mixer as well. Fortunately, singly- and doubly-balanced diode mixers function effectively in either direction, with only a few dB of conversion loss.

The greatest drawback of the diode mixer, so far as transmit conversion is concerned, is its limited power-handling capability: This normally requires the use

for separate receive and transmit amplification, TR relays K1 and K2 could then be eliminated, too. Of course, such drastic simplification would jeopardize both the receive sensitivity and transmit power. However, depending upon the application, these tradeoffs might well be justified. Such was the case with the 1296-MHz station at WA6UAM.

The details for the Simple 1296-MHz Sideband System are shown in fig. 3. Equal emphasis was placed on simplifying the system to its minimum required content, and optimizing each sub-assembly to provide reliable communications over a reasonable range. Free-space loss and receiver noise calculations indicate that ssb communications between two such stations would be practical to distances of at least 100 miles (160 km).

Note the total elimination of TR relays and feedlines (and their resultant losses) in the microwave portion of this system. This is accomplished by mounting the mixer, filter and LO chain directly at the antenna (readily accomplished, as these modules are both lightweight and relatively small), and pumping only 28-MHz energy (plus 12 Vdc for the local oscillator) up and down the tower. The rf modules and antenna might even be

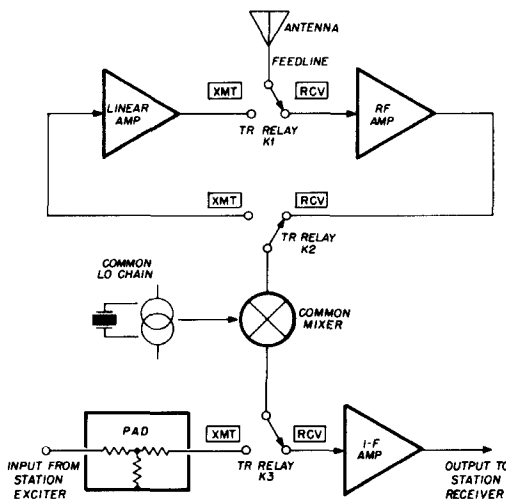


fig. 2. This uhf ssb system, which uses a common local-oscillator chain and balanced diode mixer, minimizes circuit redundancy but requires three TR relays.

combined into a single physical unit, as shown in fig. 4.

mixer

So far as design tradeoffs are concerned, the mixer is, by far, the most critical component of the Simple Sideband System. To obtain a reasonable receive noise figure, low conversion loss is of paramount importance. At the same time usable transmit rf levels dictate high power-handling capabilities. As will be shown in a minute, these two criteria tend to be mutually exclusive. With readily available Schottky-barrier (hot-carrier) diodes in a balanced mixer, the system seems to optimize at about 6-dB conversion loss, with 3 mW of usable output power. Don't scoff at these seemingly restrictive figures. Calculations (see page 21) will show that this type of performance is more than satisfactory for communications to the edge of the visual horizon, and perhaps beyond.

Several reproducible uhf balanced mixers have been published recently.^{1,2,3,4} The balanced mixer presented here is based upon a design developed by W6FZJ, and currently used by him on 2304 MHz. Versions for 1296 MHz have been built by both WA6UAM and WB6JNN, and provide a considerable improvement over the single-diode trough-line or interdigital designs frequently used by amateurs in uhf transmitting and receiving converters. An improved version of the W6FZJ mixer, which uses a commercially available balun to match the rf port of the mixer to the 50-ohm transmission line, will also be described.

Whatever mixer design is chosen, the diodes you select will determine its conversion loss and power-handling capability. One high-power, low-cost device is the Hewlett-Packard 5082-2817.*

*Hewlett-Packard 5082-2817 hot-carrier diodes are available in small quantities for \$1.50 each from any Hewlett-Packard sales office. Matched pairs (5082-2818, \$3.25), and matched quads (5082-2819, \$7.00) are also available. If you can't find a Hewlett-Packard sales office in the *Yellow Pages*, write to Hewlett-Packard, 1101 Embarcadero Road, Palo Alto, California 94303.

These diodes have a burnout rating of 4.5-watts peak, or 1-watt CW, and are capable of conversion efficiencies of better than -5 dB.

practical mixer circuit

The complete 1296-MHz mixer shown in fig. 5 uses hybrid construction (discrete components on etched microstripline), making duplication relatively straightforward for anyone with access to PC board fabrication facilities. Given sufficient time and patience, you can even "etch" your substrate with an Exacto knife and straightedge. At least four such manual efforts have been completed to date, and performance is equal in all respects to photochemically produced versions.

The equivalent circuit shown in fig. 6 will help to clarify the operation of the mixer. Rf energy injected into the delta (Δ) port is transformed by the balun so that there is a 180° phase difference between the signals applied to the two diodes. The diodes are effectively in series and of like polarity so that the applied rf simultaneously biases both diodes on and then off, for alternate half-cycles.

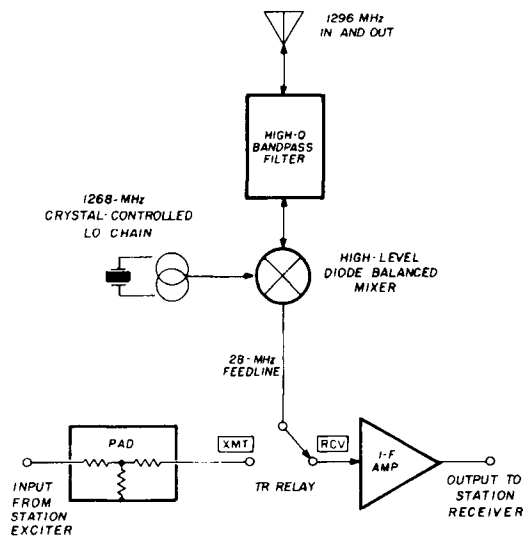


fig. 3. The Simple Sideband System, shown here, has been reduced to its ultimate simplicity but can still provide beyond-the-horizon communications when used with a high-gain antenna.

Rf energy applied to the sigma (Σ) input is transmitted down a quarter-wavelength, two-way power divider (which looks like a tuning fork) so that the signals it applies to the two diodes appear in phase. Since the diodes are in reverse polarity with respect to their bias return (common junction), this rf is applied to the two diodes out of phase. The simultaneous application of in- and out-of-phase rf signals to the diode pair results in a signal at their junction composed of the rf applied to the sigma (Σ) port, chopped at the rate of the rf applied to the delta (Δ) port. This complex repeating waveform can be shown by Fourier analysis to contain components of the sigma frequency, the delta frequency, their sum and their difference. Mixing, by the traditional definition, has occurred. The circuitry shown to the right of the diodes in **fig. 6** serves the purposes of signal conditioning (filtering out all but the difference-frequency component), dc bias return and a means of measuring diode bias current.

The most significant advantage of the balanced mixer over a single-ended design is that rf injected into the delta port is isolated from the sigma port, and vice-versa. To see how this is accomplished, consider a balanced signal applied to the two diodes through the delta port. In addition to feeding the diodes, this signal is shunted by the sigma port's power divider. Note that the power divider appears to this signal as a balanced transmission line shorted at the load end. Since this transmission line is a quarter-wavelength long at the delta frequency (assuming the delta and sigma signals are close in frequency), it transforms the short to an open, and the sigma port is effectively nonexistent so far as the delta signal is concerned.

Conversely, rf injected into the sigma port divides down the power divider, and appears to the diodes as two signals, equal in amplitude and phase. Looking toward delta, these two signals are cancelled in the balun and, thus, never reach the delta port.

It should be noted that single-balanced mixers provide no isolation whatever between the i-f port and either the delta or sigma port. Hence, filtering is required to remove the higher-frequency components from the i-f. Such filtering is accomplished in the hybrid balanced mixer by virtue of stubs at the i-f side of each diode, a quarter-wavelength long at 1296 MHz and open at the far end. These

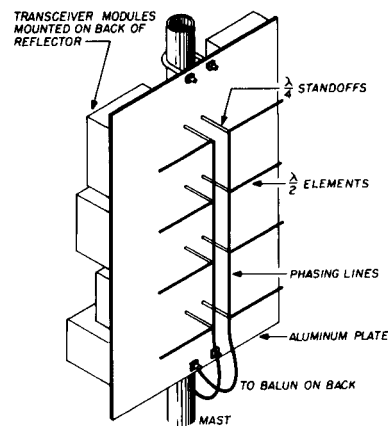
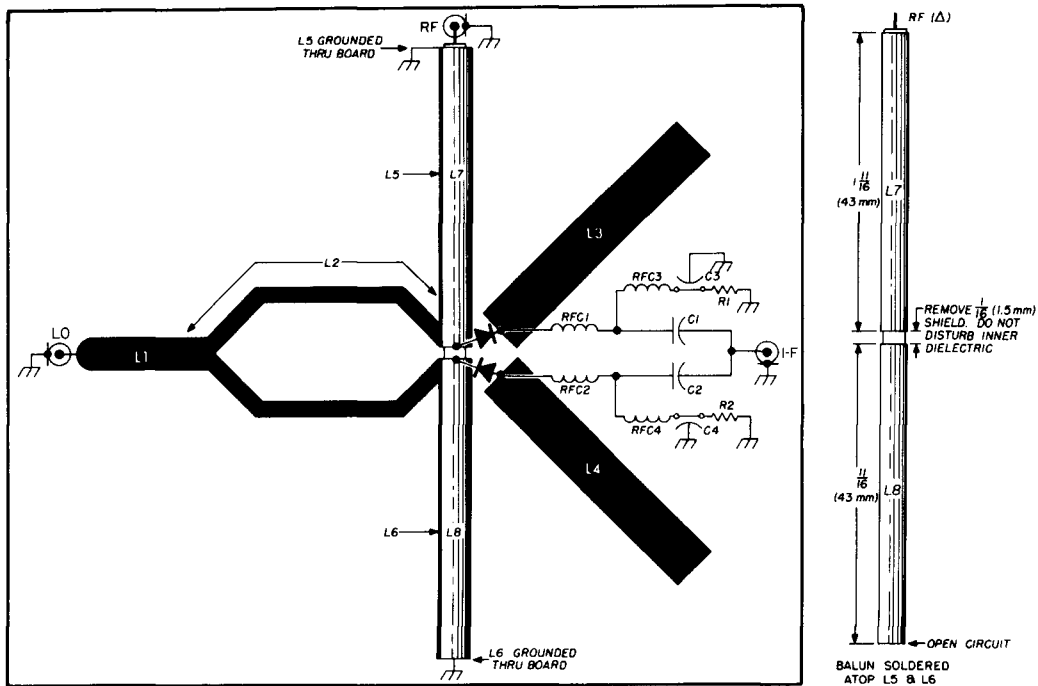


fig. 4. Antenna-mounted transceiver converter. With this arrangement, using the system of **fig. 3**, only 28-MHz energy must be run down the tower to the operating room.

quarter-wavelength sections ground out the i-f port to energy near 1296 MHz.

Derivation of the balun used in the hybrid balanced mixer is shown in **fig. 7**. **Fig. 7A** is a coaxial balun frequently used in antenna work. In **fig. 7B**, the same balun is straightened out to improve symmetry. Note that a common ground is still required between the far end of the stub and a point one quarter-wavelength back on the feedline. This ground is frequently accomplished by connecting to the walls of the half-wave-long box in which the balun is built. In the case of the hybrid balanced mixer, the return is to a ground plane, on the opposite side of a substrate to which the balun is attached.

Both bandwidth and balance may be improved by modifying the in-line balun



C1,C2	0.01 μ F disc ceramic	L5,L6	50-ohm micro-stripline, 0.168" (4.5 mm) wide, 1-19/32" (40.5 mm) long
C3,C4	1000-pF feedthrough	L7,L8	50-ohm UT-141 coaxitube, 0.141" (3.5 mm) diameter, 1-11/16" (43 mm) long
CR1,CR2	hot-carrier diodes (H-P 5082-2818)	R1,R2	10-ohm, 1/4-watt carbon composition resistors
L1	50-ohm micro-stripline, 0.168" (4.5 mm) wide, any length	RFC1,RFC2	2" (51 mm) no. 32 wire, close wound on 0.050" (1.5 mm) diameter form or ferrite beads on leads of C1 and C2
L2	75-ohm micro-stripline, 0.080" (2.0 mm) wide, 1-7/16" (36.5 mm) long, (along center)	RFC3,RFC4	22 μ H
L3,L4	38-ohm micro-stripline, 0.25" (6.5 mm) wide, 1-7/16" (36.5 mm) long		

fig. 5. This high-performance 1296-MHz balanced mixer uses etched 1/16" (1.5 mm) thick Teflon-fiberglass printed-circuit board and a coaxial balun. Full-size printed-circuit layout is shown in fig. 8. An equivalent circuit of this mixer, illustrating circuit operation, is given in fig. 6.

of fig. 7B as shown in fig. 7C. In this version the stub is a piece of coax identical to the original feedline. The center connector of the feeder is now connected to the center conductor, rather than to the shield of the stub. At the far end of the stub the center conductor is open. A quarter-wavelength toward the source (at the junction of the feeder and the stub) this open is transformed to a short, and rf sees the center conductor of the stub as being continuous with the shield. Therefore, the circuit at fig. 7C is

electrically identical to that of fig. 7B, but with improved physical symmetry. Balanced output is taken from the same point as before.

Note that at the far end of the stub the center conductor must be open, and the shield grounded. Again the balun may be constructed upon a substrate, with return through it to a groundplane.

The balun used in the 1296-MHz mixer is made from a single piece of UT-141 type semirigid coax (50-ohms, Teflon dielectric, 0.141-inch [3.5-mm]

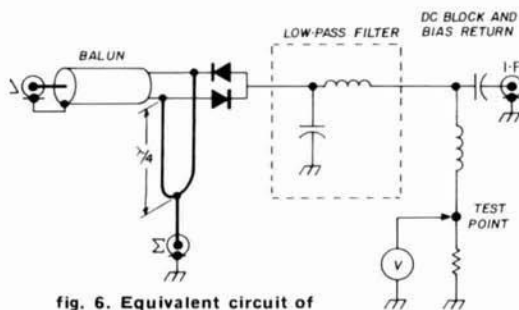


fig. 6. Equivalent circuit of the balanced mixer shown in fig. 5.

OD). Correcting for velocity factor, the quarter-wavelength sections are each 1-11/16-inches (43-mm) long. Judicious use of an Exacto knife and small tubing cutter will aid in the removal of 1/16-inch (1.5-mm) of the outer conductor at the junction of the feeder and the stub. Since no physical connection must be made to the center conductor at this junction, the Teflon dielectric should not be disturbed. Allow a short length of center conductor to extend beyond the quarter-wavelength section comprising the feeder coax. This will be attached to the center pin of the delta port's coax connector.

The mixer's substrate is etched on one side of a 1/16-inch (1.5-mm) thick, double-sided, 1 ounce, copper-clad, Teflon-fiberglass PC board. *Do not* use fiberglass-epoxy board, as its dielectric constant is not correct for the dimensions provided in fig. 5. The use of Teflon PC board is necessary in this design so that the velocity of propagation (and hence the electrical wavelength) of the striplines will approximate that of the coax balun. The full-sized PC layout is shown in fig. 8. All micro-striplines must be opposite a groundplane (the other unetched side of the double-sided board).

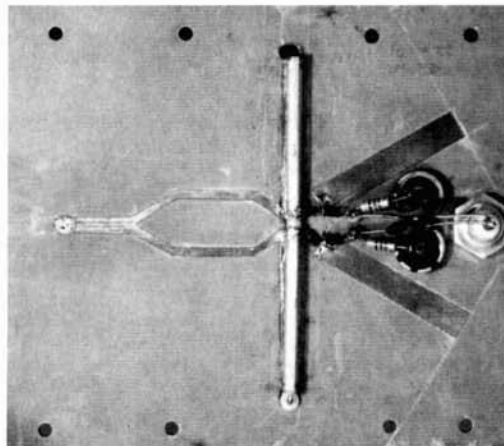
Although Teflon PC board makes an excellent substrate for micro-striplines at 1.3 GHz, it is quite expensive (and in

*Anzac model TP-101, 500 kHz to 1.5 GHz, 50-ohms balanced to 50-ohms unbalanced transformer with midband insertion loss of 0.4 dB maximum and vswr 1.6:1 maximum, \$15.50 in single quantities from Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154.

some areas, totally unobtainable). The use of fiberglass-epoxy board, though it would increase losses slightly, would bring this type of equipment within reach of many experimenters who might otherwise be deterred. Of course, a glass-epoxy substrate is incompatible with the UT-141 coax balun because of the widely different velocities of propagation of the two mediums. In order to develop a 1296-MHz balanced mixer on glass-epoxy board, a different method of unbalanced-to balanced transformation is required.

Anzac Electronics manufactures an appropriate balun of moderate cost, excellent electrical performance and small physical size which frees the mixer design from restrictions as to substrate material.* Mixers built on 1/16-inch (1.5-mm) G-10 double-clad PC board using the Anzac balun exhibited improved matching at the rf port, an effect which more than offsets any additional losses in the glass-epoxy dielectric.

A schematic of the improved 1.3-GHz balanced mixer using the commercial balun is shown in fig. 9. Note that pins 1, 3 and 5 of the balun must be grounded *through* the substrate to the groundplane. When mounting the balun, do not allow its case to short out the striplines of L2.



Construction of the doubly-balanced mixer of fig. 5, showing installation of the coaxial balun. Capacitors C1 and C2 can be seen between the two feedthrough capacitors.

Full-size artwork for a mixer board on 1/16-inch (1.5-mm) G-10 PC board ($\epsilon = 4.8$) is provided in fig. 10. Either of these two balanced-mixer designs (fig. 5 or fig. 9) will provide satisfactory performance in the Simple Sideband System for 1296 MHz.

output power - noise figure tradeoff

To avoid excessive intermodulation distortion in the transmit mode, it is desirable to inject into the i-f port a signal level at or below the mixer's 1-dB compression point. This is the level beyond which incremental increases in input power result in an ever-diminishing increase in output power. Such a situation typically occurs with i-f injection 5-dB below the local-oscillator signal level.

Due to the ready availability of any desired signal level at 28 MHz, the i-f injection level is not considered a limiting factor in system design. The following discussion assumes the use of the optimum i-f injection level in the transmit mode; that is, 5-dB below whatever LO insertion is applied.

In the transmit mode the usable output power is equal to the i-f injection level minus the mixer's conversion loss, L_c . For operation at the 1-dB compression point, this relationship can be expressed as

$$P_{out} = P_{LO(dBm)} - 5 \text{ dB} - L_c$$

This formula implies that the system power output continues to improve for increases in LO injection. This would be true were it not for the fact that conversion loss does not remain constant for all levels of LO injection. Fig. 11 demonstrates the variation in conversion loss, as well as optimum power output, as a function of LO injection level for a typical microwave diode balanced mixer. Note that optimum conversion loss occurs at an LO injection level of around +8 dBm.

Beyond this point, though conversion efficiency degrades, output power con-

tinues to increase. Indeed, within the power restriction of the HP 5082-2817 diodes, it is possible to obtain about 16 mW of clean output power. However, recall that the Simple Sideband System uses the same mixer and local-oscillator chain for both transmit and receive. Any

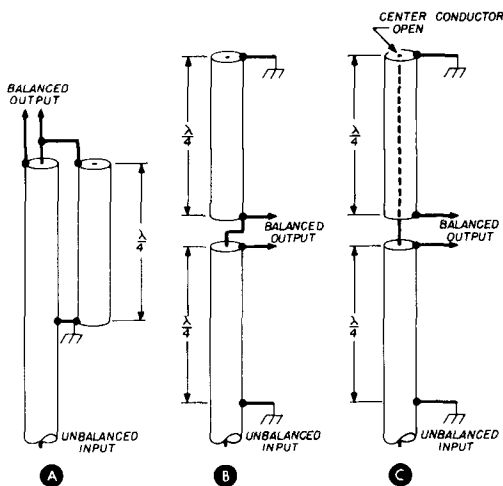


fig. 7. Derivation of the coaxial balun used in the balanced mixer of fig. 5. Conventional balun in (A) has been straightened out in (B) to improve symmetry. Balun in (C) is identical, electrically, to the circuit in (B) but symmetry has been further improved.

decrease in mixer conversion efficiency will degrade receive noise figure accordingly. Beyond +10 dBm of LO injection, transmitter power is gained only at the expense of receiver noise figure.

The break-even point occurs at an LO power of +16 dBm. Beyond this level, each dB of increase in transmitter output results in one dB of receiver degradation. Thus, the Simple Sideband System optimizes at 6-dB conversion loss, 40-mW of LO injection, 12.6-mW of i-f drive and 3-mW of output power. The tradeoff involved in determining this optimum performance point is illustrated by the $P_{out} + L_c$ curve of fig. 11. The sum of conversion loss and output power is used as a figure of merit for communications between two identical systems. This sum

represents the output power available at the i-f port of a receive mixer which is driven by an identical transmit mixer operating at the 1-dB compression point. Note that as local-oscillator power is increased, a knee is reached beyond which additional power will serve no

passband. The actual transmitter output power, after the filter, will thus be

$$+7.8\text{dBm} - 3.0\text{dB} - 0.5\text{dB} = 4.3\text{dBm} = 2.7\text{mW}$$

Note that the filter will also eliminate receiver image noise, as well as blocking those out-of-band signals which might

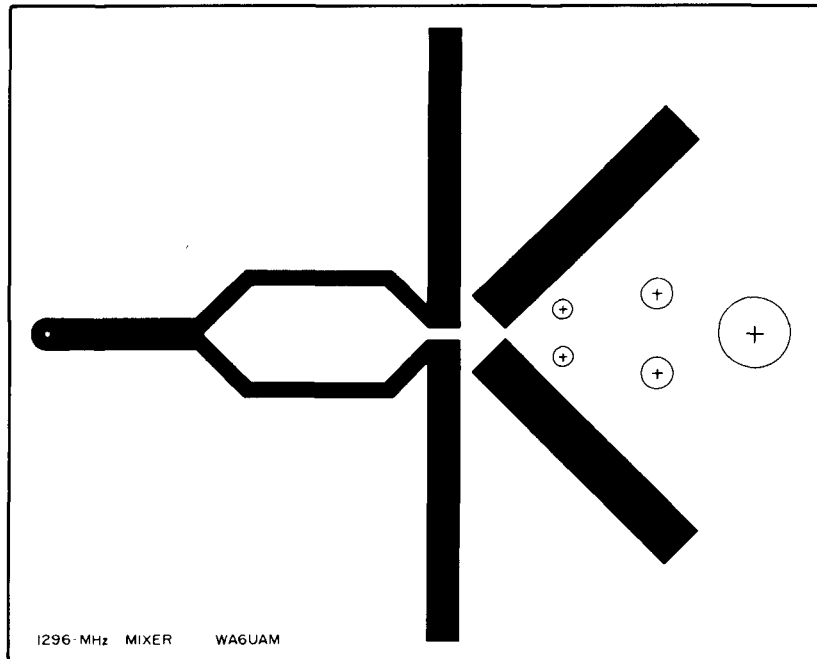


fig. 8. Full-size printed-circuit layout for the balanced mixer of fig. 5. Material is 1/16" (1.5 mm) thick, double-sided Teflon-fiberglass circuit board.

useful purpose. Thus it is desirable to operate the system at the knee of this curve, which I call the *transceive figure of merit* curve.

bandpass filter

With 40 mW of LO injection in the transmit mode, and using the balanced mixer described above, a mixer output power of 6 mW is indicated on a power meter. It must be remembered that this signal represents both the desired output signal (LO + i-f) and the image (LO - i-f). A bandpass filter with sufficient skirt selectivity to reject the image will also have about 0.5-dB insertion loss in the

otherwise enter the mixer and cause cross-modulation distortion and interference. Of course, filter insertion loss must be added to mixer conversion loss and i-f noise figure when determining receive converter performance.

Physically, the bandpass filter can be a half- or quarter-wavelength coaxial resonator, or a trough-line cavity such as has been used in previous 1296-MHz receiving converters. Coupling in and out can be accomplished by means of links, loops, taps, capacitors or even the coaxial matching scheme used by K6UQH in his latest converters.⁵ In the interest of avoiding multiple responses, a similar

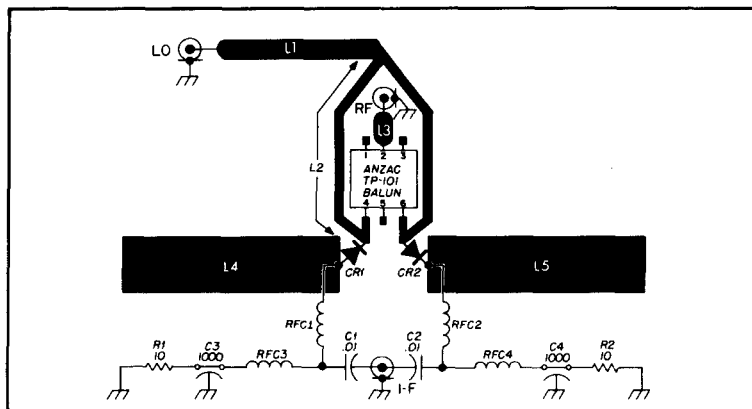
filter at the output of the LO chain may prove desirable.

local-oscillator chain

The key criteria here are stability and spectral purity. For maximum stability it is advisable to invest in the best possible low-temperature coefficient crystal you can afford.* An additional ten dollars invested in a quality crystal can do much

provide minimum crystal feedback consistent with ready starting, and should of course be buffered.

Spectral considerations dictate very careful selection of the multiplication scheme used to reach the desired injection frequency. High-order multiplication in a single stage is out, as the resultant harmonic comb requires extensive filtering. W6FZJ, whose success on both



C1,C2	0.01 μ F disc ceramic	L4,L5	25-ohm micro-stripline, 0.30" (7.5 mm) wide, 1.14" (29 mm) long
C3,C4	1000-pF feedthrough	R1,R2	10-ohm, 1/4-watt carbon composition resistor
CR1, CR2	hot-carrier diodes (H-P 5082-2818)	RFC1, RFC2	2" (51 mm) no. 32 wire, close wound on 0.050" (1.5 mm) form or ferrite beads on leads of C1 and C2
L1,L3	50-ohm micro-stripline, 0.10" (2.5 mm) wide, any length	RFC3, RFC4	22 μ H
L2	75-ohm micro-stripline, 0.045" (1.0 mm) wide, 1.24" (31.5 mm) long (along center)		

fig. 3. Improved 1296-MHz balanced mixer uses commercial balun and double-clad epoxy circuit board. Full-size printed-circuit layout is provided in fig. 10.

to alleviate slight frequency drift which (when multiplied into the microwave region) can make ssb transmission and reception a running battle between your right hand and the tuning dial. The oscillator circuit should be designed to

*The Croven C180DBX-00 5th-overtone crystal is highly recommended. This series-resonant, HC-18/U crystal has a calibration tolerance of ± 10 ppm and temperature tolerance of ± 5 ppm from 15° to 35° C. \$12.00 in single quantities, or \$8.00 each for two or more crystals of the same frequency. Write to Croven Ltd., 500 Beech Street, Whitby, Ontario, Canada.

432-MHz EME and 2304-MHz tropo-scatter speaks well for his expertise in such matters, recommends doubling in every stage of any LO multiplier. If the starting frequency (the crystal-oscillator stage) is in the 70- to 125-MHz region, any spurious responses present after repeated doubling will be sufficiently separated in frequency to be easily filtered.

The balanced mixer used at WA6UAM receives 40-mW of LO injection. Obtaining this power level at 1200 MHz is relatively simple with today's low-cost

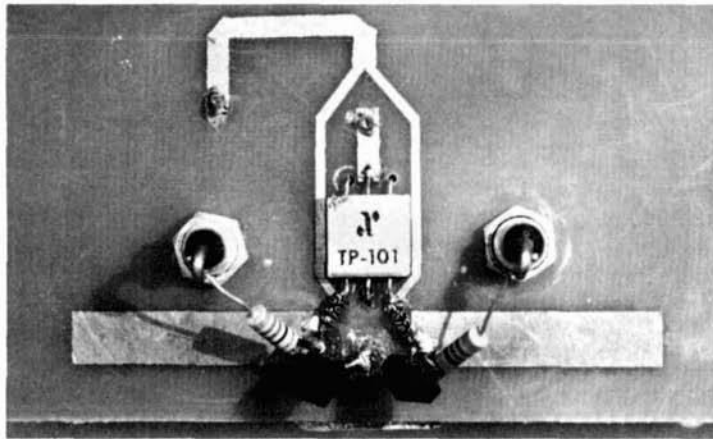
uhf transistors. Stability will be enhanced by designing the second stage of the LO chain (buffer) for 100-mW output, and running succeeding doublers at or close to unity gain. Resistive 3-dB pads between stages will alleviate excessive drive and provide impedance matching.

When designing active frequency doublers, recall that second-harmonic generation is enhanced by a collector

the Simple Sideband System could be duplicated without the use of specialized test equipment. In the case of the LO chain, alignment can be accomplished and injection measured by merely monitoring the diode current of the balanced mixer.

local-oscillator circuits

The primary advantage of the modular system which I use is the ease with which



Construction of the doubly-balanced mixer using the Anzac TP-101 balun.

conduction angle of 180° . This condition is easily met by grounded-base, zero-bias operation. Fortunately, this is also probably the simplest frequency-multiplier circuit to drive and adjust. Incidentally, it was stated earlier that the components of

substitutions can be made, and performance of various circuits compared. The Simple Sideband System has already worn three different LO chains; no doubt others will be attempted in the future. The block diagram of fig. 12 shows what

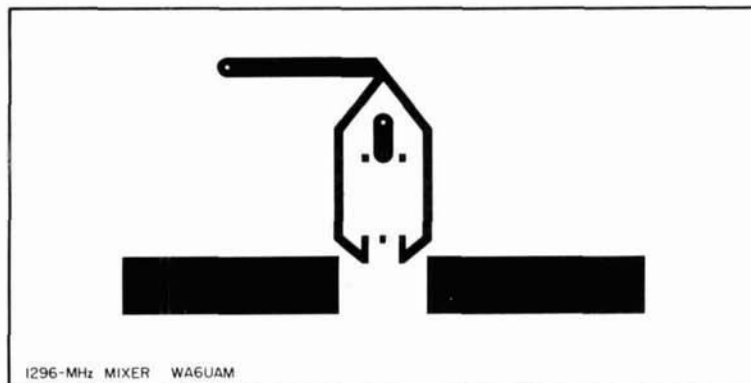


fig. 10. Full-size printed-circuit layout for the improved 1296-MHz mixer. Board is $1/16''$ (1.5 mm) thick, double-clad 1-ounce copper G-10 fiberglass-epoxy circuit board.

has thus far proved the most workable compromise between stability and spectral purity on the one hand, and low cost, simplicity and ease of alignment on the other.

However, this LO chain violates a number of the *ideal* design principles outlined above. For example, the purist will want the last multiplier to double rather than triple, and would probably use an active device rather than a diode. The decision to go with a diode multiplier was based primarily on the cost of 1.3 GHz transistors at the required power level. Tripling was used here because of the greater ease of generating a high level of power at 422 MHz, as compared to 630 MHz. The common-base configuration, although simple, was abandoned in the lower-level stages in deference to the greater gains available from common-emitter circuits.

The active multipliers are all operated at a low power level as a concession to stability, with the two 422-MHz power amplifiers providing plenty of drive to the tripler. Considering the low conversion efficiency of the diode tripler circuit, a high drive level is a must.

The low-level stages shown in **fig. 13** were designed by W6FZJ for use in a 432-MHz receiving converter, and published in his *432 Newsletter*. Except for the crystal frequency and the number of turns on two inductors, his original circuit is unchanged when used in the 1296-MHz LO chain. Copies of the original circuit have been successfully built by a number of San Francisco area 432-MHz enthusiasts, and spectrum analyzer tests prove the W6FZJ design to be superior to any of my own attempts to date.

Since the tripler which raises the injection frequency to 1267 MHz exhibits about 10-dB conversion loss,* a half-watt of drive is necessary at 422 MHz to achieve the desired 40-mW of LO injection.

*Greater tripler efficiency is possible through the use of an idler circuit, a technique avoided by the author in the interest of simplicity.

This is accomplished by applying the 10-mW output of the low-level LO module to two stages of power amplification, operating at 10- and 7-dB of gain, respectively. The 2N3866s were selected because of their low cost and ready

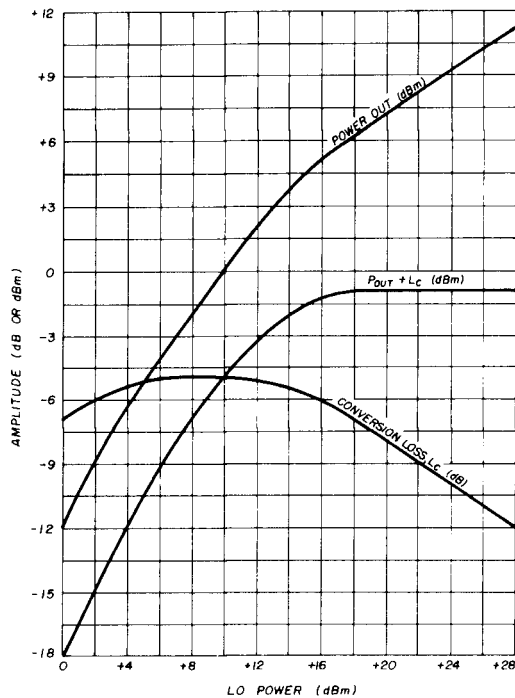


fig. 11. Balanced mixer conversion loss, output power at the 1-dB compression point and their sum (transceive figure of merit) for a typical microwave balanced mixer, all as a function of local-oscillator injection level. The transceive figure of merit represents the i-f output power available from a receive mixer driven directly by the output of an identical transmit mixer.

availability from a number of mail-order surplus component dealers.

The power amplifier circuits are shown in **fig. 14**. Care should be taken to closely duplicate the input and output tank circuits unless a spectrum analyzer is available, as adequate spectral purity occurs when these particular circuits are tuned for maximum indicated output.

The slab inductors in the collector

circuits provide high output Q, and the pi-network filter (C8, L3, C9) feeding the second stage tends to suppress any harmonics generated by the first. The first 2N3866 is biased to class AB for increased gain; the second stage is run in class C

tributed constants. Design of the poles was accomplished as follows: It was desired that the filter resonate with the tuning capacitors (C2, C3) at midrange, or 1.5 pF. Assuming an additional 0.5 pF of stray and coupling capacitance, the

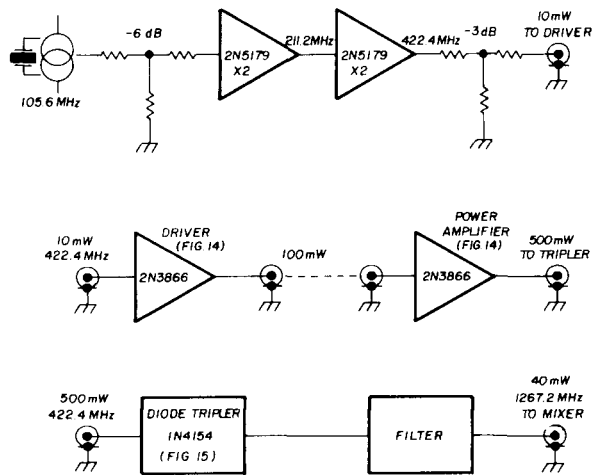


fig. 12. Block diagram of the local-oscillator chain used with the Simple Sideband System for 1296 MHz. Each of the individual sections is built into a small module; these are connected together with miniature 50-ohm coaxial cable.

for improved collector efficiency. If the stages are built separately, each can be tuned for maximum power into a 50-ohm load.

The diode tripler and filter assembly was first built in the popular trough-line configuration. Later, an interdigital filter was attempted. In both cases performance was satisfactory. However, the construction process required extensive metalworking. As many amateurs avoid projects which involve bending sheet metal or cutting brass tubing, I decided to reduce the tripler/filter to a PC board. The result is shown schematically in fig. 15.

Rf energy from the 422-MHz power amplifier, fig. 15, is applied to CR1, a GE 1N4154 high-speed switching diode through an L-network (L1, C1) similar to that used by K6UQH in his trough-line multiplier/mixers. The harmonic comb developed at the output of the diode is applied to a two-pole resonator (L2, C2, L3, C3) which blocks all but the 1267-MHz component.

The filter combines lumped and dis-

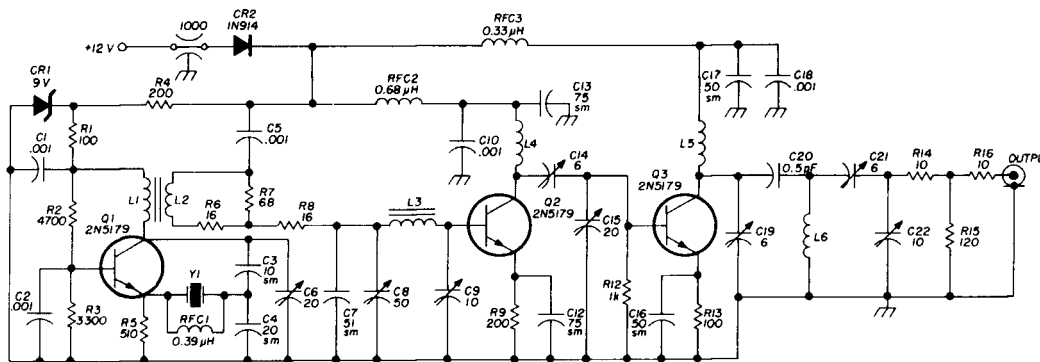
tributed constants. Design of the poles was accomplished as follows: It was desired that the filter resonate with the tuning capacitors (C2, C3) at midrange, or 1.5 pF. Assuming an additional 0.5 pF of stray and coupling capacitance, the

$$\theta = \arctan \frac{X}{Z_0}$$

where θ represents wavelength of the stripline in *degrees*. To convert to fractions of a wavelength, divide θ by 360° . Thus, for 62.8-ohm inductive reactance with an arbitrarily selected stripline characteristic impedance of 25 ohms,

$$\theta \arctan \frac{62.8}{25} = 68.3^\circ = 0.19 \text{ wavelength}$$

A 25-ohm micro-stripline on 1/16-inch (1.5-mm) G10 PC board is 0.3-inch (7.5-mm) wide. The 0.19 wavelength at 1267 MHz (correcting both for velocity factor and width-to-height ratio) is 0.865 inch (22 mm).



C6,C15	20-pF trimmer (JFD DJV 300)	L3	12 turns no. 28 on Micrometals T-25-13 toroid core
C8	50-pF trimmer (JFD DVJ 305)	L4	6 turns no. 24, close wound on 0.1" (2.5 mm) diameter form
C9,C22	10-pF trimmer (JFD DVJ 302)	L5,L6	1½ turn no. 22, 1/8" (3 mm) diameter, 1/4" (6 mm) long
C14,C19,	6-pF trimmer (JFD DVJ 301)	RFC1	0.39 μH miniature inductor (Nytronics Deci-Ductor)
C21		Y1	5th overtone crystal, series resonant, HC-18/U holder, 105.600 MHz for 28.8-MHz i-f
C20	0.5-pF ceramic (two 1-pF capacitors in series)		
L1	10 turns no. 28 on Micrometals T-25-13 toroid core		
L2	4 turns no. 28 on cold end of L1		

fig. 13. Crystal-controlled local oscillator circuit, based on a design by Joe Reisert, W6FZJ.

Matching the resonators to the relatively low impedances of the diode and the output transmission line can be accomplished by tapping up on the micro-striplines the required distance above ground. Although formulas exist for approximating the required tap position, matching in the circuit shown was determined empirically.

An important (and often neglected) consideration in diode multipliers is bias current. Resistor R1 in fig. 15 enables the diode current to be varied over a wide range. Remember that diode current will affect conduction angle, which should be 120° to maximize third-harmonic generation.

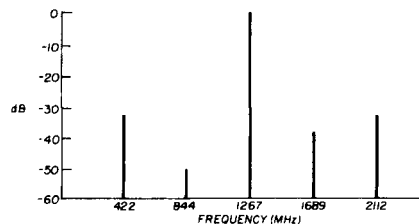
transmit i-f attenuator

Proper operation of balanced mixers requires that each port be terminated in its characteristic impedance (usually 50 ohms). Most methods used to sample a low-level ssb signal from a high-frequency transmitter would result in a horrendous impedance mismatch at the mixer's i-f port. As 12.6-mW of sideband injection is

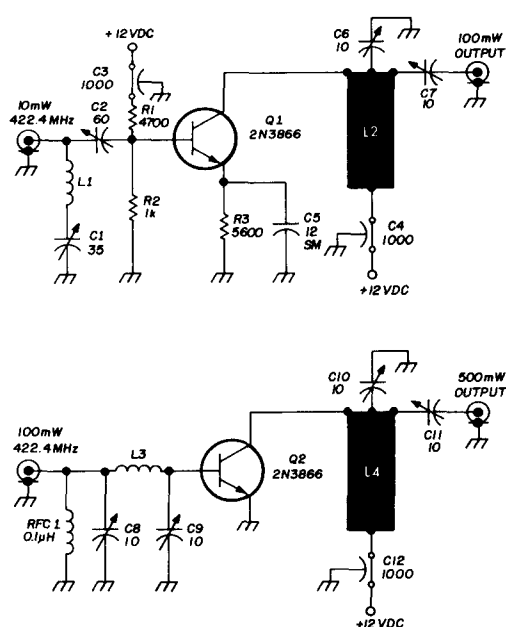
desired, one recommended method of assuring i-f impedance matching is to run about one watt of ssb into a 20-dB resistive pi- or T-pad at the mixer's i-f port. Since the attenuation pad will provide the proper termination to the mixer, the method of coupling out of the station transmitter has no effect on the mixer's operation.

i-f amplifier

The noise figure of a receiving converter without pre-conversion gain is the



Spectrum output of the 1267.2-MHz local-oscillator chain when tuned for maximum mixer diode current. The 422- and 2112-MHz spurs are down 32 dB, the 1689-MHz spur is down 38 dB and the 844-MHz spur is down 50 dB.



- C1 3-35 pF trimmer
- C2 8-60 pF trimmer
- C6-C11 10-pF concentric piston trimmers
- L1 2 turns no. 18, wound on 1/4" (6 mm) mandrel, 1/8" (3 mm) long
- L2,L4 brass strip, 0.5" (12.5 mm) wide, 1.5" (38 mm) long, mounted 1/8" (3 mm) above ground plane
- L3 2 turns 1/8" (3 mm) wide brass strip, 0.1" (2.5 mm) diameter, 0.5" (12.5 mm) long

fig. 14. The two 422-MHz power amplifiers provide one-half watt output. The first stage, Q1, provides 10-dB gain, while the second stage, Q2, provides 7-dB gain.

sum of the feedline and TR relay losses (if any), input filter insertion loss, mixer conversion loss and i-f amplifier noise figure. For the system presented here the noise contribution of these stages prior to the i-f amplifier is less than 6.5 dB. Feeding the output of the mixer into a high-frequency ssb receiver would, however, result in a system noise figure approaching 20 dB. Many amateurs are surprised to learn that the noise figure of even a high quality commercial communications receiver is seldom below 10 or 15 dB. In the high-frequency spectrum the

level of man-made and atmospheric noise exceeds that of receiver noise by several orders of magnitude, so noise figure, per se, is not usually a significant consideration in high-frequency receiver design. Of course, such is not the case in the microwave region. To achieve a reasonable noise figure in any uhf converter, a low-noise i-f amplifier must be used to mask the following receiver noise.

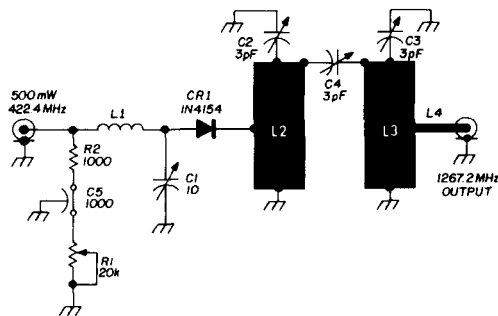
Numerous circuits have been described in the past which will yield reasonable gain at 28 MHz with a noise figure under 1 dB. WB6NMT, known for his pioneering efforts in 220-MHz EME, uses neutralized fets, while W6FZJ favors a dual-gate mosfet following the mixer. An appealing circuit by W9PRZ uses a dual jfet in a cascode configuration and is described in a Siliconix applications note.⁶

system performance calculations

It remains to be shown that the simple microwave ssb system described here lends itself to satisfactory communications over reasonable distances (my experience shows the following calculations to be somewhat on the conservative side). The overall noise figure of the receive system is the sum of filter loss (0.5 dB), mixer conversion loss (6 dB) and i-f noise figure (1 dB). Thus, a receive noise figure of 7.5 dB is assumed. With the 2.1-kHz i-f bandpass of a good sideband receiver, the graph of fig. 16 indicates the receive sensitivity to be about -133 dBm.

Good intelligibility on single sideband requires a 6-dB signal-to-noise ratio. Allowing a 10-dB signal-to-noise ratio for good measure, the receiver requires a -123 dB signal. At +4.3 dBm of transmitter power, given 10-dB of antenna gain at each end, and allowing 5-dB of additional loss for antenna aiming errors, the permissible path attenuation between transmit and receive stations is 142.3 dB. The free-space loss formula is⁷

$$L_{FS} = 36.6 + 20 \log_{10} f_{MHz} + 20 \log_{10} d_{miles}$$



- C1 1-10 pF concentric piston trimmer
- C2,C3 0.3-3 pF concentric piston trimmer
- C4
- CR1 1N4154 high-speed switching diode
- L1 2 turns no. 20, 0.1" (2.5 mm) diameter, 0.25" (6 mm) long
- L2 micro-stripline, 0.3" (7.5 mm) wide, 0.865" (22 mm) long, grounded at bottom, tapped 0.20" (5 mm) from ground end
- L3 Same as L2 but tapped 0.25" (6 mm) from ground end
- L4 50-ohm micro-stripline, 0.1" (2.5 mm) wide, any length
- R1 20k, 10-turn trimpot

fig. 15. The 422- to 1267-MHz diode tripler is built on 1/16" (1.5 mm) double-clad, glass epoxy circuit board.

which can be rewritten to solve for distance at 1296 MHz

$$\log_{10} d_{\text{miles}} = (L_{FS} - 98.85)/20$$

Given 142.3 dB of permissible free-space loss, an HP-35 calculator yields

$$d(\text{maximum}) = 161 \text{ miles (259 km)}$$

Amateurs who have operated simple equipment in the 23-cm band find it difficult to believe that a 161-mile path is possible with only 3-mW of output power. After all, they reason, that's greater than the APX-6's range, and those have 3-watts output. It should be remembered, however, that directly-modulated oscillators tend to be extremely unstable, and that reception of their emissions requires a wideband receiver. The APX-6 i-f strip is about 5-MHz wide and bandwidth has a serious

impact on receiver sensitivity. Fig. 16 indicates a deterioration in receiver sensitivity of 10 dBm for each tenfold increase in i-f bandwidth. This more than offsets any path increase afforded by the greater power of the APX-6 or similar equipment.

direct conversion

If simplicity is the goal, there is practically no limit to the evolutionary process. This last step occurred almost by accident, while developing test equipment for 1296 MHz. A weak-signal source was built for 1296.00 MHz to permit calibration and receiver testing. Since a few milliwatts of stable power was desired, I used the same techniques as I used for building the converter local oscillator. Later a key was added, and the signal source used for on-the-air CW contacts over a limited range.

Eventually, a means for monitoring transmit quality from the ssb converter was desired. An antenna and tuned cavity feeding a diode detector and audio amplifier produced the characteristic ssb "Donald Duck" squawk in a speaker. However, for high-quality signal monitoring a ssb converter was required. By using the 1296.0-MHz signal source developed earlier, driving a balanced mixer with its i-f port feeding an audio amplifier, a direct-conversion receiver was built. If the

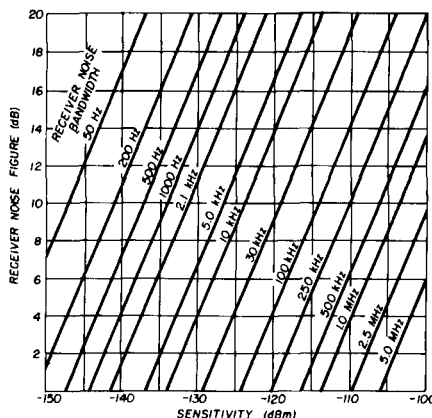


fig. 16. Receiver noise figure vs sensitivity for various receiver bandwidths.

transmitter (suppressed) carrier frequency and the receiver LO frequency are the same, the difference frequency at the i-f port of the mixer will be audio sidebands. No input filter is required because the image is merely the other sideband – and at 1296 MHz QRM is hardly a problem yet.

If a direct-conversion double-sideband receiver will work, there's no reason to expect otherwise for direct double-sideband generation. Indeed, when audio was applied at the i-f of the balanced mixer, sidebands around 1296.0 MHz appeared at the rf port. Though extremely low power, the signals could be copied in a sensitive receiver at a range of about a mile.

The resulting direct-conversion double-sideband transceiver is shown in fig. 17. Though primarily a lab accessory and demonstration rig, it provides reliable communications over moderate distances. It is simple in the extreme, and suggests the possibility of microwave double-sideband walkie-talkies. The concept may even work at greater power levels than the few microwatts attempted to date. Certainly there's no more basic a way to produce a microwave sideband station.

conclusion

The primary advantage of ssb in the microwave spectrum, as anywhere else, is in the significant increase in receiver sensitivity resulting from the narrow bandwidth which is required. The only limiting factor is stability, a criterion which can be readily achieved by judicious application of good engineering practice to the design and construction of local-oscillator chains. The design trade-offs presented here make microwave ssb feasible over considerable distances, while requiring a minimum of specialized skills, equipment or technique.

acknowledgements

I wish to express my deep appreciation to Don Farwell, WA6GYD, for giving me my first contact on 1296 MHz and kindling the fire, to Frank Pacier,

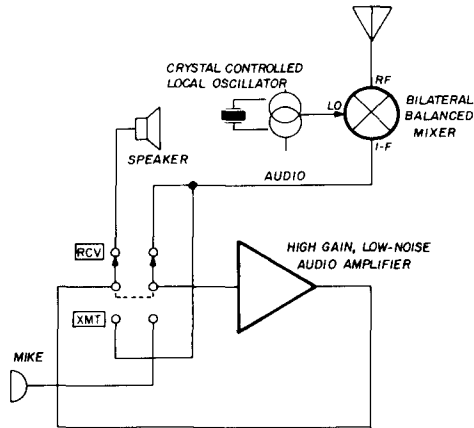


fig. 17. Simple direct-conversion double-sideband transceiver for 1296 MHz.

W6VMY, for the countless hours we spent together working on APX-6s in the early days, to Bill Troetschel, K6UQH, who shamed me into giving up modulated oscillators by his snide remarks at uhf conferences, to Bob Ney, WB6LLD, who gave me my first microwave transistors, and to Joe Reisert, W6FZJ, who taught me how to use them. And most important, a word of thanks to my wife, Suk, who for the past three years has shown more understanding of the hours spent on the 1296 project than I had any right to expect.

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