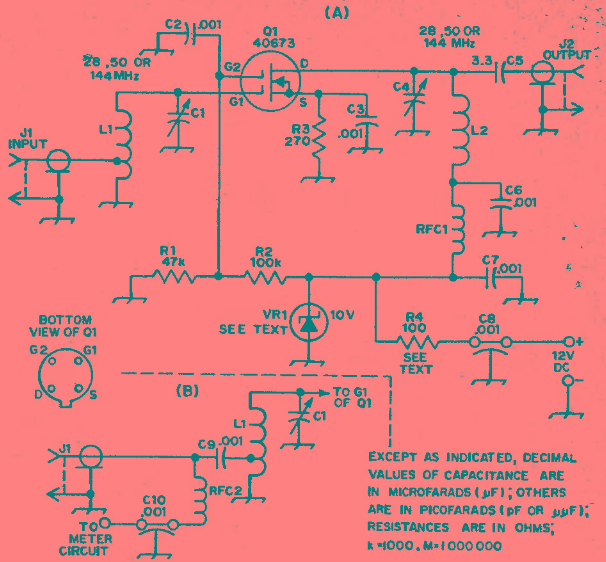


Fig. 1 — Schematic diagram for the preamplifier. Part designations not listed below are for pc board placement purposes. Alternative input circuit for use with microwave diode mixer is shown at B.

C1, C4 — See Table I.
C2, C3, C5, C6, C7, C9 — Disk ceramic.
C8 — .001 feedthrough capacitor.
J1, J2 — Coaxial connectors. Phono-type, BNC or SO-239 acceptable.
L1, L2 — See Table I.
R4 — 3 turns No. 28 enam. on ferrite bead. A 220-ohm, 1/2-watt resistor may be substituted.
RFC2 — 33 μH, iron-core inductor. Millen J300-33 or J. W. Miller 70F335A1.

is to cover the copper with masking tape, transfer the pattern with carbon paper, then cut away the tape to expose the part to be etched. On small, simple boards the masking-tape method is hard to beat.

The pc board may be mounted in almost any small enclosure. Construction is not tricky or difficult. It should take only a few minutes to complete the unit after the board is prepared. The board is fastened in the enclosure by means of one metal standoff post and a No. 4 screw and nut. Input and output connectors are not critical; phono-type jacks may be used in the interest of low cost.

Adjustment is so easy that it almost needs no description. After connecting the amplifier to a receiver, simply tune the input (C1) and the output (C4) for maximum indication on a weak signal. One possible area of concern might be that the toroids used in the ten- and six-meter versions are not always uniform in permeability, as purchased from various suppliers. However, it is an easy matter to add capacitance or remove a turn as required to make the circuits resonate at the correct frequency.

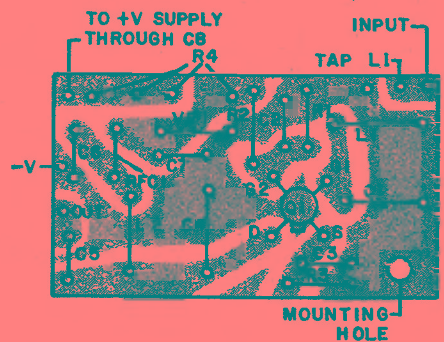


Fig. 2 — Full-scale layout and parts placement guide for the pc board. Foil side shown.

Table I

	28 MHz	*50 MHz*	*144 MHz*
L1	17 turns No. 28 enam. on Amidon T-50-6 core. Tap at 6 turns from ground end	12 turns No. 26 enam. on Amidon T-37-10 core. Tap at 5 turns from ground end.	5 turns No. 20 tinned 1/2-inch ID X 1/2-inch long. Tap at 2 turns from ground end.
L2	Same as L1, without tap.	Same as L1, without tep.	4 turns No. 20 tinned like L1, without tap.
C1, C4,	15 to 60-pF ceramic trimmar. Erie 538-002F.	1.8- to 16.7-pF air veriable. E. F. Johnson 189-506-005.	1.5- to 11.6-pF air variable. E. F. Johnson 189-504-005.

Fig. 1 — Completed six- and two-meter converters (left end center) with power supply.

CONVERTERS FOR 50 AND 144 MHz

The converters described here are designed by the Rochester VHF Group and details are presented by W2DUC and K2YCO.

Because of the nature of the project, a universal circuit-board design is used. One circuit board serves for either band, with only slight modification. Other specific design goals were:

1) Low noise figure, less than 3 dB.

2) State-of-the-art freedom from cross modulation.

3) Sufficient gain to override the front-end noise of most receivers.

4) Double-tuned bandpass interstage and output circuits to achieve a flat response over a two-MHz portion of either band.

5) Filtering of the local oscillator chain in the two-meter model to reduce spurious responses.

6) Small size and low power consumption.

7) Freedom from accidental mistuning during the life of the converter.

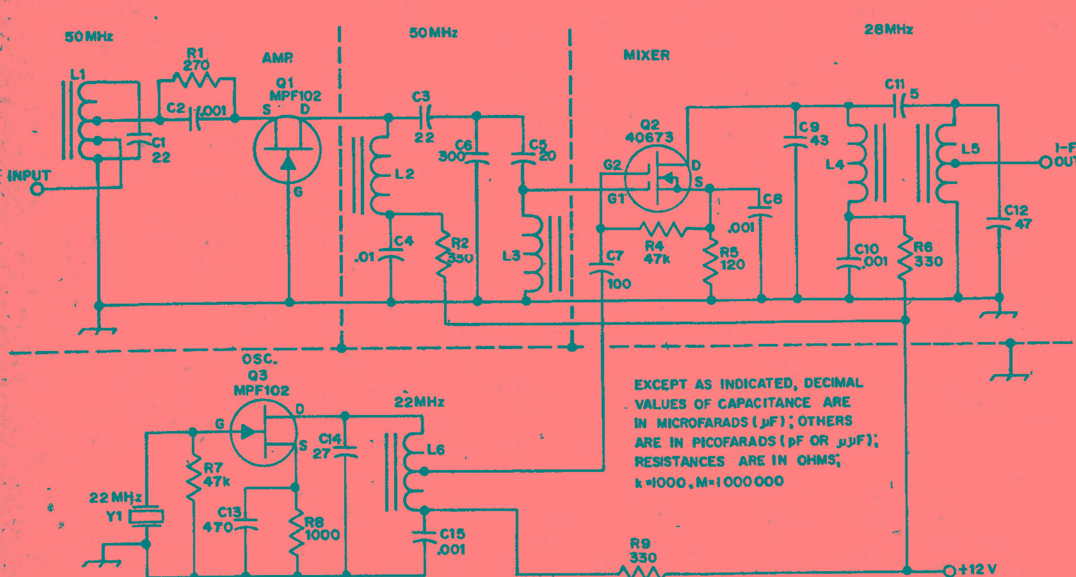


Fig. 2 — Schematic diagram of the six-meter converter. All resistors are 1/4-watt composition. C2, C8, C10 and C15 are .001 μF disk ceramic. C4 is .01-μF disk ceramic. All other capacitors are dipped mica.

L1-L6, incl. — All No. 28 enam. wire wound on Amidon T-30-6 cores as follows: L1, 14 turns tapped at 4 turns and 6 turns; L2, 13 turns; L3, 12 turns; L4, 18 turns; L5, 18 turns tapped at 4 turns from cold end; L6, 26 turns tapped at 6 turns from hot end.

Y1 — 22-MHz crystals. International Crystal Mfg. Co. type EX.

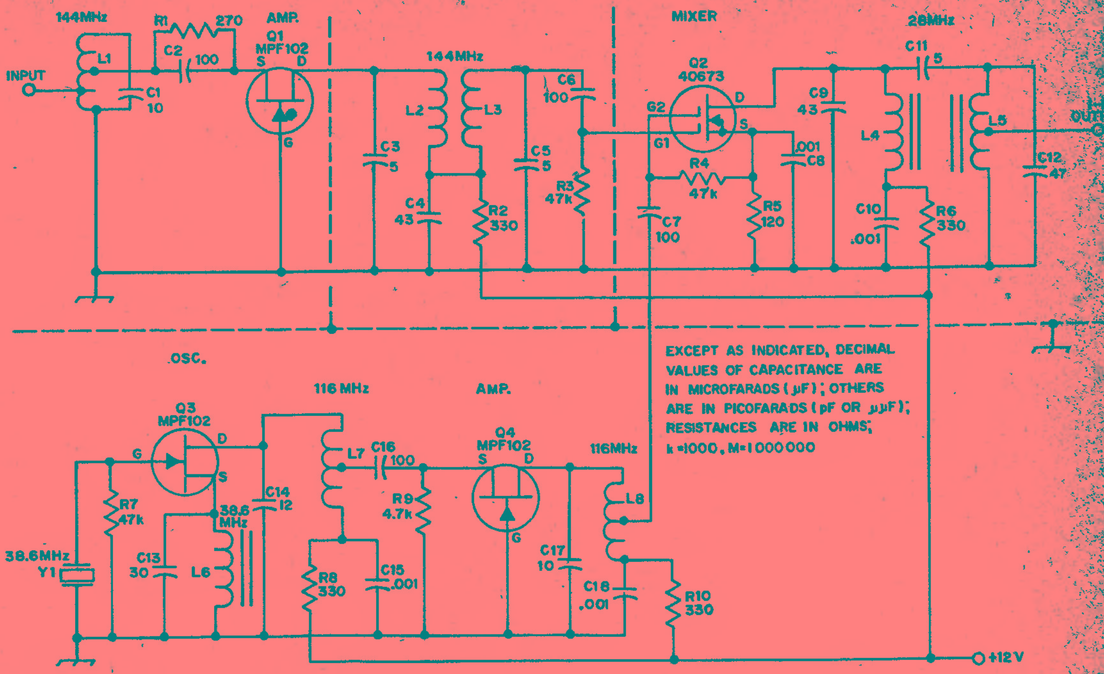


Fig. 3 — Schematic diagram of the two-meter converter. All resistors are 1/4-watt composition. C8, C10, C15 and C18 are .001-μF disk ceramic. All other capacitors are dipped mica units.

L1, L2, L3, L7, L8 — All No. 20 enam. wire formed by using the threads of a 1/4-20 bolt as a guide. L1, 5 turns tapped at 1-3/4 turns and 3/4 turn from cold end; L2, 5 turns; L3, 4 turns; L7, and L8, 5 turns tapped at 2 turns from hot end.

L4 — 18 turns No. 28 enam. wound on Amidon T-30-6 core.

L5 — 18 turns like L4, tapped at 4 turns from cold end.

L6 — 0.68 μH miniature inductor. Delevan 1025 serias or J. W. Miller 9230-16.

Y1 — 38.666-MHz crystal. International Crystal Mfg. Co. type EX.

Other points considered were such things as freedom from the necessity of neutralization and the use of moderately priced transistors.

Several breadboard models were constructed and tested as the design evolved. Fig. 1 shows two completed converters and a power supply.

Circuit Design

A schematic diagram for the six-meter converter is shown in Fig. 2, and for the two-meter model in Fig. 3. The configuration of the rf and mixer portions of the circuit are virtually identical for six and two meters, with the values of the frequency-determining components being scaled appropriately. The major difference between the two converters is a change in the local oscillator chain. A minor change in the method of interstage coupling was necessary to prevent stray-capacitance effects from making the alignment critical on the six-meter converter.

All inductors in the six-meter model and the two-meter output circuit are wound on Amidon T-30-6 toroid cores. The tuned circuits are aligned by spreading or compressing the turns around the toroid core. After alignment the coils are glued in place with Silastic compound (sold as bathtub caulk).

The rf amplifier, Q1, is used in a grounded-gate configuration. The input circuit is tapped to provide a proper match between the antenna and source of the FET while maintaining a reasonable *Q*. The six-meter interstage coupling network consists of C3, C5, L2, and L3. Band-pass coupling is controlled by the capacitive T network of C3 and C5 in ratio with C6. A 40673 dual-gate MOSFET is used in the mixer circuit (Q2). Gate 1 receives the signal, while gate 2 has the local oscillator injection voltage applied to it through C7. A slight amount of positive bias is applied to gate 2 through R2. A top-coupled configuration, using toroid inductors, serves as the 28-MHz output circuit of both converters.

The oscillator circuit in the six-meter model is straightforward, relying on the drain-to-gate capacitance of the FET for feedback. A tap at four turns from the hot end of the toroid winding provides the injection to the mixer through capacitor C7. In the two-meter converter, Fig. 3, the rf stage is identical to the six-meter version except for

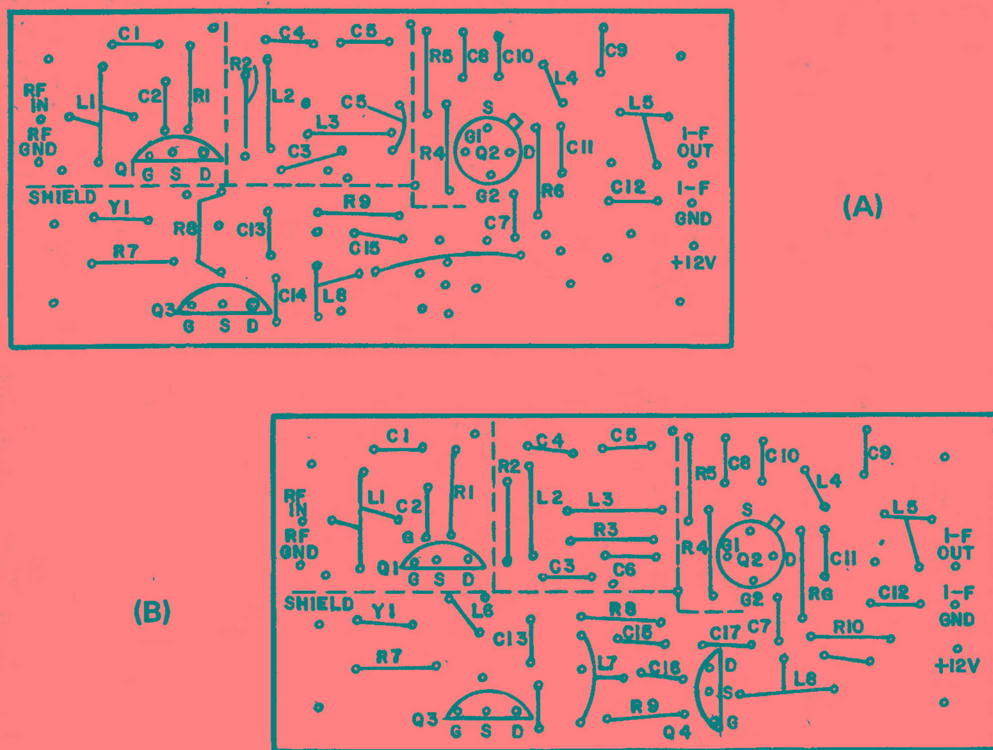


Fig. 4 — Parts-placement guide for the six-meter converter, A, and the two-meter converter, B. View is from the foil side of the board. Dashed lines show the location of shields that are soldered to short pieces of wire which project through holes in the pc board. The shields may be fabricated from sheet brass or copper, or scraps of copper-clad board material.

the tuning networks. L1, L2, and L3 are air wound, self-supporting, and are formed initially by winding wire around the threads of a 1/4-20 bolt. The turns of L1 are spread to permit adding taps prior to mounting on the board. The degree of interstage coupling in the two-meter model is controlled by the positions of L2 and L3. Since they are mounted at right angles, the coupling is very light. By changing the angle between these two coils, the passband may be optimized.

In the two-meter oscillator stage, Q3 is changed to an oscillator/tripler by replacing the source bias resistor with L6. Replace bypass capacitor, C13,

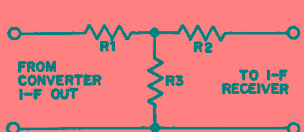


Fig. 5 — An i-f attenuator may be necessary if the receiver following the converter is exceptionally hot. Values for 6 dB: R1, R2 — 18 ohms; R3 — 68 ohms. For 10 dB: R1, R2 — 27 ohms; R3 — 39 ohms.

with a 30-pF value to resonate L6 near the crystal frequency. Source-to-gate capacitance provides the feedback in this case. The drain tank is modified to provide output at the third harmonic, thus eliminating the need for a separate tripler stage. Q4 is used as an isolation amplifier running at very low current level (as controlled by R9) to provide attenuation of the adjacent harmonics. This stage is not needed for amplification of the oscillator signal but without the additional filtering, severe "birdies" may result from nearby fm or TV stations. In both the six- and two-meter versions, a number of printed-circuit pads will be left over when construction is completed. These are the result of providing both bands on a common pc layout. For example, the isolation amplifier following the oscillator is not used on six meters. Therefore, this stage is bypassed by a jumper wire from L6 to C7. Five additional holes are located in the ground area along the centerline of the board and between rf and mixer stages. Component lead clippings are soldered into these holes to provide a mounting for the shield partitions, which are soldered to the wires where they extend through the board. Fig. 4 shows the parts layout for the six- and two-meter converters. Notice that one lead of

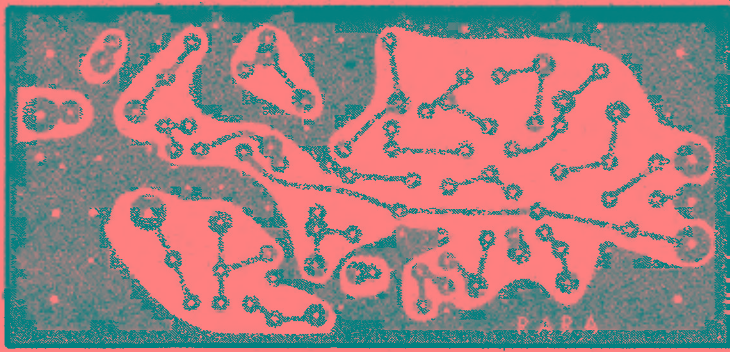


Fig. 6 — Scale-size layout for the pc board. The same pattern is used for either band. Foil side shown here.

Table I — Performance Specifications

Parameter	*6 Meters*	*2 Meters*
Noise figure, dB	1.8 – 2.3	2.0 – 2.4
Conversion gain, dB	22 – 28	17 – 24
Spurious responses, dB	–80*	–60*
	* Has a response at 6 MHz	* Responses at 107 & 181 MHz
Freq. response, ± 1 dB	49.8 – 51.5 MHz	143.9 – 146.4 MHz
Current at 12 V dc	12 – 18 mA	14 – 20 mA

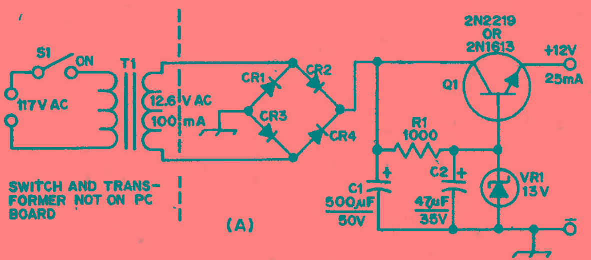


Fig. 7 — Schematic diagram and parts-placement guide for the power supply to the converters. The transformer is mounted external to the board. Pc board size is identical to the one used for the converters.

C3 must reach past the ground hole and connect to the foil. R3 is not used on the six-meter converter.

Alignment and Test

Perhaps the most difficult task in the project was the test and tune-up of the finished converter. A single test setup using a sweep generator, diode probe, and oscilloscope was a necessity to assure the flat response over the tuning range. Commercial attenuators were used to calibrate each converter by the substitution method.

Tuning of the air-wound rf circuit for two meters was accomplished by spreading or compressing the turns of the coils. After alignment, the windings were secured by a bead of Silastic compound along the oil to hold the turns in place. The noise figure of each converter was checked using the Monode noise-generator technique.[1] A final sensitivity check using a receiver (NC300) and a model 80 calibrated signal generator completed the checkout.

[1] Guentzler, "The Monode Noise Generator," *QST*, April 1967.

The transistors used in the rf stage were also subject to some variation in noise figure. When this occurred, an rf FET was carefully traded with an oscillator FET, since performance of the FET as an oscillator was always satisfactory.

The performance specification range for the converters is seen in Table I.

Small ceramic trimmers can be used in place of the fixed-value mica capacitors in the tuned circuits of these converters. The midrange of the trimmer should be approximately the value of the mica capacitors replaced. This procedure may simplify the tuning process of the converters where a sweep generator setup is not available. A little careful tweaking should give a reasonably flat response.

If trimmers are used, the rf input circuit should be tuned to the center of the desired response, 50.5 MHz as an example. This circuit tunes broadly and is not too critical. The rf interstage circuits should be stagger tuned, one at 50.0 MHz and the other at 51.0 MHz, as an example, the output i-f circuits can be tuned in a manner similar to the interstage circuits.

DOUBLE BALANCED MIXER

Advances in technology have, in recent years, provided the amateur builder with many new choices of hardware to use in the building of receivers, converters, or preamplifiers. The broadband double-balanced mixer package is a fine example of this type of progress, and as amateurs gain an understanding of the capabilities of this device, they are incorporating this type of mixer in many pieces of equipment, especially receiving mixers. The combined mixer/amplifier described here was presented originally in *QST* for March, 1975, by K1AGB,

Mixer Comparisons

Is a DBM really better than other types? What does it offer, and what are its disadvantages? To answer these questions, a look at more conventional "active" (voltages applied) mixing techniques and some of their problems is in order. The reader is referred to a recent article in *QST*[3] dealing with mixers. Briefly reiterated, common single-device active mixers with gain at vhf and uhf are beset with problems of noise, desensitization and small local-oscillator (LO) isolation from the r-f and i-f "ports." As mixers, most devices have noise figures in excess of those published for them as rf amplifiers and will not provide sufficient sensitivity for weak-signal work. To minimize noise, mixer-device current is generally maintained at a low level. This can reduce dynamic range, increasing overload potential, as defined in the terminology appendix. Gain contributions of rf amplifiers (used to establish a low system noise figure) further complicate the overload problem. LO-noise leakage to the rf and i-f ports adversely affects system performance. Mixer dynamic range can be limited by conversion of this noise to i-f, placing a lower limit on mixer system sensitivity. Generally 20 dB of mixer midband, inter-port, isolation is required, and most passive DBM can offer greater than 40 dB.

A commercially manufactured double-balanced diode mixer offers performance predictability, circuit simplicity and flexibility. Closely matched Schottky-barrier hot-carrier diodes, commonly used in most inexpensive mixers of this type, provide outstanding strong-signal mixer performance (up to about 0 dBm at the rf input port) and add little (0.5 dB or so) to the mixer noise figure. Essentially, diode conversion loss from rf to i-f, listed in Table I, represents most of the mixer

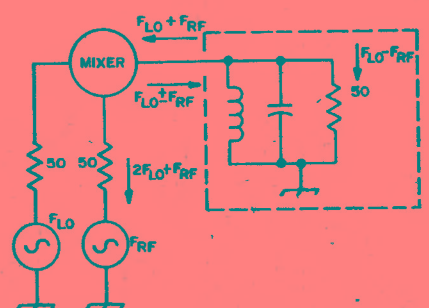


Fig. 1 — The i-f port of a double-balanced mixer is matched at $f_{LO} - f_{rf}$ and reactive at $f_{LO} + f_{rf}$. In this configuration conversion loss, rf compression and desensitization levels can vary ±3 dB while harmonic modulation and third-order IMD products can vary ±20 dB.

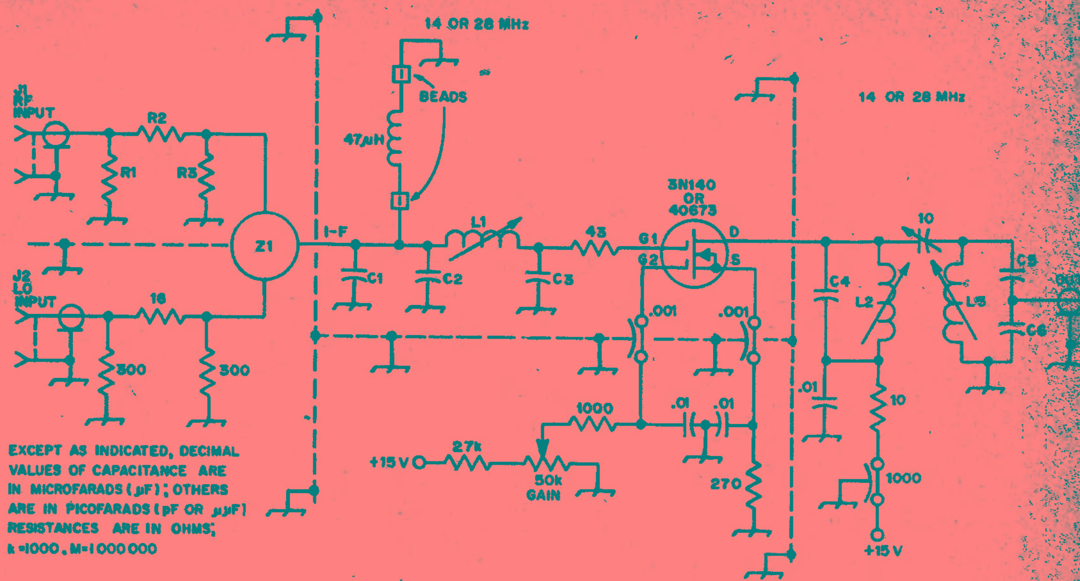


Fig. 2 — A schematic diagram for the double-balanced mixar and i-f post amplifier. Tha i-f can be either 14 or 28 MHz. Parts values are given in Tabla II.

contribution to system noise figure.† Midband isolation between the LO port and the rf and i-f ports of a DBM is typically > 35 dB — far greater than that achievable with conventional single-device active-mixing schemes. This isolation is particularly advantageous in dealing with low-level local-oscillator harmonic and noise content. Of course, selection of LO devices with low audio noise figures, and proper rf filtering in the LO output, will reduce problems from this source.

Often listed disadvantages of a diode DBM are (a) conversion loss, (b) LO power requirements, and (c) i-f-interface problems. The first two points are closely interrelated. Conversion loss necessitates some low-noise r-f amplification to establish a useful weak-signal system noise figure. Active mixers also have this requirement, as will be demonstrated later. Additional LO power is fairly easy to generate, filter, and measure. If we accept the fact that more LO power is necessary for the DBM than is used in conventional single-device active mixing circuits, we leave only two real obstacles to be overcome in the DBM, those of conversion loss and i-f output interfacing.

To minimize conversion loss in a DBM, the diodes are driven by the LO beyond their square-law region, producing an output spectrum which in general includes the terms:

- Fundamental frequencies fLO and $f$rf
- All of their harmonics
- The desired i-f output, fLO ± $f$rf
- All higher order products of nfLO ± mfrf where n and m are integers.‡

The DBM, by virtue of its symmetry and internal transformer balance, suppresses a large number of the harmonic modulation products. In the system described here, fLO is on the low side of $f$rf; therefore, numerically, the desired i-f output is $f$rf − fLO. Nonetheless, the term fLO ± $f$rf appears at the i-f-output port equal in amplitude to the desired i-f signal, and this unused energy must be effectively terminated to obtain no more than the specified mixer-conversion loss. This is not the image frequency, fLO − fi-f, which will be discussed later.

In any mixer design, all rf port signal components must be bypassed effectively for best conversion efficiency (minimum loss). Energy not "converted" by mixing action will reduce conversion gain in active systems, and increase conversion loss in passive systems such as the diode DBM. Rf bypassing also prevents spurious resonances and other undesired phenomena from affecting mixer performance. In this system, rf bypassing at the i-f-output port will be provided by the input capacitance of the i-f interface. The DBM is not a panacea for mixing ills, and its effectiveness can be reduced drastically if all ports are not properly terminated.

DBM Port Terminations

Most DBM-performance inconsistencies occur because system source and load impedances presented to the mixer are not matched at all frequencies encountered in normal operation. The terminations (attenuator pads) used in conjunction with test equipment by manufacturers to measure published performance characteristics are indeed "broadband" matched. Reactive mixer terminations can cause system problems, and multiple reactive terminations can usually compound these problems to the point where performance is very

†See appendix on noise figure.

‡See appendix for mixer terminology.

TABLE 1

Manufacturer:	Relcom	Anzac	MCL	MCL	MCL	MCL
Model	M6F*	MD-108	SRA-1	SRA-1H	RAY-1	MA-1 ***
Frequency range, LO	2-500	5-500	.5-500	.5-500	5-500	1-2500
MHz rf	2-500	5-500	.5-500	.5-500	5-500	1-2500
i-f	DC-500	DC-500	DC-500	DC-500	DC-500	1-1000
Conversion loss, Mid-range	9 dB Max.	7.5 dB Max.	6.5 dB Typ.	6.5 dB Typ.	7.5 dB Typ.	8.0 dB Typ. 1-2.5 GHz.
Isolation, LO – RF	35-40 dB Min.	40 dB Min.	45 dB Typ.	45 dB Typ.	40 dB Typ.	40 dB Typ.
mid-range LO – i-f	25-35 dB Min.	35 dB Min.	40 dB Typ.	40 dB Typ.	40 dB Typ.	40 dB Typ.
Total input power:	50 mW	400 mW	500 mW	500 mW	1 W	50 mW
LO power requirement:	+7 dBm (5 mW)	+7 dBm (5 mW)	+7 dBm (5 mW)	+17 dBm (50 mW)	+23 dBm (200 mW)	+10 dBm (10 mW)
Signal 1-dB compression level:	Not spec.	Not spec.	+1 dBm	+10 dBm	+15 dBm	+7 dBm
Impedance, all ports:	50 ohms	50 ohms	50 ohms	50 ohms	50 ohms	50 ohms
Price class:	**	$7 Single unit	$9.95 Single unit	$15.95 5+ units	$31.95 4+ units	$99.95 Single unit

All specifications apply only at stated LO power level.

* 1968 data.
** Units provided by a second source.
*** SMA connectors standard.
Relcom, Division of Watkins-Johnson, 3333 Hillview Ave., Palo Alto, CA 94304.
Anzac Electronics, 39 Green Street, Waltham, MA 02154.
MCL – Mini-Circuits Laboratory, 837–843 Utica Ave., Brooklyn, NY 11203.

difficult to predict. Let's see how we can deal with reactive terminations.

The I-F Port

The i-f port is very sensitive to mismatch conditions. Reflections from the mixer/i-f amplifier interface (the pi network in Fig. 2) can cause the conversion loss to vary as much as 6 dB. Also greatly affected are third-order intermodulation-product ratio and the suppression of spurious signals, both of which may vary ±10 dB or more. It is ironic that the i-f port is the most sensitive to a reactive termination, as this is a receiving system point where sharp-skirted filters are often desired.

Briefly, here is what happens with a reactive i-f-port termination. Fig. 1 shows a DBM with "high side" LO injection and an i-f termination matched at fLO – $f$rf but reactive to fLO + $f$rf. The latter term re-enters the mixer, again combines with the LO and produces terms that exit at the rf port, namely 2fLO + $f$rf, a dc term, and fLO + $f$rf – fLO (the original rf-port input frequency). This condition affects conversion loss, as mentioned earlier, in addition to rf-port VSWR, depending on the phase of the reflected signal. The term 2fLO + $f$rf also affects the harmonic modulation-products spectrum resulting in spurious responses.

One solution to the i-f-interface problem is the use of a broadband 50-ohm resistive termination, like a pad, to minimize reflections. In deference to increased post-conversion system noise figure, it seemed impractical to place such a termination at the mixer i-f output port. While a complimentary filter or diplexer (high-pass/low-pass filters appropriately terminated) can be used to terminate both $f$rf + fLO and $f$rf – fLO§, a simpler method can be used if $f$rf + fLO is less than 1 GHz and ($f$rf + fLO) / ($f$rf – fLO) ≥ 10. Place a short-circuit termination to $f$rf + fLO, like a simple lumped capacitance, directly at the mixer i-f terminal. This approach is easiest for the amateur to implement and duplicate, so a form of it was tried – with success. In our circuit, C1 serves a dual purpose. Its reactance at $f$rf + fLO is small enough to provide a low-impedance "short-circuit" condition to this term for proper mixer operation. Additionally, it is part of the input reactance of the mixer i-f-amplifier interface. Fortunately the network impedance-transformation ratio is large enough, and in the proper direction, to permit a fairly large amount of capacitance (low reactance) at the mixer i-f-output port. The capacitor, in its dual role, must be of good quality at vhf/uhf (specifically $f$rf + fLO), with short leads, to be effective. The mixer condition ($f$rf + fLO) / ($f$rf – fLO) ≥ 10 is met at 432 and 220 MHz with a 404/192-MHz LO (28-MHz i-f) and on 14 MHz with a 130-MHz LO (14-MHz i-f). At 50 MHz, with a 36 MHz LO, we are slightly shy of the requirement, but no problems were encountered in an operating unit. The pi-type interface circuit assures a decreasing impedance as i-f operation departs from midband, thereby lessening IMD problems.

§Presentation and calculation format of these terms is based on "low side" LO injection. See the appendix for explanation.

The LO Port

The primary effect of a reactive LO source is an increase in harmonic modulation and third-order IMD products. If the drive level is adequate, no effect is noted on conversion loss, rf compression and desensitization levels. A reactive LO source can be mitigated by simply padding the LO port with a 3- or 6-dB pad and increasing the LO drive a like amount. If excess LO power is not available, matching the LO source to the mixer will improve performance. This method is acceptable for single-frequency LO applications, when appropriate test equipment is available to evaluate matching results. For simplicity, a 3-db pad was incorporated at the LO-input port as an interface in both versions of the mixer. Thus the LO port is presented with a reasonably broadband termination, and is relatively insensitive to applied frequency, as long as it is below about 500 MHz. This implies that frequencies other than amateur assignments may be covered – and such is indeed the case when appropriate LO frequencies and rf amplifiers are used. Remotely located LOs, when adjusted for a 50-ohm load, can be connected to the mixer without severe SWR and reflective-loss problems in the transmission line.

Broadband mixers exhibit different characteristics at different frequencies, due to circuit resonances and changes in diode impedances resulting from LO power-level changes. Input impedances of the various ports are load dependent, even though they are isolated from each other physically, and by at least 35 dB electrically. At higher frequencies, this effect is more noticeable, since isolation tends to drop as frequency increases. For this reason, it is important to maintain the LO power at its appropriate level, once other ports are matched.

The RF Port

A reactive rf source is not too detrimental to system performance. This is good, since the output impedance of most amateur preamplifiers is seldom 50 ohms resistive. A 3-dB pad is used at the rf port in the 50- and 144-MHz mixer to 14 MHz, and a 2-dB pad is used in the 220/432-MHz mixer to 28 MHz, although they add directly to mixer noise figure. Rf inputs between about 80 and 200 MHz are practical in the 14-MHz i-f-output model, while the 28-MHz-output unit is most useful from 175 to 500 MHz. Mixer contribution to system noise figure will be almost completely overcome by a low-noise rf amplifier with sufficient gain and adequate image rejection.

Image Response

Any broadband mixing scheme will have a potential image-response problem. In most amateur vhf/uhf receiver systems (as in these units) single-conversion techniques are employed, with the LO placed below the desired rf channel for non-inverting down-conversion to i-f. Conversion is related to both i-f and LO frequencies and, because of the broadband nature of the DBM, input signals at the rf image frequency (numerically fLO – fi-f in our case) will legitimately appear inverted at the i-f-output port, unless proper filtering is used to reduce them at the mixer rf-input port. For example, a 144-MHz converter with a 28-MHz i-f output (116-MHz LO) will have rf image-response potential in the 84 to 88-MHz range. TV channel 6 wideband-fm audio will indeed appear at the i-f-output port near 28 MHz unless appropriate rf-input filtering is used to eliminate it. While octave-bandwidth vhf/uhf "imageless mixer" techniques can improve system noise performance by about 3 dB (image noise reduction), and image signal rejection by 20 dB – and much greater with the use of a simple gating scheme – such a system is a bit esoteric for our application. Double or multiple-conversion techniques can be used to advantage, but they further complicate an otherwise simple system. Image noise and signal rejection will depend on the effectiveness of the filtering provided in the rf-amplifier chain.

Mixer Selection

The mixer used in this system is a Relcom M6F, with specifications given in Table I. Suitable substitute units are also presented. The M6F is designed for printed-circuit applications (as are the recommended substitutes), and the lead pins are rather short. While mixers are available with connectors attached, they are more expensive. The simple package is suggested as, aside from less expense, improved interface between mixer and i-f amplifier is possible because of the short leads. The combining of mixer and i-f amplifier in one converter package was done for that reason. Along these lines, the modular-construction approach

This top view of the DBM/i-f amplifier shows the plastic mixer package plus rf/LO inputs and i-f output jacks clearly marked for cabling. The unit is mounted on the open face of a standard 6 × 4 × 2-inch aluminum chassis. This shielding is necessary to prevent the 3N140 from picking up external signals in the 14-MHz region.

The bottom view of the DBM/i-f amplifier shows component and shielding layout. L1, the mixer-amplifier interface inductance and associated components ere Indicated. C1, with its wide silver-strap leads, is connected directly between the mixer i-f output pin and the copper-clad ground plane with essentially zero lead length. Connection between the mixer output pin and other components (L1, C2 end the rf choke for d-c return) is made by using excess lead from C1. The 43-ohm, 1/4-W resistor in the 3N140 gate 1 lead is connected between the high-impedance end of L1 and a spare terminal on the coil form. The device gate No. 1 lead end resistor are joined at this point. It is important that input/output isolation of the 3N140 be mainteined as it is opereting at high gain. Mixer packages other than the M6F may have different pin connections and require slightly different input-circuit layout and shielding. Double-sided copper-cled board was used throughout.

permits good signal isolation and enables the mixer-amplifier/i-f system to be used at a variety of rf and LO-input frequencies, as mentioned earlier.

Most commonly available, inexpensive DBM are not constructed to take advantage of LO powers much above +10 dBm (10 mW). To do so requires additional circuitry which could degrade other mixer characteristics, specifically conversion loss and inter-port isolation. The advantage of higher LO power is primarily one of improved strong-signal-handling performance. At least one manufacturer advertises moderately priced "high-level" receiving DBM which can use up to +23 dBm (200 mW) LO power, and still retain excellent conversion loss and isolation characteristics, shown in Table I. The usefulness of mixers with LO power requirements above the commonly available +7 dBm (5 mW) level in amateur receiving applications may be a bit moot, as succeeding stages in most amateur receivers will likely overload before the DBM. Excessive overdesign is not necessary.

In general, mixer selection is based on the lowest practical LO level requirement that will meet the application, as it is more economical and results in the least LO leakage within the system. As a first-order approximation, LO power should be 10 dB greater than the highest anticipated input-signal level at the rf port. Mixers with LO requirements of +7 dBm are quite adequate for amateur receiving applications.

Application Design Guidelines

While the material just presented only scratches the surface in terms of DBM theory and utilization in amateur vhf/uhf receiving systems, some practical solutions to the non-ideal mixer-port-termination problem have been offered. To achieve best performance from most commercially manufactured broadband DBM in amateur receiver service, the following guidelines are suggested:

- Choose i-f and LO frequencies that will provide maximum freedom from interference problems. Don't "guesstimate," go through the numbers!
- Provide a proper i-f-output termination (most critical).
- Increase the LO-input power to rf-input power ratio to a value that will provide the required suppression of any in-band interfering products. The specified LO power (+7 dBm) will generally accomplish this.
- Provide as good an LO match as possible.
- Include adequate pre-mixer rf-image filtering at the rf port.

When the mixer ports are terminated properly performance usually in excess of published specifications will be achieved – and this is more than adequate for most amateur vhf/uhf receiver mixing applications.

This is a side view showing construction details for tha double-tuned i-f output circuit. The 3N140 drain lead passes through the shield wall via a small Teflon press-fit bushing and is connected directly to L2. A dc-input isolation compartment elong with device gate 2 biasing components (bias configuration modified slightly after photograph was taken), can be seen to the left of the i-f-output components. L2 and L3 era spaced 1–1/8-inch (2.9 cm) center-to-center in the 14-MHz model shown, and 1 inch (2.5 cm) apart in the 28-MHz unit.

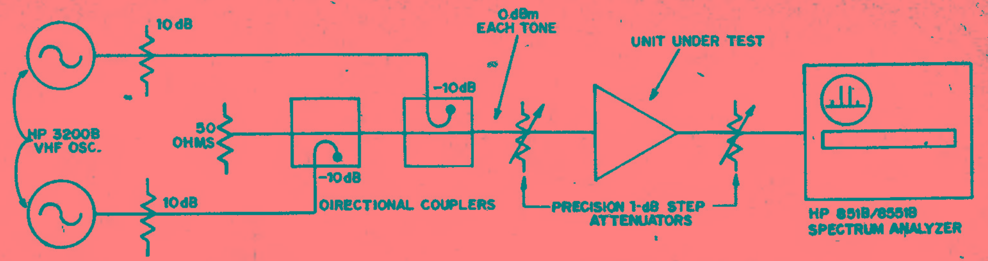


Fig. 6 — A test setup used to measure IMD. The first attenuator adjusts the input level to the unit under test. The second one provides a means of staying within the linear range or the spectrum analyzer.

The Combined DBM/I-F Amplifier

A low-noise i-f amplifier (2 dB or less) following the DBM helps ensure an acceptable system noise figure when the mixer is preceeded by a low-noise rf amplifier. A pi-network matching system used between the mixer i-f-output port and gate 1 of the 3N140 transforms the nominal 50-ohm mixer-output impedance to a 1500-ohm gate-input impedance (at 28 MHz) specifically for best noise performance. The network forms a narrow-band mixer/i-f-output circuit which serves two other important functions: It helps achieve the necessary isolation between rf-and i-f signal components, and serves as a 3-pole filter, resulting in a monotonic decrease in match imperances as the operating i-f departs from mid-band. This action aids in suppression of harmonic-distortion products.

The combined DBM/i-f amplifier is shown schematically in Fig. 2 and pictorially in the photographs. In the 14-MHz model, the 3N140 drain is tapped down on its associated inductance to provide a lower impedance for better strong-signal-handling ability. The 3N140 produces about 19 dB gain across a 700-kHz passband, flat within 1 dB between 13.8 and 14.5 MHz. A 2-MHz passband is used for the 28-MHz model, and the device drain is connected directly to the high-impedance end of its associated inductance. Both amplifiers were tuned independently of their respective mixers, and checked for noise figure as well as gain. With each i-f amplifier pretuned and connected to its mixer, signals were applied to the LO and rf-input ports. The pi-network inductance in the i-f interface was adjusted carefully to see if performance had been altered. No change was noted. I-f gain is controlled by the externally accessable potentiometer. Passband tuning adjustments in the drain circuit are best made with a sweep generator, but single-signal tuning techniques will be adequate. While there should be no difficulty with the non-gate-protected 3N140, a 40673 may be substituted directly if desired.

DBM/I-F Amplifier IMD Evaluation

Classical laboratory IMD measurements made on the DBM/i-f amplifier, using the test setup shown in Fig. 6, from both tones of a two-equal-tone rf-input test signal consisting of −10 dBm each tone. The tones were closely spaced in the 144-MHz range, and converted to 14 MHz LO. Close spacing was necessary to ensure third-order products would appear essentially unattenuated within the relatively narrow i-f-output passband. In operation, as simulated by these test conditions, equivalent output signal levels at J3 would be strong enough to severely overload most amateur receivers. Perhaps the early Collins 75A series, R390A and those systems described by Sabin[4] and Hayward[5] would still be functioning well.

A high-performance, small-signal, vhf/uhf receiving amplifier optimized for IMD reduction and useful noise figure is only as good as any succeeding receiving-system stage, in terms of overload. The DBM/i-f-amplifier combination presented significantly reduces common first-mixer overload problems, leaving the station receiver as the potentially weak link in the system. When properly understood and employed, the broadband DBM followed by a selective low-noise i-f amplifier can

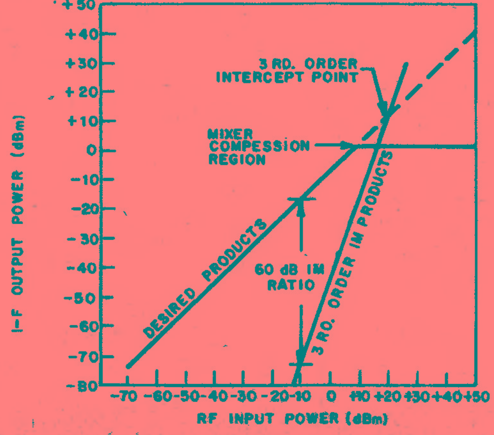


Fig. 7 — A third-order intercept point is determined by extrapolating the desired product curve beyond the mixer compression point and intersecting with the third-order IM-product curve. In this case LO power is +7 dBm, conversion loss is 5 dB.

Table II
DBM/IF AMPLIFIER
PARTS LIST

	14 MHz i-f output	*28 MHz i-f output*
C1	470 pF JFD 471J or equal.	300 pF JFD 301J or equal.
C2	390 pF SM	not used
C3	180 pF SM	51 pF SM
C4	39 pF SM	18 pF SM
C5	56 pF SM	27 pF SM
C6	300 pF SM	150 pF SM
L1	9 turns No. 18 enam., close-wound on a 3/8-inch (.95 cm) diameter red-slug coil form.	9 turns No. 24 enam., close-wound on a 1/4-inch (.65 cm) diameter green-slug coil form.
L2	18 turns No. 26 enam., close-wound on a 3/8-inch (.95 cm) diameter red-slug coil form. Tap down 7 turns from top for 3N140 drain connection. See text.	12 turns No. 26 enam., close-wound on a 1/4-inch (.65 cm) diameter green-slug coil form. No tap used.
L3	Same as L2 but no tap, spaced 1-1/8 inch (2.9 cm) center-to-center with L2.	Same as L2, spaced 1 inch (2.5 cm) center-to-center with L2.
R1, R3	300 ohm, 1/4 W carbon.	430 ohm, 1/4 W carbon.
R2	16 ohm, 1/4 W carbon.	11 ohm, 1/4 W carbon.
Note:	Ferrite beads can be replaced by a 10-ohm, 1/4 W carbon resistor at one end of the choke, if desired.	
	SM = Silver Mica.	

be a useful tool for the amateur vhf/uhf receiver experimenter.

Appendix

Mixer Terminology

$f$rf – rf input frequency
fLO – local-oscillator input frequency
fi-f – i-f output frequency

By convention, mixing signals and their products are referred to the LO frequency for calculations. In the mixer system presented, $f$rf is always above fLO, so we will refer our signals to $f$rf, with the exception of Fig. 1 which uses the fLO reference.

Overload

A generic term covering most undesired operating phenomena associated with device non-linearity.

Harmonic Modulation Products

Output responses caused by harmonics of fLO and $f$rf and their mixing products.

RF Compression Level

The absolute single-signal rf input-power level that causes conversion loss to increase by 1 dB.

RF Desensitization Level

The rf input power of an interfering signal that causes the small-signal conversion loss to increase by 1 dB, i.e. reducing a weak received signal by 1 dB.

Intermodulation Products

Distortion products caused by multiple rf signals and their harmonics mixing with each other and the LO, producing new output frequencies.

Mixer Intermodulation Intercept Point

Because mixers are nonlinear devices, all signals applied will generate others. When two signals (or tones), $F1$ and $F2$, are applied simultaneously to the rf-input port, additional signals are generated and appear in the output as $f\mathrm{LO} \pm (nF1 + mF2)$. These signals are most troublesome when $n \pm m$ is a low odd number, as the resulting product will lie close to the desired output. For n-1 (or 2) and m–2 (or 1), the result is three (3), and is called the two-tone/third-order intermodulation products When $F1$ and $F2$ are separated by 1 MHz, the third-order products will lie 1 MHz above and below the desired outputs. Intermodulation is generally specified under anticipated operating conditions since performance varies over the broad mixer-frequency ranges. Intermodulation products may be specified at levels required (i.e. 50 dB below the desired outputs for two 0-dBm input signals) or by the intercept point.

The intercept point is a fictitious point determined by the fact that an increase of level of two input tones by 10 dB will cause the desired output to increase by 10 dB, but the third-order output will increase by 30 dB. If the mixer exhibited no compression, there would be a point at which the level of the desired output would be equal to that of the third-order product. This is called the third-order intercept point and is the point where the desired-output slopes and third-order slopes intersect (Fig. 7).

Noise Figure

Noise figure is a relative measurement based on excess noise power available from a termination (input resistor) at a particular temperature (290 degrees K). When measuring the NF of a double balanced mixer with an automatic system, such as the HP 342A, a correction may be necessary to make the meter reading consistant with the accepted definition of receiver noise figure.

In a broadband DBM, the actual noise bandwidth consists of two i-f passbands, one on each side of the local-oscillator frequency (fLO + fi-f and fLO – fi-f). This double sideband (dsb) i-f response includes the rf channel and its image. In general, only the rf channel is desired for further amplification. The image contributes nothing but receiver and background noise.

When making an automatic noise-figure measurement using a wideband noise source, the excess noise is applied through both sidebands in a broadband DBM. Thus the instrument meter indicates NF as based on both sidebands. This means that the noise in the rf and image sidebands is combined in the mixer i-f-output port to give a double contribution (3 dB greater than under ssb conditions). For equal rf-sideband responses, which is a reasonable assumption, and in the absence of preselectors, filters, or other image rejection elements, the automatic NF meter readings are 3 dB lower than the actual NF for DBM measurements.

The noise figure for receivers (and most DBM) is generally specified with only one sideband for the useful signal. As mentioned in the text, most DBM diodes add no more than 0.5 dB (in the form of NF) to conversion loss, which is generally measured under single-signal rf-input (ssb) conditions. Assuming DBM conversion efficiency (or loss) to be within specifications, there is an excellent probability that the ssb NF is also satisfactory. Noise figure calculations in the text were made using a graphical solution of the well known noise-figure formula:

$$fT = f_1 + \frac{f_2 - 1}{g_1}$$

converted to dB.

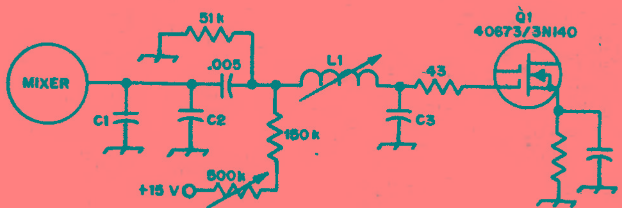
Improved Wide Band I-F Responses

The following information was developed in achieving broad-band performance in the mixer-to-amplifier circuitry. In cases where only a small portion of a band is of interest the original circuit values are adequate. For those who need to receive over a considerable portion of a band, say one to two MHz, a change of some component will provide improved performance over a broad range while maintaining an acceptable noise figure.

The term "nominal 50-ohm impedance" applied to diode DBM ports is truly a misnomer, as their reflective impedance is rarely 50 ohms $+j0$ and a VSWR of 1 is almost never achieved. Mixer performance specified by the manufacturer is measured in a 50-ohm broadband system, and it is up to the designer to provide an equivalent termination to ensure that the unit will meet specifications. Appropriate matching techniques at the rf and LO ports will reduce conversion loss and low-power requirements. Complex filter synthesis can improve the i-f output match. However, if one does not have the necessary equipment to evaluate his efforts, they may be wasted. Simple, effective, easily reproduced circuitry was desired as long as the trade-offs were acceptable, and measurements indicate this to be the case.

The most critical circuit in the combined unit is the interface between mixer and i-f amplifier. It must be low-pass in nature to satisfy vhf signal component bypassing requirements at the mixer i-f port. For best mixer IMD characteristics and low conversion loss, it must present to the i-f port a nominal 50-ohm impedance at the desired frequency, and this impedance value must not be allowed to increase as i-f operation departs from midband. The impedance at the i-f amplifier end of the interface network must be in the optimum region for minimum cross-modulation and low noise. A dual-gate device offers two important advantages over most bi-polars. Very little, if any, power gain is sacrificed in achieving best noise figure, and both parameters (gain and NF) are relatively independent of source resistance in the optimum region. As a result, the designer has a great deal of flexibility in choosing a source impedance. In general, a 3:1 change in source resistance results in only a 1-dB change in NF. With minimum cross-modulation as a prime system consideration, this 3:1 change (reduction) in source resistance implies a 3:1 improvement in cross-modulation and total harmonic distortion.

Tests on the 3N201 dual-gate MOSFET have shown device noise performance to be excellent for source impedances in the 1-kΩ to 2-kΩ region. For optimum noise and good cross-modulation



14 MHz	28 MHz
C1 — 300 pF (JFD 301).	100 pF (JFD 101).
C2 — 51 pF S.M.	Not used
C3 — 68 pF S.M.	7.5 pF S.M.
L1 — 15t No. 24 enameled on 3/8-inch dia red-slug form. 1.5-2.5 μH range, 1.95 μH for network.	19t No. 26 enameled on 1/4-inch dia red-slug form.

Fig. 1 — Suggested changes in the mixer-to 3N140 pi-network interface circuit, producing lower Q_L and better performance. See the original article for additional circuit details.

performance, the nominal 50-ohm mixer i-f output impedance is stepped up to about 1500 ohms for i-f amplifier gate 1, using the familiar low-pass pi network. This is a mismatched condition for gate 1, as the device input impedance for best gain in the hf region is on the order of 10 kΩ network loaded-Q values in the article are a bit higher than necessary, and a design for lower Q_L is preferred. Suggested modified component values are listed in Fig. 8. High-frequency attenuation is reduced somewhat, but satisfactory noise and bandwidth performance is more easily obtained. Coil-form size is the same, so no layout changes are required for the modification. Components in the interface must be of high Q and few in number to limit their noise contribution through losses. The 28-MHz values provide satisfactory interface network performance over a 2-MHz bandwidth. A higher Q_L in the 28-MHz interface can be useful if one narrows the output network and covers only a few hundred kilohertz bandwidth, as is commonly done in 432-MHz weak-signal work.

Device biasing and gain control methods were chosen for simplicity and adequate performance. Some sort of gain adjustment is desirable for drain-circuit overload protection. It is also a handy way to "set" the receiver S meter. A good method for gain adjustment is reduction of the gate-2 bias voltage from its initial optimum-gain bias point (greater than +4 V dc), producing a remote-cutoff characteristic (a gradual reduction in drain current with decreasing gate bias). The initial gain-reduction rate is higher with a slight forward bias on gate 1, than for $Vg_1s = 0$. Input and output circuit detuning resulting from gain reduction (Miller effect) is inconsequential as the gate-1 and drain susceptances change very little over a wide range of Vg_2s and I_D at both choices of i-f. Best intermodulation figure for the 3N201 was obtained with a small forward bias on gate 1, and the bias-circuit modification shown may be tried, if desired.

References

[1] Fisk, "Double-Balanced Mixers," *Ham Radio*, March, 1968.

[2] Ress, "Broadband Double-balanced Modulator," *Ham Radio*, March, 1970.

[3] DeMaw and McCoy: "Learning to Work With Semiconductors," Part IV, *QST*, July, 1974.

[6] Sabin, "The Solid-State Receiver," *QST*, July, 1970.

[7] Hayward, "A Competition-Grade CW Receiver," *QST*, March and April 1974.

AN OSCAR UP-CONVERTER

Many amateur operators who wish to receive the 10-meter signals from the Oscar satellites do so with an average receiver that is already "at hand" in the shack, sometimes adding a preamplifier for improved performance. Others use a converter to translate the signals to a lower frequency where the station receiver is more stable or more sensitive.

However, there is another approach that should be explored – that of converting the 10-meter signals *up* to a higher band. Just a very short time ago this system would have been impractical, if not ridiculous, because of the complexity and size of the equipment involved. Recent developments in two-meter transceivers make the up-conversion scheme practical and attractive. A nearly "ideal" Oscar package can be obtained by the addition of a small converter which allows a normal transceive style of operation with the vhf equipment.

There are several makes of 144-MHz ssb transceivers available, but only a few are beginning to appear on the market in the Western Hemisphere. The KLM Echo II was used here to evaluate the technique and test the performance of the converter that was assembled. This particular transceiver had been modified to permit cw operation as well as the usual ssb – a desirable feature to look for in any equipment being contemplated.

In some instances it may seem a bit redundant to convert a 28-MHz signal to 144 MHz, only to have it converted back down to 28 MHz in the receiver first mixer. However, there are reasons why this scheme is not all that bad, and a chief one is that the frequencies do not translate directly in all cases. A secondary, but important, consideration is that it may not be desirable or possible to modify the equipment to accept a 28-MHz input. And of course not all transceivers have 28 MHz as a first i-f.

The Converter

If the pc board and parts-placement layout appears familiar, it is because an existing design was modified to serve our purpose. Rather than go through the entire process of developing a new board the "Rochester" converter was rehabilitated for this project. See pages 300-304 for more details of these converters. Most suppliers of amateur radio pc boards have this pattern on hand, and many have etched and drilled boards in stock. There have been a few changes to some parts of the circuit, necessitating the placing of one capacitor on the foil side of the board. In operation the converter reverses the process of the original in that it first amplifies the 10-meter signals in Q1 (Fig. 1), then mixes them with 116.45-MHz energy in Q2, to provide an output between 145.85 and 145.95 MHz. The original oscillator and harmonic-generator circuit proved adequate with a slight modification; a third-overtone crystal with a frequency of 58.225 was used instead of the 38.6-MHz unit specified earlier. A buffer stage (Q4) is necessary to allow some rejection of unwanted harmonics while maintaining a suitable injection voltage for the mixer.

Construction

The assembly of the converter is greatly simpli-

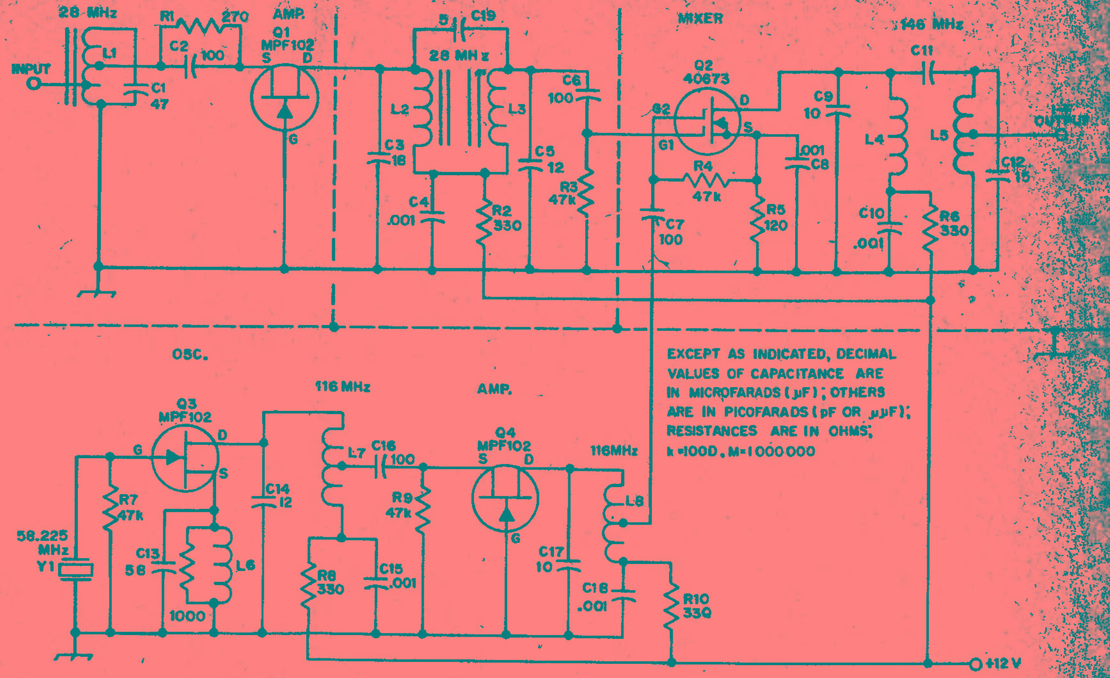


Fig. 1 — The schematic diagram of the "Rochester" converter as modified for up-conversion.

C11 — Two 1/2-inch pieces No. 18 insulated hookup wire, twisted together 1/2 turn.

L1, L2, L3 — 18 turns No. 28 enam. wound on Amidon T-30-6 core. L1 tapped at 6 turns and 11 turns from ground end.

L4, L5 — 5 turns No. 20 enem., formed by using threads of 1/4-20 bolt as a guide. L5 is tapped 2 turns from the ground end.

L6 — 10 turns No. 24 enam. close wound on the body of a 1000-ohm 1/2-watt resistor.

L7, L8 — 5 turns No. 20 enam., formed the same as L4. Both are tapped 2 turns from the hot end.

Y1 — 58.225-MHz crystal. International Crystal third-overtone type in FM-1 (wire leads) or FM-2 (pins) holder. Calibration tolerance of .0025% at a load capacitance of 20 pF used hera. Genaral-purpose (.01%) tolerance may be adequata in most instances.

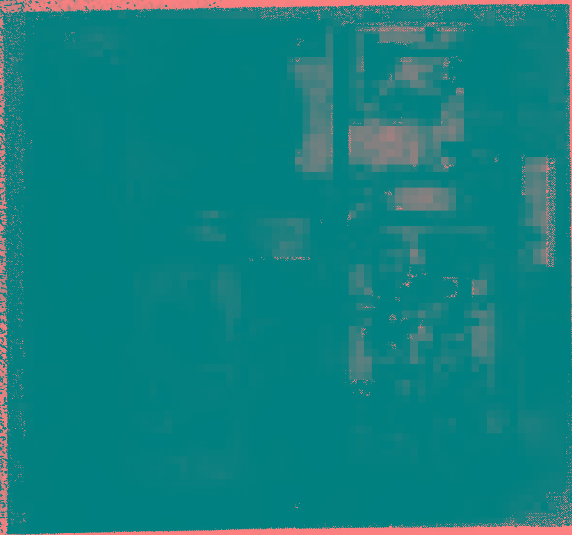
fied by the use of a ready-made board, since for the most part the work consists of placing the component leads through the holes and soldering them in place.

Because of slight differences in materials used in toroids, it will be necessary to adjust the input and interstage tuned circuits to resonance after the converter is completed. This was done by the substitution method, placing different values of capacitance across the windings while monitoring the signal on a receiver. There are miniature trimmers available that would fit into the space, but the cost of these devices is a bit high. Alternatively, larger trimmers could be mounted below the pc board, if the builder allows enough space between the board and the enclosure to which it is fastened. Proper operation of the oscillator and buffer stages can be ascertained by using a grid-dip meter to indicate output on the correct harmonic of Y1.

There is sufficient space in some transceivers to allow the converter to be mounted internally, as was the case in the Echo II unit shown in the photographs. A small bracket was fastened to the rear apron of the equipment to provide a mount for the 28-MHz input connector. Where a transceiver is too compact to allow this style of assembly, the converter could be fastened in a small Minibox for shielding. A source of +12 V. and a small-diameter coax for input and output connections makes the wiring job simple. In some equipment it may be necessary to disconnect the receiver input wiring from the T/R relay but this was not done in the Echo II. The output from the converter was wired in parallel with the receiver input lead. Any small amount of noise picked up by the receiver was masked by the output from the converter. However, in areas of high local activity or high ambient noise levels, it will be necessary to disable the normal receiver input.

Performance

Sensitivity of the converter was sufficient that the residual output of a Model 80 signal generator



In this particular model transceiver, room was available to mount the converter in an inverted position just below the speaker. A small bracket is fastened under one of the bolts that also holds a transformer to tha bottom sidewall. Cables are routed along with axisting wiring harnesses and tied in place. The 28-MHz input connector is fastened to the rear apron of the equipment.

could be heard clearly. A 1-μV signal was loud enough that it evoked the immediate reaction of turning down the audio gain control on the transceiver. Ignition noise picked up by the variety of antennas tried was strong enough to be bothersome at times, further attesting to the sensitivity of the converter – it also pointed out the usefulness of the noise blanker in the Echo II. Considering the performance of the converter/transceiver combination, the addition of a preamplifier or an i-f post amplifier was not considered necessary. Additional gain could even be detrimental by causing overloading or intermodulation problems – there was no evidence of these problems during several tests.

Since the first i-f in the transceiver is at 28 MHz, the question of possible "leak-through" of local signals was raised. No indication of this type of interference was found during receiving tests, but admittedly it could happen. The output circuit of the mixer (L4, L5, C11) has a band-pass characteristic centered on 146 MHz and should provide a high degree of attenuation to hf-band signals.

In areas where strong local operation does cause such leakage of signals through the converter it will be necessary to install a high-pass filter between the converter output and the receiver input. Designs for such filters can be found in the ARRL *Handbook*. Because the filter will be used at essentially zero power level, it can be made physically quite compact. Of course good shielding and high-quality coaxial cable is a must in any effort to keep unwanted signals out – the best filters in the world will do no good if there is a path around them.

[EDITOR'S NOTE: The parts placement for this up-converter is virtually identical to that used in the "Rochester" converters, found elsewhere in this chapter. The reader can follow that layout, keeping in mind the differences in tuned-circuit frequencies throughout.]

Top viaw of the modified Rochester Converter. A mounting bracket has been fastened to tha lower right corner. The 28-MHz input is to the upper right, with the rf amplifier along the top of the board. Oscillator and buffer stages are located along the bottom portion. The i-f output coils, L4 and L5, are at the upper left with C11 (twisted wires) just below L4, adjacent to the resistor. The shialds between stages have been omitted for e bettar view.

INTERDIGITAL CONVERTER FOR 1296 OR 2304 MHz

In a world where rf spectrum pollution is becoming more serious, even into the microwave region, it is almost as important to keep unwanted signals out of a receiver as it is to prevent radiation of spurious energy. An interdigital filter was described some years ago, featuring low insertion loss, simplicity of construction, and reasonable rejection to out-of-band signals.[1] It could be used in either transmitters or receivers.

This twice-useful principle has now been put to work again – as a mixer. Again, the ease of construction and adaptation leads many to wonder that it had not been thought of before. It was first described by W2CQH in *QST* for January, 1974.

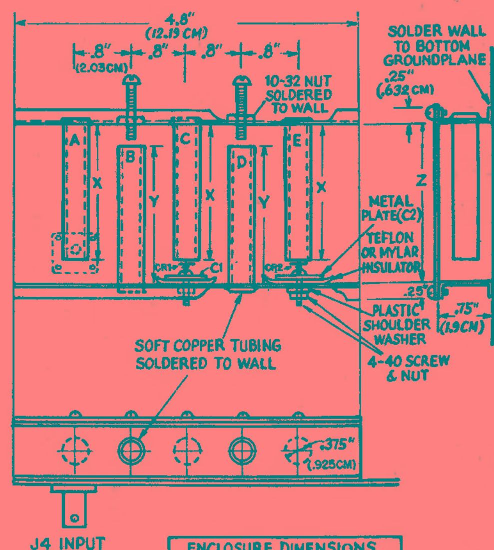
A Filter and Mixer

A layout of the microwave portions of both converters is shown in Fig. 1. The structure consists of five interdigitated round rods, made of 3/8-inch OD brass or copper tubing. They are soldered to two sidewalls and centrally located between two ground-planes made of 1/16-inch sheet brass or copper-clad epoxy fiber glass. One ground plane is made larger than the microwave assembly and thus provides a convenient mounting plate for the remainder of the converter components.

The sidewalls are bent from .032-inch thick sheet brass or they can be made from 1/4 × 3/4-inch brass rod. One edge of each sidewall is soldered to the larger ground plane. The other edge is fastened to the smaller ground plane by 4-40 machine or self-tapping screws, each located over the centerline of a rod. The sidewall edges should be sanded flat, before the ground plane is attached, to assure continuous electrical contact. Note that no end walls are required since there are no electric fields in these regions.

Electrically, rods A, B, and C comprise a one-stage, high-loaded-*Q* (Q_L = 100), interdigital filter[1] which is tuned to the incoming signal frequency near 1296 or 2304 MHz. The ungrounded end of rod A is connected to a BNC coaxial connector and serves as the coupling section to the filter input. Rod B is the high-*Q*

[1] Fisher, "Interdigital Bandpass Filters for Amateur VHF/UHF Applications," *QST* March, 1968.

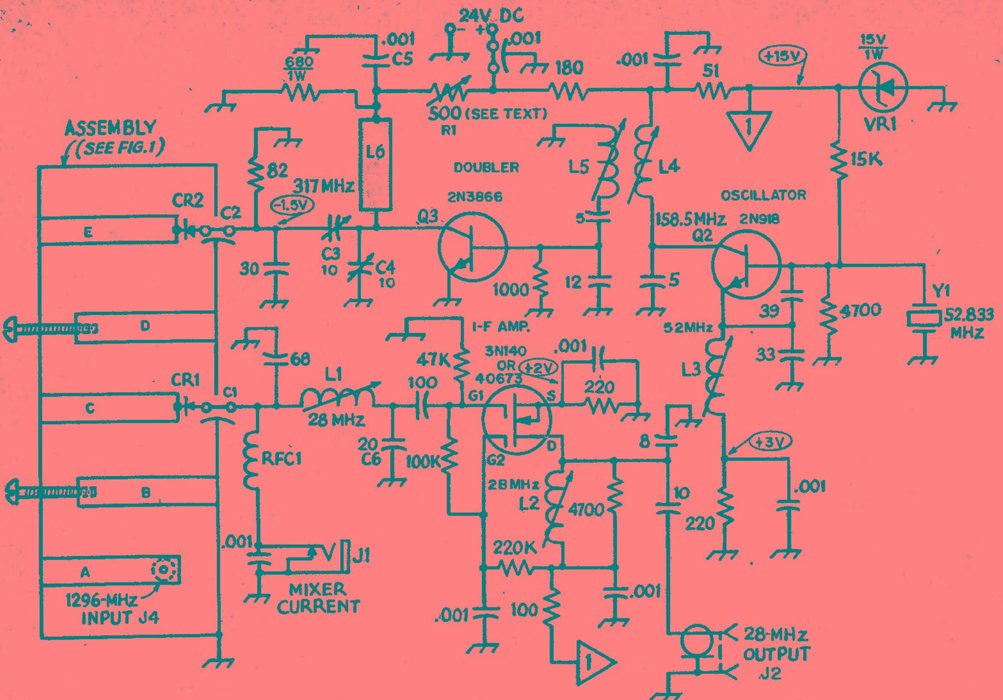


ENCLOSURE DIMENSIONS

FREQ.	X	Y	Z
1296 MHz	2.00" (5.08CM)	2.00" (5.08CM)	2.25" (5.71CM)
2304 MHz	.9375" (2.38CM)	1.00" (2.54CM)	1.25" (3.17CM)

Fig. 1 — Dimensions and layout for the filter and mixer portions of the interdigital converters. The signal input is to the left rod, labelled "A." Local-oscillator injection is through the diode to rod "E." CR1 is the mixer diode, connected to the center rod in the assembly.

The converter for 1296 MHz. This unit was built by R.E. Fisher, W2CQH. While the mixer assembly (top center) in this model has solid brass walls, it can be made from lighter material as explained in the text and shown in Fig. 1. The i-f amplifier is near the center, just above the mixer-current-monitoring jack, J1. A BNC connector at the lower left is for 28-MHz output. The local oscillator and multiplier circuits are to the lower right. Note that L6 is very close to the chassis, just above the crystal. The veriable capacitor near the crystal is an optional trimmer to adjust the oscillator to the correct frequency.



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF) ; OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS ; k = 1000, M = 1000 000.

Fig. 2 — Schematic diagram of the 1296-MHz converter with oscillator and multiplier sections included. Dimensions for the filter and mixer assembly are given in Fig. 1.

C1, C2 — 30-pF homemade capacitor. See text and Fig. 1.
C3, C4 — 0.8- to 10-pF glass trimmer, Johanson 2945 or equiv.
C5 — .001-μF button mica.
C6 — 2- to 20-pF air variable, E. F. Johnson 189-507-004 or equiv.
CR1 — Hewlett Packard 5082-2577 or 5082-2835.
CR2 — Hewlett Packard 5082-2811 or 5082-2835.
J1 — Closed-circuit jack.
J2 — Coaxial connector, chassis mount. Type BNC acceptable.
L1, L2 — 18 turns No. 24 enam. on 1/4-inch OD slug-tuned form (1.5 μH nominal).
L3 — 10 turns like L1 (0.5 μH).
L4, L5 — 6 turns like L1 (0.2 μH).
L6 — Copper strip, 1/2-inch wide × 2-1/2-inches (1.27 × 6.35 cm) long. See text and photographs.
RFC1 — 33 μH, J. W. Miller 74F33SAI or equiv.

resonator and is tuned by a 10-32 machine screw. Rod C provides the filter output-coupling section to the mixer diode, CR1.

The mixer diode is a Hewlett-Packard 5082-2577 Schottkey-barrier type which is available from distributors for about $4. The cheaper 5082-2835, selling for 90 cents, can be used instead, but this substitution will increase the 2304-MHz mixer noise figure by approximately 3 dB.

One pigtail lead of the mixer diode is tack-soldered to a copper disk on the ungrounded end of rod C. Care should be taken to keep the pigtail lead as short as possible. If rod C is machined from solid brass stock, then it is feasible to clamp one of the mixer-diode leads to the rod end with a small setscrew. This alternative method facilitates diode substitution and was used in the mixer models shown in the photographs.

Fig. 1 also shows that the other end of CR1 is connected to a homemade 30-pF bypass capacitor, C1, which consists of a 1/2-inch-square copper or brass plate clamped to the sidewall with a 4-40 machine screw. The dielectric material is a small screw passes through an oversize hole and is insulated from the other side of the wall by a small plastic shoulder washer.

In the first converter models constructed by the author and shown in the photographs, C1 was a 30-pF button mica unit soldered to the flange of a 3/8-inch diameter threaded panel bearing (H.H. Smith No. 119). The bearing was then screwed into a threaded hole in the sidewall. This provision made it convenient to measure the insertion loss and bandwidth of the interdigital filters since the capacitor assembly could be removed and replaced with a BNC connector.

Rods C, D, and E comprise another high loaded-Q ($Q_L = 100$) interdigital filter tuned to the local oscillator (LO) frequency. This filter passes only the fourth harmonic (1268 or 2160 MHz) from the multiplier diode, CR2. The two filters have a common output-coupling section (rod C) and their loaded Qs are high enough to prevent much unwanted coupling of signal power from the antenna to the multiplier diode and LO power back out to the antenna.

The multiplier diode is connected to the driver circuitry through C2, a 30-pF bypass capacitor identical to C1. CR2 is a Hewlett-Packard 5082-2811 although the 5082-2835 works nearly as well. Fifty milliwatts drive at one quarter of the LO frequency is sufficient to produce 2 mA of mixer diode current, which represents about 1 milliwatt of the local-oscillator injection. A Schottkey-barrier was chosen over the more familiar varactor diode for the multiplier because it is cheaper, more stable, and requires no idier circuit.

Table I

Converter Specifications

	1296 MHz	*2304 MHz*
Noise figure	5.5 dB	6.5 dB
Conversion gain	20 dB	14 dB
3-dB bandwidth	2 MHz	7 MHz
Imaga rejection	18 dB	30 dB
I-f output	28 MHz	144 MHz

Fig. 2 shows the schematic diagram of the 1296 to 28 MHz converter. All components are mounted on a 7 × 9-inch (17.8 × 22.9 cm) sheet of brass or copper-clad epoxy-fiber glass board. As mentioned earlier, this mounting plate also serves as one ground plane for the microwave mixer. When completed, the mounting plate is fastened to an inverted aluminum chassis which provides a shielded housing.

Oscillator and Multipliers

The nonmicrowave portion of the converter is rather conventional. Q1, a dual-gate MOSFET, was chosen as the 28-MHz i-f amplifier since it can provide 25 dB of gain with a 1.5-dB noise figure. The mixer diode is coupled to the first gate of Q1 by a pi-network matching section. It is most important that the proper impedance match be achieved between the mixer and i-f amplifier if a low noise figure is to be obtained. In this case, the approximately 30-ohm output impedance of the mixer must be stepped up to about 1500 if Q1 is to yield its rated noise figure of 1.5 dB. It is for this reason that a remote i-f amplifier was not employed as is the case with many contemporary uhf converters.

Q2 functions in a oscillator-tripler circuit which delivers about 10 milliwatts of 158.5-MHz drive to the base of Q3. The emitter coil, L3, serves mainly as a choke to prevent the crystal from oscillating at its fundamental frequency. Coils L4 and L5, which are identical, should be spaced closely such that their windings almost touch.

Q3 doubles the frequency to 317 MHz, providing about 50 milliwatts drive to the multiplier diode. It is important that the emitter lead of Q3 be kept extremely short; 1/4-inch (6.36 mm) is probably too long. L6, the strip-line inductor in the collector circuit of Q3, consists of a 1/2 × 2-1/2-inch (1.27 × 6.35 cm) piece of flashing copper spaced 1/8-inch (3.18 mm) above the ground plane. The cold end of L6 is bypassed to ground by C5, a .001-μF button mica capacitor.

The multiplier circuits are tuned to resonance in the usual manner by holding a wavemeter near each inductor being tuned. Resonance in the Q3 collector circuit is found by touching a VTVM probe (a resistor must be in the probe) to C2 and adjusting the Johanson capacitors until about −1.5 volts of bias is obtained. The 317- to 1268-MHz multiplier cavity is then resonated by adjusting the 10-32 machine screw until maximum mixer current is measured at J1. When resonance is found, R1 should be adjusted so that about 2 mA of mixer current is obtained. As an alternative to mounting a potentiometer in the converter, once a value of resistance has been found that provides correct performance it can be measured and the nearest standard fixed-value resistor substituted. Some means of adjusting the collector voltage on the multiplier stage must be provided initially to allow for the nonuniformity of transistors.

A 2304-MHz Version

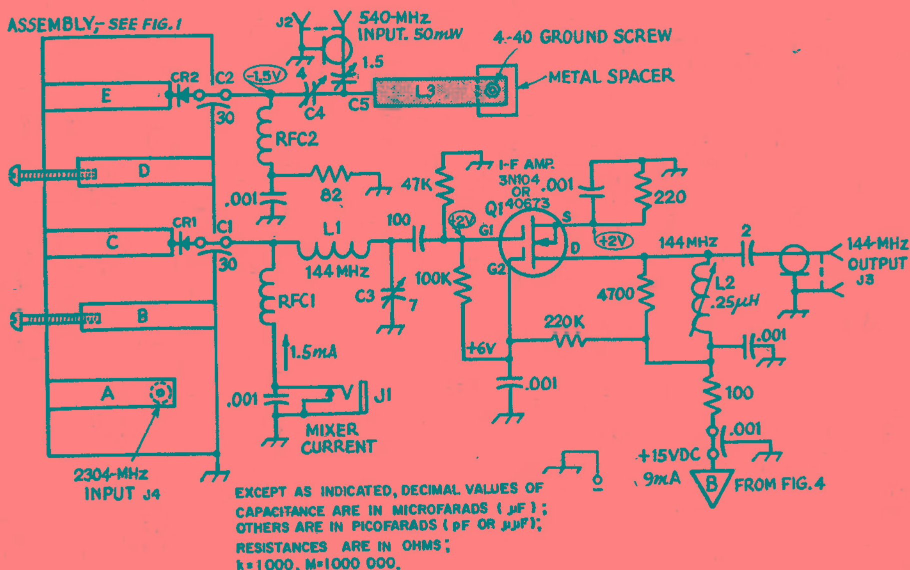
Fig. 3 and 4 show the schematic diagrams of the 2304-MHz converter and multiplier. The mixer and i-f preamplifier was built on a separate chassis since, at the time of their construction, a multiplier chain from another project was available. An i-f of 144 MHz was chosen although 50 MHz would work as well. An i-f output of 28 MHz, or lower, should not be used since this would result in undesirable interaction between the mixer and multiplier interdigital filters.

The 2304-MHz mixer and i-f amplifier section, shown in Fig. 3, is very similar to its 1296-MHz counterpart. Q1, the dual-gate MOSFET, operates at 144 MHz and thus has a noise figure about 1 dB higher than that obtainable at 28 MHz.

The multiplier chain, Fig. 4, has a separate oscillator for improved drive to the 2N3866 output stage. Otherwise the circuitry is similar to the 1296-MHz version.

Fig. 3 — Schematic diagram of the 2304-MHz version of the converter, with the i-f amplifier. The oscillator and multiplier circuits are constructed separately.

C1, C2 — 30-pF homemade capacitor. See text.
C3, C4, C5 — 0.8- to 10-pF glass trimmer, Johanson 2945 or equiv.
CR1 — Hewlett Packard 5082-2577 or 5082-2835.
CR2 — Hewlett Packard 5082-2811 or 5082-2835.
J1 — Closed-circuit jack.
J2, J3, J4 — Coaxial connector, chassis mount. Type BNC.
L1 — 5 turns No. 20 enem., 1/4-inch ID X 1/2-inch long. (6.35 X 12.7 mm).
L2 — 6 turns No. 24 enam., on 1/4-inch OD slug tuned form (0.25 μH).
L3 — Copper strip 1/2-inch wide X 2-11/16 inches (1.27 X 6.86 cm) long. See text and photographs.
RFC1 — Ohmite Z-144 or equiv.
RFC2 — Ohmite Z-460 or equiv.



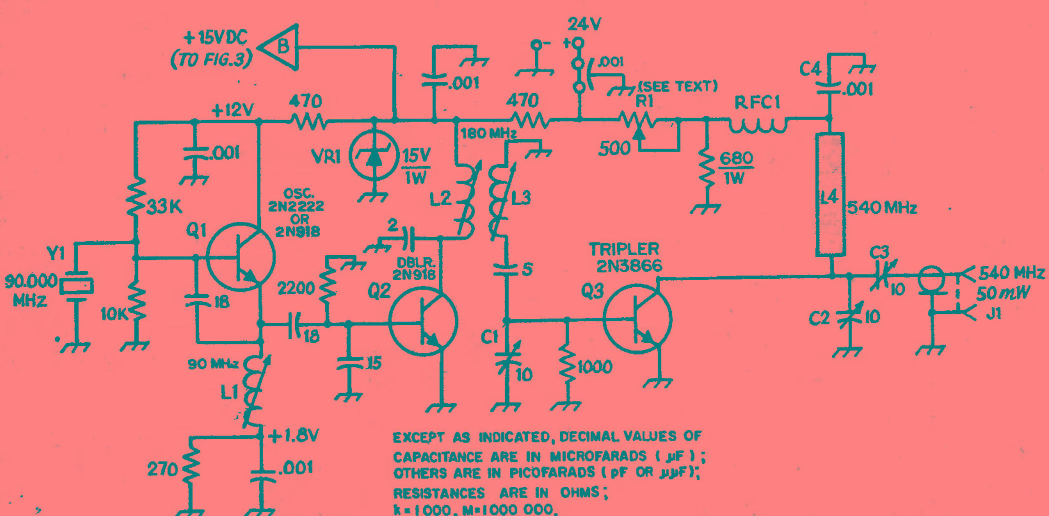


Fig. 4 — Schematic diagram of the oscillator and multipliar for the 2304-MHz converter. As explained in the text and shown in the photographs, a fixed-velue resistor may be substituted for R1 after the value that provides proper performance has been found.

C1, C2, C3 — 0.8- to 10-pF glass trimmer, Johanson 2945 or equiv.
C4 — .001-μF button mica.
J1 — Coaxiel connector, chassis mount, type BNC or equiv.
L1 — 10 turns No. 24 enam. on 1/4-inch OD slug-tuned form.
L2, L3 — 3 turns like L1.
L4 — Copper strip 1/2-inch wide X 1-1/2 inches (1.27 X 3.81 cm) long. Space 1/8 inch (3.18 mm) from chassis.
RFC1 — 10 turns No. 24 enam. 1/8-inch ID, closewound.

Mobile and Portable/Emergency Equipment and Practices

MOBILE AND PORTABLE EQUIPMENT

Amateur mobile and portable operation provides many opportunities for one to exercise his skill under less than ideal conditions. Additionally, the user of such equipment is available for public-service work when emergencies arise in his community – an important facet of amateur-radio operation. Operating skill must be better than that used at most fixed locations because the mobile/portable operator must utilize inferior antennas, and must work with low-power transmitters in many instances.

Most modern-day hf-band mobile work is done while using the ssb mode. Conversely, the fm mode is favored by mobile and portable vhf operators, though ssb is fully practical for vhf service. Some amateurs operate cw mobile, much to the consternation of local highway patrolmen, but cw operation from a *parked car* should not be overlooked during emergency operations.

High-power mobile operation has become practical on ssb because of the low duty cycle of voice operation, and because low-drain solid-state mobile power supplies lessen battery drain over that of dynamotors or vibrator packs. Most mobile a-m and fm operation is limited to 60 watts for reasons of battery drain.

Portable operation is popular on ssb, cw and fm while using battery-powered equipment. Ordinarily, the power of the transmitter is limited to less than five-watts dc input for practical reasons. Solid-state equipment is the choice of most modern amateurs because of its compactness, reliability, and low power consumption. High-power portable operation is practical and desirable when a gasoline-powered ac generator is employed.

The secret of successful operation from portable sites is much the same as that from a fixed station – a good antenna, properly installed. Power levels as low as 0.5 watt are sufficient for covering thousands of miles during hf-band ssb and cw operation. In the vhf and uhf region of operation it is common to work distances in excess of 100 miles – line of sight – with less than one watt of transmitter output power. Of course it is important to select a high, clear location for such operation on vhf, and it is beneficial to use an antenna with as much gain as is practical. Low-noise receiving equipment is the ever-constant companion of any low-power portable transmitter that provides successful long-distance communications. Careful matching of the portable or mobile antenna to obtain the lowest possible SWR is another secret of the successful operator.

All portable and mobile equipment should be assembled with more than ordinary care, assuring that maximum reliability under rough-and-tumble conditions will prevail. All solder joints should be made well, stranded hookup wire should be used for cabling (and in any part of the equipment subjected to stress). The cabinets for such gear should be rugged, and should be capable of protecting the components from dust, dirt, and moisture.

ELECTRICAL-NOISE ELIMINATION

One of the most significant deterrents to effective signal reception during mobile or portable operation is electrical impulse noise from the automotive ignition system. The problem also arises during the use of gasoline-powered portable ac generators. This form of interference can completely mask a weak signal, thus rendering the station ineffective. Most electrical noise can be eliminated by taking logical steps toward suppress-

Fig. 10-1 – Effective portable operation can be realized when using lofty locations for vhf or uhf. Here, W1CKK is shown operating a battery-powered, 150-mW output, 2-meter transceiver. With only a quarter wavelength antenna it is possibla to communicate with stations 25 miles or more away. Low-power transistor equipment like this unit will operate many hours from a dry-cell battery pack.

Fig. 10-2 — High-power portable/emergency operation can be made possible on all amateur bands by using vacuum-tube transmitters, and powering them from a gasoline-operated ac generator of one or more kW rating. (Shown here is VE7ARV/7 during a Field Day operation.)

ing it. The first step is to clean up the *noise source* itself, then utilize the receiver's built-in noise-reducing circuit as a last measure to knock down any noise pulses from passing cars, or from other man-made sources.

Spark-Plug Noise

Spark-plug noise is perhaps the worst offender when it comes to ignition noise. There are three methods of eliminating this type of interference — resistive spark-plug suppressors, resistor spark plugs, or resistance-wire cabling. By installing Autolite resistor plugs a great deal of the noise can be stopped. Tests have proved, however, that suppressor cable between the plugs and the distributor, and between the distributor and ignition coil, is the most effective means of curing the problem. Distributed-resistance cable has an approximate resistance of 5000 ohms per foot, and consists of a carbon-impregnated sheath followed by a layer of insulation, then an outer covering of protective plastic sheathing. Some cars come equipped with suppressor cable. Those which do not can be so equipped in just a matter of minutes. Automotive supply stores sell the cable, and it is not expensive. It is recommended that this wiring be used on all mobile units. The same type of cable can be installed on gasoline-powered generators for field use. A further step in eliminating plug noise is the addition of shielding over each spark-plug wire, and over the coil lead. It should be remembered that each ignition cable is an antenna by itself, thus radiating those impulses passing through it. By fitting each spark-plug and coil lead with the shield braid from a piece of RG-59/U coax line, grounding the braid at each end to the engine block, the noise reduction will be even greater. An additional step is to encase the distributor in flashing copper, grounding the copper to the engine block. This copper is quite soft and can be form-fit to the contour of the distributor. (Commercially-manufactured shielded ignition cable kits are also available.) The shield braid of the spark-plug wires should be soldered to the distributor shield if one is used. Also, the ignition coil should be enclosed in a metal shield since the top end of many of these coils is made of plastic. A small tin can can often be used as a top cover for the coil or distributor. It should be soldered to the existing metal housing of the coil. Additional reduction in spark-plug noise can be effected by making certain that the engine hood makes *positive* contact with the frame of the car when it is closed, thus offering an additional shield over the ignition system The engine block should also be bonded to the frame at several points. This can be done with the shield braid from coax cable. Feedthrough (hi-pass) capacitors should be mounted on the coil shield as shown in Fig. 10-6 to filter the two small leads leaving the assembly.

Other Electrical Noise

The automotive generator system can create an annoying type of interference which manifests itself as a "whine" when heard in the receiver. This noise results from the brushes sparking as the commutator passes over them. A dirty commutator is frequently the cause of excessive sparking, and can be cleaned up by polishing its surface with a fine grade of emery cloth. The commutator grooves should be cleaned out with a small, pointed instrument. A coaxial feedthrough capacitor of 0.1- to 0.5-μF capacitance should be mounted on the generator frame and used to filter the generator *armature* lead. In stubborn cases of generator noise a parallel *L/C* tuned trap can be used in place of the capacitor, or in addition to it, tuned to the receiver's operating frequency. This is probably the most effective measure used for curing generator noise.

Fig. 10-3 — A typical homemade shielding kit for an automotive ignition system. Tin cans have been put to use as shields for the spark coil and distributor. Additional shields have been mounted on the plug ends of the wires for shilding the spark plugs. The shield braid of the cabling protrudes at each end of the wires and is grounded to the engine block.

Fig. 10-4 — A close-up view of the distributor shield can. The shield braid over each spark-plug wire is soldered to the top of the can, and the can is grounded to the engine block.

Voltage regulators are another cause of mobile interference. They contain relay contacts that jitter open and closed when the battery is fully charged. The noise shows up in the receiver as a ragged, "hashy" sound. Coaxial feedthrough capacitors can be mounted at the *battery* and *armature* terminals of the regulator box to filter those leads. The *field* terminal should have a small capacitor and resistor, series-connected, from it to chassis ground. The resistor prevents the regulator from commanding the generator to charge constantly in the event the bypass capacitor short-circuits. Such a condition would destroy the generator by causing overheating.

Alternators should be suppressed in a similar manner as dc generators. Their slip rings should be kept clean to minimize noise. Make sure the brushes are making good contact inside the unit. A coaxial feedthrough capacitor and/or tuned trap should be connected to the output terminal of the alternator. Make certain that the capacitor is rated to handle the output current in the line. The same rule applies to dc generators. *Do not connect a capacitor to the alternator or generator field terminals.* Capacitor values as high as 0.5 μF are suitable for alternator filtering.

Some alternator regulator boxes contain solid-state circuits, while others use single or double contact relays. The single-contact units require a coaxial capacitor at the *ignition* terminal. The double-contact variety should have a second such capacitor at the *battery* terminal. If noise still persists, try shielding the field wire between the regulator and the generator or alternator. Ground the shield at both ends.

Instrument Noise

Some automotive instruments are capable of creating noise. Among these gauges and senders are the heat- and fuel-level indicators. Ordinarily, the addition of a 0.5-μF coaxial capacitor at the sender element will cure the problem.

Other noise-gathering accessories are turn signals, window-opener motors, heating-fan motors and electric windshield-wiper motors. The installation of a 0.25-μF capacitor will usually eliminate their interference noise.

Frame and Body Bonding

Sections of the automobile frame and body that come in contact with one another can create additional noise. Suspected areas should be bonded together with flexible leads such as those made from the shield braid of RG-8/U coaxial cable. Trouble areas to be bonded are:

1 – Engine to frame.
2 – Air cleaner to engine block.
3 – Exhaust lines to car fame and engine block.
4 – Battery ground terminal to frame.
5 – Steering column to frame.
6 – Hood to car body.
7 – Front and rear bumpers to frame.
8 – Tail pipe to frame.
9 – Trunk lid to frame.

Wheel and Tire Static

Wheel noise produces a ragged sounding pulse in the mobile receiver. This condition can be cured by installing static-collector springs between the spindle bolt of the wheel and the grease-retainer cap. Insert springs of this kind are available at automotive supply stores.

Tire static has a ragged sound too, and can be detected when driving on hard-surface highways. If the noise does not appear when driving on dirt roads it will be a sure indication that tire static exists. This problem can be resolved by putting

Fig. 10-5 — Gasoline-powered ac generators used for portable/emergency operation should be treated for ignition noise in the same manner as automobile engines are. The frame of the gas generator should be connected to an earth ground, and the entire unit should be situated as far from the operating position as possible. This will not only reduce ignition noise, but will minimize ambient noisa from the power unit. (Shown here is K1GTK during Field Day operations.)

Fig. 10-6 — The automobile ignition coil should be shielded as shown here. A small tin can has been soldered to the metal coil case, and coaxial feed through capacitors have been soldered to the top of the can. The "hot" lead of the coil enters the shield can through a modified audio connector.

antistatic powder inside each tire. This substance is available at auto stores, and comes supplied with an injector tool and instructions.

Corona-Discharge Noise

Some mobile antennas are prone to corona build-up and discharge. Whip antennas which come to a sharp point will sometimes create this kind of noise. This is why most mobile whips have steel or plastic balls at their tips. But, regardless of the structure of the mobile antenna, corona build-up will frequently occur during or just before a severe electrical storm. The symptoms are a high-pitched "screaming" noise in the mobile receiver, which comes in cycles of one or two minutes duration, then changes pitch and dies down as it discharges through the front end of the receiver. The condition will repeat itself as soon as the antenna system charges up again. There is no cure for this condition, but it is described here to show that it is not of origin within the electrical system of the automobile.

Electronic Noise Limiters

Many commercially built mobile transceivers have some type of built-in noise clipping or cancelling circuit. Those which do not can be modified to include such a circuit. The operator has a choice of using af or rf limiting. Circuits of this type are described in the theory section of the hf receiving chapter.

Simple superregenerative receivers, by nature of their operation, provide noise-limiting features, and no additional circuit is needed. Fm receivers, if operating properly, do not respond to noise pulses of normal amplitude; hence no additional circuitry is required.

THE MOBILE ANTENNA

The antenna is perhaps the most important item in the successful operation of the mobile installation. Mobile antennas, whether designed for single or multiband use, should be securely mounted to the automobile, as far from the engine compartment as possible (for reducing noise pickup), and should be carefully matched to the coaxial feed line which connects them to the transmitter and receiver. All antenna connections should be tight and weatherproof. Mobile loading coils should be protected from dirt, rain, and snow if they are to maintain their Q and resonant frequency. The greater the Q of the loading coil, the better the efficiency, but the narrower will be the bandwidth of the antenna system.

Though bumper-mounted mobile antennas are favored by some, it is better to place the antenna mount on the rear deck of the vehicle, near the rear window. This locates the antenna high and in the clear, assuring less detuning of the system when the antenna moves to and from the car body. *Never use a base-loaded antenna on a bumper mount* if an efficient system is desired. Many operators avoid cutting holes in the car body for fear of devaluation when selling the automobile. Such holes are easily filled, and few car dealers, if any, lower the trade-in price because of the holes.

The choice of base or center loading a mobile antenna has been a matter of controversy for many years. In theory, the center-loaded whip presents a slightly higher base impedance than does the base-loaded antenna. However, with proper impedance-matching techniques employed there is no discernible difference in performance between the two methods. A base-loading coil requires fewer turns of wire than one for center loading, and this is an electrical advantage because of reduced coil

Fig. 10-7 — Here a mobile station is used as a portable/emergency station. As such, it can be connected to a full-size stationary antenna for maximum effectiveness. The angine should be noise-suppressed, and should be kept running during operation of the station to assure full battery power. (WA3EQK operating.)

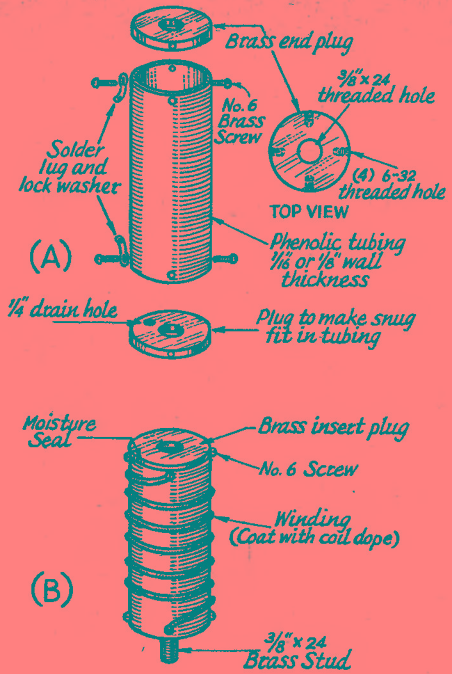


Fig. 10-8 – Details for making a home-built mobile loading coil. A breakdown view of the assembly is given at A. Brass end plugs are snug-fit into the ends of the phenolic tubing, end each is held in place by four 6-32 brass screws. Center holes in the plugs are drilled and tapped for 3/8-24 thread. The tubing can be eny diametar from one to four inches. The larger diemeters are recommended. Illustration B shows the completed coil. Resonence can be obtained by installing the coil, applying transmitter power, then pruning the turns until the lowest SWR is obtained. Pruning the coil for maximum field-strength-meter indication will elso serve as a resonance indication. The chart in Fig. 10-10 will serve as e guide in determining the number of turns required for a given frequency of operation.

losses. A base-loaded antenna is more stable from the standpoint of wind loading and sway. If a homemade antenna system is contemplated, either system will provide good results, but the base-loaded antenna may be preferred for its mechanical advantages.

Loading Coils

There are many commercially built antenna systems available for mobile operation, and some manufacturers sell the coils as separate units. Air-wound coils of large wire diameter are excellent for use as loading inductors. Large Miniductor coils can be installed on a solid phenolic rod and used as loading coils. Miniductors, because of their turns spacing, are easy to adjust when resonating the mobile antenna, and provide excellent circuit *Q*. Phenolic-impregnated paper or fabric tubing of large diameter is suitable for making homemade loading coils. It should be coated with liquid fiber glass, inside and out, to make it weather proof. Brass insert plugs can be installed in each end, their centers drilled and tapped for a 3/8 X 24 thread to accommodate the mobile antenna sections. After the coil winding is pruned to resonance it should be coated with a high-quality, low-loss compound to hold the turns securely in place, and to protect the coil from the weather. Liquid polystyrene is excellent for this. It can be made by dissolving chips of solid polystyrene in carbon-tetrachloride. *Caution*: Do not breathe the chemical fumes, and do not allow the liquid to come in contact with the skin. *Carbon tetrachloride is hazardous to health.* Dissolve sufficient polystyrene material in the liquid to make the remaining product the consistency of Q-dope or pancake syrup. Details for making a home-built loading coil are given in Fig. 10-8.

Impedance Matching

Fig. 10-9 illustrates the shunt-feed method of obtaining a match between the antenna and the coaxial feed line. For operation on 75 meters with a center-loaded whip, L2 will have approximately 18 turns of No. 14 wire, spaced one wire thickness between turns, and wound on a 1-inch diameter form. Initially, the tap will be approximately 5 turns above the ground end of L2. Coil L2 can be inside the car body, at the base of the antenna, or it can be located at the base of the whip, outside the car body. The latter method is preferred. Since L2 helps determine the resonance of the overall antenna, L1 should be tuned to resonance in the desired part of the band with L2 in the circuit. The

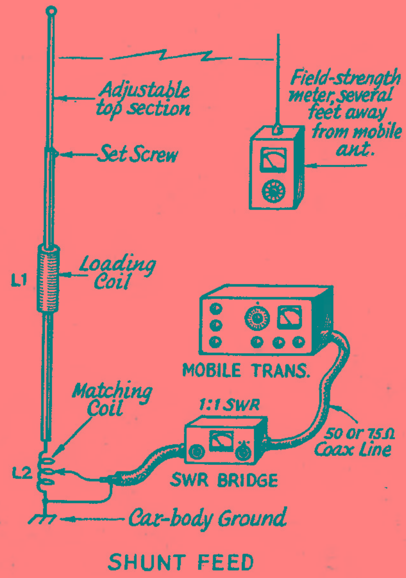


Fig. 10-9 – A mobile entenna using shunt-feed matching. Overall entanna resonance is determined by the combination of L1 and L2. Antenna resonance is set by pruning the turns of L1, or adjusting the top section of the whip, while observing the field-strength meter or SWR indicator. Then, adjust the tap on L2 for lowest SWR.

Approximate Values for 8-foot Mobile Whip

Base Loading

f kHz	Loading LμH	R_C(Q50) Ohms	R_C(Q300) Ohms	R_R Ohms	Feed R* Ohms	Matching LμH*
1800	345	77	13	0.1	23	3
3800	77	37	6.1	0.35	16	1.2
7200	20	18	3	1.35	15	0.6
14,200	4.5	7.7	1.3	5.7	12	0.28
21,250	1.25	3.4	0.5	14.8	16	0.28
29,000	—	—	—	—	36	0.23

Center Loading

f kHz	Loading LμH	R_C(Q50) Ohms	R_C(Q300) Ohms	R_R Ohms	Feed R* Ohms	Matching LμH*
1800	700	158	23	0.2	34	3.7
3800	150	72	12	0.8	22	1.4
7200	40	36	6	3	19	0.7
14,200	8.6	15	2.5	11	19	0.35
21,250	2.5	6.6	1.1	27	29	0.29

R_C = Loading-coil resistance; R_R = Radiation resistance.
* Assuming loading coil Q = 300, and including estimated ground-loss resistance.
Suggested coil dimensions for the required loading inductance are shown in a following table.

Fig. 10-10 — Chart showing inductance values used as a starting point for winding homemade loading coils. Values are based on an approximata base-loaded whip capacitance of 25 pF, and a capacitance of 12 pF for center-loaded whips. Large-diameter wire and coils, plus low-loss coil forms, are recommended for best Q.

TABLE 10-I

Suggested Loading-Coil Dimensions

Req'd L μH	Turns	Wire Size	Dia In.	Length In.
700	190	22	3	10
345	135	18	3	10
150	100	16	2 1/2	10
77	75	14	2 1/2	10
77	29	12	5	4 1/4
40	28	16	2 1/2	2
40	34	12	2 1/2	4 1/4
20	17	16	2 1/2	1 1/4
20	22	12	2 1/2	2 3/4
8.6	16	14	2	2
8.6	15	12	2 1/2	3
4.5	10	14	2	1 1/4
4.5	12	12	2 1/2	4
2.5	8	12	2	2
2.5	8	6	2 3/8	4 1/2
1.25	6	12	1 3/4	2
1.25	6	6	2 3/8	4 1/2

adjustable top section of the whip can be telescoped until a maximum reading is noted on the field-strength meter. The tap is then adjusted on L2 for the lowest reflected-power reading on the SWR bridge. Repeat these two adjustments until no further increase in field strength can be obtained; this point should coincide with the lowest SWR. The number of turns needed for L2 will have to be determined experimentally for 40- and 20-meter operation. There will be proportionately fewer turns required.

MATCHING WITH AN *L* NETWORK

Any mobile antenna that has a feed-point impedance less than the characteristic impedance of the transmission line can be matched to the line by means of a simple L network, as shown in Fig. 10-11. The network is composed of C_M and L_M. The required values of C_M and L_M may be determined from the following:

$$C_M = \frac{\gamma\ R_A (R_0 - R_A) \times 10^9}{2\pi f\ kHz\ R_A R_0}\ \text{pF and}$$

$$L_M = \frac{\gamma\ R_A (R_0 - R_A) \times 10^3}{2\pi f\ kHz}\ \mu\text{H}$$

where R_A is the antenna feed-point impedance and R_0 is the characteristic impedance of the transmission line.

As an example, if the antenna impedance is 20 ohms and the line is 50-ohm coaxial cable, then at 4000 kHz,

$$C_M = \frac{\sqrt{20\,(50-20) \times 10^9}}{(6.28)\,(4000)\,(20)\,(50)}$$

$$= \frac{\sqrt{600 \times 10^4}}{(6.28)\,(4)\,(2)\,(5)}$$

$$= \frac{24.1}{251.2} \times 10^{4} = 974 \text{ pF}$$

$$L_M = \frac{\sqrt{20\,(50-20) \times 10^3}}{(6.28)\,(4000)}$$

$$= \frac{\sqrt{600}}{25.12} = \frac{24.5}{25.12} = 0.97\ \mu\text{H}$$

The chart of Fig. 10-12 shows the capacitive reactance of C_M, and the inductive reactance of L_M necessary to match various antenna impedances to 50-ohm coaxial cable.

In practice, L_M need not be a separate inductor. Its effect can be duplicated by adding an equivalent amount of inductance to the loading coil, regardless of whether the loading coil is at the base or at the center of the antenna.

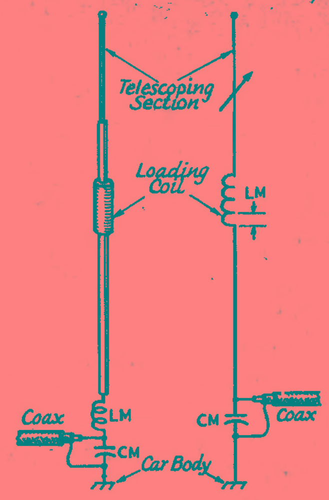


Fig. 10-11 — A whip antenna may also be matched to coax line by means of an L network. The inductive reactance of the L network can be combined in the loading coil, as indicated at the right.

Adjustment

In adjusting this system, at least part of C_M should be variable, the balance being made up of combinations of fixed mica capacitors in parallel as needed.

A small one-turn loop should be connected between C_M and the chassis of the car, and the loading coil should then be adjusted for resonance at the desired frequency as indicated by a GDO coupled to the loop at the base. Then the transmission line should be connected, and a check made with an SWR bridge connected at the transmitter end of the line.

With the line disconnected from the antenna again, C_M should be readjusted and the antenna returned to resonance by readjustment of the loading coil. The line should be connected again, and another check made with the SWR bridge. If the SWR is less than it was on the first trial, C_M should be readjusted in the same direction until the point of minimum SWR is found. Then the coupling between the line and the transmitter can be adjusted for proper loading. It will be noticed from Fig. 10-12 that the inductive reactance varies only slightly over the range of antenna resistances likely to be encountered in mobile work. Therefore, most of the necessary adjustment is in the capacitor.

The one-turn loop at the base should be removed at the conclusion of the adjustment and slight compensation made at the loading coil to maintain resonance.

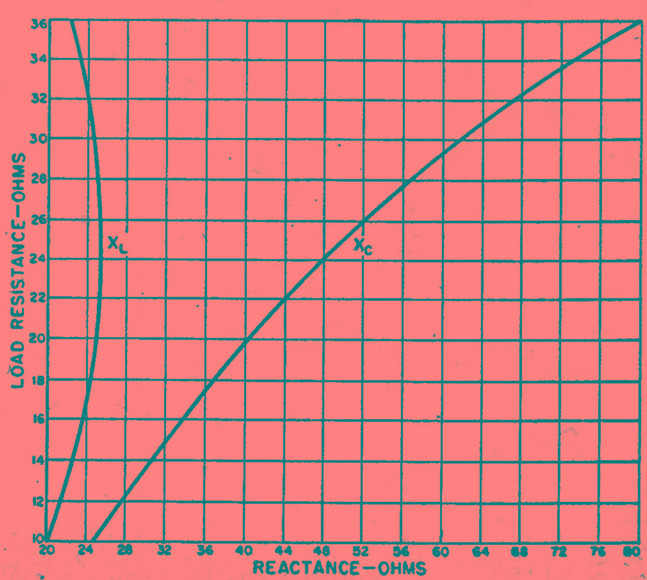


Fig. 10-12 — Curves showing inductive and capacitive reactances required to match a 50-ohm coax line to a variety of antenna resistances.

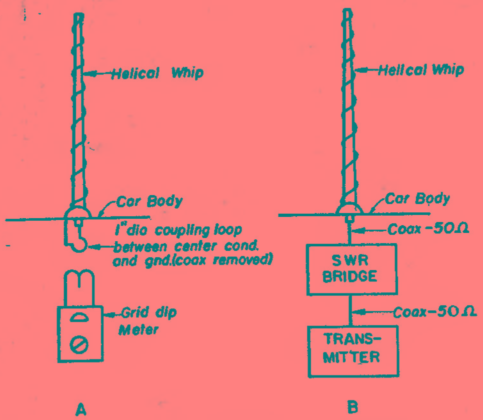


Fig. 10-13 – The resonant frequency of the antenna can be checked (A) with a grid-dip meter or (B) by finding the frequency at which minimum feed-line SWR occurs. The latter method is more accurate at high frequencies because it eliminates the effect of the coupling loop required in A.

CONTINUOUSLY-LOADED HELICAL WHIPS

A continuously-loaded whip antenna of the type shown in Fig. 10-14 is thought to be more efficient than a center- or base-loaded system (*QST*, May 1958, W9KNK). The feed-point impedance of the helically-wound whip is somewhat greater than the previously described mobile antennas, and is on the order of 20 ohms, thus providing an SWR of only 2.5 when 50-ohm coaxial feed line is used. The voltage and current distribution is more uniform than that of lumped-constant antennas. The low SWR and this feature make the antenna more efficient than the center- or base-loaded types. Antennas of this variety can be wound on a fiber glass fishing rod, then weatherproofed by coating them with liquid fiber glass, or by encapsulating them with shrinkable vinyl-plastic tubing.

Tapered Pitch

On frequencies below 28 MHz the radiation resistance falls off so rapidly that for the desired 4- and 6-foot whip lengths the resistance values are not suitable for direct operation with 50-ohm lines. It is desirable to raise the feed-point *R* to a value, approaching 50 ohms so that a matched line condition will exist. Based on extensive experimentation, a *tapered*-pitch continuous-loading antenna is recommended. Since it is not feasible to wind the helix with continuously varying pitch, a "step-tapered" design is best. A typical step-tapering technique for a variable-pitch helical whip antenna is to divide the total length of the radiator, say 4 feet, into 6 equal parts of 8 inches each. The helix is then wound with a 2-inch pitch for the first 8 inches, pitches of 1, 1/2, 1/4 and 1/8 inch, respectively, for the next four 8-inch sections, and finished with close winding of the final section. The resonant frequency will depend upon the rod diameter, wire size and number of turns. However, the variable-pitch 6-step taper approaches the ideal continuously-variable condition closely enough to give a good 50-ohm match with a 4-foot antenna at frequencies between 20 and 30 MHz.

Adjustment

With this design it is difficult to adjust the resonant frequency by changing the turns near the base; however, the frequency may be adjusted very readily by cutting off sections of the tightly-wound portion near the top of the whip. The technique to follow is to design for a frequency slightly lower than desired and then to bring the unit in on frequency by cutting small sections off the top until it resonates at the desired frequency. Resonance can be checked either by the use of a grid-dip meter or by the use of a transmitter and SWR bridge. Reflected power as low as 2 to 5 percent can easily be obtained with the units properly resonated even though it may mean cutting an inch or two off the top closely wound section to bring the unit in on frequency. These values can be obtained in the 10- and 15-meter band with overall lengths of 4 feet and in the 20- and 40-meter bands with a length of 6 feet. In the 75-meter band it has been possible to obtain an SWR of 1.5 using a 6-foot tapered-pitch helical winding, although the bandwidth is restricted to about 60 kHz. This affords operation comparable to the center coil loaded 12-foot whips. In general, the longer the radiator (in wavelengths), the greater the bandwidth. By arbitrarily restricting the physical length to 6 feet, or less, we obtain the following results:

Band	*Length*	*Resonant Freq.*	*SWR*	*Bandwidth for SWR = 2.0*
10 meters	4 feet	29.00 MHz	1,3	800 kHz
15 meters	4 feet	21.30 MHz	1.4	500 kHz
20 meters	6 feet	14.25 MHz	1.3	250 kHz
40 meters	6 feet	7.25 MHz	1.5	100 kHz
75 meters	6 feet	3.90 MHz	1.5	60 kHz

In the 15-, 20- and 40-meter bands the bandwidths of the taper-pitch designs are good enough to cover the entire phone portions of the bands. The bandwidths have been arbitrarily

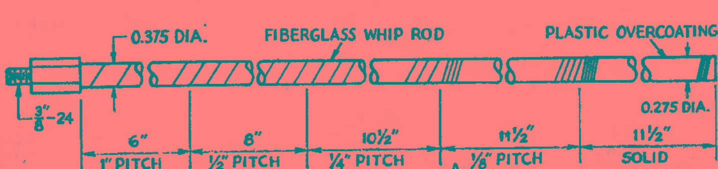


Fig. 10-14 – Dimensions for a 15-meter stepped-pitch whip, wound with No. 20 enameled wire.

Fig. 10-15 — K1MET prunes a capacity hat for antenna resonance at the low end of the 160-meter band. The Webster Big-K antenna is first tuned for the high segment of the band. The capacity hat is clipped on when operation on the "low end" is desired. Fine edjustments can be made by incraasing or decreasing the spacing between the two No. 10 wires.

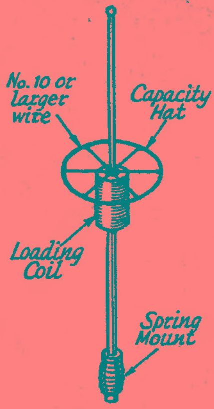


Fig. 10-16 — A capacitance "hat" can be used to improve the performance of base- or center-loaded whips. A solid metal disk can be used in place of the skeleton disk shown here.

selected as that frequency spread at which the SWR becomes 2 on a 50-ohm line, although with most equipment SWR values up to 2.5 can be tolerated and loading accomplished with ease.

Top-Loading Capacitance

Because the coil resistance varies with the inductance of the loading coil, the resistance can be reduced, beneficially, by reducing the number of turns on the coil. This can be done by adding capacitance to that portion of the mobile antenna that is *above* the loading coil. To achieve resonance, the inductance of the coil is reduced proportionally. Capacity "hats," as they are often called, can consist of a single stiff wire, two wires or more, or a disk made up from several wires, like the spokes of a wheel. A solid metal disk can also be used. The larger the capacity hat, in terms of mass, the greater the capacitance. The greater the capacitance, the smaller the amount of inductance needed in the loading coil for a given resonant frequency.

There are two schools of thought concerning the attributes of center-loading and base-loading. It has not been established that one system is superior to the other, especially in the lower part of the hf spectrum. For this reason both the base- and center-loading schemes are popular. Capacity-hat loading is applicable to either system. Since more inductance is required for center-loaded whips to make them resonant at a given frequency, capacity hats should be particularly useful in improving their efficiency.

REMOTE ANTENNA RESONATING

Fig. 10-17 shows circuits of two remote-control resonating systems for mobile antennas. As shown, they make use of surplus dc motors driving a loading coil removed from a surplus ARC-5 transmitter. A standard coil and motor may be used in either installation at increased expense.

The control circuit shown in Fig. 10-17A is a three-wire system (the car frame is the fourth conductor) with a double-pole double-throw switch and a momentary (normally off) single-pole single-throw switch. S2 is the motor reversing switch. The motor runs so long as S1 is closed.

The circuit shown in Fig. 10-17B uses a latching relay, in conjunction with microswitches, to reverse automatically the motor when the roller reaches the end of the coil. S3 and S5 operate the relay, K1, which reverses the motor. S4 is the

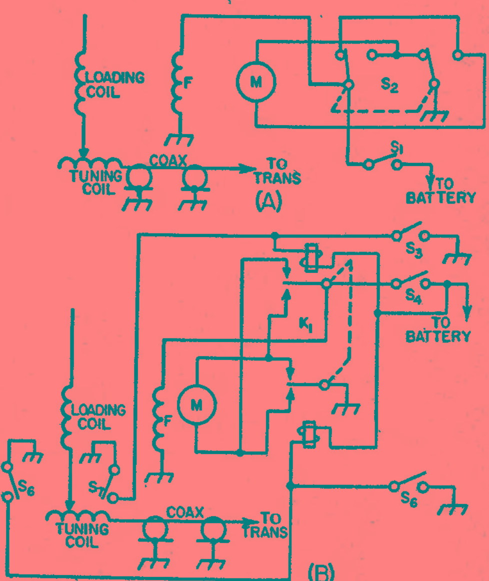


Fig. 10-17 — Circuit of the remote mobile-whip tuning systems.
K1 — Dpdt latching relay.
S1, S3, S4, S5 — Momentary-contact spst, normally open.
S2 — Dpdt toggle.
S6, S7 — Spst momentery-contact microswitch, normally open.

motor on-off switch. When the tuning coil roller reaches one end or the other of the coil, it closes S6 or S7, as the case may be, operating the relay and reversing the motor.

The procedure in setting up the system is to prune the center-loading coil to resonate the antenna on the highest frequency used without the base-loading coil. Then, the base-loading coil is used to resonate at the lower frequencies. When the circuit shown in Fig. 10-17A is used for control, S1 is used to start and stop the motor, and S2, set at the "up" or "down" position, will determine whether the resonant frequency is raised or lowered. In the circuit shown in Fig. 10-17B, S4 is used to control the motor. S3 or S5 is momentarily closed (to activate the latching relay) for raising or lowering the resonant frequency. The broadcast antenna is used with a wavemeter to indicate resonance. (Originally described in *QST*, December 1953.)

VHF MOBILE ANTENNAS

The three most popular vhf mobile antennas are the so-called halo, the turnstile, and the 1/4-wavelength vertical. The same rules apply to the installation and use of these antennas as for antennas operated in the hf bands – mounted as high and in the clear as possible, and with good electrical connections throughout the system.

The polarization chosen – vertical or horizontal – will depend upon the application and the area of the USA where operation will take place. It is best to use whatever polarity is in vogue for your region, thus making the mobile signal compatible with those of other mobiles or fixed stations. Vertically-polarized mobile antennas are more subject to pattern disturbance than horizontal types. That is to say, considerably more flutter will be inherent on the signal than with horizontal antennas. This is because such objects as trees and power poles, because of their vertical profile, tend to present a greater path obstacle to the vertical antenna. It is becoming common practice, however, to use omnidirectional, vertically-polarized vhf mobile antennas in connection with fm/repeater mobile service, even in areas where horizontal antennas are favored.

Both the turnstile and halo antennas are horizontally polarized. The halo is physically small, but is less effective than a turnstile. It is a half-wavelength dipole bent into a circle, and because the ends are in close proximity to one another, some signal cancellation occurs. This renders the antenna less efficient than a straight center-fed dipole. Halos do not offer a perfectly circular radiation pattern, though this has been a popular belief. Tests indicate that there is definite directivity, though broad, when a halo is rotated 360 degrees over a uniform plane surface.

Turnstile antennas of the type shown in Fig. 10-20 more closely approach the desired circular pattern of radiation, though the pattern is somewhat like a poorly defined four-leaf clover. Here two dipoles are fed with a 90-degree phase difference, and the antenna has a gain equal to, or better than a straight dipole. Of the three types discussed in the foregoing text, the latter is recommended.

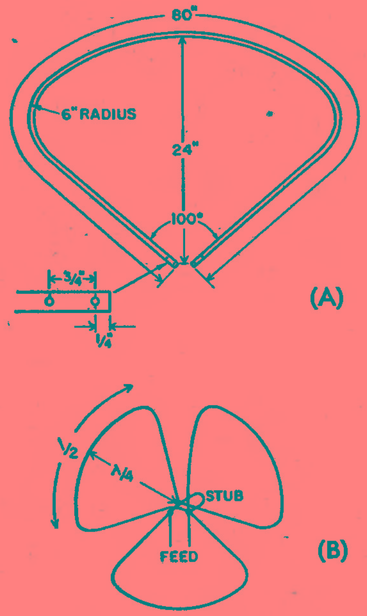


Fig. 10-19 – Schematic representation of the Big Wheel at B. Three one-wavelength elements are connected in parallel. The resulting low feed impedance is reised to 52 ohms with an inductive stub. Illustration A shows the bend details of one element for 144-MHz use.

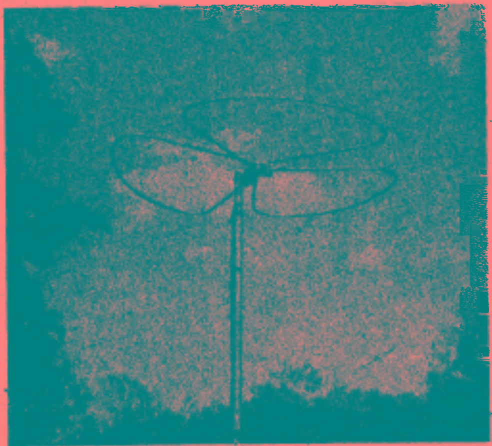


Fig. 10-18 – The Big Wheel, an omnidirectional horizontal antenna for the 144-MHz bend designed by W1FVY and W1IJD. Radiating elements occupy an area approximately 40 inches in diameter.

Fig. 10-20 — Two-meter turnstile antenna shown mounted on the front of an automobile. The miniature coax cable which feeds the antenna is taped to its 1/4-inch diameter steel supporting rod. Tha ends of the antenna elements should be flattened, or rounded, to meke them safer in the event of accidental contact with the human body.

If one does not object to having an antenna that is likely to become a conversation piece because of its size and shape, it would be well to consider using the "Big Wheel" antenna, designed by W1FVY and W1IJD (September and October *QST*, 1961, and *ARRL Radio Amateur's VHF Manual*). The "wheel" consists of three one-wavelength elements, Fig. 10-19, connected in parallel and arranged as a cloverleaf. The antenna has a low feed-point impedance which is raised to 50 ohms by means of an inductive stub. Each clover leaf is 80 inches long overall (144 MHz), and can be made from aluminum tubing. Though the radiation pattern is not perfectly circular, it offers a good approach to that goal. Its performance greatly surpasses that of the three previously described antennas. It showed an increase in signal strength, from a selected test site, of several dB over the vertical whip, the halo, and the turnstile. Polarization is horizontal, as was the fixed-station antenna used in the tests.

TWO-METER 5/8-WAVELENGTH VERTICAL

Probably the most popular antenna used by the fm group is the 5/8-wavelength vertical. As stated previously, this antenna has some gain when compared to a dipole. The antenna can be used in either a fixed location with radials or in a mobile installation. An inexpensive antenna of this type can be made from a modified CB whip. The

Fig. 1 — The new coil is tapped two turns from the base end. It may be necessary to file the coil ends so that the assembly will fit in the phenolic covering.

antenna shown in Figs. 1 and 2 is a 5/8-wavelength, 2-meter whip.

There are a number of different types of CB mobile antennas available. This particular antenna to be modified consists of a clamp-on trunk mount, a base loading coil, and a 39-inch spring-mounted, stainless-steel whip.

The modification consists of removing the loading-coil inductance, winding a new coil, and mounting a 3-30 pF trimmer in the bottom housing. The capacitor is used for obtaining a precise match in conjunction with the base coil tap.

The first step is to remove the weatherproof phenolic covering from the coil. Remove the base housing and clamp the whip side of the antenna in a vise. Insert a knife blade between the edge of the whip base and the phenolic covering. Gently tap the knife edge with a hammer to force the housing away from the whip section.

Next, remove the coil turns and wind a new coil using No. 12 wire. The new coil should have nine turns, equally spaced. The tap point is two turns up from the base (ground) end on the antenna as

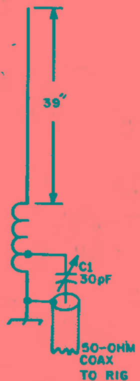
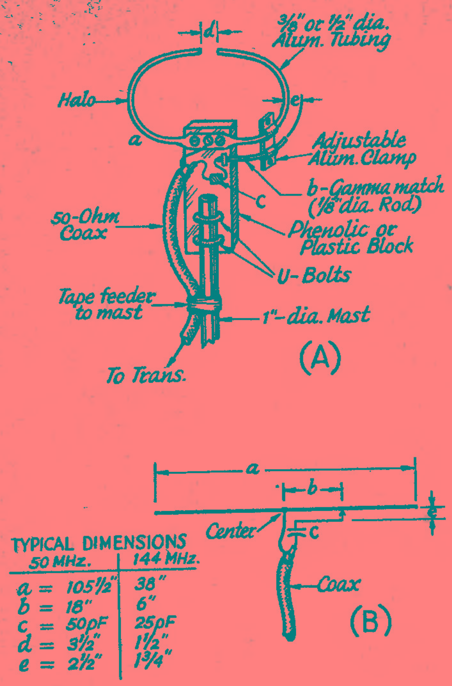


Fig. 2 — Circuit diagram of the whip antenna. C1 is a 3- to 30-pF trimmer.



TYPICAL DIMENSIONS

	50 MHz.	144 MHz.
a =	105½"	38"
b =	18"	6"
c =	50pF	25pF
d =	3½"	1½"
e =	2½"	1¾"

Fig. 10-22 — Details for building a halo antenna for 6- or 2-meter use are shown at A. Other mechanical methods are possible, and the construction technique used will be up to the builder. The open end of the coax cable should be sealed against the weather. At B, a schematic representation of the halo. Dimension *a* is set for 1/2 wavelength at the operating frequency. The chart gives approximate dimensions in inches, and will serve as a guide in building a halo.

modified. The trimmer capacitor is mounted on a terminal strip which is installed in the base housing. A hole must be drilled in the housing to allow access to the capacitor adjustment screw.

Initially, the tap on the coil was tried at three turns from the bottom. The antenna was mounted on the car, an SWR indicator was inserted in the feed line, and C1 and the whip height were adjusted for a match. A match was obtained, but when the phenolic sleeve was placed over the coil, it was impossible to obtain an adjustment that proved a match. Apparently the dielectric material used in the coil cover has an effect on the coil. After some experimenting it was found that with the tap two turns up from the bottom, and with the cover over the coil, it was possible to get a good match with 50-ohm line.

This antenna can be used in a fixed location by adding radials. The radials, three or four, should be slightly longer than 1/4-wave and should be attached to the base mounting section.

THE QUARTER-WAVELENGTH VERTICAL

Ideally, the vhf vertical antenna should be installed over a perfectly flat plane reflector to assure uniform omnidirectional radiation. This suggests that the center of the automobile roof is the best place to mount it. Alternatively, the flat portion of the auto's rear trunk deck can be used, but will result in a directional pattern because of car-body obstruction. Fig. 10-23 illustrates at A and B how a Millen high-voltage connector can be used as a roof mount for a 144-MHz whip. The hole in the roof can be made over the dome light, thus providing accessability through the upholstery. RG-59/U and matching section *L*, Fig. 10-23C, can be routed between the car roof and the ceiling upholstery and brought into the trunk compartment, or down to the dashboard of the car. Some operators install an SO-239-type coax connector on the roof for mounting the whip. The method is similar to that of drawing A.

VHF HALO ANTENNAS

The antenna of Fig. 10-22 can be built from aluminum tubing of medium tensile strength. The one-half-wavelength dipole is bent into a circle and fed with a gamma match. Capacitor *c* is shown as a fixed value, but a variable capacitor mounted in a weatherproof box will afford more precise

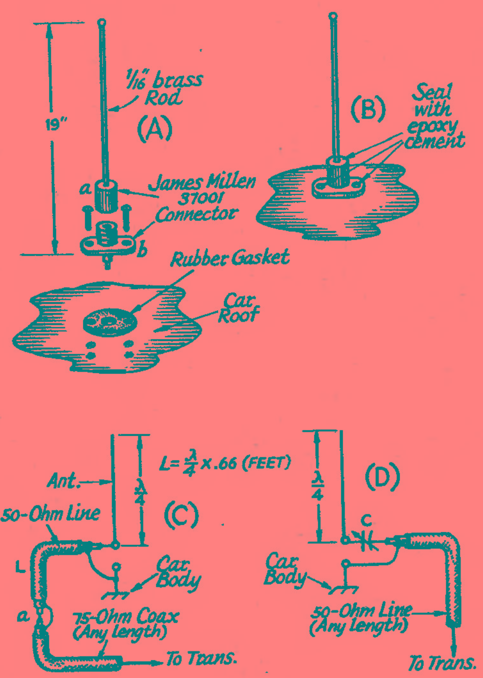
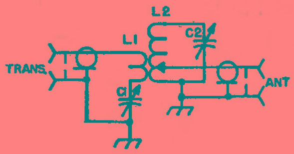


Fig. 10-23 — At A and B, an illustration of how a quarter-wavelength vertical antenna can be mounted on a car roof. The whip section should be soldered into the cap portion of the Millen connector, then screwed to the base socket. This handy arrangement permits removing the antenna when desired. Epoxy cement should be used et the two mounting screws to prevent moisture from entering the car. Diagrams C and D are discussed in the text.



50 MHz	144 MHz
C1=100pF	35pF
C2=25pF	15pF
L1= 2 Ts NO.16 ENAM., 2-1/2" DIA.	2 Ts NO.16 ENAM., 1-1/2" DIA.
L2=5 Ts NO.10 1-1/2" DIA.	4 Ts NO. 10 1" DIA.

Fig. 10-24 — Schematic diagram of the 6- or 2-meter antenna-matching circuit for use at the base of the quarter-wavelength vertical antenna. It can be housed in a Minibox and mounted permanently at the antenna base, inside or outside the car. If used outside, it should be sealed against dirt and moisture.

adjustment of the SWR. Or, a variable capacitor can be used initially for obtaining a 1:1 match, then its value can be measured at that setting to determine the required value for fixed capacitor *c*. Fixed-value capacitor *c* should be a dipped silver mica. A 75-pF variable should be used for 6-meter antennas, and a 35-pF variable will suffice for 144 MHz.

The tubing of *a* can be flattened to provide a suitable mounting surface for attachment to the insulating block of Fig. 10-22A. Gamma rod *b* can be secured to the same block by flattening its end and bolting it in place with 4-40 brass hardware. The spacing at *d* can be varied during final adjustment to secure the lowest SWR. Better physical stability will result if a high-dielectric insulator is connected across area *d*. Steatite material is recommended if an insulator/stabilizer is used.

If 75-ohm transmission line is used for the vertical, a quarter-wavelength matching transformer, L, can be used to match the feed impedance of the whip – approximately 30 ohms – to that of the feed line. A section of 50-ohm coax inserted as shown provides a close match to the antenna. Coax fittings can be used at junction *a* to assure a flat line, and to provide mechanical flexibility. BNC connectors are ideal for use with small coax lines. Illustration D shows how a series capacitor can be used to tune the reactance out of the antenna when using 50-ohm feed line. For 144-MHz use it should be 35 pF. A 75-pF variable will suffice for 6-meter antennas. An SWR bridge should be connected in the line while *c* is tuned for minimum reflected-power indication.

A more precise method of matching the line to the antenna is shown in Fig. 10-24. This antenna coupler can match 50- or 75-ohm lines to any antenna impedance from 20 ohms to several hundred ohms. It should be installed *at* the base of the vertical, and with an SWR bridge in the line C1 and C2 should be adjusted for the lowest SWR possible. The tap near the ground end of L2 should then be adjusted for the lowest SWR, readjusting C1 and C2 for minimum reflected power each time the tap is moved. A very compact tuner can be built by scaling down the coil dimensions appropriately. Trimmer capacitors can be used for C1 and C2 if power levels of less than 50 watts are used.

MOBILE POWER SUPPLIES

Most modern-day mobile installations utilize commercially-built equipment. This usually takes the form of a transceiver for ssb on the hf bands, and ssb or a-m for vhf operation. For fm operation in the vhf bands, most transceivers are surplus units which were originally used by commercial land-mobile services. Some home-built equipment is still being used, and it is highly recommended that one consider building his own mobile installation for the technical experience and satisfaction such a project can afford.

Many mobile transceivers contain their own power supplies for 6- and 12-volt dc operation. Some internal power supplies will also work off the 117-V mains. Vibrator power supplies are quite popular for low and medium power levels, but solid-state supplies are more reliable and efficient. Dynamotors are still used by some operators, but are bulky, noisy, and inefficient. The latter imposes an extremely heavy drain on the car battery, and does not contribute to long-term mobile or emergency operation without having the engine running at fairly high rpm to maintain the charge level of the battery.

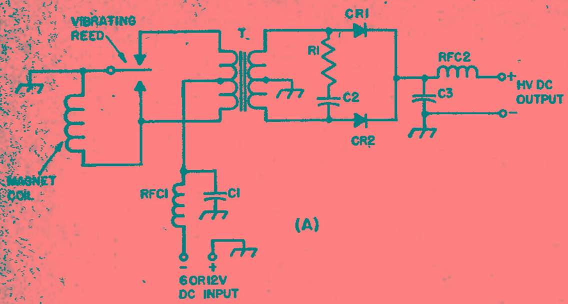
Dynamotors

A dynamotor differs from a motor generator in that it is a single unit having a double armature winding. One winding serves for the driving motor, while the output voltage is taken from the other. Dynamotors usually are operated from 6-, 12-, 28- or 32-volt storage batteries and deliver from 300 to 1000 volts or more at various current ratings.

Commutator noise is a common cause of poor reception when dynamotors are used. It can usually be cured by installing .002-μF mica bypass capacitors from the dynamotor brushes (high-voltage end of armature) to the frame of the unit, preferably inside the cover. The high-voltage output lead from the dynamotor should be filtered by placing a .01-μF capacitor in shunt with the line (a 1000-V disk), followed by a 2.5-mH rf choke (in series with the line) of adequate current rating for the transmitter or receiver being powered by the dynamotor. This network should be followed by a smoothing filter consisting of two 8-μF electrolytic capacitors and a 15- or 30-H choke having a low dc resistance. The commutator and its grooves, at both ends of the armature, should be kept clean to further minimize noise. Heavy, direct leads should be used for connecting the dynamotor to the storage battery.

Vibrator Power Supplies

The vibrator type of power supply consists of a special step-up transformer combined with a



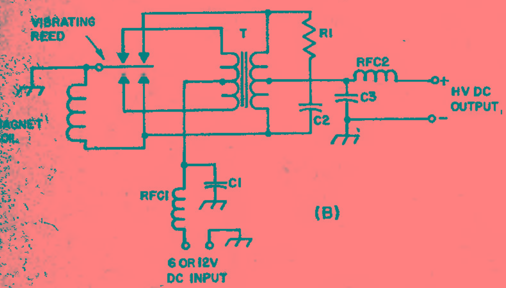


Fig. 10-25 — Basic types of vibrator power supplies. A — Nonsynchronous. B — Synchronous.

vibrating interrupter (vibrator). When the unit is connected to a storage battery, plate power is obtained by passing current from the battery through the primary of the transformer. The circuit is made and reversed rapidly by the vibrator contacts, interrupting the current at regular intervals to give a changing magnetic field which induces a voltage in the secondary. The resulting square-wave dc pulses in the primary of the transformer cause an alternating voltage to be developed in the secondary. This high-voltage ac in turn is rectified, either by silicon diode rectifiers or by an additional synchronized pair of vibrator contacts. The rectified output is pulsating dc, which may be filtered by ordinary means. The smoothing filter can be a single-section affair, but the output capacitance should be fairly large – 16 to 32 μF.

Fig. 10-25 shows the two types of circuits. At A is shown the nonsynchronous type of vibrator. When the battery is disconnected the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with one contact point, causing current to flow through the lower half of the transformer primary winding. Simultaneously, the magnet coil is short-circuited, de-energizing it, and the reed swings back. Inertia carries the reed into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil again is energized, and the cycle repeats itself.

The synchronous circuit of Fig. 10-25B is provided with an extra pair of contacts which rectifies the secondary output of the transformer, thus eliminating the need for a separate rectifier tube. The secondary center tap furnishes the positive output terminal when the relative polarities of primary and secondary windings are correct. The proper connections may be determined by experiment.

The buffer capacitor, C2, across the secondary of T, absorbs spikes that occur on breaking the current, when the magnetic field collapses almost instantly and hence causes high voltages to be induced in the secondary. Without C2 excessive sparking occurs at the vibrator contacts, shortening the vibrator life. Resistor R1 is part of the buffer and serves as a fuse if C2 should short out, thus protecting the vibrator and transformer from damage. Values between 1000 and 5600 ohms, 1 watt, are commonly used. Correct values for C2 lie between .005 and .03 μF, and for 220-350-V supplies the capacitor should be rated at 2000 V or better, dc. The exact capacitance is critical, and should be determined experimentally while observing the output waveform on an oscilloscope for the least noise output. Alternatively, though not as effective a method, the capacitor can be selected for least sparking at the vibrator contacts.

Vibrator-transformer units are available in a variety of power and voltage ratings. Representative units vary from one delivering 125 to 200 volts at 100 mA to others that have a 400-volt output rating at 150 mA. Most units come supplied with "hash" filters, but not all of them have built-in ripple filters. The requirements for ripple filters are similar to those for ac supplies. The usual efficiency of vibrator packs is in the vicinity of 70 percent, so a 300-volt 200-mA unit will draw approximately 15 amperes from a 6-volt storage battery. Special vibrator transformers are also available from transformer manufacturers so that the amateur may build his own supply if he so desires. These have dc output ratings varying from 150 volts at 40 mA to 330 volts at 135 mA.

"Hash" Elimination

Sparking at the vibrator contacts causes rf interference ("hash," which can be distinguished from hum by its harsh, sharper pitch) when used with a receiver. To minimize this, rf filters are incorporated, consisting of RFC1 and C1 in the battery circuit, and RFC2 with C3 in the dc output circuit.

Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough rf to cause interference in a sensitive amateur receiver.

TRANSISTORIZED POWER SUPPLIES

Most present-day mobile equipment is powered by solid-state dc-to-dc converters. They are somewhat similar to vibrator supplies in that they use power transistors to switch the primary voltage of the transformer. This technique eliminates sparking in the switching circuit, and offers greater reliability and efficiency. The switching transistors can be made to oscillate, by means of a feedback winding on the transformer, and by application of forward bias on the bases of the switching transistors. The switching rate can be set for any frequency between 50 Hz and several thousand Hz and depends to a great extent upon the inductance of the transformer windings. The switching waveform is a square wave. Therefore, the supply is capable of causing a buzzing sound in transmitter or receiver output in much the same fashion as with a vibrator supply. Rf filtering should be employed as a corrective measure. At higher switching rates the buzz becomes a whine which sounds like that from a dynamotor. High-frequency switching rates are preferred for dc-to-dc converters because smaller transformer cores can be used, and because less output filtering is required. The efficiency of a well-designed solid-state power suuply is on the order of 80 percent, an improvement over the usual 60 to 70 percent of vibrator supplies, or the miserable 30 to 40 percent of dynamotors.

A typical transistorized supply is shown in Fig. 10-26. The supply voltage is fed into the emitter circuit of Q1-Q2. A resistive divider is used to obtain forward bias for the transistors through base-feedback-winding 1. The primary switching takes place between the emitter and collector of each transistor. Q1 and Q2 are connected in push-pull and conduct on alternate half cycles. As each transistor is driven into conduction it saturates, thus forming a closed contact in that leg of the circuit. The induced voltage is stepped up by T, and high-voltage appears across winding 3. Zener diodes CR1 and CR2 protect Q1 and Q2 from voltage spikes. They should be rated at a voltage slightly lower than the Vce of the transistors. Diodes CR3 through CR6 form a bridge rectifier to provide dc output from winding 3. Some supplies operate at a switching rate of 2000 to 3000 Hz. It is possible to operate such units without using output rectifiers, but good filtering is needed to remove the ripple from the dc output.

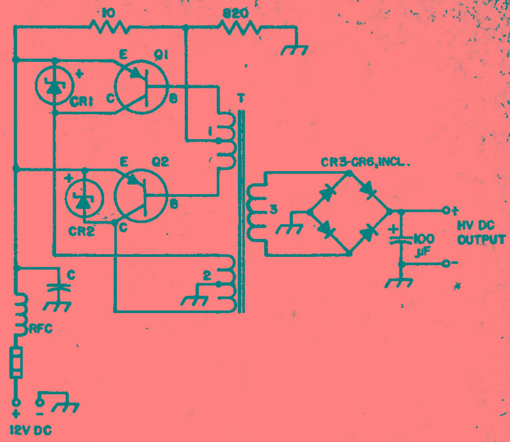


Fig. 10-26 — Typical dc-to-dc converter. Ratings for CR3-CR6, and the 100-µF filter capacitor can be selected from data in the power-supply chapter.

Transistor Selection

The switching transistors should be able to handle the primary current of the transformer. Since the feedback will diminish as the secondary load is increased, the beta of the transistors, plus the design of the feedback circuit, must be sufficient to sustain oscillation under full-load conditions. During no-load conditions, the feedback voltage will reach its highest peak at the bases of Q1 and Q2. Therefore, the transistors must be rated for whatever base-emitter reverse voltage that occurs during the cutoff period. Since the transistors must be able to handle whatever peak voltage occurs during the switching process, it is wise to stay on the safe side. Choose transistors that have a Vceo rating of three or four times the supply voltage, keeping in mind that fully charged automobile batteries can deliver as much as 14 volts. Heat sinks should be used on Q1 and Q2 to prevent damage from excessive heating. The larger the heat sink, the better. Under full-load conditions the transistors should only be slightly warm to the touch. If they are running hot, this will indicate inadequate heat sinking, too great a secondary load, or too much feedback. *Use only enough feedback to sustain oscillation under full loading*, and to assure rapid starting under the same conditions.

MOBILE POWER SUPPLY FOR TRANSCEIVERS

Transceivers, such as the Heath SB-102, and the Drake TR-4 require a separate power supply when operated from 12-volts dc. Additionally, linear amplifiers can be run from a separate dc supply to allow increased power operation from relatively low-power transceivers. The unit described here, when operated from 12-volts dc, will deliver approximately 900-volts dc at 300 mA, 250-volts dc at 200 mA, negative 150-volts dc at 40 mA, and an adjustable bias voltage from 10 to 150 volts of dc.

The Circuit

A common-emitter configuration is used with diodes to provide a return path for the feedback winding, as shown in Fig. 10-28. Assuming that Q2 conducts first, the base is driven negative by the feedback winding (connections 6 and 7 on T1). CR15 then conducts, thereby protecting the base of Q1. CR14 is back-biased to an open circuit when Q2 is conducting. When T1 saturates producing a square wave, the voltage at pins 6 and 7 of T1 reverses turning on Q1. When Q2 conducts, current flows through the primary of T2 in one direction and as Q1 conducts, current flows through the primary of T2 in the other direction. This reversal of current in the primary of T2 provides an alternating square-wave voltage which is stepped up by the secondary winding. Full-wave rectification with current limiting is used with each secondary winding.