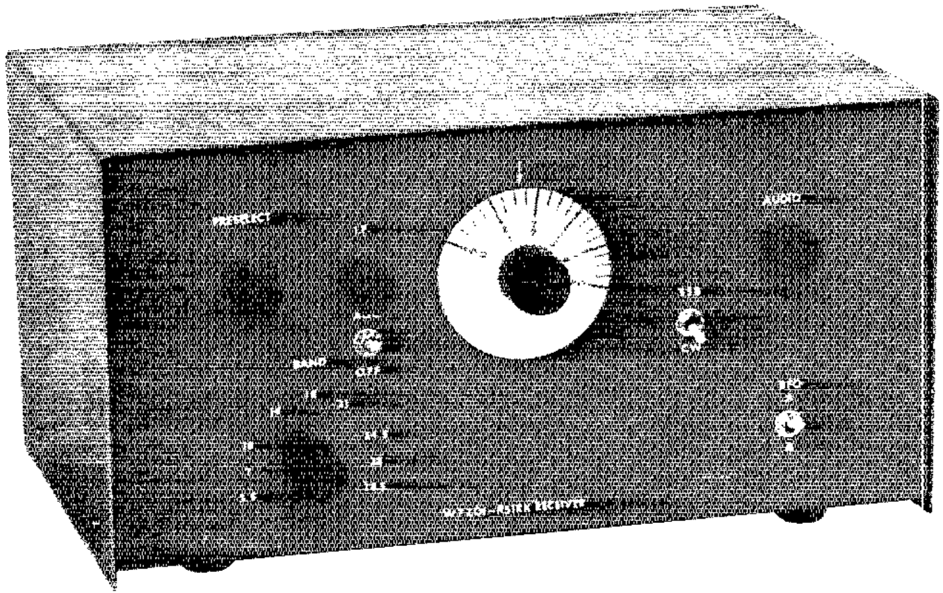
A Progressive Communications Receiver

Try your hand at building a good receiver! This two-step approach combines construction simplicity with performance that equals or exceeds that of many commercial units. You'll be proud to say, "I built it myself!"

By Wes Hayward,* W7ZOI and John Lawson,** K5IRK



The word "progressive" has a multiplicity of meanings. When applied to a receiver construction project it might imply that the work *progresses* from a simple beginning to something more elaborate. The receiver user, however, would assume that a *progressive* receiver is modern in design, having progressed with the state of the amateur art.

Both meanings apply to the receiver described here. The initial project results in a simple, but well-performing direct-conversion (D-C) receiver. Phase two adds circuits to provide an 80-meter superheterodyne. Multiband coverage is then provided for on an as-needed basis by the addition of carefully designed crystal-controlled converters. Virtually all of the components used for the D-C receiver are contained in the final version.

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saving time and money.

We should emphasize that although simple, this receiver is not a toy. The final superheterodyne features excellent stability, selectivity (consistent with the filter used), adequate sensitivity and a dynamic range that rivals or exceeds that of many commercial equivalents. The major compromise is the utilization of dual conversion on the higher bands. This penalty is small, for the gain distribution has been controlled carefully through proper design. Achieving good performance in simple equipment is not something that just happens — it must be *designed*. The reader is urged to review the thoughts on this subject presented by Roy Lewallen, W7EL.[1]

Some prospective builders may have little interest in a simple D-C receiver. Step-by-step construction of a D-C version is not shown here. However, bypassing that part of the project and proceeding directly to the final "superhet" design is *strongly* discouraged, especially if the builder lacks construction experience. The two-step method of construction facilitates later debugging and adjustment.

Simplicity and ease of duplication were considered paramount in the design. Readily available components are used throughout; alternatives are suggested where appropriate. Circuits are insensitive to the transistor type, thereby allowing component-substitution freedom.

A variety of construction methods may be used. Some builders may wish to etch their own circuit boards.[2] This is not meant to imply that etched boards are necessary or even desired. All of the circuit was initially breadboarded using "ugly" methods;[3] while not as professional in appearance, performance is vir-

[1]Notes appear on page 21.

tually identical. Where a performance difference could be detected, the ugly breadboards proved superior, usually a result of improved grounding. An added virtue of the ugly boards is that they are built in less than half the time required to lay out, etch, drill and construct the etched versions. Also, an "ugly board" is altered easily allowing a builder to incorporate design changes.

Ugly boards are built using scraps of unetched circuit-board material, which serve as the ground foil. Components are supported by those parts that are attached to the ground foil. Additional support may be provided with suitable tie points. Large-value resistors can serve this purpose well, especially in rf circuits where impedance levels are generally low.

The Progressive System

The system will be described in block-diagram form before we proceed with circuit details. This shows what the final result can be and aids in module interconnection. Fig. 1 is the block diagram of an 80-meter D-C receiver. The preselector filter is followed by the product detector and audio amplifier module. A doubly balanced diode-ring mixer serves as the detector. Four audio stages provide sufficient output to drive low-impedance headphones. An optional R-C active filter is shown. The remaining circuit section is the VFO.

A three-band superheterodyne version of the receiver is presented in Fig. 2. With the band switch in the 80-meter position, incoming signals are applied first to the preselector filter. This is the same one used in the D-C receiver. The preselector output is routed to a mixer module. This board contains a diode-ring mixer and a bipolar transistor i-f amplifier. Output of the mixer module is fed to the i-f amplifier board, which contains the crystal filter and a simple i-f derived agc system.

The i-f amplifier drives the detector/audio board used originally in the D-C receiver. Except for a couple of resistor value changes, the audio amplifier is unaltered. Detector injection voltage is provided by a crystal-controlled BFO. Use of two BFOs provides convenient sideband selection.

The 80-meter mixer is driven by the same VFO that was used in the D-C receiver. Some capacitors have been changed to move the output frequency of 3.5 to 4 MHz up to 5 to 5.5 MHz.

Multiband reception is provided by a crystal-controlled conversion process. While the receiver shown in Fig. 2 has only three bands, the band coverage and means of switching are flexible. The converter filter section *may* contain an optional rf amplifier (recommended only for the higher bands). A separate preselector filter is required for each band to be covered. The filter section is followed by a mixer module that feeds the 80-meter part of the receiver. Only one mixer module is used in the converter section. Mixer LO injection is provided by a group of crystal-controlled oscillators. One oscillator is required for each band.

There is considerable board commonality in the superhet. All of the converter filters have identical layouts; the converter mixer module is identical to that used in the 80-meter receiver. The crystal oscillators used for the BFOs and converters are identical.

80-Meter Preselector Filter

This filter (Fig. 3) consists of two cascaded sections. The first section is a 7-pole high-pass filter (3-MHz cutoff) composed of the components located between the two 650-pF capacitors. It is used to suppress spurious responses from mf broadcast signals.

The second filter section is unusual. While basically a low-pass type, it was designed for a very pronounced peak, resulting in a sharp, bandpass-like response. A front-panel mounted PRESELECTOR control is required. A 365-pF broadcast band replacement type of capacitor is used. If it is located remotely from the rest of the filter, a short length of small-diameter coaxial cable (RG-174/U) may be used for interconnection. A filter tuning range of 3 to 4 MHz may be useful for some applications to be discussed later. Data is provided for this variation.

Some builders may wish to construct a receiver without the front panel PRESELECTOR control. In that case, we suggest a 9-pole low-pass filter be designed for a 1-dB ripple (Chebyshev response) and a cutoff frequency of 4.1 MHz. This can be cascaded with the 7-pole high-pass filter of Fig. 3.

Detector-Audio Section

Shown in Fig. 4 is the backbone of the D-C receiver — the product detector and audio amplifiers. For ease of construction and component procurement, no ICs have been used. Although discrete-component detectors were used in early versions of the receiver, the Mini-Circuits Labs SBL-1 doubly balanced mixer was finally chosen. Not only does it offer excellent performance in D-C receiver applications, but the improved balance helps to eliminate problems caused by the agc system of the superhet being activated by BFO leakage. Other commercial mixers or homemade equivalents will also work well.

Detector output is fed to a diplexer network consisting of RFC1 and the related components. This network ensures that the detector is terminated properly at all frequencies from audio to vhf, providing optimum dynamic range in D-C receiver applications. Additional information about product detectors for D-C receivers is presented in *Solid-State Design for the Radio Amateur*[4] and in the paper by Lewallen.[5] We have borrowed liberally from his work in much of this design.

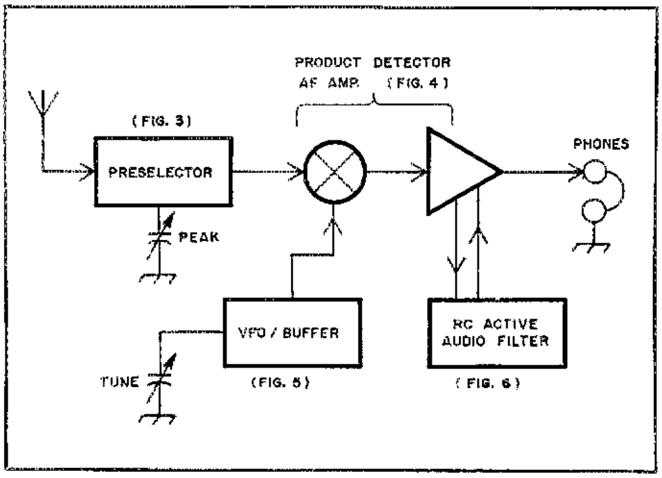
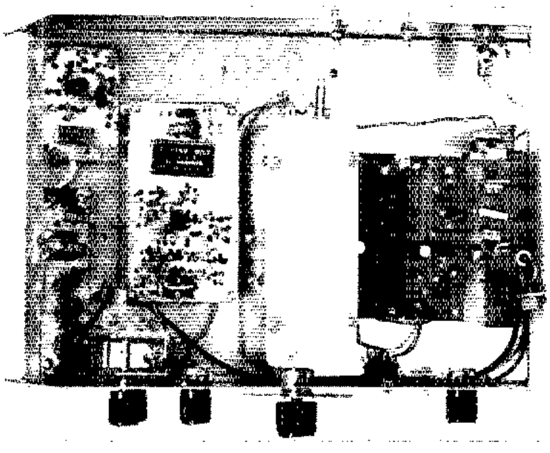


Fig. 1 — Block diagram of an 80-meter direct-conversion receiver. The receiver may be constructed for use on other bands.



An inside, top view of the Progressive Receiver. The unit shown in the photos was constructed by John Lawson, K5IRK. *(photos by Roger Hayward, KA7EXM)*

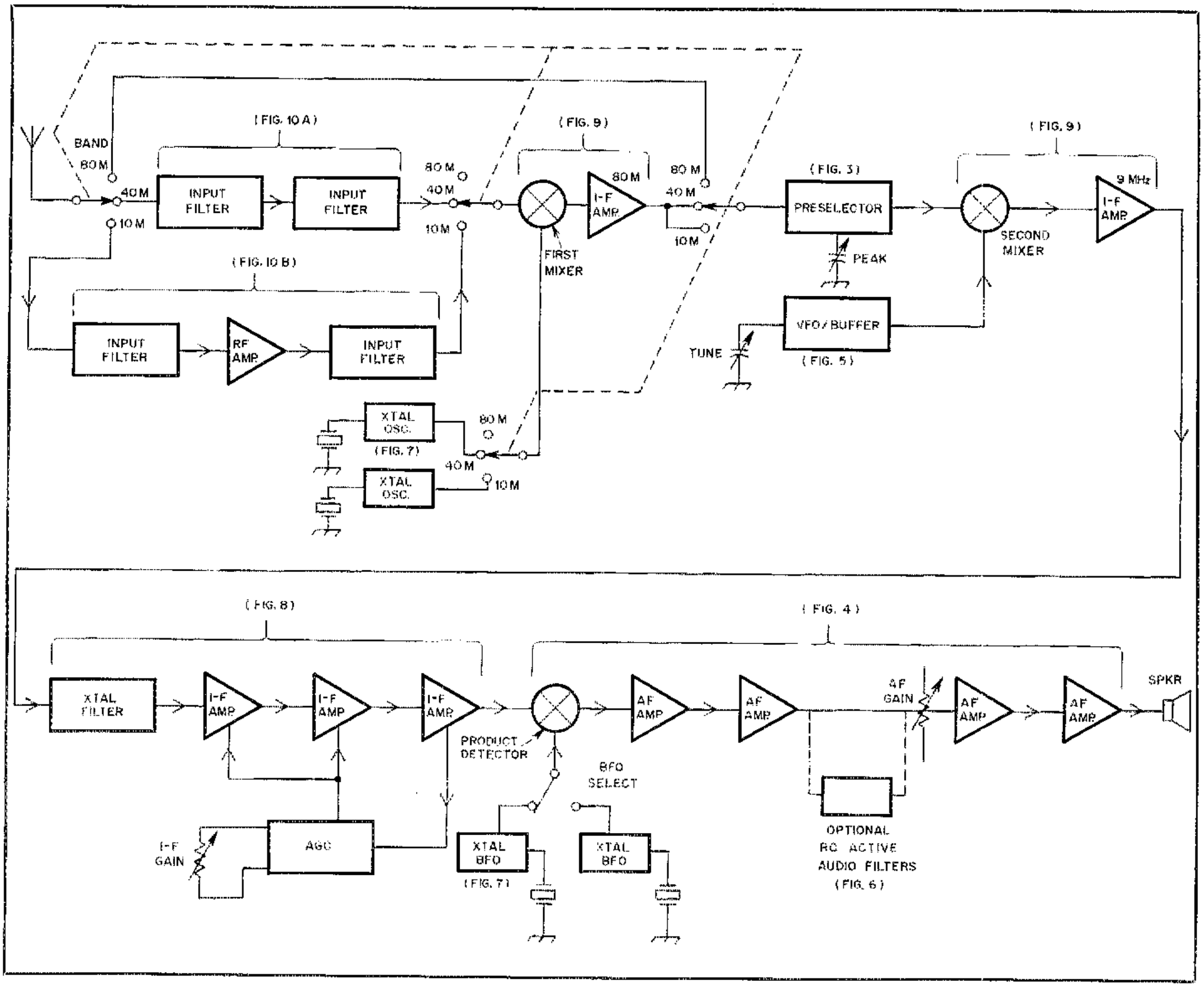


Fig. 2 — Superheterodyne receiver block diagram. As shown, the receiver covers three bands: 80, 40 and 10 meters. All bands from 80 through 10 meters (including the WARC frequencies) may be added to the basic receiver at the builder's discretion. Refer to the text for details.

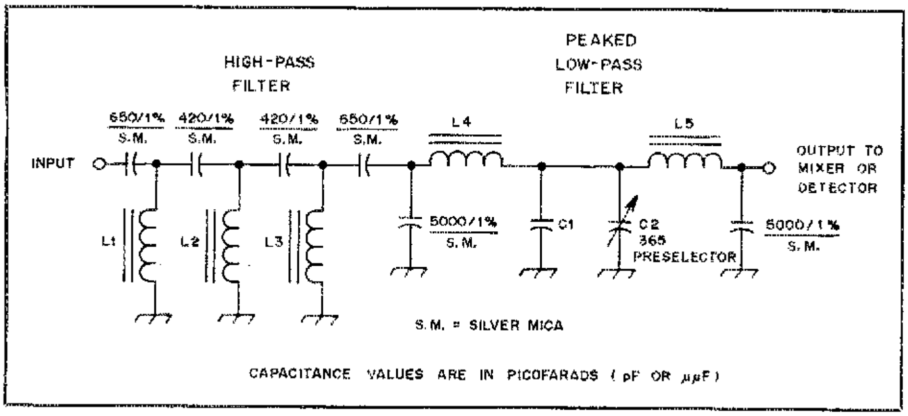


Fig. 3 — The 80-meter preselector filter. A high-pass filter is used at the input; a peaked low-pass filter comprises the output section. Nearest value 5% silver-mica capacitors may be substituted for the 1% units shown.

C1 — 560-pF silver-mica capacitor for 3.5- through 4-MHz coverage, 250-pF silver-mica unit for 3- through 4-MHz coverage (see text).
C2 — Broadcast replacement type of variable capacitor, 365 pF or more.
L1, L3 — 21 turns no. 22 enameled wire on T50-2 core.
L2 — 20 turns no. 22 enameled wire on T50-2 core.
L4, L5 — 30 turns no. 22 enameled wire on T68-2 core for 3.5- through 4-MHz coverage; 45 turns no. 24 enameled wire on T68-2 core for 3- through 4-MHz coverage.

First audio amplifier Q1 is connected in a common-base configuration to terminate properly the diode-ring detector. It is biased to an emitter current of about 0.5 mA to present an input resistance of 50 ohms. Q2 is a direct-coupled pnp amplifier that functions as the second audio stage. The receiver may be muted by shorting the collector of Q2 to ground. This is accomplished by applying a positive potential of a few volts to the muting input to saturate Q5. The output of Q2 is fed to an AUDIO GAIN control on the front panel. If the optional R-C active filter is used, it is inserted in series with the output of Q2.

Q3 is a common-emitter amplifier. Q4 is an emitter follower. Q4 has an emitter-current bias of about 30 mA to provide enough audio output to drive low-impedance (4- to 16-ohm) headphones. Two series-connected, 75-ohm, 1/4-watt emitter resistors are used to provide sufficient power dissipation and to ensure

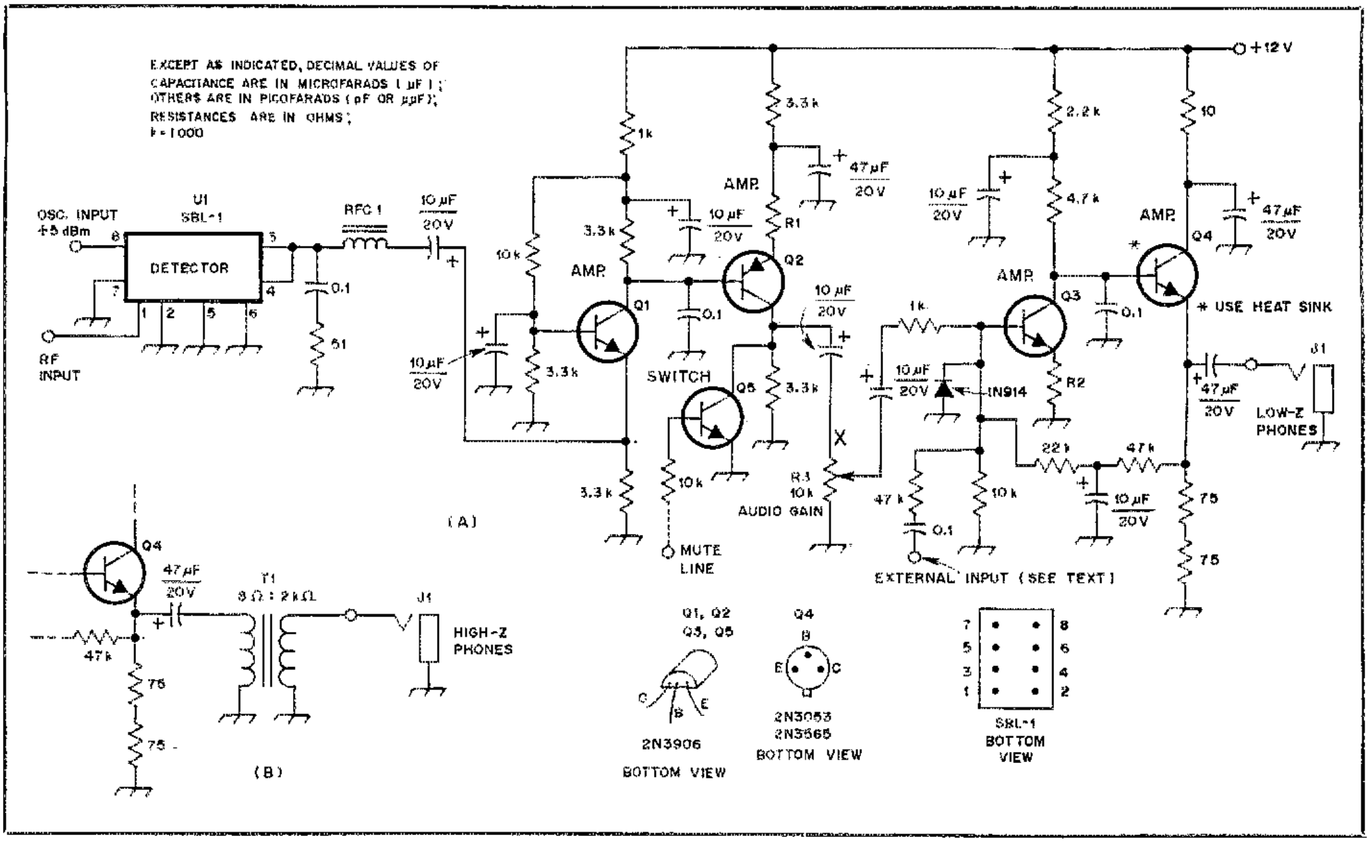


Fig. 4 — Schematic diagram of the product detector and audio-amplifier section is shown at A. The circuit at B may be added if high-impedance headphones are used. When the recommended optional audio filters are employed, they are inserted at the point marked X. All resistors are 1/4-watt composition or metal film types. Pin 1 of the SBL-1 is beneath the MCL marking on the top of the case.

Q1, Q3 — Silicon npn low-noise bipolar transistor 2N3565 or equiv.
Q2 — Silicon pnp general-purpose bipolar transistor, 2N3906 or equiv.
Q4 — Silicon npn bipolar transistor, 2N3053 or equiv., with small heat sink.
Q5 — Silicon npn general-purpose bipolar transistor, 2N3904, 2N2222, 2N3565 or equiv.
R1, R2 — 10 Ω for direct-conversion receiver, 220 Ω for superheterodyne version.
R3 — 10-kΩ audio-taper potentiometer.
RFC1 — 20 turns no. 28 enameled wire on FT37-43 core.
T1 — Reverse-connected miniature audio-output transformer, 8 Ω to 2 kΩ.
U1 — Doubly balanced diode-ring mixer, Mini-Circuits Laboratory SBL-1 or equiv. (see text). The SBL-1 may be obtained from Mini-Circuits Laboratory, 2625 East 14th St., Brooklyn, NY 11235.

reliability. A small heat sink is used on Q4. If low-impedance headphones are used, they should be the inexpensive types. Those designed for high-fidelity applications are usually too inefficient for use in this circuit. Should you wish to use high-impedance headphones, the circuit of Fig. 4B should be employed to achieve the required voltage amplification.

Two resistors in the audio system (R1 and R2) must be changed when the board is used in the superhet version. This is because higher audio gain is needed in the D-C receiver. An auxiliary input is provided in the audio-amplifier section. This is intended for cw sidetone signal injection.

The VFO

A general-purpose VFO is shown in Fig. 5. It will function well at frequencies in the 2.5- to 10-MHz range. Only the capacitors need to be changed to alter the tuning range.

Q6 is employed in a JFET Hartley circuit that has the virtues of simplicity and good stability. Recent work by Lewallen[6] has optimized this oscillator for thermal stability. Best stability results if a toroid core of the SF type (Amidon −6 code)' is used for the inductor. This material has a +50 ppm/C° temperature coefficient — much better than the usual slug-tuned inductor. All fixed-value capacitors should be NPO ceramic units. They have the lowest temperature coefficient of any of the readily available types. Use of silver-mica and polystyrene capacitors should be avoided. The latter exhibit a temperature coefficient of −150, ±50 ppm/C° and are not recommended. Their popularity in VFO applications results from frequent use with slug-tuned inductors, which often have temperature coefficients of about +150 ppm/C°.

The resonator (tuned circuit) should be lightly loaded by the FET. This is ensured by keeping the gate-coupling capacitor as small in value as possible. If the specified 2.7-pF NPØ ceramic unit cannot be found, a small air-variable of similar capacitance may be substituted.

Excellent VFO stability is obtained easily if these precautions are followed. Oscillators operating at 5 or 7 MHz have exhibited a typical warm-up drift of less than 200 Hz over a period of about 10 minutes. Afterward, the VFO does not move more than 10 or 20 Hz in a five-minute period, assuming the ambient temperature is reasonably constant. This data is not the result of a single measurement. Literally dozens of these oscillators have been built, all producing predictable results.

Table 1

VFO Capacitor Values For Different Operating Frequencies

Frequency Range (MHz)	*C3 (pF)*	*C4 (pF)*	*C5 (pF)*
3.5 to 3.8	281	200	120
3.5 to 4.0	290	200	40
3.0 to 4.0	305	1000	18
5.0 to 5.5	126	100	82
5.0 to 6.0	66	200	120
7.0 to 7.2	60	50	200

Capacitor values for the circuit of Fig. 5 may be taken from Table 1 if a 365-pF variable is used for C6, the MAIN TUNING capacitor. Considerable flexibility exists, and the equations in the appendix may be used to calculate the values for C3, C4 and C5 if another type of variable capacitor is used for C6. The Table 1

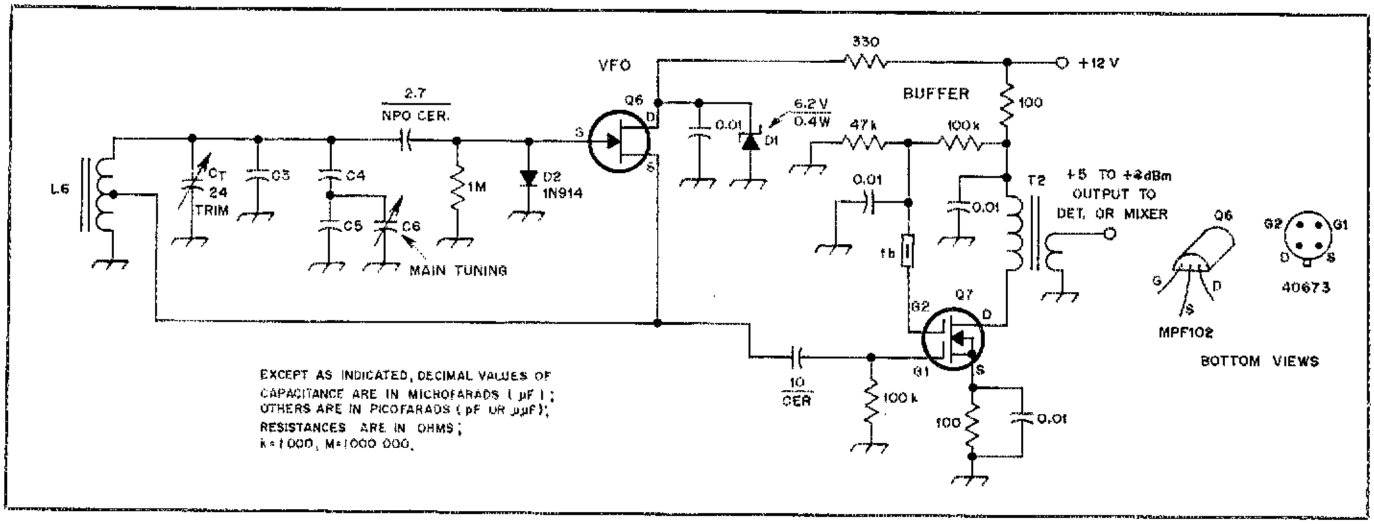


Fig. 5 — VFO schematic diagram. This circuit performs well at frequencies between and including 2.5 and 10 MHz. Refer to the text, Table 1 and the Appendix for component selection information. Resistors are 1/4-watt, 5% composition or metal film types. All fixed-value capacitors are disc-ceramic types.

C3-C5, incl. — NPØ ceramic. See Table 2 and Appendix.
C6 — Air variable capacitor, 100 pF or more total capacitance variation.
L6 — Approximately 4.9 µH, 35 turns no. 28 enameled wire on T50-6 core, tapped at 8 turns from ground end.
Q6 — Silicon n-channel high-frequency JFET, MPF-102, 2N4416, TIS-88, 2SK19GR or equiv.
Q7 — Silicon n-channel dual-gate rf-amplifier MOSFET, 40673, 3N140, 3N211, 3SK40 or equiv.
T2 — Ferrite transformer, 18-turn primary, 5-turn secondary, no. 28 enameled wire on FT37-43 core.

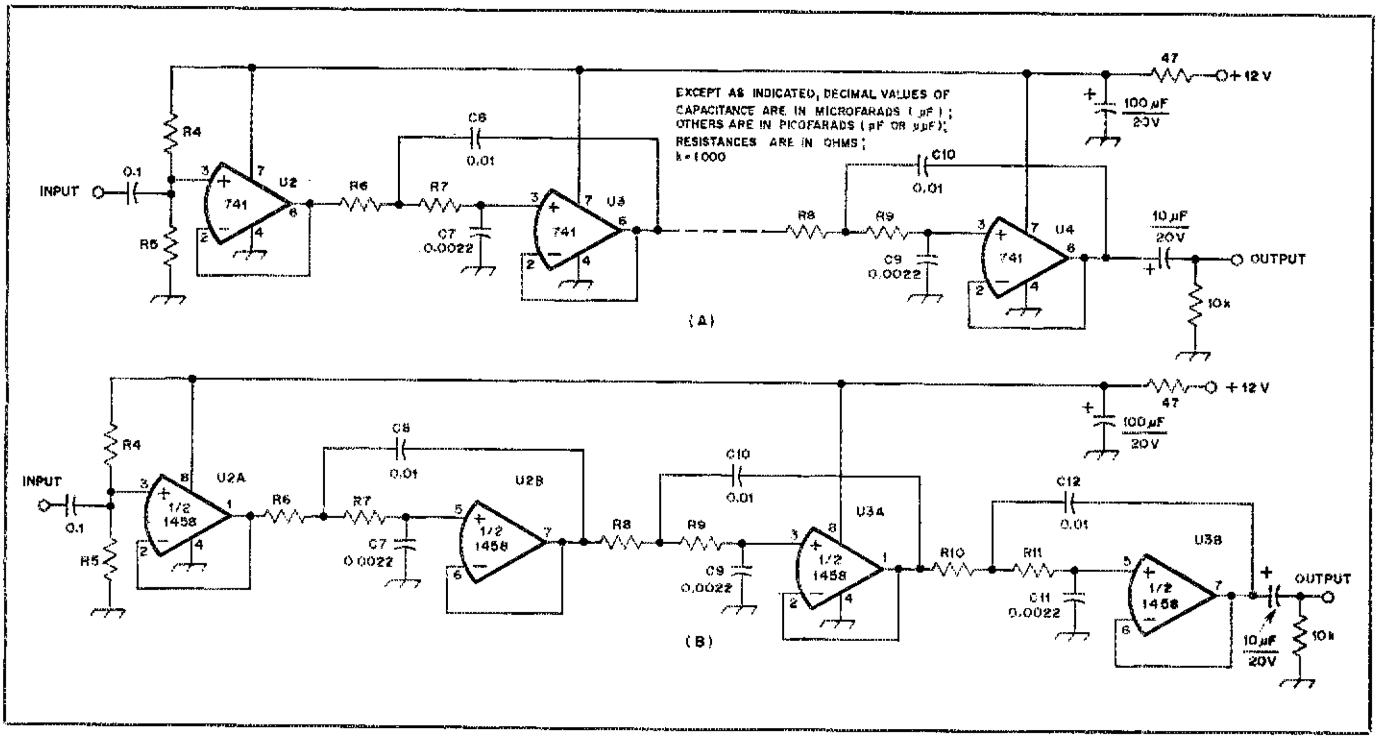


Fig. 6 — An optional R-C active audio filter. At A, a 4-pole, low-pass filter using 741 (single-section) op amps is shown. The circuit at B employs two 1458 (dual-section) op amps and provides six poles of low-pass filtering; more poles are recommended for improved performance. Other types of op amps may be substituted, but use caution as pin numbers may differ. Resistors are 1/4-watt, 5% composition or metal film types.

C7, C9, C11 — 0.0022 µF ceramic or polystyrene, 10% or better tolerance.
C8, C10, C12 — 0.01 µF ceramic or polystyrene, 10% or better tolerance.
R4-R11, incl. — 33 kΩ, 5% for cw filter, 15 kΩ for ssb filter.
U2-U4, incl. — See caption.

value for C3 includes the capacitance of the trimmer, C_T.

A dual-gate MOSFET is used for the buffer amplifier, Q7. The circuit is conventional except for the use of a broadband output transformer. Good isolation is provided, and the power output is ample for driving a diode-ring mixer or detector. Power values of +5 to +8 dBm have been measured, using a 50-ohm load. A ferrite bead is placed directly on gate 2 of Q7. While we attempt to design without "Band-Aids of anticipation," this one is worthwhile; it introduces loss, which will stop uhf oscillations.

Mechanical construction is important in a VFO. Any of the available reduction drives are suitable for the direct-conversion model of this receiver. Additional bandspread is desirable for superhet applications. A Jackson Brothers dualspeed drive, type 4511/DRF,[8] was used in the K5IRK receiver, and it is adequate. The receiver built at W7ZOI contains a surplus capacitor and drive unit from a BC-455.[9]

Optional R-C Audio Active Filter

Additional selectivity may be desired for either phase of the project. The R-C active filter shown in Fig. 6 is designed to fulfill this need. The filter has a single pole of high-pass filtering followed by a series of low-pass poles. The cutoff frequency is about 1 kHz for cw and 2 kHz for ssb. Filter bandwidth is selected by a proper choice of the resistor values. More filter sections may be added for improved skirt selectivity. The Q of the individual low-pass sections may be increased if a narrower bandwidth is desired. While the filter is shown with operational amplifiers as the active elements, discrete transistors are also suitable.[10]

The performance of the D-C receiver is excellent. Detailed measurements have not been done on this unit, but Lewallen has done some while using a similar circuit.[11] Good D-C receivers display a quality of exceptional "cleanliness" and "presence", and this one is no exception. Indeed, the effect is perhaps more pronounced because of improved audio fidelity.

Crystal Oscillators

The circuit shown in Fig. 7A is used for all of the crystal oscillators in the receiver. One or two will serve as the BFO. Additional units are required for each band to be covered by means of the crystal-controlled converters. Though the circuit may appear to be strange (something of a "trick" circuit), this is not the case. If the diagram is redrawn with the ground point placed at the transistor base, it becomes clear that this is nothing more than a Hartley oscillator with the crystal in series with the feedback tap from the coil.

A capacitor in series with the crystal permits adjusting the oscillator to the operating frequency. This is vital only for the oscillators used for the BFOs. This capacitor may be eliminated (using a jumper wire) in those modules used for converter application. C12 is tuned for maximum output and reliable oscillator starting. The +12-volt operating bias is applied through the resonator output link for ease of band switching. Power is supplied to any single oscillator, as shown in Fig. 7B. The method used for sideband and band selection is shown in Fig. 7C. Only the oscillator in use has power applied to it.

These crystal oscillators will deliver an output power of about +10 dBm. This is more than enough to drive the detector or mixer. The circuits are adjusted easily as long as they are terminated properly — they are best adjusted with the mixer or detector attached. If experimentation is done with the circuits for other than their intended application, they should be ac coupled to a 50-ohm load.

One crystal type, a KVG XF-903,[12] will function well for all 9-MHz BFO applications. Two crystals can be used in separate BFO units, or a single BFO can be mounted near the receiver front panel, with an operator-adjusted variable capacitor used for the BFO control.

A list of oscillator frequencies and component values appears in Table 2. Use of the 3.3-MHz oscillator frequency will convert the 7- through 7.3-MHz band to the 3.7- through 4-MHz tuning range. The virtue of this scheme is that all bands will

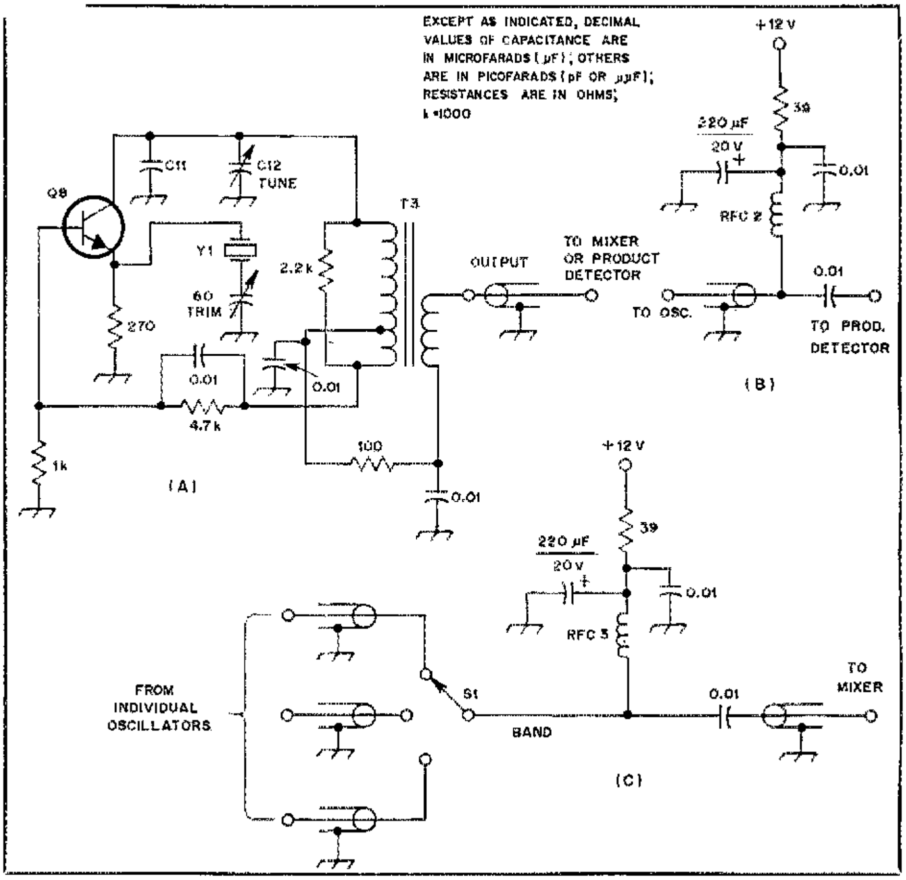


Fig. 7 — At A, the crystal oscillator schematic diagram. This module is duplicated a number of times in the receiver. It is used for the BFO as well as for the individual conversion oscillators. Refer to the text for circuit details. Component information is given in Table 2 and the parts list. Resistors are 1/4-watt composition or metal film types, 5% tolerance. B and C show the methods used to couple the oscillator output to the product detector and mixer circuits, respectively.

C11 — Silver-mica or ceramic capacitor; see Table 2.
C12 — Mica compression or similar trimmer capacitor; see Table 1.
Q8 — Silicon npn general-purpose bipolar transistor, 2N3904, 2N2222 or equiv.
RFC2, RFC3 — Approximately 20 turns no. 28 enameled wire on FT37-43 core.
S1 — Part of band switch or sideband-selector switch; see text.
T3 — Wound with no. 28 enameled wire; see Table 2.
Y1 — Series-resonant crystal for use at required frequency shown in Table 2. For 9-MHz BFO applications, the KVG XF-903 is suitable for lsb or usb.

Table 2

Crystal Oscillator Component Selection

Y1 (MHz)	*Frequency (meters)*	*C11 (pF)*	*C12 (pF)*	*Core type*	*T3 Primary turns*	*Tap turns*	*Secondary turns*
3.3	40	100	90	T68-2	65	13	10
6.5	30	100	60	T50-6	35	7	6
9	BFO	56	60	T50-6	35	7	6
10.5	20	56	60	T50-6	30	7	6
11	20/40	22	60	T50-6	30	7	6
14.5	17	33	60	T50-6	23	5	4
17.5	15	33	60	T50-6	23	5	4
20.5	12	none	60	T50-6	20	4	4
24.5	10/15	none	60	T50-6	20	4	4
32	10	none	60	T50-6	15	3	3

Note: No. 28 enameled wire is used for T3 windings.

then tune in the same direction, a subjective detail of consequence to some. Forty-meter reception may be achieved by using an oscillator frequency of 11 MHz. But the tuning direction will be opposite that on 80 meters. The latter method is used by the writers.

Many converter-construction options are available. You can build the basic 80-meter receiver to tune the wider range of 3 to 4 MHz. Then, all five bands from 80 to 10 meters may be received using only two crystal oscillators (11 and 25 MHz). This scheme would be ideal for the economy-minded builder. Data for this approach is presented for the construction of the VFO and 80-meter preselector.

I-F Amplifier

An important part of any superheterodyne receiver is the i-f section. A 9-MHz i-f is used in this receiver (Fig. 8). Selectivity is provided by a crystal filter of the builder's choice. This circuit accepts

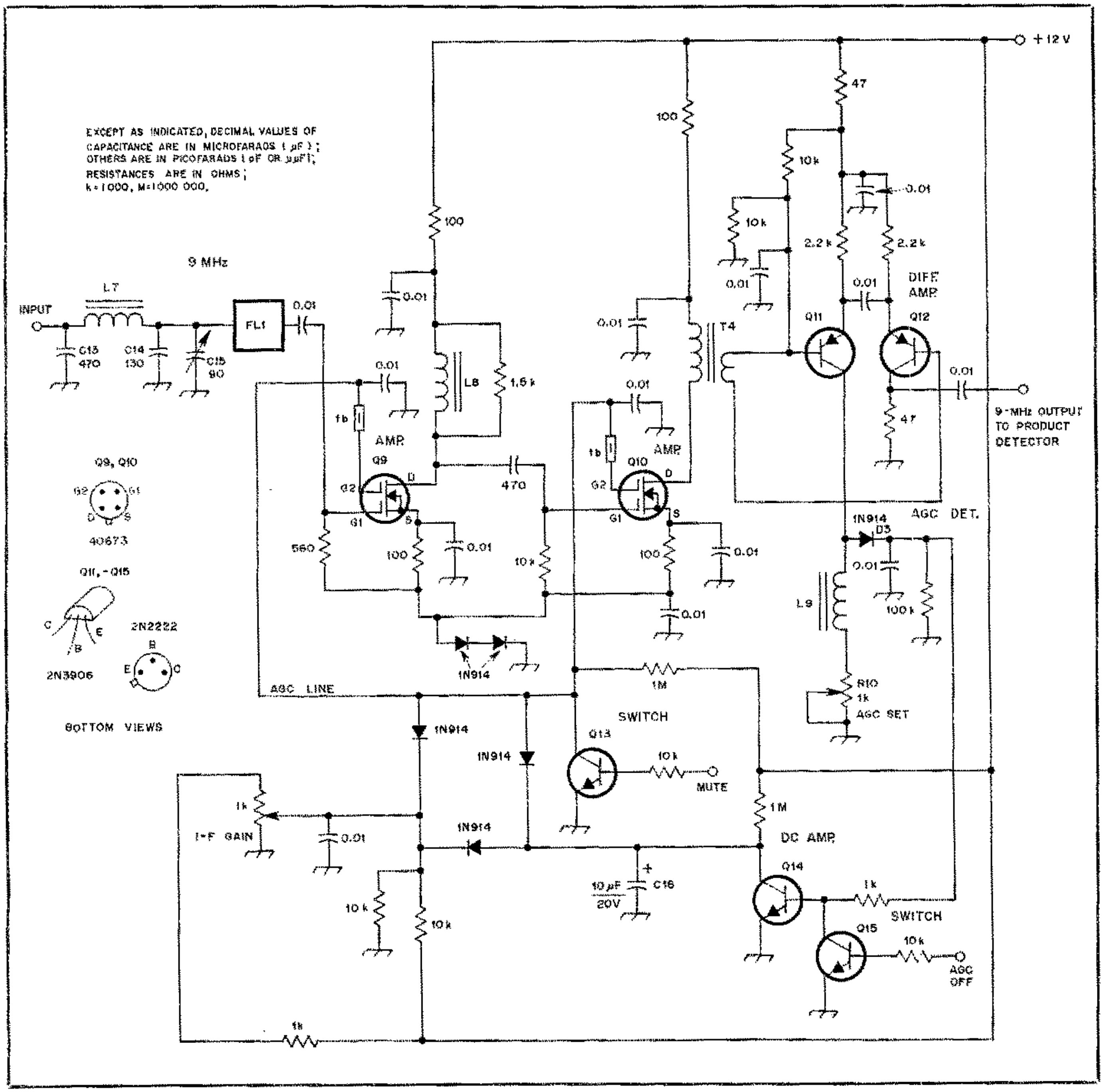
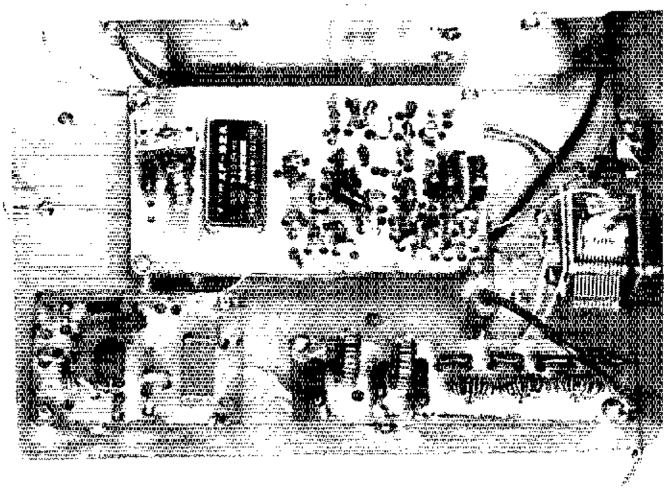
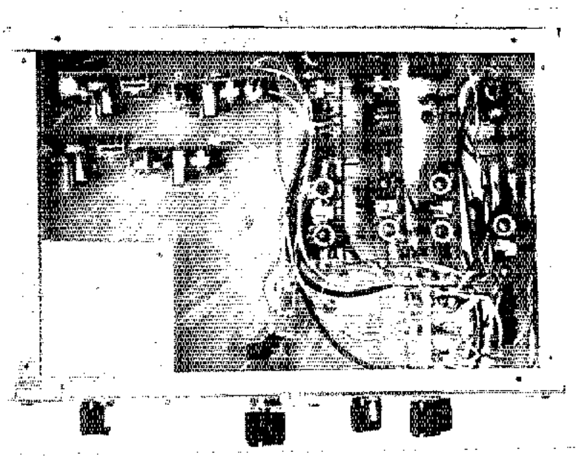


Fig. 8 — Schematic diagram of the 9-MHz i-f amplifier. All fixed-value resistors are 1/4-watt, 5% carbon or metal film types. The adjustment of R10 is discussed in the text.

C15 — 90-pF mica compression trimmer.
FL1 — 9-MHz crystal filter, 500-Ω input/output impedance; for ssb, KVG type XF-9B or Yaesu XF-92A; for cw, KVG type XF-9M, XF-9NB or Yaesu type XF-90C. KVG filters may be obtained from Spectrum International, Inc., P.O. Box 1084, Concord, MA 01742. Yaesu filters may be obtained from Yaesu Electronics Corp., P.O. Box 498, Paramount, CA 90723.
L7 — 23 turns no. 22 enameled wire on T50-6 core.
L8, L9 — 15 turns no. 28 enameled wire on FT37-43 core.
Q9, Q10 — Silicon n-channel dual-gate rf-amplifier MOSFET, 40673, 3N211, 2SK40 or equiv.
Q11, Q12 — Silicon pnp general-purpose bipolar transistor, 2N3906 or equiv.
Q13-Q15, incl. — Silicon npn general-purpose bipolar transistor, 2N3904, 2N2222 or equiv.
R10 — 1-k, pc-mount, linear-taper potentiometer.
T4 — Ferrite transformer, 15 primary turns and 3 secondary turns of no. 28 enameled wire on an FT37-43 core.



In this close-up photo, the 80-meter preselector and 2nd mixer module may be seen in the foreground. The 9-MHz i-f board is in the center, and a portion of the VFO enclosure is at the top of the photo.



At the lower left, a pc-board enclosure houses the BFO circuitry. Immediately above the BFO four crystal oscillator modules may be seen. Four input filter boards are arranged behind the BAND switch. The two input-filter boards at the extreme right are without rf amplifier components. Another mixer module is at the lower center of the chassis.

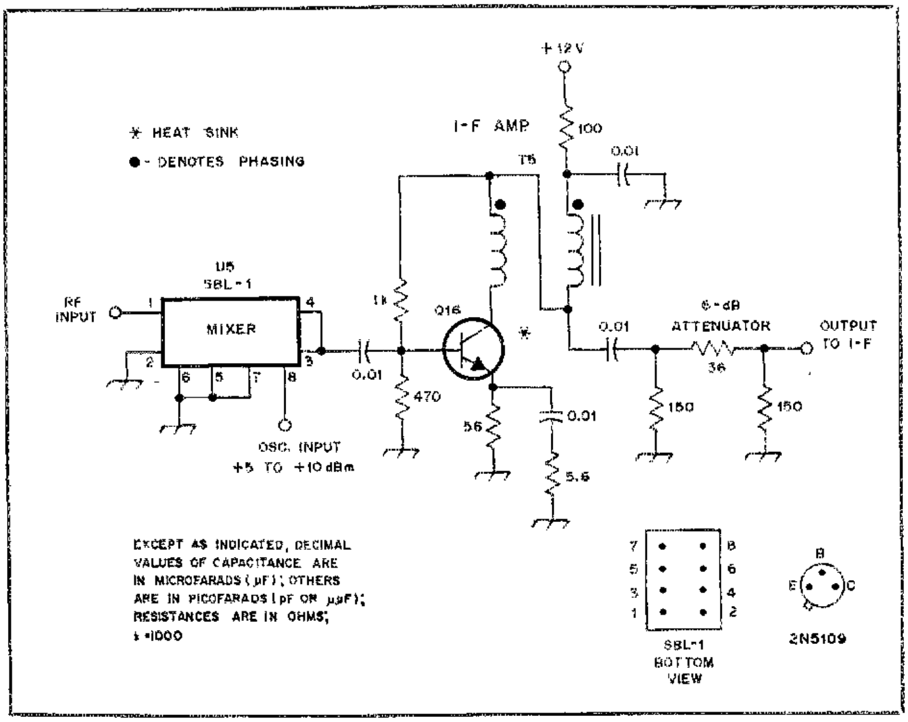


Fig. 9 — Mixer-module schematic diagram. Two of these units are required when constructing the superheterodyne version of the receiver. One module is used as a first mixer stage for input signal frequencies of 40 through 10 meters, with an output at 80 meters. The other module (the second mixer stage) converts 80-meter input signals to the 9-MHz i-f. All resistors are 1/4-watt, 5% composition or metal film types.

Q16 — Silicon npn CATV amplifier (f_T = 500 MHz or greater) bipolar transistor, type 2N5109, 2N3866, 2SC1252, 2SC1365 or equiv. A small heat sink should be used on Q16.

T5 — Broadband ferrite transformer, 10 bifilar turns no. 28 enameled wire on FT37-43 core.

U5 — Doubly balanced diode-ring mixer, Mini-Circuits Laboratory SBL-1 or equiv.

filters requiring 500-ohm terminations. A pi-network input circuit transforms the 50-ohm source impedance of the mixer module to the 500-Ω load required by the filter. The filter output is terminated in a 560-ohm resistor. An unusual feature of the i-f system is the lack of tuned circuits other than the pi network. All transformers are ferrite types that are designed for broadband performance.

A majority of the i-f gain is provided by two dual-gate MOSFETs, Q9 and Q10. The bias on these stages is raised by placing a pair of silicon diodes in the source lead. This extends the gain control range as the gate 2 bias is altered.

The last i-f stage is a differential pair of transistors, Q11 and Q12. Output is available from each of the collectors. The output of Q12 is routed to the product detector by means of small-diameter coaxial cable, while Q11 output is fed to the agc detector, D3. A voltage appears at the base of Q14 when high-amplitude signals are present. C16, the timing capacitor, is then discharged. To reduce i-f stage gain, this voltage change is coupled to the agc line by means of a diode. R10, the AGC SET, is adjusted for a dc voltage of 0.4 to 0.5 at the base of Q14. This adjustment should be made with the agc system activated and with no signals present. The potential on the agc line will then be about 6 volts with manual I-F GAIN set at maximum. A high-impedance (10 megohms or greater) voltmeter should be used for any agc-line measurements.

Two transistor switches are contained in the i-f strip. Q15 defeats the agc when a positive voltage is applied to the AGC OFF line. Q13 is attached directly to the agc line. A positive input voltage to that switch shorts the agc line to ground to mute the receiver. The agc line diodes allow muting to occur quickly without discharging C16. The i-f strip returns quickly to full gain after muting periods.

Agc response is more than adequate, with minimal overshoot. Recovery time is relatively independent of the signal level. By decreasing the value of C16 or its associated 1-megohm resistor, the recovery time may be shortened.

Mixer

Fig. 9 is the schematic diagram of the 80-meter mixer. A Mini-Circuits Lab SBL-1 or similar unit may be used. Again, homemade mixers will also work well. A suitable substitute can be built from a pair of Amidon FT37-43 toroid cores and four hot-carrier diodes.

Q16, a 9-MHz i-f amplifier, follows the mixer. This is one of the more critical stages of the receiver. It must have a low noise figure, for it will determine the receiver sensitivity. Intermodulation distortion must be low, and the input and output impedances need to be 50 ohms. A bipolar transistor with negative feedback is used to establish the gain and required impedance levels. The stage is biased to draw a moderately high current. This ensures low distortion. A 6-dB pad at the output preserves the impedances at the input and output of the amplifier.

Transistor type *is* critical for Q16 because it should be one with a high gain-bandwidth product (f_T) — at least 500 MHz. Amplifier gain, minus the loss in the pad, is about 16 dB. The mixer has a conversion loss of about 6 dB, leaving a net gain of 10 dB. The amplifier output intercept is approximately +30 dBm. Careful measurements have shown that a diplexer network is not required between the mixer and this amplifier.

Assuming the VFO frequency has been changed to cover 5 to 5.5 MHz, the receiver is now ready for operation on 80 meters. A few circuits requiring alignment are adjusted for maximum output. Homebuilt instrumentation[13] is adequate for use during alignment. It was used by the writers. The adjustments may also be performed "by ear," listening to incoming signals with the agc defeated.

Crystal-Controlled Converters

To receive other bands of interest, suitable converters can be added to the basic receiver. Net gain through the converters is kept low to preserve the overall receiver dynamic range. An additional mixer module, identical to that used in the 80-meter front end, is needed. Each band to be added will require a band-pass preselector filter and a crystal oscillator. You may wish to include an rf amplifier on the higher frequency bands; none is required for the 40-meter band.

Fig. 10 shows two versions of the input filter. That at A is without the optional rf-amplifier stage, while that at B shows a filter with the rf amplifier. The same board layout is used for both versions; a wire jumper (W) between points X and Y is used when the amplifier is omitted.

Consider first the option without the rf amplifier stage. There are two filter circuits shown that may be separated at points X-Y. The input filter (to the left of the X) is a five-pole low-pass type. Use of this filter section was necessary to eliminate spurious responses resulting from strong TV and fm broadcast station signals. This problem results from the tendency of diode-ring mixers toward harmonic mixing. The second filter (to the right of point Y) provides the major portion of the front-end selectivity. It is a double-tuned circuit composed of L12, L13 and the related capacitors. These filters were designed for a Butterworth response, but may be shifted to a Chebyshev response with a slight increase in the value of C22. The filters were designed while using the equations presented in Appendix 2 of reference 4.

A variable capacitor (C22) is used as the coupling element between the resonators. This was done in the interest of component selection. Small, nonstandard capacitance values are often difficult to obtain, whereas a 1- to 5-pF variable trimmer is common. The proper capacitance setting for C22 may be taken from Table 3.

Use a signal generator to align the filters. If one is not available, a crystal calibrator may be used. Terminate the filter input with a 50-ohm resistor. C22 is set initially near minimum capacitance, and the receiver is tuned to the band center. C23 and C25 are adjusted to provide a peak response. The capacitance of C22 is then increased, and C23 and C25 are repeaked. Filter bandwidth is estimated by tuning toward the band edges, repeating the alignment procedure until the desired bandwidth is realized.

A dual-gate MOSFET is used for the rf amplifier in the version shown in Fig. 10B, along with a modified input low-pass filter. This modification produces a pi network that transforms the 50-ohm filter

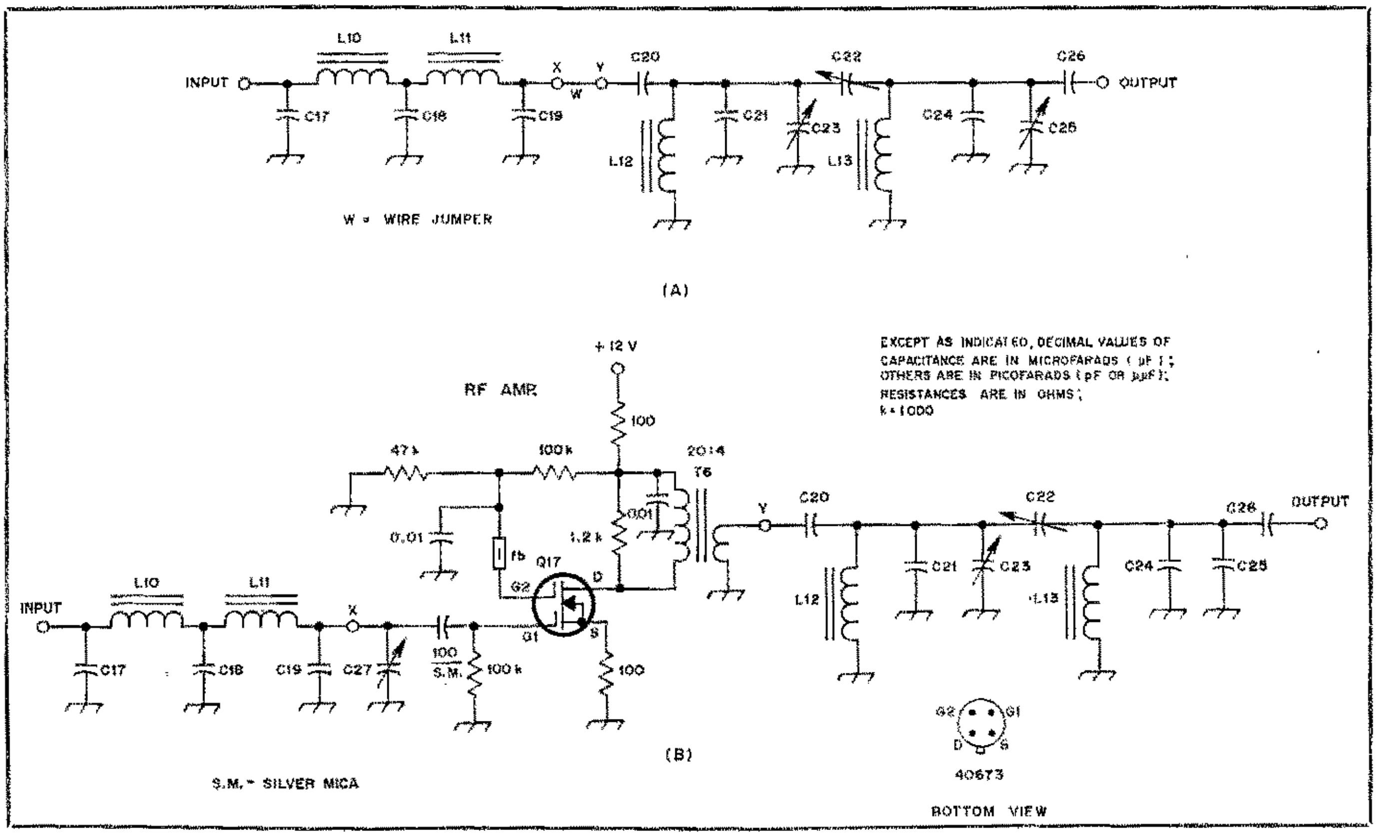


Fig. 10 — Input filter/amplifier circuitry. The same pc board pattern is used for the circuits of A and B. At A, a jumper wire is used when no rf amplifier (as shown at B) is employed. The circuit at B is recommended for operation at frequencies above 40 meters. Resistors are 1/4-watt, 5% composition or metal film types. Refer to the parts list and Table 3 for component descriptions.

Q17 — Silicon n-channel dual-gate rf-amplifier MOSFET, 40673, 3N211, 3SK40 or equiv.

T6 — Ferrite transformer, 20-turn primary, 4-turn secondary, no. 28 enameled wire on FT37-43 core.

Table 3

Input Filter Components

Without rf amplifier:

Frequency (MHz)	C17 C19 (pF)	C18 (pF)	L10* L11 (turns)	C20 C26 (pF)	C21 C24 (pF)	C23 C25 (pF)	C22 (turns)	L12* L13 (turns)
7.1	430	860	17	42	50	180	4.6	25
10.1	300	600	13	32	50	180	4.1	17
14.2	220	430	12	20	—	180	2.3	17
18.2	180	360	10	22	50	180	3.9	10
21.2	150	300	10	18	—	180	3.0	10
24.2	130	270	9	14	—	180	2.1	10
28.5	110	220	8	12	—	180	1.6	10

With rf amplifier:

Frequency (MHz)	C17 (pF)	C18 (pF)	C19 (pF)	C27 (pF)	L10* (turns)	L11* (turns)
10.1	300	680	33	50	13	29
14.2	220	500	22	50	12	25
18.2	180	390	—	50	10	22
21.2	150	330	—	50	10	20
24.2	130	300	—	50	9	19
28.5	110	250	—	50	8	17

Note: Other filter parts identical to that without rf amplifier.
*All inductors wound with no. 22 enameled wire on T50-6 cores.

Table 4

Performance Summary

Circuit	Rf amp.	Bandwidth (Hz)	NF (dB)	IP_{in} (dBm)	MDS (dBm)	DR (dB)
Single conv.*	no	500	16	+18	−131	99
Single conv.	no	2500	16	+18	−124	94
Single conv.	yes	500	5	+2	−142	96
Single conv.	yes	2500	5	+2	−135	92
Dual conv.	no	500	18	+12	−129	94
Dual conv.*	no	2500	18	+12	−122	89
Dual conv.	yes	500	6	−2	−141	92
Dual conv.*	yes	2500	6	−2	−134	88

Rf amplifier assumed to have a 3-dB noise figure, a 15-dB gain and a +22-dBm output intercept. Circuits marked (*) are measured cases. All measurements done at 14 MHz.

impedance up to about 2000 ohms, with a Q of 10, to provide a near optimum driving impedance for the amplifier. The amplifier output circuit uses a broadband ferrite transformer to present a 50-ohm output impedance to the following double-tuned circuit. This ensures a proper termination. This two-pole input preselector filter is aligned as previously described, then the pi network section is adjusted, peaking the response by means of C27 at the band center.

The amplifier uses no source-bypass capacitor. During evaluation it was found that the gain of the rf amplifier was excessive when such a bypass was included, resulting in degraded receiver dynamic range. Removal of the capacitor dropped the gain from 25 to 15 dB with little change in the distortion characteristics; noise figure was still low enough.

Thoughts, Hints and Results

A number of construction details have been left to the discretion of the builder — band switching, the power supply and an enclosure. Band switching is not critical, because all switched points are low impedance. A multiwafer rotary switch would serve nicely. Something as mundane as a group of slide switches will work just as well. Small-diameter coaxial cable, such as RG-174/U, should be used for all signal lines. Some builders may wish to include an auxiliary audio power amplifier for speaker operation. Suitable circuits have been described in *QST*.

Other refinements could be included: an S meter (using a simple high-impedance voltmeter circuit attached to the agc line), a crystal calibrator or a digital readout. The constructor looking toward the future may want to include the WARC bands. Calculated data for these bands is provided, although the writers have yet to build these circuits.

Just as the receiver can be refined and made more elaborate, simplification is also possible. For example, a converter may be used with the original D-C receiver. Or, a simple superhet may be built by eliminating the i-f amplifier. A crystal filter mounted on a board with a pair of impedance-matching, 50-ohm pi networks could be inserted between a mixer module and the detector. This could be refined by placing a single i-f stage after the filter. A single-conversion receiver for 14 MHz may be built by replacing the 80-meter preselector with one for 20 meters and by choosing the appropriate VFO frequency. There is no need to include bands that you are not interested in. Use of an ssb filter, with provisions for selecting sidebands, is not warranted for the devoted cw enthusiast.

System measurements were performed on constructed receivers at various stages of development. The data is summarized in Table 4. Measured data is extended with calculations to give the prospective builder some feel for the performance to be expected.[14] Measurements and calculations generally agreed within 1 dB.

Table 4 reveals no surprises. The nature of the trade-off between single and dual conversion is well illustrated, as is the effect of adding an rf amplifier stage. That system showing the greatest dynamic range is the single-conversion design without the rf amplifier. It should be emphasized that the data in Table 4 pertains to the modules described. Changes in gain, noise figure or intercept of any stage will change the results.

The dual-conversion systems are about 5 dB "weaker" than the single-conversion ones — a typical situation. It should be understood that this observation applies only to dual-conversion systems with a wide bandwidth, first i-f section. Modern systems that use a crystal filter at the first i-f will display performance much the same as that of a single-conversion receiver, even if they use many conversions. Still, traditional dual conversion seems to be a reasonable compromise if the overall gain distribution is well controlled.

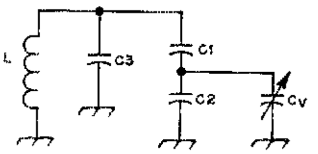
No gain compression was measurable in a single-conversion model without an rf amplifier stage while using an input signal level of −10 dBm. The VFO phase noise was measured to be −152 dBc/Hz at a spacing of 10 kHz. While the writers have access to laboratory-quality instrumentation, all of the evaluations were done with home-built test equipment. This was done to emphasize that a reasonable receiver *can* be built and evaluated without the need for exotic instruments.

The results quoted represent performance that rivals or exceeds that of most commercial equipment available to the amateur. These receivers are pleasing to use, offering a clean, crisp sound that is not always found in the commercial equivalents. Of great significance is that the receiver should be easily duplicated, with a cost well under that of a similar "appliance."

The writers gratefully acknowledge the interest and ideas of many friends. We would especially like to thank Roy Lewallen, W7EL, for all of the data he has shared with us during the two-year period devoted to this project.

Appendix

Calculation of capacitor values for a VFO using an arbitrary capacitor.



In the following equations, the inductance (L) is expressed in henrys, capacitance (C) in farads and frequency (f) in hertz. Follow the sequence of equations as shown. f_b and f_u represent the lower and upper frequency limits to be tuned. C_v is the *range* of the variable capacitor.

$$\omega_b = 2\pi f_b$$

$$\omega_u = 2\pi f_u$$

$$C_b = \frac{1}{\omega_b{}^2 L}$$

$$C_u = \frac{1}{\omega_u{}^2 L}$$

$$\Delta C = C_b - C_u \qquad \text{(Eq. 1)}$$

Pick a value for C1 and C_v. Then let:

$$x = 2C1 + C_v$$

$$y = C1C_v + C1^2\left(1 - \frac{C_v}{\Delta C}\right) \qquad \text{(Eq. 2)}$$

Then, C2 and C3 are given by:

$$C2 = 1/2\left(-x + \sqrt{x^2 - 4y}\right)$$

$$C3 = C_v - \left(\frac{1}{C1} + \frac{1}{C2}\right)^{-1} \qquad \text{(Eq. 3)}$$

Example:

$L = 3 \mu H$, $f_b = 7$ MHz and $f_u = 7.2$ MHz.

Then $C_b = 172.3 \times 10^{-12}$ F
$C_u = 162.8 \times 10^{-12}$ F.

Choose C1 = 50 pF and $C_v = 355$ pF (variation in a 365-pF variable capacitor).

Then, $x = 455 \times 10^{-12}$,
$y = -73.76 \times 10^{-21}$, yielding
$C2 = 126.8 \times 10^{-12}$ F and
$C3 = 127.02 \times 10^{-12}$ F. QST

Notes

[1] R. Lewallen, "An Optimized QRP Transceiver," *QST*, Aug. 1980, p. 14.
[2] Etched circuit boards and many of the required parts are available from Circuit Board Specialists, P.O. Box 969, Pueblo, CO 81002. Pc templates and parts overlays are available from the ARRL for $2 and an s.a.s.e.
[3] R. Hayward and W. Hayward, "The Ugly Weekender," *QST*, Aug. 1981, p. 18.
[4] W. Hayward and D. DeMaw, *Solid-State Design for the Radio Amateur* (Newington: American Radio Relay League, 1977).
[5] See note 1.
[6] See note 1.
[7] Toroid cores used in these receivers are available from Amidon Associates, 12033 Otsego St., North Hollywood, CA 91607.
[8] See note 2.
[9] Fair Radio Sales, P.O. Box 1105, Lima, OH 45805.
[10] See note 4.
[11] See note 1.
[12] Spectrum International, P.O. Box 1084, Concord, MA 01742.
[13] See note 4.
[14] The noise factor formula for two cascaded stages is given in Chapter 6 of reference 4. Also presented are the relationships between intercepts, noise figure and dynamic range. The output and input intercepts of a given stage differ only by the stage gain. Intercepts may be combined for a cascade by noting that the output intercept of one stage adds to the input intercept of the following one in exactly the same way that resistors in parallel combine so long as the intercepts are presented in milliwatts rather than in dBm. This relationship assumes that the third-order intermodulation distortion is coherent. (This relationship is derived in Hayward's ***Introduction to Radio Frequency Design***, Prentice-Hall, Englewood Cliffs, NJ 07632, to be published early in 1982. Also see B. P. Gross, "Calculating the Cascade Intercept Point of Communications Receivers," *Ham Radio*, Aug. 1980, and W. Sabin, "A BASIC Approach to Calculating Cascaded Intercept Points and Noise Figure," *QST*, Oct. 1981.)

Strays

"Thanks for the memories . . ." might better have been sung, "Thanks for jogging the public's memory circuits," as Jim Dudley, W5HYW, presented an award to comedian Bob Hope in appreciation of his efforts in promoting Amateur Radio. A plaque, outlined by Hope's famous profile, was given by the Port Arthur (Texas) ARC on the occasion of the Bum Phillips Celebrity Golf Tournament. Amateur Radio fraternities from Jefferson and Orange Counties, Texas, served as the communications arm of the event. *(photo by WB5YIF)*

APPLICATIONS FOR GENERAL MANAGER, ARRL

The present General Manager of the League is retiring in the spring of 1982, and the Board of Directors seeks a qualified replacement. Those wishing to submit their applications for this challenging post should do so prior to January 1, 1982. Send your resumes and any supporting data to Ray Wangler, W5EDZ, Chairman of the ARRL Management & Finance Committee, 642 Beryl Dr., San Antonio, TX 78213.

The General Manager, ARRL, is charged with managing and directing League operations under policies established by the Board of Directors. He employs and oversees the operations of a staff of 115, ensuring that they maintain professional standards. He maintains sound financial policies and procedures, involving an annual budget of six million dollars. He assists the president in representing the League with national and international government agencies and other Amateur Radio organizations, and serves as secretary of the International Amateur Radio Union. He is responsible for developing effective programs for the growth of the Amateur Radio population and of League membership, which currently stands at over 150,000. He is responsible for initiating plans, programs and policies for the advancement of Amateur Radio which are presented to the Board for approval. He must monitor all aspects of the Amateur Radio Service in order to advise and counsel the Board. He also serves as the Editor of *QST*.

A demonstrated knowledge of management skills, a wide-ranging interest in and knowledge of Amateur Radio and ARRL, and evidence of leadership will be in your favor.