When a receiver is satisfactory in every respect (stability and selectivity) except sensitivity on 14 through 30 MHz, the best solution for the amateur is to add a preamplifier, a stage of rf amplification designed expressly to improve the sensitivity. If image rejection is lacking in the receiver, some selectivity should be built into the preamplifier (it is then called a preselector). If, however, the receiver operation is poor on the higher frequencies but is satisfactory on the lower ones, a converter is the best solution.

Some commercial receivers that appear to lack sensitivity on the higher frequencies can be improved simply by tighter coupling to the antenna. This can be accomplished by changing the antenna feed line to the right value (as determined from the receiver instruction book) or by using a simple matching device.

Gain Control

In a receiver front end designed for best signal-to-noise ratio, it is advantageous in the reception of weak signals to eliminate the gain control from the first rf stage and allow it to run "wide open" all the time. If the first stage is controlled along with the i-f (and other rf stages, if any), the signal-to-noise ratio of the receiver will suffer. As the gain is reduced, the g_m of the first tube is reduced, and its noise figure becomes higher. A good receiver might well have two gain controls, one for the first rf stage and another for the i-f (and any other rf) stages. The first rf stage gain would be reduced only for extremely strong signals, thus assuring a good noise figure.

THE 80-10 FET PRESELECTOR

It is often necessary to put new life into tired or inexpensive receivers, especially when operation is marginal on the three higher hf bands – 14, 21 and 28 MHz. A preselector of the type described here can pep up the front end of such receivers while at the same time offering additional selectivity on all the hf bands. The latter helps to reduce images and generally improves the reception on some of the low-cost receivers. Often, signals heard on the amateur bands actually originate on quite different frequencies. They appear on ham sections of the dial as a result of image reception or overload of a receiver mixer.

Circuit Details

This preselector is self-contained, except for the power supply, and no modifications are required in the receiver used. The diagram of the unit is shown in Fig. 8-37. Input and output tuned circuits consist of the preselector tuning capacitor, C1, and high-Q coils wound on small toroid cores. Each coil has a trimmer capacitor for alignment purposes. A secondary winding is added which serves as the input or output 50-ohm link. Band changing is accomplished by S1, a multipole miniature switch. An "off" position is included so that the preselector may be bypassed when it is not required.

Two JFETs are operated in a cascode circuit. The advantage of this arrangement is that the capacitance between input and output is only a fraction of a picofarad – so low that neutralization is not required in the hf range. Current drain is low, so the preselector may be operated from a 9-volt transistor-radio battery if desired, with only a slight loss of gain and dynamic range. Otherwise, a 12-volt miniature power supply, such as the type sold for battery replacement, should be used. If battery operation is contemplated it would be well to add a power on/off switch; otherwise, current will be drawn all of the time.

The gain for each band has been set at approximately 20 dB by adjustment of the turns ratio on the rf transformers. Although the cascode circuit can provide up to 30 dB of gain, care must be taken so that the preselector does not overload the succeeding stages in the receiver.

Construction

The preselector is built on a 4 X 5-inch etched circuit board which is housed in a 7 X 5 X 3-inch homemade cabinet. The enclosure is made from two U-shaped pieces of aluminum stock. Any of the popular commercially made cabinets may be substituted. Also, point-to-point wiring using terminal strips may be employed in place of the etched board. Whatever the assembly technique chosen, good isolation between the input and output tuned circuits is of prime importance. Any stray coupling can cause instability. If trouble

Fig. 8-36 – The 80- to 10-meter preselector is constructed in a cabinet made from two U-shaped pieces of sheet aluminum. Press-on feet are used. Panel decals (obtained from H. H. Smith) lend a "finished" appearance to the unit.

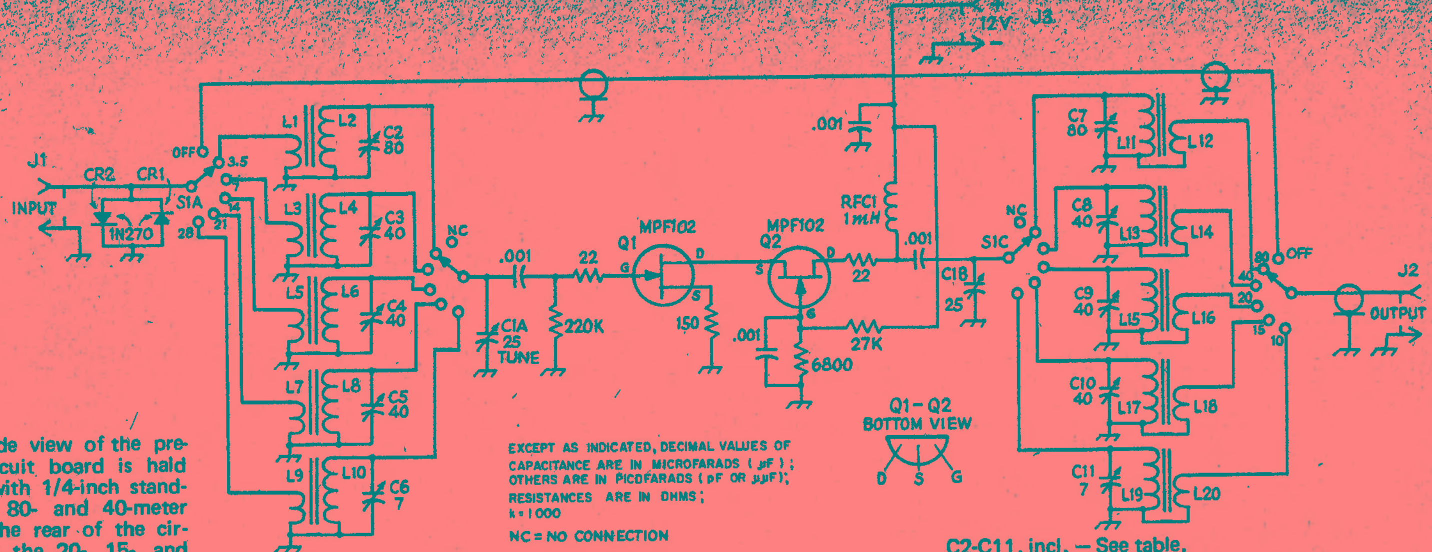


Fig. 8-37 — Schematic diagram of the selective presalector. Unless otherwise indicated, decimal values of capacitance are in μF; others are in pF. Resistors are 1/4- or 1/2-watt composition and fixed-value capacitors are disk ceramic.
C1 — Split-stator variable, duel section (Hemmarlund HFD-25).
C2-C11, incl. — See table.
CR1-CR2 — High-speed switching diodes.
J1-J3, incl. — Phone jack, panel mount.
L1-L20, incl. — See table.
Q1, Q2 — HEP 802 or 2N5486.
RFC1 — Miniature choke (Miller 70F103A1).
S1 — Ceramic miniature rotary switch, 4 pole, 6 position, 2 section (Centralab PA-2011).

Fig. 8-38 — Inside view of the preselector. The circuit board is hald off the chassis with 1/4-inch standoff pillars. The 80- and 40-meter coils are along the rear of the circuit board, with the 20-, 15-, and 10-meter input coils to the right of the switch while the output coils for the 14- to 28-MHz bands are to the left. Q1 is on the bottom left of the circuit board, with Q2 just above. ▼

COIL TABLE

80	L1, L12 — 5 turns No. 30 enam. over L2, L11, respectively.	L2, L11 — 85 turns No. 30 enam. on Amidon T-50-2 core.	C2, C7 — 7-80-pF compression trimmer, Calectro A1-247
40	L3, L14 — 3 turns No. 30 enem. over L4, L13, respectively.	L4, L13 — 40 turns No. 30 enam. on Amidon T-50-2 core.	C3, C8 — 4-40-pF compression trimmer, Calectro A1-246
20	L5, L16, — 2½ turns No. 22 enam. over L6, L15, respectively.	L6, L15 — 20 turns No. 22 enam. on Amidon T-50-2 core.	C4, C9 — 4-40-pF compression trimmer, Calectro A1-246
15	L7, L18 — 2 turns No. 22 enam. over L8, L17, respectively.	L8, L17 — 13 turns No. 22 enam. on Amidon T-50-6 core.	C5, C10 — 4-40-pF compression trimmer, Calectro A1-246
10	L9, L20 — 1½ turns No. 22 enam. over L10, L19, respectively.	L10, L19 — 10 turns No. 22 enam. on Amidon T-50-6 core.	C6, C11 — 0.9-7-pF compression trimmer, Calectro A1-245

Note: Amidon Associates, 12033 Otsego Street, North Hollywood, CA 91607 develops, a shield between Q1 and Q2 may be of help.

The band switch, S1, is mounted on an aluminum bracket which is, in turn, mounted at the center of the circuit board. The toroid coils are held in place with a drop of epoxy cement. The shield that separates the two sections of C1 must be grounded to the etched board with a short lead. This metal strip provides vital shielding between sections A and B. The PRESELECTOR capacitor is mounted directly to the front panel using hardware supplied with the unit. All of the trimmer capacitors are mounted on the circuit board.

During assembly, whether or not a circuit board is used, a heat sink should always be employed when soldering the transistor leads. If excessive heat reaches the body of the transistor, the device can be ruined.

The input, output, and power jacks are mounted on the rear apron of the chassis. The rf protection diodes, CR1CR2, are connected right across J1. Subminiature coax (RG-174A/U) is used to connect the input and output jacks to the circuit board. Sockets for the transistors were included in the original model to facilitate experimentation; they may be omitted if desired.

Alignment

The completed preselector is best adjusted with a signal generator. However, if no test equipment is available, on-the-air signals may be used. The antenna or generator should be connected to J1 and a short patch cord run from J2 to the receiver. Start with the 10-meter band, and set C2 with the plates fully unmeshed. Then tune in a signal at the uppermost point in the band. Adjust trimmers C6 and C11 for maximum indication on the receiver S meter. Repeat this procedure for the other bands, setting the appropriate trimmers. The lower-frequency bands will appear to tune more broadly. However, the selectivity provided by the high-*Q* rf transformers is about the same on each band.

If this preselector is to be used with a transceiver, the unit will have to be switched *out* of the antenna line when transmitting. Otherwise severe damage will result to the coils and transistors in the unit. If the transceiver has a separate receiving-antenna input, as some do, the preselector can be connected to this jack, and the feeder switched with an external antenna-changeover relay.

SQUELCH CIRCUITS

An audio squelch is one that cuts off the receiver output when no signal is coming through the receiver. A squelch is useful in mobile equipment where the no-signal receiver hiss noise may be as loud as some of the weak signals being copied. Noise of this kind, when listened to over a sustained period, can cause considerable operator fatigue. A squelch is useful with certain types of fixed-station equipment too, especially where continuous monitoring of a fixed vhf or uhf frequency is desired.

A practical vacuum-tube squelch circuit is given in Fig. 8-39 at A. A twin triode (12AX7) serves as an audio amplifier and a control tube. When the agc voltage is low or zero, the lower (control) triode draws plate current. The consequent voltage drop across the adjustable resistor in the plate circuit cuts off the upper (amplifier) triode and no signal or noise is passed. When the agc voltage rises to the cutoff value of the control triode, the tube no longer draws current and the bias on the amplifier triode is now only its normal operating bias, furnished by the 1000-ohm resistor in the cathode circuit. The tube now functions as an ordinary amplifier and passes signals. The relation between the agc voltage and the signal turn-on point is adjusted by varying the resistance in the plate circuit of the control triode.

The circuit shown at B employs a Schmitt-trigger input to achieve positive on-off gating of the audio signal. For ssb operation, the length of the squelch-gate on time after the input signal disappears is increased by switching in an electrolytic capacitor. The dc signal from the squelch gate controls an emitter follower which is connected between the receiver detector and the first audio amplifier.

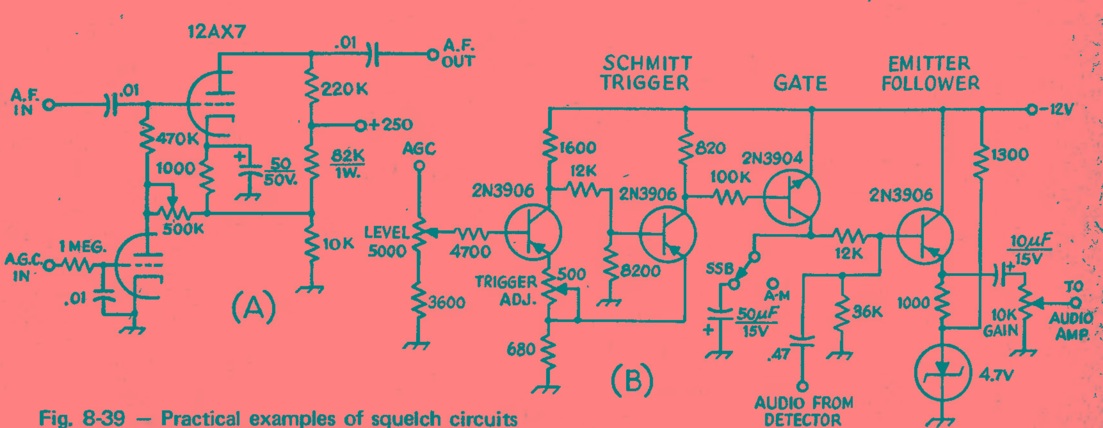


Fig. 8-39 — Practical examples of squelch circuits for cutting off the receiver output when no signal is present.

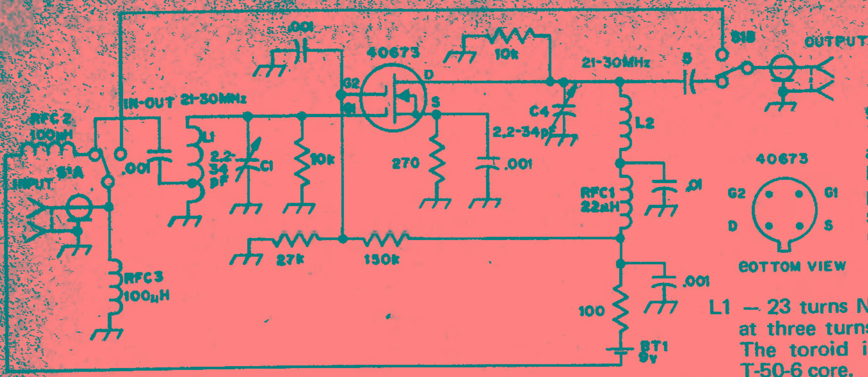


Fig. 8-41 — Circuit diagram for the 15- and 10-meter amplifier. J1 and J2 are BNC type, however conventional phono connectors are satisfactory.

C1, C4 — 2.2 to 34 pF (Johnson 193-0010-001 or equiv.).

L1 — 23 turns No. 26 enam. with a tap at three turns from the ground end. The toroid inductor is an Amidon T-50-6 core.

L2 — 20 turns No. 26 enam. on an Amidon T-50-6 core.

RFCI — 22 µH (Millen J302-22 or equiv.).

Fig. 8-40 — Inside of the receiver preamplifier. The battery is held in position with a clamp fabricated from a small strip of aluminum.

AN RF AMPLIFIER FOR 10 AND 15 METERS

Many receivers are deficient in gain when operated on the higher hf amateur bands. The problem becomes very apparent when the receiver is used as an i-f for vhf converters. The preamplifier given in Fig. 8-41 will provide a gain of about 14 dB and will perk up the performance of almost any receiver on the 15- and 10-meter amateur bands.

With the shield placed between C1 and C4, the preamplifier shows no signs of instability. The range of the capacitor-inductor combinations is sufficient to peak the output on either 10 meters or 15 meters. A band switch is not needed. Some adjustment of the turns spacing on the toroid inductors may be necessary to allow each corresponding capacitor to tune the full frequency range of the unit.

TUNING A RECEIVER

CW Reception

In a receiver without selectivity, it doesn't much matter where the BFO is set, so long as it is within the pass band of the receiver. However, in a receiver with selectivity, the BFO should be offset, to give single-signal code reception. The proper setting of the BFO is easy to find. In the absence of incoming signals, it will be found that, as the BFO control is tuned, the pitch of the background noise will go from high to low and back to high again. The setting that gives the lowest pitch represents the setting of the BFO in the *center* of the pass band. Setting the BFO for a higher pitch (to the noise) will give more or less single-signal effect on incoming signals, depending upon the selectivity of the receiver. If the receiver uses a crystal filter that has a "rejection notch" or "phasing" control, setting the notch on the audio image will improve the single-signal effect.

The best receiver condition for the reception of code signals will have the first rf stage running at maximum gain, the following rf mixer and i-f stages operating with just enough gain to maintain the signal-to-noise ratio, and the audio gain set to give comfortable headphone or speaker volume. The audio volume should be controlled by the audio gain control, not the i-f gain control. Under the above conditions, the selectivity of the receiver is being used to best advantage, and cross modulation is minimized. It precludes the use of a receiver in which the gains of the rf and i-f stages are controlled simultaneously.

Single-Sideband Phone Reception

The receiver is set up for ssb reception in a manner similar to that for single-signal code reception, except that a suitable bandwidth for ssb (2 to 3 kHz) is used. The BFO *must* be set off to one side of the passband if good use is to be made of the selectivity. To determine which side to set it, remember this rule: A selective receiver can be set up for *lower*-sideband reception by setting the BFO so that there is little or no signal on the *low*-frequency side of zero beat when tuning through a steady carrier or cw signal. Lower sideband is customarily used on 3.9 and 7 MHz, upper on the higher frequencies.

Unless the receiver has an agc system suitable for ssb reception (fast attack, slow decay), the operator must be very careful not to let the receiver overload. If the receiver does overload, it will be impossible to obtain good ssb reception.

Run the receiver with as little rf gain as possible, consistent with a good signal-to-noise ratio, and run the audio gain high.

Carefully tune in an ssb signal using only the main tuning dial. When the voice becomes natural sounding and understandable, the signal is properly tuned. If the incoming signal is on lower sideband, tuning the receiver to a lower frequency will make the voice sound lower pitched. An upper-sideband signal will sound higher pitched as the receiver is tuned to a lower frequency.

If the receiver has excellent selectivity, 2.1 kHz or less, it will be desirable to experiment slightly with the BFO setting, remembering that each adjustment of the BFO calls for a similar adjustment of the main tuning control. If the selectivity is quite high, setting the BFO too far from the pass band will limit the incoming signal to the high audio frequencies only. Conversely, setting it too close will limit the response to the low audio frequencies.

A-M Phone Reception

In reception of a-m phone signals, the normal procedure is to set the rf and i-f gain at maximum, switch on the agc, and use the audio gain control for setting the volume. This insures maximum effectiveness of the agc system in compensating for fading and maintaining constant andio output on either strong or weak signals. On occasion a strong signal close to the frequency of a weaker desired station may take control of the agc, in which case the weaker station may disappear because of the reduced gain. In this case better reception may result if the agc is switched off, using the manual rf gain control to set the gain at a point that prevents "blocking" by the stronger signal.

When receiving an a-m signal on a frequency within 5 to 20 kHz from a single-sideband signal it may also be necessary to switch off the agc and resort to the use of manual gain control, unless the receiver has excellent skirt selectivity.

A crystal filter will help reduce interference in phone reception. Although the high selectivity cuts sidebands and reduces the audio output at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility. As in code reception, it is advisable to do all tuning with the filter in the circuit. Variable-selectivity filters permit a choice of selectivity to suit interference conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat note equal to the frequency difference.

Spurious Responses

Spurious responses can be recognized without a great deal of difficulty. Often it is possible to identify an image by the nature of the transmitting station, if the frequency assignments applying to the frequency to which the receiver is tuned are known. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat note with the desired signal and is actually on the same frequency, the beat note will not change as the receiver is tuned through the signal; but if the interfering signal is an image, the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be turned off for this test. Using a crystal filter with the beat oscillator on, an image will peak on the side of zero beat opposite that on which desired signals peak.

Harmonic response can be recognized by the "tuning rate," or movement of the tuning dial required to give a specified change in beat note. Signals getting into the i-f via high-frequency oscillator harmonics tune more rapidly (less dial movement) through a given change in beat note than do signals received by normal means.

Harmonics of the beat oscillator can be recognized by the tuning rate of the beat-oscillator pitch control. A smaller movement of the control will suffice for a given change in beat note than that necessary with legitimate signals. In poorly designed or inadequately shielded and filtered receivers it is often possible to find BFO harmonics below 2 MHz, but they should be very weak or nonexistent at higher frequencies.

ALIGNMENT AND SERVICE OF SUPERHETERODYNE RECEIVERS

I-F Alignment

A calibrated signal generator or test oscillator is a useful device for alignment of an i-f amplifier. Some means for measuring the output of the receiver are required See Fig. 8-40. If the receiver has a tuning meter, its indications will serve. Lacking an S meter, a high-resistance voltmeter or a vacuum-tube voltmeter can be connected across the second-detector load resistor, if the second detector is a diode. Alternatively, if the signal generator is a modulated type, an ac voltmeter can be connected across the primary of the transformer feeding the speaker, or from the plate of the last audio amplifier through a 0.1-μF blocking capacitor. Lacking an ac voltmeter, the audio output can be judged by ear, although this method is not as accurate as the others. If the tuning meter is used as an indication, the agc of the receiver should be turned on, but any other indication requires that it be turned off. Lacking a test oscillator, a steady signal tuned through the input of the receiver (if the job is one of just touching up the i-f amplifier) will be suitable. However, with no oscillator and tuning an amplifier for the first time, one's only recourse is to try to peak the i-f transformer on "noise," a difficult task if the transformers are badly off resonance, as they are apt to be. It would be much better to haywire together a simple oscillator for test purposes.

Initial alignment of a new i-f amplifier is as follows: The test oscillator is set to the correct frequency, and its output is coupled through a capacitor to the grid of the last i-f amplifier tube. The trimmer capacitors of the transformer feeding

Fig. 8-42 — Receiver alignment requires a stable signal source (a signal generator) and voltmeter. Alignment of communications receivers should be checked at least once a year.

the second detector are then adjusted for maximum output, as shown by the indicating device being used. The oscillator output lead is then clipped on to the grid of the next-to-the-last i-f amplifier tube, and the second-from-the-last transformer trimmer adjustments are peaked for maximum output. This process is continued, working back from the second detector, until all of the i-f transformers have been aligned. It will be necessary to reduce the output of the test oscillator as more of the i-f amplifier is brought into use. It is desirable in all cases to use the minimum signal that will give useful output readings. The i-f transformer in the plate circuit of the mixer is aligned with the signal introduced to the grid of the mixer. Since the tuned circuit feeding the mixer grid may have a very low impedance at the i-f, it may be necessary to boost the test generator output or to disconnect the tuned circuit temporarily from the mixer grid.

If the i-f amplifier has a crystal filter, the filter should first be switched out and the alignment carried out as above, setting the test oscillator as closely as possible to the crystal frequency. When this is completed, the crystal should be switched in and the oscillator frequency varied back and forth over a small range either side of the crystal frequency to find the exact frequency, as indicated by a sharp rise in output. Leaving the test oscillator set on the crystal peak, the i-f trimmers should be realigned for maximum output. The necessary readjustment should be small. The oscillator frequency should be checked frequently to make sure it has not drifted from the crystal peak.

A modulated signal is not of much value for aligning a crystal-filter i-f amplifier, since the high selectivity cuts sidebands and the results may be inaccurate if the audio output is used as the tuning indication. Lacking the agc tuning meter, the transformers may be conveniently aligned by ear, using a weak unmodulated signal adjusted to the crystal peak. Switch on the beat oscillator, adjust to the suitable tone, and align the i-f transformers for maximum audio output.

An amplifier that is only slightly out of alignment, as a result of normal drift or aging, can be realigned by using any steady signal, such as a local broadcast station, instead of the test oscillator. One's 100-kHz standard makes an excellent signal source for "touching up" an i-f amplifier. Allow the receiver to warm up thoroughly, tune in the signal, and trim the i-f for maximum output as noted on the S meter, or by tuning for peak af output.

Rf Alignment

The objective in aligning the rf circuits of a gang-tuned receiver is to secure adequate tracking over each tuning range. The adjustment may be carried out with a test oscillator of suitable frequency range, with harmonics from your 100-kHz standard or other known oscillator, or even on noise or such signals as may be heard. First set the tuning dial at the high-frequency end of the range in use. Then set the test oscillator to the frequency indicated by the receiver dial. The test-oscillator output may be connected to the antenna terminals of the receiver for this test. Adjust the oscillator trimmer capacitor in the receiver to give maximum response on the test-oscillator signal, then reset the receiver dial to the low-frequency end of the range. Set the test-oscillator frequency near the frequency indicated by the receiver dial and tune the test oscillator until its signal is heard in the receiver. If the frequency of the signal indicated by the test-oscillator calibration is higher than that indicated by the receiver dial, more inductance (or more capacitance in the tracking capacitor) is needed in the receiver oscillator circuit; if the frequency is lower, less inductance (less tracking capacity) is required in the receiver oscillator. Most commercial receivers provide some means for varying the inductance of the coils or the capacitance of the tracking capacitor, to permit aligning the receiver tuning with the dial calibration. Set the test oscillator to the frequency indicated by the receiver dial, and then adjust the tracking capacitance or inductance of the receiver oscillator coil to obtain maximum response. After making this adjustment, recheck the high-frequency end of the scale as previously described. It may be necessary to go back and forth between the ends of the range several times before the proper combination of inductance and capacitance is secured. In many cases, better overall tracking will result if frequencies near but not actually at the ends of the tuning range are selected, instead of taking the extreme dial settings.

After the oscillator range is properly adjusted, set the receiver and test oscillator to the high-frequency end of the range. First adjust the mixer trimmer capacitor for maximum hiss or signal, then the rf trimmers. Reset the tuning dial and test oscillator to the low-frequency end of the range, and repeat; if the circuits are properly designed, no change in trimmer settings should be necessary. If it is necessary to increase the trimmer capacitance in any circuit, more indicatance is needed; conversely, if less capacitance resonates the circuit, less inductance is required.

Tracking seldom is perfect throughout a tuning range, so that a check of alignment at intermediate points in the range may show it to be slightly off. Normally the gain variation will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned for maximum performance in that region, even though the ends of the frequency range as a whole may be slightly out of alignment.

RECEIVER SELECTION

Beginning amateurs often find themselves faced with the dilemma of choosing between a home-built or store-bought receiver. Ideally, the new ham would elect to build his own complete amateur station, extracting the maximum value from the project through the knowledge he would gain about electronics. Additionally, home-built equipment is more familiar in detail to its owner than is a manufactured receiver. Thus, he can service his unit more rapidly and does not have to consult with the manufacturer about servicing details. If he wishes to add new circuits to the home-built receiver, or to modify existing circuitry, he need not worry about destroying the resale value of the equipment. For this reason the owner may be encouraged to experiment more with circuits, enhancing his overall knowledge of electronics.

Conversely, single-lot quantities of small parts are quite expensive these days, sometimes causing the constructor to spend more money on a simple home-built receiver than he would on a complicated commercially built unit. Modifications to factory-built ham gear generally degrade its resale value, discouraging the owner from making circuit improvements or improving his knowledge by experimenting.

The complexity of the receiver need only be such as to fill the operator's needs. Some very basic home-made receivers perform better than poorly designed multitube commercial units. The receivers described later in this chapter have been designed with the radio amateur's needs in mind, yet no unnecessary circuitry has been added simply to make them appear to be highly sophisticated. Many of the parts used in these receivers can be obtained from junked TV sets, war surplus stores, junked war surplus equipment, and from the workshop junk box. These possibilities should not be overlooked, for a considerable amount of money can be saved by garnering small parts in this manner.

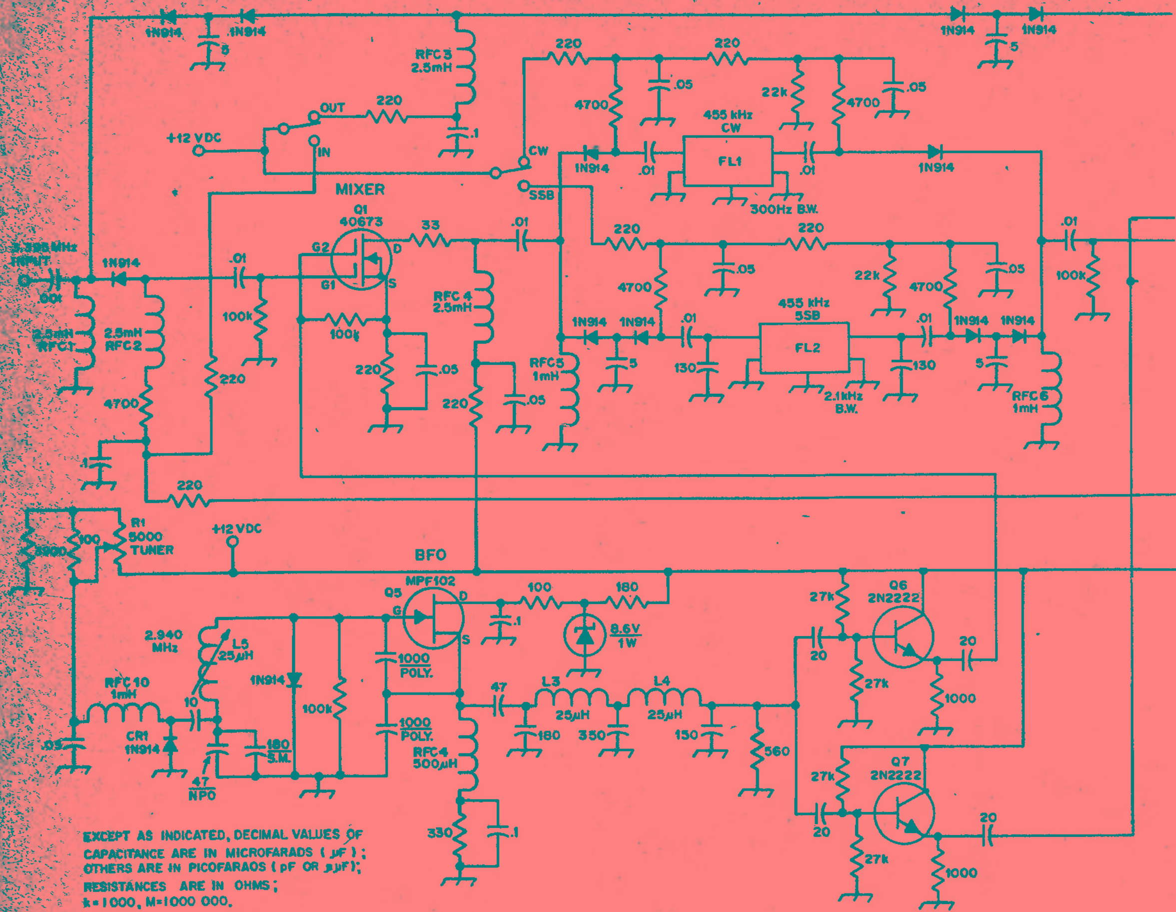
The final decision whether to buy or build will of course be up to the operator. If you're only interested in being a "communicator," then a store-bought receiver will probably suffice. If, however, you want to experience the thrill of communicating by means of home-constructed equipment, and if you want to *learn by doing*, then home-made receiving equipment should be considered. Such forthright endeavors are often the stepping stones to higher plateaus – a satisfying career in electronics, or the needed background to qualify for radio schooling when in the military service. Just having a good working knowledge of one's own station is rewarding in itself, and such knowledge contributes to an amateur's value during public service and emergency operations.

A BAND-PASS TUNER FOR ADJUSTABLE SELECTIVITY

Many modern receivers have fixed-frequency oscillator circuits for injection at the product detector during ssb and cw operation. The frequency of the fixed-frequency oscillator determines the characteristic range of the received signal. For cw operation, it determines the BFO offset frequency and accordingly the pitch of the signals heard. A receiver which has a variable BFO (sometimes called a pitch control) allows the operator to select the pitch at which the cw signals are centered in the filter response. During ssb reception, the control may be used to select a pleasing quality for the incoming signal. A pitch control is undesirable, however, if the BFO in the receiver is to act as a carrier oscillator in the transmitter for transceive operation. In this case, the BFO must be on exactly the same frequency for both receive and transmit periods; indeed many times the same oscillator is used. Transceive operation is not compatible with a pitch control, and unfortunately the operator is required to accept the resultant BFO offset. While it is possible to shift a BFO/carrier oscillator to obtain a desirable receiver response characteristic, the affect to the transmitted signal could be detrimental. These disadvantages can be overcome by incorporating

the circuit shown in Fig. 1 into the station receiver or transceiver. It may be used with any apparatus having an i-f of 3.395 MHz. Other i-f ranges could be used if appropriate changes are made to the Band-Pass Tuner VFO input and output circuits. The experienced builder should not encounter any difficulty.

The purpose of the Band-Pass Tuner is to allow the operator to adjust the range of frequencies



which pass through the filter without changing the frequency of the receiver or the transmit offset frequency. The unit could be built for cw or ssb operation only. Several filters, however, may be included in the circuit and switched to coincide with the desired mode of transmission. When the Band-Pass Tuner is interconnected to a transceiver, one needs to assure that the transmit signal does not pass through the Band-Pass Tuner system. This could be accomplished with either a double-pole relay or a diode switching array as shown in the circuit diagram. Generally speaking, the latter provides more isolation than the mechanical relay and is recommended.

Circuit Functions

The main function of the Band-Pass Tuner is to convert the i-f signal to a different frequency where sharp-skirt filters may be used to increase the selectivity of a receiving system. After the signal is filtered, it is then converted back to the receiver (or transceiver) i-f. The technique used to obtain variable band-pass response is to employ a VFO for the conversion oscillator and use this energy not only to convert the signal from 3.395 MHz down to 455 kHz, but also to convert the signal back to the receiver i-f range. Since the down conversion is equal to the up conversion (the same oscillator is used for both), changing the VFO frequency does not change the frequency of the received signal. The output frequency is always equal to the input frequency; the VFO only changes the position of the signal around the 455 kHz filter system.

A Circuit diagram for the Band-Pass Tuner is given in Fig. 1. Dual-gate MOSFETs are used to accomplish both the down and the up conversion. Since the phone-band mechanical filter has considerable insertion loss (about 10 dB), an RCA 40673 amplifier is included to bring the signal up to the proper level for re-entering the receiver i-f system. The 40673 could be controlled by voltage supplied from the receiver agc bus; however, it was not necessary. Normal agc action of the receiver seemed unaffected by the inclusion of the Band-Pass Tuner.

The VFO must be capable of reaching stability from a cold start in just a few minutes. A VFO which drifts more than a kilohertz during warmup will cause the operator to have to readjust the

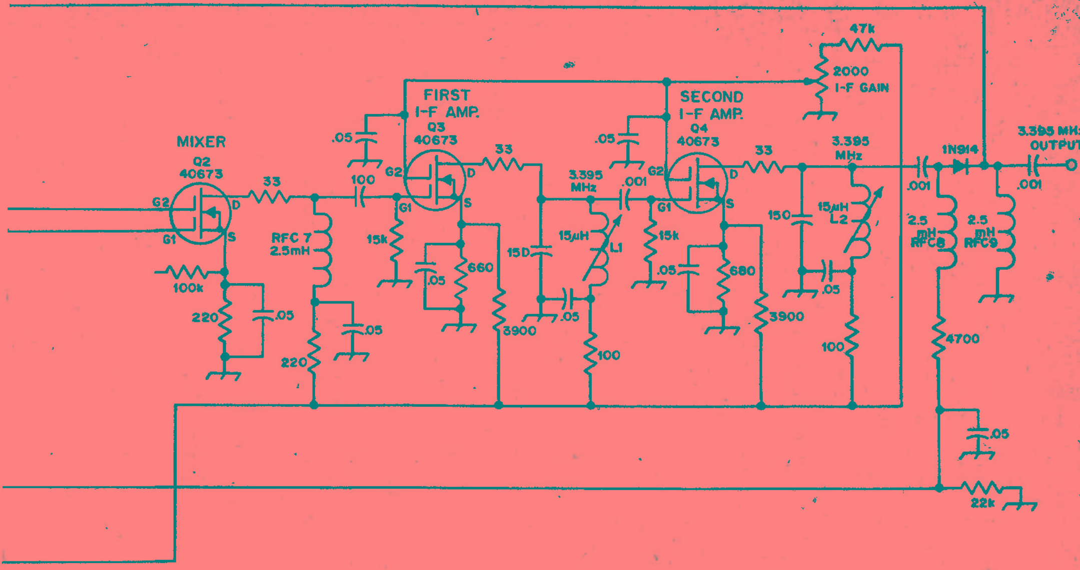


Fig. 1 — Circuit diagram for the Band-Pass Tuner. All resistors are 1/2-watt composition. FL1 and FL2 are Collins types. RFC's are from Millan. All capacitors are disc ceramic. The output of the Pass-Band Tuner should be connected in the receiver at a point before the receiver i-f filter system. The i-f control should be adjusted to give suitebla gain whan the unit is placed in operetion as outlined in the text.

tuning during this period. Since the tuning is accomplished with a varactor diode which has a limited range, drift must be held to within a few hundred Hertz for warmup and within about 25 Hz for normal operation. The voltage source for the VFO must be regulated. CR1 in Fig. 1 serves that function.

Most receivers with a 3.395 MHz second i-f have a first i-f in the range of 8 MHz. Since the third harmonic of the VFO falls in that range, a low-pass filter was incorporated to eliminate any receiver spurious response. The tenth harmonic of the VFO is the only one to fall in an hf amateur band. It was not detectable on the receiver shown in the photographs.

A pair of emitter followers are incorporated in the VFO injection line; each one offers isolation between the VFO and the associated mixer. Additionally, isolation is provided between the input mixer, U1 and the output mixer, U3. Insufficient isolation on this line could cause poor selectivity characteristics (parallel path to the filters) and instability.

Construction

Several methods may be used to enclose the Band-Pass Tuner. If sufficient space is available in the receiver or transceiver, and if the control functions can be made part of the front-panel layout (without drilling holes!), then mounting the circuit boards internally would be desirable. The other option is to fabricate a cabinet or obtain a sufficiently large Minibox to house the circuit boards and controls.

All of the components for the VFO, converters, and amplifiers are included on one board. The filter switching network and the filters are mounted on the same circuit board.

Shielded leads should be used between the input and output points on the circuit board and the station receiver.

Receiver Interconnection and Alignment

The input terminal of the Band-Pass Tuner should be connected to the output of the second mixer in the receiver. The output of the Band-Pass Tuner connects directly to the input of the receiver crystal filter.

Alignment is simple. Before the Band-Pass Tuner is installed in the receiver, tune in the receiver crystal calibrator and log the S-meter reading. Install the Band-Pass Tuner. Set the VFO for 2.940 MHz. A frequency counter is handy for this step, however a general-coverage receiver is suitable. The band-pass tuning control R5, should cause a frequency change of approximately three kilohertz either side of 2.940 MHz.

With the receiver and the Band-Pass Tuner turned on, and the calibrator tuned in, the band-pass control (VFO) should be adjusted to show a peak reading on the S-Meter. Then the gain

Top view of the Band-Pass Tuner

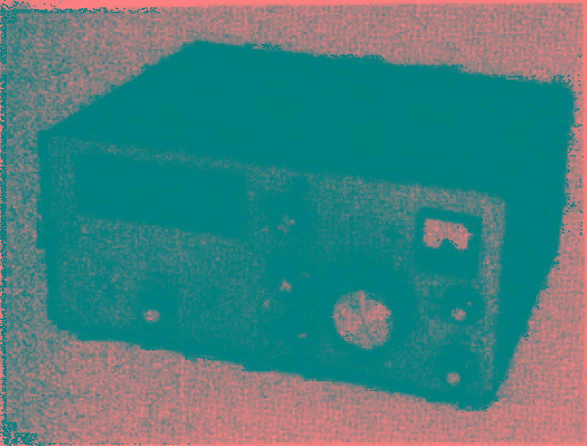
control, R1, should be set so that the receiver S-meter reading under the above conditions is the same as the reading taken before the Band-Pass Tuner was installed. The gain control will no doubt have to be changed during normal operation if more than one filter is used.

Operation

The incorporation of the Band-Pass Tuner does not change the normal operation of the receiver. S1 may be used to select an out condition (Band-Pass Tuner out of the circuit), the ssb filter, or a sharper cw filter. The band-pass tuning control may be adjusted to give the desired filter response in relation to the signal being received.

With most receivers, the change from one sideband to the other at the receiver mode switch will not require a change in the position of the band-pass tuning control. It will, however, change the tuning direction of the response. For instance, if the tuning control is rotated in one direction to favor a higher pitch with usb operation, when used on lower sideband, the same direction of rotation will cause the favored pitch to become lower.

A COMMUNICATIONS RECEIVER WITH DIGITAL FREQUENCY READOUT



Front view of the communications receiver built by WA1JZC. The receiver controls are grouped at the right side of the front panel, with the digital readout occupying the left side.

This solid-state receiver will enable the operator to tune the amateur bands from 160 to 10 meters in 500 kHz segments. An integral part of the unit is a digital frequency counter that may be used to display the received frequency directly to the nearest 100 Hz. The basic receiver consists of a single-conversion tunable i-f which covers a range that includes the 160-meter band. Converters for each additional desired band are placed ahead of the tunable i-f and may be built into the same enclosure. This approach combines the virtues of high performance, moderate complexity, and reasonable cost with plenty of flexibility.

Circuit Overview

The design objectives incorporated in this receiver include optional coverage of all of the amateur bands below 30 MHz, ability to withstand strong signals without cross modulation or overloading, selectable phone or cw bandwidth i-f filters, extensive use of diode switching, and direct display of the frequency of incoming signals. The signal path through the receiver may be traced with the aid of the block diagram, Fig. 1.

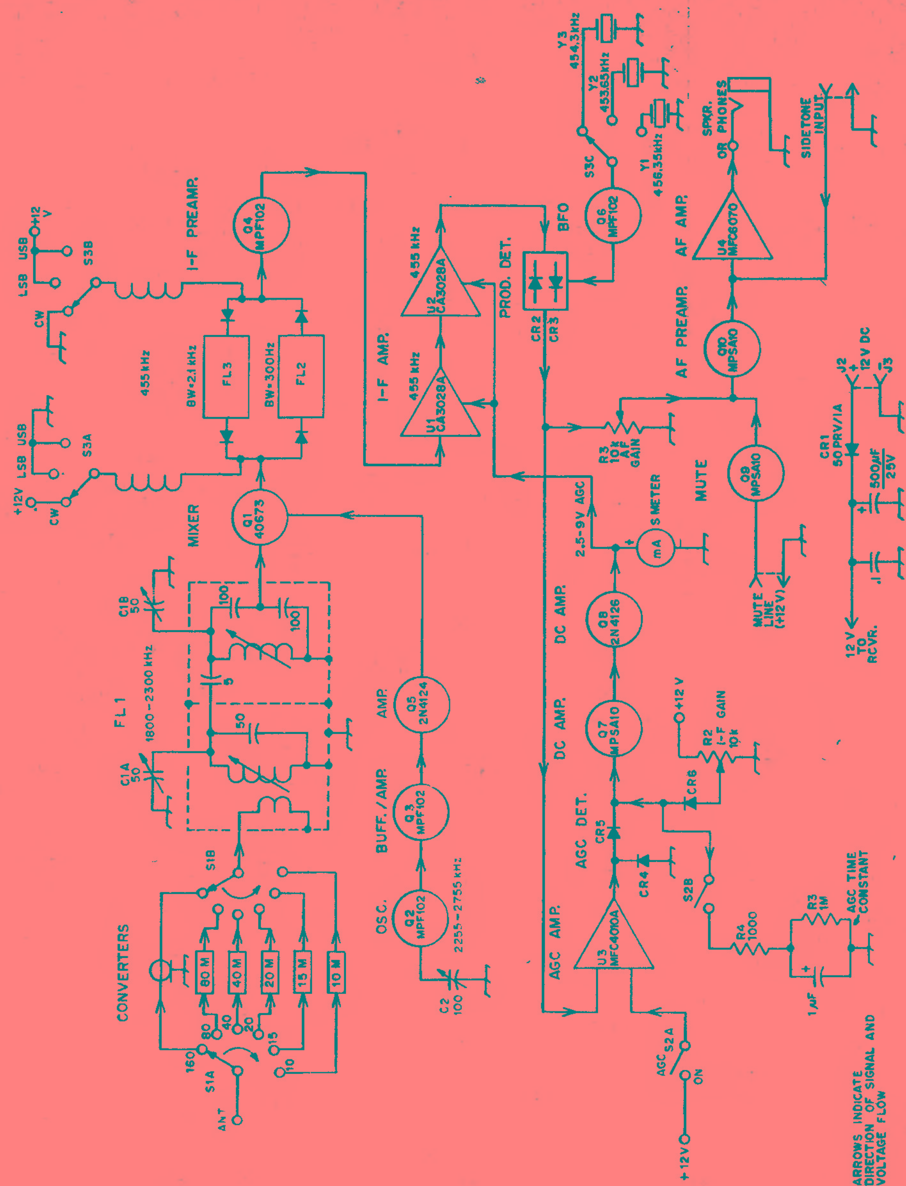


Fig. 1 — Block diagram of the solid-state communications receiver. A polarity-guarding diode, CR1, is shown at the lower right. This diode and the two bypass capacitors associated with it are mounted just inside the receiver cabinet near J2 and J3. These components do not appear in the schematic diagrams of this article.

Band switch S1 connects the rf input, i-f input, and positive supply voltage to the desired converter board. The circuits used in the 40-, 20-, 15-, and 10-meter converters are of similar design. Each one consists of an FET rf amplifier, a crystal-controlled heterodyne oscillator (HFO), and a dual-gate MOSFET mixer stage. No rf amplifier is included in the 80-meter design; an input bandpass network takes its place. The HFO and mixer in the 80-meter circuit are similar to those of the other converters. In each case, the desired output from the mixer is the difference between the incoming signal frequency and the HFO frequency. This difference frequency will fall within the range covered by the tunable i-f, 1800 to 2300 kHz. For 160-meter operation, S1 connects the antenna lead straight through to the input of the tunable i-f, bypassing the converters. No rf amplifier is needed (or used) on 160 meters. In the tunable i-f, a bandpass filter, FL1, is used ahead of a dual-gate MOFSET mixer, Q1. Oscillator injection to Q1, supplied by a three-stage variable frequency oscillator assembly (Q2, Q3, and Q5), may be tuned over a 500-kHz wide range. In this case, the desired mixer output is at 455 kHz. I-f selectivity following the mixer is established by means of a narrow-bandwidth crystal filter, FL2, for cw operation, or by a mechanical filter with a somewhat broader response, FL3, for use with ssb. A diode-switching network permits either filter to be chosen. An i-f preamplifier (Q4) following the filters compensates for filter insertion loss. Two stages of i-f amplifica-

View of the receiver compartment prior to installation of the converter boards.

tion (U1 and U2) after the preamplifier provide ample gain and dynamic range. A passive product detector (CR2, CR3) at the i-f output receives injection from a crystal-controlled BFO, (Q6). The BFO incorporates diode switching of three crystals, one for cw, one for lower sideband reception, and one for upper sideband reception. The audio output from the product detector is split into two independent paths. One path leads to an a-f preamplifier (Q10) followed by an integrated audio power amplifier (U4) which drives a speaker or headphones. The other path runs to an audio-derived agc circuit consisting of U3, CR4 and CR5, Q7, and Q8 plus associated components. The agc is applied to U1 and U2, and additionally provides an S-meter indication on M1.

Tunable I-F Section

A tuning range of 1800 to 2300 kHz was chosen for this portion of the receiver (Fig. 2A). A front-end filter, peaked by means of a front-panel mounted variable capacitor, is used to assure rejection of potentially troublesome out-of-band signals on 160-meters, particularly those of local bc stations. Inexpensive loopstick inductors are used at L2 and L3. It was necessary to remove 125 turns of wire from each so that they would be suitable for tuning from 1800 to 2300 kHz with the split-stator variable capacitor used in the filter. The unloaded *Q* of each, after modification, is 125 at 1.9 MHz. An active mixer was chosen for the first stage of the tunable i-f section to limit the number of receiver stages required for suitable overall gain. Q1, the dual-gate MOSFET, operates with the signal gate tapped down on FL1 to reduce the possibility of over-loading and cross modulation. The conversion gain is roughly 10 dB, and that more than compensates for the insertion loss of the tunable filter, FL1. It is wise to keep the signal voltage at gate 1 as low as practicable to assure good mixer performance. A small amount of forward bias is applied to both gates of the 40673 mixer to increase the linearity and conversion gain of that stage. A 33-ohm resistor is used in the drain to prevent vhf parasitic oscillations.

Local Oscillator

Perhaps the most important consideration when choosing a vfo circuit is its freedom from drift. The incorporation of narrow-bandwidth filters (FL2, FL3) in this receiver makes local oscillator stability a must, as any change in the LO frequency can be easily detected. The local oscillator depicted in Fig. 2B is unconditionally stable, from both a mechanical and an electrical standpoint. Oscillator drift, measured from a cold start, was less than 75 Hz after 30 minutes of operation. VFOs patterned after this circuit have been used in several different applications, and comparable stability has been achieved in every case.

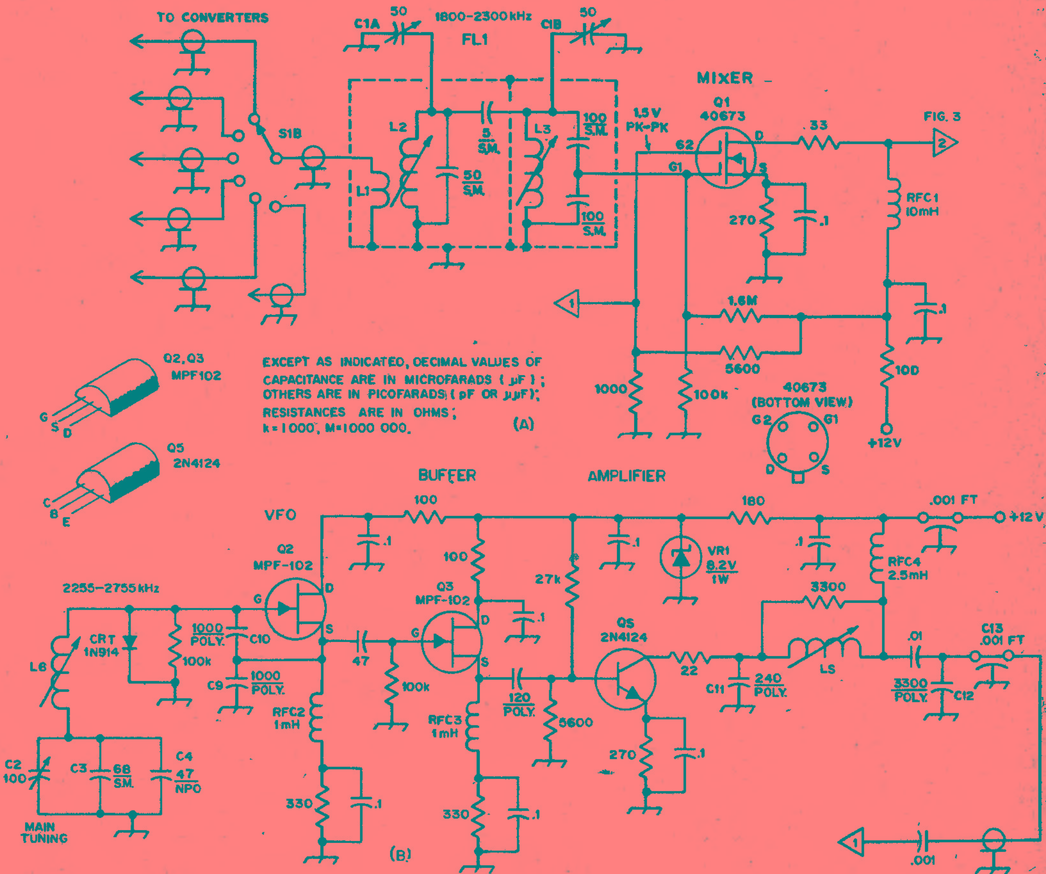
With the circuit constants shown in Fig. 2B, the vfo covers the desired frequency range of 2255 kHz to 2755 kHz, with an additional 15 kHz on either side of that range. VFO linearity across the entire tuning range is not as exacting a requirement when an "electronic dial" is used to display the receiver frequency as it is when some type of mechanical readout is used. Using a 10:1 ratio ball bearing drive to turn the shaft of C2, a tuning rate of 80 kHz per main-tuning-knob revolution is obtained at the low frequency end of the range, and 125 kHz per revolution at the high frequency end of the range. If personal preference dictates a slower tuning rate, dual ratio 36:1 and 6:1 drives are available.

Referring to the circuit diagram, Q2 is used in a series-tuned Colpitts oscillator configuration. Polystyrene capacitors were chosen for their temperature stability at C9 and C10 because these components are part of the frequency-determining network. If the inductor specified for L6 is not available, its replacement must exhibit a similarly high *Q* (approximately 150 measured at 2.5 MHz) to assure proper oscillation. With the exception of C2, C3, C4, and L6, all VFO components are mounted on a single printed-circuit board. C3 and C4 are soldered directly between the C2 stator terminal and the (grounded) capacitor body to minimize lead inductance effects. Inasmuch as the position of the powdered iron core of L6 is variable, the added inductance of the wire from C2 to L6, and from L6 to the pc board is taken into account during the initial VFO alignment. Nevertheless, from the standpoint of mechanical and thermal stability, it is desirable to keep the connecting leads short and direct, using a stiff gauge of wire.

Q3 is connected as a source follower, and is intended to act as a buffer in order to isolate the oscillator circuit from later stages. The buffer drives a single-transistor amplifier stage, Q5. Output from the VFO is taken off of the Q5 collector through a pi-network tank circuit, composed of C11, L5, C12, and C13. Operating voltages for the three transistorized vfo stages are derived from a Zener regulated 8.2-volt bus which is, in turn, derived from the 12-volt receiver power supply. A variation of plus or minus 15% in the receiver supply voltage results in less than a 5 Hz change in the VFO frequency. Additional details concerning the design of the VFO appear in the November, 1974 issue of *QST*, page 22.

Fig. 2 — Schematic diagram of the 160-meter front-end mixer and local oscillator. The mixer is at A, and the oscillator and buffer are at B. Fixed-value capacitors are disk ceramic unless otherwise indicated. Resistors can be 1/4- or 1/2-watt composition types unless specified differently.

C1 — Dual-section 50-pF variable.
C2 — Single-section 100-pF variable (J.W. Miller 2101 or equiv.).
L1 — 3 turns small-diameter insulated wire wound over ground end of L2.
L2, L3 — Radio Shack No. 270-376 ferrite bc antenna with 125 turns of wire removed (see text).
L5 — Pc-mount slug-tuned coil, 10.0 - 18.7 µH (J.W. Miller 23A155RPC).
L6 — Slug-tuned coil — 3/8-inch diameter ceramic form, red core, 18.8 — 41.0 µH (Miller 42A335CBI).
RFC1 — 10-mH miniature encapsulated inductor (Millen Mfg. Co. J302-10,000).
RFC2, RFC3 — 1-mH miniature encapsulated inductor (Millen J302-1000).
RFC4 — 2.5-mH miniature encapsulated inductor (Millen J302-2500).
S1 — Four-pole, six-position rotary switch.
VR1 — 8.2-V, 1-W Zener diode.



I-F Strip

The output of the mixer, Q1, contains not only the desired difference between the VFO and the (post-conversion) incoming signal frequencies, but the sum of these frequencies and higher order products of these frequencies as well. I-f selectivity is developed at the output of the mixer by means of sharp filters resonant at 455 kHz, which both reject the unwanted mixing products and establish the skirt selectivity of the receiver. A choice of i-f bandwidths, one suitable for cw reception and the other tailored for ssb, is made available by the use of two separate filters, FL2 and FL3. FL2 is a crystal filter with a bandwidth of 300 Hz measured at −6 dB from the peak response. FL3 is a 2.1 kHz bandwidth mechanical filter.

A diode switching arrangement, shown schematically in Fig. 4, is used to choose the appropriate filter. When the mode switch (S3) is set to cw, a dc path is created from the positive supply bus through R8, RFC5, CR11, RFC7, RFC8, RFC10, CR12, RFC11, and R9 to ground. Simultaneously, CR9 and CR10 become reverse biased, so that they look like very high impedances. When diodes CR11 and CR12 are forward biased, they appear as very low impedances, thereby opening an rf path from the drain of Q1, through C14, CR11, C15, FL2, C16, CR12, and C17, to the gate of the

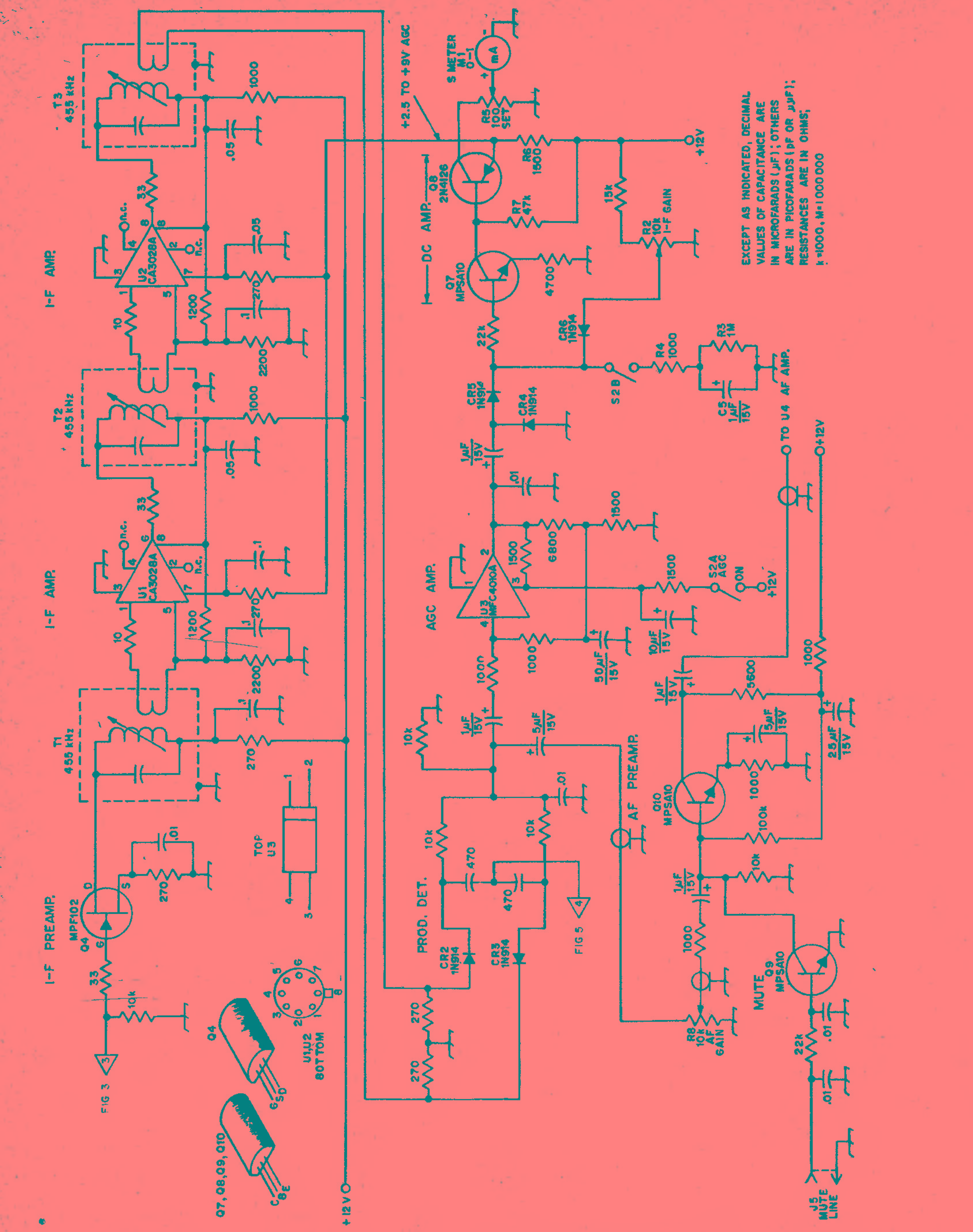


Fig. 3 — Schematic diagram of the i-f, agc, and audio preamplifier circuits. Capacitors are disk ceramic except those with polarity marked, which are electrolytic. Fixed-value resistors can be 1/4- or 1/2-watt composition unless otherwise noted. Numbered components not appearing in the parts list are so numbered for text discussion.

CR2-CR6, incl. — High-speed silicon switching diode, 1N3063, 1N914, or equiv.
J5 — Phono jack, single-hole mount.
M1 — 0 to 1-mA meter (Simpson No. 2121).
R2, R8 — 10,000-ohm linear-taper control.
R5 — 100-ohm pc-board-mount control (Mallory MTC-12L1 or equiv.).
S2 — Two-pole, single-throw toggle. Subminiature type used in this example.
T1-T3, incl.— Single-tuned miniature 455-kHz i-f transformer, 30,000-ohm primary to 500-ohm secondary (Radio Shack No. 273-1383). Use the black core at T1, yellow core at T2, and white core at T3.
U1, U2 — RCA integrated circuit.
U3 — Motorola integrated circuit.

i-f preamp (Q4). When the mode switch is placed in the lower sideband or upper sideband position, CR11 and CR12 become reverse biased, while CR9 and CR10 become forward biased, opening an rf path through FL3, and closing off the path through FL2. With this system, isolation between the filter inputs and outputs as well as between the filters is good, and since the mode switch carries dc only, no special precautions need be taken with the switching lead dress. If insufficient isolation exists between the filters, the characteristics of the narrower filter will be degraded by the wider filter. If this is the case, the use of two series-connected diodes in place of the single diodes on either side of the wider filter (and the possible addition of a small capacitor from the junction of those diodes to ground) should improve the isolation. However, no such degradation was observed when the circuit, as depicted in Fig. 4, was used. The values for C18 and the series combination of C19 and C20 are chosen to resonate with the inductance of the mechanical filter input and output transducers at 455 kHz.

An FET preamplifier stage follows the i-f filter assembly to compensate for the insertion loss of the filters, and establishes the noise figure of the i-f strip. An MPF102 was used at Q4 because of its low noise characteristics. A single-tuned 455-kHz i-f transformer is used to couple the output from Q4 to the input of U1.

Two stages of i-f amplification are provided by U1 and U2 (Fig. 3). RCA CA3028A integrated circuits were chosen for use in the i-f chain because they are inexpensive and easy to work with. In this circuit they are connected as differential amplifiers. Audio-derived agc is applied to terminal 7 of each IC (+2.5 to +9 volts), the constant-current-source bases. The dynamic range of the i-f system is approximately 60 dB.

A passive product detector was chosen over an active one because of its simplicity and good signal-handling capability. A pair of high-speed switching diodes (1N3063) were chosen because of their low cost and easy availibility.

BFO

An MPF102 JFET (Q6) functions in a Pierce crystal controlled BFO (Fig. 5). Separate crystals are used for cw, lower sideband, and upper sideband. Diode switching is used to select the proper crystal, with one section of the mode switch, S3C, performing the function. Again, lead dress to the mode switch is not critical, because only dc is being carried by those leads. The BFO crystal frequency chosen for cw operation is 454.3 kHz, 700 Hz below the center frequency of the i-f. This results in an audio beat note of 700 Hz when a cw signal is peaked in the passband. The lower sideband crystal frequency is 453.650 kHz, and the upper sideband crystal frequency is 456.350 kHz. Note the (suppressed) carrier frequency that the receiver is tuned to changes by 2.7 kHz when the mode switch is changed to the opposite sideband. BFO injection to the product detector is 7 volts peak to peak.

Audio-Derived AGC

Audio output from the product detector is split into two channels, one line feeding the agc strip and the other running to the audio amplifier circuit. An MFC4010A low-cost IC provides 60 dB of gain and serves as the agc amplifier (U3 of Fig. 3). Output from U3 is rectified by means of a voltage doubler consisting of two 1N914 diodes. Because of the high-gain capability of U3 it tends to be unstable at frequencies above the audio range. Addition of the .01-μF bypass capacitor

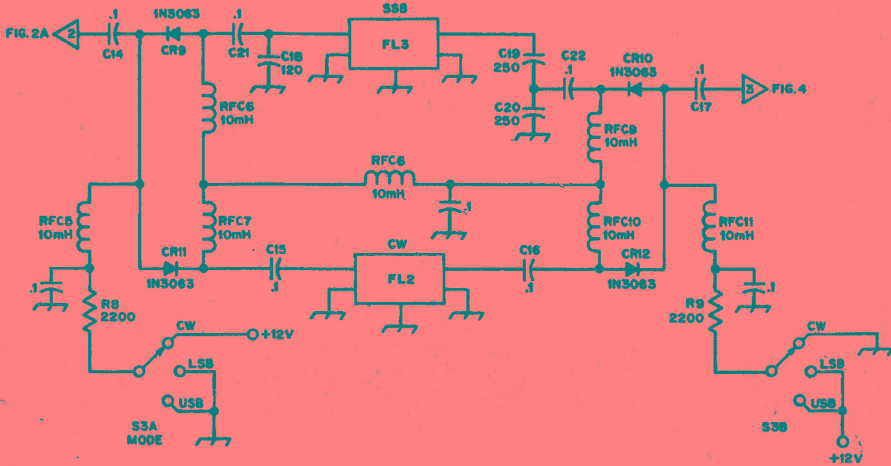


Fig. 4 — Circuit diagram of the i-f filter diode switching network. Numbered components not appearing in the parts list are so numbered for text reference purposes.

FL3 — 2.1-kHz bandwidth mechanical filter, 455-kHz center frequency (Collins F455FA21, Collins Radio Company, 4311 Jamboree Blvd., Newport Beach, CA 92663).

FL2 — 300-Hz bandwidth crystal filter, 455-kHz center frequency (Collins X455KF300, see *QST* Ham-Ads to obtain the names of suppliers).

RFC5 — RFC11 — incl. 10-mH miniature encapsulated inductor (Millen J302-10,000).

S3 — Four pole, three position rotary switch.

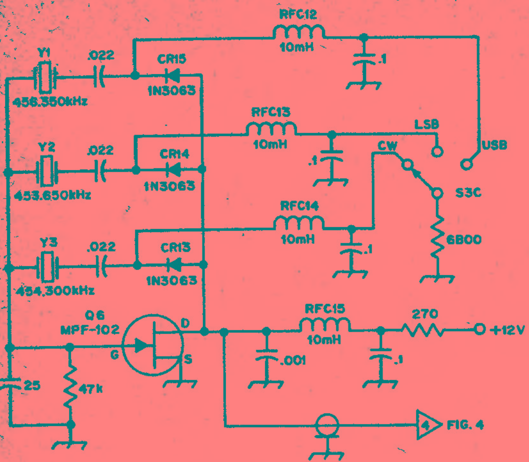


Fig. 5 — Diagram of the receiver beat frequency oscillator showing the use of diode switching of BFO crystals.

CR13—CR15, incl. — High-speed silicon switching diode, 1N3063, 1N914, or equiv.

RFC 12-15, incl. — 10-mH miniature encapsulated inductor (Millen J302-10,000).

Y1 — 456.350-kHz crystal in HC-6/U holder (International Crystal Mfg. Co. type CS, 10 North Lee, Oklahoma City, OK 73102).

Y2 — 453.650-kHz crystal in HC-6/U holder (International type CS).

Y3 — 454.300-kHz crystal in HC-6/U holder (International type CS).

from terminal 2 to ground cured all signs of unstable operation in this circuit. Stubborn cases may require some additional bypassing at terminal 4 of U3. If so, use only that amount necessary to assure stability.

Rectified audio voltage from CR4 and CR5 is supplied to a two-transistor dc amplifier, Q7 and Q8. Agc voltage is taken from the emitter of Q8. Its amount varies with the incoming signal level, and changes as the current-caused voltage drop across the 1500-ohm emitter resistor, R6, shifts in value. S-meter M1 follows the same excursions in current at Q8.

Manual i-f gain control is possible by means of potentiometer R2. It supplies dc voltage to the base of Q7, thereby causing a voltage drop across R7, which causes Q8 to conduct more heavily. As a result, the voltage drop across R6 increases and reduces the agc voltage to lower the gain of the i-f system. The same action takes place during normal agc action. Diode CR6 acts as a gate to prevent the dc voltage provided by CR4 and CR5 from being disturbed by the presence of R2. Maximum i-f gain occurs when the arm of R2 is closest to ground. R3 and C5 establish the agc time constant. The value of R4 can be tailored to provide the attack-time characteristics one prefers. Slower or faster agc time constants can be obtained by changing the values of R3 and C5. The final values will be a matter of operator preference; no two people seem to agree on which time constant is best.

Audio System

Low-cost components are used in the audio system of Fig. 3 and Fig. 6. The circuit performs well and delivers undistorted af output up to one watt in level. An MPSA10 transistor is employed as an audio preamplifier. Muting is provided for by means of another MPSA10, Q9. A positive-polarity voltage is fed to the base of Q9 from the transmitter changeover system to saturate the muting transistor. When in the saturated mode, Q9 shorts out the base of Q10 to silence the receiver. The audio output circuit, U4 of Fig. 6, was borrowed from MFJ Enterprises and is that used in their 1-watt module, No. 1000. Those wishing to do so may order the assembly direct from MFJ.

Provisions are made for feeding a side-tone signal into terminal 3 of U4. This will permit monitoring one's sending even though the receiver is muted by means of Q9. U4 remains operative at all times.

HF-Band Converters

The same pattern is followed for the individual crystal-controlled converters used from 40 through 10 meters (Fig. 7). The 80-meter design is slightly different and is seen in Fig. 8. Separate converters were incorporated to eliminate the need for complicated band switching, and also to permit

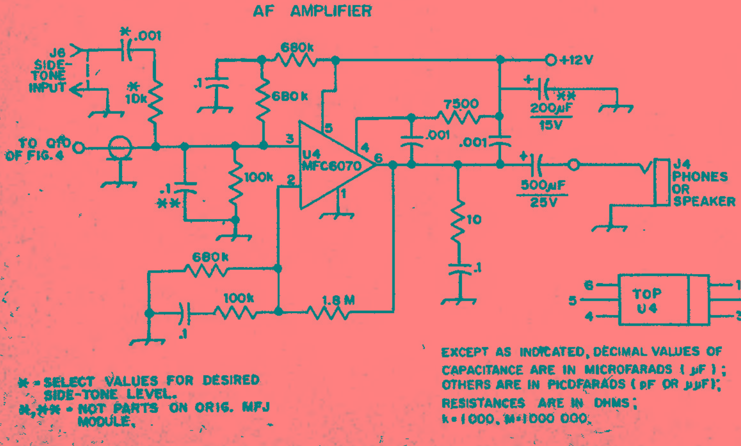
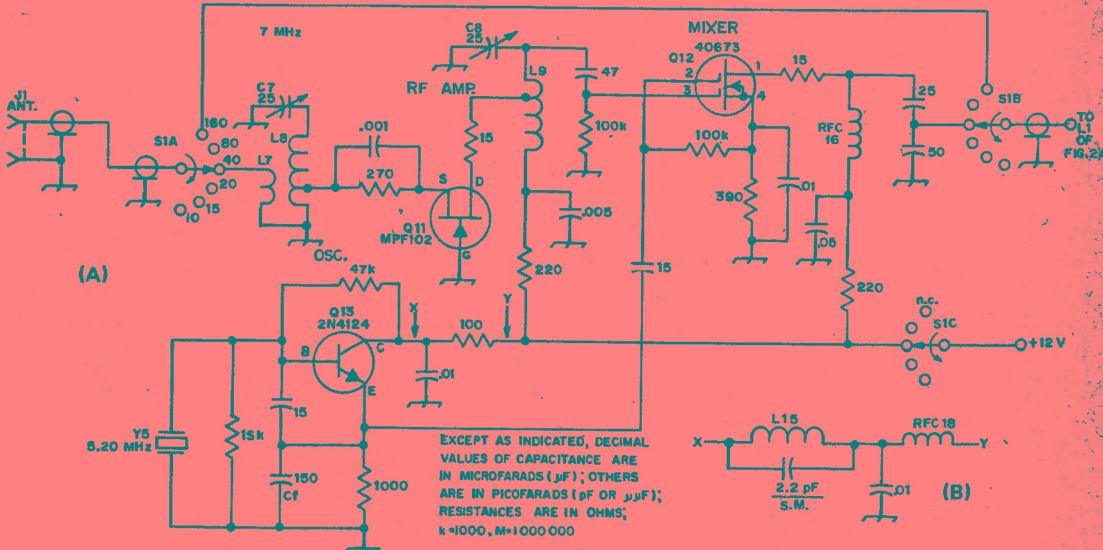


Fig. 6 — Schematic diagram of the MFJ Enterprises 1-watt audio module used in the receiver.

Fig. 7 — Schematic diagram of the 40-meter converter at A, with 10-meter oscillator modification at B. Capacitors are disc ceramic (fixed-value types). Resistors can be 1/4- or 1/2-watt composition types.

C7, C8 — Miniature ceramic or compression trimmer, 5- to 20-pF range.
CF — 39 pF for 20-, 15-, and 10-meter converters.
J1 — Coax connector of builder's choice.
L7 — 2 turns of No. 28 enam. wire over grounded end of L8.
L8 — 40 meters, 50 turns of No. 28 enam. on Amidon T-50-2 toroid core. Tap 8 turns above ground (13 μH, Qu = 180). 20 meters, 44 turns No. 28 enam. on Amidon T-50-6 toroid core. Tap 6 turns above ground (8 μH, Qu = 180). 15 meters, 25 turns No. 28 enam. on Amidon T-50-6 toroid core. Tap 4 turns above ground (4 μH, Qu = 150). 10 meters, 20 turns No. 28 enam. on Amidon T-50-6 toroid core. Tap 3 turns above ground (3.5 μH Qu = 150)
L9 — 40 meters, same as L8, but tap at 25 turns. 20 Meters, same as L8, but tap at 22 turns. 15 meters, same as L8, but tap at 12 turns. 10 meters, same as L8, but tap at 10 turns.
RFC16 — 500-μH miniature encapsulated inductor (Millen J302-500).
Y5 — International Crystal type CS crystal in type FM-1 holder; 40 meters, 5.2 MHz, fundamental mode; 20 meters, 12.2 MHz, fundamental mode; 15 meters, 19.2 MHz, fundamental mode; 10 meters (28.0- to 28.5-MHz coverage) 26.2 MHz, (28.5-to 29.0-MHz coverage) 26.7 MHz, (29.0-to 29.5-MHz coverage) 27.2 MHz, (29.5-to 30-MHz coverage) 27.7 MHz, all third overtone.
L15 — Pc mount slug-tuned coil, 1.5 μH nominal. Miller 46A156CPC or equiv.
RFC18 — 2.2 μH rf choke.



optimization of circuit values for each band of interest. The system used in this receiver calls for switching of only dc and 50-ohm circuitry. Low-impedance switching eliminates problems caused by long switch leads. Switching at high-impedance points, which is the usual technique in multiband receivers, can impair the quality of the tuned circuits and makes isolation of critical circuits more difficult.

A common-gate JFET rf amplifier provides 10 dB of gain in these converters and has good IMD and overload immunity. A 40673 MOSFET is used as the mixer in each converter. Output is taken at the i-f from a broadly resonant circuit formed by a 500-μH rf choke and an rf voltage divider which uses a series capacitor combination (25 and 50 pF). The divider provides a low-impedance pickoff point for the i-f output line to the tunable i-f receiver section.

The 40- through 15-meter converters employ simple Colpitts oscillators. A high-beta transistor is used for the oscillator. It has an f_T of approximately 200 MHz. The circuit for the 10-meter converter oscillator differs slightly from the others in that the HFO uses third overtone rather than fundamental mode crystals, necessitating the insertion of a collector tank circuit tuned to the overtone frequency.

A different design is used in the 80-meter converter, wherein a bandpass filter is used as the input fixed-tuned circuit. This technique was necessary to assure ample bandwidth from 3.5 to 4.0 MHz without the need to have a panel-mounted peaking control. The bandwidth is usable for an 80- and 75-meter frequency spread of 1 MHz.

A Pierce oscillator is used in the 80-meter front-end module to assure plenty of feedback for the 1700-kHz crystal.

Frequency Display Design Approach

The operation of the frequency display may be followed with the aid of the simplified block diagram, Fig. 9. With the conversion scheme utilized in the receiver, the received zero-beat frequency is equal to the sum of the VFO and HFO frequencies minus the BFO frequency, or in the special case of 160-meter operation, the difference between the VFO and BFO frequencies. Accordingly, the display has three identical input

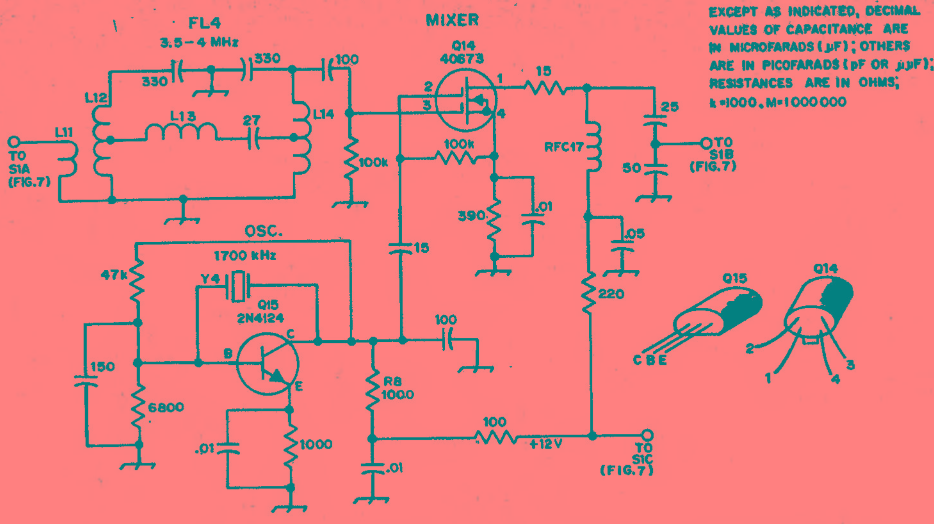


Fig. 8 — Schematic diagram of the 80-meter converter. Capacitors are disc ceramic. Resistors can be 1/4- or 1/2-watt composition types.

L11 — 4 turns No. 28 enam. wire over grounded end of L12.

L12 — 36 turns No. 28 enam. wire on Amidon T-50-2 toroid core (5.5 µH, Q_u = 175). Tap at 18 turns.

L13 — 68-µH miniature rf choke (Q_v of 50 or greater). Millen 34300-68 used in this example.

L14 — Same as L12.

RFC17 — 500-µH miniature encapsulated inductor (Millen J302-500).

Y4 — International type CS crystel in F-700 holder, 1.7 MHz.

networks — one for each oscillator — which amplify and square the incoming waveforms. The outputs of these preconditioning networks are sequentially gated into a chain of seven type SN74192 presettable up/down decade counters, U12-U18. The oscillator frequency addition and subtraction functions are performed by this counter chain. The events in a standard count sequence occur in this order: The counter chain is reset to zero and then placed in the count-up mode. The VFO signal is gated in and counted for 100 milliseconds. The counter is then placed in the count-down mode and the BFO signal is gated in and counted for 100 milliseconds (effectively subtracting the BFO frequency from the VFO frequency digitally). During the next 100 millisecond period, the counter is again placed in the count-up mode and the HFO signal is gated in and counted. At this point, the output of the counter chain ICs represents the receiver zero beat fre-

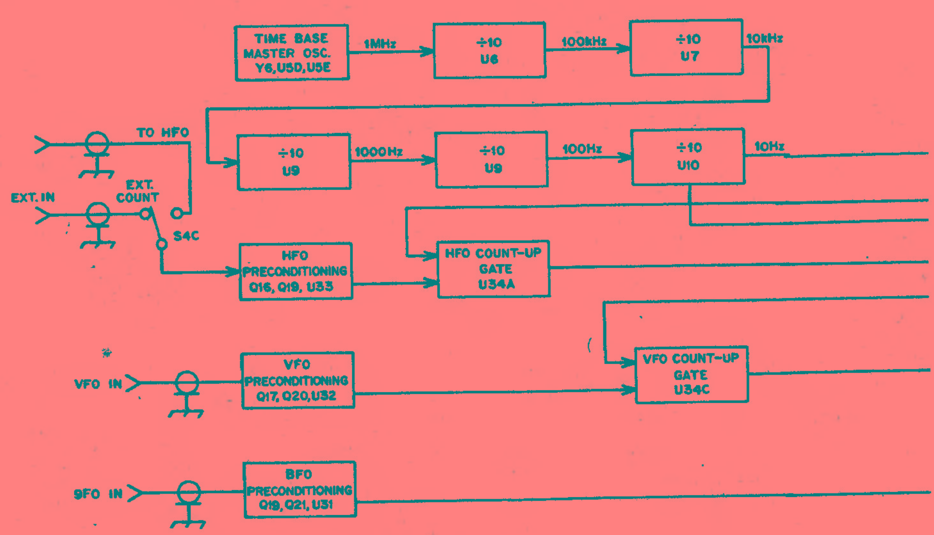
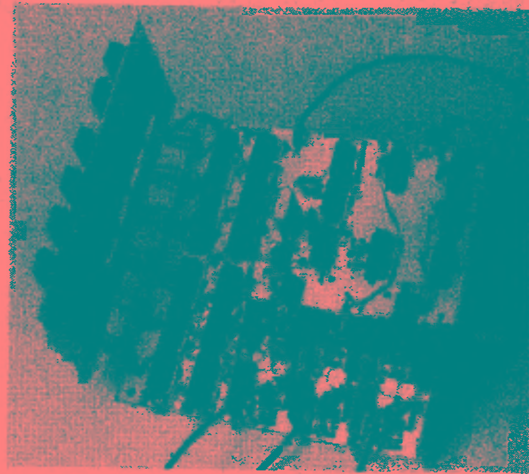


Fig. 9 — Simplified block diagram of the receiver frequency display.

quency. This output is in Binary Coded Decimal (BCD) form. The information from the last six SN74192s in the chain, U13-U18, is stored in six type SN7475 latches (in BCD form) and the entire counter chain is reset to zero during the ensuing 100 millisecond period. The standard count sequence is then repeated. Each of the latches is followed by a type SN7447 IC which decodes the BCD input and provides an output suitable for driving a seven-segment readout. Thus, the display is updated every 400 milliseconds. The output of U12 is not displayed, in order to avoid a distracting last-digit flicker. Accordingly, the 6-digit frequency display reads accurately to the nearest 100-Hz. The timing of the oscillator gating, the latch pulse, and the counter reset pulse is determined by a crystal-controlled oscillator/divider chain consisting of U5D, U5E, and U6-U11.

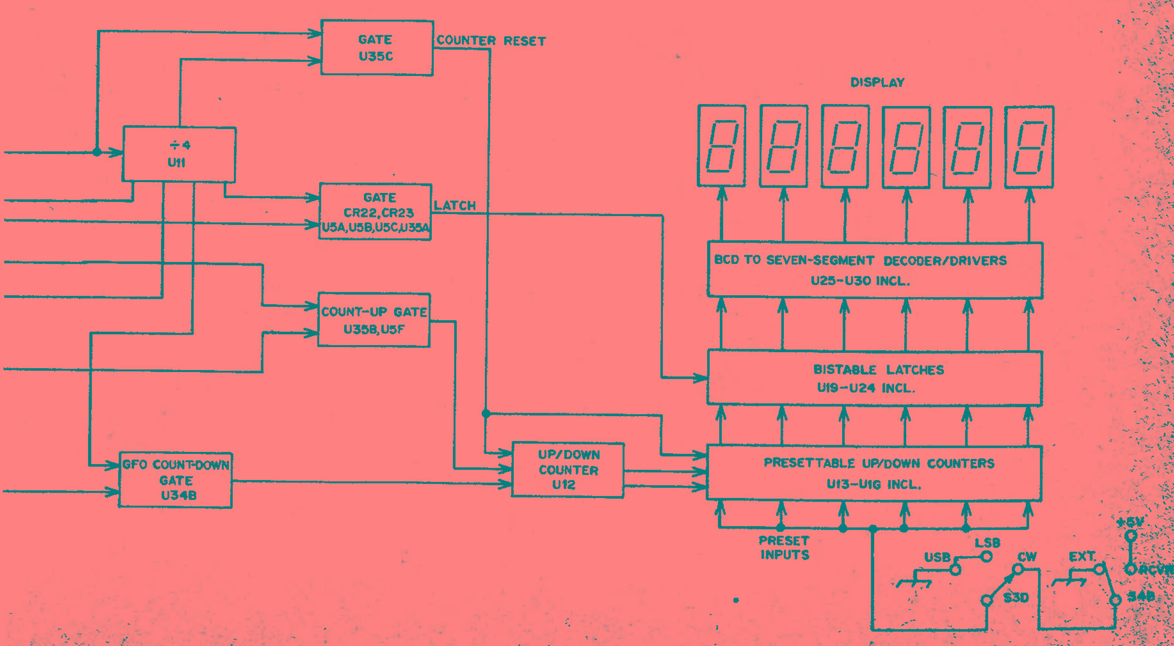
Although this system is satisfactory for ssb reception, it has the drawback that it is necessary for the operator to zero beat an incoming cw signal to read its exact frequency. By taking advantage of the presettable input feature of the SN74192 ICs, the display can be altered during cw reception to read the signal's zero beat frequency while the signal is peaked in the crystal filter passband. Inasmuch as the BFO frequency is 700 Hz below the passband center frequency, in order to obtain the desired readout it is necessary to start counting from "negative 700 Hz" rather than from zero by presetting the counter chain to 999930 every time a count sequence is begun.



Top view of frequency display circuit board. Visible in the foreground are the three shielded input wave-shaping networks.

Frequency Display Operation

Fig. 10 is a complete schematic diagram of the frequency display. The design of each of the input preconditioning networks is identical to that used by Blakeslee (*QST* for June, 1972, pages 31-32). In the HFO shaping network, protection against possible damage to Q16 caused by the application of too great an input voltage is accorded by CR16 and CR17, which conduct if the absolute value of the input signal exceeds approximately 0.6 volts. Q16 and Q19 form a two-stage amplifier that presents a high impedance to the input signal and a low impedance to the succeeding stage. Four sections of a type SN74HO4 high-speed TTL hex inverter are used to convert the incoming sinusoidal HFO signal to a square wave. U33F, operating as an amplifier, drives a Schmitt Trigger composed of U33E and U33D. U33C acts as an output buffer. A type SN7404 can be used in place of the SN74HO4 in the VFO and BFO shaping networks inasmuch as they operate at relatively low frequen-



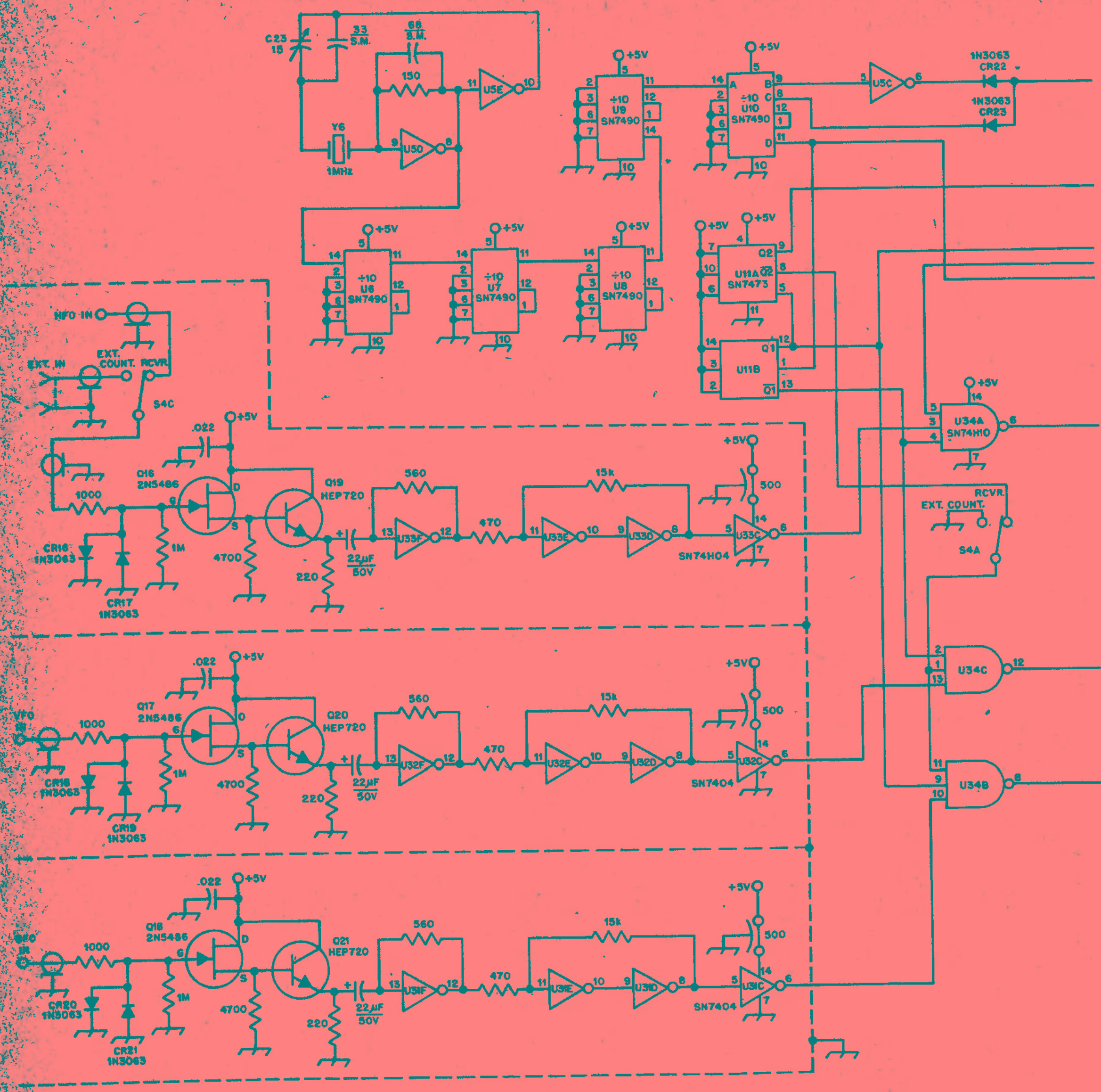
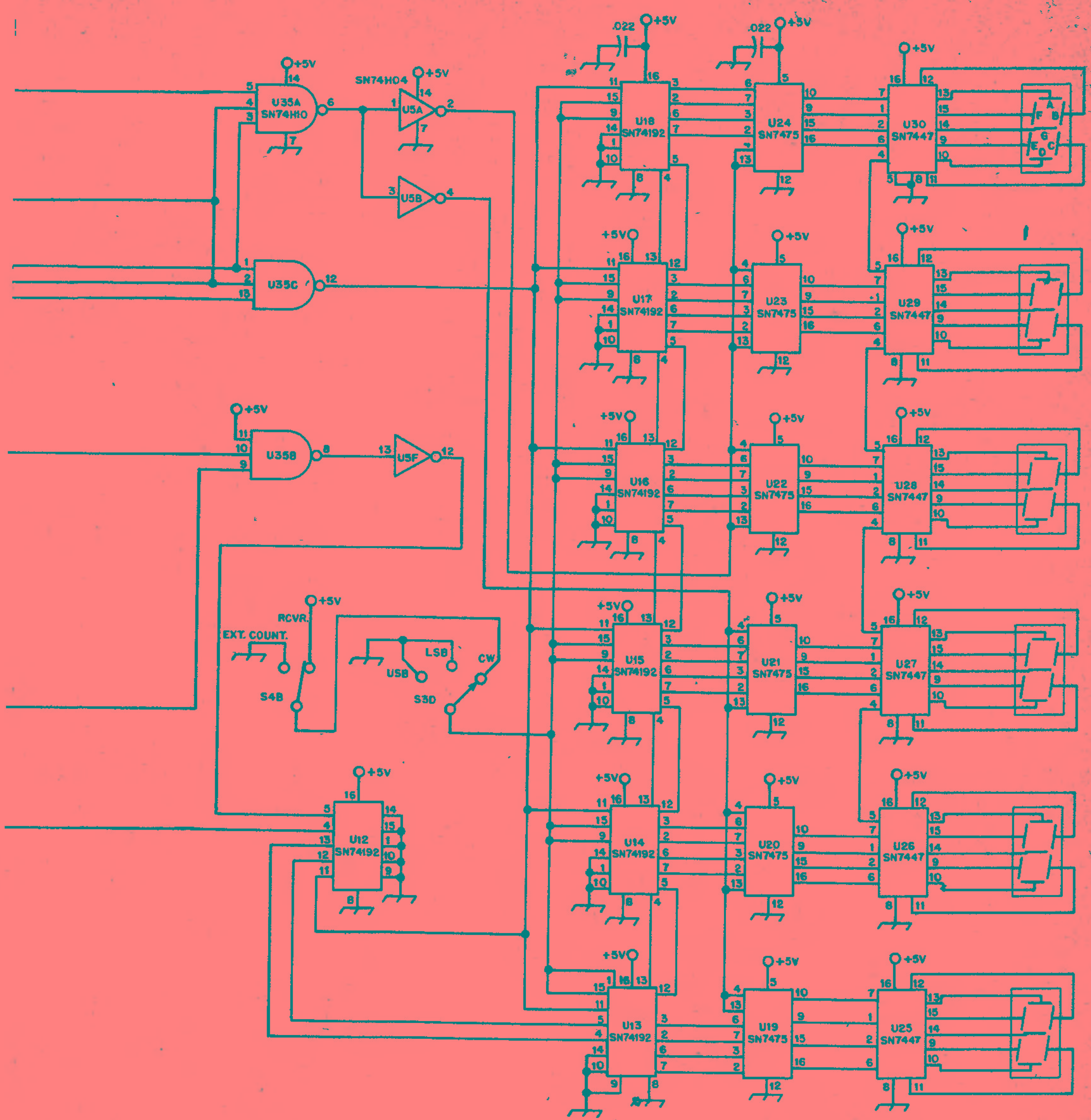


Fig. 10 — Schematic diagram of the receiver frequency display. U5-U30 are Texas Instruments SN7400 series TTL integrated circuits, or equiv.

cies. Each input network is connected to its appropriate oscillator through a short length of RG-174 miniature coaxial cable and a small value coupling capacitor. The smallest value of capacitance that will provide reliable counting should be employed. For example, 27 pF was used to couple from the VFO. Each heterodyne oscillator crystal will require its own coupling capacitor, and one of the receiver band-switch decks may be used to select the proper capacitor for each band. For use of the counter to display an external signal, such as a transmitter's frequency, the HFO input network is switched to receive signals from a front-panel mounted BNC jack, by means of S4C. The typical input circuit sensitivity is 50 to 100 mV.

The three preconditioning networks operate continuously and independently of each other. The shaped oscillator signals are gated sequentially to the input of the counter chain, each for a 100 millisecond interval. Both the duration of the interval and the order in which the receiver oscillators are sampled are governed by a crystal-controlled time base. Two sections of a type SN7404 hex inverter (U5D and U5E), a 1-MHz crystal (Y6), plus a handful of other parts constitute the master oscillator. C23 is a pc-mount air



variable capacitor that permits an incremental adjustment of the oscillator frequency.

The master oscillator is followed by a frequency divider network, U6-U11. U6-U10 are type SN7490 decade counters. Each SN7490 is composed of a divide-by-two section and a divide-by-five section. In this application, the sections are cascaded resulting in divide-by-ten operation. Four outputs are available from each SN7490 connected in this configuration, corresponding to a BCD representation of the number of input pulses previously applied to the IC, from zero to nine and then back to zero again. These outputs are labeled "A" (pin 12), "B" (pin 9), "C" (pin 8), and "D" (pin 11) representing the 2^0, 2^1, 2^2, and 2^3 bits respectively. For every ten pulses applied to the input (pin 14), one pulse appears at the "D" output, giving the effect of dividing the input frequency by ten (see Fig. 11). Five such divide-by-ten stages are cascaded in the time base resulting in division by 100,000 – with an input frequency of 1 MHz from the master oscillator, the frequency of the square wave appearing at the "D" output of U10 is 10 Hz. A slight shift of the master-oscillator frequency away from 1 MHz, due to perhaps temperature or voltage fluctuations, will also show up at U10, but divided by 100,000 – a good reason for starting with a high crystal frequency. It is obvious that with large-amplitude harmonic-rich square waves of several different frequencies present in the time base, the frequency display must be well shielded from the receiver in order to prevent the appearance of strong birdies all across the dial.

The output pulses from U10 are asymmetric – low for 80 milliseconds and high for 20 milliseconds. U11 is a type SN7473 dual flip-flop with both sections connected for divide-by-two operation and cascaded. Each time a negative-going pulse edge from U10 appears at the clock input of the first section of U11 (pin 1), the Q1 output (along with its complement, $\overline{Q1}$) changes state. Since U11 encounters a negative-going pulse edge once every 100 milliseconds, the Q1 output is alternately low for 100 milliseconds and high for 100 milliseconds. The Q1 output is used to establish the duration of the receiver-oscillator sampling interval.

Four distinct 100 millisecond intervals are necessary (VFO count, BFO count, HFO count, latch and reset). A means for distinguishing between these four intervals is provided by dividing the time base frequency (at Q1) by two in the second section of U11. Thus the Q2 output of U11 (along with its complement $\overline{Q2}$) is alternately low for 200 milliseconds and high for 200 milliseconds. The four distinct states occur with Q1 low and Q2 low, Q1 high and Q2 low, Q1 low and Q2 high, and Q1 high and Q2 high. An equivalent description of these states is that $\overline{Q1}$ and $\overline{Q2}$ are high, Q1 and $\overline{Q2}$ are high, $\overline{Q1}$ and Q2 are high, and Q1 and Q2 are high respectively. The sequential gating to the counter of the preconditioned VFO, BFO, and HFO signals is performed by U34, a type SN74H10 triple three-input NAND gate. The output of a three-input positive logic NAND gate is high under all input conditions except when all three inputs are in a high state simultaneously, when the output is forced low. The preconditioned VFO signal is applied continuously to one of the inputs of U34C. The other two inputs are tied to the $\overline{Q1}$ and $\overline{Q2}$ outputs respectively of U11. Whenever either $\overline{Q1}$ or $\overline{Q2}$ or both $\overline{Q1}$ and $\overline{Q2}$ are low, the output of U34C is held high. During the 100 millisecond long interval when both $\overline{Q1}$ and $\overline{Q2}$ are simultaneously high, however, the output of U34C follows the excursions of the VFO input, going low when the square wave VFO input is high and vice versa. Similarly, there are inputs to U34B from Q1, $\overline{Q2}$, and the BFO; to U34A from $\overline{Q1}$, Q2, and the HFO. In each case, when a signal is not being gated by a section of U34, the output of that section is held high.

The ICs used in the counter chain are SN74192 presettable up/down decade counters. Like the SN7490, each SN74192 has four terminals, "A" (pin 3), "B" (pin 2), "C" (pin 6), and "D" (pin 7) for BCD output. Additionally, the IC has "A" input, "B" input, "C" input", and "D input" terminals (pins 15, 1, 10, and 9 respectively) which may be used to initialize or preset the BCD output of the counter to some particular desired nonzero state from which point the count begins. For example, to load the preset inputs with the count of nine (the BCD representation of nine is 1001) the A and D inputs would be tied to +5 volts and the B and C inputs would be grounded. The information at the preset inputs is transferred to the output terminals when a negative-going pulse is applied to the counter reset or load terminal (pin 11), and as long as this terminal is held low, the counter is inhibited from counting pulses applied to either of its clock inputs. Each SN74192 has two clock inputs, one labeled count up (pin 5) and the other labeled count down (pin 4). When pin 4 is held at +5 volts and pulses are applied to pin 5, each incoming pulse increases the BCD output by 1 count. Likewise, when pin 5 is at a high logic level and pulses are applied to pin 4, each incoming pulse decreases the BCD output by 1 count. A single SN74192 can count from 0 to 9. If a greater number of pulses is to be counted, several of the ICs may be cascaded. Borrow and carry output terminals (pins 13 and 12 respectively) are provided for this purpose. To cascade several counters, the borrow output of the first IC is connected to the count down input of the next IC in line and

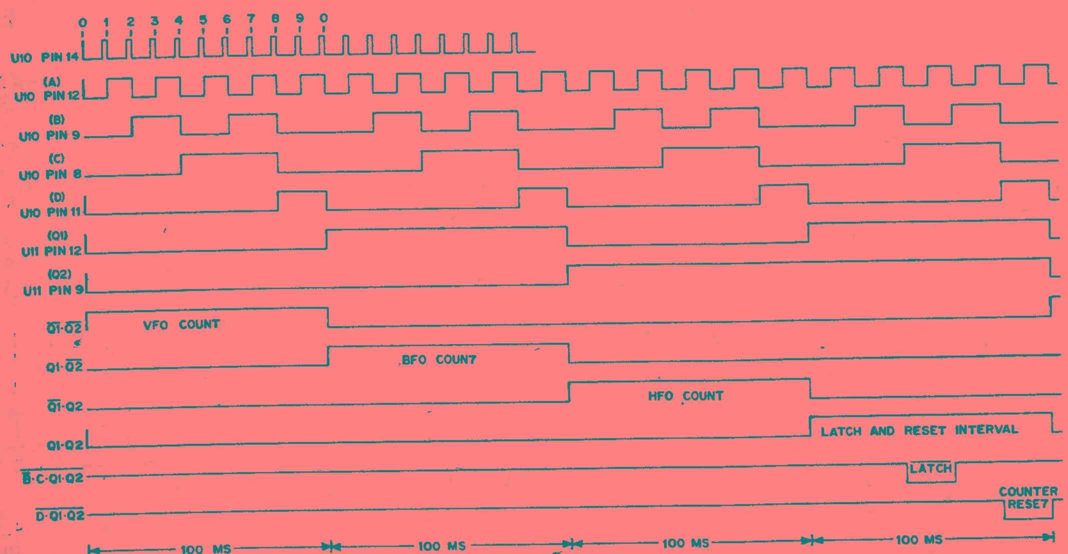


Fig. 11 — Chart showing waveforms associated with a standard count sequence.

the carry output is connected to the count up input of the next IC. In this configuration, if the output of a counter is nine and one pulse is applied to the count up input, the output of the first counter will go to zero and one pulse will be transferred to the count up input of the next counter in line. The case of the count down and borrow terminals is analogous.

In this frequency display, seven counters are cascaded with the result that a maximum of 9,999,999 counts can be registered before the chain resets to zero. If a 10-MHz signal was gated into the counter chain for 100 milliseconds, one million counts would be registered. The output of the last counter in the chain would be 1, while the other counter outputs would be zero. Thus, with a 100 millisecond sampling interval, the last counter indicates the tens of Megahertz, the counter immediately before indicates the units of Megahertz, the counter before that indicates the number of hundreds of kilohertz, and so on up to the first counter which indicates the number of tens of Hertz. Because of the plus-or-minus one count accuracy in the sampling process, which takes place three times in the making of each combined count, there is a tendency for the first counter in the chain to reflect this uncertainty by taking on several different nearby values even though the input frequency is kept constant. For this reason, only the outputs of U13-U18 are displayed on the seven-segment readouts (the display reads to the nearest 100 Hz rather than to the nearest 10 Hz). The maximum rated clock input frequency of the SN74192 is at least 20 MHz, but it is listed in the data books as typically 30 MHz, so it may be necessary to hand pick the first IC in the chain for proper high frequency operation.

The outputs of the VFO and HFO gates (U34C and U34A) are combined by U35B and U5F, and applied to the count up input of U12, while the output of the BFO gate (U34B) is applied directly to the count down input of U12, so that during the first 100 milliseconds of a count sequence the counter chain counts up the VFO frequency, during the second 100 ms period the BFO frequency is counted down, and during the third 100 ms period the HFO frequency is counted up. The result of this activity is that at the end of 300 ms, the outputs of the counter chain represent the receiver zero beat frequency. During the fourth 100 ms period, the outputs of the SN74192s are stored in six SN7475 latches, and then after 80 ms of this period have passed, the counter ICs are reset to zero (or to their preset values) by a 20 ms long reset pulse applied to pin 11. The entire sequence then repeats. The reset pulse is derived from the time base. U35C is one section of a SN74H10 triple three-input NAND gate. The inputs to U35C come from the "D" output of U10, and the Q1 and Q2 outputs of U11. By referring to Fig. 11, it can be seen that the only time that these three inputs are simultaneously high is during this 20 ms interval, forcing the output of U35C and the reset inputs of U12-U18 low.

When the receiver mode switch (S3) is in the cw position, the preset inputs of U13-U18 are set to 999930, while for ssb operation all of the preset inputs are set to 0. S4 chooses either receiver frequency display or external frequency count operation. When set to external, the counter preset inputs are set to 000000 regardless of the receiver mode switch position, the input to the HFO preconditioning network is removed from the HFO and connected to the front panel BNC jack, and U34B and U34C are disabled so that only the external signal source is counted up. The rest of the counter functions are unchanged for external operation.

The action of the SN7475 quad latches (U19-U24) is straightforward. Each SN7475 has four one-bit latches, each with an input and output terminal plus two clock lines (one for each pair of latches). When the clock line is at a high logic level, the logic level applied to each input will appear at its own output, and the output will follow any changes in the input. When the clock line is at a low logic level, the level that was present at each output at the time of the high to low transition of the clock pulse remains at that level regardless of the state of the input. In this application, the BCD output from each of the counter ICs, U13-U18, is applied to the inputs of a SN7475 latch. During the 300 ms long counting period, the clock line is kept at a low level. About midway into the fourth 100 ms period in the standard sequence, the clock line goes positive for 20 ms, transferring the BCD information present at the counter outputs to the latch outputs, and then the clock line goes negative again before the reset pulse occurs. The latching pulse is derived from the time base. Referring to Fig. 11, the pulse labeled $\overline{\text{LATCH}}$ is the output from U35A when the "C" output of U10, the Q1 and Q2 outputs of U11, and the complement of the "B" output of U10 are at a high logic level simultaneously. Note that in order to "synthesize" a four-input NAND gate from a three-input NAND gate, diodes CR22 and CR23 were added in front of one of the SN74H10 inputs. The LATCH pulse is obtained by inverting the $\overline{\text{LATCH}}$ pulse. U5A drives the clock lines of three of the SN7475s, while U5B drives the other three.

Each of the SN7475s drives a type SN7447 BCD to seven segment decoder/driver. The function of these ICs (U25-U30) is to convert the BCD code from the latches into a form suitable for driving seven segment display devices.

Construction Details

The completed receiver was assembled inside a "wrap around" style cabinet (LMB CO-1) measuring 6-1/2 × 14-1/2 × 13-inches (HWD). A homemade aluminum inner enclosure, attached to the front panel, was partitioned with an aluminum shield into two equal-size compartments – one side to house the receiver boards and the other to house the frequency display. The liberal use of perforated aluminum screening in the top and outside walls of the frequency counter compartment allows for adequate ventilation of that subsystem. No ventilation of the receiver section was deemed necessary as no heat-generating components are involved there.

The physical location of the front-panel controls was dictated as much by functional requirements as by operational considerations. The layout depicted in the photograph of the front panel is compact yet uncluttered. As far as possible, each individual control is placed near the circuit being controlled. A "commercial look" is imparted to the unit by the use of Kurz-Kasch series 1657 and 700 knobs. A Digbezel model 930-70 (Nobex Components, 1027 California Drive, Burlingame, CA 94010) filter/bezel assembly measuring approximately 5-1/4 X 1-3/4-inches is used to frame the seven-segment readouts. A red filter was chosen for this application, but filters are available for the bezel in amber, green, and neutral tints as well.

Printed circuit boards are used extensively in the construction of this receiver. The pattern used for interconnecting the twenty-five ICs on the main frequency-display board is sufficiently complex to warrant the use of photosensitive resist or silkscreening techniques in the preparation of the board. While the less complex layout of the receiver boards does not require the use of such methods of pc fabrication, the high quality of the results that may be obtained more than justify the slight additional costs involved. G-10 glass-epoxy board, 1/16-inch thick, with 1-oz. (per square foot) copper is used in all cases.

With the exception of U25 through U30 and the seven-segment displays, all of the frequency counter components are mounted on a 6 × 6-inch double-sided pc board. The pattern of interconnecting copper paths is etched on the underside of the board, while the top side is left as a solid ground plane, broken only at the places where component leads project through the board. If sockets are used, short bus-wire jumpers from the donut pads provided on the underside of the board to the top foil may be used to ground the necessary IC pins. Alternatively, individual component leads may be grounded by directly soldering them to the foil. A No. 65 drill was used for the IC and transistor pins, while the remaining holes were drilled with a No. 60 bit. The removal of copper foil from around the ungrounded leads is best accomplished by drilling those holes first, and then scraping away the top ground foil around each hole with a hand-held large-diameter drill bit (1-4-inch is satisfactory for the purpose). This process may be speeded up by the use of a drill press, although extreme caution must be exercised in order to prevent the drill bit from going completely through the board. The holes for grounded leads may then be drilled.

The liberal use of bypass capacitors in combination with a low-impedance ground return is vital in order to prevent clock pulses, switching transients, and other "garbage" from travelling along the dc supply lines to the sensitive counter input wave-shaping networks, resulting in improper operation. The configuration of the +5-volt dc supply bus resembles a five-toothed comb of jumper wires, laid across the board's upper side. The far end of each of the "teeth" is bypassed with a .022 μF disc ceramic capacitor. Shield compartments made from inch-high strips of double-sided pc board isolate the three input-shaping networks from each other and from the rest of the counter. Signal leads from the three receiver oscillators are carried to the input networks via short lengths of miniature coaxial cable (RG-174 or equiv.), and the dc supply line to each compartment passes through a .001-μF feedthrough capacitor.

Space is at a premium inside the input compartments, requiring the use of miniature vertical-mount electrolytics and 1/4-watt resistors. Low-profile sockets (Texas Instruments type C93) were used with all of the ICs on the counter board. The clock oscillator crystal, Y6, was soldered directly into the circuit with its case grounded to the top foil. A three-section, two-position rotary switch (S4), mounted with an "L" bracket to the chassis bottom to the rear of the counter board, serves as the external count/receiver count switch. Input signals from external sources are routed from the front-panel-mounted BNC jack to S4 via a length of RG-174. Standoff insulators, 1-1/2-inch high, are used to support the counter board at its corners.

The display board containing U25 through U30, as well as the seven-segment readout devices, is mounted vertically behind the front panel in such a manner that only the readouts are visible through the bezel. This board is supported partly by the stiffness of the leads that interface it with the main board, and partly by means of two No. 6 screws and metal spacers affixed to the panel. No circuit board pattern is supplied for the display board because of the wide variety of display devices and BCD decoder/driver ICs available to the builder at modest cost. LED readouts, fluorescent readouts and gas-discharge devices, to name a few, are competitive in price to the incandescent seven-segment displays that were used here, and each type has its own pin-numbering scheme and driver requirements, as well as physical mounting requirements.

The receiver is built from a number of subassemblies, with the principal one incorporating the better part of the tunable i-f. This 6 × 6-inch pc board contains the front-end mixer stage, the i-f filters and amplifiers, the product detector, and the BFO. Additionally, it supports the local oscillator module. Separate and smaller boards are used for the agc and audio circuits and each of the converters. A complete set of templates for the boards used in the receiver and in the frequency display is available from ARRL for $2 and a self-addressed, stamped business size envelope.

Physically small components are employed wherever possible, resulting in compact board layouts. Vertical format low voltage electrolytics such as the Sprague 503D series are to be preferred over axial lead types, which must be mounted upright to fit on the board. Sprague Hypercon low-voltage ceramic disc capacitors are ideally suited due to their miniature size for use wherever a 0.1- or .01-μF value is called for. Millen J302 series encapsulated inductors are used liberally. They are compact and are designed for high-density pc mounting. Either quarter-watt or halfwatt resistors can be used, although with the larger size it may be necessary to stand some of the resistors on end. T1-T3 are inexpensive imported 455-kHz i-f transformers that come four to a package. International Crystal type F-605 nylon sockets were used for Y1-Y3, which are not available with wire leads. All of the semiconductor devices used in the receiver are soldered directly into the circuit board, although sockets are available for U1 and U2 if desired.

An outboard power supply capable of providing 12 volts at 150 mA maximum for the receiver and 5.0 volts at 1.5A for the frequency display is required. It is adviseable to use separate receiver and display supplies rather than a common transformer and rectifier feeding two voltage regulators in order to minimize the chances of coupling counter "hash" into the receiver via the +12-volt supply line. A front-panel-mounted miniature toggle switch is used to apply 117-v ac to the remote supply transformer primaries. Proper bypassing of all supply leads entering the cabinet will aid in rejection of extraneous signals. A loudspeaker may be installed in the same cabinet as the power supply for operating convenience.

The local oscillator enclosure is made from four pieces of double-sided circuit board material soldered together at the edges. The assembly measures approximately 2-1/4 × 1-3/4 × 4-1/8-inches (HWD), and is bolted down to the main receiver board by four spade lugs. The main tuning capacitor, C2, is first attached to one of the side walls of the compartment with the two specially threaded machine screws provided with the capacitor. A 3/8-inch diameter hole drilled in the front compartment wall provides adequate clearance for the capacitor shaft. L6 is mounted on the same wall as C2, centered in the enclosure and directly above the VFO printed-circuit board which is affixed by soldering its ground foil along three edges to the enclosure walls. A 10:1 reduction epicyclic drive. Jackson Brothers number 5857 (available from M. Swedgal, 258 Broadway, New York, NY 10007) is used to turn the main-tuning capacitor shaft. This drive unit requires 1-1/8-inch clearance behind the front panel. A one-piece nickel-plated brass coupling is used between the drive and the capacitor shaft. Four 2-inch long No. 6 machine screws are used to support the large receiver board, which must be installed at the height that permits precise alignment between these shafts.

The main receiver board and the agc/audio preamp boards are fabricated from double-sided copper clad material. All of the other receiver boards use single-sided stock. The agc/audio preamp board and the MFJ audio board (available from MFJ Enterprises, P.O. Box 494, Mississippi State, MS 39762) are mounted with No. 6 hardware and half-inch spacers on one of the aluminum enclosure walls. The input bandpass filter, FL1, is enclosed in an alumimum Minibox measuring 1-1/8 × 2-1/8 × 3-1/4-inches HWD (Bud CU-2117A) which is mounted to the opposite enclosure wall. All interconnections between boards are made with RG-174 miniature coaxial cable. Spade lugs support the converter boards from the rear wall of the receiver compartment.

Alignment

The only adjustment that should be necessary with the frequency display is the setting of the time base crystal oscillator frequency to 1 MHz. A clip lead from the antenna input jack of a general-coverage receiver loosely coupled to the oscillator will pick up enough harmonic energy to allow for precise zero beating of the oscillator frequency to WWV at 10 MHz. A nonmetallic alignment tool should be used to turn the shaft of C23, which should provide a wide enough incremental frequency adjustment.

The 160-meter tunable i-f should be adjusted before the converters are checked out. After verifying with an oscilloscope or a general-coverage receiver that the VFO is indeed oscillating, the VFO tuning range may be set properly with the aid of the frequency display (set to external count). With the plates of C2 fully meshed, the core of L6 should be moved in or out until the frequency display reads approximately 2240 kHz. Then note the displayed frequency with C2 rotated to minimum capacitance, which should be close to 2770 kHz. If the upper frequency obtained departs significantly from the above reading, the slug of L6 should be moved to a point that provides approximately equal amounts of overlap on either side of the desired 2255-2755 kHz range.

The BFO should be checked for proper operation. At this point, the display may be connected in the receiver count mode. If all is well it will indicate a received frequency between 1800 and 2300 kHz. With a signal generator set to 1800 kHz connected to the input of FL1 and the plates of C1 fully meshed, a beat note should be heard in the speaker or headphones as the receiver is tuned to the signal generator frequency. L2 and L3 may now be adjusted to peak the signal on the S-meter or in the loudspeaker. If these coils are not carefully tuned the two filter sections will not track well, resulting in an undesirable broad response or perhaps even a double peak. Now, tune T1, T2, and T3 for maximum signal output. Spot checks across the 1800 to 2300 kHz range can be made to assure that the input bandpass filter is optimized and is tunable across the entire frequency spread. Check to see that the mode switch selects both the proper crystals in the BFO and the correct i-f filter. All that remains to be adjusted now is the S-meter control, R5. With the agc on, but with the i-f gain (R2) set at minimum sensitivity, adjust R5 to give full-scale deflection of M1. This procedure will complete the tune-up of the main portion of the receiver.

Checkout of the converters is similarly easy. The trimmers should be adjusted for peak response at the center of each frequency band of interest. A signal generator is useful for this procedure, although on-the-air signals, preferably weak ones, will suffice.

Alignment of the receiver is complete at this point. The frequency display should be operative and accurate on each band.

VHF and UHF Receiving Techniques

Adequate receiving capability is essential in vhf and uhf communication, whether the station is a transceiver or a combination of separate transmitting and receiving units, and regardless of the modulation system used. Transceivers and fm receivers are treated separately in this *Handbook*, but their performance involves basic principles that apply to all receivers for frequencies above 30 MHz. Important attributes are good signal-to-noise ratio (low noise figure), adequate gain, stability, and freedom from overloading and other spurious responses.

Except where a transceiver is used, the vhf station often has a communications receiver for lower bands, with a crystal-controlled converter for the vhf band in question ahead of it. The receiver serves as a tunable i-f system, complete with detector, noise limiter, BFO and audio amplifier. Unless one enjoys work with communications receivers, there may be little point in building this part of the station. Thus our concern here will be mainly with converter design and construction.

Choice of a suitable communications receiver for use with converters should not be made lightly, however. Several degrees of selectivity are desirable: 500 Hz or less for cw, 2 to 3 kHz for ssb, 4 to 8 kHz for a-m phone and 12 to 36 kHz for fm phone are useful. The special requirements of fm phone are discussed in Chapter 14. Good mechanical design and frequency stability are important. Image rejection should be high in the range tuned for the converter output. This may rule out 28 MHz with receivers of the single-conversion type having 455-kHz i-f systems.

Broad-band receiving gear of the surplus variety is a poor investment at any price, unless one is interested only in local work. The superregenerative receiver, though simple to build and economical to use, is inherently lacking in selectivity. With this general information in mind, this section will cover vhf and uhf receiver "front ends" stage by stage.

RF AMPLIFIERS

Signal-to-Noise Ratio: Noise of one kind or another limits the ability of any receiving system to provide readable signals, in the absence of other kinds of interference. The noise problem varies greatly with frequency of reception. In the hf range man-made, galactic and atmospheric noise picked up by the antenna and amplified by all stages of the receiver exceeds noise generated in the receiver itself. Thus the noise figure of the receiver is not of major importance in weak-signal reception, up to at least 30 MHz.

At 50 MHz, external noise still overrides receiver noise in any well-designed system, even in a supposedly "quiet" location. The ratio of external to internal noise then drops rapidly with increasing signal frequency. Above 100 MHz or so external noise other than man-made is seldom a problem in weak-signal reception. Noise characteristics of transistors and tubes thus become very important in receivers for 144 MHz and higher bands, and circuit design and adjustment are more critical than on lower frequencies.

The noise figure of receivers using rf amplifiers is determined mainly by the first stage, so solving the internal-noise problem is fairly simple. Subsequent stages can be designed for selectivity, freedom from overloading, and rejection of spurious signals, when a good rf amplifier is used.

Gain: It might seem that the more gain an rf amplifier has, the better the reception, but this is not necessarily true. The primary function of an rf amplifier in a vhf receiver is to establish the noise figure of the system; that is, to override noise generated in later stages. One good rf stage is usually enough, and two is the usual maximum requirement.

Once the system noise figure is established, any further gain required may be more readily obtained in the intermediate frequency stages, or even in the audio amplifier. Using the minimum rf gain needed to set the overall noise figure of the receiver is helpful in avoiding overloading and spurious responses in later circuits. For more on rf gain requirements, see the following section on mixers.

Stability: Neutralization or unilaterialization (see chapter on semiconductors) may be required in rf amplifiers, except where the grounded-gate circuit or its tube equivalent is used. Amplifier neutralization is accomplished by feeding the energy from the output circuit back into the input, in such amount and phase as to cancel out the effects of device capacitance and other unwanted input-output coupling that might cause oscillation or other regenerative effects. Inductive neutralization is shown in Fig. 9-1B and C. Capacitive arrangements are also usable. Examples of both will be seen later in this chapter.

An rf amplifier may not actually oscillate if operated without neutralization, but noise figure and bandwidth of the amplifier may be better with

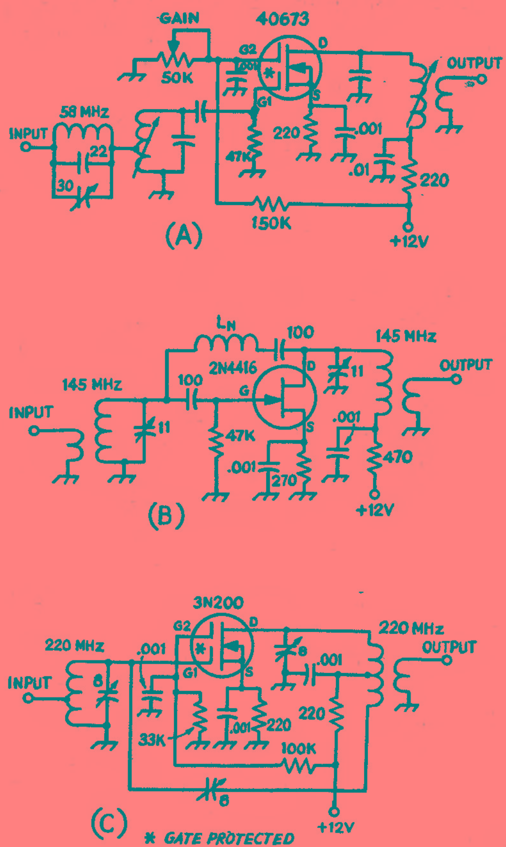


Fig. 9-1 — Typical grounded-source rf amplifiers. The dual-gate MOSFET,A, is useful below 500 MHz. The junction FET,B, and neutralized MOSFET,C, work well on all vhf bands. Except where given, component values depend on frequency.

Using RF Preamplifiers

It is important to design the front-end stages of a vhf receiver for optimum performance, but we often want to improve reception with equipment already built. Thousands of fm receivers formerly in commercial service, now revamped for amateur work in the 50-, 144- and 420-MHz bands, were built before modern low-noise tubes and transistors were available. Though otherwise useful, these receivers have excessively-high noise figure. Many other commercial and home-built vhf converters and receivers are also not as sensitive as they might be.

Though it would be better to replace the rf stages of such equipment with more modern devices, the simpler approach is usually to add an outboard rf amplifier using a low-noise tube or transistor. In the fm example, the quieting level of some receivers can be improved by as much as 10 dB by addition of a simple transistor amplifier. Similar improvement in noise figure of some receivers for other modes is also possible; particularly band-switching communications receivers that have vhf coverage.

Common circuits for rf preamplifier service are shown in Figs. 9-1, 2 and 3. Examples of amplifier construction are given later in this chapter. Circuits shown in the vhf converters described can also be adapted to preamplifier service.

Circuit discussion is cumbersome if we use strictly-correct terms for all tube and transistor amplifiers, so tube terminology will be used here for simplification. The reader is asked to remember it. Any neutralization adjustment reacts on the tuned circuits of the stage, so the process is a repetitive cut-and-try one. The objective should be greatest margin of signal over noise, rather than maximum gain without oscillation. A noise generator is a great aid in neutralization, but a weak signal can be used if the job is done with care.

Overloading and Spurious Signals: Except when some bipolar transistors are used, the rf amplifier is not normally a major contributor to overloading problems in vhf receivers, though excessive rf gain can cause the mixer to overload more readily. Overloading is usually a matter of mixer design, with either transistors or tubes. Images and other spurious responses to out-of-band signals can be kept down by the use of double-tuned circuits between the rf and mixer stages, and in the rf amplifier input circuit. In extreme cases, such as operation near to fm or TV stations, coaxial or other high-*Q* input circuits are helpful in rejecting unwanted frequencies.

Amplifier Circuitry

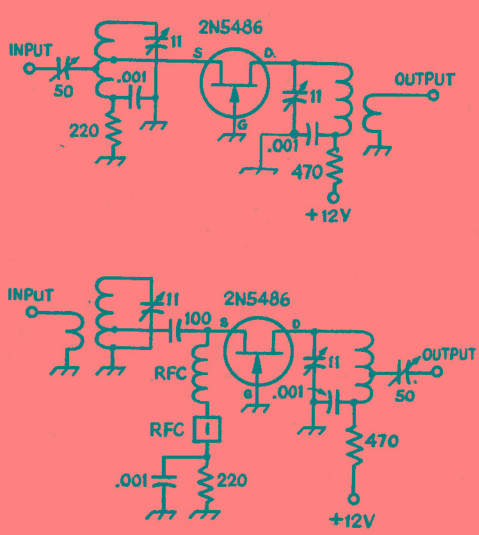


Fig. 9-2 — Grounded-gate FET preamplifier tends to have lower gain and broader frequency response than other amplifiers described.

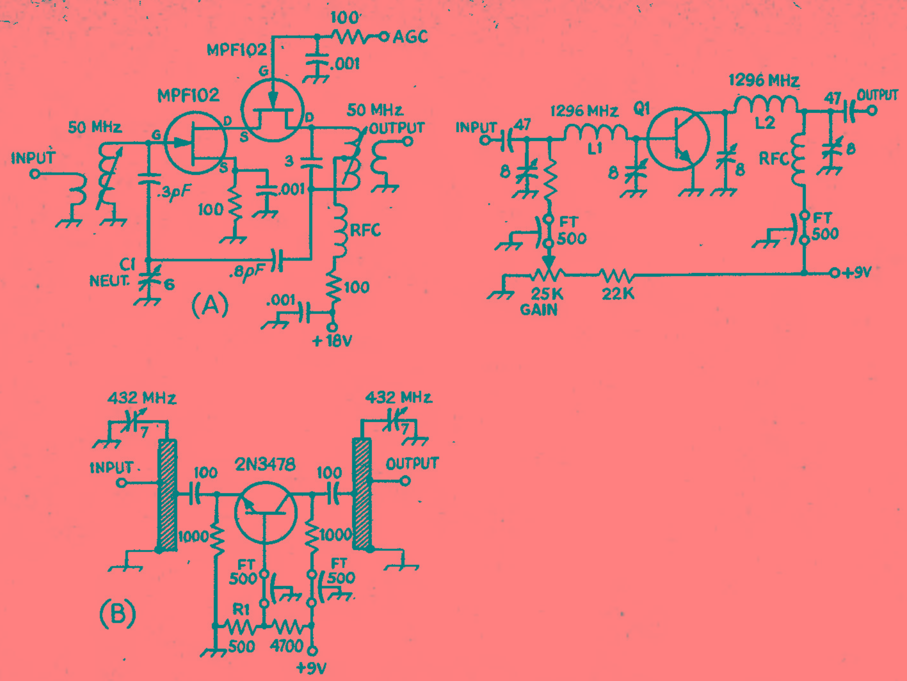


Fig. 9-3 — (A) Cascode amplifier circuit combines grounded-source and grounded-gate stages, for high gain and low noise figure. Though JFETs are shown, the cascode principle is usable with MOSFETs as well. (B and C) Examples of uhf preamplifier construction using bipolar transistors.

that "gate" may also imply "base" for bipolar transistors, or "grid" for tubes. "Source" should be read as "emitter" for the bipolar, and as "cathode" for the FET.

Amplifiers may be the grounded-source type, Fig. 9-1; grounded-gate, 9-2; or a combination of both, 9-3. The dual-gate MOSFET circuit, 9-1A, works well up to 300 MHz, but JFET and bipolar devices are superior for 420 MHz and higher. The gain and noise figure of a dual-gate MOSFET are adequate at 300 MHz, and it is simple and readily adapted to automatic gain control.

Triode tubes and FET transistors usually require neutralization for optimum noise figure with the grounded-cathode circuit. Inductive neutralization is shown in Fig. 9-1B, and the capacitive method shown at C works equally well. Examples will be seen later in this chapter. The 58-MHz trap circuit in Fig. 9-1A is discussed in the following section on mixers.

An alternative to neutralization lies in use of the grounded-gate circuit, Fig. 9-2. Its stage gain is lower and its bandwidth generally greater than with the grounded-cathode circuit. The input impedance is low, and the input circuit is tapped to provide a proper impedance match. A broad-band amplifier may be made with a low-impedance line connected directly to the input element, if selectivity is not required at this point for other reasons. Tubes designed for grounded-grid service include the 417A/5842, 416B, 7768 and the various "lighthouse" types, though almost any triode or triode-connected tetrode can be used. JFETs work well in grounded-gate circuits. In the grounded-grid amplifier, the tube heater becomes effectively a part of the tuned circuit, so some form of high-current rf choke is required. Ferrite-bead chokes work well.

The cascode circuit, Fig. 9-3, combined grounded-source and grounded-gate stages, securing some of the advantages of both. Fig. 9-3B shows a grounded-base bipolar transistor amplifer. The value of R1 should be chosen experimentally to achieve best sensitivity.

Front-End Protection

The first amplifier of a receiver is susceptible to damage or complete burnout through application of excessive voltage to its input element by way of the antenna. This can be the result of lightning discharges (not necessarily in the immediate vicinity), rf leakage from the station transmitter through a faulty send-receive relay or switch, or rf power from a nearby transmitter and antenna system. Bipolar transistors often used in low-noise uhf amplifiers are particularly sensitive to this trouble. The degradation may be gradual, going unnoticed until the receiving sensitivity has become very poor.

No equipment is likely to survive a direct hit from lightning, but casual damage can be prevented by connecting diodes back-to-back across the input circuit. Either germanium or silicon vhf diodes can be used. Both have thresholds of conduction well above any normal signal level, about 0.2 volt for germanium and 0.6 volt for silicon. The diodes used should have fast switching times. Computer diodes such as the 1N914 and hot-carrier types are suitable. A check on weak-signal reception should be made before and after connection of the diodes.

RF SELECTIVITY

The weakest point in any vhf or uhf receiver is the front-end circuit. Solid-state devices with high sensitivity, wide dynamic range and freedom from overload are now available. Thus, the quality of a front-end circuit is usually determined by how the active devices are used and the degree of rf selectivity included. High selectivity at vhf and uhf is not easy to achieve. Many lumped-constant tuned circuits are needed for even a moderate degree of selectivity at the signal frequency. Several tuned circuits before the first active stage (rf amplifier or mixer) will have sufficient loss to limit the sensitivity of the receiver. If lumped-constant circuits are employed, rf amplifiers can be interspaced between the *LC* elements to make up losses. High gain is not needed or desirable, so FETs operated grounded-gate are preferred.

For improved rf selectivity a helical resonator, a device which consists of a shield and a coil may be employed. One end of the coil is attached to the shield, as shown in Fig. 9-4, and the other end is open-circuited, except for a tuning capacitor. Helical resonators are electrically equivalent to a

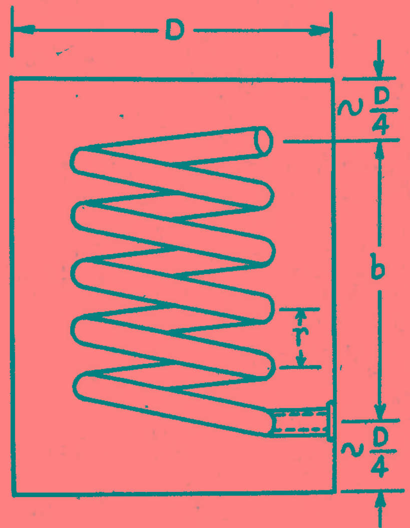


Fig. 9-4 — Outline sketch of resonator.

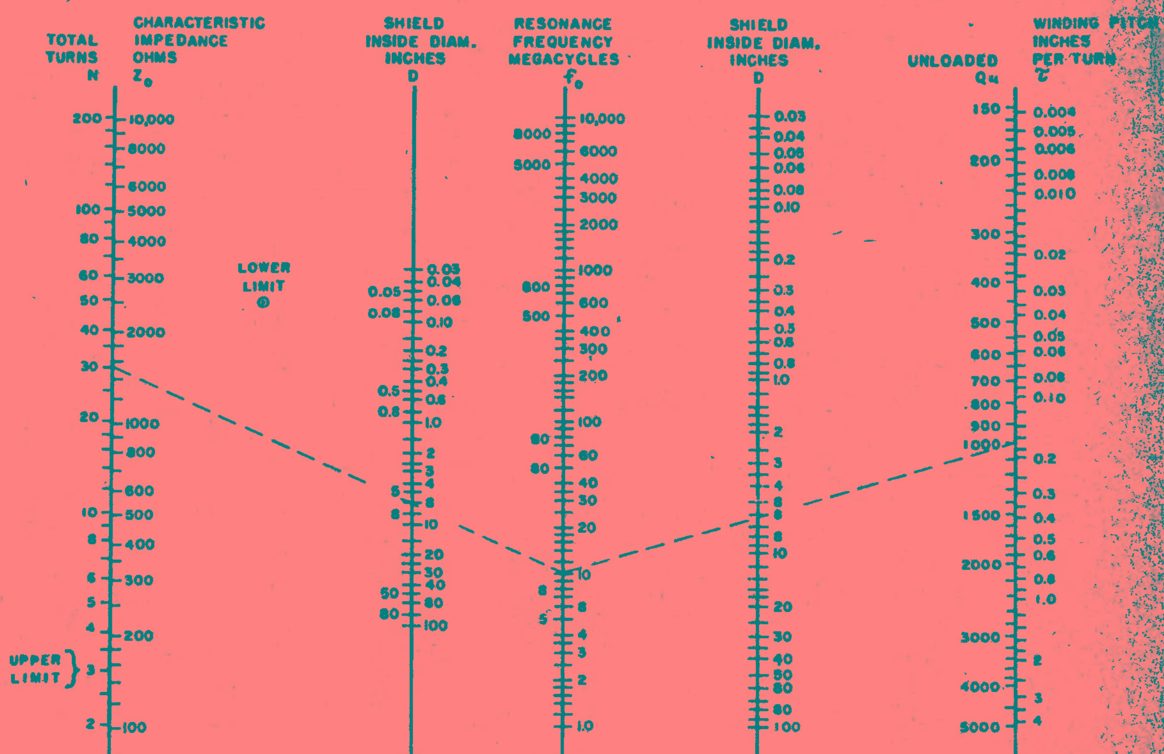


Fig. 9-5 — Design chart for quarter-wave helical resonators.

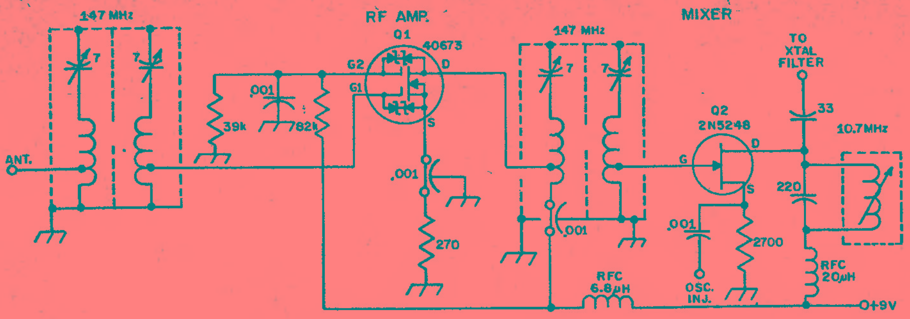


Fig. 9-6 — Schematic diagram of the Johnson 504 front-end circuit.

quarter-wave transmission-line resonator but are physically much smallter. Resonators can be built exhibiting *Q* of 1000 or more at vhf and uhf. Because the *Q* is so high, front-end circuits can be designed using helical resonators which provide a high degree of selectivity without high losses, at least a low and moderate power levels.

The inductance element in a helical resonator should be made as large as possible and capacitance kept to a minimum for best performance. Probe, tap or aperture coupling may be employed. The basic form of a helical resonator is shown in Fig. 9-4. A low-loss air-insulated trimmer or disk plunger may be used to tune the resonator. The capacitor must be much higher *Q* than the resonator to be useable. The usual precautions for fabricating high-Q coils must be observed when building a helical resonator. A protective silver plating is recommended for the coil and shield for units to be used above 100 MHz. The shield should be seamless and all joints should be effectively soldered to keep resistance to a minimum. The coil and shield should be made using heavy stock to assure mechanical stability.

Fig. 9-5 can be used to obtain approximate design information accurate to plus or minus ten percent. Complete design equations for helical resonators are beyond the scope of this text, but they may be found in Macapline and Schildknecht, "Coaxial Resonators with Helical Inner Conductor," *Proceeding of the IRE*, December, 1959.

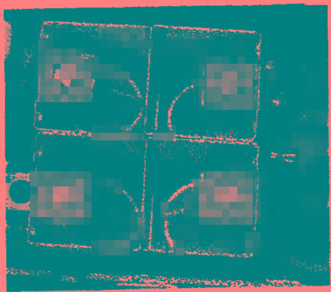


Fig. 9-7 — Close-up view of the helical resonators with the covers removed. The rf amplifier stage is constructed on the outside wall of the upper-right-hand resonator. Details are given in the text.

An application of helical resonators in a 146-MHz front-end circuit is shown in Figs. 9-6 and 9-7. This circuit is used in the Johnson 504 transceiver. The helical resonators consist of 5-3/4 turns of No. 12 wire contained in a rectangular 1 × 1 × 2-inch cavity. Both the coil and enclosure are silver plated. The coil is 5/8 inch inside diameter and 5/8 inch long, tuned with a 7-pF miniature air-variable capacitor. The 50-ohm input tap is at 1/4 turn from the ground end of the coil, an indication of the high impedance achieved. Coupling between individual resonators is through a 1/2 ×1/4 -inch aperture, or "window." Layout details can be seen in Fig. 9-7.

MIXERS

Conversion of the received energy to a lower frequency, so that it can be amplified more efficiently than would be possible at the signal frequency, is a basic principle of the superheterodyne receiver. The stage in which this is done may be called a "converter," or "frequency converter," but we will use the more common term, *mixer*, to avoid confusion with *converter*, as applied to a complete vhf receiving accessory. Mixers perform similar functions in both transmitting and receiving circuits, and mixer theory and practice are treated in considerable detail elsewhere in this *Handbook*.

A receiver for 50 MHz or higher usually has at least two such stages; one in the vhf or uhf converter, and usually two or more in the communications receiver that follows it. We are concerned here with the first mixer.

Diode Mixer: There are many types of mixers, the simplest being merely a diode with the signal and energy on the heterodyning frequency fed into it, somewhat in the manner of the 1296-MHz example, Fig. 9-8A. The mixer output includes both the sum and difference frequencies. Either can be used, but in this application it is the difference, since we are interested in going lower in frequency.

With a good uhf diode in a suitable circuit, a diode mixer can have a fairly low noise figure, and this is almost independent of frequency, well into the microwave region. The effectiveness of most active mixers falls off rapidly above 400 MHz, so the diode mixer is almost standard practice in amateur microwave communication. All diode mixers have some conversion *loss*. This must be added to the noise figure of the i-f amplifier following, to determine the overall system noise figure. Low-noise design in the first i-f stage is thus mandatory, for good weak-signal reception with a diode mixer having no rf amplifier preceding it. Purity of the heterodyning energy and the level of injection to the mixer are other factors in the performance of diode mixers.

Balanced mixers using hot-carrier diodes are capable of noise figures 1 to 2 dB lower than the best point-contact diodes. Hot-carrier diodes are normally quite uniform, so tedious selection of matched pairs (necessary with other types of diodes) is eliminated. They are also rugged, and superior in the matter of overloading.

The i-f impedance of a balanced hot-carrier diode mixer (Fig. 9-8B) is on the order of 90 ohms, when the oscillator injection is about one milliwatt. Thus the mixer and a transistorized i-f amplifier can be separated physically, and connected by means of 93-ohm coax, without an output transformer.

Conversion loss, around 7 dB, must be added to the noise figure of the i-f system to determine the overall system noise figure. Unless a low-noise preamplifier is used ahead of it, a communications receiver may have a noise figure of about 10 dB, resulting in an overall noise figure of 17 dB or worse for a vhf system with any diode mixer. A good i-f preamplifier could bring the receiver noise figure down to 2 dB or even less, but the system noise figure would still be about 9 dB; too high for good reception.

An amplifier at the signal frequency is thus seen to be required, regardless of mixer design, for optimum reception above 50 MHz. The rf gain, to override noise in the rest of the receiver, should be greater than the sum of noise figures of the mixer and the i-f system. Since the noise figure of the better rf amplifiers will be around 3 dB, the gain should be at least 20 dB for the first example in the previous paragraph, and 12 dB for the second.

Tube and Transistor Mixers: Any mixer is prone to overloading and spurious responses, so a prime design objective should be to minimize these problems. FET mixers have become standard practice at vhf. JFETs are slightly better than MOSFETs, although the junction types require

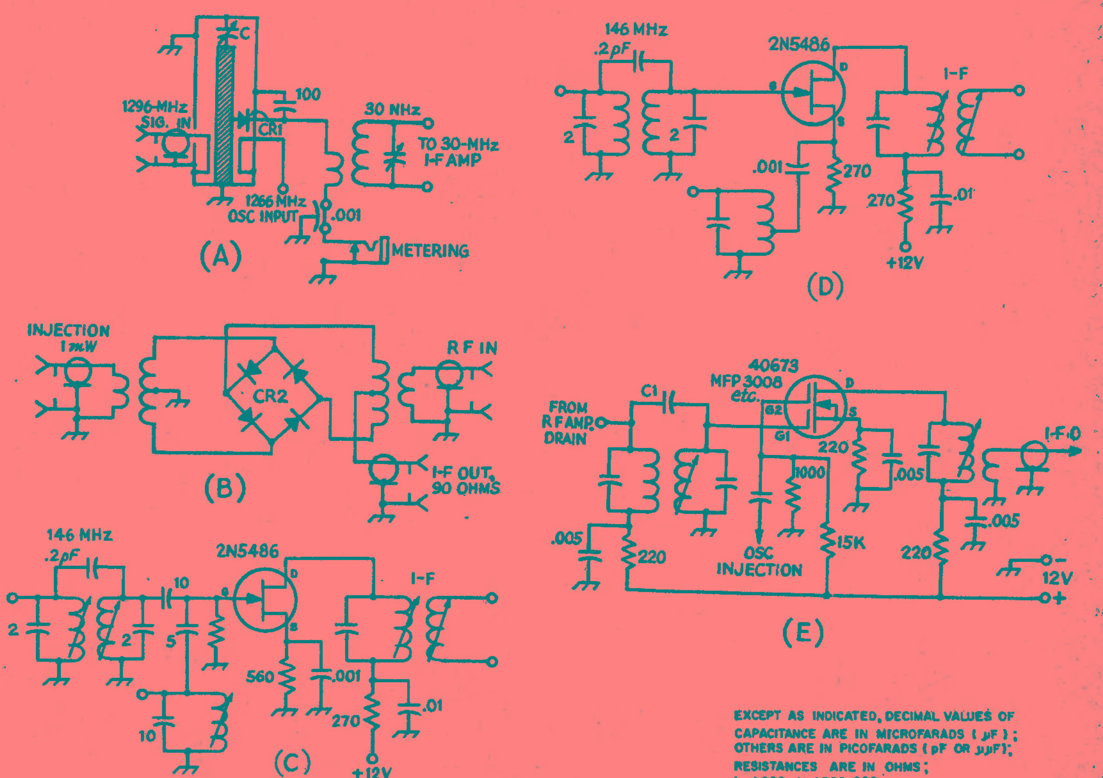


Fig. 9-8 — Vhf and uhf mixer circuits. A diode mixer for 1296 MHz, with a coaxial circuit for the signal frequency, is shown in A. CR1 is a uhf diode, such as the 1N21 series. A balanced mixer, as in B, gives improved rejection of the signal and injection frequencies. If hot-carrier diodes are used for CR2, sorting for matched characteristics is eliminated. Gate and source injection of a JFET mixer are shown at C and D, respectively.

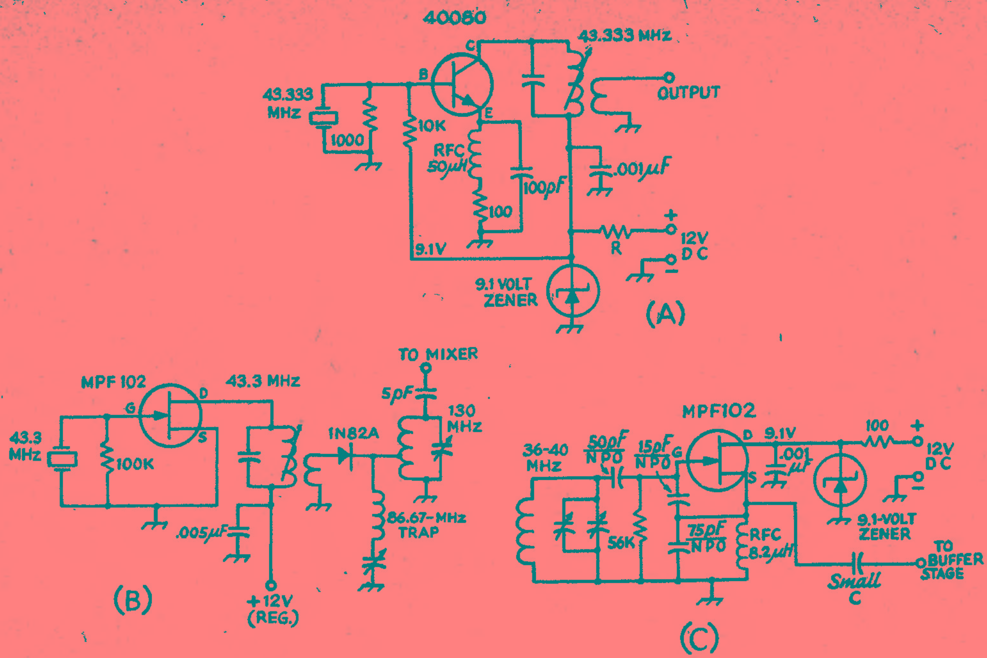


Fig. 9-9 — A simple overtone crystal oscillator for vhf converters, (A) has Zener voltage regulation. An FET overtone oscillator and diode multiplier, (B) supply injection for e 144-MHz converter with a 14-MHz i-f. Serias trap absorbs unwanted second harmonic at 86 MHz. A triode oscillator would use essentially the same circuit. A tunable oscilletor, as shown at C, would be suitable for a simple 50-MHz receiver with a broad i-f system.

more power from the injection source. When the local-oscillator frequency is far removed from the input frequency, the scheme of Fig. 9-8C can be used. The diagram at 9-8D is needed if the oscillator frequency is within 20 percent of the signal frequency.

The injection level from the oscillator affects mixer performance. Until it affects the mixer adversely in other ways, raising the injection level raises the mixer conversion gain. A simple check is made by observing the effect on signal-to-noise ratio as the injection is varied. At preferred injection levels, the gain will vary but the signal-to-noise ratio will not change. The injection should then be set for conversion gain a few decibels above that at which lower injection causes a drop in signal-to-noise ratio.

Double-tuned circuits in the mixer and the rf amplifier, as shown in several of the schematic diagrams in this chapter, help to keep down mixer response to signals outside the intended tuning range.

The insulated-gate FET is superior to other transistors for mixer service in the matter of overloading. An example is given in Fig. 9-8E. An objection to the MOSFET, the ease with which it can be damaged in handling, has been taken care of by building-in protective diodes in devices such as the MPF122, 40673, and 3N187. Units so designed require no special care in handling, and they work as well as their more fragile predeccessors. Insulated-gate MOSFETs have resistance to overloading which, while superior to most tubes, is not as good as the best JFETs.

Pentode or tetrode tubes make simple and effective mixers, up to 150 MHz or so. Triodes work well at any frequency, and are preferred in the high vhf range. Diode mixers are common in the 420-MHz band and higher.

INJECTION STAGES

Oscillator and multiplier stages that supply heterodyning energy to the mixer should be as stable and free of unwanted frequencies as possible. Stability is no great problem in crystal-controlled converters, if the oscillator is run at low input and its supply voltage is regulated. Simple Zener regulation, as in Fig. 9-9A, is adequate for a transistorized overtone oscillator. A higher order of regulation is needed for tunable oscillators. See Chapter 5 for suitable regulated power supplies.

Unwanted frequencies generated in the injection stages can beat with signals that are outside the intended tuning range. In a typical example, Fig. 9-9B, an FET overtone oscillator on 43.333 MHz feeds a diode tripler to 130 MHz. This frequency beats with signals between 144 and 148 MHz, to give desired responses at 14 to 18 MHz. The multiplier stage also has some output at twice the crystal frequency, 86.666 MHz. If allowed to reach the mixer, this can beat with fm broadcast signals in the 100-MHz region that leak through the rf circuits of the converter. There are many such annoying possibilities, as any vhf enthusiast living near high-powered fm and TV stations has found out.

Spurious frequencies can be kept down by using the highest practical oscillator frequency, no multiplier in a 50-MHz converter, and as few as possible for higher bands. Some unwanted harmonics are unavoidable, so circuit precautions are often needed to prevent both these harmonics and the unwanted signals from reaching the mixer. Selective coaxial or helical-resonator circuits are practical aids in uhf receivers. Trap circuits of various kinds may be needed to "suck out" energy on troublesome frequencies.

The series trap in Fig. 9-5B reduces the level of the 86-MHz second harmonic of the crystal frequency. A 58-MHz parallel-tuned trap, Fig. 9-1A, prevents the entry of Channel 2 TV signals that could otherwise beat with the second harmonic of a 36-MHz oscillator in a 50-MHz converter that works into a 14-MHz i-f (36 × 2 - 14=58).

Unwanted frequencies also increase the noise output of the mixer. This degrades performance in a receiver having no rf amplifier, and makes the job of an amplifier, if used, more difficult.

Frequency multipliers in vhf receivers generally follow transmitting practice, except for their low power level. The simple diode multiplier of Fig. 9-9B will often suffice. Its parallel-tuned 130-MHz circuit emphasizes the desired third harmonic, while the series circuit suppresses the unwanted second harmonic. The trap is tuned by listening to a spurious fm broadcast signal and tuning the series capacitor for minimum interference. The tripler circuit should be peaked for maximum response to a 2-meter signal. Do not detune this circuit to lower injection level. This should be controlled by the voltage on the oscillator, the coupling between the oscillator and multiplier, or by the coupling to the mixer from the 130-MHz circuit.

Tunable Oscillators

Any tunable vhf receiver must employ a variable oscillator. At this point the intermediate frequency is fixed, and the oscillator tunes a range higher or lower than the signal frequency by the amount of the i-f. In the interest of stability, it is usually lower. In Fig. 9-9C a simple JFET oscillator tunes 36 to 40 MHz, for reception of the 50-MHz band with a fixed 14-MHz i-f. Its stability should be adequate for a-m or fm reception with a relatively broad i-f, but it is unlikely to meet the requirements for ssb or cw reception, even for 50 MHz, and certainly not for higher bands.

Practically all vhf reception with high selectivity uses double-conversion schemes, with the tunable oscillator serving the second conversion. Such hf oscillators are treated in Chapter 6. They should run at the lowest practical input level, to minimize drift caused by heating. The supply should be well-regulated pure dc. Mechanically-rugged components and construction are mandatory. The circuits should be shielded from the rest of the receiver, and coupling to the mixer should be as light as practical. Drift cycling due to heating can be minimized if the oscillator is kept running continuously.

THE SUPERREGENERATIVE RECEIVER

Though the newcomer may not be too familiar with the superregenerative detector, the simple "rushbox" was widely used in early vhf work. Nothing of comparable simplicity has been found to equal its weak-signal reception, inherent noise-limiting and agc action, and freedom from overloading and spurious responses. But like all simple devices the superregenerator has limitations. It has little selectivity. It makes a high and unpleasant hissing noise, and it radiates a broad interfering signal around its receiving frequency.

Adding an rf amplifier will improve selectivity and reduce detector radiation. High-Q tuned circuits aid selectivity and improve stability. Use of superregeneration at 14 to 18, 26 to 30 MHz, or some similar hf range, in the tunable element of a simple superheterodyne receiver, works fairly well as a simple tuner for vhf converters. None of these steps corrects the basic weaknesses entirely, so the superregenerator is used today mainly where simplicity, low cost and battery economy are major considerations. Cw and narrow-band fm signals cannot be received using a superregenerative receiver.

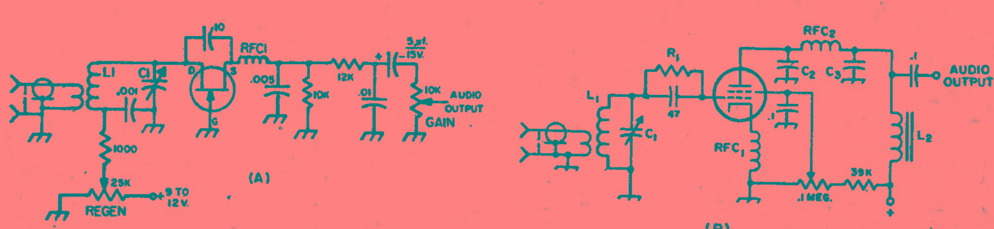


Fig. 9-10 — Circuits of typical superregenerative detectors using a field-effect transistor, A, and a tatrode tube, B. Regenaration is controlled by varying tha drain voltage on tha detector in the transistor circuit, and tha screen voltage in the tetrode or pentoda. Values of L1 and C1 should be adjusted for the frequancy involved, as should the size of the rf choke, RFC1.

C2, C3 — .001-µF disk ceramic. Try different values up to .005 for desired audio quality.
R1 — 2 to 10 megohms.
L2 — Small audio or filter choka; not critical.
RFC1 — Singla-layer rf choke, to suit frequency.
RFC2 — 85-mH rf choke.

Typical superregenerative detector circuits are shown in Fig. 9-10. High-transconductance FETs and high-beta vhf transistors are favored. The power source should be well-filtered and of low impedance. Fresh or well-charged batteries are ideal. Regeneration is controlled by varying the gain of the stage.

SERIES-RESONANT BYPASSING

Inexpensive disk-ceramic and "dog-bone" types of capacitors are relatively ineffective for bypassing above approximately 100 MHz. This is because of their considerable lead inductance, even when they are connected as close to the elements to be bypassed as possible. Actually this lead inductance can be used to advantage by selecting lead lengths that make the capacitor series-resonant at the frequency to be bypassed.

This approach is recommended by WA2KYF, who supplied the information in Table 9-I, showing capacitor and lead-length combinations for effective bypassing of rf energy at frequencies commonly encountered in vhf work. The values are not particularly critical, as a series-resonant circuit is broad by nature. The impedance of a series-resonant bypass is very close to zero ohms at the frequency of resonance, and it will be lower than most conventional capacitors for a considerable range of frequency either side of resonance.

A high-capacitance short-lead combination is preferable to a lower value with longer leads, because the former will be less likely to allow unwanted coupling to other circuits. For example, a 100-pF capacitor with 1/4-inch leads is a better bet than a 25-pF with 1-inch leads, for bypassing at 144 MHz. The series-resonant bypass is worth a try in any circuit where instability is troublesome, and conventional bypassing has been shown to be ineffective.

TABLE 9-I

Values of capacitance in pF required for resonance of frequencies commonly encountered in amateur-band vhf work, for leads of 1/4, 1/2 and 1 inch in length.

Frequency MHz	*1/4-Inch Leads*	*1/2-Inch Leads*	*1-Inch Leads*
48–50	800	400	200
72	390	180	91
96	220	100	56
144	100	47	25
220	39	20	10

MOSFET PREAMPLIFIERS FOR 10, 6, AND 2 METERS

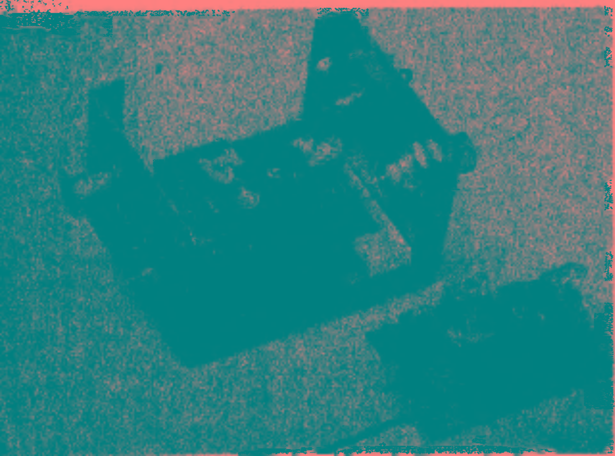
Where an hf or vhf receiver lacks gain, or has a poor noise figure, an external preamplifier can improve its ability to detect weak signals. This preamplifier uses an RCA 40673 dual-gate MOSFET. Designs for using this device as a mixer or as a preamplifier abound and many of them are excellent.

When it comes to simplicity, small size, good performance, low cost, and flexibility, a design by Gerald C. Jenkins, W4CAH, certainly qualifies.

Where the preamplifier really shines is in pepping up the performance of some of the older ten-meter receivers that many have pressed into service. A six-meter version is also very useful for any of the modes of communication available on that band.

The voltage dropping resistor, R4, and the Zener diode, VR1, may be of the value necessary to obtain 9 to 12 V dc for operation of the unit. By increasing the resistance and dissipation rating of R4 and VR1, the preamplifier may be operated from the 150- to 200-V supply found in many tube-type receivers.

The layout of the board is so simple that it is hardly worth the effort of making a negative for the photo-etch process. A Kepro resist-marking pen was used with success on several boards. Another approach – and one that is highly recommended –



Two versions of the preamplifier. The one in the box is for 2-meter use. Toroids are used in the six-meter version (right) and in the ten-meter preamplifier (not shown). Input is at the right on both units. The extra rf choke and feedthrough capacitor on the right end of the Minibox are for decoupling a crystal-current metering circuit that is part of a 2304-MHz mixer.