Receivers, Transmitters, Transceivers and Projects

17

This chapter, by William E. Sabin, W0IYH, discusses the "system design" of Amateur Radio receivers, transmitters and transceivers. "A Single-Stage Building Block" reviews briefly a few of the basic properties of the various individual building block circuits, described in detail in other chapters, and the methods that are used to combine and interconnect them in order to meet the requirements of the completed equipment. "The Amateur Radio Communication Channel" describes the relationships between the equipment system design and the electromagnetic medium that conveys radio signals from transmitter to receiver. This understanding helps to put the radio equipment mission and design requirements into perspective. Then we discuss receiver, transmitter, transceiver and transverter design techniques in general terms. At the end of the theory discussion is a list of references for further study on the various topics. The projects section contains several hardware descriptions that are suitable for amateur construction and use on the ham bands. They have been selected to illustrate system-design methods. The emphasis in this chapter is on analog design. Those functions that can be implemented using digital signal processing (DSP) can be explored in other chapters, but an initial basic appreciation of analog methods and general system design is very valuable.

A Single-Stage Building Block

We start at the very beginning with **Fig 17.1**, a generic single-stage module that would typically be part of a system of many stages.

A signal source having an "open-circuit" voltage V_{gen} causes a current I_{gen} to flow through Z_{gen}, the impedance of the generator, and Z_{in}, the input impedance of the stage. This input current is responsible for an open-circuit output voltage V_d (measured with a high-impedance voltmeter) that is proportional to I_{gen}. V_d produces a current I_{out} and a voltage drop across Z_{out}, the output impedance of the stage and Z_{load}, the load impedance of the stage. Observe that the various Zs may contain reactance and resistance in various combinations. Let's first look at the different types of gain and power relationships that can be used to describe this stage.

Actual Power Gain

Current I_{gen} produces a power dissipation P_{in} in the resistive component of Z_{in} that is equal to $I_{gen}^2\ R_{in}$. The current I_{out} produces a power dissipation P_{load} in the resistive component of Z_{load} that is equal to $I_{out}^2\ R_{load}$. The actual power gain in dB is 10 log (P_{load} / P_{in}). This is the conventional usage of dB, to describe a power ratio.

Voltage Gain

The current I_{gen} produces a voltage drop across Z_{in}. V_d produces a current I_{out} and a voltage drop V_{out} across Z_{load}. The voltage gain is the ratio

$$V_{out} / V_{in} \qquad (1)$$

In decibels (dB) it is

$$20 \log (V_{out} / V_{in}) \qquad (2)$$

This alternate usage of dB, to describe a voltage ratio, is common practice. It is *different* from the power gain mentioned in the previous section because it does *not* take into account the power ratio or the resistance values involved. It is a voltage ratio only. It is used in troubleshooting and other instances

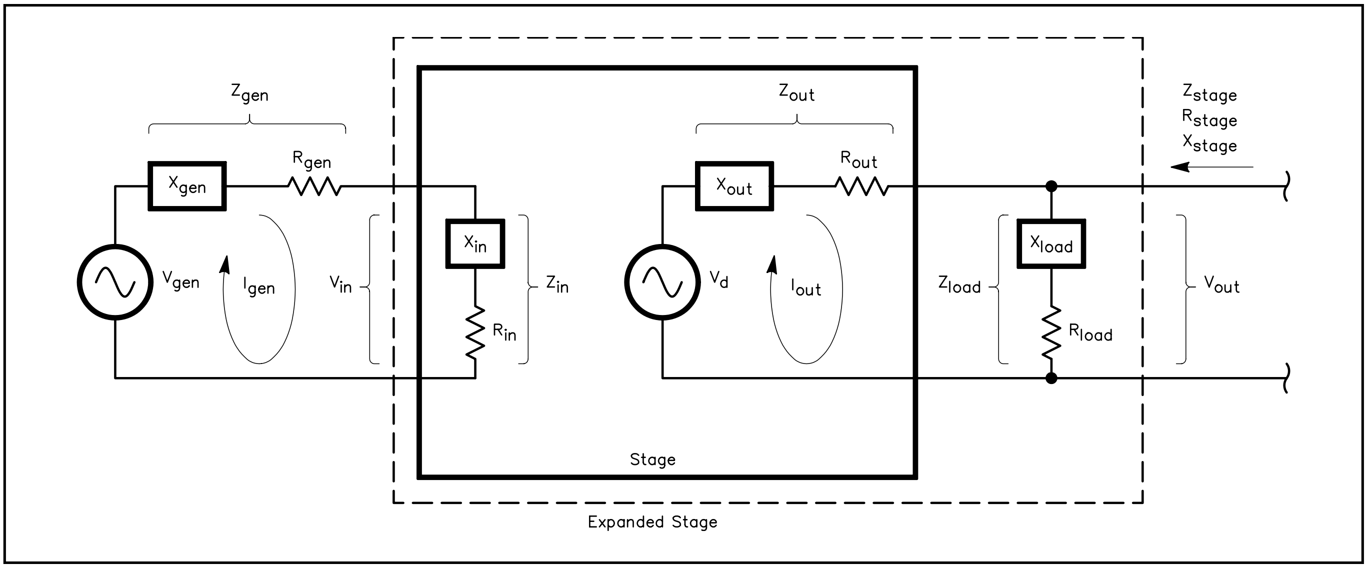


Fig 17.1—A single-stage building-block signal processor. The properties of this stage are discussed in the text.

where a rough indication of operation is needed, but precise measurement is unimportant. Voltage gain is often used in high-impedance circuits such as pentode vacuum tubes and is also sometimes convenient in solid-state circuits. Its improper usage often creates errors in radio circuit design because many calculations, for correct answers, require power ratios rather than voltage ratios. We will see several examples of this throughout this chapter.

Available Power

The maximum power, in watts, that can be obtained from the generator is $V_{gen}^2 / (4\ R_{gen})$. To see this, suppose temporarily that X_{gen} and X_{in} are both zero. Then let R_{in} increase from zero to some large value. The maximum power in R_{in} occurs when $R_{in} = R_{gen}$ and the power in R_{in} then has the value mentioned above (plot a graph of power in R_{in} vs R_{in} to verify this, letting $V_{gen} = 1$ V and $R_{gen} = 50$ W). This is called the "available power" (we're assuming sine wave signals). If X_{gen} is an inductive (or capacitive) reactance and if X_{in} is an equal value of capacitive (or inductive) reactance, the net series reactance is nullified and the above discussion holds true. If the net reactance is not zero the current I_{gen} is reduced and the power in R_{in} is less than maximum. The process of tuning out the reactance and then transforming the resistance of R_{in} is called "conjugate matching." A common method for doing this conjugate matching is to put an impedance transforming circuit of some kind, such as a transformer or a tuned circuit, between the generator and the stage input that "transforms" R_{in} to the value R_{gen} (as seen by the generator) and at the same time nullifies the reactance. **Fig 17.2** illustrates this idea and later discussion gives more details about these interstage networks. A small amount of power is lost within any lossy elements of the matching network.

This same technique can be used between the output of the stage in Fig 17.1 and the load impedance Z_{load}. In this case, the stage delivers the maximum amount of power to the load resistance. If both input and output are processed in this way, the stage utilizes the generator signal to the maximum extent possible. It is very important to note, however, that in many situations we do not want this maximum utilization. We deliberately "mismatch" in order to achieve certain goals that will be discussed later (Ref 1).

The dBm Unit of Power

In low-level radio circuitry, the watt (W) is inconveniently large. Instead, the milliwatt (mW) is commonly used as a reference level of power. The dB with respect to 1 mW is defined as

$$\text{dBm} = 10 \log (P_W / 0.001) \qquad (3)$$

where

dBm = Power level in dB with respect to 1 mW

P_W = Power level, watts.

For example, 1 W is equivalent to 30 dBm. Also

$$P_W = 0.001 \times 10^{\text{dBm}/10} \qquad (4)$$

Maximum Available Power Gain

The ratio of the power that is available from the stage, $V_d^2 / (4\ R_{out})$, to the power that is available from the generator, $V_{gen}^2 / (4\ R_{gen})$, is called the

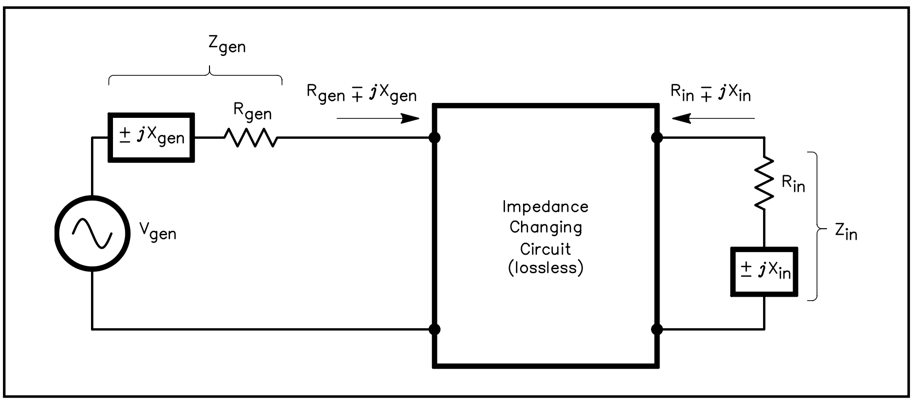


Fig 17.2—The conjugate impedance match of a generator to a stage input. The network input impedance is $R_{gen} \pm jX_{gen}$ and its output impedance is $R_{in} \pm jX_{in}$ (where either R term may represent a dynamic impedance). Therefore the generator and the stage input are both impedance matched for maximum power transfer.

maximum available power gain. In some cases the circuit is adjusted to achieve this value, using the conjugate-match method described above. In many cases, as mentioned before, less than maximum gain is acceptable, perhaps more desirable.

Available Power Gain

Consider that in Fig 17.1 the stage and its output load Z_{load} constitute an "expanded" stage as defined by the dotted box. The power available from this new stage is determined by V_{out} and by R_{stage}, the resistive part of Z_{stage}. The available power gain is then $V_{out}^2 / (4\ R_{stage})$ divided by $V_{gen}^2 / (4\ R_{gen})$. This value of gain is used in a number of design procedures. Note that Z_{load} can be a physical network of some kind, or it may be partly or entirely the input impedance of the stage following the one shown in Fig 17.1. In the latter case it is sometimes convenient to "detach" this input impedance from the next stage and make it part of the expanded first stage, as shown in Fig 17.1, but we note that Z_{out} is still the generator (source) impedance that the input of the next stage "sees."

Transducer Power Gain

The transducer gain is defined as the ratio of the power actually delivered to R_{load} in Fig 17.1 to the power that is available from the generator V_{gen} and R_{gen}. In other words, how much more power does the stage deliver to the load than the generator could deliver if the generator were impedance matched to the load? We will discuss how to use this kind of gain later.

Transformation Methods

Let's digress briefly to discuss two valuable techniques that are often used in designing radio circuits. **Fig 17.3A** shows a voltage generator in series with an impedance. Fig 17.3B shows how to represent Fig 17.3A as a current generator in parallel with the same value of impedance. We can also see how to move from 3B to 3A. Fig 17.3C shows a complicated network consisting of generators, resistors, capacitors and inductors. At any single frequency, this network can be replaced with a single generator and a single impedance as shown in Fig 17.3. The methods for doing this may be learned from a study of ac circuit analysis, and a personal computer math-equation-solving program that handles complex numbers and matrices is a valuable tool for quickly performing these operations over a range of frequencies and plotting the results. There are also simulation programs such as SPICE (which is available under many different brand names and versions) that are very effective and easy to learn and use. Student radio circuit designers should learn these techniques.

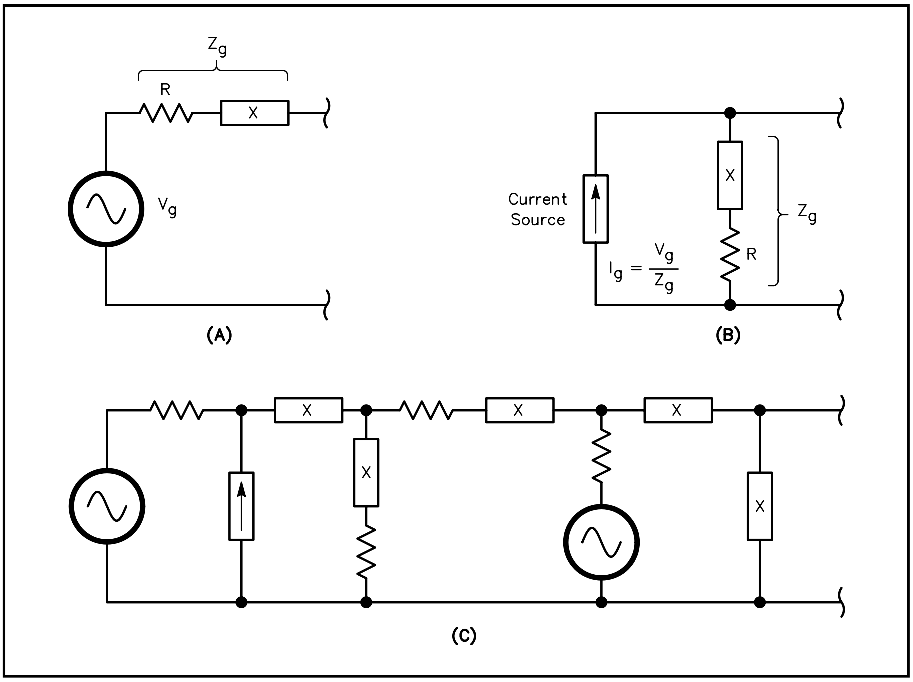


Fig 17.3—Transformations of networks that are useful in radio circuit design.

Feedback (Undesired)

One of the most important properties of the single-stage building block in Fig 17.1 is that changes in the load impedance Z_{load} cause changes in the input impedance Z_{in}. Changes in Z_{gen} also affect Z_{out}. These effects are due to reverse coupling, within the stage, from output to input. For many kinds of circuits (such as networks, filters, attenuators, transformers and so on) these effects cause no unexpected problems.

But, as the chapter on **RF Power Amplifiers** explains in detail, in active circuits such as amplifiers this reverse coupling within one stage can have a major impact not only on that stage but also on other stages that follow and precede. It is the effect on system performance that we discuss here. In particular, if a stage is expected to have certain gain, noise factor and distortion specifications, all of these can be changed either by reverse coupling (undesired feedback) within the stage or adjacent stages. For example, internal feedback can cause the input impedance of a certain stage "A" (Fig 17.1) to become very large. If this impedance is the load impedance for the preceding stage, the gain of the preceding stage can become excessive, creating problems in both stages. This same feedback can cause the gain of stage "A" to become greater, thereby causing the next stage to be driven into heavy distortion. A very common event is that stage "A" goes into oscillation. All of these occurrences are common in poorly designed radio equipment. Changes in temperature and variations in component tolerances are major contributors to these problems.

One particular example is shown in **Fig 17.4**, a transistor amplifier, shown in skeleton form, with sharply tuned resonators at input and output.

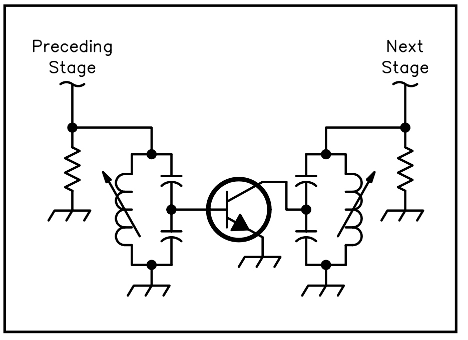


Fig 17.4—A double tuned transistor amplifier circuit that may oscillate due to excessive amplification and reverse coupling.

Because of reverse coupling, the two tuned circuits interact, making adjustments difficult or even impossible. The likelihood of oscillation is very high. There are two solutions: drastically reduce the gain of the amplifier, or use an amplifier circuit that has very little reverse coupling. Usually, both methods are used simultaneously (in the right amount) in order to get predictable performance. The object lesson for the system designer is that a combination of reduced gain and low reverse coupling is the safe way to go when designing a radio system. More stages may be required, but the price is well worthwhile. The cascode amplifier, grounded-gate amplifier, dual-gate FET and many types of IC amplifiers are examples of circuits that have little reverse coupling and good stability. "Neutralization" methods are used to cancel reverse coupling that causes instability. All such circuits are said to be "unilateral," which means "in one direction" and both input and output can be independently tuned as in Fig 17.4 if the gain is not too high.

Feedback (Desired)

The **RF Power Amplifiers** chapter explains how negative feedback (good feedback) can be used to stabilize a circuit and make it much more predictable over a range of temperature and component tolerances. Here we wish to point out some system implications of negative feedback. One is that the gain, noise-figure and distortion performances within a stage are made much more constant and predictable. Therefore a system designer can put building blocks together with more confidence and less guesswork.

There are some problems, though. In some circuits the amount of feedback depends on both the output impedance of the driving circuit and the input impedance of the next stage. A classic example is the cascadable amplifier shown in **Fig 17.5**.

In this circuit, if the output load impedance becomes very low the amplifier input impedance becomes high, and vice versa (a "teeter-totter" effect). Other amplifier properties also can change. With amplifiers of this type it is important to maintain the correct impedances at the input and output interfaces. Any

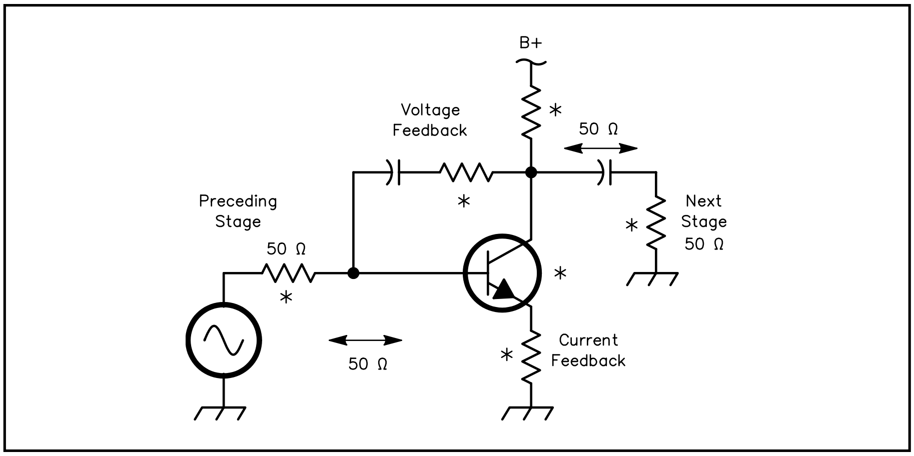


Fig 17.5—A cascadable amplifier using feedback. The feedback and therefore the amplifier performance depends on the load and driving-stage impedances.

building block should be examined for effects of this kind. Data sheets frequently specify the reverse transfer values as well as those for forward transfer. Often, lab measurements are needed. Apply a signal to the output and measure the reverse coupling to the input. Where varying load and source impedances are involved, look for a circuit that is less vulnerable (that is, has less reverse coupling).

Another problem is that feedback networks often add thermal noise sources to a circuit and so degrade its noise figure. In systems where this is a consideration, use so-called "loss-less feedback" circuits. These circuits use very efficient transformers instead of resistors or lossy networks that introduce thermal noise into a system.

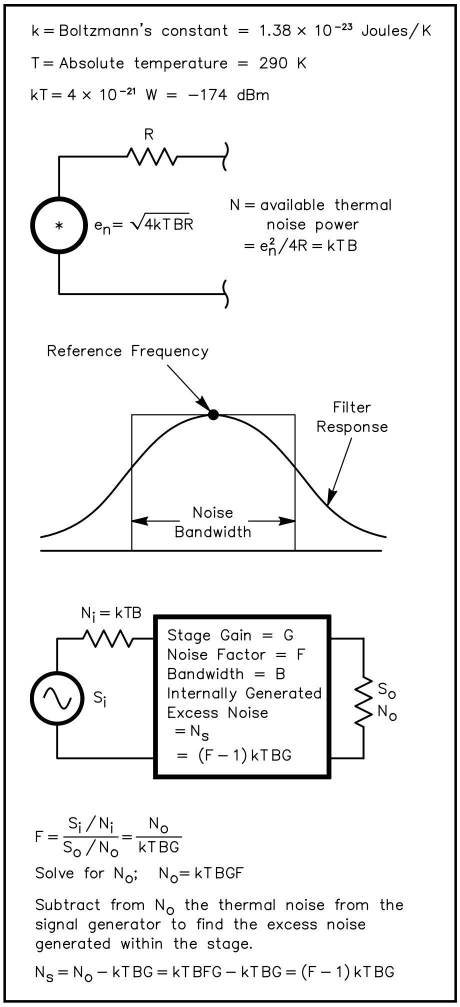


Fig 17.6—Diagram and equations that explain F, the noise factor of a single stage. The excess noise generated within the stage is also indicated. The definition of noise bandwidth is included.

Noise Factor and Noise Figure

The output resistance of the signal generator that drives a typical signal processing block such as shown in Fig 17.5 is a source of thermal noise power, which is a natural phenomenon occurring in the resistive component of any impedance. It is caused by random motion of electrons within a conducting (or semiconducting) material. Note that the reactive part of an impedance is not a source of thermal noise power because the voltage across a pure reactance and the current through the reactance are in phase quadrature (90°) at any one frequency. The average value of the product of these two (the power) is zero. If this is true at any frequency, then it is true at all frequencies. Also, a purely "dynamic resistance" such as R_e, the dynamic resistance ($\Delta V/\Delta I$) of a perfect forward conducting PN junction, is also not a source of thermal noise. However, the junction is a source of "shot noise" power that, by the way, is only 50% as great as the thermal noise that R_e would have if it were an actual resistor (Ref 2).

Each "*" in Fig 17.5 indicates a noise source. Passive elements generate thermal noise. Active components such as transistors generate thermal noise and other types, such as shot noise and flicker (1/f) noise, internally. These "excess" noises are all superimposed on the signal from the generator. Therefore the noise factor of a single stage is a measure of how much the signal to noise ratio is degraded as a signal passes through that stage.

Refer now to the diagram and equations in **Fig 17.6**. F is noise factor and S_i/N_i is the input signal to noise ratio from the signal generator. S_o/N_o is signal to noise ratio at the output and kTB is the thermal noise power that is available from any value of resistance (kT = –174 dBm in a 1-Hz bandwidth at room temperature). G is S_o/S_i, the available power gain of the stage and B is the noise bandwidth at the *output* of the stage, assumed to be not wider than the noise bandwidth at the input. The case where the output noise bandwidth is wider will be considered in a later section.

Noise bandwidth is defined in Fig 17.6. An ideal rectangular frequency response has a maximum value that is defined at the reference frequency. The area under the rectangle is the same as the area under the actual filter response, therefore the noise within the rectangle and within the actual filter response are equal. The width of the rectangle is called the noise bandwidth. Various kinds of filters have certain ratios of signal bandwidth to noise bandwidth that can be measured or calculated.

Part of the output noise is amplified thermal noise from the signal generator. To find the noise that is generated within the stage, we must subtract the amplified signal generator noise from the total output noise. Fig 17.6 shows the equation that performs this operation and the quantity (F-1)kTBG is the excess noise that the stage contributes.

In general, the excess noise of a stage is the output noise minus the amplified noise from the previous stage. A thorough understanding of this concept is very important for any one who designs radio systems that employ low-level signals. Finally, noise figure, NF, is 10 times the logarithm of F, the noise factor (Ref 3).

Noise Factor of a Passive Device

Often the stage in Fig 17.6 is a filter, an attenuator or some other passive device (no amplification) that contains only thermal noise sources. In a device of this kind, the output noise is thermal noise of the same value as the thermal noise of the generator alone. That is, using the method of Fig 17.3, all the thermal noise sources inside the device and also the generator resistance can be combined into a single resistor whose available noise power is kTB, the same as that of the generator alone. Therefore no additional noise is added by the device. But the available signal power is reduced by the attenuation (loss of signal) of the device. Therefore, using the equation for noise factor in Fig 17.6, the noise factor F of the device is numerically equal to its attenuation. For example, a 3-dB attenuator has a 3-dB noise figure, or a noise factor of 2. This important fact is very useful in radio design. It applies only when there is no amplification and no shot noise or 1/f noise sources within the device. This discussion assumes that all components and the generator are at the same temperature. If not, a slightly more complicated procedure involving Equivalent Noise Temperatures (T_E), to be discussed later in this chapter, can be used.

Sensitivity

Closely related to the concept of noise figure (or noise factor) is the idea of sensitivity. Suppose a circuit, a component or a complete system has a noise figure NF (dB) and therefore a noise output N_o (dBm). Then the value in dBm of a signal generator input that increases the total output (signal + noise) by 10 dB is defined as the "sensitivity." That is, 10 × log[(signal + noise) / (noise)] =10 dB. The ratio (signal) / (noise) is then equal to 9.54 dB (=10 × log(9)). N_o is equal to kTBFG as shown in Fig 17.6. B is noise bandwidth. Using this information, the sensitivity is

$$S\ (\text{dBm}) = -174\ (\text{dBm}) + 9.54\ (\text{dB}) + NF\ (\text{dB}) + 10\log(B)\ (\text{dB}) \quad (5)$$

In terms of the "open circuit voltage" from a 50-W signal generator (twice the reading of the generator's voltmeter) the sensitivity is

$$E\ (\text{volts open circuit}) = 0.4467 \times 10^{S/20} \quad (6)$$

N_o is 9.54 dB below the sensitivity value. This is sometimes referred to in specifications as the "noise floor." The signal level that is equal to the noise floor is sometimes referred to as the minimum detectable signal (MDS). Also associated with N_o is the concept of "noise temperature" which we discuss later under Microwave Receivers (Ref 3).

Distortion in a Single Stage

Suppose the input to a stage is called X. If the stage is perfectly linear the output is Y, and Y = AX, where A is a constant of proportionality. That is, Y is a perfect replica of X, possibly changed in size. But if the stage contains something nonlinear such as a diode, transistor, magnetic material or other such device, then $Y = AX + BX^2 + CX^3 + \ldots\ldots$. The additional terms are "distortion" terms that deliver to the output artifacts that were not present in the generator. Without getting too mathematical at this point, if the input is a pure sine wave at frequency f, the output will contain "harmonic distortion" at frequencies 2f, 3f and so on. If the input contains two signals at f1 and f2, the output contains *intermodulation distortion* (IMD) products at f1 + f2, f1 – f2, 2f1 + f2, 2f2 – f1, just to name a few. All semiconductors, vacuum tubes and magnetic materials create distortion and the radio designer's job is to limit the distortion products to acceptable levels. We wish to look at distortion from a system-design standpoint.

There are several ways to reduce distortion. One is to use a high-power device operated well below its maximum ratings. This leads to devices that dissipate more power in the form of heat. Unfortunately, these devices also tend to be noisier; so high power levels and low noise tend to be incompatible goals in most cases. (Some modern devices, such as certain GaAsFETs, achieve improved values of dynamic range.) Also, a large reduction in distortion is not always assured with this method, especially in transmitters.

Second, reduce the signal level into the device. This allows a lower power device to be used that will tend to be less noisy. To get the same output level, though, we must increase the gain of the stage. Then we run into another problem: if the signal at the output of this lower power stage becomes too large, distortion is generated at the output. Also, as mentioned before, high-gain stages tend to be unstable at RF.

Third, reduce the stage gain. But then we must add another stage in order to get the required output level. This additional stage turns out to be a high-power stage. The addition of another stage adds more noise and distortion contamination to the signal.

Fourth, use negative feedback. This is a powerful technique that is discussed in detail in the **RF Power Amplifiers** chapter. In general, if we increase the stage gain and perhaps make it more powerful, we can use feedback to reduce distortion and stabilize performance with respect to component variations. The feedback stage may be noisier, although the use of loss-less feedback can improve this situation. Negative feedback is the preferred method for reducing distortion in radio design, but the gain reduction due to feedback means that more stages are needed. This tends to reintroduce some noise and distortion.

A fifth way reduces distortion by increasing selectivity. For example, harmonics of an RF amplifier can be eliminated by a tuned circuit. Products such as f1 + f2 and f1 – f2 can often also be eliminated. Third-order products such as 2f1 – f2 and so on (and higher odd-order products), frequently are sufficiently close to f1 and f2 that selectivity does not help much, but if they are somewhat removed in frequency these so-called "adjacent channel" products can be greatly reduced.

A sixth way is to use push-pull circuits (see the **RF Power Amplifiers** chapter) that tend to greatly reduce "even-order" products such as 2f, 4f, f1 + f2, f1 – f2, and so on. But "odd-orders" such as 3f, 5f, 2f1 + f2, 2f1 – f2 are not reduced by this method except as noted later in the Modules in Combination section.

A seventh way uses diplexers to absorb undesired harmonics or other spurious products.

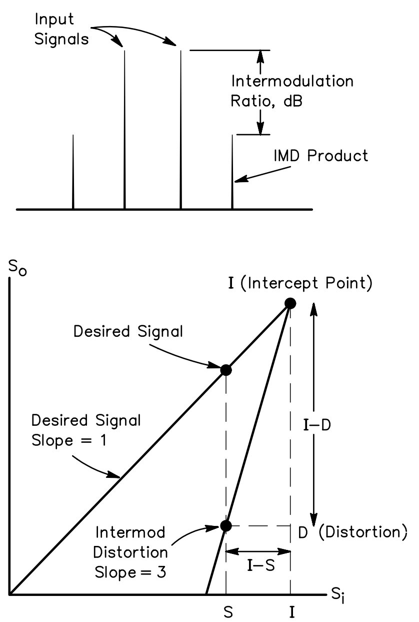
So there are compromises to be made. The designer must look for the compromise that gets the job done in an acceptable manner and is optimal in some sense. For example, devices are available that are optimized for linearity.

IMD Ratio

If a pair of equal-amplitude signals creates IMD products, the IMD ratios (IMR) are the differences, in dB, between each of the two tones and each of the IMD products (see **Fig 17.7**).

Intercept Point

The intercept point is a figure of merit that is commonly used to describe the IMD performance of an individual stage or a complete system. For example, third-order products increase at the rate of 3:1. That is, a 1-dB increase in the level of each of the two-tone input signals produces (ideally, but not always exactly true) a 3 dB increase in third-order IMD products. As the input levels increase, the distortion products seen at the output on a spectrum analyzer could catch up to, and equal, the level of the two desired signals, if the circuit did not go into a limiting process (see next topic). The input level at which this occurs is the input intercept point. Fig 17.7 shows the concept graphically, and also derives from the geometry an equation that relates signal level, distortion and intercept point. A similar process is used to get a second-order intercept point for second-order IMD. These formulas are very useful in designing radio systems and circuits. If the input intercept point (dBm) and the gain of the stage (dB) are added the result is an output intercept point (dBm). Receivers are specified by input intercept point, referring distortion back to the receive antenna input. Transmitter specifications use output intercept, referring distortion to the transmit antenna output.



From the geometry of the figure

$$\frac{I-D}{I-S} = 3$$

Solve for I to get
$I = S + 0.5\,(S/D)$
where S/D is signal to distortion ratio in dB, S is in dBm and I is in dBm.

For second–order intermodulation,
$I = S + (S/D)$

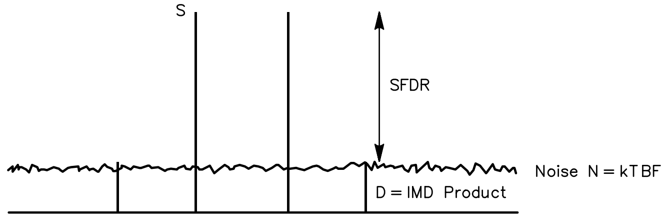
Fig 17.7—A: IMD ratio (as displayed on a spectrum analyzer), and B: intercept point.

Gain Compression

The gain of a circuit that is linear and has little distortion products deteriorates rapidly when the instantaneous input or output level reaches a critical point where the peak or trough of the waveform begins to "clip" or "saturate." The 1-dB compression point occurs when the output is 1 dB less than it would be if the stage were still linear. Some circuits do not need to be linear (and should not be linear), and we will look at several examples. In many applications linearity is necessary, especially in SSB receivers and transmitters. The situation for a linear circuit is optimum when the input and output become nonlinear simultaneously. This means that the gain, bias and load impedance are all properly coordinated. We will study this more closely in later sections.

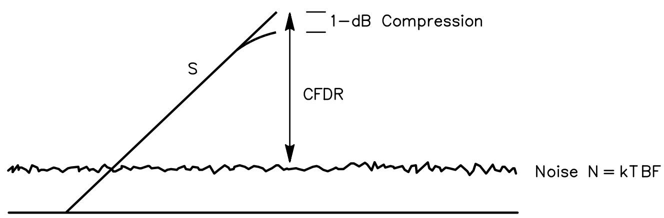
Dynamic Range

There is a relationship between noise factor, IMD, gain compression and bandwidth in a building block stage. In general, an active circuit that has a low noise factor tends to have a poor intercept point and vice versa. A well-designed transistor or circuit tries to achieve the best of both worlds. Dynamic range is a measure of this capability. Suppose that a circuit has a third-order input intercept of +10 dBm, a noise factor of 6 dB and a noise bandwidth of 1000 Hz. We want to determine its dynamic range. At a certain level per tone of a two-tone input signal the third-order IMD products are equal to the noise level in the 1000-Hz band. The ratio, in dB, of each of the two tones to the noise level is called the "spurious free dynamic range" (SFDR). **Fig 17.8** illustrates the problem and derives the proper formula. Note that the bandwidth is an important player. For the example above, the dynamic range is DR = 0.67(10 – (–174 + 10 log(1000) + 6)) = 99 dB. Often the dynamic range is calculated using a 1.0-Hz bandwidth. This is called "normalization." Another kind of dynamic range compares the 1-dB compression level with the noise level. This is the CFDR (compression-free dynamic range). Fig 17.8 illustrates this also.



The third-order intercept is
$I(dBm) = 1.5S - 0.5D$
Let intermod product (dBm) equal the noise level (dBm)
$D = N = 10\ Log(kTBF)$
Substitute N for D and solve for S
$1.5S = I + 0.5N$
Find S – N, which is the dynamic range (DR)
$1.5S - 1.5N = I - N$
$S - N = SFDR3\,(dB) = 0.67\,(I3 - N) = 0.67\,(I3\,(dBm) - 10Log\,(kTBF))$

The second-order SFDR is
$SFDR2\,(dB) = 0.50\,(I2\ (dBm) - 10Log\,(kTBF))$
For equal levels of spurious from second-order and third-order IMD, for a given value of S, the *minimum* value of I2 is
$I2 = (4 \times I3 - kTBF)/3$
A sub-octave band-pass filter will improve I2 by 1 dB for each dB of attenuation of the *out-of-band* interfering signal.



As signal level increases to the 1-dB gain-compression point, measure the difference in dB between S and N

Fig 17.8—The definitions of spurious-free dynamic range (SFDR) and compression-free dynamic range (CFDR). The derivation yields a very useful equation for SFDR.

Modules in Combination

Quite often the performance of a single stage can be greatly improved by combining two identical modules. Because the input power is split evenly between the two modules the drive source power can be twice as great and the output power will also be twice as great. In transmitters, especially, this often works better than a single transistor with twice the power rating. Or, for the same drive and output power, each module need supply only one-half as much power, which usually means better distortion performance. Often, the total number of stages can be reduced in this manner, with resulting cost savings. If the combining is performed properly, using hybrid transformers, the modules interact with each other much less, which can avoid certain problems. These are the system-design implications of module combining.

Three methods are commonly used to combine modules: parallel (0°), push-pull (180°) and quadrature (90°). In RF circuit design, the combining is often done with special types of "hybrid" transformers called splitters and combiners. These are both the same type of transformer that can perform either function. The splitter is at the input, the combiner at the output. We will only touch very briefly on these topics in this chapter and suggest that the reader consult the **RF Power Amplifiers** chapter and the very considerable literature for a deeper understanding and for techniques used at different frequency ranges. **Fig 17.9** illustrates one example of each of the three basic types.

In a 0° hybrid splitter at the input the tight coupling between the two windings forces the voltages at A and B to be equal in amplitude and also equal in phase if the two modules are identical. The 2R resistor between points A and B greatly reduces the transfer of power between A and B via the transformer, but only if the generator resistance is closely equal to R. The output combiner separates the two outputs C and D from each other in the same manner, if the output load is equal to R, as shown. No power is lost in the 2R resistor if the module output levels are identical.

The 180° hybrid produces push-pull operation. The advantages of push pull were previously discussed. The horizontal transformers, 1:1 balun transformers, allow one side of the input and output to be grounded. The R/2 resistors improve isolation between the two modules if the 2R resistors are accurate, and dissipate power if the two modules are not identical.

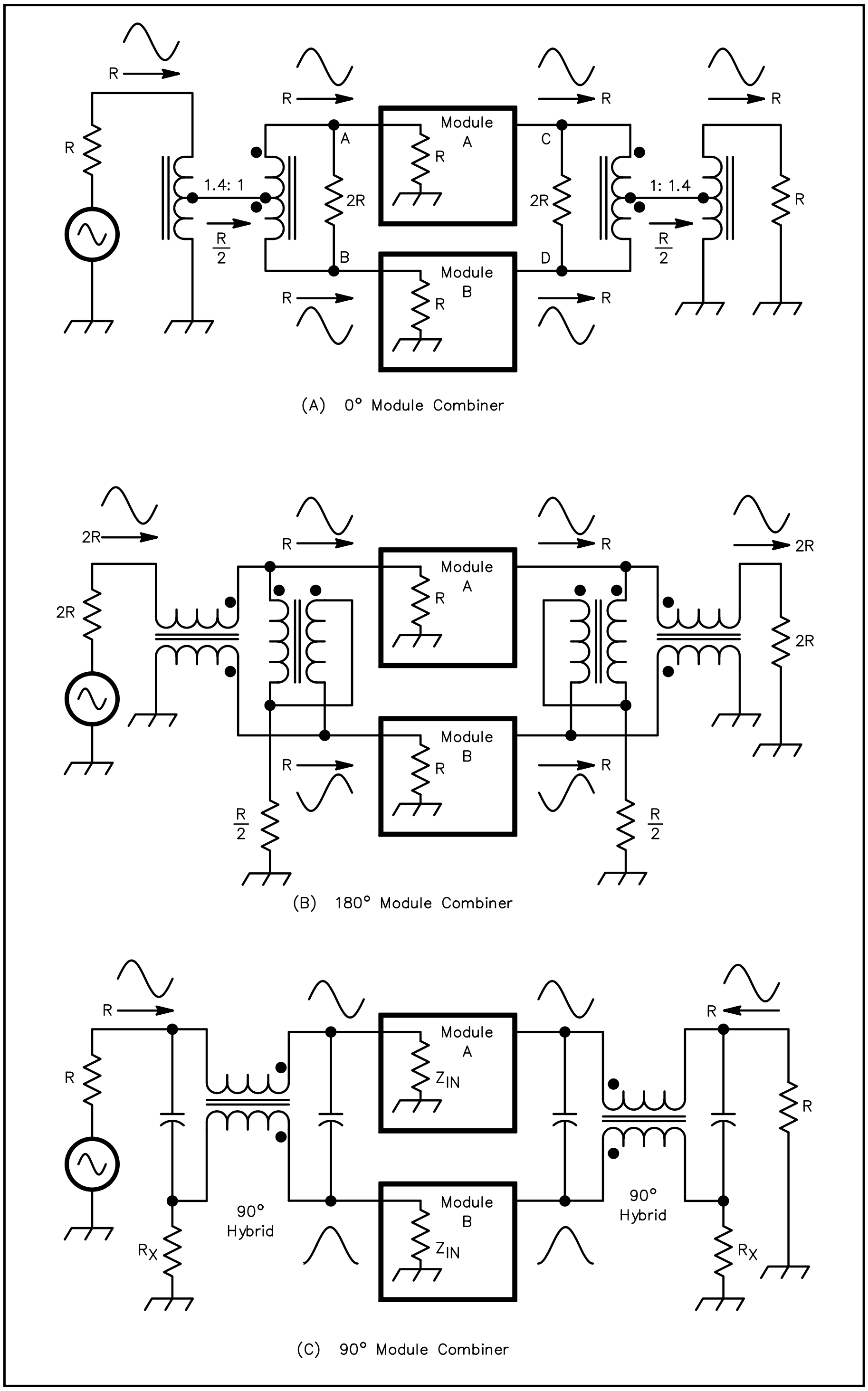


Fig 17.9—The three basic techniques for combining modules.

In a 90° hybrid splitter, if the two modules are identical but their identical input impedance values may not be equal to R, the hybrid input impedance is nevertheless R Ω, a fact that is sometimes very useful in system design. The power that is "reflected" from the mismatched module input impedance is absorbed in RX, the "dump" resistor, thus creating a virtual input impedance equal to R. The two module inputs are 90° apart. At the output, the two identical signals, 90° apart, are combined as shown and the output resistance is also R. This basic hybrid is a narrowband device, but methods for greatly extending the frequency range are in the literature (Refs 4, 5, 6, 7, 8). One advantage of the 90° hybrid is that catastrophic failure in one module causes a loss of only one half of the power output.

Multistage Systems

As the next step in studying system design we will build on what we've learned about single stages, and look at the methods for organizing several building block circuits and their interconnecting networks so that they combine and interact in a desirable and predictable manner. These methods are applicable to a wide variety of situations. Further study of this chapter will reveal how these methods can be adapted to various situations. We will consider typical receiving circuits and typical transmitting circuits.

PROPERTIES OF CASCADED STAGES

Fig 17.10 shows a simple receiver "front end" circuit consisting of a preselector filter, an RF amplifier, a second filter and a double balanced diode mixer. We want to know the gain, bandwidth, noise factor, second and third-order intercept points, SFDR and CFDR for this combination, when the circuitry following these stages has the values shown. Let's consider one item at a time.

Gain of Cascaded Stages

The antenna tuned circuit L1C1C2 has some resistive loss; therefore the power that is available from it is less than the power that is available from the generator. Let's say this loss is 2.0 dB.

Next, find the available power gain of the RF amplifier. First, note that the generator voltage V_s is transformed up to a larger voltage V_g by the input tuned circuit, according to the behavior of this kind of circuit. This step-up increases the gain of the RF amplifier because the FET now has a larger gate voltage to work with. (A bit of explanation: The FET has a high input impedance therefore, since the generator resistance R_s is only 50 Ω, a voltage step-up will utilize the FET's capabilities much better. But an excessive step-up opens up the possibility that the FET and other "downstream" circuits can be overdriven by a moderately large signal. So this step-up process should not be carried too far). The gain also depends on the drain load resistance, which is the mixer input impedance, stepped up by the circuit L2C3C4. Again, there is some loss within this tuned circuit, say 2.0 dB. If the drain load is too large the FET drain voltage swing can become excessive, creating distortion. The RF amplifier can become unstable due to excessive gain. Note also that the unbypassed source resistor provides negative feedback, to help make the RF amplifier more predictable. Dual-gate FETs have relatively little reverse coupling.

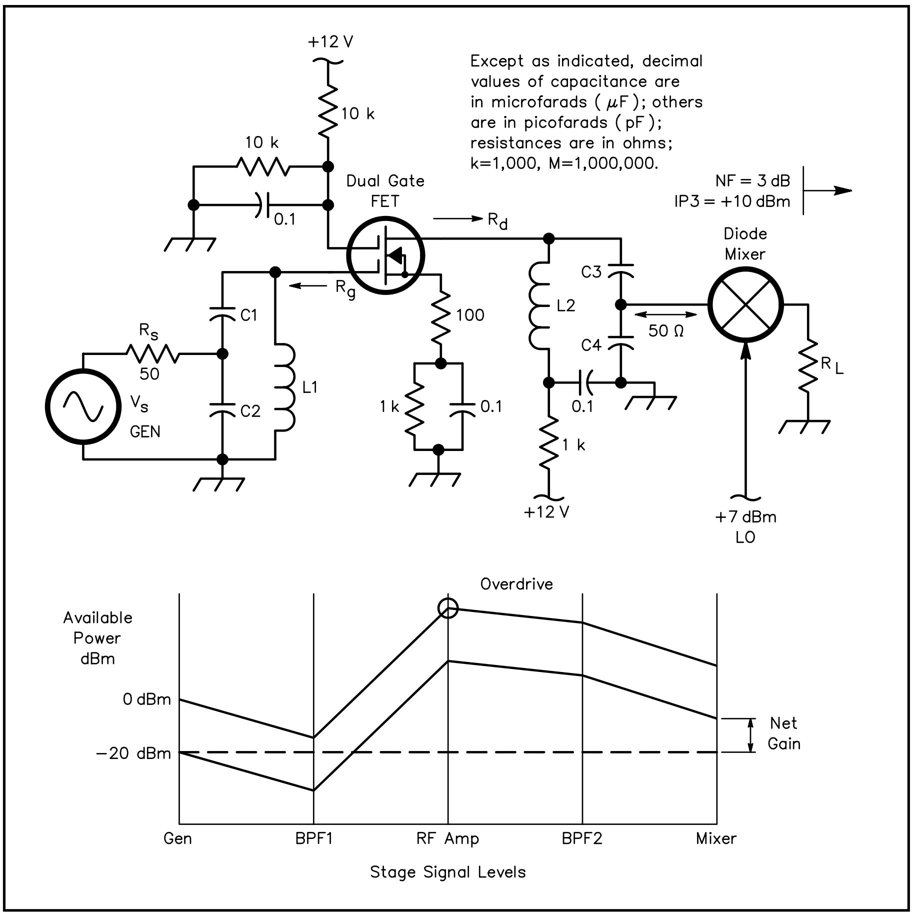


Fig 17.10—An example of cascaded stage design, a simple receiver front end.

We come now to the mixer, whose available gain is about –6 dB. This is the difference between its available IF output power and its available RF input power. This is a fairly low-level mixer, so it can be easily overdriven if the RF gain is too high. Harmonic IMD and two-tone IMD can become excessive (see later discussion in this chapter). On the other hand, as we will discuss later, too little RF gain will yield a poor receiver noise figure.

The concepts of available gain and transducer gain were introduced earlier. If we multiply the available gains of the input filter, RF amplifier, interstage filter and mixer, we have the available gain of the entire combination. The transducer gain is the ratio of the power actually delivered to R_L to the power that is available from the generator. To get the transducer gain of the combination, multiply the available gain of the first three circuits by the transducer gain of the last circuit (the mixer). This concept may require some thought on your part, but it is one that is frequently used and it adds understanding to how circuits are cascaded. One example, the transducer gain of a receiver, compares the signal power available from the antenna with the power into the loudspeaker (a perfectly linear receiver is assumed).

Fig 17.10 also shows an example of a commonly used graphical method for the available gain of a cascade. The loss or increase of available power at each step is shown. As the input increases the other values follow. But at some point, measurements of linearity or IMD will show that some circuit is being driven excessively, as the example indicates. To improve performance at that point, we may want to make gain changes or take some other action. If the overload is premature, a more powerful amplifier or a higher-level mixer may be needed. It may be possible to reduce the gain of the RF amplifier by reducing the step-up in the input LC circuit or the drain load circuit, but this may degrade noise figure too much. This is where the "optimization" process begins.

A method that is often used in the lab is to plot the voltage levels at various points in the system. These voltages are easily measured with an RF voltmeter or spectrum analyzer, using a high-impedance probe. This is a convenient way to make comparative measurements, with the understanding that voltage values are not the same thing as power-gain values, although they may be mathematically related. Many times, these voltage measurements quickly locate excessive or deficient drive conditions during the design or troubleshooting process. Comparisons of measured values with previous measurements of the same kind on properly functioning equipment are used to locate problems.

Selectivity of Cascaded Stages

The simplified receiver example of Fig 17.10 shows two resonant circuits (filters) tuned to the signal frequency. They attenuate strong signals on adjacent frequencies so that these signals will not disturb the reception of a desired weak signal at center frequency. **Fig 17.11** shows the response of the first filter and also the composite response of both filters at the mixer input.

Consider first the situation at the output of the first filter. If a strong signal is present, somewhat removed from the center frequency, the selectivity of the first filter may just barely prevent excessive signal level in the RF amplifier. When this signal is amplified and filtered again by the second filter, its level at the input of the mixer may be excessive. Our system design problem is to coordinate the amplifier gain and second filter selectivity so that the mixer level is not too great. (A computer simulation tool, such as *ARRL Radio Designer*, can be instructive and helpful.) Then we can say that for that level of undesired signal at that frequency offset the cascade is properly designed.

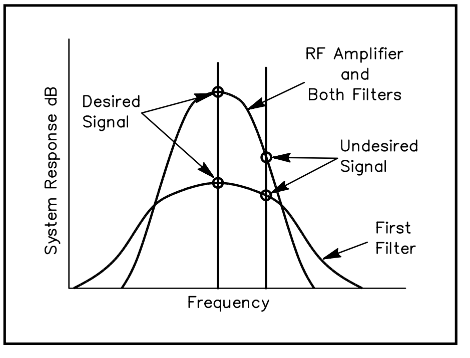


Fig 17.11—The gain and cumulative selectivity, between the generator and the mixer, of the example circuit in Fig 17.10.

The decisions regarding the "expected" maximum level and minimum frequency offset of the undesired signal are based on the operating environment for the equipment, with the realistic understanding that occasionally both of these values may be violated. If improvement is needed, it may then be necessary to (a) improve the selectivity, (b) use a more robust amplifier and mixer or (c) reduce amplifier gain. Very often, increases in cost, complexity and system noise factor are the byproducts of these measures.

Cascaded signal filters are often used to obtain a selectivity shape that has a flat top response and rapid or deep attenuation beyond the band edges. This method is often preferred over a single, more complex "brick wall" filter that has a very steep rate of attenuation outside the passband.

Noise Factor of a Cascade

In the example of Fig 17.10, the overall noise factor of the two-stage circuit is defined in the same way as for a single stage. It is the degradation in signal-to-noise ratio (S/N) from the signal generator to the output. This total noise factor can be found by direct measurement or by a stage-by-stage analysis. If we wish to optimize the total noise factor or look for trade-offs between it and other things such as gain and distortion, a stage-by-stage analysis is needed.

The definition of noise factor for a single stage applies as well to each stage in the chain. For each stage there is a signal and thermal noise generator, internal sources of excess noise and a noise bandwidth. In a cascade, the signal and thermal noise sources for a particular stage are found in the previous stage, as shown in **Fig 17.12A**. But this thermal noise source has already been accounted for as part of the excess noise for the previous stage, therefore this thermal noise must not be counted twice in the calculation. On this basis, Fig 17.12A derives the formula for the noise factor of a two-stage system. This formula can then be used to find the noise factor of the multisegment system in Fig 17.12B by applying it repetitively, first to stage N+1 and N, then to N and N–1,

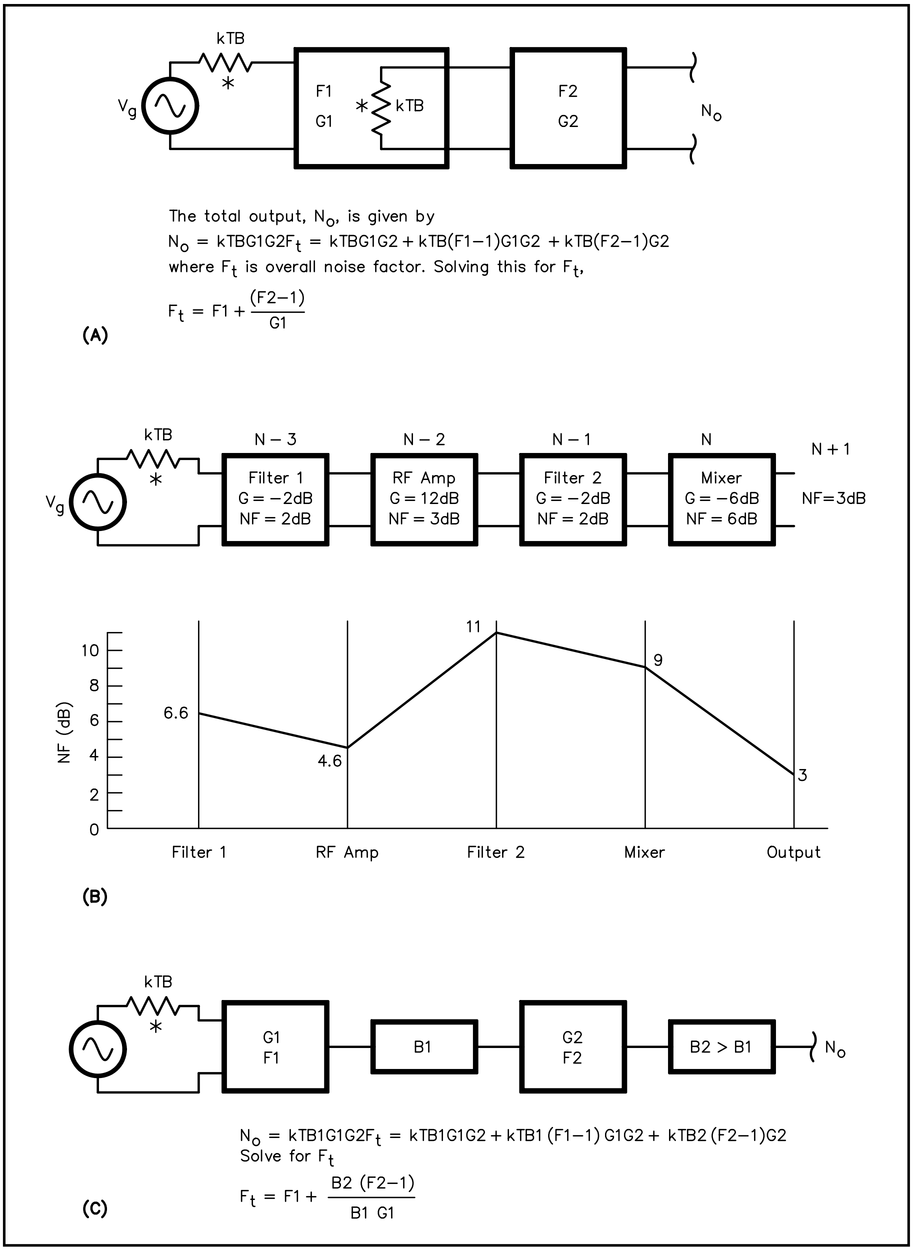


Fig 17.12—A: the noise factor of a two-stage network. B: cumulative noise factor for the example in Fig 17.10. C: noise factor when the bandwidth increases toward the output.

then to N–1 and N–2 and so on, where N is the total number of stages and N+1 is the rest of the system after the last stage. Fig 17.12B shows the cumulative noise figure (dB) at each point in the example of Fig 17.10. The graphical method aids the analysis visually.

Note that the diode mixer's noise figure approximately equals its gain loss. In applying the formula, if G1 is a lossy device, not an amplifier, then F1 equals its attenuation factor and G1 = 1/F1. Also, observe the critical role that values of RF amplifier gain and noise figure play in establishing the overall noise factor (or noise figure), despite the high noise figure that follows it.

In Fig 17.12A and B, we assumed that the noise bandwidth does not increase toward the output. If the noise bandwidth does increase toward the output a complication occurs. Fig 17.12C provides a modified formula that is more accurate under these conditions. This situation is often encountered in practice, as we will see, especially in the discussion of receiver design (Refs 9, 10).

Distortion in Cascaded Circuits

The IMD created in one stage combines with the distortion generated in following stages to produce a cumulative effect at the output of the cascade. The phase relationships between the distortion products of one stage and those of another stage can vary from 0° (full addition) to 180° (full subtraction). It is customary to assume that they add in-phase as a worst case. Under these conditions, **Fig 17.13** shows how to determine distortion at the input of a stage. Formulas are given for finding the third-order and second-order input intercept points in dBm. These formulas can be applied repetitively, in a manner similar to the noise-factor formula, to get the cumulative intercept point at each stage of the cascade. The output intercept point, in dBm, of a stage is equal to its input intercept point, in dBm, plus the gain, in dB, of the stage. When a purely passive, linear stage is part of the analysis, use a large value of intercept such as 100 dBm (10^7 W, Ref 11).

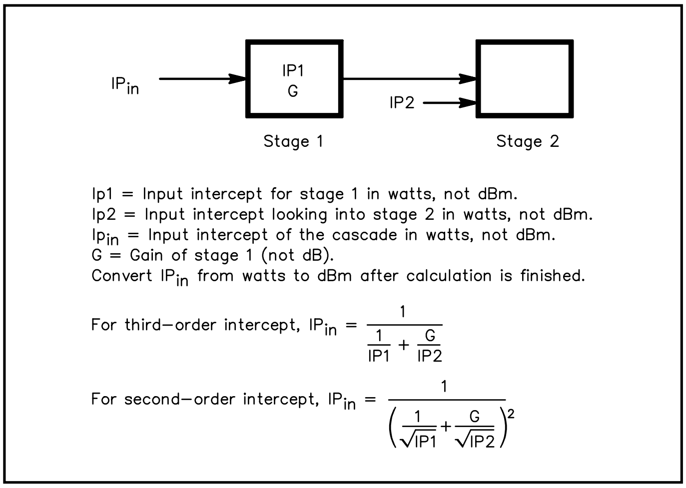


Fig 17.13—This is how IMD is calculated in cascade circuits.

Personal Computer Methods of Analysis

The methods of analysis of cascaded gain, noise factor, intercept point and dynamic range that we have described are very tedious when performed manually, especially when the effects of "what if?" value changes in various stages are desired. With the right kind of personal computer software that a home experimenter might be likely to possess, these calculations are performed almost instantaneously and a manual "optimization" process can be performed by the experimenter, using realistic values. In particular, the many available spreadsheet programs that have graph plotting features are very nice for this work. As values are changed, the modified tables and graph plots show up almost immediately. **Fig 17.14** illustrates these procedures for the example of Fig 17.10. The calculations are performed by subroutines called "macros." This particular example uses Microsoft Excel, but several other spreadsheets are available, and they work in a similar manner. When using these methods, keep in mind that changing one value can affect other values. For instance the gain, noise-factor and intercept-point values of an amplifier all interact.

For other computer analysis approaches, see Chapter 4 of *Personal Computers in the Ham Shack*, published by the ARRL.

Stage	Name	Gain(DB)	NF(DB)	IIP3 (DBM)	OIP3 (DBM)	NF TOT (DB)	IIP3 TOT (DBM)	DR(DB)	BW
1	Input filter	-2	2	100	100	6.4	6.2	95.9	1000
2	RF Amplifier	12	3	10	22	4.4	4.2	95.9	
3	RF Filter	-2	2	100	100	11.0	17.5	100.3	
4	Mixer	-6	6	25	19	9.0	15.5	100.3	
5	Output		3	10				106.7	
	Total Gain	2							

Example data sheet for cumulative noise figure, third-order intercept and spurious-free dynamic range for a 1000-Hz noise bandwidth.

NFTOT	IIPTOT	SFDR
=ARGUMENT("NF1")	=ARGUMENT("IIP1")	=ARGUMENT("BW")
=ARGUMENT("NF2")	=ARGUMENT("IIP2")	=ARGUMENT("NF")
=ARGUMENT("GAIN1")	=ARGUMENT("GAIN1")	=ARGUMENT("IP")
=10^(NF1/10)	=0.001*10^(IIP1/10)	=2/3*(IP+174-10*LOG10(BW)-NF)
=10^(NF2/10)	=0.001*10^(IIP2/10)	=RETURN(C5)
=10^(GAIN1/10)	=10^(GAIN1/10)	
=A5+(A6-1)/A7	=1/(1/B5+B7/B6)	
=10*LOG10(A8)	=10*LOG10(B8/0.001)	
=RETURN(A9)	=RETURN(B9)	

Example listing of macro subroutines to calculate cumulative noise factor, intercept point and spurious-free dynamic range.

Fig 17.14—An example of a personal computer Microsoft Excel spreadsheet analysis of the example in Fig 17.10.

Coupling Networks

In radio systems, the design of the transition network between one building block and the next is as important as the design of the blocks themselves. Impedance transformation, frequency response, adjustability and complexity are the main issues to consider. There are two general types of networks:

- Resonant networks with tuned circuits that are centered at some frequency, f, and cover perhaps no more than ±0.3 f, and
- Broadband networks that cover at least a 2:1 frequency ratio.

Complex multielement filters are treated in the **Filters** chapter. Here we are concerned with simpler networks that, although they may have some selectivity, are used mostly to transform impedances and tune out reactance.

Resonant Networks

Resonant matching circuits are examples of what are called "ladder" networks, consisting of alternating shunt and series elements. **Fig 17.15** is one simple and very useful example, the pi network, which we will consider briefly in order to get a general idea of how resonant matching circuits function.

In 17.15A the driver circuit, stage 1, has an output capacitance C_{out} that we would like to "tune out" or "absorb" into the pi network so that the impedance to ground of stage 1 will be a pure resistance, that is, R_{load} in parallel with R_{out}. The next stage, stage 2, has an input resistance shunted by a capacitance, C_{in}. We would like to absorb this capacitance into the pi network also. The pi network then looks like 17.15B. The problem is then to define C1, L and C2 so that the input resistance to the pi network is the desired value of R_{load}, which establishes the desired value of gain for stage 1. The procedures for finding these values are given in the **RF Power Amplifiers** chapter and the reader is encouraged to study these methods and become competent in their use.

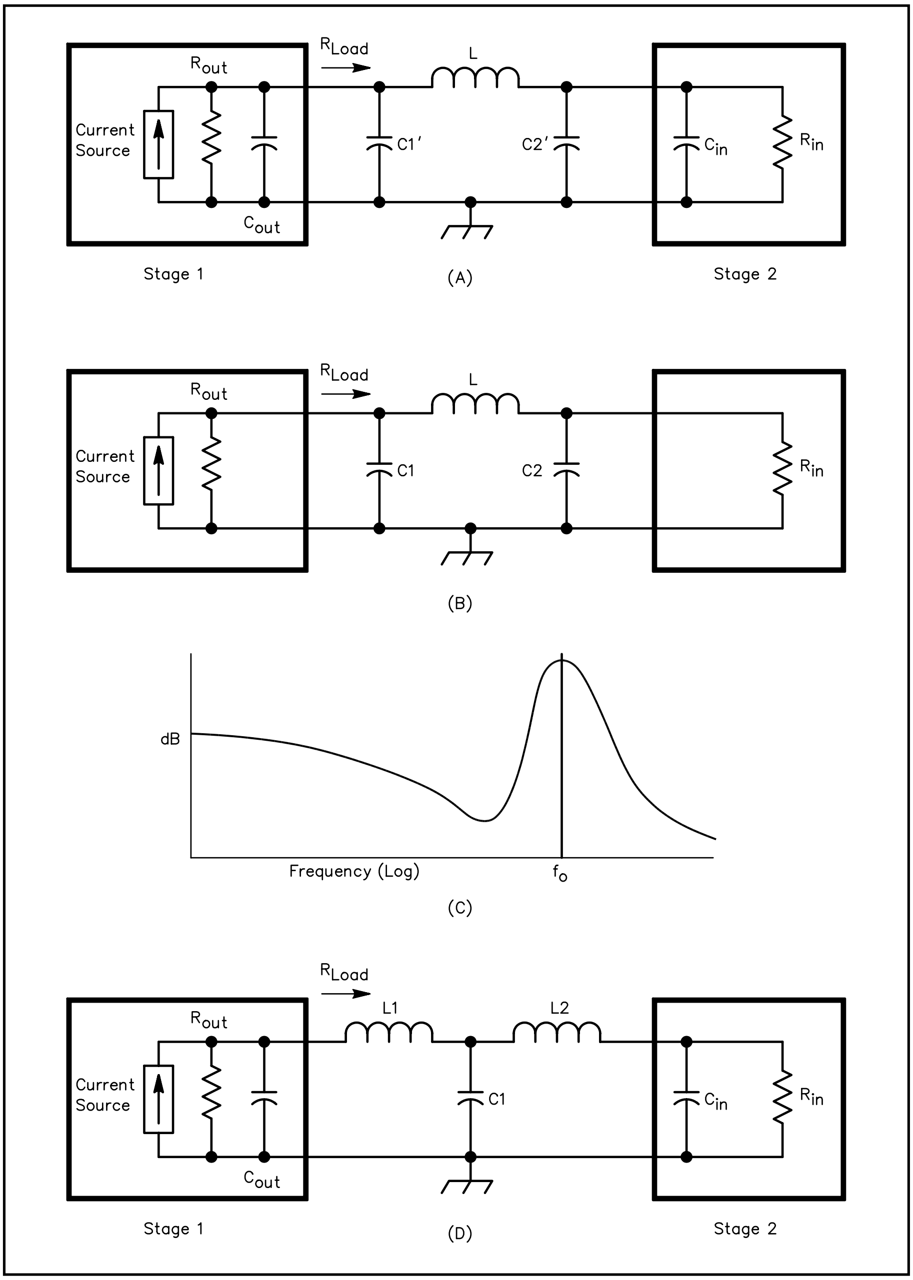


Fig 17.15—A pi network is used to absorb stray C, provide some selectivity and transform impedances at frequency f_0.

From a system design point of view, we note that the selectivity of the network depends on the values of C1 and C2. If they are large, the selectivity becomes sharp, perhaps too sharp. The answer is to reduce the C values and if necessary (if C_{out} and C_{in} are large) load the network resistively at input or output or both so that the selectivity can be reduced. More complex networks can expand the frequency response and these are also covered in the **RF Power Amplifiers** chapter.

Note that Fig 17.15B resembles a conventional three element low-pass filter. The difference is that the values are chosen to get the resonance effect at frequency f_0 (Fig 17.15C) and also the desired impedance transformation at f_0. These filters, for 2, 4, 6 and so on elements, are known as "Impedance Transforming Chebyshev" filters (Ref 12). For filters with 3, 5, 7 and so on elements, see Ref 13. The circuit has "low-pass" properties, as seen in Fig 17.15C. With shunt inductors and series capacitance, the response of Fig 17.15C would be reversed, left to right, and would have "high-pass" properties.

Another frequently used simple ladder network has a "T" configuration, shown in Fig 17.15D. It is preferred for certain applications and is discussed in detail later in this chapter under "Network Equations." Also, the tuned circuits in Fig 17.4 and Fig 17.10 are ladder networks of a slightly different kind, shown in a slightly different way. Knowledge in depth of tuned ladder networks, with the aid of the Smith Chart and the *ARRL MicroSmith* software, is a valuable resource for the radio circuit designer (Refs 14, 15).

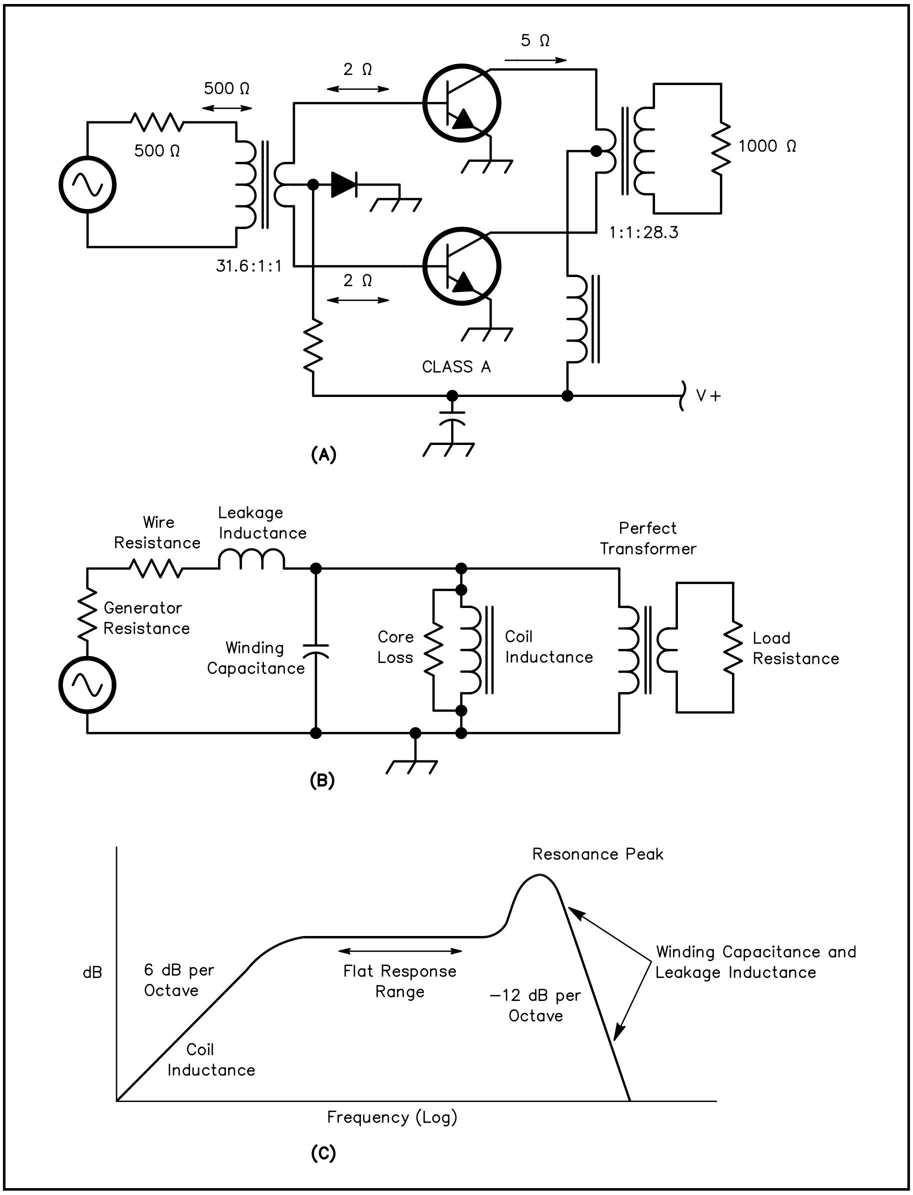


Fig 17.16—Conventional transformers in an RF power amplifier. Leakage reactances, stray capacitances and core magnetizations limit the bandwidth and linearity, and also create resonance peaks.

Broadband Transformers

The most important broadband impedance transforming device is the transformer. We will briefly consider two types: the so-called "conventional" transformer and the transmission-line transformer. Here we briefly review the merits and shortcomings of each, but this subject is so large and complex that we must defer to other chapters and the references for the most part. We are mainly interested in some frequently encountered examples of how transformers are used to solve radio system-design problems.

Conventional Transformers

Fig 17.16 shows a push-pull amplifier that we will use to point out the main properties and the problems of conventional transformers. The medium of signal transfer from primary to secondary is magnetic flux in the core. If the core material is ferromagnetic then this is basically a nonlinear process that becomes increasingly nonlinear if the flux becomes too large or if there is a dc current through the winding that biases the core into a nonlinear region. Nonlinearity causes harmonics and IMD. Push-pull operation eliminates the dc biasing effect if the stage is symmetrical. The magnetic circuit can be made more linear by adding more turns to the windings. This reduces the ac volts per turn, increases the reactance of the windings and therefore reduces the flux. For a given physical size, however, the wire resistance, distributed capacitance and leakage reactance all tend to increase as turns are added. This reduces efficiency and bandwidth. Higher permeability core materials and special winding techniques can improve things up to a point, but eventually linearity becomes more difficult to maintain. Fig 17.16B is an approximate equivalent circuit of a typical transformer. It shows the leakage reactance and winding capacitance that affect the high-frequency response and the coil inductance that affects the low-frequency response. Fig 17.16C shows how these elements determine the frequency response, including a resonance peak at some high frequency.

The transformers in a system are correctly designed and properly coordinated when the total distortion caused by them is at least 10 dB less than the total distortion due to all other nonlinearities in the system. Do not over-design them, in relation to the rest of the equipment. During the design process, distortion measurements are made on the transformers to verify this.

The main advantage of the conventional transformer, aside from its ability to transform between widely different impedances over a fairly wide frequency band, is the very high resistance between the windings. This isolation is important in many applications and it also eliminates coupling capacitors, which can sometimes be large and expensive.

In radio circuit design, conventional transformers with magnetic cores are often used in high-impedance RF/IF amplifiers, in high-power solid-state amplifiers and in tuning networks such as antenna couplers. They are seldom used any more in audio circuits. Hybrid transformers, such as those in Fig 17.9, are often "conventional."

Transmission-Line Transformers

A transmission line, consisting of two parallel or concentric (coaxial) conductors separated by a dielectric, is a self-contained system. That is, an electric field and a magnetic flux exist between the conductors, and these fields both decay very rapidly everywhere else. In **Fig 17.17**A, the generator is grounded and the power travels through the transmission line to the load. The load resistance is equal to the characteristic impedance, Z_0, of the transmission line. The study of transmission lines at high frequencies tells us that the bottom of the load, point D, may not be at ground potential because of the impedance along the length of the line, called the "common-mode impedance," between points A and D (and also between B and C). Points C and D "float" above ground, except at exact multiples of a half wavelength, where the impedances to ground at A and D (or B and C) would be the same.

This ungrounding effect is greatly enhanced over a very wide frequency range by using a high-permeability magnetic material in conjunction with a transmission line that is only a small fraction of a wavelength (at the highest frequency of interest), in the manner shown in Fig 17.17B. The first important feature of this arrangement is that when operated as shown there is much less magnetic flux in the core; the transmission line does the work. Therefore the common-mode impedance can be made quite large using a relatively small core (or cores). The load is therefore well isolated from ground (it floats). Observe also that any point of the output load resistor can be connected to ground with impunity (except as noted below). Often the center point is grounded as shown in Fig 17.17B.

The second feature is that one or multiple lengths of core-loaded transmission line can be used to

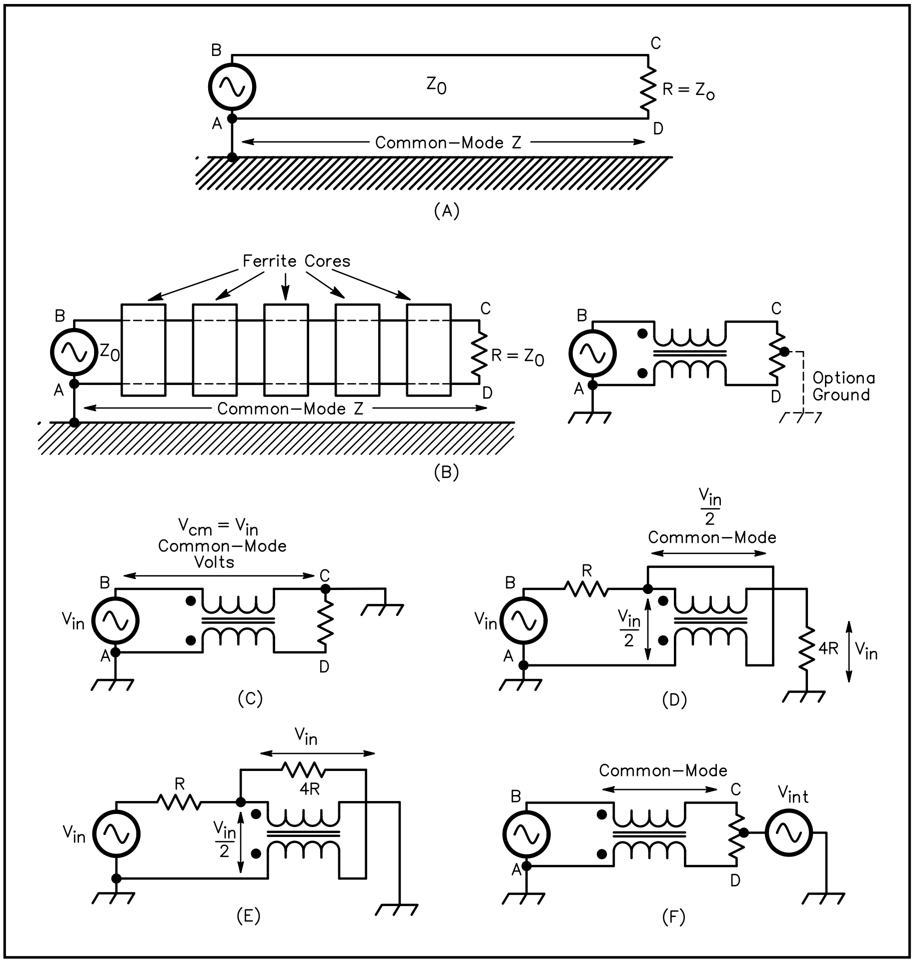


Fig 17.17—Transmission-line transformers and some of their properties. A: common-mode impedance along the length of a transmission line. B: "loading" the line with ferrite cores helps isolate the load from ground and the generator. C: a common-mode voltage appears aross the core due to a ground connection at the output. Such common-mode voltages bring Faraday's Law into play. D and E: a 4:1 impedance transformer with a common-mode voltage. F: an unexpected voltage source causes a common-mode voltage.

design impedance transforming transmission-line transformers that have some very desirable properties. These transformers utilize the principles of traveling waves in a dielectric medium for their operation. They can operate between any combination of balanced or unbalanced input or output.

The most important property of these transformers is that core magnetization can be greatly reduced, as compared to conventional transformers. Transformers of relatively small size can convey large signals with low loss and low distortion. However, problems due to increased amounts of flux in the magnetic core(s) can occur when there is a return path to ground at the output end, as shown in Fig 17.17B when the optional ground is connected, and also in Fig 17.17C. This grounding is responsible for an increased common-mode voltage V_{cm} across the core(s), which in turn increases both magnetic flux in the core(s) and perhaps also dielectric losses in the core(s). The problem also occurs in various transformer connections as shown in Fig 17.17D and E. Flux problems are most prevalent at the lower end of the frequency range. Power loss in the cores increases and in transmitters core heating becomes a greater problem. An extraneous voltage or current source in the circuit also causes problems as shown in Fig 17.17F. In all of these situations, use cores of sufficient cross-sectional area, winding length and permeability to get the desired common-mode impedance and linearity (Refs 16, 17, 18, 19). Also, methods are often feasible (see References) that reduce the common-mode voltage drops described above. In conclusion, the correct use of transmission-line transformers requires an understanding of the common-mode voltage gradients across the cores and how to deal with them. Note also that the generator "ground" and the load "ground" may be totally unrelated to each other. For example, the load "ground" may be at the top of a tower.

There is a very large collection of literature on this subject and space does not permit a detailed rundown on every aspect. In particular, the phase lags that occur in the short lengths of transmission lines can sometimes cause certain kinds of problems. The "equal delay" or Guanella design is then used.

Transmission-line transformers usually operate at low impedance levels, require transmission lines of certain Z_0 values (for maximum frequency response) and can be easily built for low ratios of input to output impedance. The literature (also see the **RF Power Amplifiers** chapter) shows clever ways to build more complex designs for higher impedance ratios. One variation of the transmission-line transformer discussed in the literature is the "twisted wire" transformer, where the requirement for the exactly correct Z_0 is relaxed and a smaller frequency range is acceptable. These are often easier to build than true transmission-line transformers and, if properly done, will work over two or three octaves (Refs 16, 17, 18, 19). A later section of this chapter gives more details about transmission-line transformers.

The Amateur Radio Communication Channel

In order to design radio equipment it is first necessary to know what specifications the equipment must have in order to establish and maintain communication. This is a very large and complex subject that we cannot fully explore here, however, it is possible to point out certain properties of the communication channel, especially as it pertains to Amateur Radio, and to discuss equipment requirements for successful communication. The “channel” is:

- the frequency band that is being transmitted and to which the distant receiver is tuned, and
- the electromagnetic medium that conveys the signal.

The Amateur Radio bands are, in fact, a very difficult arena for communications and a severe test of radio-equipment design. The very wide range of received signal levels, the high density of signals whose channels often overlap or are closely adjacent, the relatively low power levels and the randomness (the lack of formal operating protocols) are the main challenges for Amateur Radio equipment designers. An additional challenge is to design the equipment for moderate cost, which often implies technical specifications that are somewhat below commercial and military standards. These relaxed standards sometimes add to the amateur’s problems.

Received Noise Levels

There are three major sources of noise arriving at the receive antenna:

- atmospheric noise generated by disturbances in the Earth’s environment,
- galactic noise from outer space and
- noise from transmitters other than the desired signal.

Let’s briefly discuss each of these kinds of noise.

Atmospheric noise (including man-made noise) is maximum at frequencies below 10 MHz, where it has *average* values about 40 dB above the thermal noise at 290 K (K = Kelvins, absolute temperature). Above 10 MHz, its strength decreases at 20 dB per *octave*. At VHF and above, it is of little importance (Ref 20). However, various studies have found that at certain times and locations and in certain directions this noise approaches the level of thermal noise at 290 K, even at the lower frequencies. Therefore the conventional wisdom that a low receiver noise figure is not important at low HF is not completely true. Amateurs, in particular, exploit these occurrences, and most amateur HF receivers have noise figures in the 8 to 12-dB range for this reason, among others. A very efficient antenna at a low frequency can modify this conclusion, though, because of its greater signal and noise gathering power (for example, a half-wave dipole gathers about 12 dB more power at 1.8 MHz than a half-wave dipole at 30 MHz (Ref 21)). When the noise level is high, an attenuator in the antenna lead can reduce receiver vulnerability to strong interfering signals without reducing the S/N ratio of weaker signals. In other words, the system (that is, receiver plus noisy antenna) dynamic range is improved (the receiver intercept point increases and the system noise is reduced). The antenna noise, after attenuation, should be several dB above the receiver internal noise. This is a typical example of a communication-link design consideration that may not be necessary if the receiver is of high quality.

Receive Antenna Directivity

If the receive antenna has gain, and can be aimed in a certain direction, it often happens that atmospheric noise is less in that direction. A lower receiver noise figure may then help. Or, if the noise arrives uniformly from all directions but the desired signal is increased by the antenna gain, then the S/N ratio is increased. That is, the noise is constant but the signal is greater. (Explanation: if the noise is the same from all directions the high-gain antenna receives more noise from the desired direction but rejects noise from other directions; therefore the total received noise tends to remain constant.) This is one of the advantages of the HF rotary beam antenna. The same gain can also cause strong undesired signals to challenge the receiver’s dynamic range (or null out an undesired signal).

Galactic Noise

The *average* noise level from outer space is about 20 dB above that of thermal noise at 20 MHz and decreases at about 20 dB per *decade* of frequency (Ref 20). But at microwave frequencies, high-gain antennas with very-low-noise-figure receivers are able to locate sources of relatively intense (and very low) galactic noise. Amateurs working at microwave frequencies up to 10 GHz go to great lengths to get their antenna gains and receiver sensitivities good enough to take advantage of the high and low noise levels.

Transmitter Noise

Fig 17.18A shows the spectral output of a typical amateur transmitter. The desired modulation lies within a certain well defined bandwidth, which is determined by the type of modulation. Because of unavoidable imperfections in transmitter design, there are some out-of-band modulation artifacts such as high-order IMD products. The signal filter (SSB, CW and so on) response also has some slope outside the passband. There is also a region of phase noise generated in the various mixers and local oscillators (LOs). These phase noise sidebands are "coherent." That is, the upper frequency sidebands have a definite phase relationship to the lower frequency sidebands. At higher values of frequency offset, a noncoherent "additive" noise shelf may become greater than phase noise and it can extend over a considerable frequency band. Other outputs such as harmonics and other transmitter-generated spurious emissions are problems.

The general design goals for the transmitter are:

1. Make the unavoidable out-of-band distortion products as small as technology and equipment cost and complexity will reasonably allow,
2. Design the synthesizers and other local oscillators and mixers so that phase noise, as measured in a bandwidth equal to that of the desired modulation, is less than the out-of-band distortion products in goal 1 and,
3. Make the wideband noise sufficiently small that the noise will be less than any unavoidable receiver noise at nearby receivers with the same bandwidth.

If the additive noise is very small, LO phase noise may come back into the picture. In narrowband systems such as Morse code (CW) it can be very difficult or impractical to make transmitted phase noise less than the normal Morse code sidebands (see later discussion on this topic). The general method to reduce wideband noise from the transmitter is to place the narrowband modula-

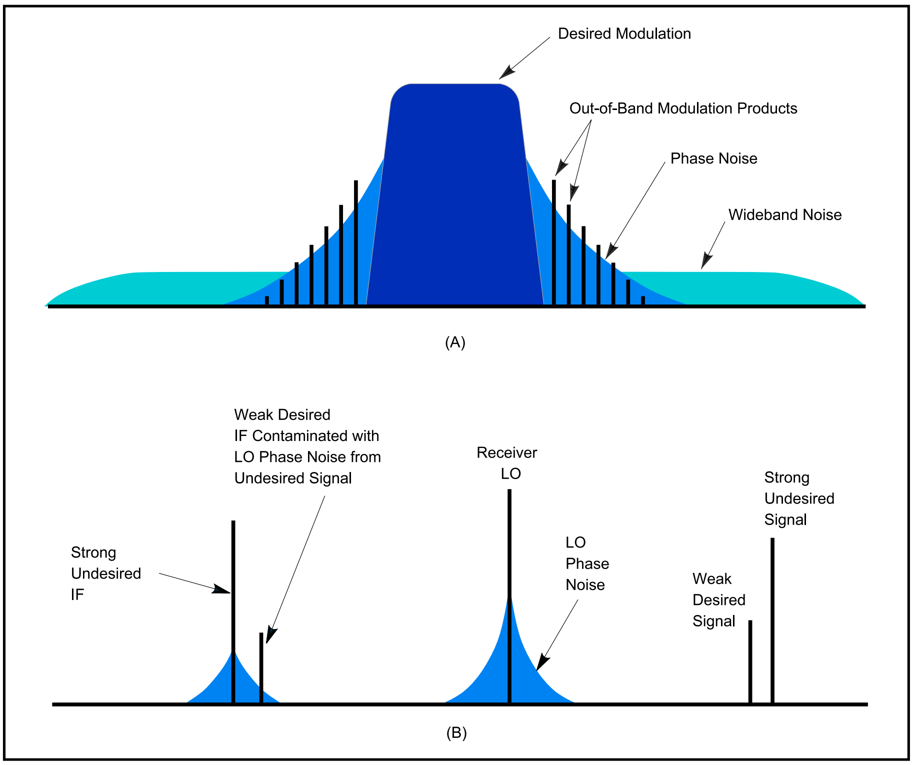


Fig 17.18—A: transmitter spectrum with discrete out-of-band products, phase noise and white noise. B: reciprocal mixing of LO phase noise onto an incoming signal.

tion band-pass filter at as high a signal level as possible and to follow that with a high-level mixer and then a low-noise first-stage RF amplifier.

Phase-noise amplitude varies with modulation. That is, the LO phase noise is modulated onto the outgoing signal by the "reciprocal mixing" process (the signal becomes the "LO" and the LO phase noise becomes the "signal"). If the actual LO to phase noise ratio is X dB, the ratio of the transmit signal to its phase noise is also X dB. In SSB the magnitude of the phase-noise sidebands is maximum only on modulation peaks. In CW it exists only when the transmitter is "key down." The additive noise, on the other hand, may be much more constant. If the power amplifiers are Class A or Class AB, additive noise does not require any actual signal and tends to remain more nearly constant with modulation.

In a communication link design, the receiver's culpability must also be considered. The receiver's LOs also generate phase noise that is modulated onto an adjacent-channel signal (reciprocal mixing) to produce an in-band noise interference, as shown in Fig 17.18B. In view of this, the transmitter and receiver share equal responsibility regarding phase noise, and there is little point in making either one a great deal better unless the other is improved also. Nevertheless, high-quality receivers with low phase noise exist, and they are vulnerable to transmitter phase noise. The converse situation also exists; receiver phase noise can contaminate a clean incoming signal (Refs 22, 23, 24).

Receiver Gain and Transmitter Power Requirements

The minimum level of a received signal is a function of the antenna noise level and the bandwidth. As just one example, for an HF SSB system with a 2.0-kHz bandwidth and a noise level 10 dB above thermal (–131 dBm in a 2.0-kHz band) the minimum readable signal, say 3 dB above the noise level, would be –128 dBm. Assume the receiver-generated noise is negligible. If the audio output to a loudspeaker is, say +20 dBm (0.1 W), then the total required receiver gain is 148 dB, which is an enormous amount of amplification. For a CW receiver with 200-Hz bandwidth, the minimum signal would be –138 dBm and the gain would be 158 dB. Receivers at other frequencies with lower noise levels can require even higher gains to get the desired audio output level. If the transmission path attenuation can be predicted or calculated, the required transmitter power can be estimated. These kinds of calculations are often done in UHF and microwave amateur work, but less often at HF (see the microwave receiver section for an example).

We do, however, get some "feel" for the receiver gain requirements, how the receiver interacts with the "channel," and that the minimum signal power is an almost incredibly small 1.6×10^{-16} W. On the other hand, amateur receiver S-meters are calibrated up to an input signal level of –13 dBm (60 dB above 100 mV from a 50-W source). Therefore the receiver must deal with a desired signal range of at least 115 dB (128 – 13) for the SSB example or at least 125 dB for the CW example (assuming that AGC limits the signal levels within the receiver).

Fading

Radio signals very often experience changes in strength due either to reflections from nearby objects (multipath) or, in the case of HF, to multiple reflections in the ionosphere. At a particular frequency and at a certain time, a signal arriving by multipath may decrease severely. The effect is noticed over a narrow band of frequencies called the "fading bandwidth." At HF, the center frequency of this fade band drifts slowly across the spectrum. Communication links are degraded by these effects, so equipment design and various communication modes are used to minimize them. For example, SSB is less vulnerable than conventional AM. In AM, loss of carrier or phase shift of the carrier, relative to the sidebands, causes distortion and reduces audio level.

The UHF/Microwave Channel

At frequencies above about 300 MHz, we need to account for the interaction of the Earth environment with the transmitted and received signals. Here are some of the things to consider:

1. Line-of-sight communications distance, as a function of receiver and transmitter antenna height.
2. Losses from atmospheric gasses and water vapor (above several gigahertz).
3. Temperature effects on paths: reflections, refractions, diffractions and transmission "ducts."
4. Atmospheric density inversions due to atmospheric pressure variations and weather fronts.
5. Tropospheric reflections and scattering.
6. Meteor scattering (mostly at VHF but occasionally at UHF).
7. Receive-antenna sky temperature.

Competitive amateur operators who are active at these frequencies become proficient at recognizing and dealing with these communication channel effects and learn how they affect equipment design. They become proficient at estimating channel performance, including path loss, receive system noise figure (or noise temperature), antenna radiation patterns and gain (Ref 25).

IMPROVING THE RADIO LINK

Aside from the above considerations, there are many other ways to get better communication. Here are a few of them:

Automatic Link Evaluation (ALE)

The personal computer has brought about the development of computer-supervised radios that search for a clear frequency that has the right propagation to support communication between two or more distant locations. These also use propagation forecasts. (Refs 26, 27, 28.)

Automatic Power Level Adjustment

By measuring the received signal level in comparison with noise and interference, it is possible to automatically tell the distant station to adjust power level (Refs 29,30).

Digital Communication

The use of digital modes of communication is increasing rapidly among radio amateurs. Using error correction, acknowledgments and repeats, very reliable work in noise and interference, using low power is being accomplished. An excellent overview by Steve Ford, WB8IMY, "Globe Hopping Digitally" in Jan 1998 *QST*, describes the various modes and their properties.

Operator Skill

In any mode of Amateur Radio communication, the operating and technical skills and also the social responsibility of the transmitting and receiving operators are still, and will always be, crucial parts of the communication link. The traditional SSB and CW modes are still very popular and enjoyable (and also effective if properly used), even though they are theoretically less effective than sophisticated digital modes. In this traditional environment, the operating skills of the individuals are even more important. The point we make in this chapter is that equipment features that enhance the operator's ability to "get the job done" are important parts of the equipment and system design considerations.

Repeaters

VHF and UHF communication distances for low-power gear are greatly increased by strategically located repeater stations. The **Repeaters, Satellites, EME and Direction Finding** chapter of this *Handbook* gives more coverage of this very popular technology.

Spread Spectrum

A few amateurs are experimenting with spread-spectrum techniques. This rapidly expanding technology is being used more and more in military, industrial, commercial and law-enforcement organizations. The rapidly oncoming "information highway," personal communications service (PCS) and computer-network systems are now using spread-spectrum methods to improve privacy and effectiveness, especially where many users share the spectrum. Powerful error-correction methods and automatic transmitter power-level management help to reduce information loss due to cochannel interference. (Refs 31, 32.)

Satellites

Communication over long terrestrial distances using Earth satellite transponders has become a popular mode, due to the pioneering efforts of a group of advanced amateurs over the past 40 years or so.

Modulation Methods

Quadrature amplitude modulation (QAM) has recently become more popular because of the advent of high-definition television (HDTV). QAM uses several discrete amplitude levels, combined with several discrete phase values, to permit a high information rate within a narrow bandwidth. This has potential for Amateur Radio.

Receiver Design Techniques

We will now look at the various kinds of receivers that are used by amateurs and at specific circuit designs that are commonly used in these receivers. The emphasis is entirely on analog approaches. Methods that use digital signal processing (DSP) for various signal processing functions are covered in the **DSP** chapter.

EARLY RECEIVER DESIGN METHODS

Fig 17.19 shows some early types of receivers. We will look briefly at each. Each discussion contains information that has wider applicability in modern circuit design, and is therefore not merely of historical interest. A lot of good old ideas are still around, with new faces.

The Crystal Set

In Fig 17.19A the antenna circuit (capacitive at low frequencies) is series resonated by the primary coil to maximize the current through both, which also maximizes the voltage across the secondary. The semiconductor crystal rectifier then demodulates the AM signal. This demodulation process utilizes the carrier of the AM signal as a "LO" and frequency translates (mixes) the RF signal down to baseband (audio). The rectifier and its output load impedance (the headphones) constitute a loading effect on the tuned circuit. For maximum audio output, a certain amount of coupling to the primary coil provides an optimum impedance match between the rectifier circuit and the tuning circuit. The selectivity is then somewhat less than the maximum obtainable. To improve the selectivity, reduce the secondary L/C ratio and/or decrease the coupling to the primary. Some decrease in audio output will usually result.

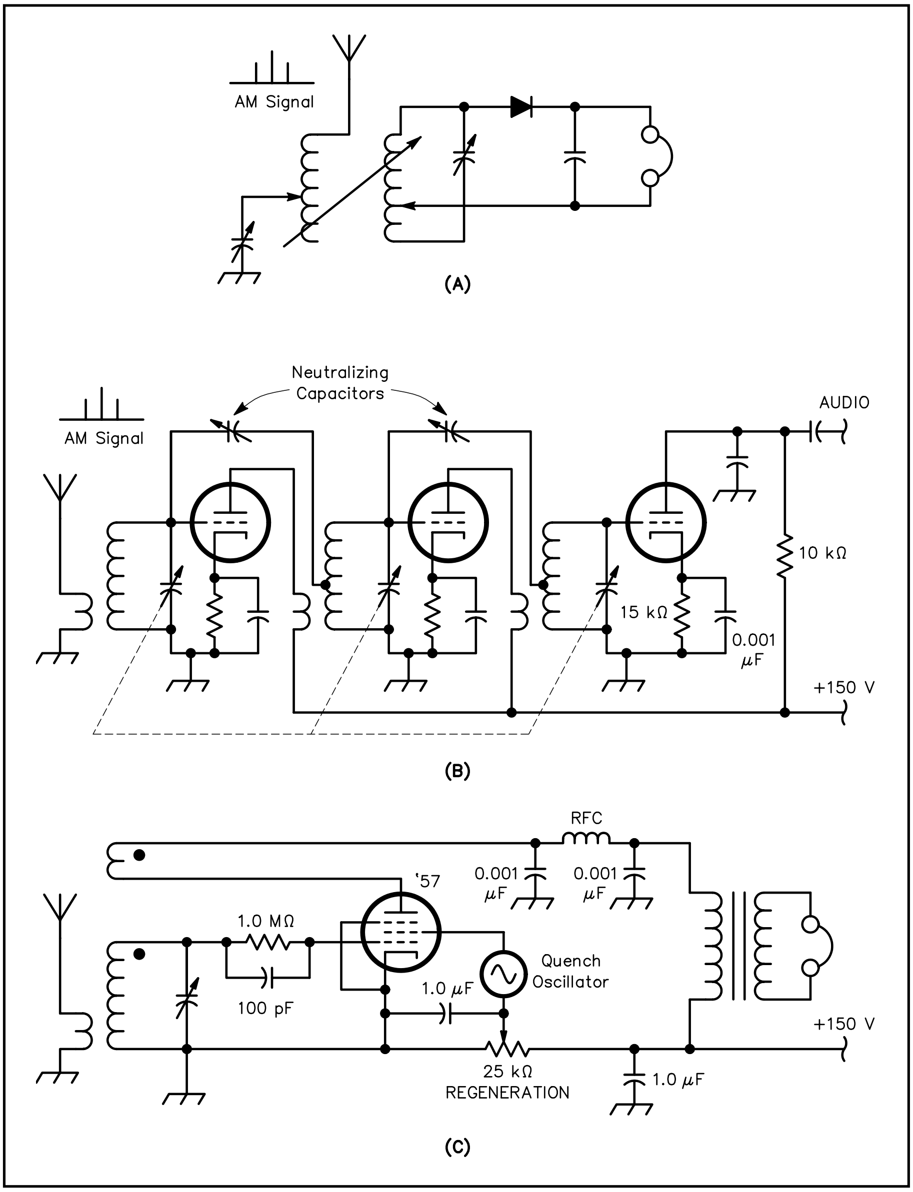


Fig 17.19—Early receiver designs. A: crystal set. B: tuned radio-frequency (TRF) receiver. C, a regenerative receiver.

This basic mechanism for demodulating an AM signal by using a rectifier is identical to that used in nearly all modern AM receivers. One important feature of this rectifier is a signal level "threshold effect" below which rectification quickly ceases. Therefore the crystal set, without RF amplification, is not very good for very weak signals. Early crystal receivers used large antennas to partially solve this problem, but they were vulnerable to strong signals (their dynamic range was not very good). However, a large antenna does make greater tuner selectivity possible (if needed) because looser coupling can be used in the tuner. That is, the loading of the secondary resonator by the antenna and rectifier can both be reduced somewhat.

There is one other interesting property of this detector. The two AM sidebands add in phase (coherently) at the audio output, but noise above and below the carrier frequency add in random phase (noncoherently). Therefore the detector provides a 3-dB improvement in signal-to-noise ratio, for a given average sideband power (Ref 33).

The Tuned Radio Frequency (TRF) Receiver

Fig 17.19B shows a TRF receiver. One, or possibly many, tuned RF stages are followed by a vacuum-tube implementation of the crystal rectifier (infinite-impedance square-law detector) described in the previous section. The RF amplification overcomes the threshold effect of the detector and the multiple tuned circuits (called "synchronous tuning"), usually isolated from each other by amplifier devices, allow much better selectivity. This selectivity is greatly reduced at the high end of the tuning range because the tuning capacitance becomes small. At the low end of the tuning dial, modulation sidebands may be rolled off or "clipped" due to excessive selectivity. This variation in selectivity, and also in gain, are the TRF's main drawbacks, which helped popularize the superheterodyne approach. The figure shows triode-tube amplifiers (they could also be three lead transistors: single-gate FETs or bipolar) that are neutralized to prevent self oscillation. This receiver is called a "Neutrodyne." Multigrid tubes or dual gate FETs do not need neutralization.

The Regenerative Receiver

Fig 17.19C (with the quench oscillator inactive) is a simple example of a regenerative receiver. The basic principle is that positive RF feedback, via the plate winding, is used to increase the RF gain up to and slightly beyond the point of self oscillation. With no signal, the internal shot and thermal noise peaks and a small bias voltage on the grid capacitor (caused by a very slight grid current) combine to produce a stable and self-limiting oscillation that is similar to the behavior of an ordinary oscillator, except that right at the peak of the oscillation cycle the amplification is extremely large and the detector is therefore very sensitive. The Q of the tuned circuit is greatly increased by the introduction of negative resistance (see the **AC/RF Sources** chapter), but the small grid-current loading limits the increase. With greater or lesser feedback, or if a strong signal appears, the gain and Q drop rapidly.

The self oscillation heterodynes with an incoming CW or SSB signal to produce an audio beat note. The main advantages of the "regen" are the beat note, the absence of a weak-signal threshold effect and amplification. Slightly below the oscillation point, makeshift AM reception is possible. From about 1920 to 1935, the regen was the favorite HF receiver among amateurs. It was the subject of much design and development by them. It also required considerable operating skill (Refs 34, 35).

The Superregenerative Receiver

Fig 17.19C (with the quench oscillator active) is a simple superregenerative receiver. The idea here is that the output of the 20-kHz quench oscillator modulates the detector and drives it through the point of oscillation and through the point of maximum gain and sensitivity at a super-audible rate. This results in an audio signal that is free of the audible heterodyne that the ordinary regen produces. Therefore, the receiver is more useful for AM or FM voice reception. Amplitude limitations, distortion and poor selectivity are inherent in super-regenerative receivers—this is the price paid for relative simplicity.

Super regens were used for many years for 30-100 MHz voice reception, where the high ratio of signal frequency to quench frequency made the circuit more manageable. The classic amateur article on this subject by Ross Hull (at ARRL) appears in the July 1931 *QST*. See also any edition of the famous *Radiotron Designer's Handbook* by Langford-Smith, circa 1953.

Modern Receiver Design Methods

The superheterodyne and the direct-conversion receiver are the most popular modern receivers and the chief topic of this discussion. Both were conceived in the 1915-1922 time frame. Direct conversion was used by Bell Labs in 1915 SSB experiments; it was then called the "homodyne" detector. E. H. Armstrong devised the superhet in about 1922, but for about 12 years it was considered too expensive for the (at that time) financially strapped amateur operators. The advent of "single signal reception," pioneered by J. Lamb at ARRL, and the gradual end of the Great Depression era brought about the demise of the regenerative receiver. We begin with a discussion of the direct conversion receiver, which has been rediscovered by amateur equipment builders and experimenters in recent years.

Direct Conversion (D-C) Receivers

The direct conversion (D-C) receiver, in its simplest form, has some similarities to the regenerative receiver:

- The signal frequency is converted to audio in a single step,
- An oscillator very near the signal frequency produces an audible beat note,
- Signal bandwidth filtering is performed at baseband (audio),
- Signals and noise (both receiver noise and antenna noise) on both sides of the oscillation frequency appear equally in the audio output. The image (on the other side of zero beat) noise is an "excess" noise that degrades the noise factor and dynamic range (Ref 36).

There are three major differences favoring the D-C receiver:

- There is no delicate state of regeneration involved. A low-gain or passive mixer of high stability is used instead.
- The oscillator is a separate and very stable circuit that is buffered and coupled to the mixer.
- The D-C (that uses modern circuit design) has much better dynamic range.

The regen has enormous RF gain (and Q multiplication) and therefore little audio gain is needed. The D-C delivers a very low-level audio that must be greatly amplified and filtered. RF amplification, band-pass filtering and automatic gain control (AGC) can be easily placed ahead of the mixer with beneficial results.

An enhancement of the D-C concept can perform a fairly large reduction of the signal and noise image responses mentioned above. It is a major technical problem, however, to get a degree of reduction over a wide frequency range, say 1.6-30 MHz, that compares with that easily obtainable using superheterodyne methods.

D-C RECEIVER DESIGN EXAMPLE

Fig 17.20 is a schematic diagram of a simple D-C receiver that utilizes all of the principles mentioned above except image rejection. The emphasis is on simplicity for both SSB and CW reception on the 14-MHz band. The LO is standard *Handbook* circuitry and is not shown. **Fig 17.21** shows a simple example of an active CW band-pass filter centered at 450 Hz.

The input RF filter shields the receiver from large out-of-ham-band signals and has a noise figure of 2 dB. The grounded-gate RF amplifier has a gain of 8 dB, a noise figure of 3 dB and an input third-order-intercept point of 18 dBm. Its purpose is to improve a 14-dB noise figure (at the antenna input) without the amplifier to about 8.5 dB. It also eliminates any significant LO conduction to the antenna and provides opportunities for RF AGC. The total gain ahead of the mixer is about 6 dB, which degrades the IMD performance of the mixer and subsequent circuitry somewhat. However, the receiver still has a third-order intercept (IP3) of about 6 dBm for two tones within the range of the audio filter. The IP3 is 11 dBm (quite respectable) when one of the two tones is outside the range of the low-pass audio filter that precedes the first audio amplifier. The low-pass filter protects the audio amplifier from wideband signals and noise. The intercept point could be improved by eliminating the RF amplifier, but the antenna input noise figure would then be much worse. This is a common trade-off decision that receiver designers must make.

The above analysis would be correct for a conventional receiver, but in this case there is a small complication that we will mention only briefly. The noise sources ahead of the mixer, both thermal and excess, that are on the image side (the side of the carrier opposite a weak desired signal) are translated to the baseband and appear as an increased noise level at the input of the first audio amplifier. If the SFDR (previously defined) in a 1000-Hz bandwidth were ordinarily 95 dB, using the above numbers, the actual SFDR would be perhaps 2.5 dB less. An image-reject mixer would correct this problem. Observe that the low noise figure of the RF amplifier minimizes the gain needed to get the desired overall noise figure. Also, its good intercept point minimizes strong-signal degradation contributed by the amplifier. The

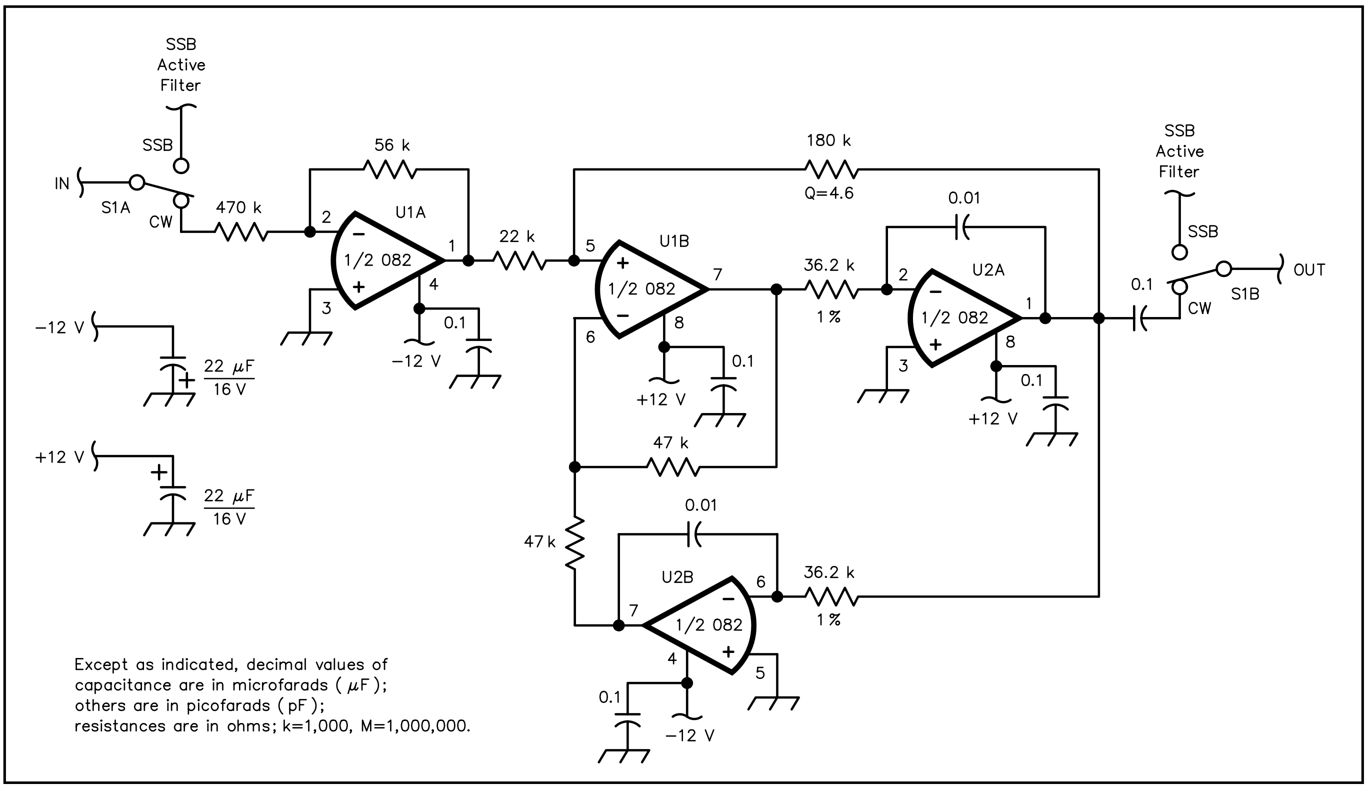


Fig 17.21—Schematic diagram of an active CW band-pass filter for the D-C receiver in Fig 17.20.

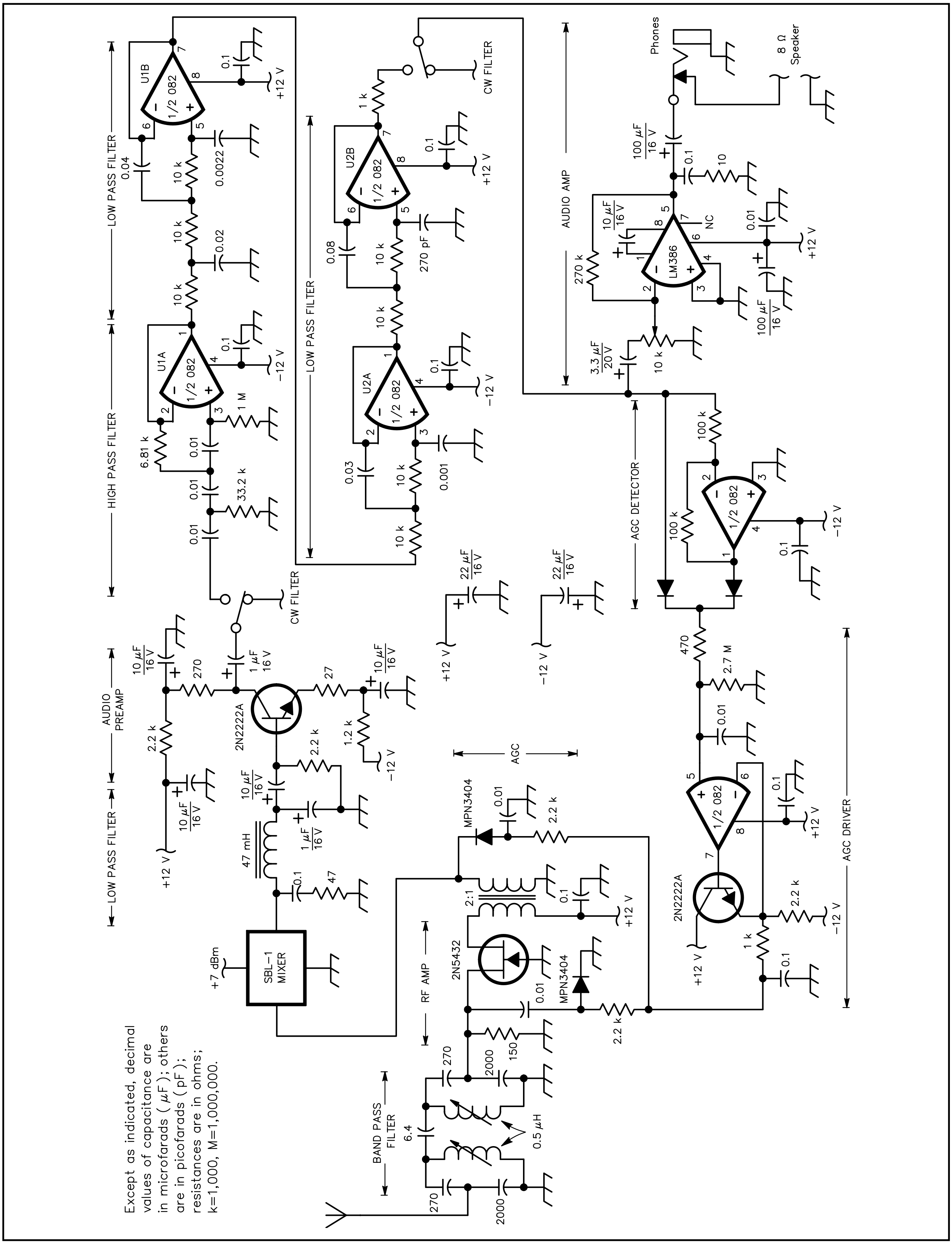


Fig 17.20—Schematic diagram of a simple D-C receiver for 20 m SSB and CW. Image cancellation is not used, but RF amplification and audio derived AGC are included.

input BPF eliminates problems from second-order IMD. Flicker-effect (1/f) noise in the mixer audio output may also be a problem, which the RF amplification reduces and the mixer design should minimize.

The preceding analysis illustrates the kind of thinking that goes into receiver design. If we can quantify performance in this manner we have a good idea of how well we have designed the receiver.

For the circuit in Fig 17.20 and the numbers given above, the gain ahead of the first audio amplifier is 0 dB. As stated before, this amplifier is protected from wideband interference by the 2-element low-pass audio filter ahead of it, which attenuates at a rate of 12 dB per octave. This filter could have more elements if desired. By minimizing front-end gain, the tendency for the audio stages to overload before the earlier stages do so is minimized—if the audio circuitry is sufficiently "robust." This should be checked out using two-tone and gain-compression tests on the audio circuits. Audio-derived AGC helps prevent signal-path overload by strong desired (in-band) signals. Additional AGC can be applied in the audio section by using a variable-gain audio op amp (MC3340P).

The audio SSB and CW band-pass filters are simplified active op-amp filters that could be improved, if desired, by using methods mentioned in some of the references. In an advanced design, digital signal processing (DSP) could be used. Good shape factor is the main requirement for good adjacent-channel rejection. Good transient response (maximally flat group delay) would be a "runner-up" consideration. Digital FIR filters and analog elliptic filters are good choices.

Image Rejection in the D-C Receiver

Rejection of noise and signals on one side of the LO is a major enhancement and also a major complication of the D-C receiver (Refs 37, 38).

Fig 17.22 shows two correct ways to build an image canceling mixer and one incorrect way. The third way does not perform the required phase cancellations for image reduction. In practice, two ±45° phase shifters are used, rather than one 90° stage. As mentioned before, it is very difficult to get close phase tracking over a wide band of signal and LO frequencies. In amateur equipment, front panel "tweaker" controls would be practical.

The block diagram in **Fig 17.23** is a typical approach

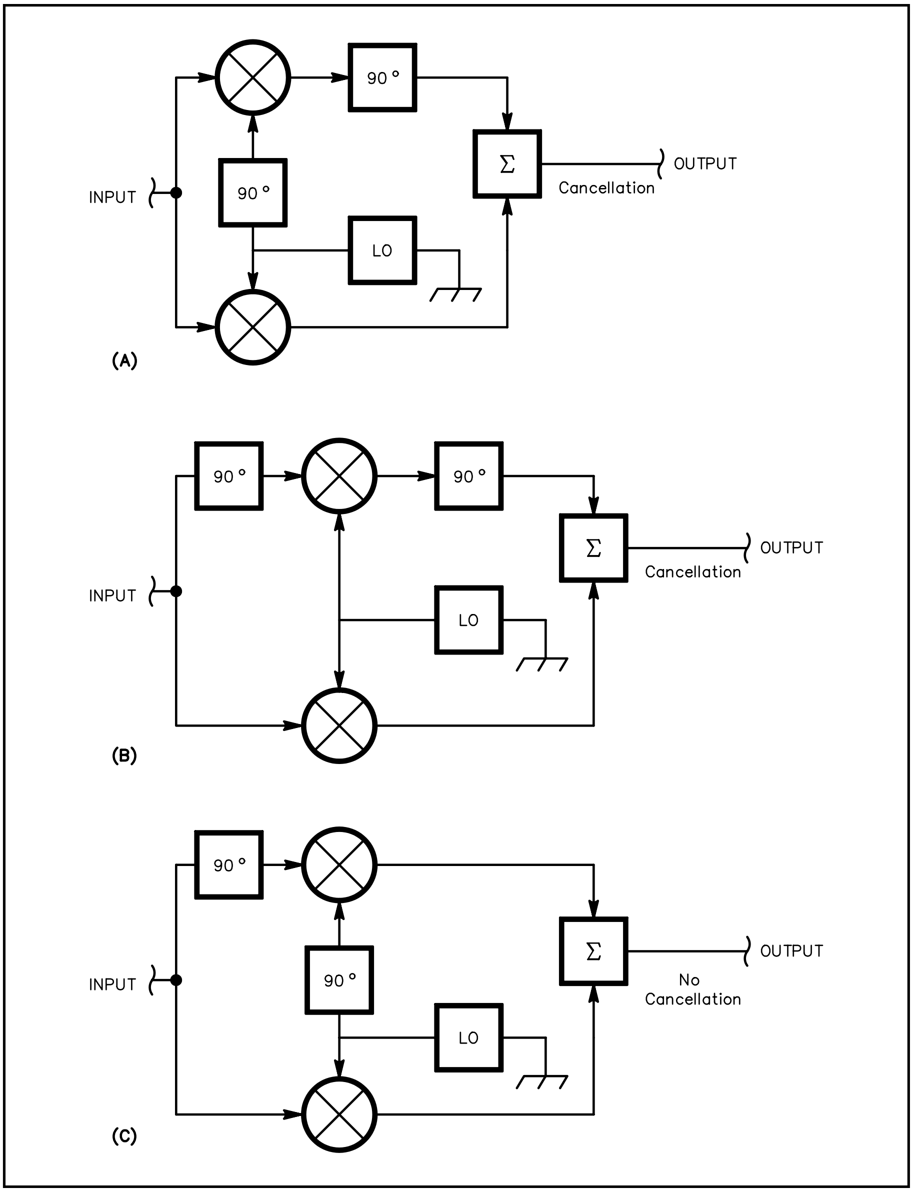


Fig 17.22—Both A and B are workable image cancelling mixer stages. The scheme at C will not cancel image signals.

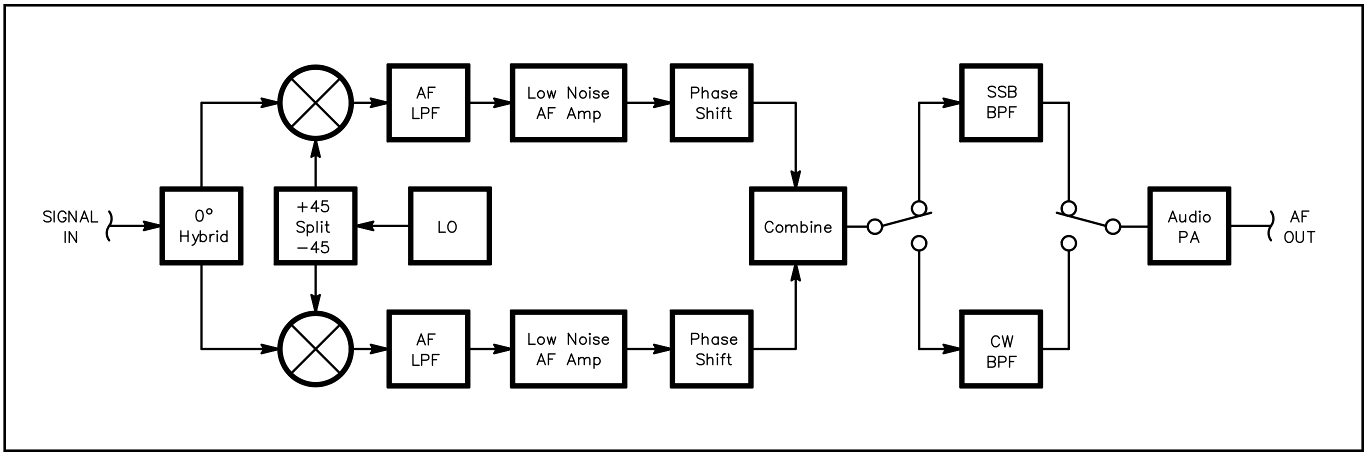


Fig 17.23—Typical block diagram of an image cancelling D-C receiver.

to an image-canceling D-C receiver. The two channels, including RF, mixers and audio must be very closely matched in amplitude and phase. The audio phase-shift networks must have equal gain and very close to a 90° phase difference. **Fig 17.24** relates phase error in degrees and amplitude error in dB to the rejection in dB of the opposite (image) sideband. For 30 or 40 dB of rejection, the need for close matching is apparent.

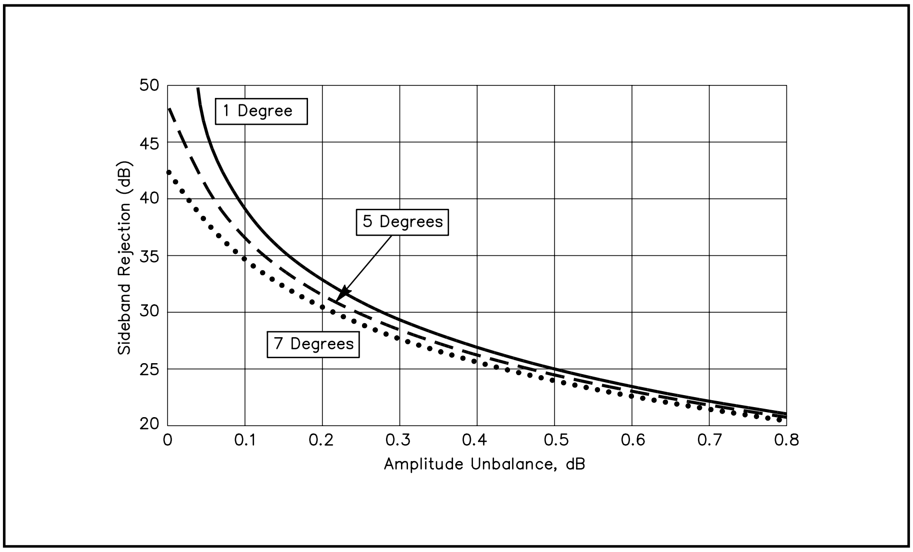


Fig 17.24—A plot of sideband rejection versus phase error and amplitude unbalance.

Audio Phase Shifters

Fig 17.25A shows an example of an audio phase-shift network. The stage in 25B is one section, an active "all-pass" network that has these properties:

- The gain is exactly 1.0 at all frequencies and
- The phase shift changes from 180° at very low frequency to 0° at very high frequency.

The shift of this single stage is +90° at $f=1/(2\pi RC)$. By cascading several of these with carefully selected values of RC the set of stages has a smooth phase shift across the audio band. A second set of stages is chosen such that the phase difference between the two sets is very close to 90°. The choices of R and C values have been worked out using computer methods; you can also find them in other handbooks (Ref 39). Fig 17.25C shows the phase error for two circuits like the one shown in Fig 17.25A. Note the rapid increase in error at very low audio frequencies (an improvement would be desirable for CW work). These frequencies should be greatly attenuated by the audio band-pass filters that follow.

D-C Receiver Problem Areas

Because of the high audio gain, microphonic reactions due to vibration of low-level audio stages are

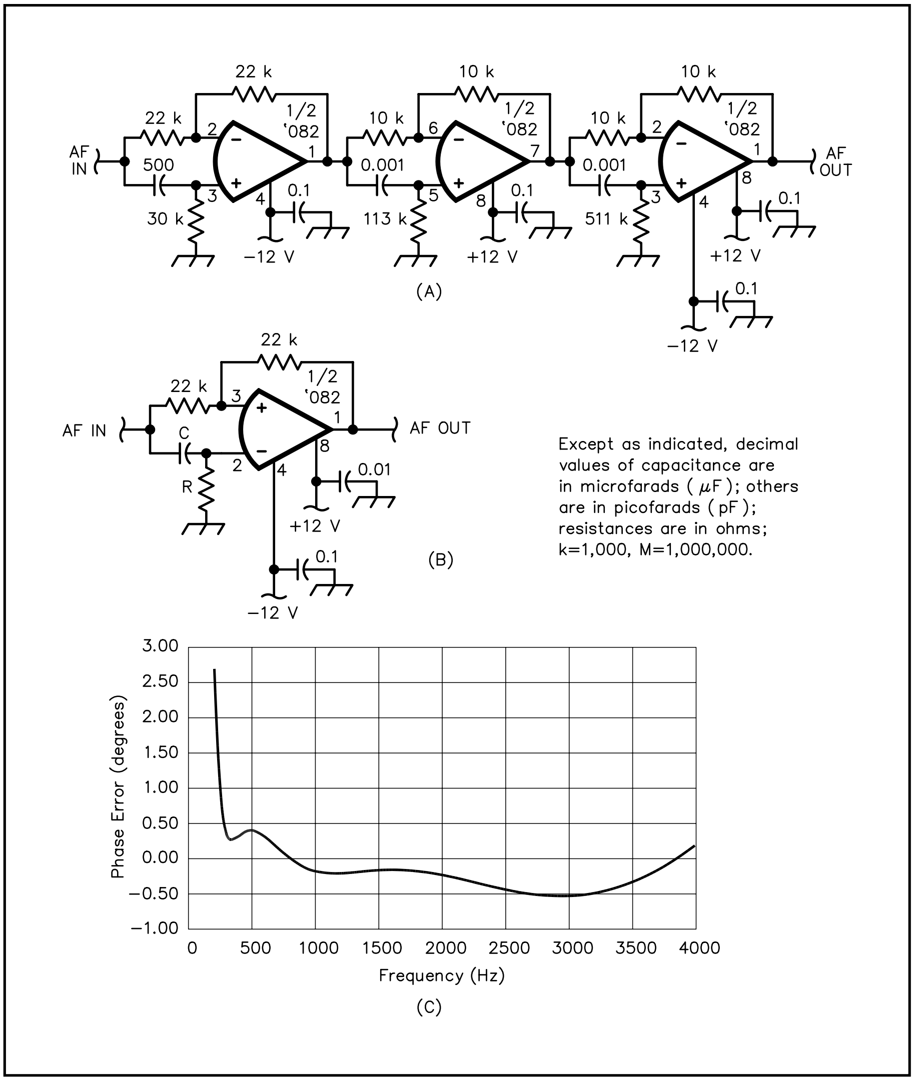


Fig 17.25—A: an example of an audio phase-shift circuit. B: a single all-pass stage. C, phase error vs audio frequency for a pair of circuits like that in A: with appropriate values of R and C.

common. Good, solid construction is necessary. Another problem involves leakage of the LO into the RF signal path by conduction and/or radiation. The random fluctuations in phase of the leakage signal interact with the LO to produce some unpleasant modulation and microphonic effects. Hum in the audio can be caused by interactions between the LO and the power supply; good bypassing and lead filtering of the power supply are needed. A small amount of RF amplification is beneficial for all of these problems.

The Superheterodyne Receiver

GENERAL DISCUSSION

The superheterodyne ("superhet") method is by far the most widely used approach to receiver and transmitter design. **Fig 17.26A** shows the basic elements as applied to an SSB/CW receiver, which we will consider first. We will consider a superhet transmitter later in this chapter.

RF from the antenna is filtered (preselected) by a band-pass filter of some kind to reduce certain kinds of spurious responses and then (possibly) RF amplified. A mixer, or frequency converter circuit, *multiplies* (in the time domain) its two inputs, the signal and the LO. The result of this multiplication process is a pair of output intermediate frequencies (IFs) that are the sum and difference of the signal and LO frequencies. If the mixer is"linear," it is a perfect multiplier, these are the only output frequencies present, and it has all of the properties of any other linear circuit except for the change of frequency. Fig 17.26B includes the mathematics that verify this linear behavior.

One of these outputs is selected to be the "desired" IF by the designer. It is then band-pass filtered and amplified. The bandwidths and shape factors of these filters are optimized for the kind of signal being received (AM, SSB, CW, FM, digital data). Two of the main attributes of the superhet are that this signal filtering band shape and also the IF amplification are constant for any value of the receive signal frequency. An excessively narrow preselector filter could, however, have some effect on the desired signal, as we saw in the case of the TRF receiver.

A second mixer, or "detector" as it is usually called, translates the IF signal to baseband (audio) where it is further amplified, possibly filtered, and applied to an output transducer (headphones, loudspeaker, some other signal processor or display).

A superhet receiver may also contain multiple frequency conversions (IFs). Later discussion will focus on strategies used to select these IFs.

Let's begin with a detailed discussion of the classic down-conversion superhet. Almost all of the topics apply as well to the various other kinds of receiver designs in subsequent sections.

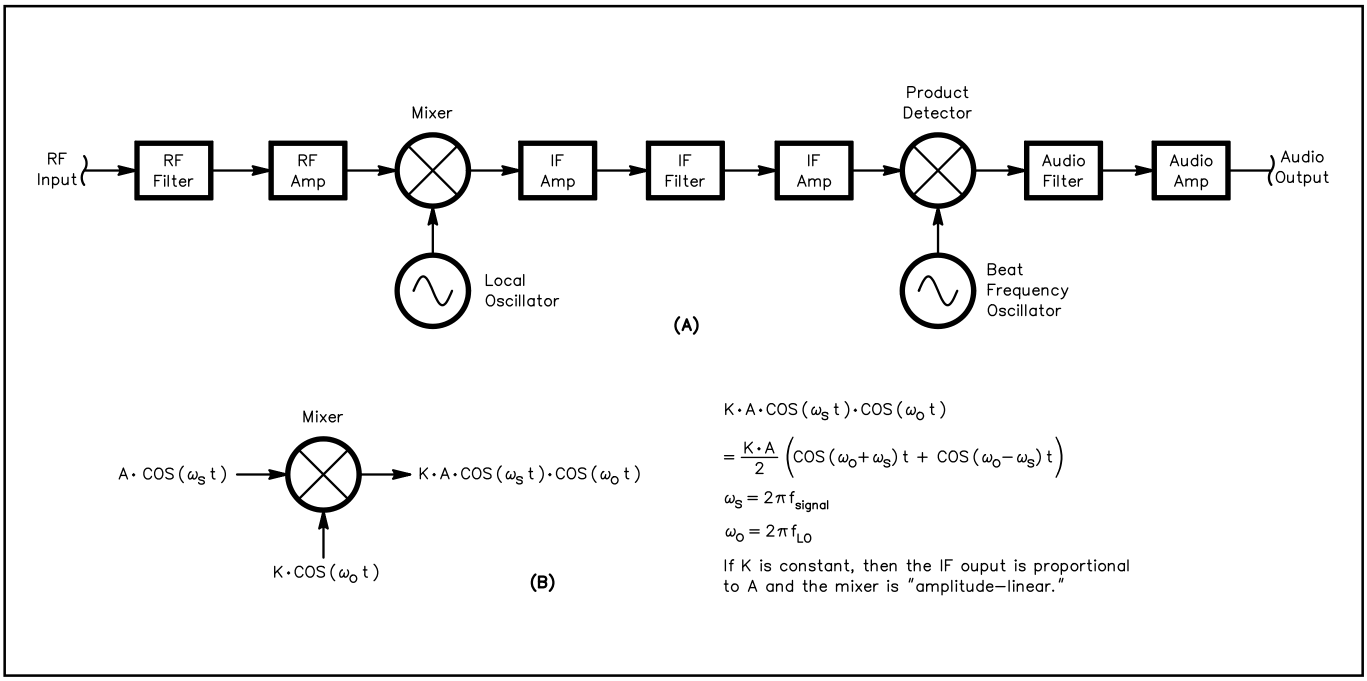


Fig 17.26—(A) Basic block diagram of a superhet receiver. (B) Showing how the input signal and a constant LO input produce a linear mixing action.

Superhet Characteristics: A Down-Conversion Example

The desirability of the superhet approach is offset somewhat by certain penalties and problem areas. As a vehicle for mentioning these difficulties, seeing how to deal with them and discussing analysis and design methods; we use the tutorial example in **Fig 17.27.** That is a "down converting" single-conversion 14-MHz superhet with a 1.5-MHz IF. This receiver is simple and capable of fairly good performance in the 1.8 to 30-MHz frequency range. Fig 17.27 is intentionally incomplete and meant for instructional purposes only; do not attempt to duplicate it as a project.

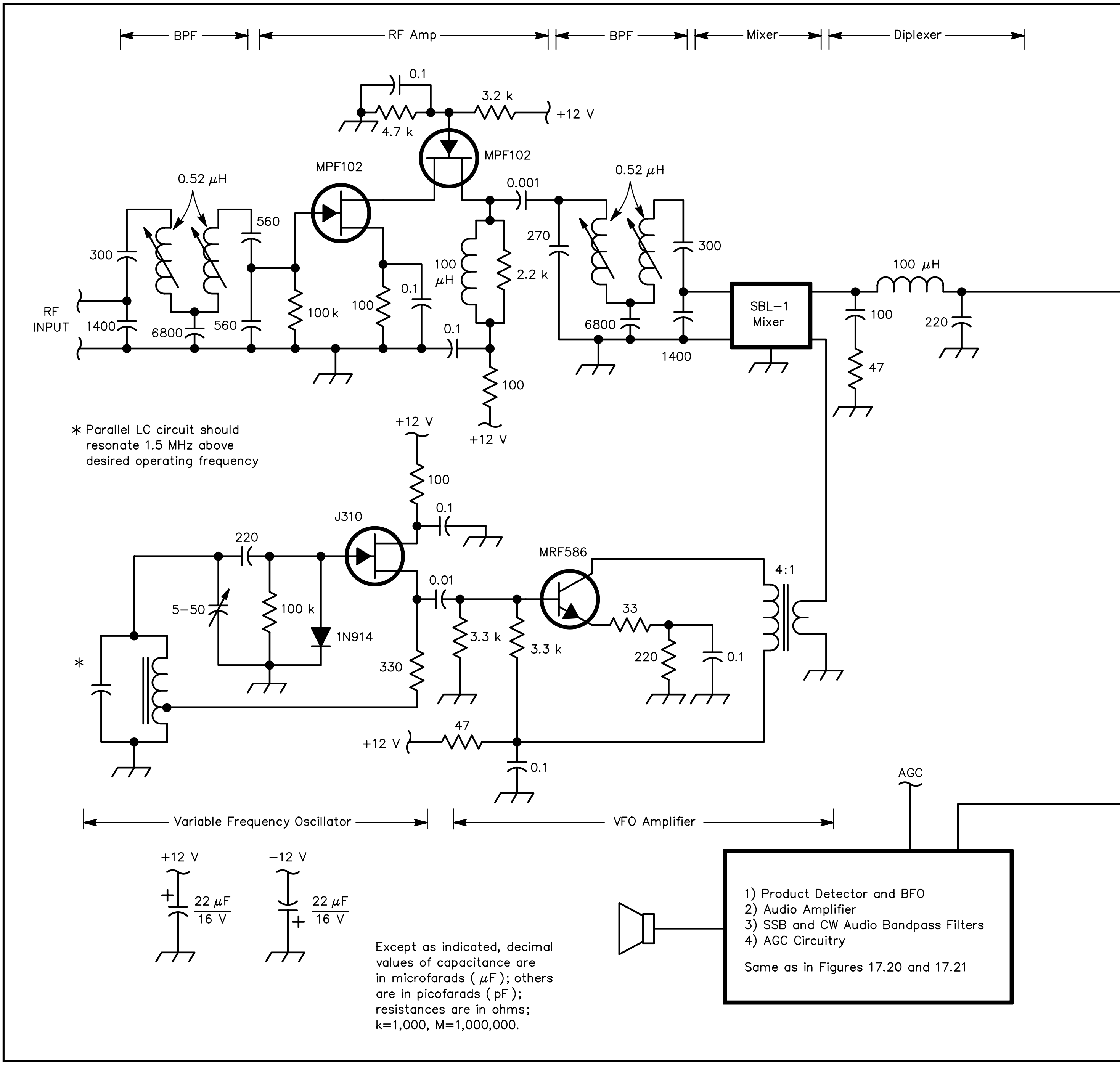
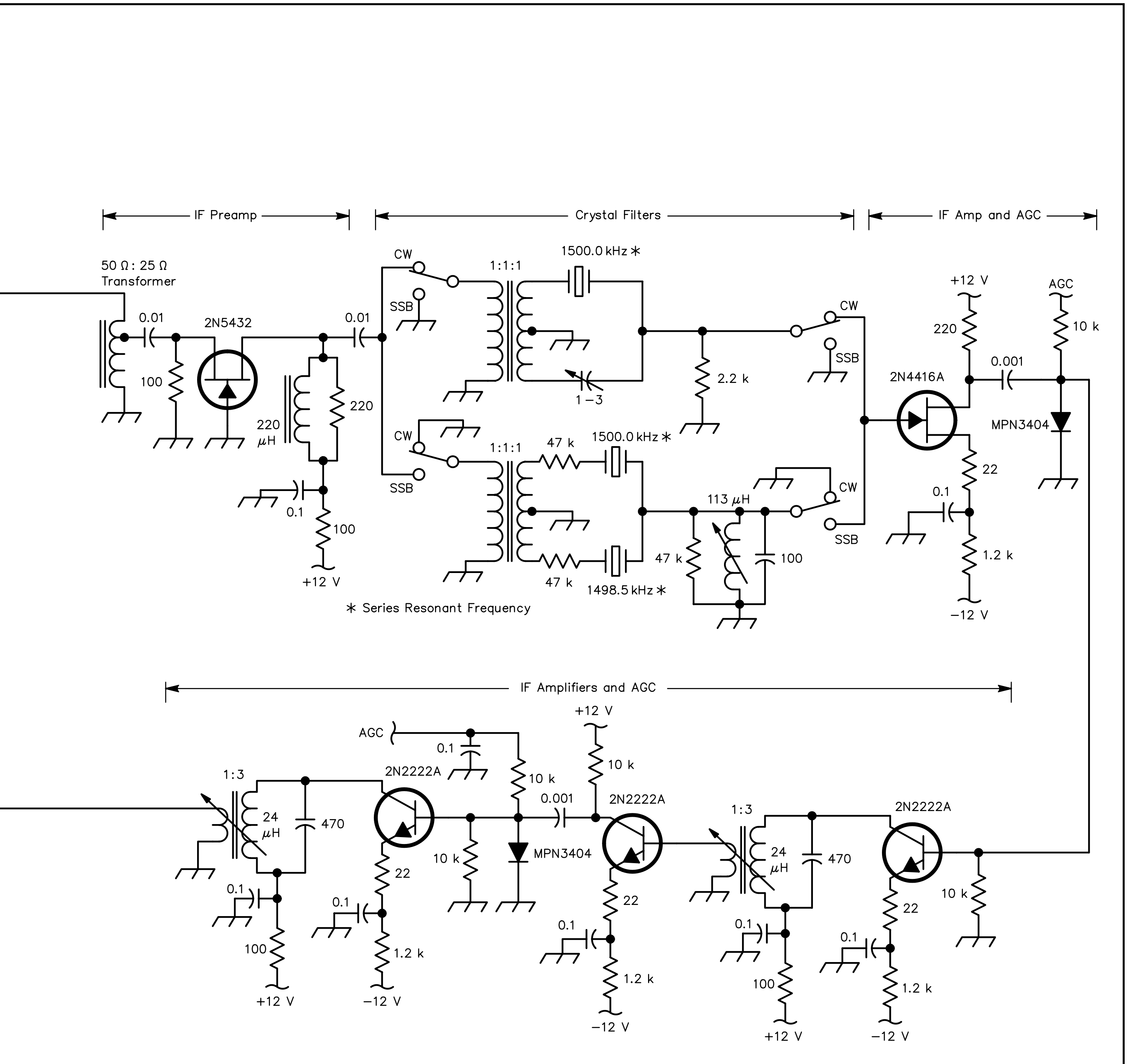


Fig 17.27—Specific example of a down-conversion superhet that is used to explain and analyze superhet behavior and design.

Block Diagram

The block labels of Fig 17.27 show that a preselector and RF amplifier are followed by downward frequency conversion to 1.5 MHz. This is followed by IF amplification and crystal filtering, a product detector, audio band-pass filters and an audio power output stage. Equal emphasis is given to SSB and CW. AGC circuitry is included. The audio and AGC circuits are the same as those in Fig 17.20 and Fig 17.21. As a first step, let's look at spurious responses of the mixer.



Mixer Spurious Responses

Mixers and their spurious responses are covered in detail in the **Mixers** chapter, but we will present a brief overview of the subject for our present purposes. We then will see how this information is used in the design process.

The mixer is vulnerable to RF signals other than the desired signal. Various harmonics of any undesired RF signal and harmonics of the LO combine to produce spurious IF outputs (called harmonic IMD). If these spurious outputs are within the IF passband they appear at the receiver output. The strength of these outputs depends on: harmonic number and strength of the RF signal as it appears at the mixer input, the harmonic number of the LO, the LO power rating (7 dBm, 17 dBm, and so on) and the design of the mixer.

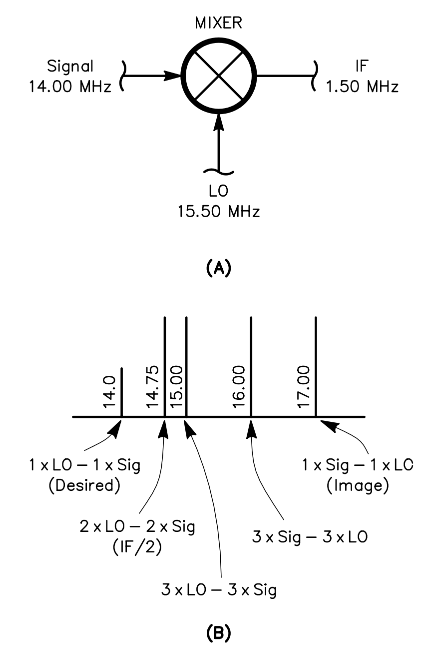
Commercially available double-balanced diode mixers are so convenient, easy to use and of such low cost and high quality that they are used in many Amateur Radio receiver and transmitter projects. These mixers also do a good job of rejecting certain kinds of spurious responses. Our numerical examples will be based on typical published data for one of these mixers (Ref 40).

Fig 17.28A shows an example for the mixer tuned to a desired signal at 14.00 MHz with the LO at 15.50 MHz. The locations of undesired signals that cause a spurious response are shown in Fig 17.28B; they are at 14.75, 15.00, 16.00 and 17.00 MHz (there are many others of lesser importance). Each of these undesired signals produces a 1.50-MHz output from the mixer. The figure indicates the harmonics of the undesired signal and the harmonics of the LO that are involved in each instance. The "order" of the spurious product is the sum of these harmonic numbers, for example the one at 16.00 MHz is a sixth-order product. The spurious at 17.00 MHz is called the "image" because it is also 1.50 MHz away from the LO, just as the 14.00 desired signal is 1.5 MHz away from the LO. It is a second-order response, as is also the response at 14.00 MHz.

Fig 17.28C is a chart that shows the relative responses for various orders of harmonic IMD products for a signal level (desired or undesired) of 0 dBm and an LO level of +7 dBm. The values are typical for a great many +7-dBm mixers having various brand names and they improve greatly for higher level mixers (at the same RF levels). The second-order (desired and image) both have a reference value of 0 dB and the others are in dB below those two.

We can now consider the receiver design that suppresses these spurious responses so that they do not interfere with a weak desired signal at 14.00 MHz. If an interfering signal is reduced in amplitude at the mixer RF input by 1.0 dB, the suppression of that spur is improved by 1.0 × Signal Harmonic Number dB. This is true in principle, but in reality the reduction may be somewhat less. For example, the spur produced by 15.00 MHz is reduced 3 dB for each dB that we reduce its level.

We accomplish this task by choosing the right mixer, limiting the amount of RF amplification and designing adequate selectiv-



RF Level = 0 dBm

RF Harmonic					
1	0	39	12	47	22
2	55	71	60	69	53
3	33	49	35	59	36
4	70	66	58	75	62
5	46	62	50	59	60
	1	2	3	4	5

LO Harmonic

(C)

Fig 17.28—A: mixer at 14.00 MHz with LO at 15.50 and IF at 1.50. B: locations of strong signals that interfere with desired signal at 14.00 MHz due to harmonic IMD. C: typical chart of harmonic IMD products for a double balanced diode mixer with 0 dBm signal and 7 dBm LO.

ity into the preselector circuitry. With respect to selectivity, though, note that in many other mixing schemes the interfering signal is so close to the desired frequency that selectivity does little good. Then we must use a mixer with a higher LO level and/or reduce RF gain.

The design method is illustrated by the following numerical example. Suppose that a signal at 14.75 MHz (the IF/2 spur) is at –20 dBm (very strong) at the antenna and –10 dBm at the mixer RF port. From the chart, this spur will be reduced by 71 + 2 (0 – (–10)) = 91 dB to a level of –10 – 91 = –101 dBm. If this is not enough then a preselector will help. If the preselector attenuates 14.75 MHz by 5 dB, the total spur reduction will be 71 + 2(0 – (–15)) = 101 dB to a level of –15 – 101 = –116 dBm, a 15-dB improvement. Notice that spurs involving high harmonics of the signal frequency attenuate more quickly as the input RF level is reduced.

On the other hand, if we consider the image signal at 17 MHz, all of the reduction of this spur must come from the preselector. In other words, selectivity is the only way to reduce the image response unless an image reducing mixer circuit is used (Ref 41). In this example, additional spur reduction is obtained by using a preselector circuit topology that has improved attenuation *above* the passband.

In designing *any* receiver we must be reasonable about spur and image reduction. Receiver cost and complexity can increase dramatically if we are not willing to accept an occasional spurious response due to some very strong and seldom occurring signal. In the case of a certain persistent interference some specific cure for that source can usually be devised. A sharply tuned "trap" circuit, a special preselector or a temporary antenna attenuator are a few examples. In practice, for down-conversion superhets, 90 dB of image reduction is excellent and 80 dB is usually plenty good enough for amateur work.

In classical down-conversion superhets, the preselection circuits are tuned and bandswitched in unison with the LO. They must all "track" each other across the dial. The cost and complexity of this arrangement have made this approach prohibitive in modern commercial multiband designs (Ref 42). For amateur work the approach in Fig 17.27 is more practical, using switched or even plug-in band-pass preselectors and oscillator coils. A frequency counter, offset by the 1.5 MHz IF and connected to the LO, eliminates the need for a calibrated dial.

Two-Tone Intermodulation Distortion

Another important mixer spurious response is two-tone IMD. This distortion has been covered previously in this chapter, and the **Mixers** chapter gives more detail. From a system design standpoint, the trade-offs between receiver noise figure and IMD have been covered in this chapter, and the choices of mixer, RF gain (if any) and selectivity are decided in a study exercise of performance, cost and complexity. A receiver that has a 10 to 20-dBm third-order intercept point for two signals 20 kHz and 40 kHz removed is an excellent receiver in many applications. Some advanced experimenters have built receivers with 25 to 40-dBm values of IP3. Values of 40 dBm are near the state of the art (Refs 43, 44, 45).

A matter of considerable interest concerns the way that IMD varies as the separation between the two tones increases. In Fig 17.27, for example, if one tone is 1.0 kHz (or 100 kHz) above 14.00 the other is 2.0 kHz (or 200 kHz) above. We see that for very close tone separations the IF filter may not prevent the tones from reaching the circuits following the IF filter. As the separation increases, first one, then both, tones fall outside the IF filter passband and the IMD becomes much less. However the mixer and the amplifier after the mixer are still vulnerable. At greater separations the preselector starts to protect these two stages, but the RF amplifier is not well protected by the first RF filter until the tone separation becomes greater, perhaps 200 kHz. It is a common procedure to plot a graph of receiver third-order input intercept point vs tone separation and then look for ways to improve the overall performance.

The stages after the IF filter are protected by AGC so that, hopefully, tones in the IF passband do not overdrive the circuits after the IF filter. But in the example of Fig 17.27 there is also a narrowband audio filter and the AGC is derived from the output of this filter. This means that circuits *after* the IF filter but *ahead of* the audio filter may not always be as well protected as we would like. Strong tones that get through the IF filter may be stopped by the audio filter and not affect the AGC. This particular example illustrates a very common problem in all kinds of receivers that have *distributed* selectivity. It is also found universally in multiple conversion receivers, as we will discuss later.

Gain and Noise Figure Distribution

Based on the information given so far, the approach to designing a superhet receiver, whether a downconverter or any other kind, can now be summarized by the following guidelines:

1. Try to keep the gain ahead of the mixer and the narrow band-pass filters (SSB, CW and so on) as low as possible. For a fixed components cost (such as mixers and amplifiers), this minimizes the IMD, both two tone and harmonic.
2. Reducing the gain implies that the noise figure may be a little higher. It is always best to avoid making the noise figure any lower than necessary. Noise figure is usually more important at microwave frequencies than at HF, and strong signal interference is usually less important. Where interference is a problem an increase in noise figure is almost always mandatory, except possibly when a higher-level mixer is used. A narrowband preselector, for example, will increase the noise figure (and also the intercept point) because of its passband attenuation.
3. Amplifier circuits and modules always involve a trade-off of some kind between intercept point and noise figure. Designers look for devices and circuits that optimize the SFDR for the particular kind of receiver under design.
4. If the receiver has distributed selectivity, make the first IF filter good enough that the AGC/IF-overload problem mentioned above is minimized.
5. To *minimize* the gain ahead of the mixer, *follow* the mixer with a low-noise, high-dynamic-range amplifier with no more gain than necessary, say 10 dB or so (see Fig 17.27).
6. Terminate the mixer in such a way that its IMD is minimized. Fig 17.27 shows a simple IF diplexer that absorbs the output image at 29.5 MHz (14.0 + 15.5).
7. The RF terminal of the mixer should be short circuited at the image frequency so that noise at the image frequency (from the preceding circuitry) is minimized.
8. Because a large amount of overall gain is needed, reducing front-end gain implies that the gain after the first IF filter must be very large. The problem of IF and audio noise then arises. It is very desirable to use a low-noise amplifier right after the first IF filter (see Fig 17.27) and to restrict the bandwidth of the IF/AF amplifiers. A second IF/AF filter downstream, and also possibly an image-reducing product detector, are excellent ways to accomplish this. This step also minimizes the degradation of receiver noise figure that can be caused by this wideband noise.
9. The LO must have very low phase noise to reduce reciprocal mixing. Also, the mixer must have good balance (meaning isolation or rejection) from LO port to RF and IF ports so that broadband additive noise from the LO amplifiers does not degrade the mixer noise figure. This is especially important when the RF amplifier gain has been minimized. If the mixer is not balanced in this sense at the LO port, a band-pass filter between LO and mixer is very desirable (Refs 46, 47).

AUTOMATIC GAIN CONTROL (AGC)

The amplitude of the desired signal at each point in the receiver is controlled by AGC. Each stage has a distortion vs signal-level characteristic that must be known, and the stage input level must not become excessive. The signal being received has a certain signal-to-distortion ratio that must not be degraded too much by the receiver. For example, if an SSB signal has –30 dB distortion products the receiver should have –40 dB quality. A correct AGC design ensures that each stage gets the right input level. It is often necessary to redesign some stages in order to accomplish this (Ref 48).

The AGC Loop

Fig 17.29A shows a typical AGC loop that is often used in amateur receivers. The AGC is applied to the stages through RF decoupling circuits that prevent the stages from interacting with each other. The AGC amplifier helps to provide enough AGC loop gain so that the gain-control characteristic of Fig 17.29B is achieved. The AGC action does not begin until a certain level, called the AGC threshold, is reached. The Threshold Volts input in Fig 17.29A serves this purpose. After that level is exceeded, the audio level slowly increases. The audio rise beyond the threshold value is usually in the 5 to 10-dB range. Too much or too little audio rise are both undesirable for most operators.

As an option, the AGC to the RF amplifier is held off, or "delayed," by the 0.6-V forward drop of the diode so that the RF gain does not start to decrease until larger signals appear. This prevents a premature increase of the receiver noise figure. Also, a time constant of one or two seconds after this diode helps keep the RF gain steady for the short term.

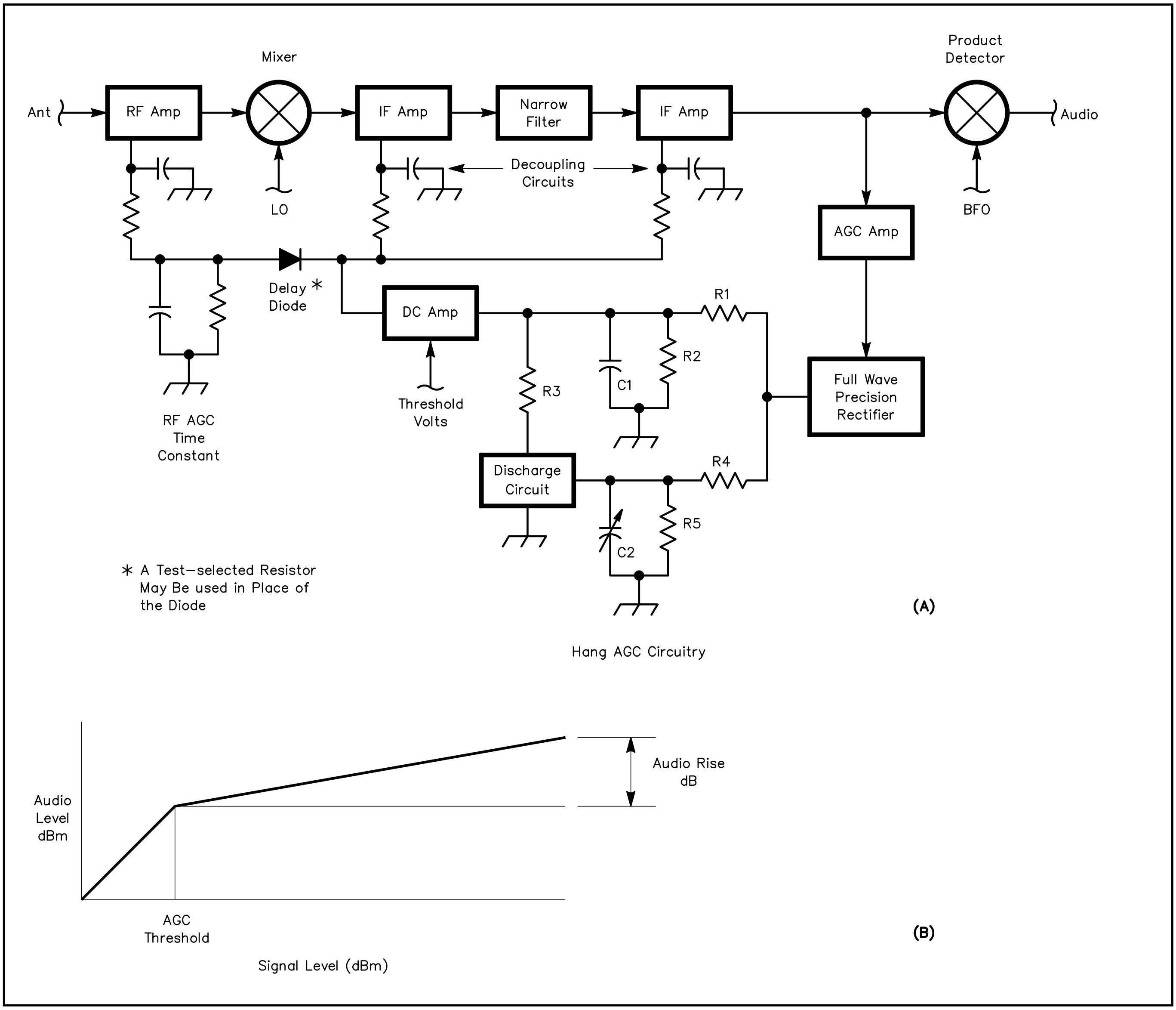


Fig 17.29—AGC principles. A: typical superhet receiver with AGC applied to multiple stages of RF and IF. B: audio output as a function of antenna signal level.

Fig 17.30 is a typical plot of the signal levels at the various stages of a certain ham band receiver. Each stage has the proper level and a 115-dB change in input level produces a 10-dB change in audio level. A manual gain control would produce the same effect.

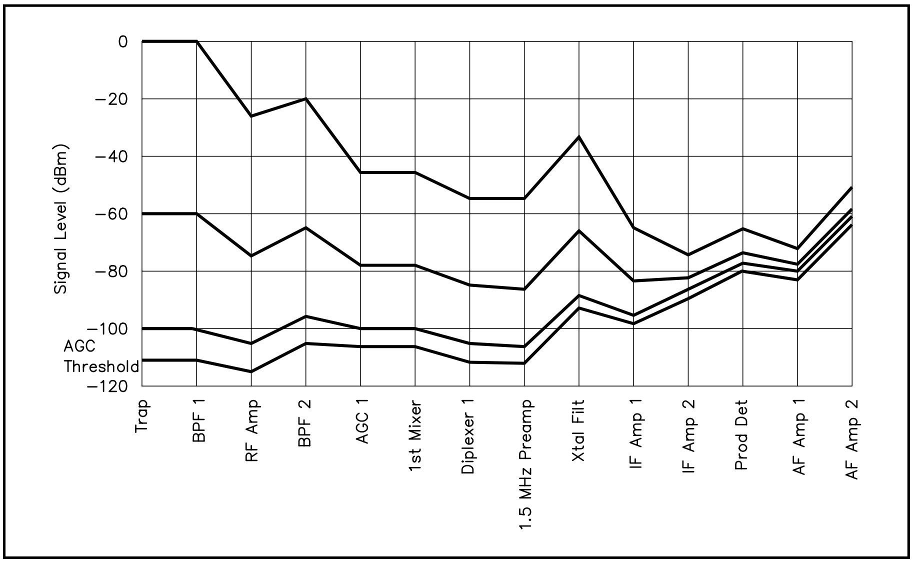


Fig 17.30—Gain control of a ham-band receiver using AGC. A manual gain control could produce the same result.

AGC Time Constants

In Fig 17.29, following the precision rectifier, R1 and C1 set an "attack" time, to prevent excessively fast application of AGC. One or two milliseconds is a good value for the R1 × C1 product. If the antenna signal suddenly disappears, the AGC loop is opened because the precision rectifier stops conducting. C1 then discharges through R2 and the C1 × R2 product can be in the range of 100 to 200 ms. At some point the rectifier again becomes active, and the loop is closed again.

An optional modification of this behavior is the "hang AGC" circuit (Ref 49). If we make R2 × C1 much longer, say 3 seconds or more, the AGC voltage remains almost constant until the R5C2 circuit decays with a switch selectable time constant of 100 to 1000 ms. At that time R3 quickly discharges C1 and full receiver gain is quickly restored. This type of control is appreciated by many operators because of the lack of AGC "pumping" due to modulation, rapid fading and other sudden signal level changes.

AGC Loop Problems

If the various stages have the property that each 1-V change in AGC voltage changes the gain by a constant amount (in dB), the AGC loop is said to be "log linear" and regular feedback principles can be used to analyze and design the loop. But there are some difficulties that complicate this textbook model. One has already been mentioned, that when the signal is rapidly decreasing the loop becomes "open loop" and the various capacitors discharge in an open-loop manner. When the signal is increasing beyond the threshold, or if it is decreasing slowly enough, the feedback theory applies more accurately. In SSB and CW receivers rapid changes are the rule and not the exception.

Another problem involves the narrow band-pass IF filter. The group delay of this filter constitutes a time lag in the loop that can make loop stabilization difficult. Moreover, these filters nearly always have much greater group delay at the edges of the passband, so that loop problems are aggravated at these frequencies. Overshoots and undershoots, called "gulping," are very common. Compensation networks that advance the phase of the feedback help to offset these group delays. The design problem arises because some of the AGC is applied before the filter and some after the filter. It is a good idea to put as much fast AGC as possible after the filter and use a slower decaying AGC ahead of the filter. The delay diode and RC in Fig 17.29A are helpful in that respect. Complex AGC designs using two or more compensated loops are also in the literature. If a second cascaded narrow filter is used in the IF it is usually a lot easier to leave the second or "downstream" filter out of the AGC loop.

Another problem is that the control characteristic is often not log-linear. For example, dual-gate MOSFETs tend to have much larger dB/V at large values of gain reduction. Many IC amplifiers have

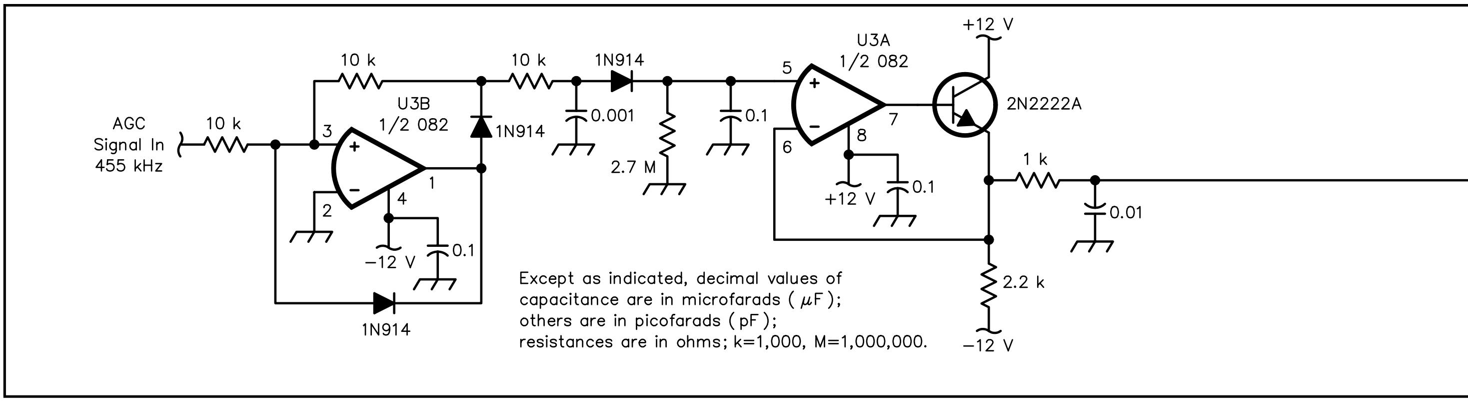


Fig 17.31—Some gain controllable amplifiers and a rectifier suitable for audio derived AGC.

the same problem. The result is that large signals cause instability because of excessive loop gain. There are variable gain op amps and other ICs available that are intended for gain control loops.

Audio frequency components on the AGC bus can cause problems because the amplifier gains are modulated by the audio and distort the desired signal. A hang AGC circuit can reduce or eliminate this problem.

Finally, if we try to reduce the audio rise to a very low value, the required loop gain becomes very large, and stability problems become very difficult. It is much better to accept a 5 to 10 dB variation of audio output.

Because many parameters are involved and many of them are not strictly log-linear, it is best to achieve good AGC performance through an initial design effort and finalize the design experimentally. Use a signal generator, attenuator and a signal pulser (2-ms rise and fall times, adjustable pulse rate and duration) at the antenna and a synchronized oscilloscope to look at the IF envelope. Tweak the time constants and AGC distribution by means of resistor and capacitor decade boxes. Be sure to test throughout the passband of each filter. The final result should be a smooth and pleasant sounding SSB/CW response, even with maximum RF gain and strong signals. Patience and experience are helpful.

Audio-Derived AGC

The example in Fig 17.20 shows audio-derived AGC. There is a problem with this approach also. At low audio frequencies the AGC can be slow to develop. That is, low-frequency audio sine waves take a long time to reach their peaks. During this time the RF/IF/AF stages can be overdriven. If the RF and IF gains are kept at a low level this problem can be reduced. Also, attenuating low audio frequencies prior to the first audio amplifier should help. With audio AGC, it is important to avoid so-called "charge pump" rectifiers or other slow-responding circuits that require multiple cycles to pump up the AGC voltage. Instead, use a peak-detecting circuit that responds accurately on the first positive or negative transition.

AGC Circuits

Fig 17.31 shows some gain controllable circuits and also a suitable rectifier for audio derived AGC.

Fig 17.31A shows a precision half-wave rectifier and a two-stage 455-kHz IF amplifier with PIN-diode gain control. This circuit is a simplified adaptation from a production receiver, the Collins 651S. The IF amplifier section shown is preceded and followed by selectivity circuits and additional gain stages with AGC. The 1.0-μF capacitors aid in loop compensation. The favorable thing about this approach is that the transistors remain biased at their optimum operating point. Right at the point where the diodes start to conduct, a small increase in IMD may be noticed, but that goes away as diode current increases slightly. Two or more diodes can be used in series, if this is a problem (it very seldom is).

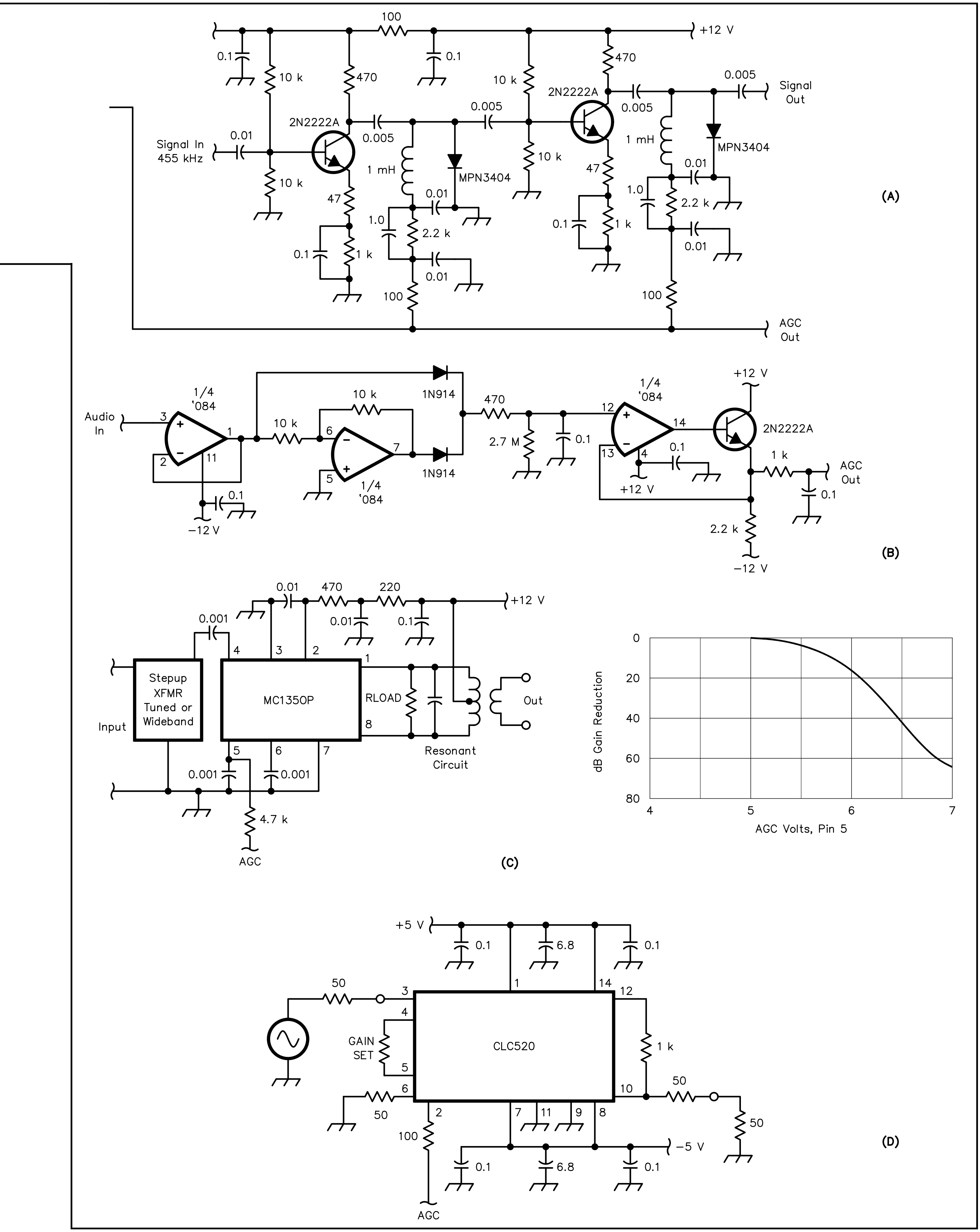

+12 V
100
0.1
10 k
470
2N2222A
Signal In
455 kHz
0.01
0.005
1 mH
MPN3404
47
1.0
2.2 k
1 k
Signal Out
AGC Out
(A)
Audio In
1/4 '084
10 k
1N914
470
2.7 M
2N2222A
AGC Out
+12 V
−12 V
(B)
Stepup XFMR Tuned or Wideband
Input
MC1350P
RLOAD
Resonant Circuit
Out
4.7 k
AGC
0.001
220
dB Gain Reduction
AGC Volts, Pin 5
(C)
+5 V
6.8
CLC520
GAIN SET
50
−5 V
AGC
(D)

Fig 17.31B is an audio-derived AGC circuit using a full-wave rectifier that responds to positive or negative excursions of the audio signal. The RC circuit follows the audio closely.

Fig 17.31C shows a typical circuit for the MC1350P RF/IF amplifier. The graph of gain control vs AGC volts shows the change in dB/V. If the control is limited to the first 20 dB of gain reduction this chip should be favorable for good AGC transient response and good IMD performance. Use multiple low-gain stages rather than a single high-gain stage for these reasons. The gain control within the MC1350P is accomplished by diverting signal current from the first amplifier stage into a “current sink.” This is also known as the “Gilbert multiplier” architecture. Another chip of this type is the NE/SA5209. This type of approach is simpler to implement than discrete-circuit approaches, such as dual-gate MOSFETs that are now being replaced by IC designs.

Fig 17.31D shows the high-end performance Comlinear CLC520AJP (14-pin DIP plastic package) voltage controlled amplifier. It is specially designed for accurate log-linear AGC from 0 to 40 dB with respect to a preset maximum voltage gain from 6 to 40 dB. Its frequency range is dc to 150 MHz. It costs about $11.50 in small quantities and is an excellent IF amplifier for high-performance receiver or transmitter projects.

IF FILTERS

There are some aspects of IF-filter design that influence the system design of receivers and transmitters. The influence of group delay, especially at the band-pass edges, on AGC-loop performance has been mentioned. Shape factor is also significant (the ratio of two bandwidths, usually 60-dB:6-dB widths). To get good adjacent-channel rejection, the transition-band response should fall very quickly. Unfortunately, this goal aggravates group-delay problems at the passband edges. It also causes poor transient response, especially in CW filters. Another filter phenomenon can cause problems: at sharp passband edges signals and noise produce a raspy sound that is annoying and interferes with weak signals.

A desirable filter response would be slightly rounded at the edges of the passband, say to –6 dB, with a steep rolloff after that. This is known as a “transitional filter” (Ref 50). Cascaded selectivity with two filters, each having fewer “poles” (than a single filter would) is also a good approach. Both methods have a smoother group delay across the passband and reduce the problems mentioned above.

Ultimate Attenuation

In a high-gain receiver with as much as 110 dB of AGC the ultimate attenuation of the filter is important. Low-level leakage through or around the filter produces high-pitch interference that is especially noticeable on CW. Give special attention to parts layout, wiring and shielding. (Filter selector switches are often leakage culprits.) Cascaded IF filters also help very considerably.

Audio Filter Supplement

An audio band-pass filter can be used to supplement IF filtering. This can help to improve signal-to-noise ratio and reduce adjacent-channel interference. Supplementary audio filtering also helps reduce the high-frequency leakage problem mentioned above. Another significant problem: If AGC is made in the IF section, strong signals inside the IF passband but outside the audio passband can “pump” or modulate the AGC, rendering weak desired signals hard to copy. This is especially noticeable during periods of high band activity, such as in a contest. These filters can use analog (see Fig 17.20 and Fig 17.21) or digital (DSP) technology.

Some Simple Crystal Filters

Fig 17.32 and **Fig 17.33** present two crystal filters to consider for a simple down-conversion receiver with a 1.5-MHz IF (see Fig 17.27). The crystals are a set of three available from JAN Crystals. (See the Address List in the **References** chapter for their current address. Mention the ARRL *Handbook*.) The filters are both driven from a low-impedance source (200 Ω, for example).

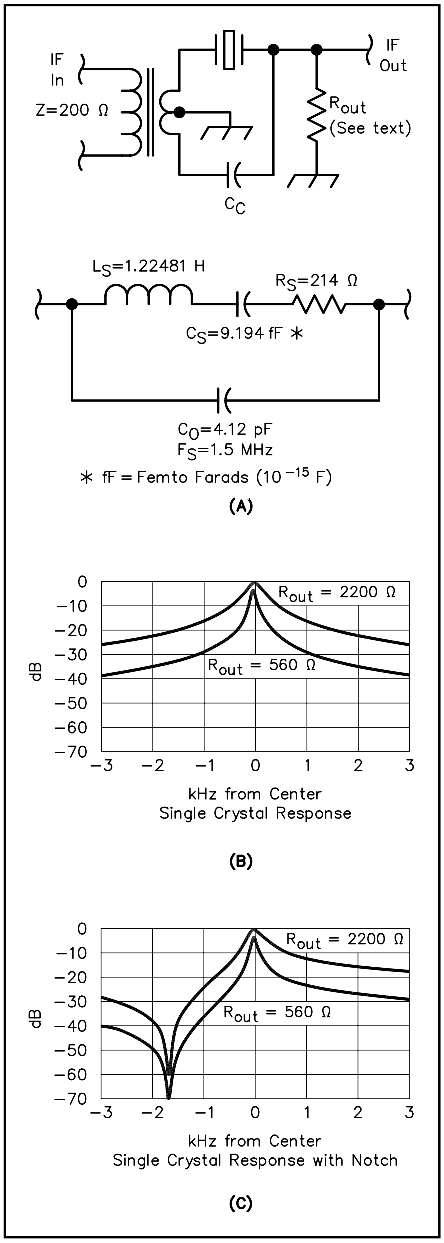


Fig 17.32—A single-crystal filter circuit for a simple CW receiver design. See also Fig 17.27.

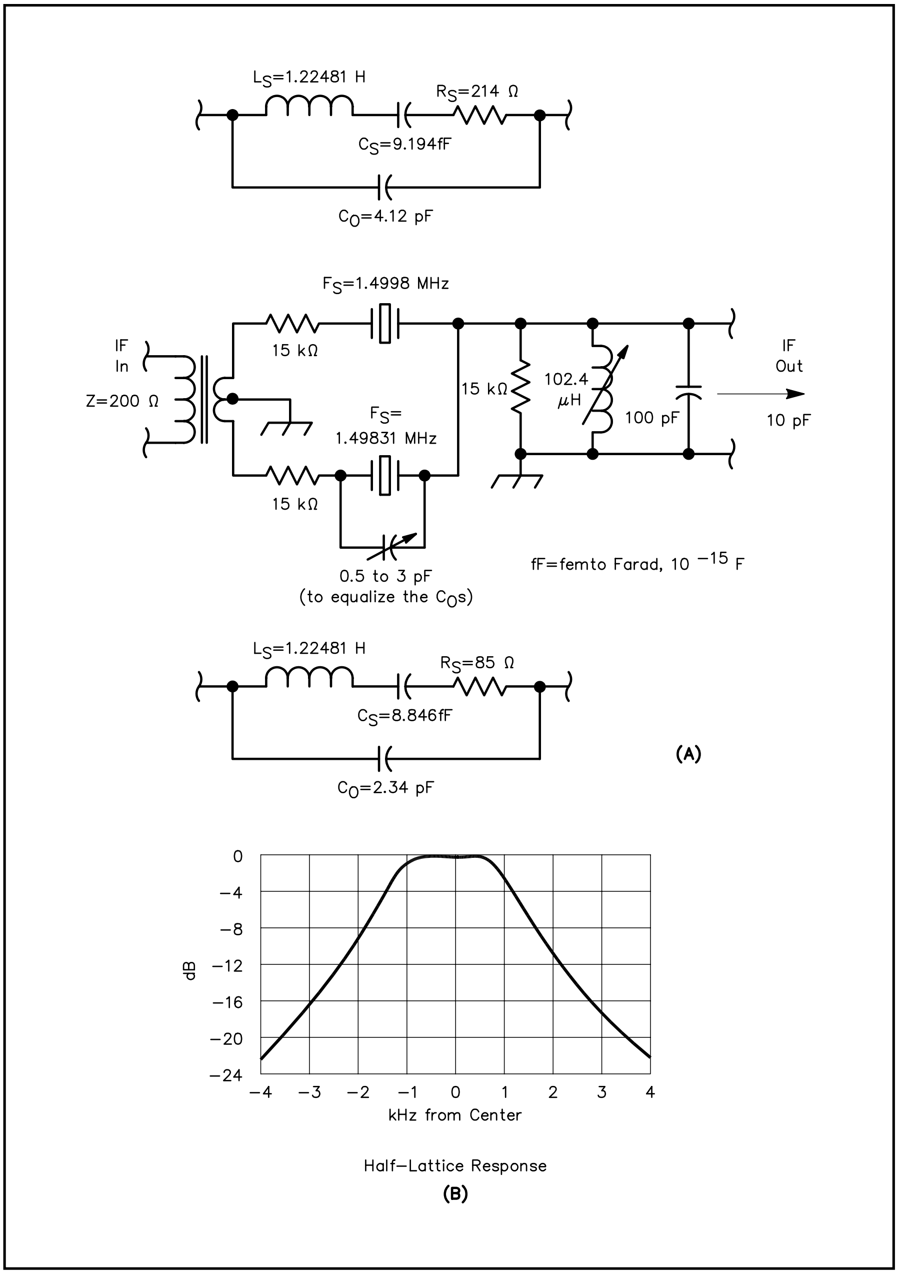


Fig 17.33—A two-crystal half-lattice filter for a simple SSB receiver. See also Fig 17.27.

CW Filter

Fig 17.32A is a "semilattice" filter using a single crystal for CW work (Ref 50). Capacitor C_c balances the bridge circuit at the crystal's parallel-resonant frequency because it is equal to the holder capacitance C_o of the crystal. The response is then symmetrical around the series resonant frequency of the crystal. The selectivity is determined by the value of R_{out}. As the value decreases the selectivity sharpens as shown in 32B. If this filter is combined with an audio band-pass filter as in Fig 17.21, pretty good CW selectivity is possible. In Fig 17.32C, the capacitor is increased to 8.3 pF and a notch appears at –1.7 kHz. This is the "single signal" adjustment. Also, note that the response on the high side is degraded quite a bit. The notch can be located above or below center frequency by adjusting the capacitor value; the degradation is on the opposite side of center.

The Sounds of Amateur Radio — **CW reception with a 2.4 kHz IF filter.**

The Sounds of Amateur Radio — **CW reception with a 2.0 kHz IF filter.**

The Sounds of Amateur Radio — **CW reception with a 500 Hz IF filter.**

The Sounds of Amateur Radio — **CW reception with a 250 Hz IF filter.**

SSB Filter

Fig 17.33 is a "half lattice" filter (Ref 50). The schematic diagram shows the LCR values and the series resonant frequencies of the two crystals. One of these (1.4998 MHz) is the same type as the one used in the CW filter. The trimmer capacitor equalizes the two values of C_o, the crystal shunt capacitance (very important) in case they are not already closely matched. Place the trimmer across the crystal that has the lowest value of C_o. The response curve shows good symmetry and modest adjacent-channel rejection. The output tuned circuit absorbs load capacitance to get a pure R_{load} (also important). The follow-up audio speech band-pass filter in Fig 17.20 will improve the overall response very considerably.

The Sounds of Amateur Radio — **SSB reception with a 2.4 kHz IF filter.**

The Sounds of Amateur Radio — **SSB reception with a 2.0 kHz IF filter.**

The Sounds of Amateur Radio — **SSB reception with a 500 Hz IF filter.**

The Sounds of Amateur Radio — **SSB reception with a 250 Hz IF filter.**

Mechanical Filters

Mechanical filters use transducers and the magnetostriction principle of certain materials to obtain a multiresonator narrow band-pass filter in the 100 to 500-kHz range. They are very frequency stable, accurate and reliable. An interesting example, for radio amateurs, is the Rockwell-Collins (Costa Mesa, California) "Low Cost Series" of miniaturized torsional-mode filters for 455.0 kHz. They come in three styles with 3 dB/60 dB bandwidths of 0.5/2.0, 2.5/5.2 and 5.5/11 kHz. In small quantities (1-4), they sell for $92 each (1997 price—mention the ARRL *Handbook* article), including tax and shipping. Used filters are sometimes available from various sources (Ref 42).

Multielement Crystal Filters

A discussion of more complex crystal filters appears in the **Filters and Projects** chapter of this *Handbook*. In this chapter we have considered only two very simple examples that might appeal especially to student designers and builders of a receiver that downconverts to an IF less than 2 MHz or so.

From the system design standpoint, note that for voice reception amateurs often use optional IF filters with less than the conventional bandwidth for SSB (for example 1.8 kHz), even though they reduce higher frequency speech components. This helps to improve adjacent-channel interference, which is a severe problem on some amateur bands.

It is common practice to use multipole crystal filters in the range from 5 to 10 MHz, because they can be economically designed for that frequency range. It is also common to cascade these filters with other types, such as mechanical or LC filters, at lower IFs (more about this later).

Filter Switching

Filter switching for different modes (AM, SSB, CW, RTTY and so on) requires some careful design to prevent impedance mismatching, leakage (discussed before) and spurious coupling to other circuitry. There are three general methods for switching: mechanical, relays, solid-state (diodes or transistors).

The latter two are needed in radios controlled by computers or logic circuits.

Mechanical switches require good isolation between the wafers and usually some kind of shielding arrangements. **Fig 17.34** shows examples of relay and diode switching that work quite well. The relays can be inexpensive miniature Radio Shack 275-241 SPDT units, one at each input and one at each output. The diodes can be inexpensive Motorola MPN3404 PIN diodes. These circuits assume that all filters are terminated with the same impedance values (Ref 42).

In PIN diode applications, IMD *can* be a problem with inadequate bias or excessive signal levels. The application (PIN diode and circuit) should be tested at the highest expected signal level.

One major problem involves high-level IF-output-signal leakage or BFO leakage into the input of the filter, which can produce high passband ripple and other unpleasant problems such as AGC malfunctions.

THE VLF IF RECEIVER

An approach to IF selectivity that has been used frequently over the years in both home built and factory made amateur receivers uses a second down conversion from an IF at, say 4 or 5 MHz or even 455 kHz, to a very low frequency, usually 50 to 85 kHz. At these frequencies several double tuned LC filters, separated by amplifier stages, make possible excellent improvements in SSB/CW frequency response and ultimate attenuation along with a relatively flat group delay. These amplifier stages can also have AGC. Low-cost (four-pole) crystal filters (SSB and CW) at the higher IF followed by two lower-IF channels (SSB and CW) make a very desirable combination. This is also an effective way to assure a narrow noise bandwidth

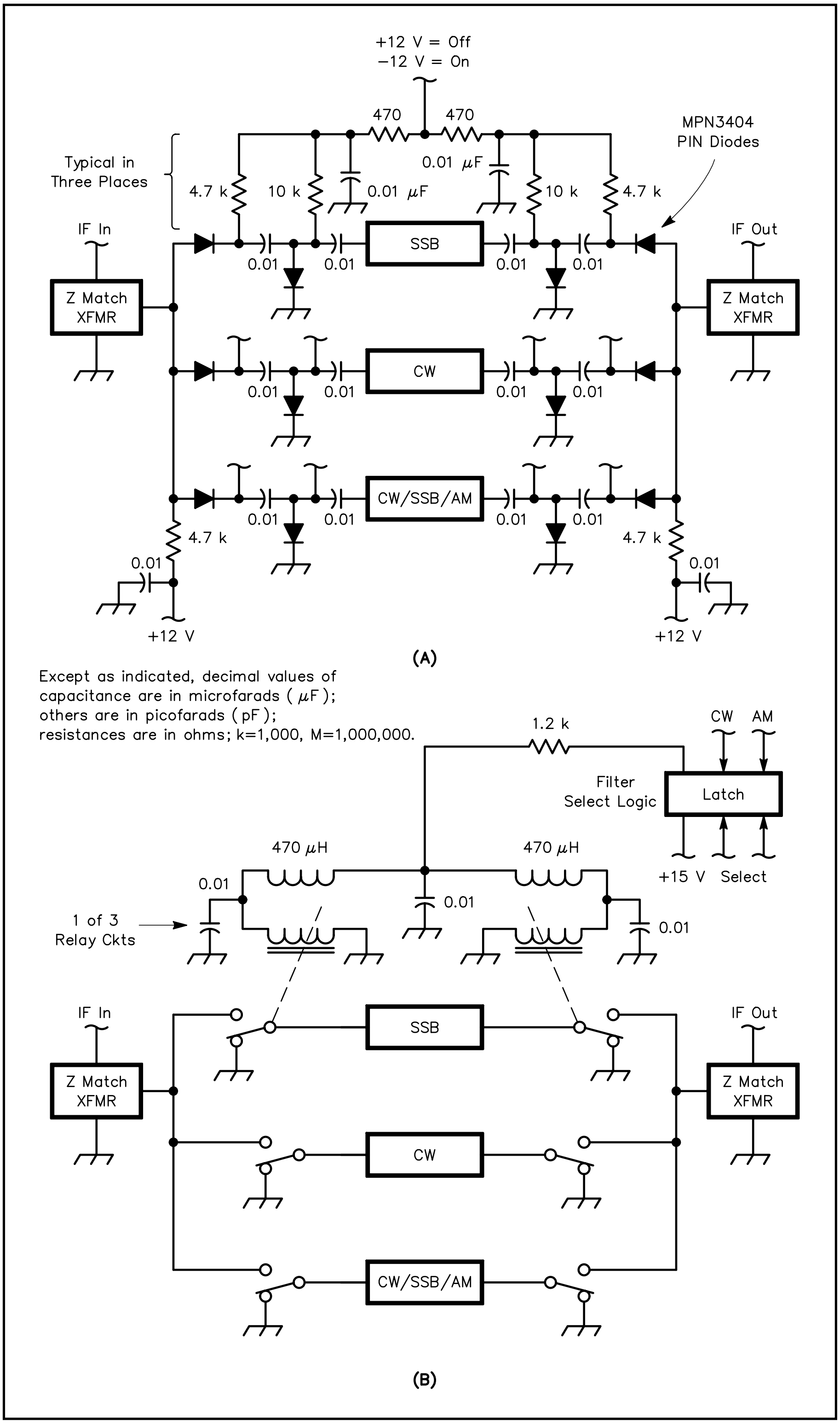


Fig 17.34—IF filter switching using relays or PIN diodes.

for the overall receiver. One requirement is that the circuitry ahead of the VLF downconverter must provide good rejection of an image frequency that is only 100 to 170 kHz away (Refs 41, 46, 51, 52, 53, 54, 55).

AM DEMODULATION

There is some interest among amateurs in double-sideband AM reception on the HF broadcast bands. Coherent AM detection is a way to reduce audio distortion that is caused by a temporary reduction of the carrier. This "selective fading" is due to phase cancellations caused by multipath propagation. By inserting a large, locally created carrier onto the signal this effect is reduced. The term "exalted carrier reception" is sometimes used. In reception of a double-sideband AM signal, the phase of the inserted carrier must be identical to that of the incoming carrier. If not, reduced audio and also audio distortion result. Therefore the common method is to use a phase-locked loop (PLL) to coordinate the phases of the incoming carrier and the locally generated carrier. This requires a Type II PLL, which drives, or integrates, the phase difference to zero degrees (Refs 41, 56, 57).

MULTIPLE CONVERSION SUPERHETS

There are a couple of drawbacks to the downconverting receiver just described. First, the LO must be bandswitched. Also, its tuning must track with the preselector tuning even though the preselector is offset from the signal frequency by the amount of the IF. A tuning dial scale is required for each band, and the receiver must be fitted to it at the factory. This adds a lot of cost and complexity.

A solution to these problems is shown in **Fig 17.35**. A crystal controlled first mixer is preceded by a gang-tuned preselector and is followed by a wideband first IF that is 200-kHz wide. The second mixer has a VFO that tunes a 200-kHz range. To change bands, the crystal is switched and the preselector is bandswitched. An additional tuned circuit removes the wideband additive noise from the crystal LO, so that it does not degrade the noise figure of the *unbalanced* mixer circuit.

One of the main design problems is to select the first IF, its bandwidth and the second mixer design so that harmonic IMD products (involving the signal, crystal frequency, first IF, second IF and VFO frequency) do not cause appreciable interference. In the example of Fig 17.35, a first IF at 2.9275 MHz (the signal frequency would be 14.2275 MHz) and a VFO at 2.7 MHz produce a fourth-order spurious response at 455 kHz, therefore the first IF filter must attenuate 2.9275 MHz sufficiently and the second mixer must reject the fourth-order response sufficiently. We have discussed the fourth-order (IF/2) response previously.

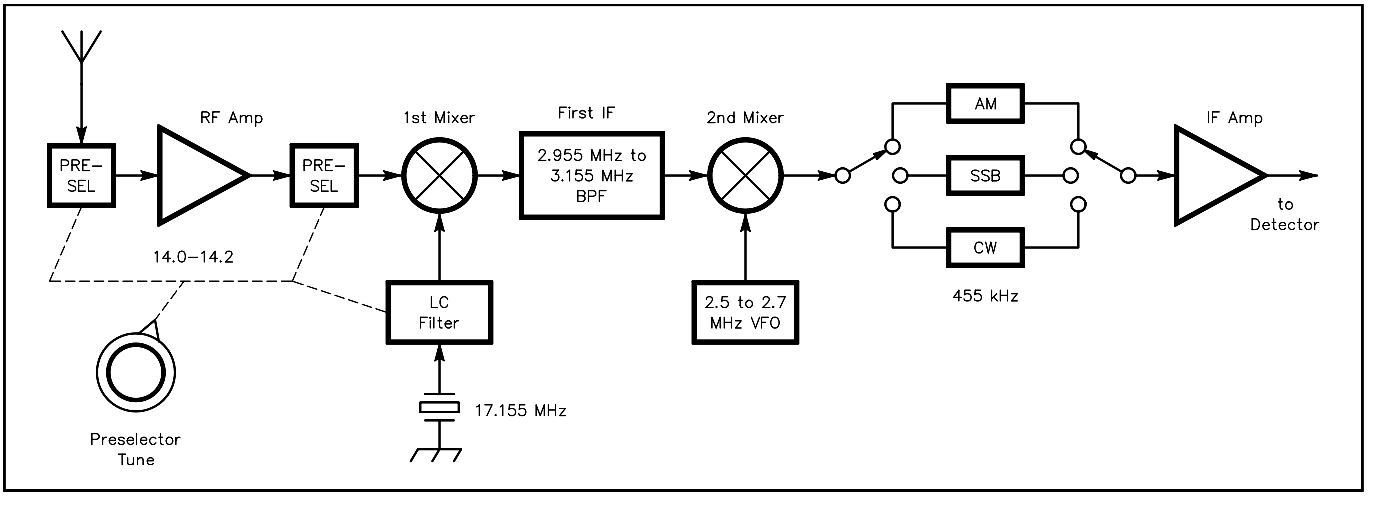


Fig 17.35—Block diagram of a double-conversion superhet that eliminates some of the tracking problems of the conventional superhet.

One of the main bonuses of this approach is that the tunable second LO can be very stable and accurately calibrated. This calibration is the same for any signal band. Another advantage is that the first crystal LO is very stable and has little phase noise. A third bonus is that the high value of the first IF simplifies the preselector design for good image rejection in the first mixer (Ref 58).

The second mixer is vulnerable to two-tone IMD caused by strong interfering signals that lie within, or near, the 200-kHz-wide first-IF bandwidth, and that have been amplified by the circuitry preceding it. They do not make AGC because they are outside the narrow signal filters.

This cascaded-selectivity problem, which we have discussed previously, makes it necessary to very carefully control the gain and noise-figure distribution ahead of the second mixer. Also, put the narrow signal filter right after the second mixer and follow that with a low-noise IF amplifier, so that "front end" gain can be minimized. In more expensive receivers of this kind, the first IF is sharply gang-tuned along with the second LO in order to reduce this problem (Ref 59).

This general approach has been extended in order to make a general-coverage receiver that has acceptable spurious responses. The first IF can be switched between two different frequency ranges and various combinations of up conversion and down conversion are used. This subject is interesting, but more complex than we can cover here. This approach is also not frequently used at this time.

THE UP CONVERSION SUPERHET

The most common approach to superhet design today is the "up converter." This designation is reserved for receivers in which the first IF is greater than the highest receive frequency. First IF values can be as low as 35 MHz for low-cost HF receivers or as high as 3 GHz for wideband receivers (and spectrum analyzers) that cover the 1 MHz to 2.5 GHz range. Let's begin by discussing the general properties of all up conversion receivers (Ref 42).

An Up Converter Example

The block diagram in **Fig 17.36** is one example for HF amateur SSB/CW use. The input circuit responds uniformly to a wide frequency band, 1.8 to 30 MHz. A 1.8 to 30-MHz band-pass filter is at the input. The absence of any narrow preselection is typical, but in difficult environments an electronically tuned or electromechanically tuned preselector is often used. Another option is a set of "half octave" (2 to 3 MHz, 3 to 4.5 MHz and so on) filters switched by PIN diodes or relays. This type of filter eliminates second-order IMD. For example, if we are listening to a weak signal at 2.00 MHz, two strong signals at 2.01 and 4.01 MHz would not create a spur at 2.00 MHz because the one at 4.01 MHz would be greatly attenuated. Fig 17.8 gives more information on this subject.

Wideband Interference

The wideband circuitry in the front end is vulnerable to strong signals over the entire frequency range if no preselection is used. Therefore the strong-signal performance is a major consideration. Total receiver noise figure is usually allowed to increase somewhat in order to achieve this goal. Double balanced passive (or often active) mixers with high intercept points (second and third-order) and high LO levels are common. A typical high-quality up conversion HF receiver has a third-order intercept (IP3) of 20 to 30 dBm and a noise figure of 10 to 14 dB. High-end performers will have an IP3 of 32 to 40 dBm and a noise figure of 8 to 12 dB.

As a practical matter, in all but the most severe situations with collocated transmitters, there is very little need in Amateur Radio for the most advanced receiver specifications. One reason for this involves statistics. To get two-tone IMD interference on a *high-quality* receiver at some particular frequency there must be two strong signals, or perhaps one very strong and a second weaker signal, on just the right pair of frequencies and at the same time. In nearly all cases, the "chances" of this are small. In Amateur Radio contest situations, these kinds of interactions are more probable. For persistent cases, other remedies are usually available.

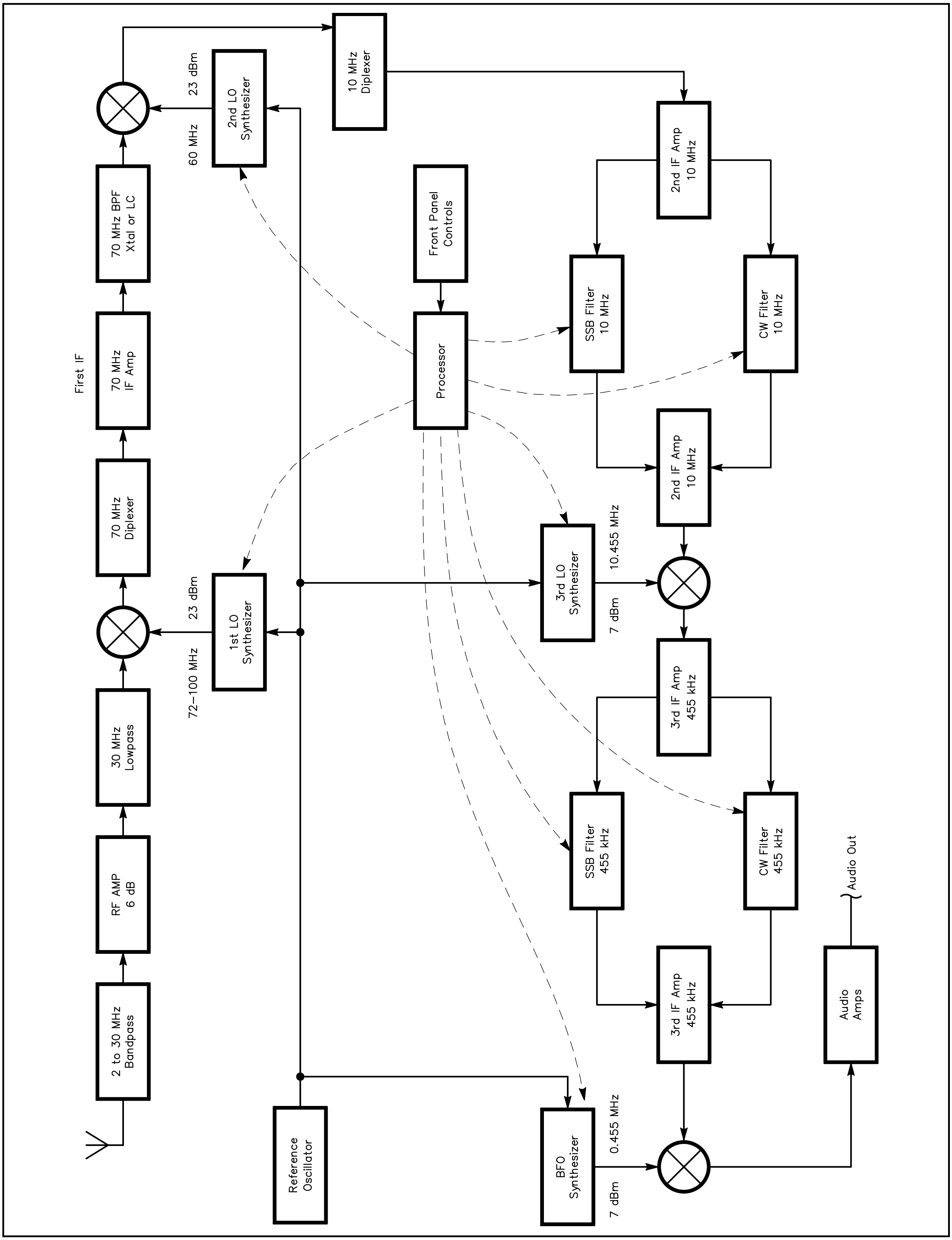


Fig 17.36—Block diagram of an HF up-conversion receiver for SSB and CW. Microprocessor control of receiver functions is included. LOs are from synthesizers.

After the Mixer

We would really like to go from the mixer directly into an SSB or CW filter, but at the high frequency of the first IF this is not realistic. Therefore we run into a major compromise: It is necessary to have at least one additional wide-band frequency conversion before getting to the narrow filters. The first IF filter can be as narrow as cost and technology will permit. In the 35 to 110-MHz range crystal filters with bandwidths of 10 to 20 kHz are available, but they are somewhat expensive in small quantities. Fig 17.36 shows an option with far less cost. The LC filter in the first IF is about 1.0-MHz wide but it has enough attenuation at 50 MHz to yield excellent image rejection in the second mixer. If we use a high-input-intercept, low-gain, low-noise amplifier followed by a strong second mixer (minimize the gain ahead of the second mixer and let the receiver noise figure go up a couple of dB) the overall receiver performance will be excellent, especially with the kinds of efficient antennas that amateurs use.

Terminating the Mixers

In the upconversion receiver, getting a pure wideband resistive termination for the mixer IF port is a problem. The output of the first mixer in Fig 17.36 contains undesired frequencies. For example, a 10-MHz signal produces 70-MHz (desired IF), 90-MHz (image) and 80-MHz (LO leakage). For a 2-MHz signal there would be 70, 72 and 74-MHz outputs. A filter that passes 70-MHz, rejects the others and at the same time terminates the mixer resistively over a wide band is a complicated band-pass diplexer.

Usually the termination is an amplifier input impedance plus a much simpler band-pass diplexer. The amplifier input should be a pure resistance, and it may then be required to deal with the vector sum of all three products. Diode upconverter mixers have typically 30 to 40 dB LO-to-IF isolation. If the LO level is 23 dBm, the amplifier may be looking at –7 to –17 dBm of LO feedthrough, which is fairly strong. The output of the amplifier and also the next stage must deal with these amplified values. The second mixer is much easier to terminate with a diplexer, as Fig 17.36 shows (Ref 60).

At lower signal and LO levels (7 dBm or less), MMIC amplifiers like the MAV-11 may provide a good termination across a wide bandwidth. However, susceptibility to IMD must be checked carefully.

At the RF terminal of the mixer, any noise at the image frequency from previous stages (such as RF amplifier, antenna or even thermal noise) must not be allowed to enter the mixer because it degrades the mixer noise figure. The RF terminal should be short circuited at the image frequency if possible.

Choosing the First IF

The choice of the first IF is a compromise between cost and performance. First, consider harmonic IMD. Published data for several high-level diode-mixer models show that if the IF is greater than three times the highest signal frequency (greater than 90 MHz for a 0 to 30-MHz receiver) the rejection of harmonics of the signal frequency increases considerably. For example 3 times an interferer at 33 MHz produces a 99 MHz IF. The input 30-MHz low-pass filter would attenuate the 33-MHz signal and so would help considerably. On the other hand, 24.75 times 4 is also 99 MHz, but the mixer does a better job of rejecting this fourth harmonic. Other spurious responses tend to improve also.

However, other factors are involved, most important of which are the LO designs for the first and second mixers. In up-converters, the LOs are invariably synthesizers whose output frequencies are phase locked to a low-frequency reference crystal oscillator. As the LO frequencies increase, two other things increase: cost and quantity of high-frequency synthesizer components, and synthesizer phase noise. Also, the exact choice is interwoven with the details of the synthesizer design. Special IFs, such as 109.350 MHz, are chosen after complex trade-off studies. The cost of the first-IF signal-path components, especially filters, tends to increase also.

For all of these reasons, the IF is quite often chosen at a lower frequency. In Fig 17.36, a 70-MHz IF is shown. Crystal filters at this frequency are widely available at reasonable cost. LC filters with a

1.0 MHz (or less) bandwidth are easy to construct and get working. A 45-MHz IF is also popular. Helical resonator filters are excellent candidates at higher IF frequencies, although they can be a bit large.

The problems associated with lower IFs can be greatly improved by using a higher performance mixer. The costs are a better mixer and a more powerful LO amplifier.

The "Gray Area"

In the upconverter front end we encounter the "cascaded selectivity" problem that was mentioned previously in this chapter. Strong signals that are *within* the first IF passband but *outside* the SSB/CW passband (the gray area) do not make AGC and therefore are not controlled by AGC. These signals intermodulate in the mixers and in the amplifiers that precede the SSB/CW filters. It is important to realize that the receiver gray-area IMD performance is the composite of all these stages, and not just the first mixer alone. It is not until the first IF filter takes effect that things get a lot better. This degradation for strong adjacent-channel signals is an artifact of the upconverting superhet receiver. In practice, it often happens that reciprocal mixing with the adjacent channel signals (caused by receiver synthesizer phase noise) also gets into the act. Phase noise on the interfering signals is also not uncommon. The design problem is to make IMD in the gray area as small as reasonably possible. The example of Fig 17.36 is an acceptable compromise in this respect for a fairly high quality receiver.

Triple Conversion

The block diagram of Fig 17.36 shows a 10-MHz second IF. This IF is selected for two reasons:

- There are narrowband crystal filters available (including homemade ladder filters) in that vicinity.
- The image frequency for the second mixer is at 50 MHz, which can be highly attenuated by the first IF filter.

It is entirely possible to let the rest of the IF remain at this frequency, and many receivers do just that. Nonetheless, there are some advantages to having a third conversion to a lower IF. For example, it is desirable to get the large amount of IF gain needed at two different frequencies, so that stability problems and leakage from the output to input are reduced. Also, a wide variety of excellent filters are readily available at 455 kHz and other low frequencies.

Note that there is a cascade of the SSB and CW filters at the two frequencies. The desirability of this was discussed earlier and it is a powerful concept for narrow-band receiver design. Placing cascaded filters at two different frequencies has another advantage that we will look into presently.

For the third mixer the image frequency is at 10.910 MHz, therefore the 10-MHz filter must have good rejection at that frequency. If the first IF filter has good rejection at 70.910 MHz it will help reduce this image also. We are looking for 90 to 100 dB rejection of this potentially serious image problem. If the first IF uses a 1.0 MHz wide LC filter, as we have suggested, some additional 10 MHz LC filtering will probably be needed.

The 50 kHz IF

The third IF could be the cascaded LC filter circuits at VLF, one for SSB and one for CW, which we discussed earlier. This is an excellent approach that has been used often. The image frequency for the third mixer is only 100 kHz away, so that problem needs careful attention. Image canceling mixers have been used to help get the required 90 to 100 dB (Ref 41).

Local Oscillator (LO) Leakage

It is easy to see that with four LOs running, some at levels of 0.2 W or more, interactions between the mixers can occur. In a multiple-conversion receiver mechanical packaging, shielding, circuit placement and lead filtering are very critical areas. As one example of a problem, if the 60-MHz second LO leaks into the first mixer, a vulnerability to strong 10-MHz input signals results. It is called "IF feedthrough" because the second IF is 10 MHz. Other audible "tweets" are very common; they occur at various frequencies that involve harmonics of LOs beating together in various mixers. It is a major exercise to devise a "frequency scheme" that minimizes tweets, or at least puts them where they do not cause too much trouble (for amateurs, outside the ham bands). After that the "dog work" of reducing the remaining tweets below the noise level begins. It is a very educational experience. Synthesizers produce numerous artifacts that can also be very troublesome. It is a very common dilemma to build a receiver using "cheap" construction and poorly conceived packaging, and then try to bully the thing into good behavior.

Frequency Tuning

The synthesized first and second LOs present several different ways to tune the receiver frequency. This chapter cannot get into the details of synthesizer design, so these are only a few brief remarks. Let's discuss two options from a system-design standpoint:

- Do all of the tuning in the first LO. If steps of 10 Hz (or 1 Hz) are needed, a single-loop synthesizer that tunes in, say 500-kHz or 1.0-MHz steps, can be used. Then, a direct digital synthesizer (DDS) that tunes in 10-Hz (or 1-Hz) steps is included in the main loop in what is termed a "translation loop." The DDS frequency is added into the loop in such a manner that its imperfections are not increased by frequency multiplication. Because the reference frequency for the loop is high (500 kHz or 1.0 MHz) the phase noise of the main loop is quite small, if the loop and the circuitry are correctly designed and if the LO frequency is not extremely high. The digital frequency readout is obtained from the bits that program the synthesizer. A simpler approach might use a free-running VFO plus a low-cost frequency counter instead of the DDS. The counter can be designed to display the receiver signal frequency.
- If the first IF filter is sufficiently wideband it is possible to tune the first LO in steps of 500 kHz or 1.0 MHz and tune the second LO in 10-Hz steps. This may be a simpler method because the second LO need only be tuned over a small range.

With this second approach the first LO could be a crystal oscillator with switched overtone crystals, one for each ham band. The second LO could be a combination of crystal and VFO. One disadvantage (not extremely serious for amateur work) is that the LOs are not locked to a very accurate reference. Another is that a separate crystal (easy to get) is needed for each band. A frequency counter on the second LO could be used to get a close approximation to the signal frequency. This approach might be of interest to the home builder who is not yet ready to get involved with synthesizers. A 500-kHz crystal calibrator in the receiver would mark the band edges accurately.

Passband Tuning

While listening to a desired signal in the presence of another partially overlapping and interfering signal, whether in SSB or CW mode, it is often possible to "move" the interference at least somewhat out of the receiver passband without affecting the tune frequency (pitch) of the desired signal. In Fig 17.36, if the processor has independent fine tuning control of the second and third LOs and also the BFO, it is a matter of software design to accomplish this. It is done by controlling the overlap or intersection of the passbands of the 10-MHz filters and the 455-kHz filters. There are three things that can be done: the bandwidth can be decreased, the center frequency of the passband can be moved and both can be done simultaneously (Ref 61).

This scheme works best when both SSB or both CW filters are of high quality and have the same bandwidth (for example, 2.5 kHz and 500 Hz), fairly flat response and the same shape factor. As the passband is made narrower by decreasing the overlap, however, the composite shape factor is degraded somewhat (it gets larger). For CW especially, this is not detrimental. A very steep-sided response at narrow bandwidths is not desirable from a transient-response standpoint. The effect is not serious for SSB either.

Later discussion in the Transceivers section of this chapter will present another method of passband tuning using a variable frequency mixer scheme rather than software control. This method is commonly found in manufactured equipment.

Noise Blanking

The desire to eliminate impulse noise from the receiver audio output has led to the development of special IF circuits that detect the presence of a noise impulse and open the signal path just long enough to prevent the impulse from getting through. Most often, a diode switch is used to open the signal path. An important design requirement is that the desired IF signal must be delayed slightly, ahead of the switch, so that the switch is opened precisely when the noise arrives at the switch. The circuitry that detects the impulse and operates the switch has a certain time delay, so the signal in the mainline IF path must be delayed also. The Transceivers section of this chapter describes how a noise blanker is typically implemented. (See also Refs 48, 62.)

VHF and UHF Receivers

The basic ideas presented in previous sections all operate equally well in receivers that are intended for the VHF and UHF regions. One difference, however, is that narrow-band frequency modulation (NBFM) is commonly used. In recent times, however, SSB has increased in popularity because of its potentially narrower channel spacing, (Ref 63). The commercial use of SSB at VHF has been encouraged by the use of a –10 dB pilot carrier combined with amplitude compandoring. This is known as amplitude-compandered SSB or ACSSB. Improvements in synthesizer design and frequency stability have also contributed to the growth of VHF SSB. We will focus the differences between VHF/UHF and HF receivers.

Narrowband FM (NBFM) Receivers

Fig 17.37A is a block diagram of an NBFM receiver for the VHF/UHF amateur bands.

Front End

A low-noise front end is desirable because of the decreasing atmospheric noise level at these frequencies and also because portable gear uses short rod antennas at ground level. Nonetheless, the possibilities for gain compression and harmonic IMD, multitone IMD and cross modulation are also substantial. So dynamic range is an important design consideration, especially if large, high-gain antennas are used. FM limiting should not occur until after the NBFM signal filter. Because of the high occupancy of the VHF/UHF spectrum by powerful broadcast transmitters and nearby two-way radio services, front-end preselection is desirable, so that low noise figure can be achieved economically within the amateur band. Fig 17.37B is an example of a simple front end for the 144- to 148-MHz band.

Downconversion

Downconversion to the final IF can occur in one or two stages. Favorite IFs are in the 5 to 10-MHz region, but at the higher frequencies rejection of the image 10 to 20-MHz away can be difficult, requiring considerable preselection. So at the higher frequencies an intermediate IF in the 30 to 50-MHz region is a better choice. Fig 17.37A shows dual downconversion.

IF Filters

The customary frequency deviation in amateur NBFM is about 5 kHz RMS (7-kHz peak) and the audio speech band extends to 5 kHz. This defines a minimum modulation index (defined as the deviation ratio) of 7/5 = 1.4. An inspection of the Bessel function plots shows that this condition confines most of the 300 to 5000-Hz speech information sidebands within a 12 to 15-kHz-wide bandwidth filter. With this bandwidth, channel separations of 20 or 25 kHz are achievable.

Many amateur NBFM transceivers are channelized in steps that can vary from 5 to 25 kHz. For low distortion of the audio output (after FM detection), this filter should have good phase linearity across the bandwidth. This would seem to preclude filters with very steep descent outside the passband, that tend to have very nonlinear phase near the band edges. But since the amount of energy in the higher speech frequencies is naturally less, the actual distortion due to this effect may be acceptable for speech purposes. A possible qualifier to this may be when preemphasis of the higher speech frequencies occurs at the transmitter and deemphasis compensates at the receiver (a commonly found feature).

Limiting

After the filter, hard limiting of the IF is needed to remove any amplitude components. In a high-quality receiver, special attention is given to any nonlinear phase shift that might result from the limiter circuit design. This is especially important in phase-coherent data receivers. In amateur receivers for

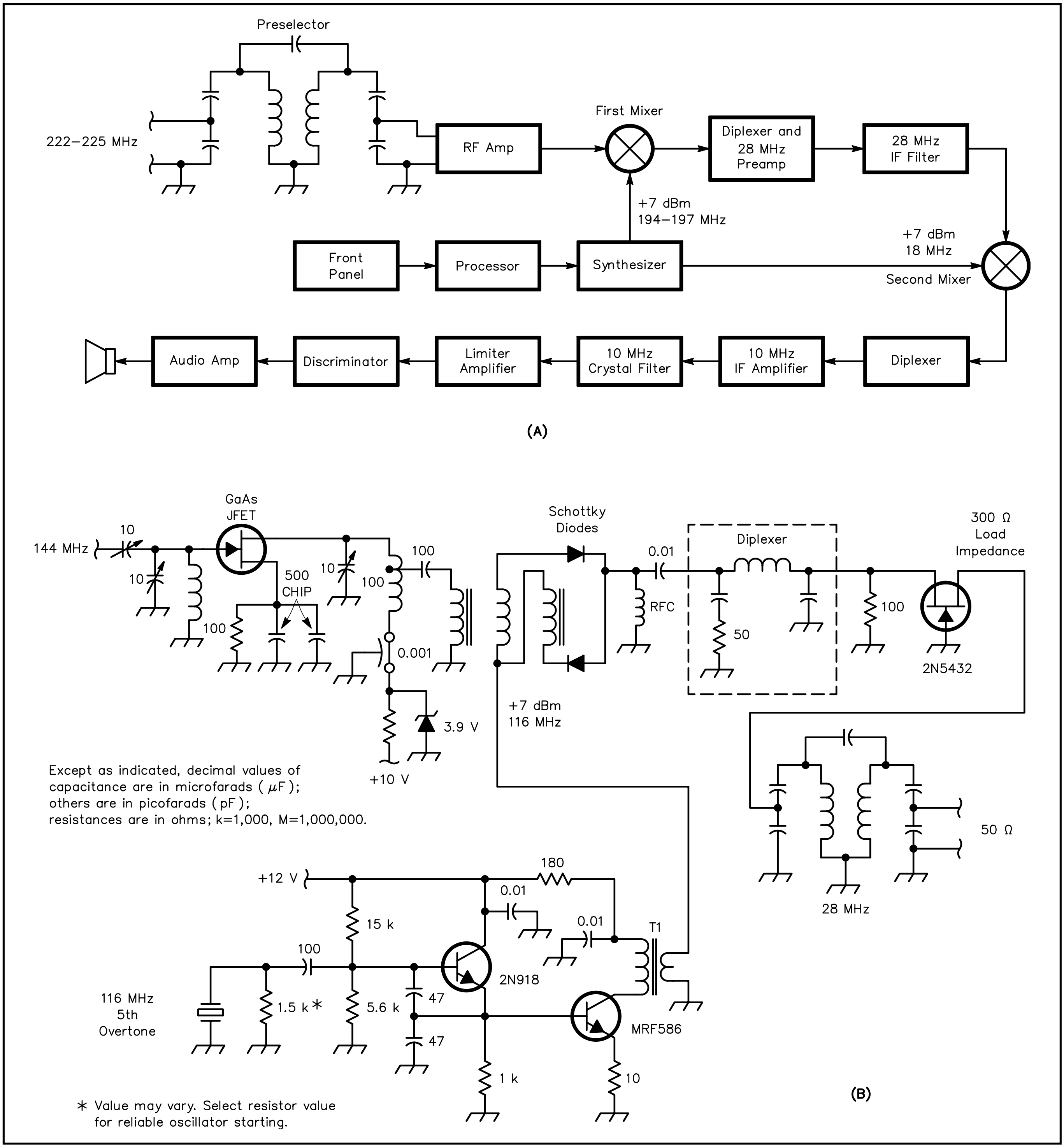


Fig 17.37—Some narrowband FM circuits. A: block diagram of a typical NBFM receiver. B: a front-end circuit with preselection and down conversion.

speech it may be less important. Also, the "ratio detector" (see the **Mixers, Modulators and Demodulators** chapter) largely eliminates the need for a limiter stage, although the limiter approach is probably still preferred.

FM Detection

The discussion of this subject is deferred to the **Mixers, Modulators and Demodulators** chapter.

Quadrature detection is used on some popular NBFM multistage ICs. An example receiver chip will be presented later. Also see the Transceivers section of this chapter.

NBFM Receiver Weak-Signal Performance

The noise bandwidth of the IF filter is not much greater than twice the audio bandwidth of the speech modulation, as it would be in wideband FM. Therefore such things as capture effect, the threshold effect and the noise quieting effect so familiar to wideband FM are still operational, but somewhat less so, in NBFM. For NBFM receivers, sensitivity is specified in terms of a SINAD (see the **Test Procedures** chapter) ratio of 12 dB. Typical values are –110 to –125 dBm, depending on the low-noise RF preamplification that often can be selected or deselected (in strong signal environments).

LO Phase Noise

In an FM receiver, LO phase noise superimposes phase modulation, and therefore frequency modulation, onto the desired signal. This reduces the ultimate signal-to-noise ratio within the passband. This effect is called "incidental FM (IFM)." The power density of IFM (W/Hz) is equal to the phase noise power density (W/Hz) multiplied by the square of the frequency offset from the carrier (the familiar parabolic effect in FM). If the receiver uses high-frequency deemphasis at the audio output (–6 dB per octave from 300 to 3000 Hz, a common practice), the IFM level at higher audio frequencies can be reduced. Levels of total (integrated) IFM from 10 to 50 Hz are high quality for amateur voice work. Ordinarily, as the signal increases the noise would be "quieted" (that is, "captured") in an FM receiver, but in this case the signal and the phase noise riding "piggy back" on the signal increase in the same proportion. As the signal becomes large the signal-to-noise ratio therefore approaches some final value (Ref 23). A similar ultimate SNR effect occurs in SSB receivers. On the other hand, a perfect AM receiver tends to suppress LO phase noise (Ref 33).

NBFM ICs

A wide variety of special ICs for NBFM receivers are available. Many of these were designed for "cordless" or cellular telephone applications and are widely used. **Fig 17.38** shows some popular versions for a 50-MHz NBFM receiver. One is an RF amplifier chip (NE/SA5204A) for 50-Ω input to 50-Ω output with 20 dB of gain. This gain should be reduced to perhaps 8 dB. The second chip (NE/SA602A) is a front-end device (Ref 75) with an RF amplifier, mixer and LO. The third is an IF amplifier, limiter and quadrature NBFM detector (NE/SA604A) that also has a very useful RSSI (logarithmic received signal strength indicator) output and also a "mute" function. The fourth is the LM386, a widely used audio-amplifier chip. Another NBFM receiver chip, complete in one package, is the MC3371P.

The NE/SA5204A plus the two tuned circuits help to improve image rejection. An alternative would be a single double tuned filter with some loss of noise figure. The Mini-Circuits MAR/ERA series of MMIC amplifiers are excellent devices also. The crystal filters restrict the noise bandwidth as well as the signal bandwidth. A cascade of two low-cost filters is suggested by the vendors (Ref 64). Half-lattice filters at 10 MHz are shown, but a wide variety of alternatives, such as ladder networks, are possible.

Another recent IC is the MC13135, which features double conversion and two IF amplifier frequencies. This allows more gain on a single chip with less of the cross-coupling that can degrade stability. This desirable feature of multiple downconversion was mentioned previously in this chapter.

The diagram in Fig 17.38 is (intentionally) only a general outline that shows how chips can be combined to build complete equipment. The design details and specific parts values can be learned from a careful study of the data sheets and application notes provided by the IC vendors. Amateur

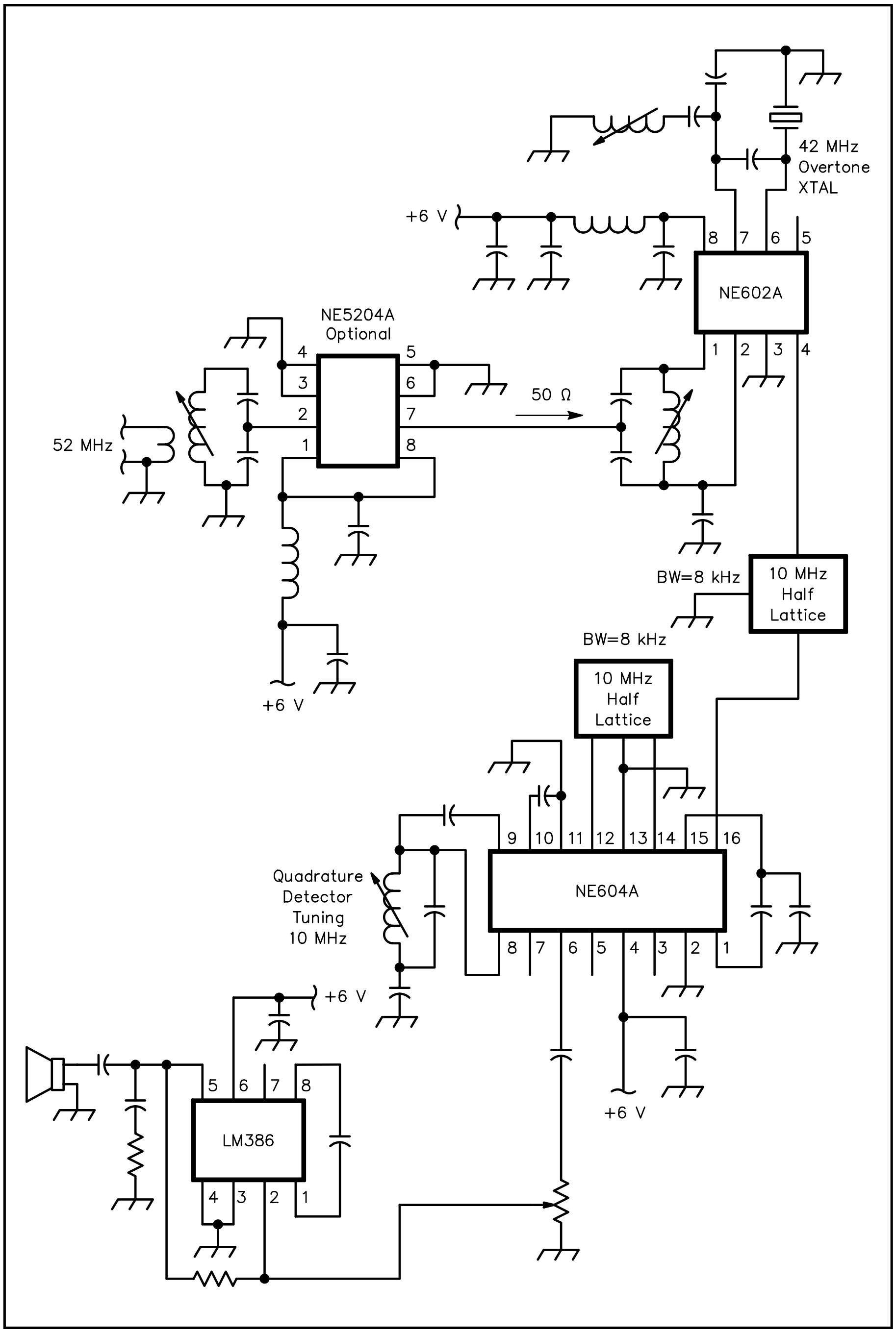


Fig 17.38—The NE/SA5204A, NE/SA602A, NE/SA604A NBFM ICs and the LM386 audio amplifier in a typical amateur application for 50 MHz.

designers should learn how to use these data sheets and other information. The best places to learn about data sheets are data books and application notes.

UHF Techniques

The Ultra High Frequency spectrum is considered to be the range from 300 MHz to 3 GHz. All of the basic principles of radio system design and circuit design that have been discussed so far apply as well in this range, but the higher frequencies require some special thinking about the methods of circuit design and the devices that are used.

GaAs FET PREAMP FOR 430 MHZ

Fig 17.39 shows the schematic diagram and the physical construction of a typical RF circuit at 430 MHz. It is a GaAsFET preamplifier intended for low noise Earth-Moon-Earth or satellite reception.

The construction uses chip capacitors, small helical inductors and a stripline surface-mount GaAsFET, all mounted on a G10 (two layers of copper) glass-epoxy PC board. The very short length of interconnection leads is typical. The bottom of the PC board is a ground plane. At this frequency, lumped components are still feasible, while microstrip circuitry tends to be rather large.

At higher frequencies, microstrip methods become more desirable in most cases because of their smaller dimensions. However, the advent of tiny chip capacitors and chip resistors have extended the frequency range of discrete components. For example, the literature shows methods of building LC filters at as high as 2 GHz, using chip capacitors and tiny helical inductors (Refs 65, 66). Commercially available amplifier and mixer circuits operate at 2 GHz, using these types of components on ceramic substrates.

ICs for UHF

In recent years a wide variety of highly miniaturized monolithic microwave ICs (MMIC) have become available at reasonable cost. Among these are the Avantek MODAMP and the Mini Circuits MAR and MAV/ERA lines (see Ref 89). Designer kits containing a wide assortment of these are available at reasonable cost. They come in a wide variety of gains, intercepts and noise figures for frequency ranges from dc to 2 GHz. A more expensive option, for more sophisticated receiver applications, are the hybrid "cascadable" amplifiers, built on ceramic substrates and mounted in TO5 or TO8 metal cans. Most of these circuits are intended for a 50-Ω to 50-Ω interface.

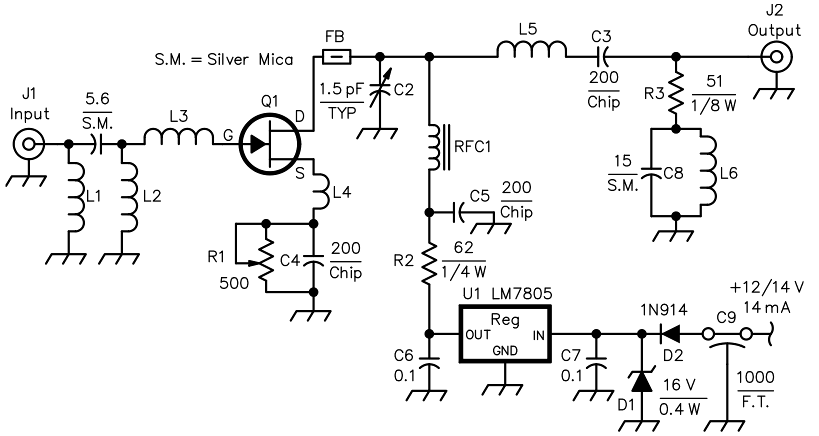
A wide variety of hybrid amplifiers, designed for the Cable TV industry, are available for the frequency range from 1 to 1200 MHz (for example, the Motorola CA series, in type 714x packages). These have gains from 15 to 35 dB, output 1-dB compression points from 22 to 30 dBm and noise figures from 4.5 to 8.5 dB. Such units are excellent alternatives to discrete home-brew circuits for many applications where very low noise figures are not needed. In small quantities they may be a bit expensive sometimes, but the total cost of a home-built circuit, including labor (even the amateur experimenter's time is not really "free") is often at least as great. Home-built circuits do, however, have very important educational value.

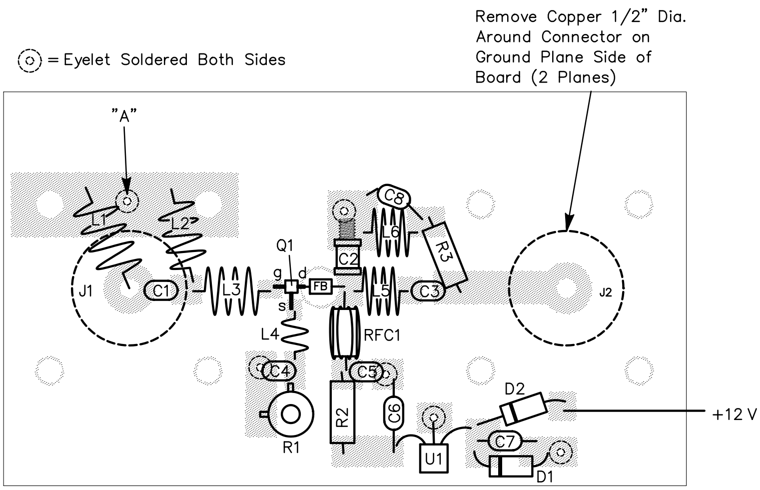
UHF Design Aids

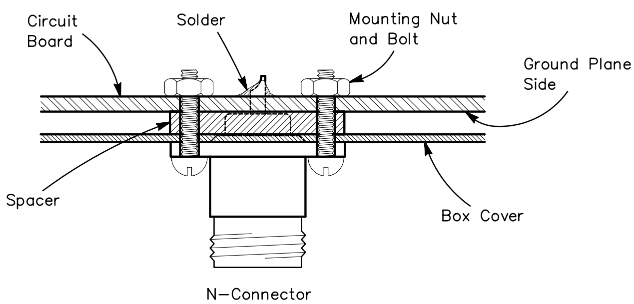
Circuit design and evaluation at the higher frequencies usually require some kind of minimal lab facilities, such as a signal generator, a calibrated noise generator and, hopefully, some kind of simple (or surplus) spectrum analyzer. This is true because circuit behavior and stability depend on a number of factors that are difficult to "guess at," and intuition is often unreliable. The ideal instrument is a vector network analyzer with all of the attachments (such as an S-parameter measuring setup), but very few amateurs can afford this. Another very desirable thing would be a circuit design and analysis program for the personal computer. Software packages created especially for UHF and microwave circuit design are available. They tend to be somewhat expensive, but worthwhile for a serious designer. Inexpensive SPICE programs are a good compromise. ARRL Radio Designer is an excellent, low cost choice.

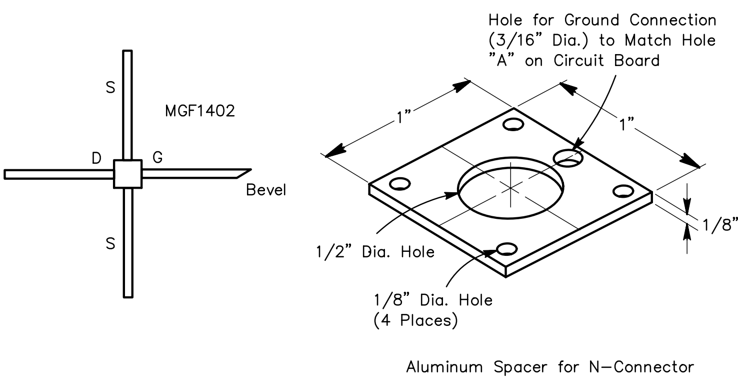
A 902 to 928 MHz (33 cm) Receiver

Fig 17.40A is a block diagram of a 902-MHz down-converting receiver. A cavity resonator at the antenna input provides high selectivity with low loss. The first RF amplifier is a GaAsFET. Two additional 902-MHz band-pass microstrip filters and a BFR96 transistor provide more gain and image rejection (at RF–56 MHz) for the Mini Circuits SRA12 mixer. The output is at 28.0 MHz.









C1—5.6-pF silver-mica capacitor or same as C2.

C2—0.6- to 6-pF ceramic piston trimmer capacitor (Johanson 5700 series or equiv).

C3, C4, C5—200-pF ceramic chip capacitor.

C6, C7—0.1-mF disc ceramic capacitor, 50 V or greater.

C8—15-pF silver-mica capacitor.

C9—500- to 1000-pF feedthrough capacitor.

D1—16- to 30-V, 500-mW Zener diode (1N966B or equiv).

D2—1N914, 1N4148 or any diode with ratings of at least 25 PIV at 50 mA or greater.

J1, J2—Female chassis-mount Type-N connectors, PTFE dielectric (UG-58 or equiv).

L1, L2—3t, #24 tinned wire, 0.110-inch ID spaced 1 wire diam.

L3—5t, #24 tinned wire, $^{3}/_{16}$-inch ID, spaced 1 wire diam. or closer. Slightly larger diameter (0.010 inch) may be required with some FETs.

L4, L6—1t #24 tinned wire, $^{1}/_{8}$-inch ID.

L5—4t #24 tinned wire, $^{1}/_{8}$-inch ID, spaced 1 wire diam.

Q1—Mitsubishi MGF1402.

R1—200- or 500-Ω cermet potentiometer (initially set to midrange).

R2—62-Ω, $^{1}/_{4}$-W resistor.

R3—51-Ω, $^{1}/_{8}$-W carbon composition resistor, 5% tolerance.

RFC1—5t #26 enameled wire on a ferrite bead.

U1—5-V, 100-mA 3-terminal regulator (LM78L05 or equiv. TO-92 package).

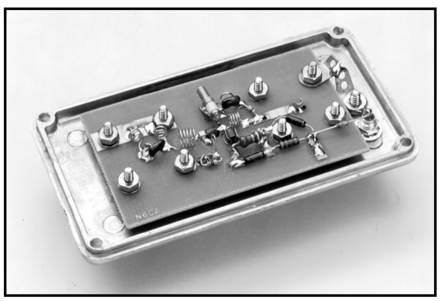
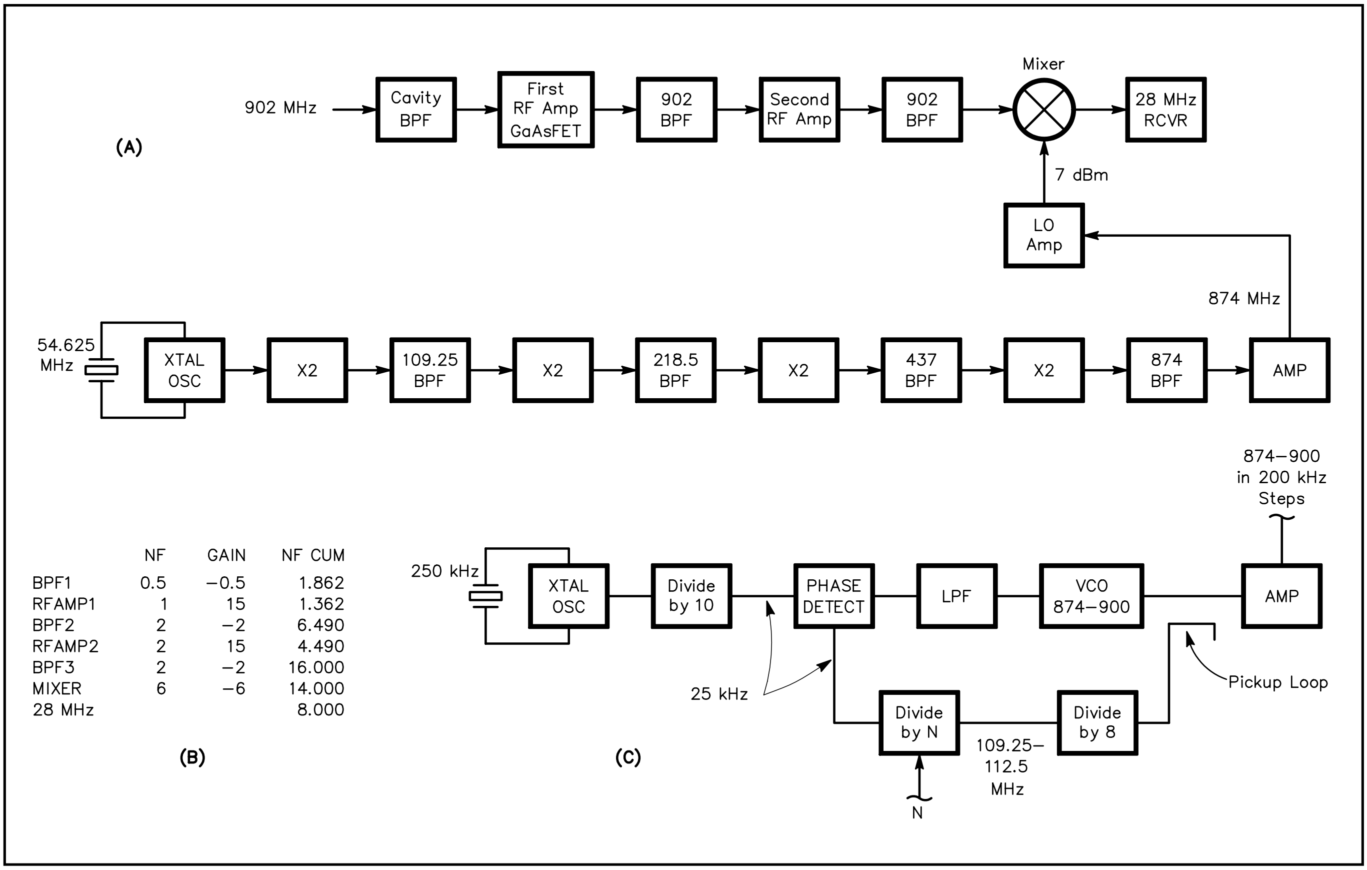


Fig 17.39—GaAsFET preamplifier schematic and construction details for 430 MHz. Illustrates circuit, parts layout and construction techniques suitable for 430-MHz frequency range.



	NF	GAIN	NF CUM
BPF1	0.5	−0.5	1.862
RFAMP1	1	15	1.362
BPF2	2	−2	6.490
RFAMP2	2	15	4.490
BPF3	2	−2	16.000
MIXER	6	−6	14.000
28 MHz			8.000

Fig 17.40—A down converter for the 902 to 928 MHz band. A: block diagram; B, cumulative noise figure of the signal path; C, alternative LO multiplier using a phase locked loop.

Cumulative Noise Figure

Fig 17.40B shows the cumulative noise figure (NF) of the signal path, including the 28-MHz receiver. The 1.5-dB cumulative NF of the input cavity and first-RF-amplifier combination, considered by itself, is degraded to 1.9 dB by the rest of the system following the first RF amplifier. The NF values of the various components for this example are reasonable, but may vary somewhat for actual hardware. Also, losses prior to the input, such as transmission-line losses (very important), are not included. They would be part of the complete receive-system analysis, however. It is common practice to place a low-noise preamp outdoors, right at the antenna, to overcome coax loss (and to permit use of inexpensive coax).

Local Oscillator (LO) Design

The +7 dBm LO at 874 to 900 MHz is derived from a set of crystal oscillators and frequency multipliers, separated by band-pass filters. These filters prevent a wide assortment of spurious frequencies from appearing at the mixer LO port. They also enhance the ability of the doubler stage to generate the second harmonic. That is, they have a very low impedance at the input frequency, thereby causing a large current to flow at the fundamental frequency. This increases the nonlinearity of the circuit, which increases the second-harmonic component. The higher filter impedance at the second harmonic produces a large harmonic output.

For very narrow-bandwidth use, such as EME, the crystal oscillators are often oven controlled or otherwise temperature compensated. The entire LO chain must be of low-noise design and the mixer should have good isolation from LO port to RF port (to minimize noise transfer from LO to RF).

A phase-locked loop using GHz-range prescalers (as shown in Fig 17.40C) is an alternative to the multiplier chain. The divide by N block is a simplification; in practice, an auxiliary dual-modulus divider in a "swallow count" loop would be involved in this segment.

The cascaded 902-MHz band-pass filters in the signal path should attenuate any image frequency noise (at RF–56 MHz) that might degrade the mixer noise figure.

Summary

This example is fairly typical of receiver design methods for the 500 to 3000 MHz range, where down conversion to an existing HF receiver is the most-convenient and cost-effective approach for amateurs. At higher frequencies a double down conversion with a first IF of 200 MHz or so, to improve image rejection, might be necessary. Usually, though, the presence of strong signals at image frequencies is less likely. Image reducing mixers plus down conversion to 28 MHz is also coming into use, when strong interfering signals are not likely at the image frequency.

"No-Tune" Techniques

In recent years, a series of articles have appeared that emphasize simplicity of construction and adjustment. The use of printed-circuit microstrip filters that require little or no adjustment, along with IC or MMIC devices, or discrete transistors, in precise PC-board layouts that have been carefully worked out, make it much easier to "get going" on the higher frequencies. Several of the References at the end of this discussion show how these UHF and microwave units are designed and built (Refs 67, 68, 69, 70, 71). See also the projects at the end of this chapter.

Microwave Receivers

The world above 3 GHz is a vast territory with a special and complex technology and an "art form" that are well beyond the scope of this chapter. We will scratch the surface by describing a specific receiver for the 10-GHz frequency range and point out some of the important special features that are unique to this frequency range.

A 10-GHz Preamplifier

Fig 17.41A is a schematic and parts list, B is a PC-board parts layout and C is a photograph for a 10-GHz preamp, designed by Zack Lau at ARRL HQ. With very careful design and packaging techniques a noise figure approaching the 1 to 1.5-dB range was achieved. This depends on an accurate 50-Ω generator impedance and noise matching the input using a microwave circuit-design program such as Touchstone or Harmonica. Note that microstrip capacitors, inductors and transmission-line segments are used almost exclusively. The circuit is built on a 15-mil Duroid PC board. In general, this kind of performance requires some elegant measurement equipment that few amateurs have. A detailed discussion appears in Ref 72; it is recommended reading. On the other hand, preamp noise figures in the 2 to 4-dB range are much easier to get (with simple test equipment) and are often satisfactory for terrestrial communication.

Articles written by those with expertise and the necessary lab facilities almost always include PC board patterns, parts lists and detailed instructions that are easily duplicated by readers. So it is possible to "get going" on microwaves using the material supplied in the articles. Complete kits are available from suppliers listed in the ARRL Address List in the **References** chapter. Microwave ham clubs and their publications are a good way to get started in microwave amateur technology.

Because of the frequencies involved, dimensions of microstrip circuitry must be very accurate. Dimensional stability and dielectric constant reliability of the boards must be very good.

System Performance

At microwaves, an estimation of system performance can often be performed using known data about

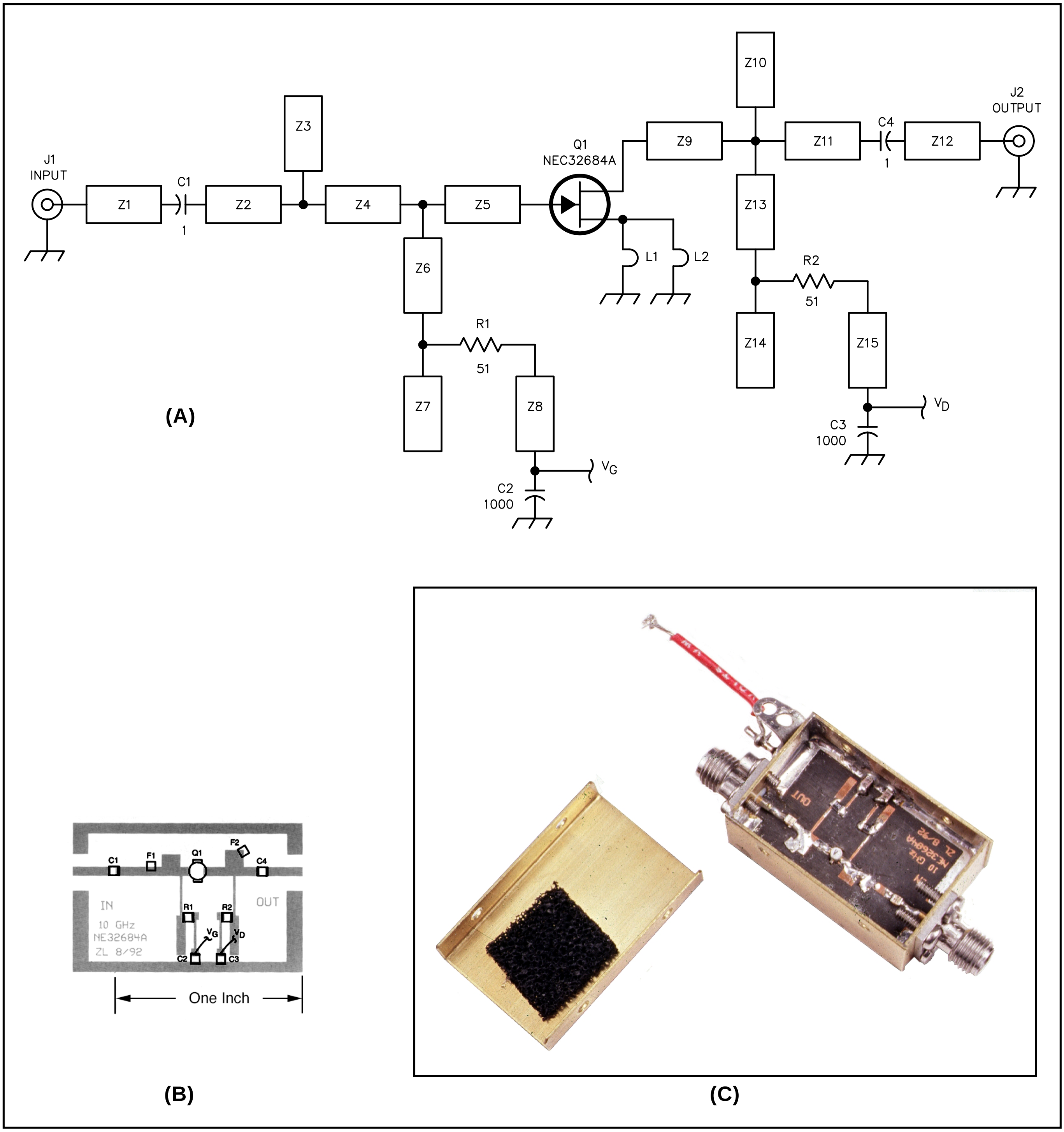


Fig 17.41—A low-noise preamplifier for 10 GHz, illustrating the methods used at microwaves. A: schematic. B: PC board layout. Use 15-mil 5880 Duroid, dielectric constant of 2.2 and a dissipation factor of 0.0011. For a negative of the board write, phone or e-mail the Technical Department Secretary at ARRL HQ and request the template from the December 1992 *QEX* "RF" column. C: A photograph of the completed preamp.

C1, C4—1-pF ATC 100-A chip capacitors. C1 must be very low loss.

C2, C3—1000-pF chip capacitors. (Not critical.) The ones from Mini Circuits work fine.

F1, F2—Pieces of copper foil used to tune the preamp.

J1, J2—SMA jacks. Ideally these should be microstrip launchers. The pin should be flush against the board.

L1, L2—The 15-mil lead length going through the board to the ground plane.

R1, R2—51-Ω chip resistors.

Z1-Z15—Microstriplines etched on the PC board.

the signal path terrain, atmosphere, transmitter and receivers systems. **Fig 17.42** shows a simplified example of how this works. This example is adapted from December 1980 *QST* (also see ARRL *UHF/ Microwave Experimenter's Manual*, p 7-55). In the present context of receiver design we wish to establish an approximate goal for the receiver system, including the antenna and transmission line.

A more detailed analysis includes terrain variations, refraction effects, the Earth's curvature, diffraction effects and interactions with the atmosphere's chemical constituents and temperature gradients. *The ARRL UHF/Microwave Experimenter's Manual* is a good text for these matters.

In microwave work, where very low noise levels and low noise figures are encountered, experimenters like to use the "effective noise temperature" concept, rather than noise factor. The relationship between the two is given by

$$T_E = 290\,(F - 1) \tag{7}$$

T_E is a measure, in terms of temperature, of the "excess noise" of a component (such as an amplifier). A resistor at 290 + T_E would have the same available noise power as the device (referred to the device's

Analysis of a 10.368 GHz communication link with SSB modulation:

Free space path loss (FSPL) over a 50 mile line-of-sight path (S) at F = 10.368 GHz:
FSPL = 36.6 (dB) + 20 log F (MHz) + 20 log S (Mi) = 36.6 + 80.3 + 34 = 150.9 dB.

Effective isotropic radiated power (EIRP) from transmitter:
EIRP (dBm) = P_{XMIT} (dBm) + Antenna Gain (dBi)

The antenna is a 2 ft diameter (D) dish whose gain GA (dBi) is:
GA = 7.0 + 20 log D (ft) + 20 log F (GHz) = 7.0 + 6.0 + 20.32 = 33.3 dBi

Assume a transmission-line loss L_T, of 3 dB
The transmitter power P_T = 0.5 (mW PEP) = –3 (dBm PEP)
P_{XMIT} = P_T (dBm PEP) – L_T (dB) = –3 –3 = –6 (dBm PEP)
EIRP = P_{XMIT} + G_A = –6 + 33.3 = 27.3 (dBm PEP)
Using these numbers the received signal level is:
P_{RCVD} = EIRP (dBm) – Path loss (dB)
= 27.3 (dBm PEP) – 150.9 (dB) = –123.6 (dBm PEP)
Add to this a receive antenna gain of 17 dB. The received signal is then P_{RCVD} = –123.6 +17 = –106.6 dBm

Now find the receiver's ability to receive the signal:
The antenna noise temperature T_A is 200 K. The receiver noise figure NF_R is 6 dB (FR = 3.98, noise temperature T_R = 864.5 K) and its noise bandwidth (B) is 2400 Hz. The feed line loss L_L is 3 dB (F=2.00, noise temperature T_L = 288.6 K). The system noise temperature is:
$T_S = T_A + T_L + (L_L)\,(T_R)$
T_S = 200 + 288.6 + (2.0) (864.5) = 2217.6 K
$N_S = kT_SB = 1.38 \times 10^{-23} \times 2217.6 \times 2400 = 7.34 \times 10^{-17}$ W = –131.3 dBm
This indicates that the PEP signal is –106.6 –(–131.3) = 24.7 dB above the noise level. However, because the average power of speech, using a speech processor, is about 8 dB less than PEP, the average signal power is about 16.7 dB above the noise level.

To find the system noise factor F_S we note that the system noise is proportional to the system temperature T_S and the "generator" (antenna) noise is proportional to the antenna temperature T_A. Using the idea of a "system noise factor": $F_S = T_S / T_A$ = 2217.6 / 200 = 11.09 = 10.45 dB.
If the antenna temperature were 290 K the system noise figure would be 9.0 dB, which is precisely the sum of receiver and receiver coax noise figures (6.0 + 3.0).

Fig 17.42—Example of a 10-GHz system performance calculation. Noise temperature and noise factor of the receiver are considered in detail.

input) specified by T_E. For a lossy device (such as a lossy transmission line) T_E is given by T_E = 290 (L – 1), where L is the loss factor (same as its noise factor). The cascade of noise temperatures is similar to the Friis formula for cascaded noise factors and its use is illustrated in Fig 17.42.

$$T_S = T_G + T_{E1} + T_{E2}/G_1 + T_{E3}/(G_1G_2) + T_{E4}/(G_1G_2G_3) + \ldots \qquad (8)$$

where T_S is the system noise temperature (including the generator, which may be an antenna) and T_G is the temperature of the antenna.

The 290 number in the formulas for T_E is the standard ambient temperature (kelvins) at which the noise factor of a two-port transducer is defined and measured, according to an IEEE recommendation. So those formulas relate a noise factor F, measured at 290 K, to the temperature T_E. In general, though, it is perfectly correct to say that the ratio $(S_I/N_I)/(S_O/N_O)$ can be thought of as the ratio of total system output noise to that system output noise attributed to the "generator" alone, regardless of the temperature of the equipment or the nature of the generator, which may be an antenna at some arbitrary temperature, for example. This ratio is, in fact, a special "system noise factor (or figure), F_S" that need not be tied to any particular temperature such as 290 K. The use of the F_S notation avoids any confusion. As the example of Fig 17.42 shows, the value of this system noise factor F_S is just the ratio of the total system temperature to the antenna temperature.

Having calculated a system noise temperature, the receive system noise floor (that is, the antenna input level of a signal that would exactly equal system noise, both observed at the receiver output) associated with that temperature is:

$$N = k\, T_S\, B_N \qquad (9)$$

where
$k = 1.38 \times 10^{-23}$ and
B_N = noise bandwidth.

The system noise figure F_S is indicated in the example also. It is higher than the sum of the receiver and coax noise figures.

The example includes a loss of 3 dB in the receiver transmission line. The formula for T_S in the example shows that this loss has a double effect on the system noise temperature, once in the second term (288.6) and again in the third term (2.0). If the receiver (or high-gain preamp with a 6 dB NF) were mounted at the antenna, the receive-system noise temperature would be reduced to 1064.5 K and a system noise figure, F_S, of 7.26 dB, a very substantial improvement. Thus, it is the common practice to mount a preamp at the antenna.

Microwave Receiver for 10 GHz

Ref 73 provides a good example of modern amateur experimenter techniques for the 10-GHz band. The intended use for the radio is narrowband CW and SSB work, which requires extremely good frequency stability in the LO. Here, we will discuss the receiver circuit (see also Ref 74).

Block Diagram

Fig 17.43 is a block diagram of the receiver. Here are some important facets of the design.

1. The antenna should have sufficient gain. At 10 GHz, gains of 30 dBi are not difficult to get, as the example of Fig 17.42 demonstrates. A 4-ft dish might be difficult to aim, however.
2. For best results a very low-noise preamp at the antenna reduces loss of system sensitivity when antenna temperature is low. For example, if the antenna temperature at a quiet direction of the sky is 50 K and the receiver noise figure is 4 dB (due in part to transmission-line loss), the system temperature is 488 K for a system noise figure of 9.9 dB. If the receiver noise figure is reduced to 1.5 dB by adding a preamp at the antenna the system temperature is reduced to 170 K for a system noise figure of 3.4 dB, which is a very big improvement.

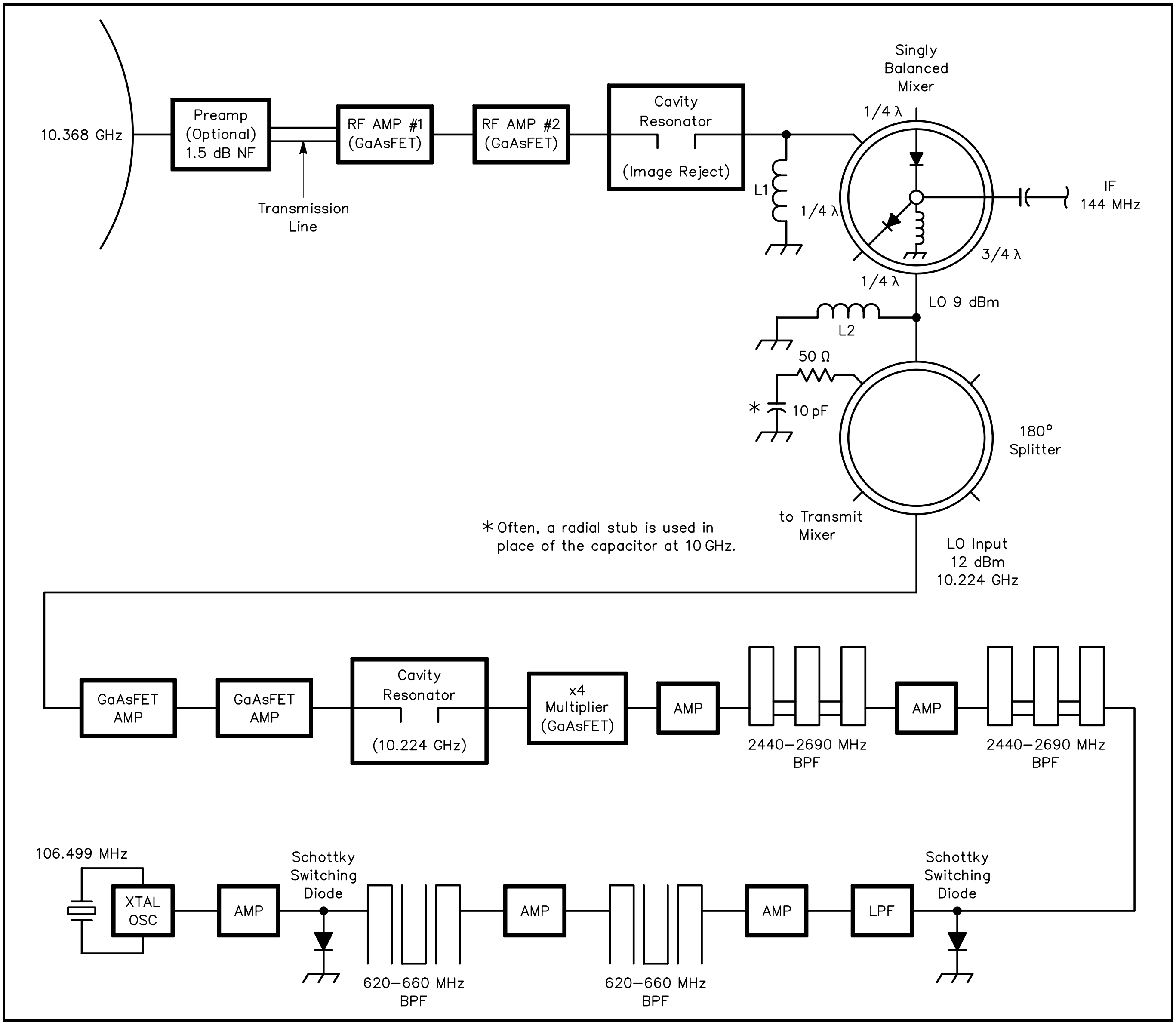


Fig 17.43—A block diagram of the microwave receiver discussed in the text.

3. After two stages of RF amplification using GaAsFETs, a probe-coupled cavity resonator attenuates noise at the mixer's image frequency, which is 10.368-0.288 =10.080 GHz. An image reduction of 15 to 20 dB is enough to prevent image frequency noise generated by the RF amplifiers from affecting the mixer's noise figure.
4. The singly balanced diode mixer uses a "rat-race" 180° hybrid. Each terminal of the ring is $^1/_4$ wavelength (90°) from its closest neighbors. So the anodes of the two diodes are 180° ($^1/_2$ wavelength) apart with respect to the LO port, but in-phase with respect to the RF port. The inductors (L1, L2) connected to ground present a low impedance at the IF frequency. The mixer microstrip circuit is carefully "tweaked" to improve system performance. Use the better mixer in the transmitter.
5. The crystal oscillator is a fifth-overtone Butler circuit that is capable of high stability. The crystal frequency error and drift are multiplied 96 times (10.224 / 0.1065), so for narrowband SSB or CW work it may be difficult to get on (and stay on) the "calling frequency" at 10.368 GHz. One acceptable (not perfect) solution might be to count the 106.5 MHz with a frequency counter whose internal clock is constantly compared with WWV. Adjust the 106.5 as required. At times there may be a small

Doppler shift on the WWV signal. It may be necessary to switch to a different WWV frequency, or WWV's signals may not be strong enough. Surplus frequency standards of high quality are sometimes available. Many operators just "tune" over the expected range of uncertainty.

6. The frequency multiplier chain has numerous band-pass filters to "purify" the harmonics by reducing various frequency components that might affect the signal path and cause spurious responses. The final filter is a tuned cavity resonator that reduces spurs from previous stages. Oscillator phase noise amplitude is multiplied by 96.0 also, so the oscillator must have very good short term stability to prevent contamination of the desired signal.
7. A second hybrid splitter provides an LO output for the transmitter section of the radio. The 50-Ω resistor improves isolation between the two output ports.

The two-part *QST* article (Ref 73) is recommended reading for this very interesting project, which provides a fairly straightforward (but not extremely simple) way to get started on 10 GHz.

Transmitter Design

TRANSMITTER DESIGN VS RECEIVER DESIGN

Many of the building blocks used in transmitter design are either identical to or very similar to those used in receiver design. Such things as mixers, oscillators, low-level RF/IF/AF amplifiers are the same. There is one major difference in the usage of these items, though. In a transmitter, the ratio of maximum to minimum signal levels for each of these is much less than in a receiver, where a very large ratio exists routinely. In a transmitter the signal, as it is developed to its final frequency and power level, is carefully controlled at each stage so that the stage is driven close to some optimum upper limit. The noise figures and dynamic ranges of the various stages are somewhat important, but not as important as in a receiver.

The transmitter design is concerned with the development of the desired high level of output power as cleanly, efficiently and economically as possible. Spurious outputs that create interference are a major concern. Protection circuitry that prevents self-destruction in the event of parts failures or mishandling by the operator help the reliability in ways that are unimportant in receivers.

THE SUPERHET SSB/CW TRANSMITTER

The same mixing schemes, IF frequencies and IF filters that are used for superhet receivers can be, and very often are, used for a transmitter. **Fig 17.44** is a block diagram of one approach. Let's discuss the various elements in detail, starting at the microphone.

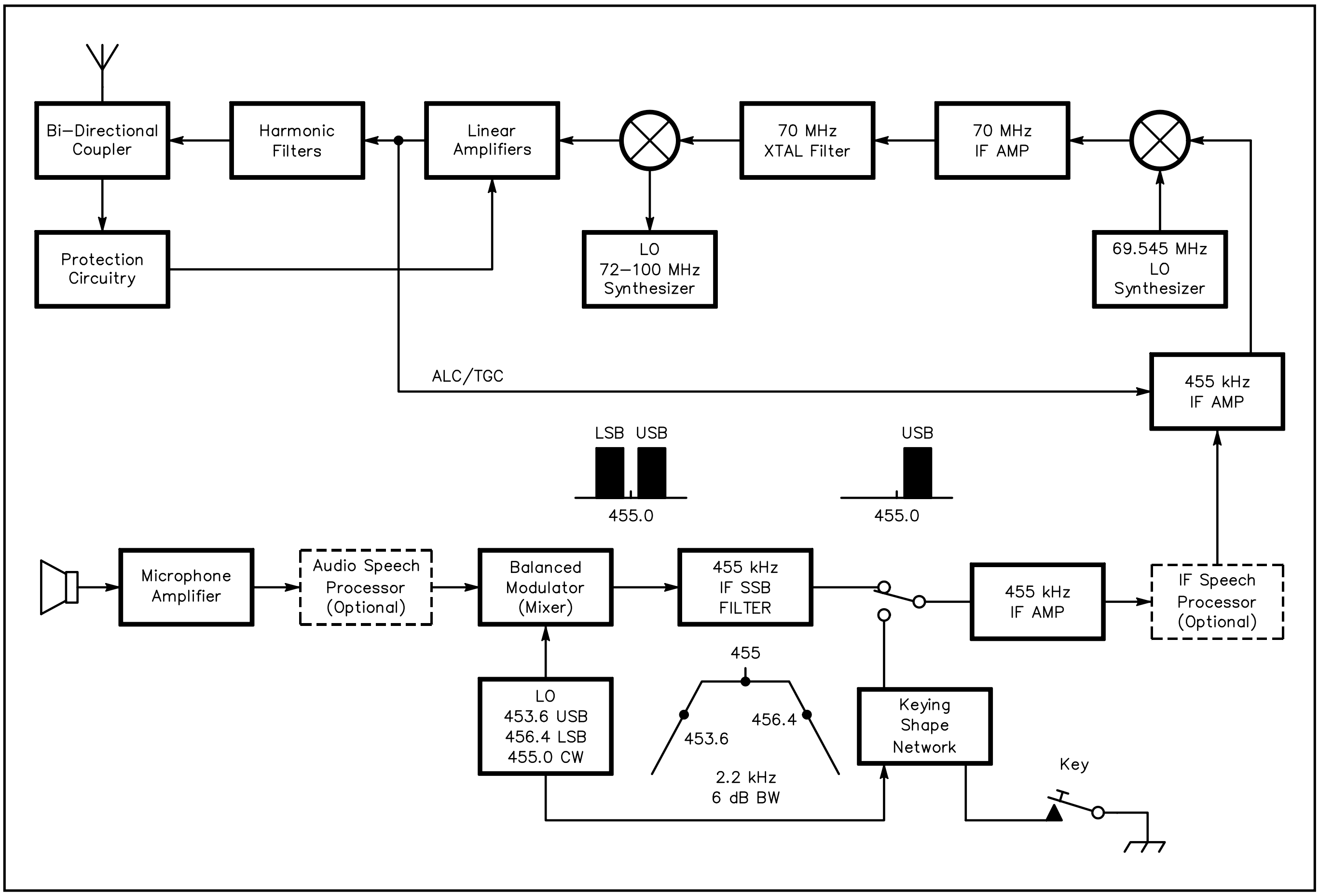


Fig 17.44—Block diagram of an up conversion SSB/CW transmitter. The various functions are discussed in the text.

Microphones (Mics)

A microphone is a transducer that converts sound waves into electrical signals. For speech, its frequency response should be as flat as possible from below 200 to above 3500 Hz. Response peaks in the microphone can increase the peak to average ratio of speech, which then degrades (increases) the peak to average ratio of the transmitted signal. If a transmitter uses speech processing, most microphones pick up a lot of background ambient noises because the speech amplification, whether it be at audio or IF/RF, may be as much as 20 dB greater than without speech processing. A "noise canceling" microphone is recommended to reduce this background pickup (Ref 22, 76). Microphone output levels vary, depending on the microphone type. Typical amateur mics produce about 10 to 100 mV.

Ceramic

Ceramic mics have high output impedances but low level outputs. They require a high-resistance load (usually about 50 kΩ) for flat frequency response and lose low-frequency response as this resistance is reduced (electrically, the mic "looks like" a small capacitor). These mics vary widely in quality, so a "cheap" mic is not a good bargain because of its effect on the transmitted power level and speech quality.

Dynamic

A dynamic mic resembles a small loudspeaker, with an impedance of about 680 Ω and an output of about 12 mV on voice peaks. In many cases a transformer (possibly built-in) transforms the impedance to 100 kΩ or more and delivers about 100 mV on voice peaks. Dynamic mics are widely used by amateurs.

Electret

"Electret" mics use a piece of special insulator material that contains a "trapped" polarization charge (Q) at its surfaces and a capacitance (C). Sound waves modulate the capacitance (dC) of the material and cause a voltage change (dV) according to the law dV/V = –dC/C. For small changes (dC) the change (dV) is almost linear. A polarizing voltage of about 4 V is required to maintain the charge. The mic output level is fairly low, and a preamp is sometimes required. These mics have been greatly improved in recent years.

Microphone Amplifiers

The balanced modulator and (or) the audio speech processor need a certain optimum level, which can be in the range of 0.3 to 0.6 V ac into perhaps 1 kΩ to 10 kΩ. Excess noise generated within the microphone amplifier should be minimized, especially if speech processing is used. The circuit in **Fig 17.45** uses a low-noise BiFET op amp. The 620-Ω resistor is se-

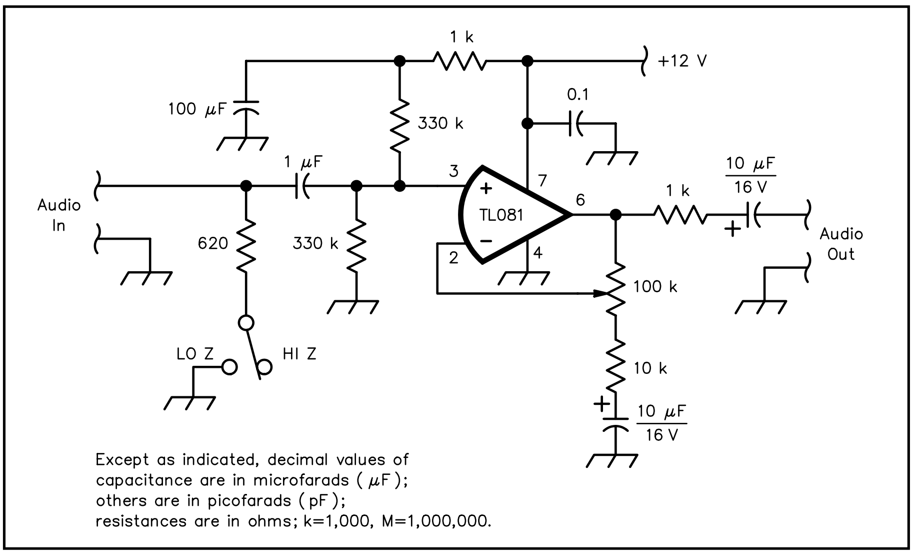


Fig 17.45—Schematic diagram of a simple op-amp microphone amplifier for low- and high-impedance microphones.

lected for a low impedance microphone, and switched out of the circuit for high-impedance mics. The amplifier gain is set by the 100-kΩ potentiometer.

It is also a good idea to experiment with the low-and high-frequency responses of the mic amplifier to compensate for the frequency response of the mic and the voice of the operator (Ref 77).

Audio Speech Clipping

If the audio signal from the microphone amplifier is further amplified, say by as much as 12 dB, and then if the peaks are clipped (sometimes called “slicing”) by 12 dB by a speech clipper, the output peak value is the same as before the clipper, but the average value is increased considerably. The resulting signal contains harmonics and IMD but the speech intelligibility, especially in a white-noise background, is improved by 5 or 6 dB.

The clipped waveform frequently tends to have a square-wave appearance, especially on voice peaks. It is then band-pass filtered to remove frequencies below 300 and above 3000 Hz. The filtering of this signal can create a “repeaking” effect. That is, the peak value tends to increase noticeably above its clipped value.

An SSB generator responds poorly to a square-wave audio signal. The Hilbert Transform effect, well known in mathematics, creates significant peaks in the RF envelope. These peaks cause out-of-band splatter in the linear amplifiers unless Automatic Level Control (ALC, to be discussed later) cuts back on the RF gain. The peaks increase the peak-to-average ratio and the ALC reduces the average SSB power output, thereby reducing some of the benefit of the speech processing. The square-wave effect is also reduced by band-pass filtering (300 to 3000 Hz) the input to the clipper as well as the output.

Fig 17.46 is a circuit for a simple audio speech clipper. A CLIP LEVEL potentiometer before the clipper controls the amount of clipping and an OUTPUT LEVEL potentiometer controls the drive level to the balanced modulator. The correct adjustment of these potentiometers is done with a two-tone audio input or by talking into the microphone, rather than a single tone, because single tones don’t exhibit the repeaking effect (Ref 48).

Audio Speech Compression

Although it is desirable to keep the voice level as high as possible, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use an automatic gain control that follows the average variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the resultant dc to a control terminal in an early stage of the amplifier.

The circuit of **Fig 17.47A** works on this AGC principle. One section of a Signetics 571N IC is used. The other section can be connected as an

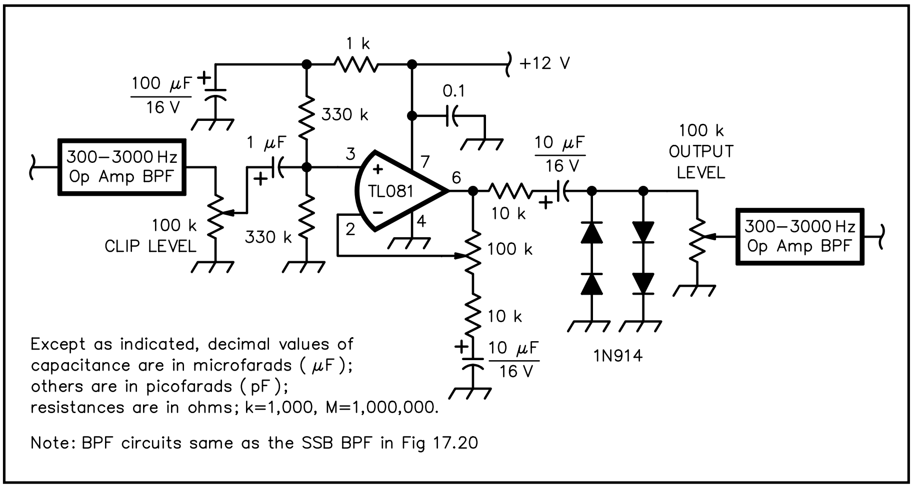


Fig 17.46—A simple audio speech clipper. The input signal is band pass filtered, amplified by 20 dB, clipped and band pass filtered again.

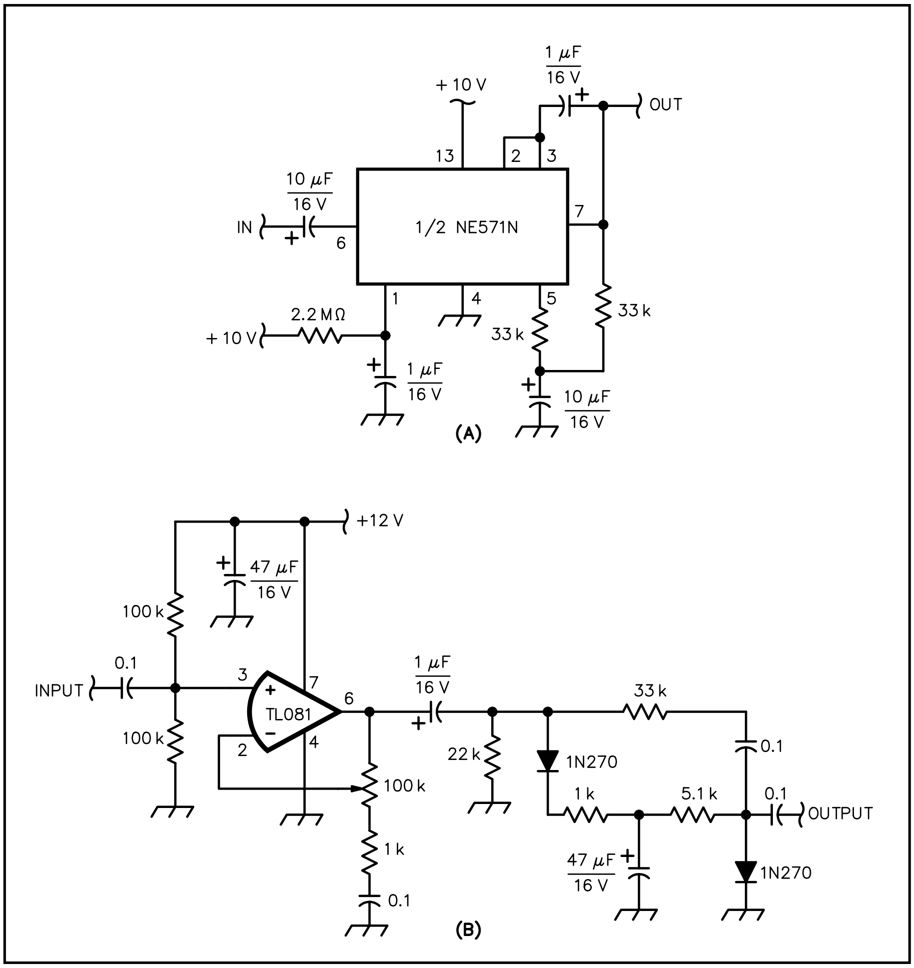


Fig 17.47—Typical solid-state compressor circuits. The circuit at A works on the AGC principle, while that at B is a forward-acting compressor.

expander to restore the dynamic range of received signals that have been compressed in transmission. Operational transconductance amplifiers such as the CA3080 are also well suited for speech compression.

When an audio AGC circuit derives control voltage from the output signal the system is a closed loop. If short attack time is necessary, the rectifier-filter bandwidth must be opened up to allow syllabic modulation of the control voltage. This allows some of the voice frequency signal to enter the control terminal, causing distortion and instability. Because the syllabic frequency and speech-tone frequencies have relatively small separation, the simpler feedback AGC systems compromise fidelity for fast response.

Problems with loop dynamics in audio AGC can be sidestepped by eliminating the loop and using a forward-acting system. The control voltage is derived from the input of the amplifier, rather than from the output. Eliminating the feedback loop allows unconditional stability, but the trade-off between response time and fidelity remains. Care must be taken to avoid excessive gain between the signal input and the control voltage output. Otherwise the transfer characteristic can reverse; that is, an increase in input level can cause a decrease in output. A simple forward-acting compressor is shown in Fig 17.47B.

An additional interesting method of speech compression involves the technology of “Homomorphic Signal Processing.” This subject is covered prominently in digital signal processing textbooks (Ref 78). An excellent example of an analog instrument that performs this function is described in Ref 79.

Balanced Modulators

A balanced modulator is a mixer. A more complete discussion of balanced modulator design is provided in the **Mixers** chapter. Briefly, the IF frequency LO (455 kHz in the example of Fig 17.44) translates the audio frequencies up to a pair of IF frequencies, the LO plus the audio frequency and the LO minus the audio frequency. The balance from the LO port to the IF output causes the LO frequency to be suppressed by 30 to 40 dB. Adjustments are provided to improve the LO null.

The filter method of SSB generation uses an IF band-pass filter to pass one of the sidebands and block the other. In Fig 17.44 the filter is centered at 455.0 kHz. The LO is offset to 453.6 kHz or 456.4 kHz so that the upper sideband or the lower sideband (respectively) can pass through the filter. This creates a problem for the other LOs in the radio, because they must now be properly offset so that the final transmit output's carrier (suppressed) frequency coincides with the frequency readout on the front panel of the radio. Various schemes have been used to do this. One method uses two crystals for the 69.545-MHz LO that can be selected. In synthesized radios the programming of the microprocessor control moves the various LOs. Some synthesized radios use two IF filters at two different frequencies, one for USB and one for LSB, and a 455.0-kHz LO. These radios are often designed to transmit two independent sidebands (ISB, Ref 48).

In times past, balanced modulators using diodes, balancing potentiometers and numerous components spread out on a PC board were universally used. These days it doesn't make sense to use this approach. ICs and packaged diode mixers do a much better job and are less expensive. The most famous modulator IC, the MC1496, has been around for more than 15 years and is still one of the best and least expensive (Ref 80). **Fig 17.48** is a typical balanced modulator circuit using the MC1496.

The data sheets for balanced modulators and mixers specify the maximum level of audio for a given LO level. Higher audio levels create excessive IMD. The IF filter after the modulator removes higher-order IMD products that are outside its passband but the in-band IMD products should be at least 40 dB below each of two equal test tones. Speech clipping (AF or IF) can degrade this to 10 dB or so, but in the absence of speech processing the signal should be clean, in-band.

IF Filters

The desired IF filter response is shown in **Fig 17.49A**. The reduction of the carrier frequency is augmented by the filter response. It is common to specify that the filter response be down 20 dB at the carrier frequency. Rejection of the opposite sideband should (hopefully) be 60 dB, starting at 300 Hz below the carrier frequency, which is the 300-Hz point on the opposite sideband. The ultimate attenuation should be at least 70 dB. This would represent a very good specification for a high quality transmitter. The filter passband should be as flat as possible (ripple less than 1 dB or so).

Special filters, designated as USB or LSB, are designed with a steeper rolloff on the carrier frequency side, in order to improve rejection of the carrier and opposite sideband. Mechanical filters are available that do this. Crystal-ladder filters (see the **Filters** chapter) are called "single-sideband" filters because they also have this property. The steep skirt can be on the low side or the high side, depending on whether the crystals are across the signal path or in series with the signal path, respectively.

Filters require special attention to their terminations. The networks that interface the filter with surrounding circuits should be accurate and stable over temperature. They should be easy to adjust. One very good way to adjust them is to build a narrowband sweep generator and look at the output IF envelope with a logarithmic amplifier, as indicated in Fig 17.49B. There are three goals: The driver stage must see the desired load impedance: the stage after the filter must see the desired source

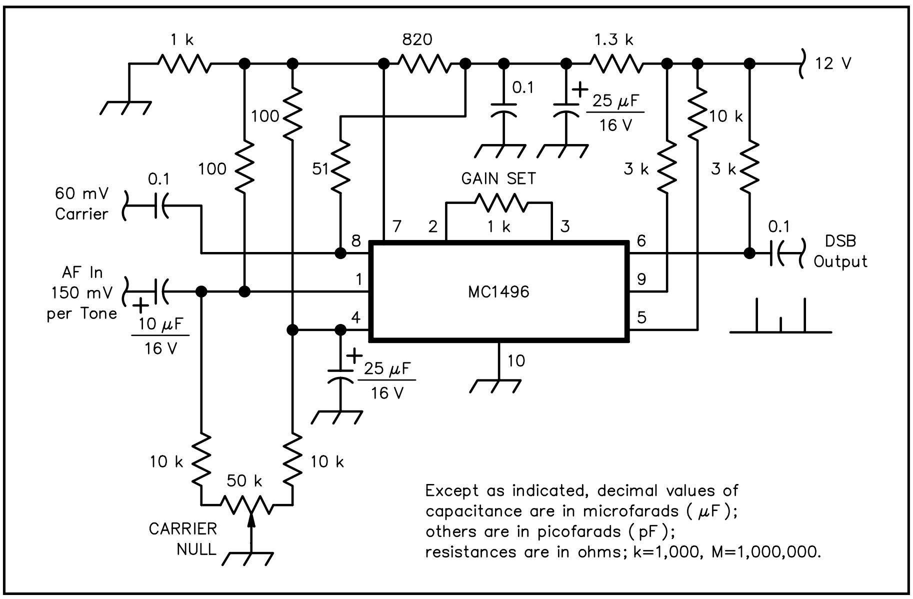


Fig 17.48—An IC balanced modulator circuit using the MC1496. The resistor from pin 2 to pin 3 sets the conversion gain.

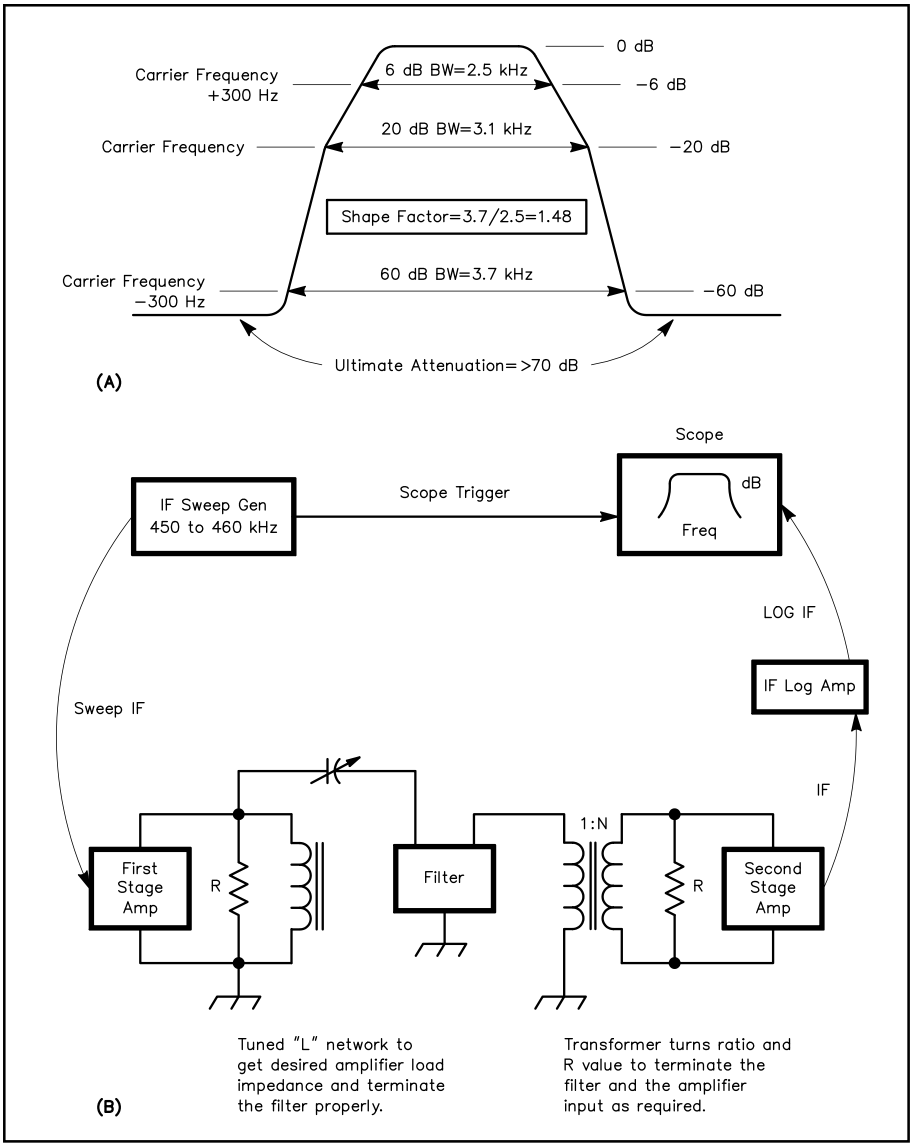


Fig 17.49—A: desired response of an SSB IF filter. B: one method of terminating a mechanical filter that allows easy and accurate tuning adjustment and also a possible test setup for performing the adjustments.

(generator) impedance and the filter must be properly terminated at both ends. Fig 17.49B shows two typical approaches (Ref 42). This kind of setup is a very good way to make sure the filters and other circuitry are working properly.

Finally, overdriven filters (such as crystal or mechanical filters) can become nonlinear and generate distortion. So it is necessary to heed the manufacturer's instructions. Magnetic core materials used in the tuning networks must be sufficiently linear at the signal levels encountered. They should be tested for IMD separately.

IF Speech Clipper

Audio-clipper speech processors generate a considerable amount of in-band harmonics and IMD (involving different simultaneously occurring speech frequencies). The total distortion detracts somewhat from speech intelligibility. Other problems were mentioned in the section on audio processing. IF clippers overcome most of these problems, especially the Hilbert Transform problem (Refs 48, 77).

Fig 17.50A is a diagram of a 455-kHz IF clipper using high-frequency op amps. 20 dB of gain precedes the diode clippers. A second amplifier establishes the desired output level. The clipping produces a wide band of IMD products close to the IF frequency. Harmonics of the IF frequency are generated that are easily rejected by subsequent selectivity. "Close-in" IMD distortion products are band limited by the 2.5-kHz-wide IF filter so that out-of-band splatter is eliminated. The in-band IMD products are at least 10 dB below the speech tones.

Fig 17.51 shows oscilloscope pictures of an IF clipped two-tone signal at various levels of clipping. The level of clipping in a radio can be estimated by comparing with these photos.

Listening tests verify that the IMD does not sound nearly as bad as harmonic distortion. In fact, processed speech sounds relatively clean and crisp. Tests also verify that speech intelligibility in a noise background is improved by 8 dB.

The repeaking effect from band-pass filtering the clipped IF signal occurs, and must be accounted for when adjusting the output level. A two-tone audio test signal or a speech signal should be used. The ALC

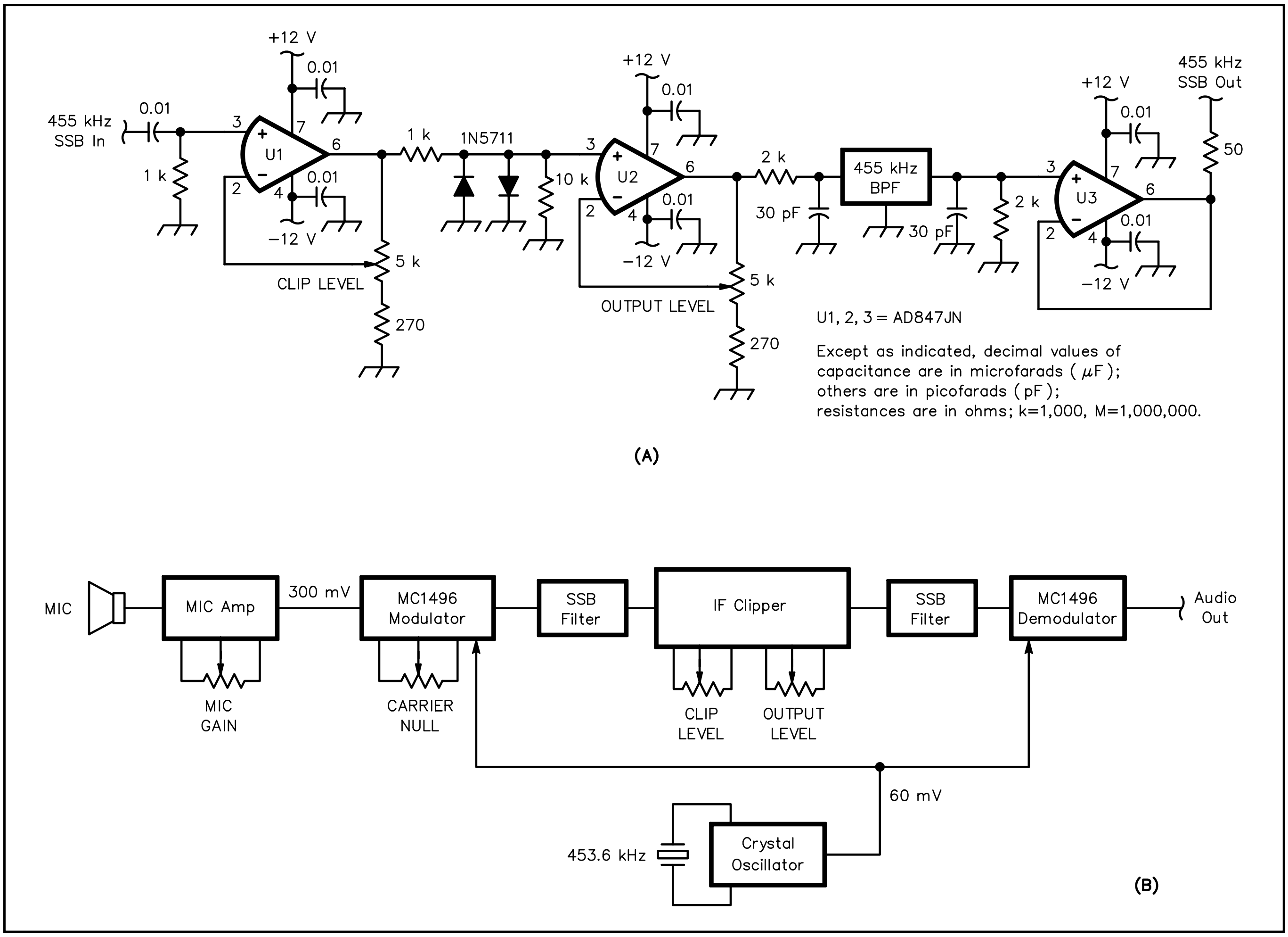


Fig 17.50—IF speech clipping. A: an IF clipper circuit approach. B: the audio signal is translated to 455 kHz, processed, and translated back to audio.

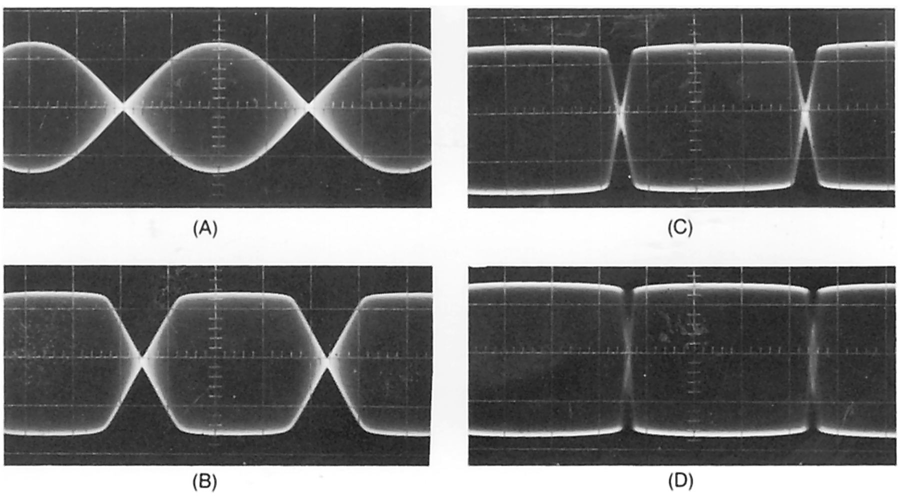


Fig 17.51—Two-tone envelope patterns with various degrees of RF clipping. All envelope patterns are formed using tones of 600 and 1000 Hz. At A, clipping threshold; B, 5 dB of clipping; C, 10 dB of clipping; D, 15 dB of clipping (from “RF Clippers for SSB,” by W. Sabin, July 1967 *QST*, pp 13-18).

circuitry (discussed later) will cut back the IF gain to prevent splattering in the power amplifiers. If the IF filter is of high quality and if subsequent amplifiers are "clean," the transmitted signal is of very high quality, very effective in noisy situations and often also in "pile-ups."

The extra IF gain implies that the IF signal entering the clipper must be free of noise, hum and spurious products. The cleanup filter also helps reduce the carrier frequency, which is outside the passband.

An electrically identical approach to the IF clipper can be achieved at audio frequencies. If the audio signal is translated to, say 455 kHz, processed as described, and translated back to audio, all the desirable effects of IF clipping are retained. This output then plugs into the transmitter's microphone jack. Fig 17.50B shows the basic method. The mic amplifier and the MC1496 circuits have been previously shown and the clipper circuit is the same as in Fig 17.50A.

Another method, performed at audio, synthesizes mathematically the function of the IF clipper. This method is mentioned in Ref 48, and was an accessory for the Collins KWM380 transceiver.

The interesting operating principle in all of these examples is that the characteristics of the IF-clipped (or equivalent) speech signal do not change during frequency translation, even when translated down to audio and then back up to IF in a balanced modulator.

IF Linearity and Noise

Fig 17.44 indicates that after the last SSB filter, whether it is just after the SSB modulator or after the IF clipper, subsequent BPFs are considerably wider. For example, the 70-MHz crystal filter may be 15 to 30 kHz wide. This means that there is a "gray region" in the transmitter just like the one that we saw in the up conversion receiver, where out-of-band IMD that is generated in the IF amplifiers and mixers can cause adjacent-channel interference.

A possible exception, not shown in Fig 17.44, is that there may be an intermediate IF in the 10-MHz region that also contains a narrow filter, as we saw in the triple-conversion receiver in Fig 17.36.

The implication is that special attention must be paid to the linearity of these circuits. It's the designer's job to make sure that distortion in this gray area is much less than distortion generated by the PA and also less than the phase noise generated by the final mixer. Recall also that the total IMD generated in the exciter stages is the resultant of several amplifier and mixer stages in cascade, therefore each element in the chain must have at least 40 to 50 dB IMD quality. The various drive levels should be chosen to guarantee this. This requirement for multistage linearity is one of the main technical and cost burdens of the SSB mode.

Of interest also in the gray region are white, additive thermal and excess noises originating in the first IF amplifier after the SSB filter and highly magnified on their way to the output. This noise can be comparable to the phase noise level if the phase noise is low, as it would be in a high-quality radio. Recall also that phase noise is at its worst on modulation peaks, but additive noise may be (and often is) present even when there is no modulation. This is a frequent problem in colocated transmitting and receiving environments. Many transmitter designs do not have the benefit of the narrow filter at 70 MHz, so the amplified noise can extend over a much wider frequency range.

CW Mode

Fig 17.44 shows that in the CW mode a carrier is generated at the center of the SSB filter passband. There are two ways to make this carrier available. One way is to unbalance the balanced modulator so that the LO can pass through. Each kind of balanced modulator circuit has its own method of doing this. The approach chosen in Fig 17.44 is to go around the modulator and the SSB filter.

A shaping network controls the envelope of the IF signal to accomplish two things: control the shape of the Morse code character in a way that limits wideband spectrum emissions that can cause interference, and makes the Morse code signal easy and pleasant to copy.

RF Envelope Shaping

On-off keying is a special kind of low-level amplitude modulation (a low-signal-level stage is turned on and off). It generates numerous sidebands around the carrier frequency whose amplitudes are influenced by the shapes of the RF envelope rise and fall intervals. Consider an unprocessed keying waveform, a string of equal-length rectangular RF pulses separated by equal-length "dead times." Its spectrum consists of the carrier frequency, a pair of the usual modulation sidebands at the carrier frequency plus and minus the repetition rate of the pulses, and numerous other sidebands at multiples of the repetition rate. The higher frequency sidebands can create "key clicks" that extend many kHz either side of the carrier frequency.

Adjusting an exponentially shaped rise and fall time to about 5 ms (a value recommended by ARRL) controls key clicks, yet allows a wide range of practical keying speeds.

Fig 17.52A is a computer simulation of a single dot-space that is part of a continuous sequence, 20 ms on and 20 ms off (a 60 wpm rate), that has been shaped in this manner. This fast dot sequence is probably a "worst case" as far as Morse code interference is concerned. Fig 17.52B shows the line spectrum (Discrete Fourier Transform, DFT) of this periodic sequence. The carrier component is reduced because the on-off duty cycle is 50%, and some power is transferred from the carrier to the sidebands. (Note: This is a little different from ordinary AM, in which an additional source of power is used to create sidebands and the carrier remains constant.) The first pair of spectrum lines are ±25 Hz from the carrier. These are the normal modulation sidebands, determined only by the keying rate (25 Hz = 1/(40 ms)). The subsequent pairs are odd-order sidebands starting at ±75 Hz from the carrier (the symmetry of the on-off reduces even-order sidebands).

If the duration of the dot is held constant but the time between dots is greatly increased (the keying rate is greatly reduced) many more lines, much closer together and now including the even-order (50, 100, 150 Hz, and so on) sidebands, would appear in Fig 17.52B. But the general shape and also the bandwidth of the spectrum remain the same as shown. For example, a single, unrepeated dot would have essentially the same spectrum shape and width as in Fig 17.52B but would be a continuous spectrum (the Fourier Transform, FT) instead of the line spectrum (DFT) shown. In all cases the bandwidth of the interference, measured at the –20 dB points, is approximately 150 Hz. Finally, a single "key-down-and-hold" or "key-up-and-hold" operation would produce a similar spectrum of brief duration (a single key click) due to the shaping of the leading or trailing edge.

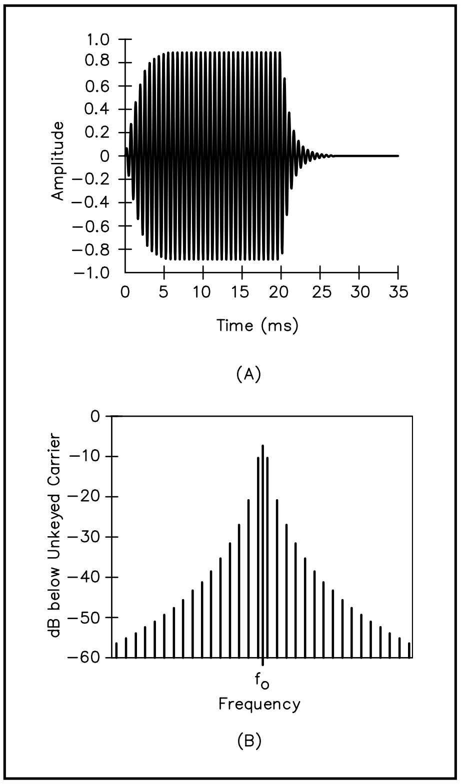


Fig 17.52—A: envelope, and B: spectrum of a 20-ms dot that has been shaped to 5 ms rise and fall times.

The rolloff rate of the higher sidebands in the shaped pulse is much faster than for a rectangular pulse. The example shown could be improved further by making the transitions "sinusoidal." The higher sidebands are still present but they are further reduced, as verified by computer simulations. This waveform (or other kinds of "windowing") could be generated with DSP or with an op-amp state-variable analog computer in the Key Shape block of Fig 17.44. The linearity of the SSB amplifier chain will then preserve this envelope shape right up to the antenna.

Keying Circuits

The circuit in **Fig 17.53** is one example of an approach that will

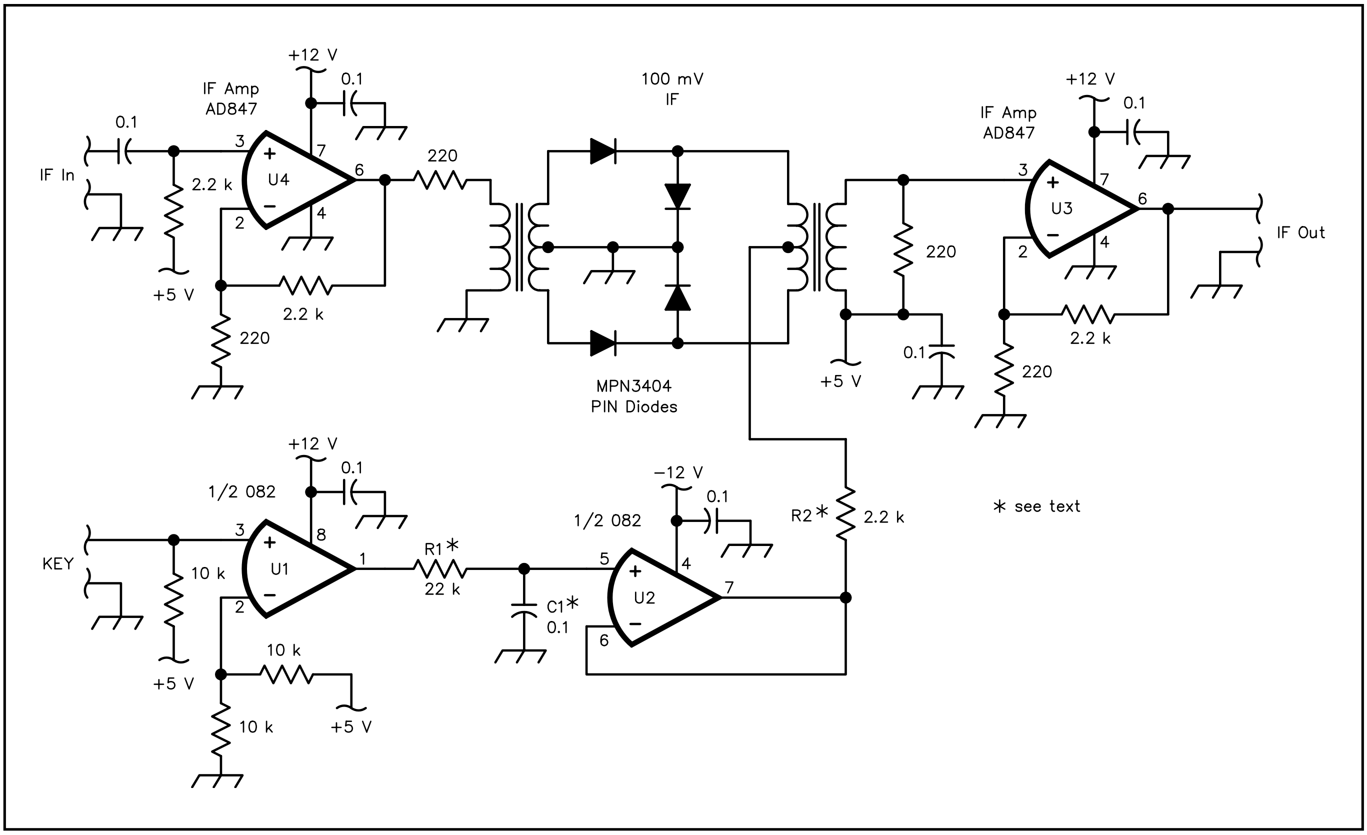


Fig 17.53—A keying circuit for the transmitter diagram in Fig 17.44.

provide accurate control of the IF envelope. The PIN diodes are connected either to the signal source or to ground. They are turned on or off through U2 according to the exponential charging or discharging of the capacitor C1 through R1. R1 connects to +12 V or –12 V through U1. Some experimentation with R1 and R2 will be necessary, depending on the IF drive level, to get the desired rise and fall characteristics. Look at the IF envelope on an oscilloscope while sending a string of dots. Try an input IF level of 100 mV. The transformers are optimized for the IF in use.

Another approach is to use gain controllable IC amplifiers of the type discussed in connection with receiver AGC. See, for example, Fig 17.31. The key-up "turnoff" must be complete to avoid "backwave," residual output.

Wideband Noise

In the block diagram of Fig 17.44 the last mixer and the amplifiers after it are wideband circuits that are limited only by the harmonic filters and by any selectivity that may be in the antenna system. Wide band phase noise transferred onto the transmitted modulation by the last LO can extend over a wide frequency range, therefore LO (almost always a synthesizer of some kind) cleanliness is always a matter of great concern (Ref 23).

The amplifiers after this mixer are also sources of wide-band “white” or additive noise. This noise can be transmitted even during times when there is no modulation, and it can be a source of local interference. To reduce this noise: use a high-level mixer with as much signal output as possible, and make the noise figure of the first amplifier stage after the mixer as low as possible.

Transmitters that are used in close proximity to receivers, such as on shipboard, are always designed to control wideband emissions of both additive noise and phase noise, referred to as “composite” noise.

Transmit Mixer Spurious Signals

The last IF and the last mixer LO in Fig 17.44 are selected so that, as much as possible, harmonic IMD products are far enough away from the operating frequency that they fall outside the passband of the low-pass filters and are highly attenuated. This is difficult to accomplish over the transmitter's entire frequency range. It helps to use a high-level mixer and a low enough signal level to minimize those products that are unavoidable. Low-order crossovers that cannot be sufficiently reduced are unacceptable, however; the designer must "go back to the drawing board."

The **Regulations** chapter gives information regarding FCC requirements for harmonic and spurious levels. The **Filters** chapter describes the design of harmonic filters.

Automatic Level Control (ALC)

The purpose of ALC is to prevent the various stages in the transmitter from being overdriven. Overdrive can generate too much out-of-band distortion or cause excessive power dissipation, either in the amplifiers or in the power supply. ALC does this by sampling the peak amplitude of the modulation (the envelope variations) of the output signal and then developing a dc gain-control voltage that is applied to an early amplifier stage, as suggested in Fig 17.44.

ALC is usually derived from the last stage in a transmitter. This ensures that this last stage will be protected from overload. However, other stages prior to the last stage may not be as well protected; they may generate excessive distortion. It is possible to derive a composite ALC from more than one stage in a way that would prevent this problem. But designers usually prefer to design earlier stages conservatively enough so that, given a temperature range and component tolerances, the last stage can be the one source of ALC. The gain control is applied to an early stage so that all stages are aided by the gain reduction.

Speech Processing with ALC

A fast response to the leading edge of the modulation is needed to prevent a transient overload. After a strong peak, the control voltage is "remembered" for some time as the voltage in a capacitor. This voltage then decays partially through a resistor between peaks. An effective practice provides two capacitors and two time constants. One capacitor decays quickly with a time constant of, say 100 ms, the other with a time constant of several seconds. With this arrangement a small amount of speech processing, about 1 or 2 dB, can be obtained. (Explanation: The dB of improvement mentioned has to do with the improvement in speech intelligibility in a random noise background. This improvement is equivalent to what could be achieved if the transmit power were increased that same number of dB). The gain rises a little between peaks so that weaker speech components are enhanced. But immediately after a peak it takes awhile for the enhancement to take place, so weak components right after a strong peak are not enhanced very much. **Fig 17.54A** shows a complete ALC circuit that performs speech processing.

ALC in Solid-State Power Amplifiers

Fig 17.54B shows how a dual directional coupler can be used to provide ALC for a solid-state power amplifier (PA). The basic idea is to protect the PA transistors from excessive SWR and dissipation by monitoring both the forward power and the reflected power. The projects section of this chapter includes a 50-W amplifier project; the protection circuitry is discussed in detail there. Thermal protection is also included.

Transmit Gain Control (TGC)

This is a widely used feature in commercial and military equipment. A calibrated "tune-up " test carrier of a certain known level is applied to the transmitter. The output carrier level is sampled, using a diode detector. The resulting dc voltage is used to set the gain of a low-level stage. This control voltage is digitized and stored in memory so that it is semipermanent. A new voltage may be generated and stored after each frequency change, or the stored value may be "fetched." A test signal is also used to do automatic antenna tuning. A dummy load is used to set the level and a low-level signal (a few mW) is used for the antenna tune-up.

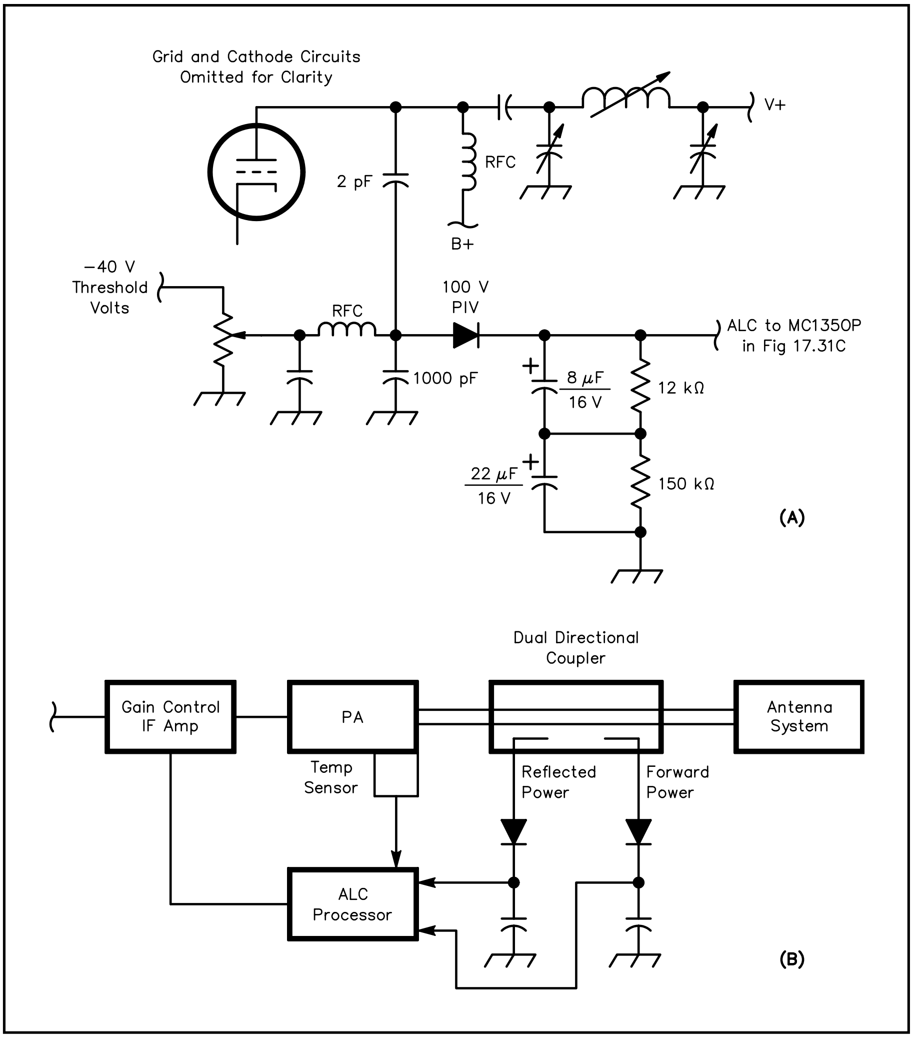


Fig 17.54—A: an ALC circuit with speech processing capability. B: protection method for a solid-state transmitter.

Transmitter Output Load Impedance

The following logical processes are used to tune and load the final PA of a transmitter:

1. The RF input power requirement to the input terminal of the PA has been determined. That power level is applied to the input.
2. The desired load impedance of the plate/collector/drain of the PA has been determined, either graphically or by calculation, from the power to be delivered to the load, the dc power supply voltage and the ac voltage on the plate/collector/drain (see **RF Power Amplifiers** chapter).
3. The input impedance, looking toward the antenna, of the transmission line which is connected to the transmitter is adjusted by a network of some kind to its Z_0 value (if it is not already equal to that value).
4. A network of some kind is designed, which transforms the transmission-line Z_0 to the impedance required in step 2. This may be a sharply tuned resonator with impedance transforming capability, or it may be a wideband transformer of some kind.

Under these conditions the PA is performing as intended. Note that a knowledge of the output impedance of the PA is not needed to get these results. That is, we are interested mostly in the actual power gain of the PA, which does not require a knowledge of the amplifier's output impedance.

The output impedance, looking backward from the plate/collector/drain terminal of the network, in step 4, will have some influence on the selectivity of a resonant tuned circuit or the frequency response of a low-pass filter. This must (or should) be considered during the design process, but it is not needed during the "tune and load" process (Ref 81).

A Classic Amateur Vacuum-Tube Transmitter

Cascaded stage transmitter designs became very popular after about 1930. Home-brew solid-state CW transmitters often use the same general idea today. **Fig 17.55** is a typical example. The first stage is an oscillator, which may be either crystal controlled or a variable frequency oscillator (VFO) of high stability. The oscillator is followed by a combination of buffer amplifiers and frequency multipliers, usually doublers, to arrive at the final frequency and power level. Mixer circuits and crystal-VFO "mix-master" frequency generator arrangements are not used, so spurious mixer products are not a problem. Each stage provides the power input that the next stage needs and each stage increases in size and power dissipation. The vacuum tube '47 crystal osc, '46 doubler, 210 ("five watter") buffer, 203A ("fifty watter") final (200 W input) was a famous CW "rig" during the mid 1930s. HF CW DXers would usually have one favorite crystal that was just several Hz inside the band edge.

The process of frequency multiplication also multiplies the frequency drift, frequency/phase modulation and microphonic effects of the VFO. So for many years amateurs relied on crystal control, until they learned how to design very stable, low-noise VFOs. The frequency multiplier is an item worthy of discussion.

Frequency Multipliers

A passive multiplier using diodes is shown in **Fig 17.56A**. The full-wave rectifier circuit can be recognized, except that the dc component is shorted to ground. If the fundamental frequency ac input is 1.0 V RMS the second harmonic is 0.42 V RMS or 8 dB below the input, including some small diode losses. This value is found by calculating the Fourier Series coefficients for the full-wave-rectified sine wave, as shown in many textbooks.

Transistor and vacuum-tube frequency multipliers operate on the following principle: if a sine wave input causes the plate/collector/drain current to be distorted (not a sine wave) then harmonics of the input are generated. If an output resonant circuit is tuned to a harmonic the output at the harmonic is emphasized and other frequencies are attenuated. For a particular harmonic the current pulse should be distorted in a way that maximizes that harmonic. For example, for a doubler the current pulse should look like a half sine wave (180° of conduction). A transistor with Class B bias would be a good choice. For a tripler use 120° of conduction (Class C).

An FET, biased at a certain point, is very nearly a "square law" device. That is, the drain-current change is proportional to the square of the gate-voltage change. It is then an efficient frequency doubler that also deemphasizes the fundamental.

A push-push doubler is shown in Fig 17.56B. The FETs are biased in the square-law region and the

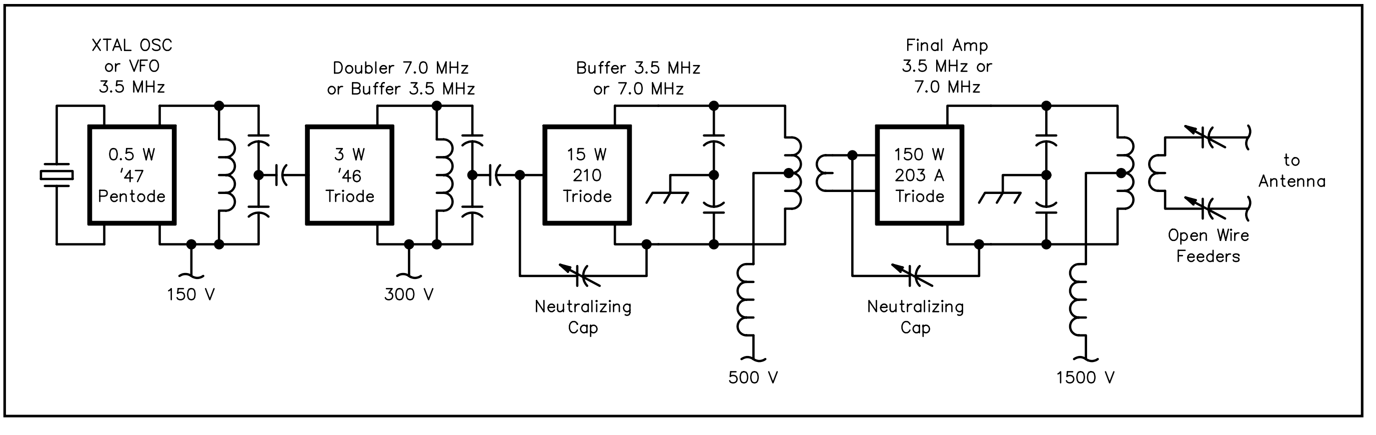


Fig 17.55—Example block diagram of MOPA transmitter, typical of the 1930s. Crystal control is shown, but VFOs replaced them during the 40s.

BALANCE potentiometer minimizes the fundamental frequency. Note that the gates are in push-pull and the drains are in parallel. This causes second harmonics to add in-phase at the output and fundamental components to cancel.

Fig 17.56C shows an example of a bipolar-transistor doubler. The efficiency of a doubler of this type is typically 50%, a tripler 33% and a quadrupler 25%. Harmonics other than the one to which the output tank is tuned will appear in the output unless effective band-pass filtering is applied. The collector tap on L1 is placed at the point that offers the best compromise between power output and spectral purity.

A push-pull tripler is shown in Fig 17.56D. The input and output are both push-pull. The balance potentiometer minimizes even harmonics. Note that the transistors have no bias voltage in the base circuit; this places the transistors in Class C for efficient third-harmonic production. Choose an input drive level that maximizes harmonic output.

The step recovery diode (SRD) is an excellent device for harmonic generation, especially at microwave frequencies (Ref 25). The basic idea of the SRD is as follows: When the diode is forward conducting, a charge is stored in the diode's diffusion capacitance; and if the diode is quickly reverse-biased, the stored charge is very suddenly released into an LC harmonic-tuned circuit.

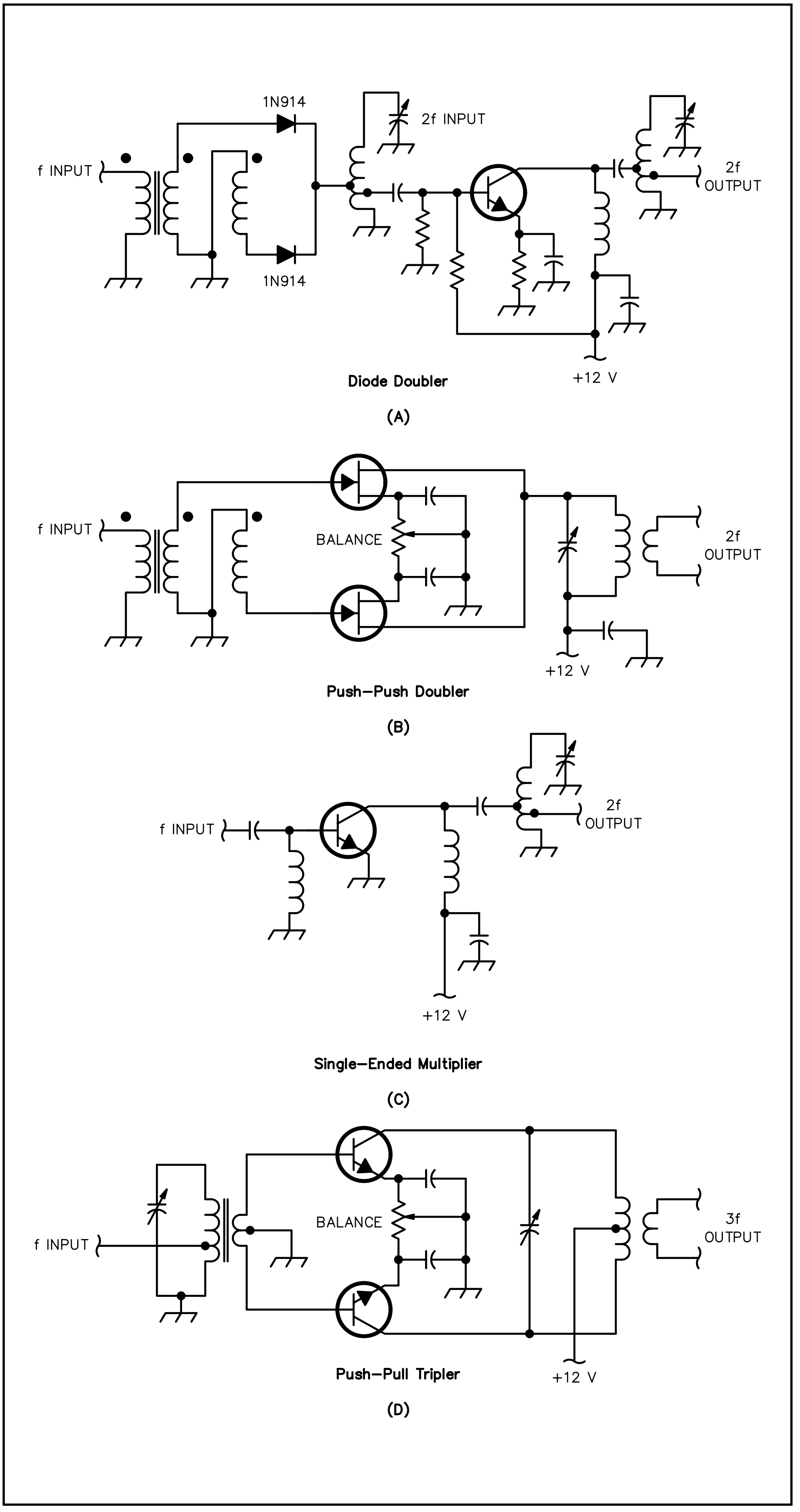


Fig 17.56—A: diode doubler. B: push-push doubler using JFETS. C: single-ended multiplier using a BJT. D: push-pull tripler using BJTs.

The circuit is also called a "comb generator" because of the large number of harmonics that are generated. (The spectral display looks like a comb.) Phase-locked loops (PLLs) can be made to lock onto these harmonics. A typical low-cost SRD is the HP 5082-0180, found in the HP Microwave & RF Designer's Catalog. **Fig 17.57A** is a typical schematic. For more information regarding design details there are two References: Hewlett-Packard application note AN-920 and Ref 25.

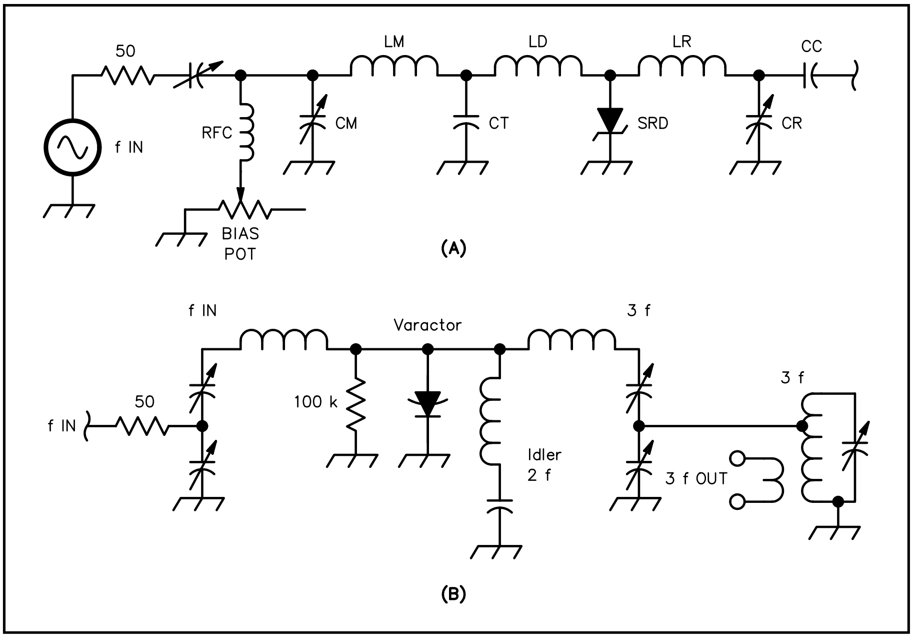


Fig 17.57—Diode frequency multipliers. A: step-recovery diode multiplier. B: varactor diode multiplier.

The varactor diode can also be used as a multiplier. Fig 17.57B shows an example. This circuit depends on the fact that the capacitance of a varactor changes with the instantaneous value of the RF excitation voltage. This is a nonlinear process that generates harmonic currents through the diode. Power levels up to 25 W can be generated in this manner.

Coupling Between Transmitter Stages

Most of the methods described here are also widely used in receiver circuits. Correct impedance matching between a stage and its load provides maximum transfer of power. The load can be an antenna or the next stage. The input impedance of the next stage is transformed to the value of load impedance that the preceding stage wants to "see" for proper operation. If the next stage is a bipolar transistor, determining the base input impedance can be difficult without adequate test equipment. Generally, when the drive power is 2 W or more the base impedance will be less than 10 Ω and often in the 1 to 2-Ω range. Under these conditions, some kinds of LC matching networks are difficult to adjust. It is desirable to use a network that can be easily adjusted over a wide range of impedances, where both L and C are variable. Sometimes a deliberate mismatch is preferred for stability or for accurate power-level control. This subject was discussed earlier in this chapter.

In the interest of stability, it is common practice to use low-Q networks between stages in a solid-state transmitter. Loaded Q values of five are common. The penalty for this is poor attenuation of harmonics and other spurious outputs. Conversely, vacuum-tube stages operate at relatively high-impedance levels in the grid and plate circuits. Loaded Q values of 15 are used. The rejection of spurs is greatly improved. In all cases the matching network is used to "absorb" circuit and tube or transistor reactances into the network, so that pure resistive impedance values can be achieved. The impedance values depend on frequency and power level, and tend to be difficult to predict. So experimental methods, using wide-range networks, are usually preferred.

The interstage coupling method in **Fig 17.58A** is a common one for vacuum tubes. A low-value coupling capacitor goes to the grid. RFC1 and RFC2 have a very high impedance over the operating frequency range.

Band-pass coupling is shown in Fig 17.58B. C1 is selected to give a flat response over a frequency range, such as a ham band, and high attenuation outside the band. This response can be changed, though,

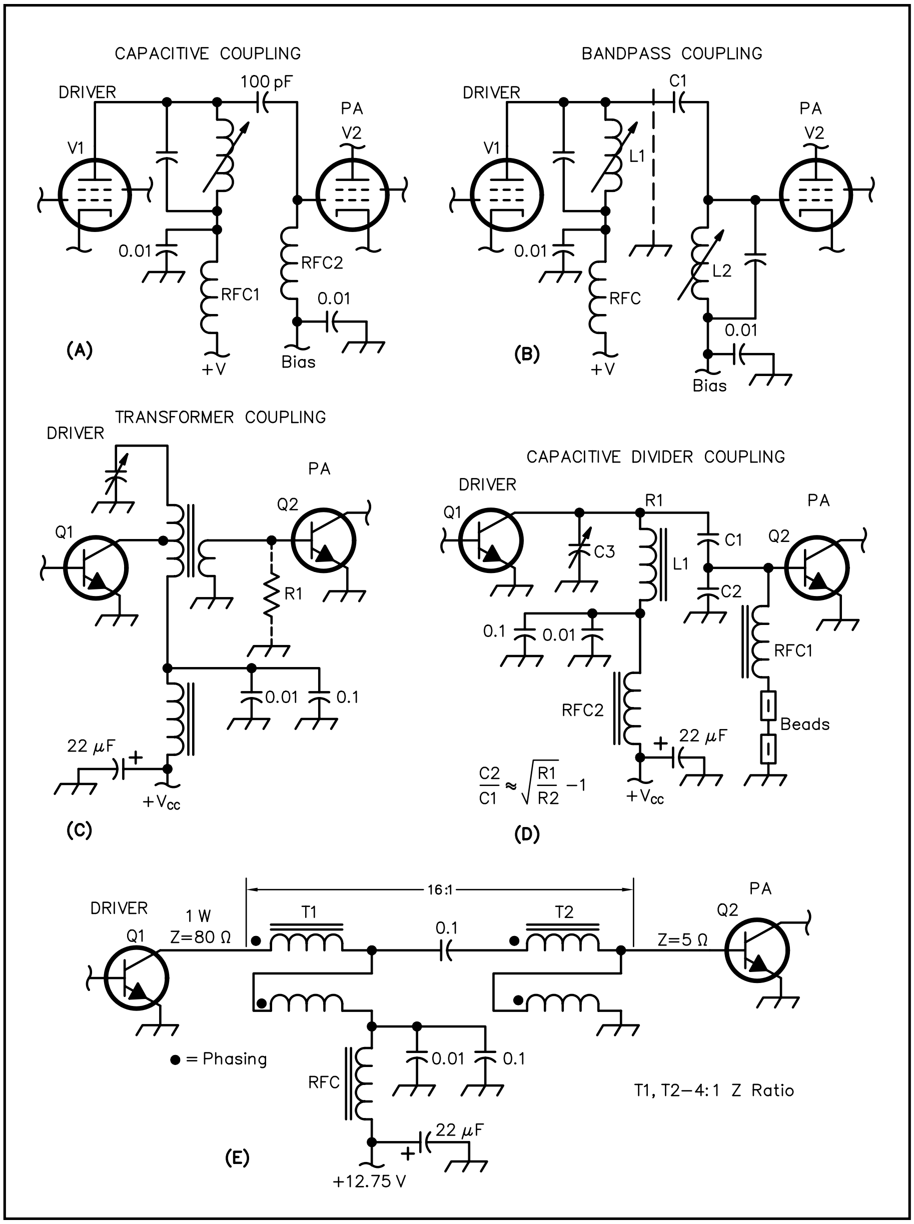


Fig 17.58—Five typical coupling methods for use between amplifier stages.

by changes in the loading of the input and output terminals of the network. It works best between stages that have constant loading. Often, shunt resistors are added to improve the constancy of the response. As an alternative to the coupling capacitor, link coupling can be used at the "cold" end of each coil. Similar band-pass networks can be used in transistor circuits. The collector of the driver and the base of the next stage would be tapped down on their respective coils, to maintain the loaded Q and the selectivity.

A common form of transformer coupling is shown in Fig 17.58C. T1 is usually a toroid for frequencies up to 30 MHz. At higher frequencies, it is sometimes difficult to get accurate control of the impedance ratios, because fractions of a turn cannot be achieved. The collector tap sets the desired load impedance for the transistor and also improves the loaded Q. The impedance ratio from primary to secondary is approximately the square of the ratio of the number of turns at the primary tap point to the number of turns on the secondary. R1 (dotted) can be added to provide a more constant load and improve stability. R1 is usually in the range of 5 to 27 Ω.

A method for coupling between stages with a capacitive divider is shown in Fig 17.58D. The net value of C1 and C2 in series must be added to C3 to determine the value of L1 for resonance. The formula for the ratio of C2 to C1 is included. The reactance of C2 should be about 0.25 of the transistor input resistance. C2 also helps to discourage parasitic oscillations. The ferrite beads (950 µ) reduce the Q of RFC1 at frequencies where it might be involved in an oscillation. The circuit in Fig 17.58C is more susceptible to instability.

When the impedance levels to be transformed are of the proper value to allow an integer turns ratio, two 4:1 impedance- ratio transformers can get a 16:1 impedance ratio as shown in Fig 17.58E. This transforms the 5-Ω base impedance to an 80-Ω collector load impedance. This approach has no selectivity worth mentioning but is widely used in broadband amplifiers where selectivity is not needed or wanted.

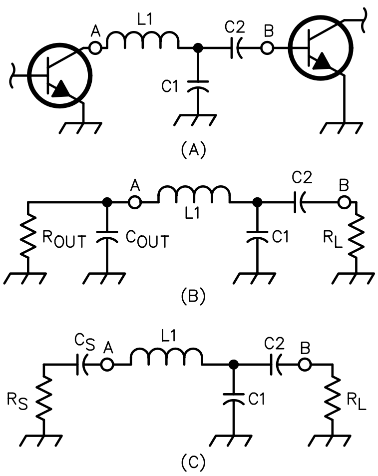
Network Equations

The three networks shown in **Figs 17.59** through **17.61** provide practical solutions to many of the impedance-matching problems encountered by amateurs. In each figure, it is assumed that the desired impedance to the right of the network is higher than that to the left. If the opposite case is true, reverse the network, left to right.

These networks also assume that two resistance values (R_L and R_{out}) are being "matched" by the network, and the discussion proceeds along those lines. But as pointed out previously, the value of R_{out} is usually the desired value of transistor load resistance, which seldom corresponds exactly to the actual output impedance of the transistor. Still, the design method leads to correct answers. It provides the desired resistive load for the transistor output. It also absorbs the transistor output capacitance into the network.

Normally the output impedance of a transistor is given as a resistance R_{out} in parallel with a capacitance C_{out}. To use the design equations for these three networks the output impedance must first be converted from the parallel form to the series form R_s and C_s. These equivalent circuits and the equations for conversion are given in **Fig 17.62**. Often the output capacitance is small enough that it can be neglected. The resulting error is compensated by minor adjustments of the tunable components.

The low-pass T network in Fig 17.61 has the advantage of matching a wide range of impedances with practical component values. Some designers believe that this is the best network in terms of collector efficiency. The harmonic suppression provided by the T network varies with the transformation ratio and the loaded Q of the network. Stages feeding an antenna will need additional harmonic suppression. This is also true for Fig 17.59 and



When $R_S < R_l$ Network 1

1—Select Q_l

2—$X_{L1} = Q_l R_S + X_{CS}$

3—$X_{C2} = Q_L R_L$

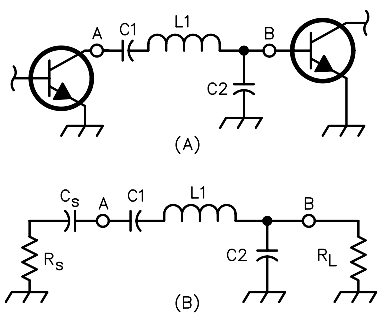
4—$X_{C1} = \dfrac{R_V}{Q_l - Q_L}$

Where

$$Q_L = \sqrt{\left[\frac{R_S\left(1+Q_l^2\right)}{R_L}\right] - 1} = \sqrt{\frac{R_V}{R_L} - 1} = \text{Load Q}$$

$R_V = R_S(1 + Q_l^2)$ = Virtual Resistance of the Network

Fig 17.59—Circuit and mathematical solution for matching network number 1. Actual circuit (A), parallel equivalent (B) and series equivalent (C).



When $R_S < R_L$ Network 2

1—Select Q

2—$X_{C1} = Q_l R_S$

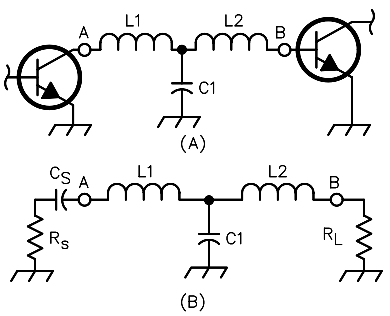
3—$X_{C2} = R_L\sqrt{\dfrac{R_S}{R_L - R_S}}$

4—$X_{L1} = X_{C1} + \left(\dfrac{R_S R_L}{X_{C2}}\right) + X_{C_S}$

Fig 17.60—Circuit and design equations for matching network number 2. Actual circuit (A) and series equivalent (B).

Fig 17.60. These three networks are covered in detail in Motorola Application Notes AN-267 and AN-721. There is also an excellent program, NBMATCH.BAS, on the software disk for *The ARRL UHF/Microwave Experimenter's Manual*. This program and a diagram showing 14 possible network configurations is part of the *Handbook* software, included on this CD. The diagram is also included in the Templates section of the **References** chapter.

Fig 17.63 illustrates, in skeleton form, how transmission line transformers can be used in a push-pull solid state power amplifier. The idea is to maintain highly balanced stages so that each transistor shares equally in the amplification in each stage. The balance also minimizes even-order harmonics so that low-pass filtering of the output is made much easier. In the diagram, T1 and T5 are current (choke) baluns that convert a grounded connection at one end to a balanced (floating) connection at the other end, with a high impedance to ground at both wires. T2 transforms the 50 Ω generator to the 12.5 Ω (4:1 impedance) input impedance of the first stage. T3 performs a similar step-down transformation from the collectors of the first stage to the gates



Network 3

1—Select Q_i

2—$X_{L1} = (R_S Q_i) + X_{C_S}$

3—$X_{L2} = R_L Q_L$

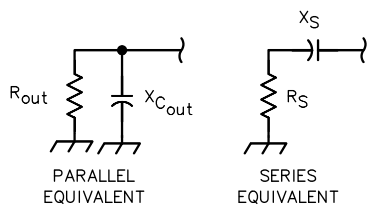
4—$X_{C1} = \dfrac{R_V}{Q_i + Q_L}$

Where

$$R_V = R_S\left(1 + Q_i^2\right)$$

$$Q_L = \sqrt{\left(\frac{R_V}{R_L}\right) - 1} = \text{Load } Q$$

Fig 17.61—Circuit and design equations for matching network number 3. Actual circuit (A) and series equivalent (B).



$$R_S = \frac{R_{out}\ X_{C_{out}}^2}{R_{out}^2 + X_{C_{out}}^2}$$

$$X_S = \frac{R_{out}^2\ X_{C_{out}}}{R_{out}^2 + X_{C_{out}}^2}$$

Fig 17.62—Parallel and series equivalent circuits and the formulas used for conversion.

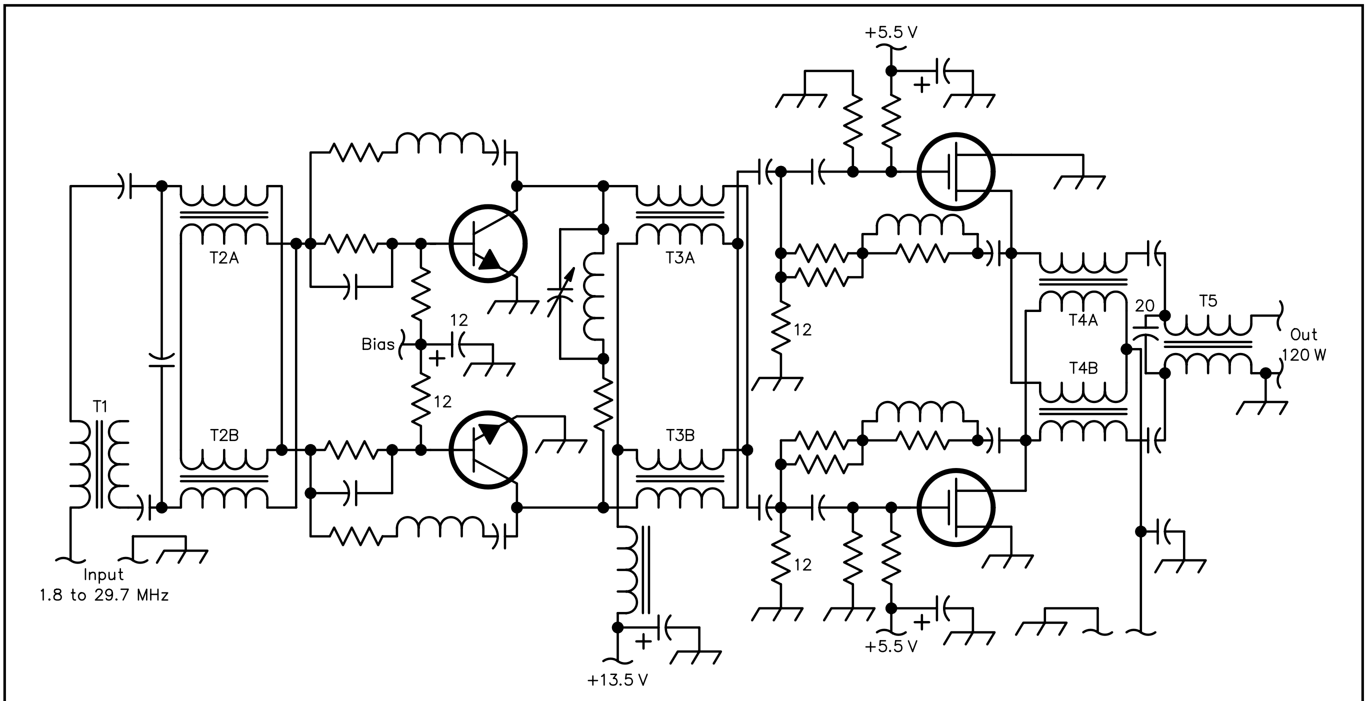


Fig 17.63—This diagram illustrates how transmission line transformers can be used in a push-pull solid state amplifier.

of the second stage. The MOSFETs require a low impedance from gate to ground. The drains of the output stage require an impedance step up from 12.5 Ω to 50 Ω, performed by T4. Note how the choke baluns and the transformers collaborate to maintain a high degree of balance throughout the amplifier. Note also the various feedback and loading networks that help keep the amplifier frequency response flat. **Figs 17.64**, **17.65** and **17.66** give other useful data on transmission line transformers.

Some notes about toroid coils: Toroids do have a small amount of leakage flux, despite rumors to the contrary. Toroid coils are wound in the form of a helix (screw thread) around the circular length of the core. This means that there is a small component of the flux from each turn that is perpendicular to the circle of the toroid (parallel to the axis through the hole) and is therefore not adequately linked to all the other turns. This effect is responsible for a small leakage flux and the effect is called the "one-turn" effect, since the result is equivalent to one turn that is wound around the outer edge of the core and not through the hole. Also, the inductance of a toroid can be adjusted, despite rumors to the contrary. If the turns can be pressed closer together or separated a little, inductance variations of a few percent are possible.

A grounded aluminum shield between adjacent coils can eliminate any significant capacitive or inductive (at high frequencies) coupling. These effects are most easily noticed if a network analyzer is available during the checkout procedure (how many of us are that lucky?), but spot checks with an attenuator ahead of a receiver that is tunable to the harmonics are also very helpful.

Transmission-line transformers were introduced briefly earlier in this chapter. Here we consider practical details for transmitter use and winding instructions. In transmitters, core heating and non-linearity (from excessive voltage or current) determine the choice of core material, its permeability, wire size and the dimensions of the core. A detailed discussion of all of these factors is available in the second edition of ARRL's *Transmission Line Transformers*, by Jerry Sevick. Additional information is contained in the **Analog Theory** chapter of this *Handbook*.

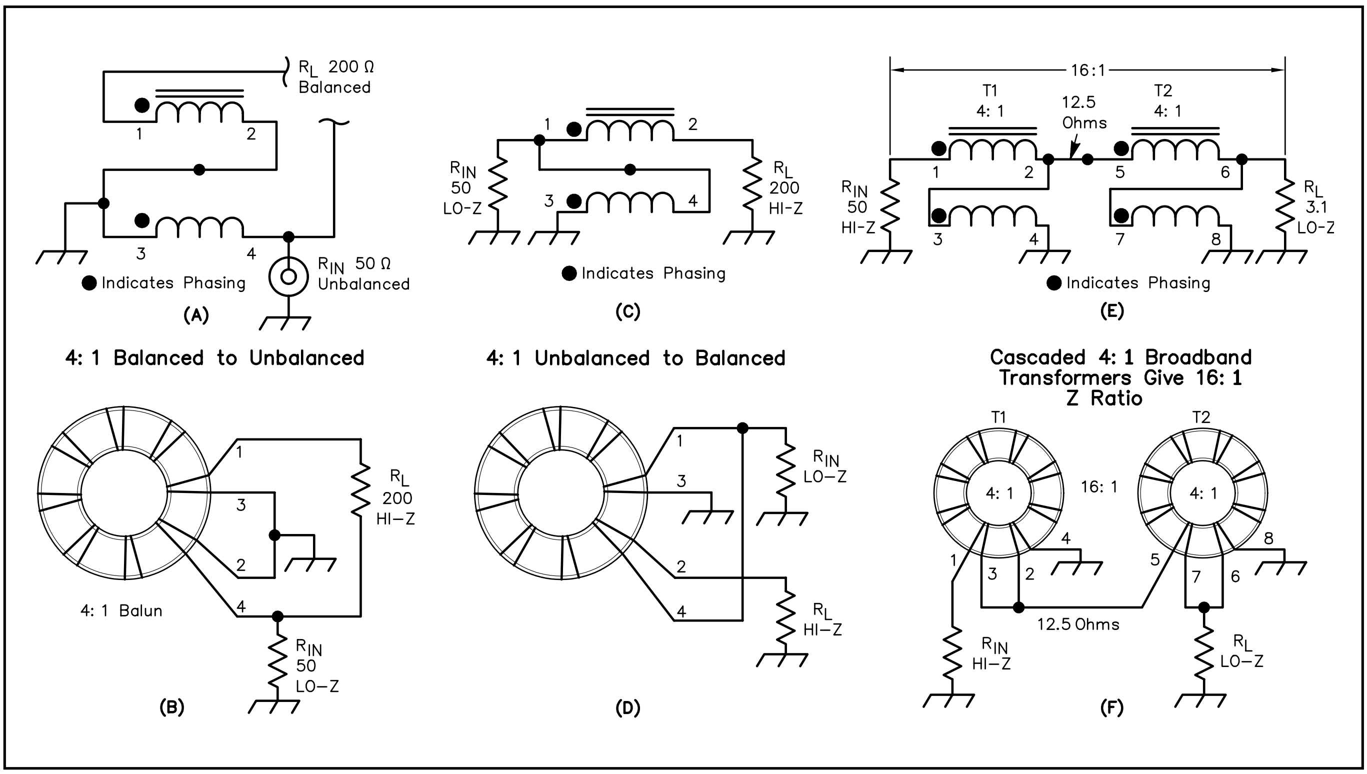


Fig 17.64—Circuit illustrations of 4:1 broadband transmission-line transformers.

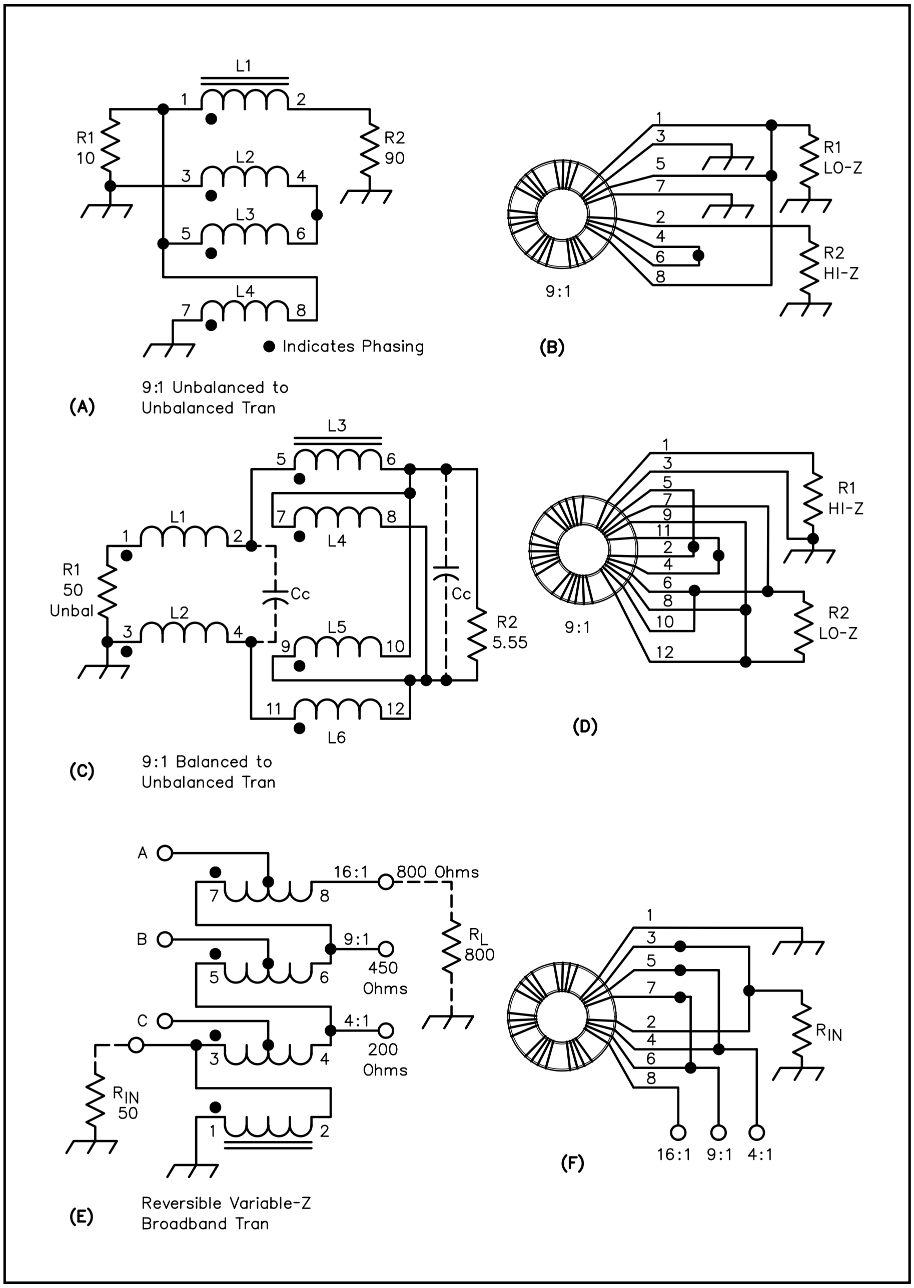


Fig 17.65—Circuit examples of 9:1 broadband transmission-line transformers (A and C) and a variable-impedance transformer (E).

The broadband transformers shown in **Fig 17.64**, **Fig 17.65** and **Fig 17.66** are suitable for use in low-level and high-level solid-state circuits. They can be used as matching devices between circuit modules and in antenna matching networks. For low-power levels and in receivers the cores are usually ferrites with permeabilities from 100 (30 MHz) to 2000 (0.1 MHz). Ferrites are usually preferred because they result in smaller transformers than powdered-iron cores. Small physical size implies fewer turns, greater bandwidth (the length of the winding is a smaller fraction of a wavelength at higher frequencies) and more compact mechanical design. At higher frequencies materials with lower permeabilities are more efficient. For higher power levels, ferrites with permeabilities from 40 to 120 or powdered iron are preferred. Also involved in heating are wire losses, dielectric losses and hysteresis losses (in the core material). IMD and harmonic production in the cores are frequent problems in transmitters. The core size and permeability should be selected to minimize these problems at the signal levels in the equipment. IMD from cores is often noticed in high-signal-level receiver circuits as well.

Fig 17.64 shows two kinds of 4:1 transformers and a method (E) to series connect two of them for a 16:1 impedance ratio. The circuit at E is often used between a 50-Ω source and the base of an RF power transistor.

Two styles of 9:1 (impedance ratio) transformers are shown in Fig 17.65 A and C. They also are used at the inputs of transistor amplifiers and in collector/load circuits. The variable-ratio transformer in Fig 17.65E is excellent for obtaining many values of impedance transformation. It was developed by Jerry Sevick, W2FMI, for a ground-mounted vertical antenna.

Transformers for phase-reversal, 1:1-balun and hybrid-combiner/splitter use are shown in Fig 17.66. This hybrid transformer was discussed earlier in this chapter.

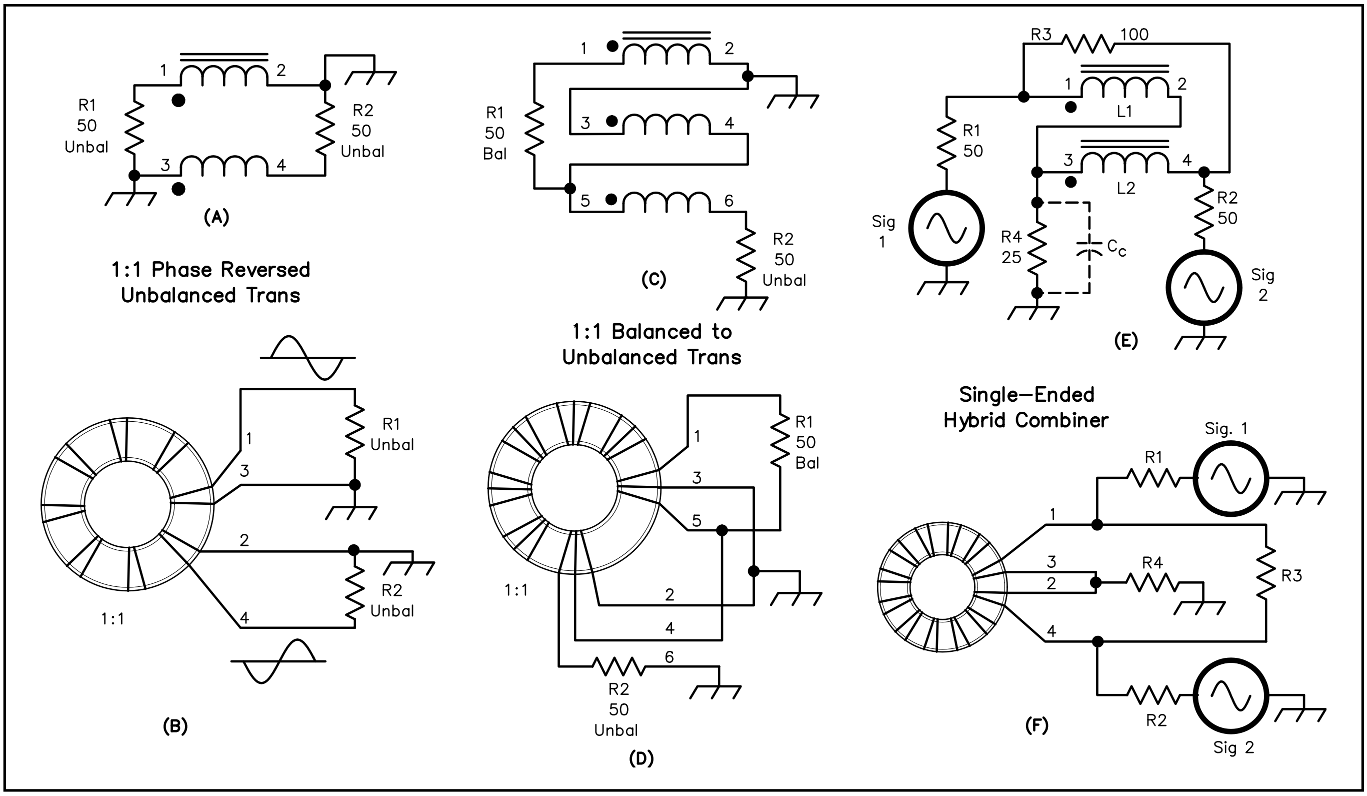


Fig 17.66—Assorted broadband transmission-line transformers.

NBFM Transmitter Block Diagram

Fig 17.67 shows the phase-modulation method, also known as indirect FM. It is the most widely used approach to NBFM. Phase modulation is performed at a low IF, say 455 kHz. Prior to the phase modulator, speech filtering and processing are performed to achieve four goals:

1. Convert phase modulation to frequency modulation (see below),
2. Preemphasize higher speech frequencies for improved signal-to-noise ratio at the receive end,
3. Perform speech processing to emphasize the weaker speech components and
4. Adjust for the microphone's frequency response and possibly also the operator's voice characteristics.

Multiplier stages move the signal to some desired higher IF and also multiply the frequency deviation to the desired final value. If the FM deviation generated in the 455-kHz modulator is 250 Hz, the deviation at 9.1 MHz is 20 × 250, or 5 kHz. A second conversion to the final output frequency is then performed. Prior to this final translation, IF band-pass filtering is performed in order to minimize adjacent-channel interference that might be caused by excessive frequency deviation. This filter needs good phase linearity to assure that the FM sidebands maintain the correct phase relationships. If this is not done, an AM component is introduced to the signal, which can cause nonlinear distortion problems in the PA stages. The final frequency translation retains a constant value of FM deviation for any value of the output signal frequency.

The IF/RF amplifiers are Class C amplifiers because the signal in each amplifier contains, at any one instant, only a single value of instantaneous frequency and not multiple simultaneous frequencies as in SSB. These amplifiers are not sources of IMD, so they need not be "linear." The sidebands that appear in the output are a result only of the FM process (the Bessel functions).

In phase modulation, the frequency deviation is directly proportional to the frequency of the audio signal. To make the deviation independent of the audio frequency, an audio-frequency response that rolls off at 6 dB per octave is needed. An op-amp integrator circuit in the audio amplifier accomplishes this.

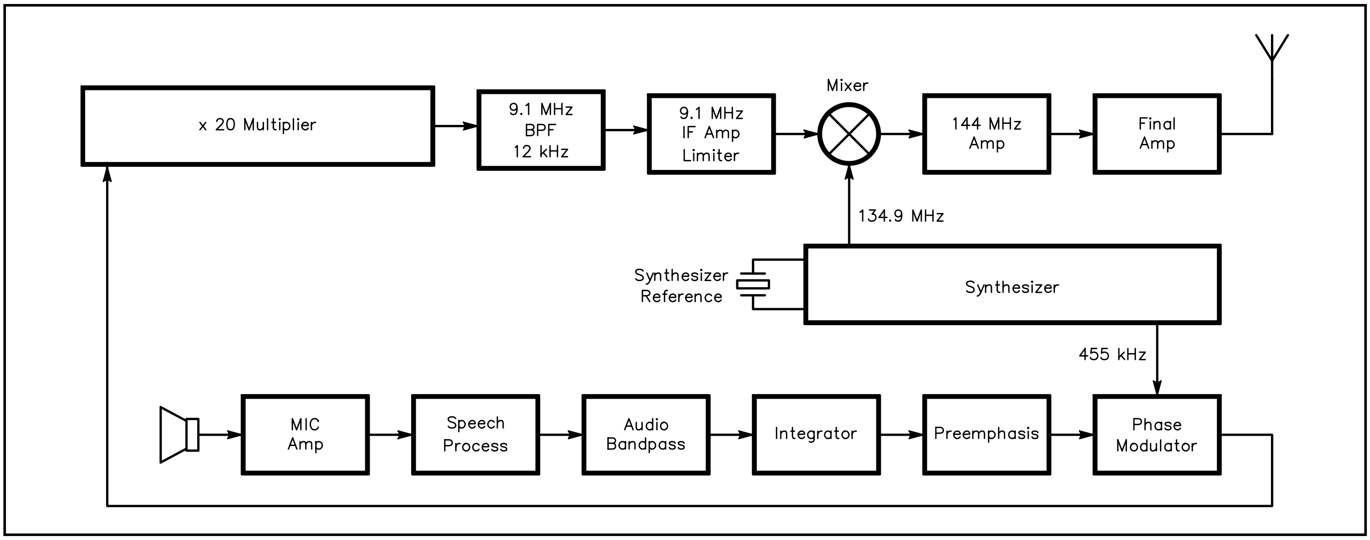


Fig 17.67—Block diagram of a VHF/UHF NBFM transmitter using the indirect FM (phase modulation) method.

This process converts phase modulation to frequency modulation. In addition, audio speech processing helps to maintain a constant value of speech amplitude, therefore constant IF deviation, with respect to audio speech levels. Also, preemphasis of the speech frequencies (6 dB per octave from 300 to 3000 Hz) is commonly used to improve the signal-to-noise ratio at the receive end. Analysis shows that this is especially effective in FM systems when the corresponding deemphasis is used at the receiver (Ref 82).

An IF limiter stage may be used to ensure that any amplitude changes created during the modulation process are removed. The indirect-FM method allows complete frequency synthesis to be used in all the transmitter LOs, so that the channelization of the output frequency is very accurate. The IF and RF amplifier stages are operated in a highly efficient Class-C mode, which is helpful in portable equipment operating on small internal batteries.

NBFM is more tolerant of frequency misalignments, between the transmitter and receiver, than is SSB. In commercial SSB communication systems, this problem is solved by transmitting a pilot carrier that is 10 or 12 dB below PEP. The receiver phase locks to this pilot carrier. The pilot carrier is also used for squelch and AGC purposes. A short-duration “memory” feature in the receiver bridges across brief pilot-carrier dropouts, caused by multipath nulls (Ref 83).

“Direct FM” frequency modulates a high-frequency (say, 9 MHz or so) crystal oscillator by varying the voltage on a varactor. The audio is preemphasized and processed ahead of the frequency modulator. The Transceivers section of this chapter describes such a system (see Fig 17.69).

Transverters

At VHF, UHF and microwave frequencies, transverters that interact with factory-made transceivers in the HF or VHF range are common and are often home-built. These units convert the transceiver transmit signal up to a higher frequency and convert the receive frequency down to the transceiver receive frequency. The resulting performance and signal quality at the higher frequencies are enhanced by the frequency stability and the signal processing capabilities of the transceiver. For example, SSB and narrowband CW from 1.2 to 10 GHz are feasible, and becoming more popular. Some HF and VHF transceivers have special provisions such as connectors, signal-path switching and T/R switching that facilitate use with a transverter.

VHF Transverters

The methods of individual circuit design for a transverter are not much different than methods that have already been described. The most informative approach would be to study carefully an actual project description.

The interface between the transceiver and transverter requires some careful planning. For example, the transceiver power output must be compatible with the transverter's input requirements. This may require an attenuator or some modifications to a particular transverter or transceiver.

When receiving, the gain of the transverter must not be so large that the transceiver front-end is overdriven (system IMD is seriously degraded). On the other hand, the transverter gain must be high enough and its noise figure low enough so that the overall system noise figure is within a dB or so of the transverter's own noise figure. The formulas in this chapter for cascaded noise figure and cascaded third-order intercept points should be used during the design process to assure good system performance. The transceiver's performance should be either known or measured to assist in this effort.

Microwave Transverters

The microwave receiver section of this chapter discussed a 10-GHz transverter project and gave references to the *QST* articles that give a detailed description. The reader is encouraged to refer to these articles and to review the previous material in this chapter.

Other Information

The ARRL UHF / Microwave Experimenter's Manual and ARRL's *Microwave Projects* contain additional interesting and valuable descriptions of transverter and transponder requirements. See also Refs 67-71 in this chapter for valuable information on transverter projects and design.

Transceivers

In recent years the transceiver has become the most popular type of purchased-equipment among amateurs. The reasons for this popularity are:

1. It is economical to use LOs (especially synthesizers), IF amplifiers and filters, power supplies, DSP modules and microprocessor controls for both transmit and receive.
2. It is simpler to perform transmit-receive (T/R) switching functions smoothly and with the correct timing within the same piece of equipment.
3. It is convenient to set a receive frequency and the identical (or properly offset) transmit frequency simultaneously.

In addition, transceivers have acquired very impressive arrays of operator aids that help the operator to communicate more easily and effectively.

The complex design, numerous features and the very compact packaging have made the transceiver essentially a "store bought" item that is extremely difficult for the individual amateur to duplicate at home. The complexity of the work done by teams of design specialists at the factories is incompatible with the technical backgrounds of nearly all individual amateur operators.

The result of this modern trend is that amateur home-built equipment tends to be simpler, with less power output and more specialized (one-band, QRP, CW only, direct conversion, no-tune, receive only, transmit only and so on). Or, the amateur designs and builds add-on devices such as antenna couplers, active adaptive filters, computer interfaces and such.

Transceiver Example

As a way of providing a detailed, in-depth description of modern high-quality transceiver design, let's discuss one recent example, the Ten-Tec Omni VI Plus, an HF ham-band-only solid-state 100 W (output) transceiver, shown in **Fig 17.68**. Ref 84 is a Product Review of this radio. Let's consider first the signal-path block diagram in **Fig 17.69**, one section at a time.

Receiver Front End

The receive antenna can be either the same as the transmitting antenna or an auxiliary receive antenna. A 20-dB attenuator can be switched in as needed. A 1.6-MHz high-pass filter attenuates the broadcast band. A 9.0-MHz trap attenuates any very strong signals at 9.0 MHz that might create interference in the form of blocking or harmonic IMD, especially when tuned to the 10.1-MHz (30-m) or 7-MHz (40-m) bands.

A set of band-pass filters, one for each HF amateur band, eliminates image responses and other spurs in the first mixer. These filters are also used in the low-level transmit stages. A low-noise, high-dynamic-range, grounded-gate JFET RF amplifier with about 9 dB of gain precedes the double balanced diode mixer, which uses 17 dBm of LO in a high-side mix.

Fig 17.68—Photograph of Ten-Tec Omni VI Plus HF transceiver.

First IF

The first IF is 9.0 MHz.

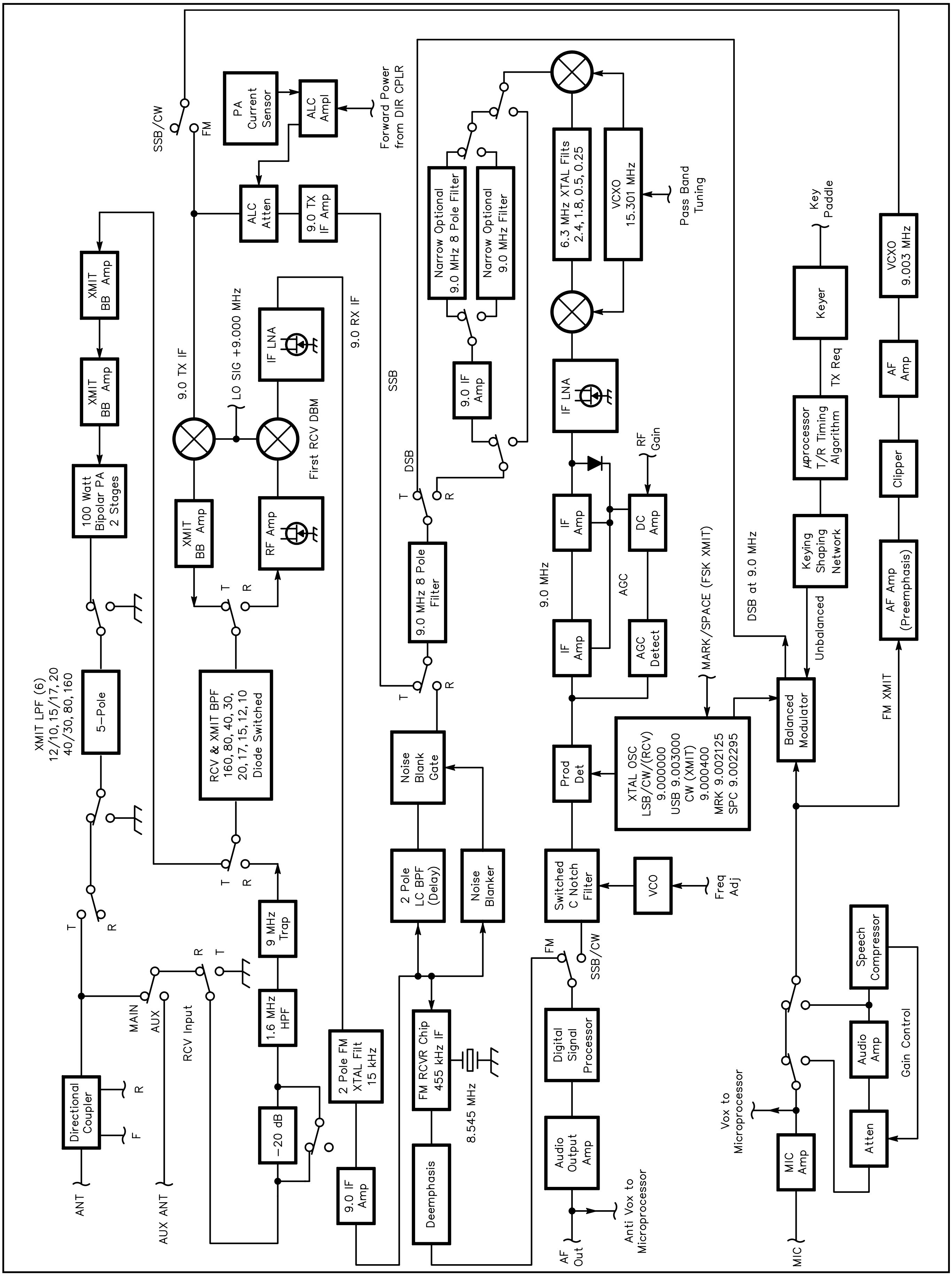


Fig 17.69—Signal path block diagram, receive and transmit, for the Omni VI Plus transceiver.

Because the LO is on the high side, there is a sideband inversion (USB becomes LSB and so on) after the first mixer. A grounded-gate, low-noise JFET amplifier terminates the first mixer in a resistive load and provides 6 dB of gain. This preamp helps to establish the receiver sensitivity (0.15 µV) with minimum gain preceding the mixer.

The preamp is followed by a 15-kHz-wide two-pole filter, which is used for NBFM reception. It is also a roofing filter for the IF amplifier and the noise-blanker circuit that follow it.

The noise blanker gathers impulse energy from the 15-kHz filter, amplifies and rectifies it, and opens a balanced diode noise gate. The IF signal ahead of the gate is delayed slightly by a two-pole filter so that the IF noise pulse and the blanking pulse arrive at the gate at the same time (Ref 48).

The standard IF filter for SSB/CW has 8 poles, is centered at 9.0015 MHz and is 2.4 kHz wide at the –6 dB points. Following this filter, two optional 9.0-MHz filters with the following bandwidths can be installed: 1.8 kHz, 500 Hz, 250 Hz or a 500 Hz RTTY filter. The optional filters are in cascade with the standard filter, for improved ultimate attenuation.

Passband Tuning Section

A mixer converts 9.0 MHz to 6.3 MHz and drives a standard 2.4 kHz wide filter. One of three optional filters, 1.8 kHz, 0.5 kHz (CW) or 250 Hz can be selected instead. A second mixer translates back to the 9.0-MHz frequency. A voltage tuned crystal oscillator at 15.300 MHz (tunable ± 1.5 kHz) is the LO for both mixers. This choice of LO and the 6.3-MHz IF results in very low levels of harmonic IMD products that might cross over the signal frequency and cause spurious outputs. The passband can be tuned ±1.5 kHz.

The composite passband is the intersection of the fixed 9.0-MHz passband and the tunable 6.3-MHz passband. If the first filter is wide and the second much narrower the passband width remains constant over most of the adjustment range. If both have the same bandwidth the resultant bandwidth narrows considerably as the second filter is adjusted. This can be especially helpful in CW mode. **Fig 17.70** shows how passband tuning works.

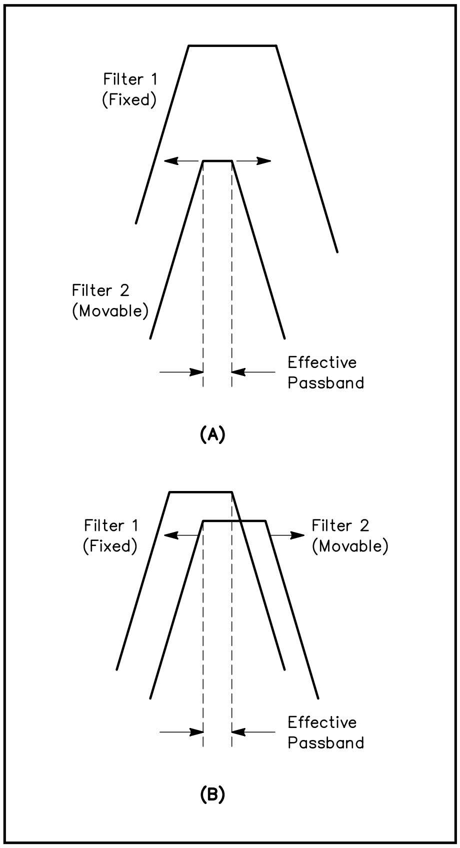


Fig 17.70—Explanation of passband tuning. A: wide first filter and narrow second filter. B: two narrow filters.

IF Amplifiers after Passband Tuning

A low-noise grounded-gate JFET amplifier, with PIN-diode AGC, establishes a low noise figure and a low level of IF noise after the last IF filter. Two IC IF amplifiers (MC1350P) provide most of the receive IF gain. These three stages provide all of the AGC for the receiver. The AGC loop does not include the narrowband IF filters. Two AGC recovery times (Fast and Slow) are available. AGC can be switched off for manual RF gain control as well. The AGC drives the S-meter, which is calibrated at 50 µV for S9 and 0.8 µV for S3.

Product Detector

The IC product detector (CA3053E) uses LO frequencies of 9.000 MHz for LSB and CW (in receive only), 9.003 MHz for USB. When switching between USB and LSB, for a constant value of signal carrier frequency (such as 14.20000 MHz), the LO of the first mixer is moved 3.00 kHz in order to keep the signal within the passband of the IF SSB filters. More about this later.

Audio Notch Filter

In the CW and digital modes, a switched-capacitor notch filter (MF5CN) places a narrow notch in the audio band. The location of the notch is determined by the clock rate applied to the chip. This is determined by a VCO (CD4046BE) whose frequency is controlled by the front panel NOTCH control.

NBFM Reception

After the 15 kHz wide IF filter at 9.0 MHz and before the noise blanker, the IF goes to the NBFM receiver chip (MC3371P). A mixer (8.545-MHz LO) converts it to 455 kHz. The signal goes through an off-chip ceramic band-pass filter, then goes back on-chip to the limiter stages and a quadrature detector. A received signal strength indicator (RSSI) output provides a dc voltage that is proportional to the dB level of the signal. This voltage goes to the front panel meter when in the NBFM mode. A squelch function (NBFM only) is controlled from a potentiometer on the front panel.

Audio Digital Signal Processing (DSP)

The DSP is based on the Analog Devices ADSP 2105 processor. The DSP program is stored in an EPROM and loaded into the 2105's RAM on power-up.

DSP can be used in both SSB and CW. In USB or LSB (not CW or data) the DSP automatically locates and notches out one or several interfering carriers. In SSB or CW the manual audio notch filter described previously is also available, either as a notch filter or to reduce high frequency response (hiss filter). In the CW mode the DSP can be instructed to low-pass filter the audio with several corner-frequency values. A DSP noise reduction function tracks desired signals and attenuates broadband noise by as much as 15 dB, depending on conditions.

Audio Output

The 1.5-W audio output uses a TDA2611 chip. Either FM audio or SSB/CW audio or, in transmit, a CW sidetone, can be fed to the speaker or headphones. The sidetone level (a software adjustment) is separate from the volume control. The audio output, after A/D conversion, is also fed to the Anti-VOX algorithm in the microprocessor.

Transmit Block Diagram

Now, let's look at the path from microphone or key to the antenna, one stage at a time.

Microphone Amplifier

The suggested microphone is 200 Ω to 50 kΩ at 5 mV (–62 dB). A polarizing voltage for electret mics is provided. The Mic Amp drives the balanced modulator, either directly or through the speech compressor. It also supplies VOX information to the microprocessor, via an A/D converter. The microprocessor software sets VOX hang time and sensitivity, as well as the Anti-VOX, via the keypad. Timing and delays for T/R switching are also in the software.

Speech Processor

The audio speech processor is a compressor, as discussed in a previous section of this chapter. A dc voltage that is proportional to the amount of compression is sent to the front panel meter so that compression can be set to the proper level. Clipper diodes limit any fast transients that might overdrive the signal path momentarily.

Balanced Modulator

The balanced modulator generates a double-sideband, suppressed carrier IF at 9.0 MHz. The LO is that used for the receive product detector. There is a carrier nulling adjustment. In CW and FSK modes, the modulator is unbalanced to let the LO pass through. A built-in iambic keyer (Curtis style A or B) is adjustable from 10 to 50 wpm. An external key or keyer can also be plugged in.

IF Filter

The standard 9.0015 MHz, 2.4 kHz-wide 8-pole filter (also used in receive) removes either the lower or upper sideband. The output is amplified at 9.0 MHz.

ALC

The forward-power measurement from the PA output directional coupler is used for ALC, which is applied at the output of the first 9.0-MHz IF amplifier by a PIN diode attenuator. A front panel LED lights when ALC is operating. An additional circuit monitors dc current in the PA and cuts back on RF drive if the PA current exceeds 22 A.

Output Mixer

This mixer translates to the output signal frequency. The same LO frequency is used for each transmit mode (USB/LSB) and the same 3.00000-kHz frequency shift is used to assure that the frequency readout is always correct.

Band-pass Filters

The mixer output contains the image, at 9.0 plus LO and low values of harmonics and harmonic IMD products. A band-pass filter for each ham band, the same ones used in receive, eliminates all out of band products from the transmitter output.

PAs

Four stages of amplification, culminating in push-pull bipolar MRF454s in Class AB, supply 100 W output, CW or PEP from 160 to 10 m. Temperature sensing of the transistors in the last two stages helps to prevent thermal damage. Full output can be maintained, key down, for 20 minutes. Forced air heat-sink cooling allows unlimited on-time.

LO Frequency Management

The LO goals are to achieve low levels of phase noise, high levels of frequency stability and at the same time keep equipment cost within the reach of as many amateurs as possible. As part one of the LO analysis we look at the method used to adjust the LOs in order to keep the speech spectrum of a USB or LSB signal within the 9.0-MHz IF filter passband. This method is somewhat typical for many equipment designs. Refer to **Fig 17.71**.

First Mixer and Product Detector

An SSB signal whose carrier frequency is at 14.20000 MHz, which may be LSB or USB, is translated so that the modulation (300 to 2700 Hz) in either case falls within the passband of the 9.0015-MHz 8-pole crystal filter. For a USB signal this is accomplished by increasing the first LO 3 kHz, as indicated. Since the LO is on the high side of the signal frequency there is a sideband “inversion” at the first mixer. The passband tuning module does not change this relationship. At the product detector, the LO is increased 3 kHz in USB so that it is the same as the carrier (suppressed) frequency of the IF signal. Note that the designators “USB” or “LSB” at the product detector LO refer to the antenna signal, not the IF signal. The jog of the first mixer LO is accomplished partly within the crystal oscillator and partly within the VFO. The microprocessor sends the frequency instructions to both of these oscillators. Despite the frequency offsets, the digital readout displays the correct carrier fre-

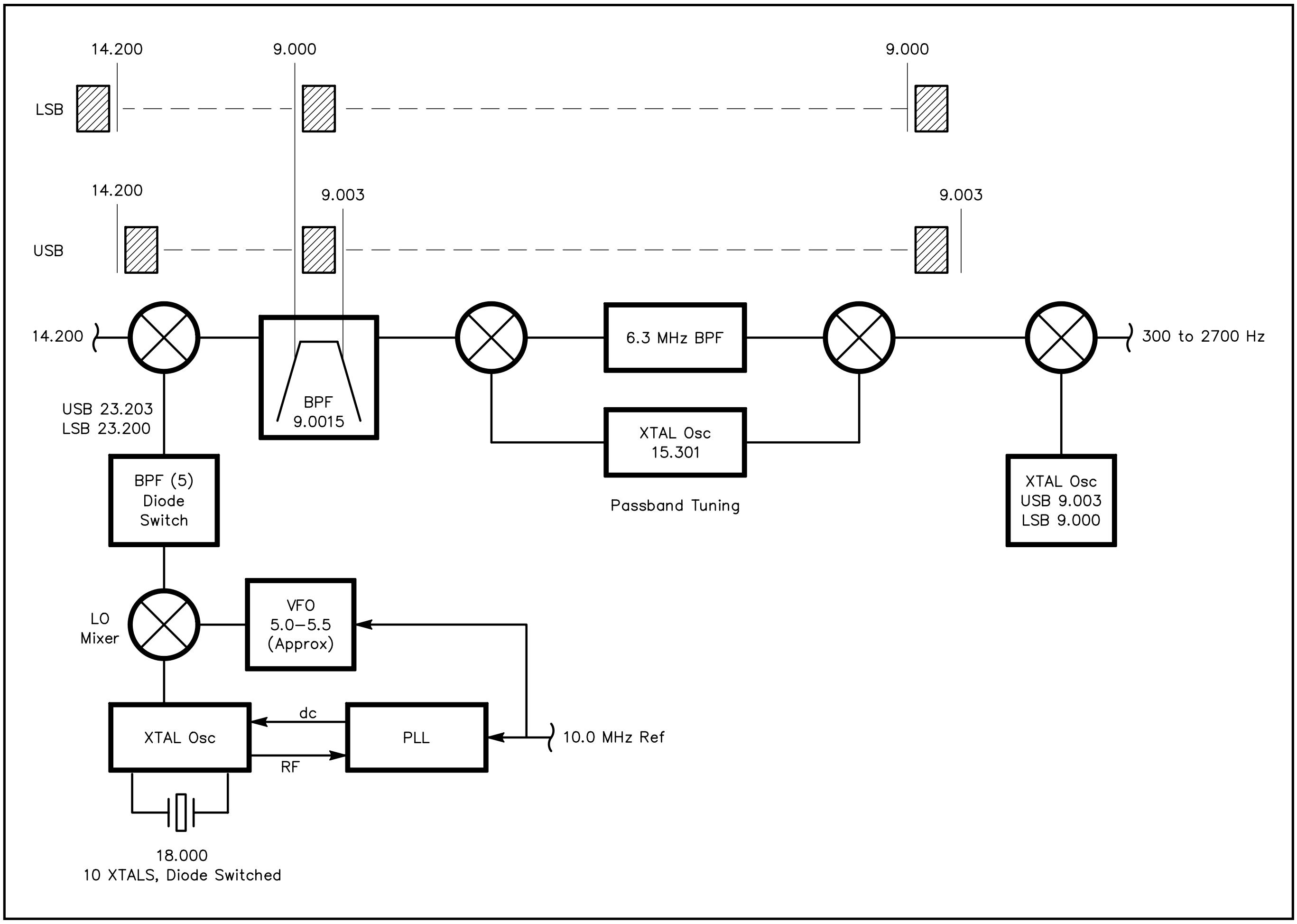


Fig 17.71—LO frequency management in USB and LSB.

quency, in this example 14.20000 MHz. And, of course, the same procedures apply to the transmit mode.

Another interesting idea involves the first LO mixer. The 18-MHz crystal and the VFO are added to get 23.0 to 23.5 MHz, for the 20-m band. But the output of the LO mixer also contains the difference, 12.5 to 13.0 MHz, which is just right for the 80-m band. For the 20-m band one BPF selects 23.0 to 23.5 MHz. For the 80-m band the 12.5 to 13.0-MHz BPF is selected. A problem occurs, though, because now the direction of frequency tuning is reversed, from high to low. The microprocessor corrects this by reversing the direction of the tuning knob (an optical encoder). Other "book keeping" is performed so that the operation is transparent to the operator. A similar trick is used on the 17-m band and the 28.5 to 29.0-MHz segment of the 10-m band.

CW-Mode LO Frequencies

In CW mode the "transceiver problem" shows up. See **Fig 17.72** for a discussion of this problem. If the received carrier is on exactly 14.00000 MHz and if we want to transmit our carrier on that same exact frequency then the transmitter and the receiver are both "zero beat" at 14.00000 MHz. In receive we would have to retune the receiver, say up 700 Hz, to get an audible 700-Hz beat note. But then when we transmit we are no longer on 14.0000 MHz but are at 14.00070 MHz. We would then have to reset to 14.00000 when we transmit.

The transceiver's microprocessor performs all of these operations automatically. Fig 17.72 shows that

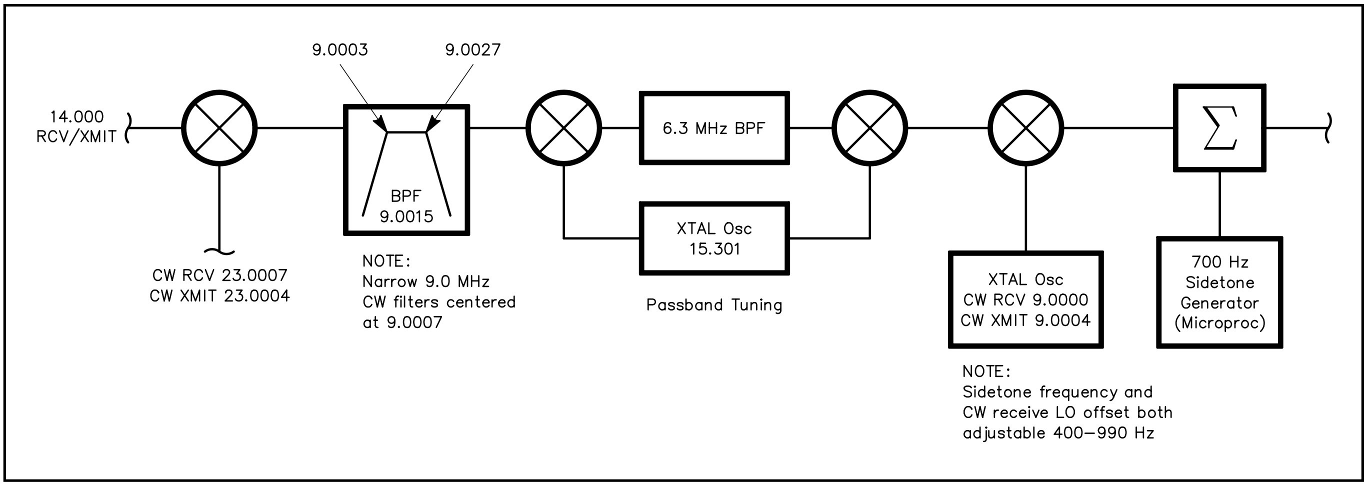


Fig 17.72—LO management in the CW receive and transmit modes.

in receive the first-LO frequency is increased 700 Hz. This puts the first IF at 9.0007 MHz, which is inside the passband of the 9.0 MHz IF filter. The BFO is at 9.0000 MHz and an audio beat note at 700 Hz is produced. This 700-Hz pitch is compared to a 700-Hz audio oscillator (from the microprocessor). When the two pitches coincide the signal frequency display of our transceiver coincides almost exactly with the frequency of the received signal. The digital frequency display reads "14.00000" at all times. The value of the 700-Hz reference beat can be adjusted between 400 and 950 Hz by the user. The receive LO shift matches that value.

When the optional 500/250 Hz CW 9.0 MHz IF filters are used, these are centered at 9.0007 MHz. These filters are used in receive but not in transmit.

When we transmit, the transmit frequency is that which the frequency display indicates, 14.00000 MHz. However, there is a slight problem. The 9.0-MHz transmit IF must be increased slightly to get the speech signal within the passband of the 9.0 MHz IF filter. The transmit BFO is therefore at 9.0004 MHz. The mixer LO is also moved up 400 Hz so that the transmit output frequency will be exactly 14.0000 MHz.

In addition to the above actions, the RIT (receive incremental tuning) and the XIT (transmit incremental tuning) knobs permit up to ±9.9 kHz of independent control of the receive and transmit frequencies, relative to the main frequency readout.

Local Oscillators

Fig 17.71 indicates that the crystal oscillator for the first mixer is phase locked. Each of the 10 crystals is locked to a 100-kHz reference inside the PLL chip. This reference is derived from a 10-MHz system reference.

Let's go into some detail regarding the very interesting 5.0 to 5.5 MHz VFO circuitry. **Fig 17.73** is a block diagram.

The VCO output, 200 to 220 MHz, is divided by 40 to get 5.0 to 5.5 MHz. The reference frequency for the PLL loop is 10 kHz, so each increment of the final output is 10 kHz / 40, which is 250 Hz. Phase noise in the PLL is also divided by 40, which is equivalent to 32 dB (20 × log(40)).

To get 10-Hz steps at the output, the voltage-tuned crystal oscillator at 39.94 MHz is tuned in 200-Hz steps (the division ratio from oscillator to final output is 20 instead of 40 because of the frequency doubler). To get 200-Hz steps in this oscillator, serial data from the microprocessor is fed into a latch. This data is sent from a RAM lookup table that has the correct values to get the 200-Hz increments very accurately. The outputs of the latch are fed into an R/2R ladder (D/A conversion) and the dc voltage tunes the VCXO. Adjustment potentiometers calibrate the tuning range of the oscillator over 5000 Hz in

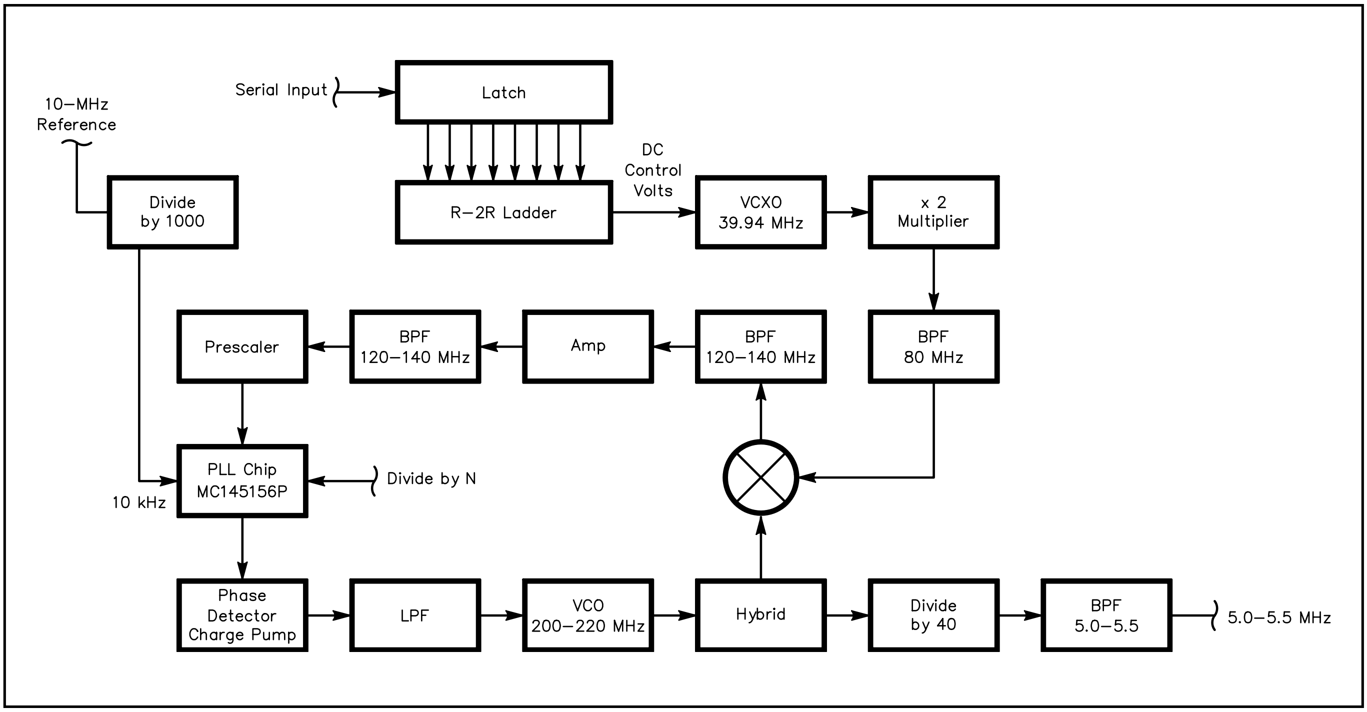


Fig 17.73—Block diagram of the VFO.

200-Hz increments. At the final output, this tuning range fills in the 10-Hz steps between the 250-Hz increments of the PLL. Although this circuit is not phase locked to a reference (it's an open-loop), the resulting frequency steps are very accurate, especially after the division by 20. This economical approach reduces the complexity and cost of the VFO considerably, but performs extremely well (very low levels of phase noise and frequency drift).

CW Break-In (Fast QSK) Keying

The radio is capable of break-in keying at rates up to 25 wpm when it is in the FAST QSK mode. This mode is also used for AMTOR FSK. **Fig 17.74** explains the action and the timing involved in this T/R switching. The sequence of events is as follows:

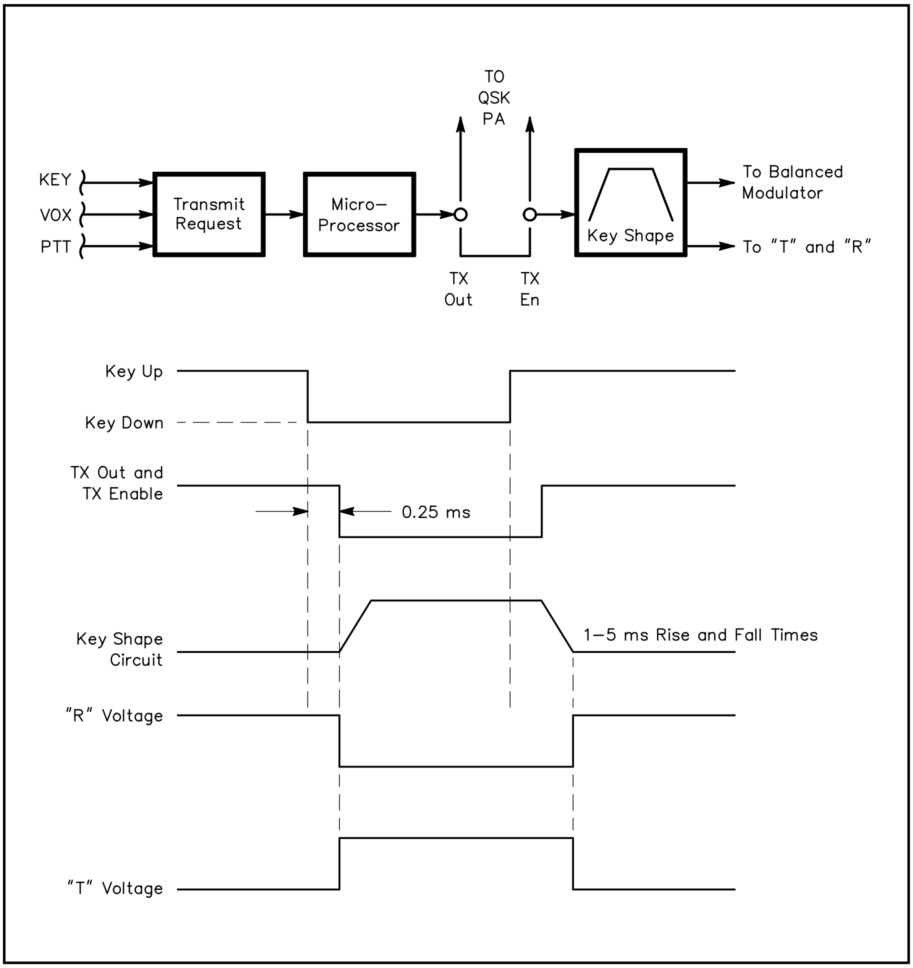


Fig 17.74—Fast QSK operation of the Omni VI transceiver.

1. The key is pressed.
2. A Transmit Request is sent to the microprocessor.
3. The microprocessor changes the LO and BFO to their transmit frequencies.
4. After a 0.25-ms delay a Transmit Out logic level is sent to a jack on the rear panel.
5. The Transmit Out signal is jumpered to the TRANSMIT ENABLE jack. If an external Fast QSK PA is being used, an additional short delay is introduced while it is being switched to the transmit mode.
6. The Transmit Enable signal starts the keying-waveform circuit, which ramps up in 3 ms.
7. Near the bottom of this ramp, the “T” (transmit B+) voltage goes high and the “R” (receive B+) voltage goes low. The T/R reed relay (very fast) at the Omni VI PA output switches to transmit.
8. The shaped keying waveform goes to the balanced modulator and the RF envelope builds up. There is a very brief delay from balanced modulator to Omni VI output. The T/R relay is switched before the RF arrives.
9. The key is opened.
10. The Transmit Out and Transmit Enable lines go high.
11. The keying waveform ramps down. RF ramps down to zero.
12. After 5 ms (a fixed delay set by the microprocessor) “T” goes low and “R” goes high and the microprocessor returns the LO and BFO to their receive frequencies.

Slow QSK

In the Slow QSK mode the action is as described above. The radio reverts quickly to the receive mode. However, the receive audio is muted until the end of an extended (adjustable) delay time.

There is also a relay in the Omni VI that can be used to T/R switch a conventional (not Fast QSK) external PA. This option is selected from the operator’s menu and is available only in the Slow QSK mode. The relay is held closed for the duration of the delay time in the slow QSK mode. When using this option the operator must ensure that the external PA switches fast enough on the first “key-down” so that it is not “hot switched” or that the first dot is missed. Many older PAs do not respond well at high keying rates. If the PA is slow, we can still use the Fast QSK mode with the Omni VI (“barefoot”) and the PA will be bypassed because the optional relay is not energized in the Fast QSK mode.

VOX

In SSB mode, VOX and PTT perform the same functions and in the same manner as the CW Slow QSK described above. The VOX hang-time adjustment is separate from the CW hang-time adjustment. A MUTE jack allows manual switching (foot switch, and so on) to enable the transmit mode without applying RF. The key or the VOX then subsequently applies RF to the system. This arrangement helps if the external PA has slow T/R switching.

Operating Features

Modern transceivers have, over the years, acquired a large ensemble of operator’s aids that have become very popular. Here are some descriptions of them:

Key Pad

The front panel key pad is the means for configuring a wide assortment of operating preferences and for selecting bands and modes.

Frequency Change

1. Use the tuning knob. The tuning rate can be programmed to 5.12, 2.56, 1.71, 1.28, 1.02 or 0.85 kHz per revolution. The knob has adjustable drag.

2. UP and DOWN arrows give 100 kHz per step.
3. Band selection buttons.
4. Keyboard entry of an exact frequency.

Mode Selection

1. ***Tune:*** puts the rig in CW mode, key down, for various "tune-up" operations.
2. ***CW:*** An optional DSP low-pass filter can be selected. Cutoff frequencies of 600, 800, 1000, 1200 or 1400 Hz can be designated. A SPOT function generates a 700-Hz audio sidetone that can be used for precise frequency setting (the received signal pitch matches the 700-Hz tone). The pitch of the sidetone can be adjusted from 400 to 950 Hz. Audio level is adjustable also. FAST QSK and SLOW QSK are available as previously described. Cascaded CW filters are available, with passband tuning of one of the filters. CW filter options: 500/250 Hz at 6.3 MHz IF and 500 Hz at 9.0-MHz IF.
3. ***USB or LSB***: Standard SSB IF bandwidth is 2.4 kHz (two 8-pole filters in cascade). The second filter can be passband-tuned ±1.5 kHz. Additional IF filters with 1.8-kHz BW are available.
4. ***FSK and AFSK:*** Special FSK filter for receive. AMTOR operation with FAST QSK capability. AFSK generator can be plugged into microphone jack.
5. ***FM:*** 15-kHz IF filters at 9.0 MHz and 455 kHz. Quadrature detection, RSSI output and squelch. Adjustable transmit deviation. FM transmit uses the direct method, as shown in Fig 17.69.
6. ***VOX:*** VOX sensitivity and hang time adjustable via the key pad. Anti-VOX level adjustable.

Time of Day Clock

There is a digital readout on front panel.

Built-In Iambic Keyer

Curtis type A or B, front panel speed knob. Adjustable dot-dash ratio. Also external key or keyer.

Dimmer

Adjusts brightness of front panel display.

Dual VFOs

Select A or B. Independent frequency, mode, RIT and filter choices stored for each VFO. Used for split-frequency operation.

Receiver Incremental Tuning (RIT)

Each VFO has its own stored RIT value.

Frequency Offset Display

RIT value adjusted with knob. RIT can be toggled on and off, or cleared to zero.

Transmitter Incremental Tuning (XIT)

Same comments as RIT. Simultaneous RIT and XIT.

Cross-Band and Cross-Mode Operation

For cross-band operation, use PTT for SSB and manual switch for CW.

Scratch-Pad Memory

Stores a displayed frequency. Restores that frequency to the VFO on command.

Band Register

Allows toggling between two frequencies on each band.

Memory Store

Store 100 values of frequency, band, mode, filter, RIT, XIT. Memory channels can be recalled by channel number (key pad), "scrolling" the memory channels or "memory tune" using the main tuning knob.

Lock

Locks the main tuning dial.

User Option Menu

Enables configuration of the radio via the keypad.

Meter

Select between receive signal strength (on SSB/CW or NBFM), speech processor level, forward power, SWR and PA dc current.

AGC

Fast, slow, off and manual RF gain control.

FM Squelch Adjust

Passband Tuning Knob

Notch

Automatically notch out several heterodynes on SSB/FM or manually notch on CW/digital modes. Adjustable low-pass filter in CW mode.

Antenna Switch

Auxiliary antenna may be selected in receive mode.

Interface Port

25-pin D connector for interface to personal computer.

OTHER TRANSCEIVERS

Other transceivers vary in cost, complexity and features. The one just described is certainly one of the best, at a reasonable price. For reviews of other transceivers (to see the differences in cost, features and performance specs) refer to the Product Reviews in *QST* and ARRL's *Radio Buyer's Sourcebook* series. For example, see Refs 85, 86, 87, 88.

Let's look quickly at a more expensive transceiver, to get some feel for other options and design approaches at higher prices.

The Kenwood TS-950SDX

This radio has a "subreceiver," containing an SSB filter and a 500-Hz CW filter. Simultaneous reception on both receivers is possible. Both receivers can be tuned independently.

A stored-message memory allows automatic transmission of short CW messages.

It performs harmonic sampling and DSP at a 100-kHz IF. Digital detection is followed by digital filtering. The 100-kHz IF also has a voltage-tuned notch filter. SSB is generated by a DSP phasing method.

The synthesizer uses a DDS that can tune in 1-Hz steps. The receiver covers from 100 kHz to 30 MHz. The radio uses up conversion to a first IF of 73 MHz, followed by IFs at 8.8 MHz, 455 kHz and 100 kHz.

The PA uses FET technology. A cooling fan is provided. The radio also contains a built-in automatic antenna tuner with memory retention of tuner settings.

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A ROCK-BENDING RECEIVER FOR 7 MHz

This simple receiver by Randy Henderson, WI5W, originally published in Aug 1995 *QST*, is a direct-conversion type that converts RF directly to audio. Building a stable oscillator is often the most challenging part of a simple receiver. This one uses a tunable crystal-controlled oscillator that is both stable and easy to reproduce. All of its parts are readily available from multiple sources and the fixed-value capacitors and resistors are common components available from many electronics parts suppliers.

THE CIRCUIT

This receiver works by mixing two radio-frequency signals together. One of them is the signal you want to hear, and the other is generated by an oscillator circuit (Q1 and associated components) in the receiver. In **Fig 17.75**, mixer U1 puts out sums and differences of these signals and their harmonics. We don't use the sum of the original frequencies, which comes out of the mixer in the vicinity of 14 MHz. Instead, we use the frequency *difference* between the incoming signal and the receiver's oscillator—a signal in the audio range if the incoming signal and oscillator frequencies are close enough to each other. This signal is filtered in U2, and amplified in U2 and U3. An audio transducer (a speaker or headphones) converts U3's electrical output to audio.

How the Rock Bender Bends Rocks

The oscillator is a tunable crystal oscillator—a variable crystal oscillator, or *VXO*. Moving the oscillation frequency of a crystal like this is often called *pulling*. Because crystals consist of precisely sized pieces of quartz, crystals have long been called *rocks* in ham slang—and receivers, transmitters and transceivers that can't be tuned around due to crystal frequency control have been said to be *rockbound*. Widening this rockbound receiver's tuning range with crystal pulling made *rock bending* seem just as appropriate!

L2's value determines the degree of pulling available. Using FT-243-style crystals and larger L2 values, the oscillator reliably tunes from the frequency marked on the holder to about 50 kHz below that point with larger L2 values. (In the author's receiver a 25-kHz tuning range was achieved.) The oscillator's frequency stability is very good.

Inductor L2 and the crystal, Y1, have more effect on the oscillator than any other components. Breaking up L2 into two or three series-connected components often works better than using one RF choke. (The author used three molded RF chokes in series—two 10-µH chokes and one 2.7-µH unit.) Making L2's value too large makes the oscillator stop.

The author tested several crystals at Y1. Those in FT-243 and HC-6-style holders seemed more than happy to react to adjustment of C7 (TUNING). Crystals in the smaller HC-18 metal holders need more inductance at L2 to obtain the same tuning range. One tiny HC-45 unit from International Crystals needed 59 µH to eke out a mere 15 kHz of tuning range.

Input Filter and Mixer

C1, L1, and C2 form the receiver's input filter. They act as a peaked *low-pass* network to keep the mixer, U1, from responding to signals higher in frequency than the 40-meter band. (This is a good idea because it keeps us from hearing video buzz from local television transmitters, and signals that might mix with harmonics of the receiver's VXO.) U1, a Mini-Circuits SBL-1, is a passive diode-ring mixer. Diode-ring mixers usually perform better if the output is terminated properly. R11 and C8 provide a resistive termination at RF without disturbing U2A's gain or noise figure.

Audio Amplifier and Filter

U2A amplifies the audio signal from U1. U2B serves as an active low-pass filter. The values of C12,

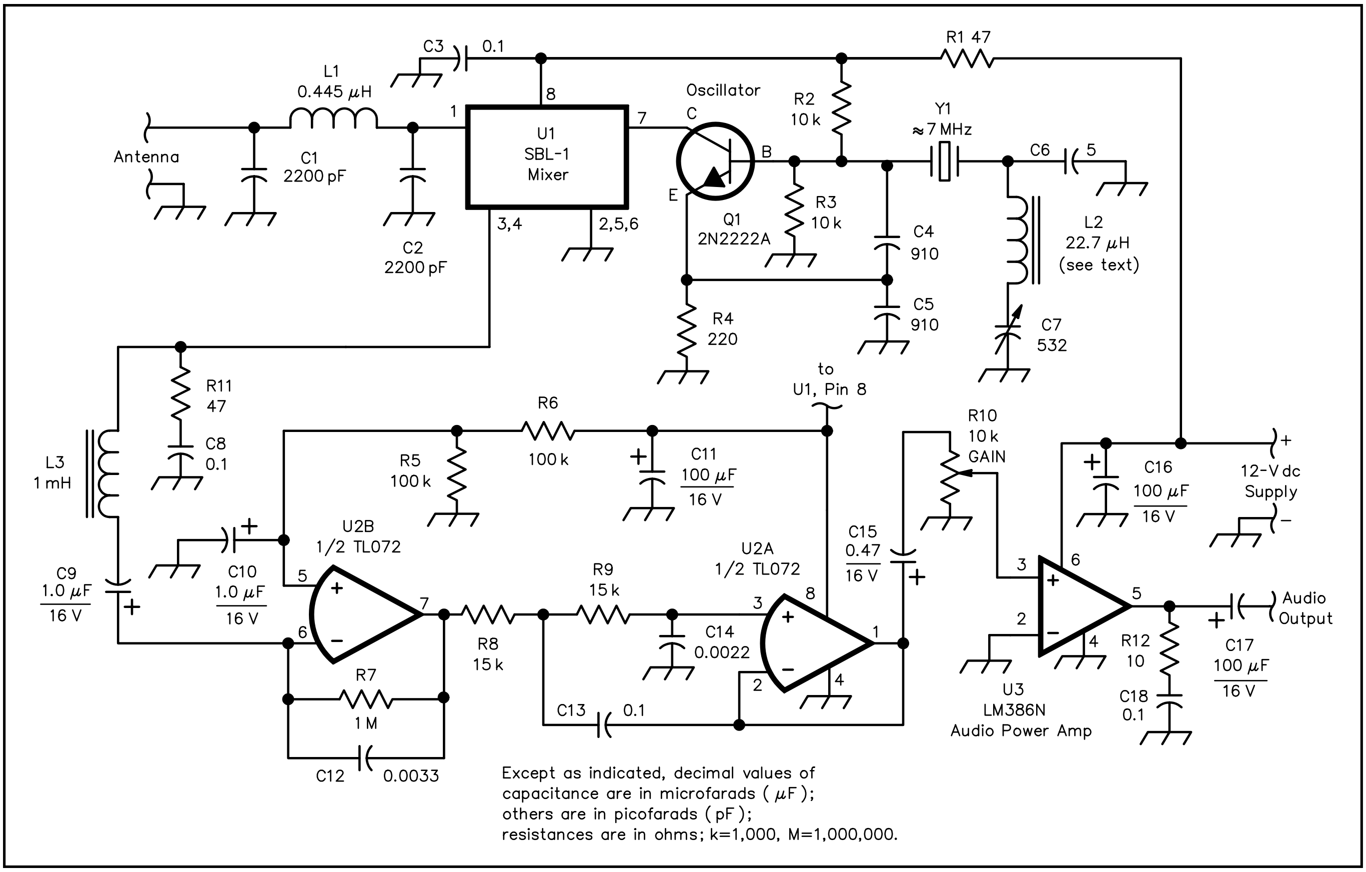


Fig 17.75—An SBL-1 mixer (U1, which contains two small RF transformers and a Schottky-diode quad), a TL072 dual op-amp IC (U2) and an LM386 low-voltage audio power amplifier IC (U3) do much of the Rock-Bending Receiver's magic. Q1, a variable crystal oscillator (VXO), generates a low-power radio signal that shifts incoming signals down to the audio range for amplification in U2 and U3. All of the circuit's resistors are $^1/_4$-W, 5%-tolerance types; the circuit's polarized capacitors are 16-V electrolytics, except C10, which can be rated as low as 10 V. The 0.1-µF capacitors are monolithic or disc ceramics rated at 16 V or higher.

C1, C2—Ceramic or mica, 10% tolerance.

C4, C5, and C6—Polystyrene, dipped silver mica, or C0G (formerly NP0) ceramic, 10% tolerance.

C7—Dual-gang polyethylene-film variable (266 pF per section) available as #24TR218 from Mouser Electronics (see the Address List in the References chapter). Screws for mounting C7 are Mouser #48SS003. A rubber equipment foot serves as a knob. (Any variable capacitor with a maximum capacitance of 350 to 600 pF can be substituted; the wider the capacitance range, the better.)

C12, C13, C14—10% tolerance. For SSB, change C12, C13 and C14 to 0.001 µF.

L1—4 turns of AWG #18 wire on $^3/_4$-inch PVC pipe form. Actual pipe OD is 0.85 inch. The coil's length is about 0.65 inch; adjust turns spacing for maximum signal strength. Tack the turns in place with cyanoacrylic adhesive, coil dope or Duco cement. (As a substitute, wind 8 turns of #18 wire around 75% of the circumference of a T-50-2 powdered-iron core. Once you've soldered the coil in place and have the receiver working, expand and compress the coil's turns to peak incoming signals, and then cement the winding in place.)

L2—Approximately 22.7 µH; consists of one or more encapsulated RF chokes in series (two 10-µH chokes [Mouser #43HH105 suitable] and one 2.7-µH choke [Mouser #43HH276 suitable] used by author). See text.

L3—1-mH RF choke. As a substitute, wind 34 turns of #30 enameled wire around an FT-37-72 ferrite core.

Q1—2N2222, PN2222 or similar small-signal, silicon NPN transistor.

R10—5 or 10-kΩ audio-taper control (Radio Shack No. 271-215 or 271-1721 suitable).

U1—Mini-Circuits SBL-1 mixer.

U2—TL072CN or TL082CN dual JFET op amp.

Y1—7-MHz fundamental-mode quartz crystal. Ocean State Electronics carries 7030, 7035, 7040, 7045, 7110 and 7125-kHz units.

PC boards for this project are available from FAR Circuits (see the Address List in the References chapter). Price \$5, plus \$1.50 shipping (for up to four boards).

C13, and C14 are appropriate for listening to CW signals. If you want SSB stations to sound better, make the changes shown in the caption for **Fig 17.76**.

U3, an LM386 audio power amplifier IC, serves as the receiver's audio output stage. The audio signal at U3's output is more than a billion times more powerful than a weak signal at the receiver's input, so don't run the speaker/earphone leads near the circuit board. Doing so may cause a squealy audio oscillation at high volume settings.

CONSTRUCTION

If you're already an accomplished builder, you know that this project can be built using a number of construction techniques, so have at it! If you're new to building, you should consider building the Rock-Bending Receiver on a printed circuit (PC) board. (The parts list tells where you can buy one ready-made.)

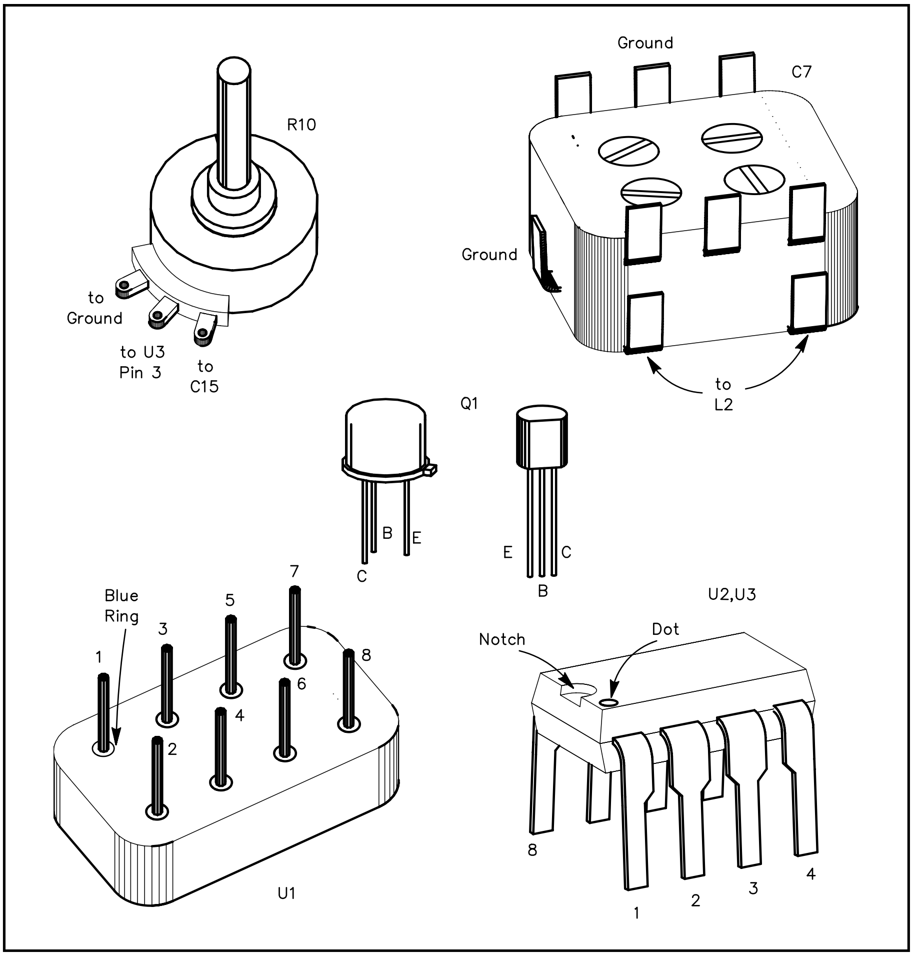
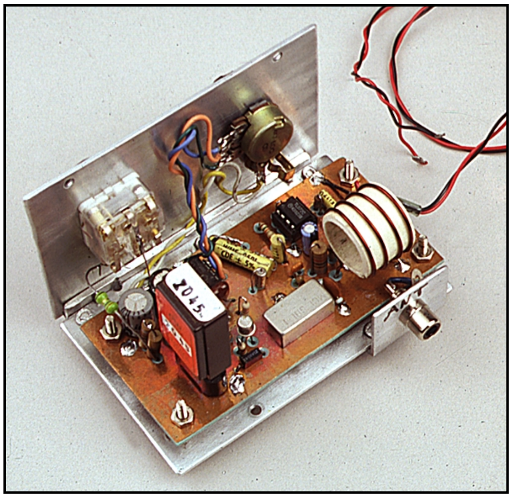


Fig 17.76—The Mouser Electronics part suggested for C7 has terminal connections as shown here. (You can use any variable capacitor with a maximum capacitance of 350 to 600 pF for C7, but its terminal configuration will differ from that shown here.) Two Q1-case styles are shown because plastic or metal transistors will work equally well for Q1. If you build your Rock-Bending Receiver using a prefab PC board, you should mount the ICs in 8-pin mini-DIP sockets rather than just soldering the ICs to the board.

If you use a homemade double-sided circuit board based on the PC pattern in the **References** chapter, you'll notice that it has more holes than it needs to. The extra holes (indicated in the part-placement diagram with square pads) allow you to connect its ground plane to the ground traces on its foil side. (Doing so reduces the inductance of some of the board's ground paths.) Pass a short length of bare wire (a clipped-off component lead is fine) into each of these holes and solder on both sides. Some of the circuit's components (C1, C2 and others) have grounded leads accessible on both sides of the board. Solder these leads on both sides of the board.

Another important thing to do if you use a homemade double-sided PC board is to countersink the ground plane to clear all ungrounded holes. (Countersinking clears copper away from the holes so components won't short-circuit to the ground plane.) A $^1/_4$-inch-diameter drill bit works well for this. Attach a control knob to the bit's shank and you can safely use the bit as a manual countersinking tool. If you countersink your board in a drill press, set it to about 300 rpm or less, and use very light pressure on the feed handle.

Mounting the receiver in a metal box or cabinet is a good idea. Plastic enclosures can't shield the TUNING capacitor from the presence of your hand, which may slightly affect the receiver tuning. You don't have



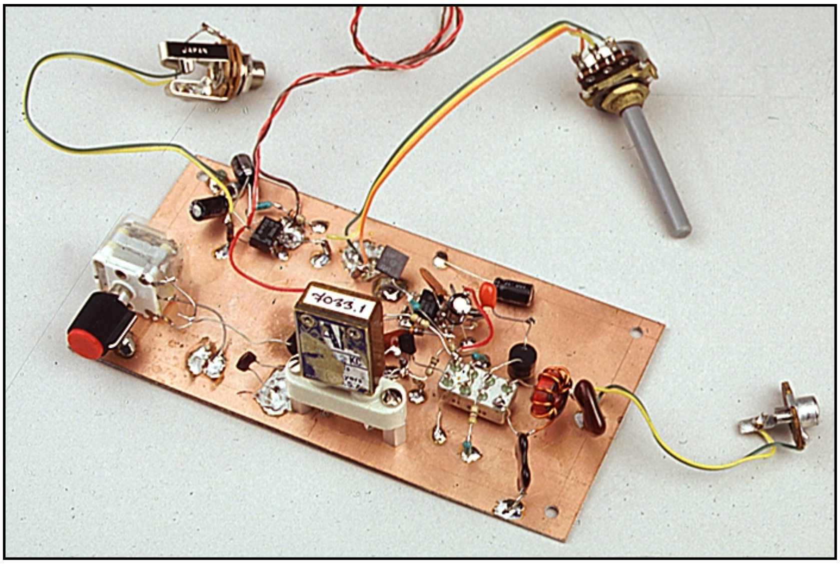


Fig 17.77—Ground-plane construction, PC-board construction—both of these approaches can produce the same good Rock-Bending Receiver performance. (WI5W built the one that looks nice, and W9VES—who wrote this caption—built the one that doesn't.)

to completely enclose the receiver—a flat aluminum panel screwed to a wooden base is an acceptable alternative. The panel supports the tuning capacitor, GAIN control and your choice of audio connector. The base can support the circuit board and antenna connector.

CHECKOUT

Before connecting the receiver to a power source, thoroughly inspect your work to spot obvious problems like solder bridges, incorrectly inserted components or incorrectly wired connections. Using the schematic (and PC-board layout if you built your receiver on a PC board), recheck every component and connection one at a time. If you have a digital voltmeter (DVM), use it to measure the resistance between ground and everything that should be grounded. This includes things like pin 4 of U2 and U3, pins 2, 5, 6 of U1, and the rotor of C7.

If the grounded connections seem all right, check some supply-side connections with the meter. The connection between pin 6 of U3 and the positive power-supply lead should show less than 1 Ω of resistance. The resistance between the supply lead and pin 8 of U1 should be about 47 Ω because of R1.

If everything seems okay, you can apply power to the receiver. The receiver will work with supply voltages as low as 6 V and as high as 13.5 V, but it's best to stay within the 9 to 12-V range. When first testing your receiver, use a current-limited power supply (set its limiting between 150 and 200 mA) or put a 150-mA fuse in the connection between the receiver and its power source. Once you're sure that everything is working as it should, you can remove the fuse or turn off the current limiting.

If you don't hear any signals with the antenna connected, you may have to do some troubleshooting. Don't worry; you can do it with very little equipment.

TROUBLE?

The first clue to look for is noise. With the GAIN control set to maximum, you should hear a faint rushing sound in the speaker or headphones. If not, you can use a small metallic tool and your body as a sort of test-signal generator. (If you have any doubt about the safety of your power supply, power the Rock-Bending Receiver from a battery during this test.) Turn the GAIN control to maximum. Grasp the metallic part of a screwdriver, needle or whatever in your fingers, and use the tool to touch pin 3 of U3. If you hear a loud scratchy popping sound, that stage is working. If not, then something directly related to U3 is the problem.

You can use this technique at U2 (pin 3, then pin 5) and all the way to the antenna. If you hear loud pops when touching either end of L3 but not the antenna connector, the oscillator is probably not working. You can check for oscillator activity by putting the receiver near a friend's transceiver (both must be in the same room) and listening for the VXO. Be sure to adjust the TUNING control through its range when checking the oscillator.

The dc voltage at Q1's base (measured without the RF probe) should be about half the supply voltage. If Q1's collector voltage is about equal to the supply voltage, and Q1's base voltage is about half that value, Q1 is probably okay. Reducing the value of L2 may be necessary to make some crystals oscillate.

OPERATION

Although the Rock-Bending Receiver uses only a handful of parts and its features are limited, it performs surprisingly well. Based on tests done with a Hewlett-Packard HP 606A signal generator, the receiver's minimum discernible signal (by ear) appears to be 0.3 µV. The author could easily copy 1-µV signals with his version of the Rock-Bending Receiver.

Although most HF-active hams use transceivers, there are advantages in using separate receivers and transmitters. This is especially true if you are trying to assemble a simple home-built station.

A WIDEBAND MMIC PREAMP

This project illustrates construction techniques used in the microwave region (at and beyond 1 GHz). It also results in a neat "dc to daylight" preamplifier with many uses around your shack, not the least of which is monitoring the downlinks from Amateur Radio satellites. The original article was written by William Parmley, KR8L, in Nov 1997 *QST*.

The preamplifier uses the MAR-6 monolithic microwave integrated circuit (MMIC) manufactured by Mini-Circuits Labs. The MAR-6 is a four terminal, surface mount device (SMD) with an operating frequency range from dc to 2 GHz, a noise figure of 3 dB, a gain of up to 20 dB, and input and output impedances of 50 Ω. The basic concept for the preamplifier and the construction techniques used to build it came from *The ARRL UHF/Microwave Experimenter's Manual*. The parts and circuit board material in this project are readily available from sources such as Ocean State Electronics.

CIRCUIT DESCRIPTION

Fig 17.78A is the schematic for the preamplifier. C1 and C2 are dc blocking capacitors. The device receives V_{cc} at the output lead, through RF choke L1 and limiting resistor R1. The only other components used are the bypass capacitors on the V_{cc} lead. C1 and C2 should present a low impedance at the lowest signal frequency of interest. The author designed his preamplifier for 435 MHz, a downlink frequency for many amateur satellites. Two 220 pF disc ceramic capacitors were used for C1 and C2. To use the preamplifier at 29 MHz for downlink signals from Russian RS-series satellites, C1 and C2 become 0.001 µF disc ceramic capacitors.

The power-supply voltage determines R1's value. The MAR-6 draws about 16 mA, and needs a V_{cc} of about 3.5 V. Use Ohm's Law to calculate the necessary voltage drop from your power supply voltage down to 3.5 V. The author's power supply provided about 14.6 V, so a 680 W, $^1/_2$-W resistor was used for R1. RF choke L1 helps isolate the power supply from the MMIC output. L1 is a homemade 0.12 mH choke, consisting of 8 turns of #30 enameled wire around the shank of a $^3/_{16}$ inch drill bit, spaced for a total length of 0.3 inches. (Remove the drill bit; it's only a winding mandrel!) This value of L1 was left in place when the preamplifier was used at lower frequencies.

The remaining three essential parts are bypass capacitors. Because capacitors have self-resonant frequencies (resulting from unavoidable inductances in the devices and their leads), it is a common practice to use capacitors of several different values in parallel. This design uses a 0.001 µF feedthrough capacitor passing through the circuit board ground plane. The parallel 0.01 µF and 0.1 µF capacitors are disc ceramics.

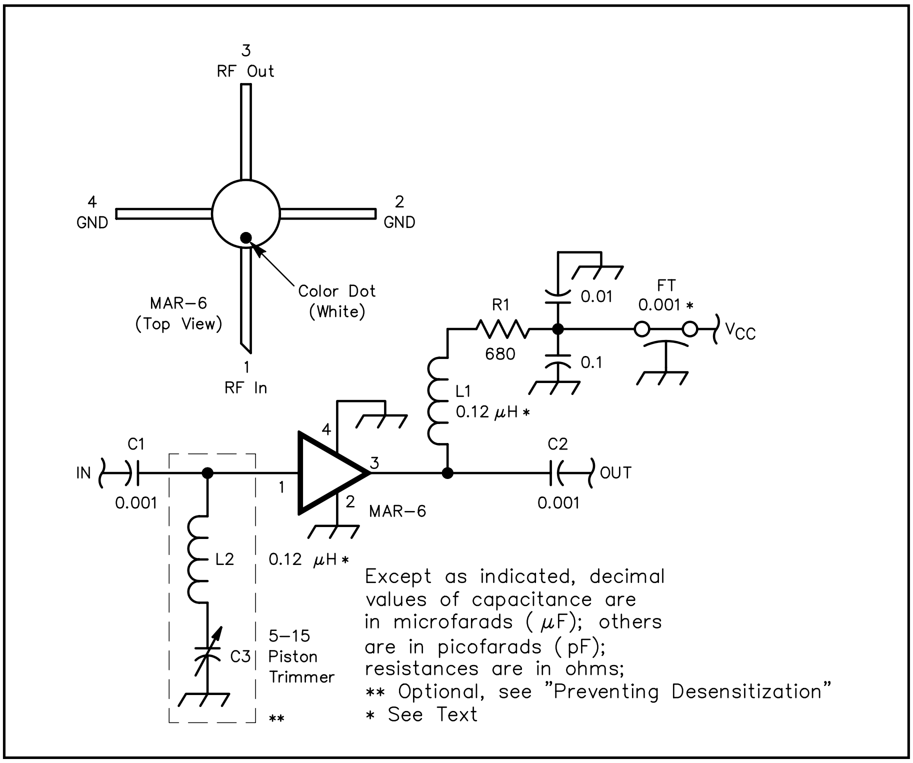


Fig 17.78A—A schematic of the preamp circuit. Equivalent parts may be substituted.

L2 and C3 are optional components for 432 MHz, used to provide some selectivity against desensitization when transmitting on 144 MHz for satellite Mode J, at the expense of wideband coverage, of course.

CIRCUIT CONSTRUCTION

Fig 17.78B shows the circuit-board layout. The material used is double-sided, glass-epoxy board with a thickness of 0.0625 inches, known as FR-4 or G-10. This is the least expensive board material suited for microwave use. (The board I used is a product of GC/Thorsen in Rockford, Illinois.) Notice that most of the top of the board, and all the bottom side of the board, serves as circuit ground.

The signal-conducting part of the circuit is a “microstrip.” (That is a strip-type transmission line: a conductor above or between extended conducting surfaces.—*Ed.*) The line width, board thickness and board dielectric constant determine the microstrip’s characteristic impedance. A 0.1-inch-wide line and the ground plane on 0.0625-inch-thick G-10 form a 50-Ω transmission line, which matches the MMIC’s input and output impedance.

The author fabricated his board by laying out the traces with a machinist’s rule. Then he cut through the copper foil with a knife and lifted off the unwanted copper areas while heating them with a 100-W soldering gun. You could etch the board if you prefer, or use any other method you like. The single mounting pad was “etched” by grinding away the copper with a hand-held grinder.

The MMIC is tiny. Connect it to the traces with the shortest possible distances between the traces and the body. (The author managed to achieve about 0.03 inch.) Also, the device leads are very delicate—if possible, do not bend them at all. To fit the MMIC leads flat on the PC-board traces without bending, a small depression was ground in the board dielectric for the MMIC body. Remember that, viewed from the top, the colored dot (white on the MAR-6) on the body marks pin 1, the input lead. The other leads are numbered counterclockwise; pin 3 is the output lead.

Mount the blocking capacitors as close to the board as possible. To do this, the capacitor leads were cut to about $^{1}/_{16}$ inch long. Both the capacitor leads and circuit traces were tinned and then the capacitors were soldered in place. This method of mounting minimizes lead inductance.

The author installed N connectors for his unit. To achieve a “zero lead length,” he notched the ends of the board to fit the profile of the connectors and installed the connectors directly to the board. The center pin was laid on top of the microstrip and soldered. Then the connector body was soldered to the ground foil in four places: two on the top of the board and two on the bottom. Another very good technique is to drill a hole in the microstrip and insert the center pin from the bottom of the board. The center pin is then soldered to the microstrip, and the body is soldered to the ground foil or mounted with machine screws. (If you do this, be sure to remove a bit of foil from around the hole on the bottom side so the center pin

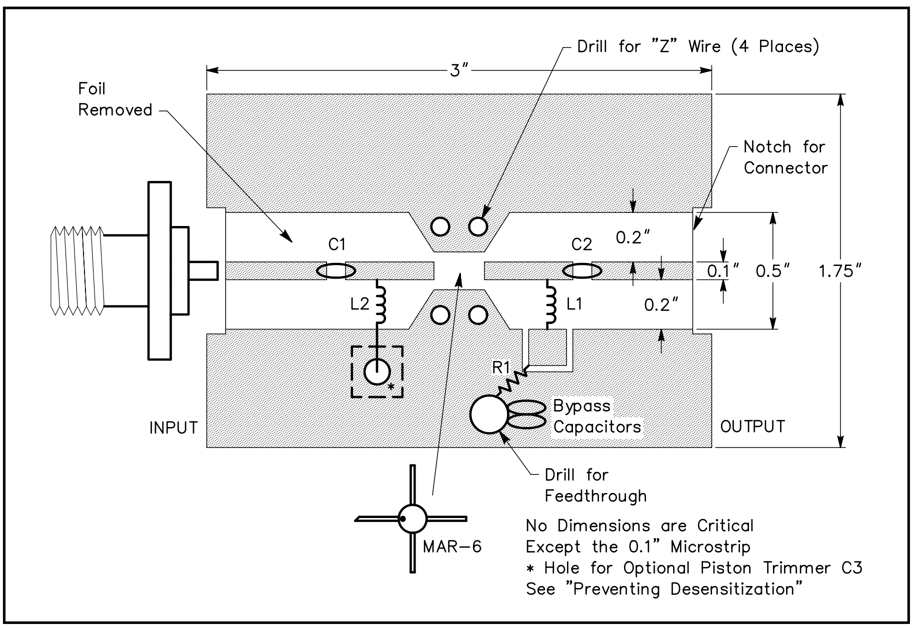


Fig 17.78B—A part-placement diagram for KR8L’s MMIC preamp. Dark areas are copper on the component (near) side. The reverse is a copper ground plane.

doesn't short to ground.) The latter approach is much better if you want to mount the preamplifier into a box. You can mount the board on the inside of the lid with the connectors projecting through.

It is important that all portions of the ground foil be at equal potential, particularly near the MMIC and the board edges. To achieve this, wrap the long edges of the board with pieces of 0.003-inch-thick brass shim stock and solder them on both top and bottom. Thinner or thicker material is suitable (up to about 0.005 inch), as is copper flashing. Two small holes were drilled on either side of each MMIC ground lead, and a small Z-shaped wire was soldered to each side of the board. (A Z wire is a short, small-gauge, solid copper wire bent 90º, inserted through the hole, bent 90º again and soldered on both sides of the board.)

The inductor is also mounted using minimal-length leads. One lead connects to the microstrip and the other to the square pad. The resistor connects from the pad to the feedthrough capacitor, and the other two bypass capacitors connect from the feed-through to the ground foil.

HOOKUP AND OPERATION

For the basic preamplifier design there is nothing to align or adjust. Simply connect the preamplifier between your antenna and receiver and apply power. If you connect the preamp to a transceiver, take precautions to prevent transmitting through the preamp! This preamplifier is very handy for many uses: adding gain to an older 10-meter receiver or scanner, boosting signal-generator output or for casual monitoring of the Amateur Radio satellites on 29, 145 and 435 MHz. A commercial metal box, home-made PC board or thin sheet-metal boxes make suitable cabinets for this project.

A BINAURAL I-Q RECEIVER

This little receiver was designed and built by Rick Campbell, KK7B. It was first described in the March 1999 issue of *QST*. It replaces the narrow filters and inteference-fighting hardware and software of a conventional radio with a wide-open *binaural I-Q detector*. If you liken a conventional receiver to a high-powered telescope, this receiver is a pair of bright, wide-field binoculars. The receiver's classic junk-box-available-parts construction approach achieves better RF integrity than that of much commercial ham gear. A PC board and parts kit is available for those who prefer to duplicate a proven design.[1]

A receiver with presence . . . to fully appreciate this receiver, you've got to hear it! "Once my ears got used to the effect, they had to drag me away from this radio. This is one I gotta have!"
—Ed Hare, W1RFI, ARRL Lab Supervisor

The total construction time was only 17 hours. There are a number of toroids to wind, and performance was not compromised to simplify construction or reduce parts count.

BINAURAL I-Q RECEPTION

Modern receivers use a combination of band-pass filters and digital signal processing (DSP) to select a single signal that is then amplified and sent to the speaker or headphones. When DSP is used, the detector often takes the form shown in **Fig 17.79**. The incoming signal is split into two paths, then mixed with a pair of local oscillators (LOs) with a relative 90° phase shift. This results in two baseband signals: an in-phase, or *I* signal, and a quadrature, or *Q* signal. Each of the two baseband signals contains all of the information in the upper and lower sidebands. The baseband pair also contains all of the information needed to determine whether a signal is on the upper or lower sideband before multiplication. An analog signal processor consisting of a pair of audio phase-shift networks and a summer could be used to reject one sideband. In a DSP receiver, the *I* and *Q* baseband signals are digitized and the resulting sets of numbers are phase-shifted and added.

The human brain is a good processor for infor-

[1] The complete kit version, available from Kanga US, uses a different VFO circuit than the one shown here. The kit VFO runs at one-half the desired output frequency, and is followed by a balanced frequency doubler and driver amplifier.

Steel chassis such as the Hammond 1441-12 (2 × 7 × 5 inches [HWD]) with 1431-12 bottom plate and the Hammond 1441-14 (2 × 9 × 5 inches [HWD]) with 143-14 bottom plate are suitable enclosures. These chassis and bottom plates are not available in single quantities directly from Hammond, but are available from Allied Electronics and Newark Electronics. See the **References** chapter for contact information.

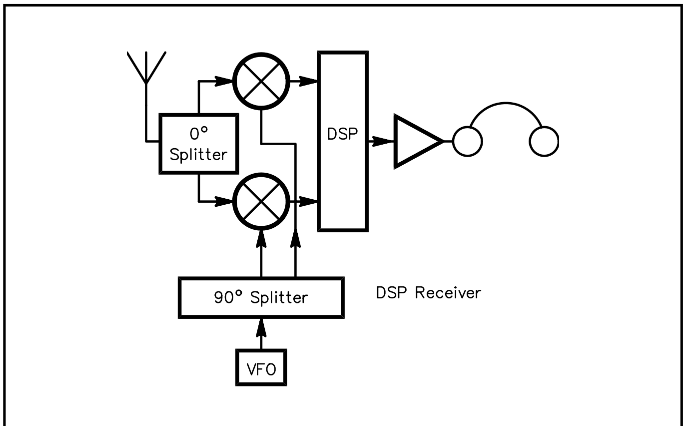


Fig 17.79—The simplified block diagram of a receiver using a DSP detector; see text.

mation presented in pairs. We have two eyes and two ears. Generally speaking, we prefer to observe with both eyes open, and listen with both ears. This gives us depth of field and three-dimensional hearing that allows us to sort out the environment around us. The ear/brain combination can be used to process the output of the I-Q detectors as shown in **Fig 17.80**.

The sound of CW signals on a binaural I-Q receiver is like listening to a stereo recording made with two identical microphones spaced about six inches apart. The same information is present on each channel, but the *relative phase* provides a stereo effect that is perceived as three-dimensional space. Signals on different sidebands—and at different frequencies—appear to originate at different points in space. Because SSB signals are composed of many audio frequencies, they sound a little spread in the perceived three-dimensional sound space. This spreading also occurs with most sounds encountered in nature, and is pleasant to hear.

To keep the receiver as simple as possible, a single-band direct-conversion (D-C) approach is used. A crystal-controlled converter can be added for operation on other bands, changing the receiver to a single-conversion superhet. Alternatively, the binaural I-Q detector

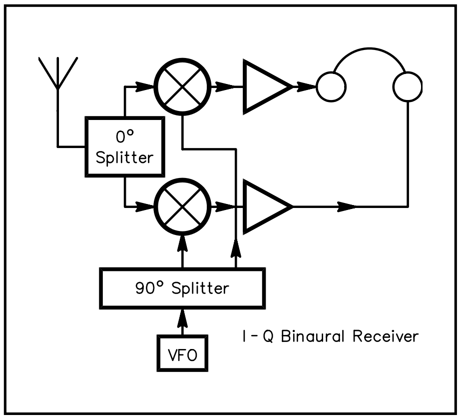


Fig 17.80—The block diagram of a binaural I-Q receiver that allows the ear/brain combination to process the detector output, resulting in stereo-like reception.

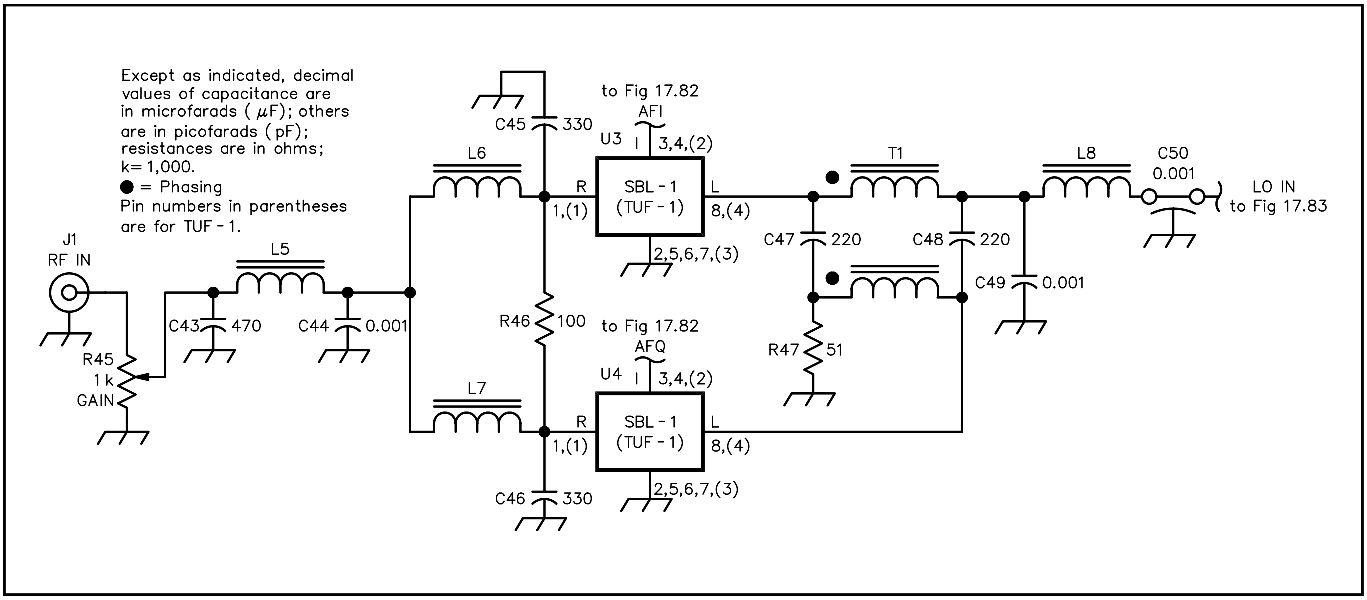


Fig 17.81—This diagram shows the front end and *I* and *Q* demodulators of the Binaural Weekender receiver. Unless otherwise specified, resistors are $^{1}/_{4}$ W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted. Pin connections for the SBL-1 and TUF-1 mixers at U3 and U4 are shown; the TUF-1 pin numbers are in parentheses. A kit is available (see Note 1). Parts are available from several distributors including Digi-Key Corp, Mouser Electronics and Newark Electronics. See the References chapter for contact information.

C43—470 pF disc ceramic
C44, C49—0.001 µF metal polyester
C45, C46—330 pF disc ceramic
C47, C48—220 pF disc ceramic
C50—0.001 µF feed-through capacitor
J1—Chassis-mount female BNC connector
L5—1.6 mH, 24 turns #28 enameled wire on T-30-6 powdered-iron core
L6, L7—1.3 µH, 21 turns #28 enameled wire on T-30-6 powdered-iron core
L8—350 nH, 11 turns #28 enameled wire on T-30-6 powdered-iron core
R45—1 kΩ panel-mount pot
T1—17 bifilar turns #28 enameled wire on T-30-6 powdered-iron core
U3, U4—Mini-Circuits SBL-1 or TUF-1 mixer

can be used in a conventional superhet, with a tunable first converter and fixed-frequency BFO. If proper receiver design rules are followed, there is no advantage to either design over the other.

THE RECEIVER

Figs 17.81, 17.82 and 17.83 show the complete receiver schematic. In **Fig 17.81**, signals from the antenna are connected directly to a 1 kΩ GAIN pot on the front panel. J1 is a BNC antenna connector, popular with QRP builders. Adjusting the gain before splitting the signal path avoids the need for a two-gang volume control, and eliminates having to use separate RF and AF-gain adjustments. This volume-control arrangement leaves the "stereo background noise" constant and varies the signal-to-noise ratio. The overall gain is selected so that the volume is all the way up when the band is quiet. Resistor values R9 and R31 may be changed to modify the overall gain if required. After the volume control, the signal is split with a Wilkinson divider and connected to two SBL-1 diode-ring mixers. (The TUF-1 is a better mixer choice, but I had more SBL-1s in my junk box.) The VFO signal is fed to the two mixers through a quadrature hybrid, described by Reed Fisher.[2] All of the circuitry under the chassis is broadband, and there are *no* tuning adjustments.

The audio-amplifier design of **Fig 17.82** is derived from that used in the R1 High-Performance Direct-Conversion Receiver,[3] with appropriate simplifications. The R1 high-power audio output is not needed to drive headphones, the low-pass filter is eliminated, and the diplexer has fewer components. Distortion performance is not compromised—well over 60 dB of in-band two-tone dynamic range is available. The original article, and the additional notes in Technical Correspondence for February 1996,[4] describe the audio-amplifier chain in detail.

THE VFO

Fig 17.83 is the schematic of the receiver VFO, a JFET Hartley oscillator with a JFET buffer amplifier. Components for the VFO tuned circuit are chosen for linear tuning from 7.0 to 7.3 MHz with the available junk-box variable capacitor. Setting up the VFO is best done with a frequency counter, receiver and oscilloscope. The frequency counter makes it easy to select the parallel NP0 capacitors and squeezing and spreading the wire turns on L1 achieves the desired tuning range. After the tuning range is set, listen to the VFO signal with a receiver to make sure the VFO

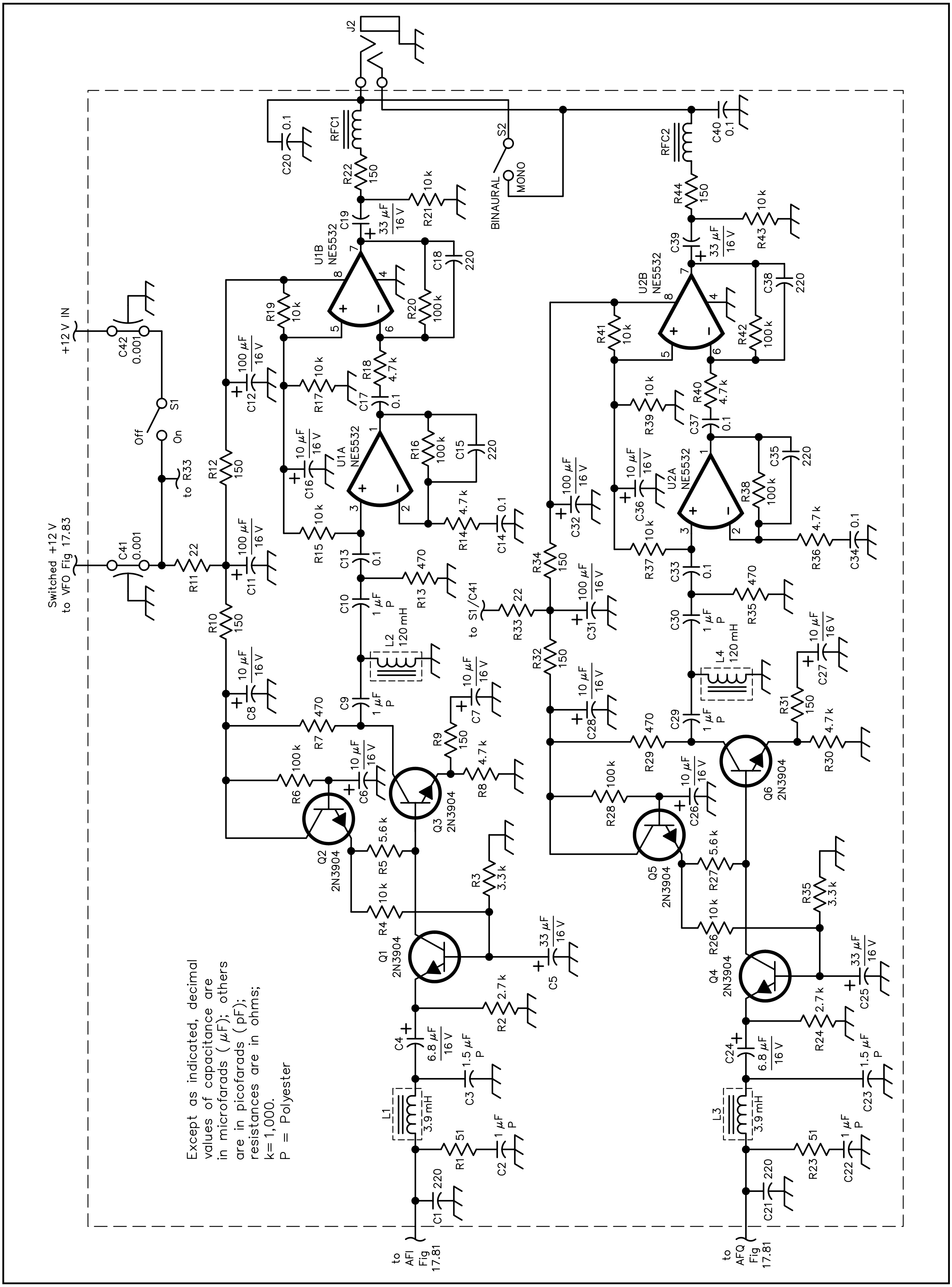
[2] Reed Fisher, W2CQH, "Twisted-Wire Quadrature Hybrid Directional Couplers," *QST*, Jan 1978, pp 21-23. See also *IEEE Transactions MTT*, Vol MTT-21, No. 5, May 1973, pp 355-357.

[3] Rick Campbell, KK7B, "High-Performance Direct-Conversion Receivers," *QST*, Aug 1992, pp 19-28.

[4] Rick Campbell, KK7B, "High-Performance, Single-Signal Direct-Conversion Receivers," *QST*, Jan 1993, pp 32-40. See also Feedback, *QST*, Apr 1993, p 75.

Fig 17.82—This diagram shows the receiver audio-amplifier design.

C1, C15, C18, C21, C35, C38—220 pF disc ceramic
C2, C9, C10, C22, C29, C30—1 µF metal polyester (Panasonic ECQ-E(F) series)
C3, C23—1.5 µF metal polyester (Panasonic ECQ-E(F) series)
C4, C24—6.8 µF, 16 V electrolytic (Panasonic KA series)
C5, C19, C25, C39—33 µF, 16 V electrolytic (Panasonic KA series)
C6, C7, C8, C16, C26, C27, C28, C36—10 µF, 16 V electrolytic (Panasonic KA series)
C11, C12, C31, C32—100 µF, 16 V electrolytic (Panasonic KA series)
C13, C14, C17, C20, C33, C34, C37, C40—0.1 µF metal polyester (Panasonic V series)
C41, C42, C50—0.001 µF feed-through capacitor
J2—1/8-inch stereo phone jack
L1, L3—3.9 mH Toko 10RB shielded inductor
L2, L4—120 mH Toko 10RB shielded inductor
Q1 through Q6—2N3904
RFC1, RFC2—10 turns #28 enameled wire on Amidon ferrite bead FB 43-2401 (six-hole bead)
S1, S2—SPST toggle switch
U1, U2—NE5532 dual low-noise high-output op amp


Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k= 1,000.
P = Polyester
Switched +12 V to VFO Fig 17.83
+12 V IN
to AFI Fig 17.81
to AFQ Fig 17.81
BINAURAL
MONO

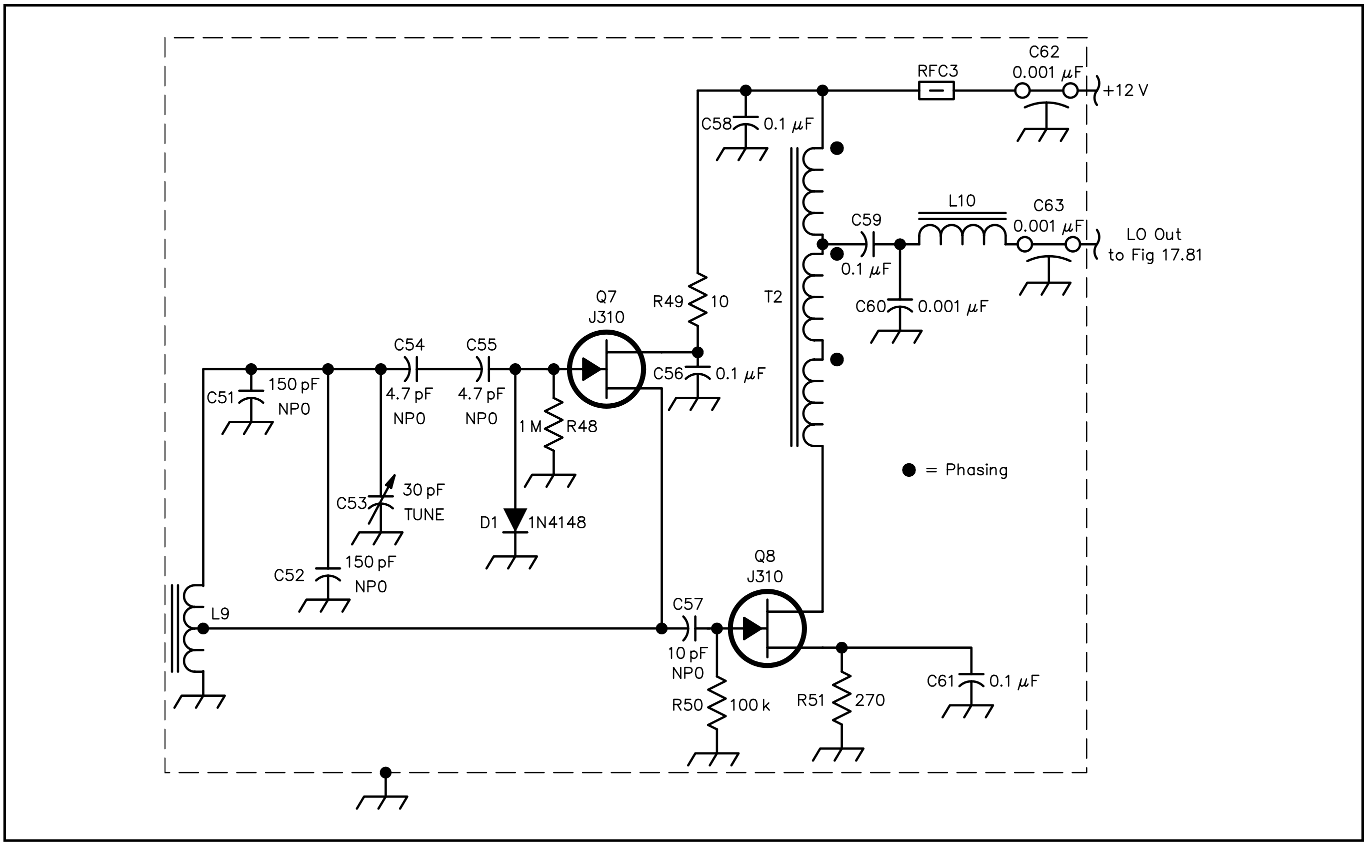


Fig 17.83—The diagram shows the prototype binaural receiver's VFO. The LO output is +10 dBm. This simple VFO works exceptionally well, but must be completely shielded for good D-C receiver performance. A receiver with an open PC-board VFO will work better if the variable oscillator is not running on the received frequency. As noted elsewhere, the kit version of the receiver uses a different VFO.

C51, C52—150 pF, NP0 disc ceramic
C53—30 pF air-dielectric variable
C54, C55—4.7 pF NP0 disc ceramic
C56, C57, C59, C61—0.1 µF metal polyester (Panasonic V series)
C57—10 pF NP0 disc ceramic
C60—0.001 pF metal polyester
C62, C63—0.001 µF feed-through capacitor
D1—1N4148
L9—1.5 mH, 22 turns #22 enameled wire on T-37-6 powdered-iron core; tap 5 turns from ground end
L10—350 nH, 11 turns #28 on T-30-6 powdered-iron core
Q7, Q8—J310 (U310 used in prototype)
RFC3—10 turns #28 enameled wire on Amidon ferrite bead FB 43-2401 (six-hole bead used in prototype)
T2—10 trifilar turns #28 enameled wire on Amidon ferrite bead FB 43-2401 (six-hole bead used in prototype)

tunes smoothly and has a good note. Interrupt the power to hear its start-up chirp. The signal may sound ratty with the frequency counter on, so turn it off. The VFO is one area where craftsmanship pays off. Solid construction, a self-aligning variable-capacitor mounting, complete RF and air shielding and good capacitor bearings all contribute to a receiver that is a joy to tune.

Both connections to the VFO compartment are made with feed-through capacitors. The power supply connection is self-explanatory, but passing RF through a feed-through capacitor (at LO Out) may seem a bit unusual. Electrically, the capacitor is one element of a low-pass pi network. Using feed-through capacitors keeps local VHF signals (high-powered FM broadcast and TV signals near my location) out of the VFO compartment. A second pi network feeds the VFO signal to the detector circuit below the chassis. The use of VHF construction techniques in a 40 meter receiver may seem like overkill, but the present KK7B location is line-of-sight to broadcast towers serving the Portland, Oregon area. Using

What Do You Hear?

Even the earliest solid-state direct-conversion (D-C) receivers had a *presence* or *clarity* that is rarely duplicated in more elaborate receivers. Many of us remember the first time we heard this crispness in a "homebrewed" D-C receiver. As we try to "enhance" our rigs through the addition of IF filters and other "features," we still hope that the result will be as clean as that first D-C receiver.

This binaural D-C receiver is such an experience—but even better. The binaural processing supplies the ears with additional information without compromising what was already there, enhancing the presence.

As you tune through a CW signal on a quiet band (best done with your eyes closed while sitting in a solid chair), a centered signal enters, but moves to the left background, undergoes circular motions at the back of your head as you tune through zero beat, repeats the previous gyrations on the right side, fades to the right background, and finally drops away in the center. Multiple signals within the receiver passband are distributed throughout this perceived space. With training, concentration on one signal allows it to be copied among the many. An SSB signal seems to occupy parts of the space, left and right, with clarity when properly tuned, leaving others vacant. Static crashes and white noise appear distributed throughout the entire space without well defined position. Receiver noise, although present, has no perceived position.

It's vital that this receiver include a front-panel switch to shift between binaural and monaural output. Although useful during the learning process, it becomes indispensable for the demonstrations that you will want to do. I used the switch to set up my son, Roger, KA7EXM, for the experience. We entered the shack and I handed him the headphones. He put one phone to just one ear, but I told him that he had to use both, that it would not work with just one. He put the phones on his head, casually tuned the receiver through the 40-meter CW band, removed the phones and commented, "Well, it sounds just like a direct-conversion receiver: A good one, but still just a direct-conversion receiver." I smiled and asked him to put the headphones on again. As I flipped the switch to the binaural position his hand reached out, seeking the support of the workbench. His facial expression became more serious. He eased into the chair and began tuning the receiver, very slowly at first. After a minute he took the headphones off, but remained speechless for a while—an unusual condition for Roger. Finally, he commented, "Wow! The appliance guys have never heard that!"

A builder of the Binaural Weekender should prepare for some truly unusual experiences.—*Wes Hayward, W7ZOI*

commercial HF gear with conventional bypassing under these circumstances provided disappointing results.

The lead photo shows the prototype receiver front panel. Receiver controls are simple and intuitive. The ear/brain adjusts so naturally to binaural listening that I added a BINAURAL/MONO switch to provide a quick reminder of how signals sound on a conventional receiver. The switch acts much like the STEREO-MONO switch on an FM broadcast receiver—given the choice, it always ends up in the STEREO position!

The uses a pair of Koss SG-65 headphones with his receiver. They are not necessary, but have some useful features. First, at about $32, they are relatively inexpensive. Second, they have relatively high-impedance drivers, (90 Ω) so they can be driven at reasonable volume directly from an op amp. Finally, they make an attempt at low distortion. Other headphones in the same price bracket are acceptable, but some have much lower impedance and won't provide a very loud audio signal using the component values given in the schematic. Those $2.95 bubble-packed, throw-away headphones are not a good choice! Audiophile headphones are fine, but don't really belong on an experimenter's bench. A stray clip-lead brushing across the wrong wire in the circuit can instantly burn out a driver and seriously ruin your day.

BUILDING A BINAURAL WEEKENDER

A few construction details are generally important, while others were determined by the components that happened to be in my junk box. The big reduction drive is delightful to use, but doesn't contribute to electrical performance. I purchased it at a radio flea market. The steel chassis provides a significant reduction in magnetic hum pickup, something that can be a problem if the receiver is operated near a power transformer. (Steel chassis are available from parts houses that cater to audio experimenters.) The VFO mounting and mushroom-can shield shown in **Fig 17.84** are a simple way to eliminate mechanical backlash, keep radiated VFO energy off the antenna, prevent hand capacitance from shifting the tuning, and reduce VFO drift caused by air currents.

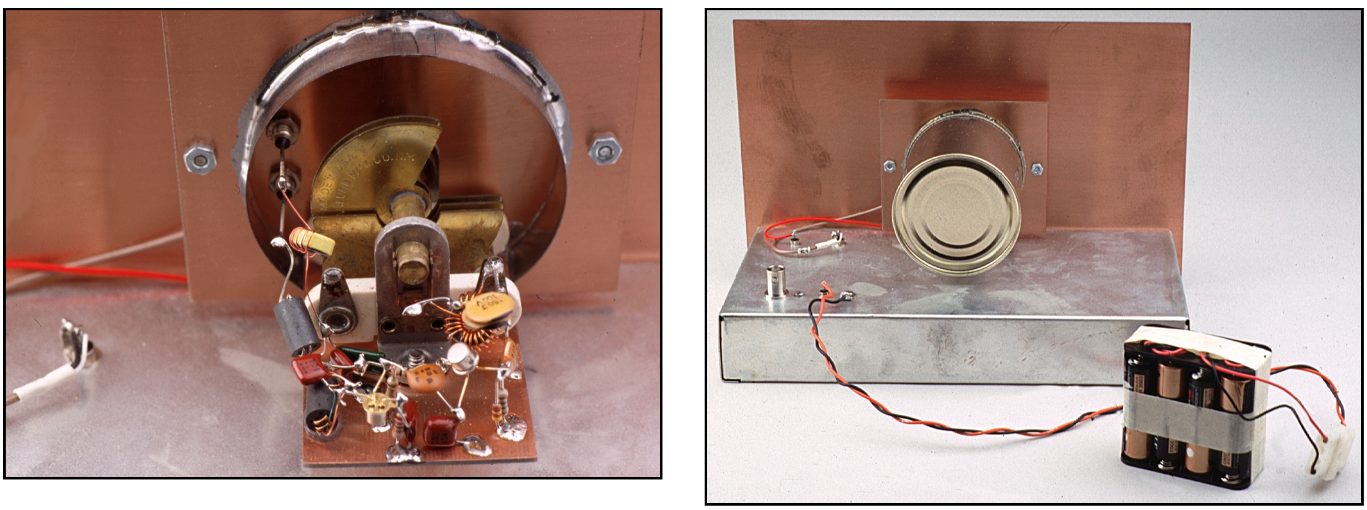
Experienced builders can duplicate this receiver simply using the schematic and construction techniques described here. Unlike a phasing receiver, there is no need to precisely duplicate the exact amplitudes and phases between the two channels. The ear/brain combination is the ultimate adaptive processor, and it quickly learns to focus on a desired signal and ignore interference. Small errors in phase and amplitude balance are heard as slight shifts in a signal's position. Standard-tolerance components may be used throughout.

One note about the kit version: A very good VFO can be built on an open PC board if the variable oscillator is not running on the desired output frequency. The Kanga kit VFO runs at one-half the desired frequency, and is followed by a balanced frequency doubler and driver amplifier.

OTHER EXPERIMENTS

My earliest experiments with binaural detectors feeding stereo audio amplifiers were done in 1979, using two antennas. The technique works very well, but requires two antennas either physically spaced some distance apart, or of different polarization. Listening to OSCAR 13 on a binaural receiver with cross-polarized Yagis was an unsettling experience. The need for two antennas is a liability—these days most of us struggle to put up one. A number of experiments have also been done with binaural independent sideband (ISB) reception. These are profoundly interesting for AM broadcast reception, and could be used for amateur AM or DSB reception using a Costas Loop for carrier recovery. Binaural ISB detection of shortwave AM broadcasting can be analyzed as a form of spread spectrum with the ear/brain combination serving the despreading function, or as a form of frequency diversity, with the ear/brain as an optimal combiner.

The binaural techniques described here are analogous to binocular vision: They present the same



(A) (B)

Fig 17.84 — A shows a close-up of the VFO. The simple VFO used in the prototype works exceptionally well, but must be completely shielded for good D-C receiver performance. B shows how an empty mushroom can lives again as a VFO shield in the prototype receiver.

information to each ear, but from a slightly different angle. This provides a very natural sound environment that the brain interprets as three-dimensional space. There are other “binaural” techniques that involve the use of different filter responses for the right and left ears. My experiments with different filter responses for the left and right ears have not been particularly interesting, and I have not pursued them.

SUMMARY

This little receiver is a joy to tune around the band. It is a serious *listening* receiver, and allows digging for weak signals in a whole new way. Digging for weak signals in a three-dimensional sound field is sometimes referred to as the “cocktail party effect.” It is difficult to quantify the performance of a binaural receiver, because the final signal processing occurs in the brain of the listener—you. The experimental literature of psycho-acoustics suggests that the ear/brain combination provides a signal-to-noise advantage of approximately 3 dB when listening to speech or a single tone in the presence of uncorrelated binaural noise. The amount of additional noise in the opposite sideband is also 3 dB, so it appears that the binaural I-Q detector breaks even. In some applications, such as UHF weak-signal work, the binaural I-Q detector may have an advantage, as it permits listening to a larger slice of the band without a noise penalty. In other situations, such as CW sweepstakes, the “cocktail party” may get entirely out of hand. Binoculars and telescopes both have their place.

References

Campbell, Rick, KK7B, “Direct Conversion Receiver Noise Figure,” Technical Correspondence, *QST*, Feb 1996, pp 82-85.

Campbell, Richard L., “Adaptive Array with Binaural Processor,” *Proceedings of the IEEE Antennas and Propagation Society International Symposium*, Philadelphia, PA, June 1986, pp 953-956.

Campbell, Rick, KK7B, “Binaural Presentation of SSB and CW Signals Received on a Pair of Antennas,”*Proceedings of the 18th Annual Conference of the Central States VHF Society*, Cedar Rapids, IA, July 1984, pp 27-33.

A SUPERREGENERATIVE VHF RECEIVER

Introduction

The complexity of many published receiver circuits has "scared off" some would-be home builders. Yet, making a receiver from scratch is an extremely rewarding experience. Charles Kitchin, N1TEV, put together this VHF receiver that can easily be built by the average person, and does not require any special components or test equipment. It covers roughly 118-170 MHz, and receives AM, WBFM and NBFM.

Don't expect to squeeze out the ultimate in selectivity or stability with this receiver, although it does provide a sensitivity of around 1 mV. It uses the principle of superregeneration for this high sensitivity with a low parts count. This design differs from other superregenerative circuits with an addition of a *quench waveform* control, which greatly helps to improve selectivity for NBFM reception. Parts cost should be less than $20, and the receiver is powered by a standard 9-V battery.

This simple circuit is not going to compete with your modern digitally synthesized, multi-programmable transceiver but it is very useful (and fun) for monitoring your local 2-meter repeater, the aircraft band (118-137 MHz), marine radio, weather, police, snowplows, fire stations, telephone paging systems and many other types of local communications. As with any simple receiver, a certain amount of practice and patience is required in learning how to tune and adjust the set.

Regeneration and Superregeneration

Regenerative detectors are basically oscillators to which an input signal has been coupled. In a straight regenerative circuit, the input signal is coupled to the detector and then *regenerated* to very high levels by feeding back a portion of the output signal, in phase, back to its input, until just before or just at a critical point where a self sustaining oscillation begins. After that point, the circuit's gain stops increasing and starts going down, as most of the transistor's energy is devoted to generating its own oscillations.

The regeneration control is included to allow the operator to maintain the feedback at a point close to oscillation. With modern components, practical circuit gains of 20,000 are easily achieved in a single stage.

The superregenerative circuit uses an oscillating regenerative detector, which is periodically shut off or *quenched*. Superregeneration allows the input signal to be regenerated over and over again, providing single-stage circuit gains of close to 1 million, even at UHF. Note the oscillations must be completely quenched each time, before being allowed to start-up again. Superregenerative detectors can use either a separate lower frequency oscillator to interrupt the detector (separately quenched) or as in this design, a single JFET is used to produce both oscillations (self quenched circuit).

Circuit Details

The receiver circuit shown in **Fig 17.86** uses a JFET RF stage with a separate dc return path for RF stage and detector. This uses a few more components than simply soldering a tap onto the coil, but it reduces the current drawn through the regeneration control. This allows the use of a Zener voltage regulator for the detector, which reduces the frequency drift due to changes in battery voltage.

The dual gang variable capacitor and toggle switch controls the band selected. A single section is used for the high band and the second section is simply switched in for low-band reception. The switch was soldered right onto the hot (not grounded) terminal of the tuning capacitor. A series (mica) capacitor, with short leads was used to reduce the second section's total capacitance and set the desired frequency range of the low band.

Construction Details

Since this is a superregenerative receiver used at very high frequencies, stray circuit capacitances and

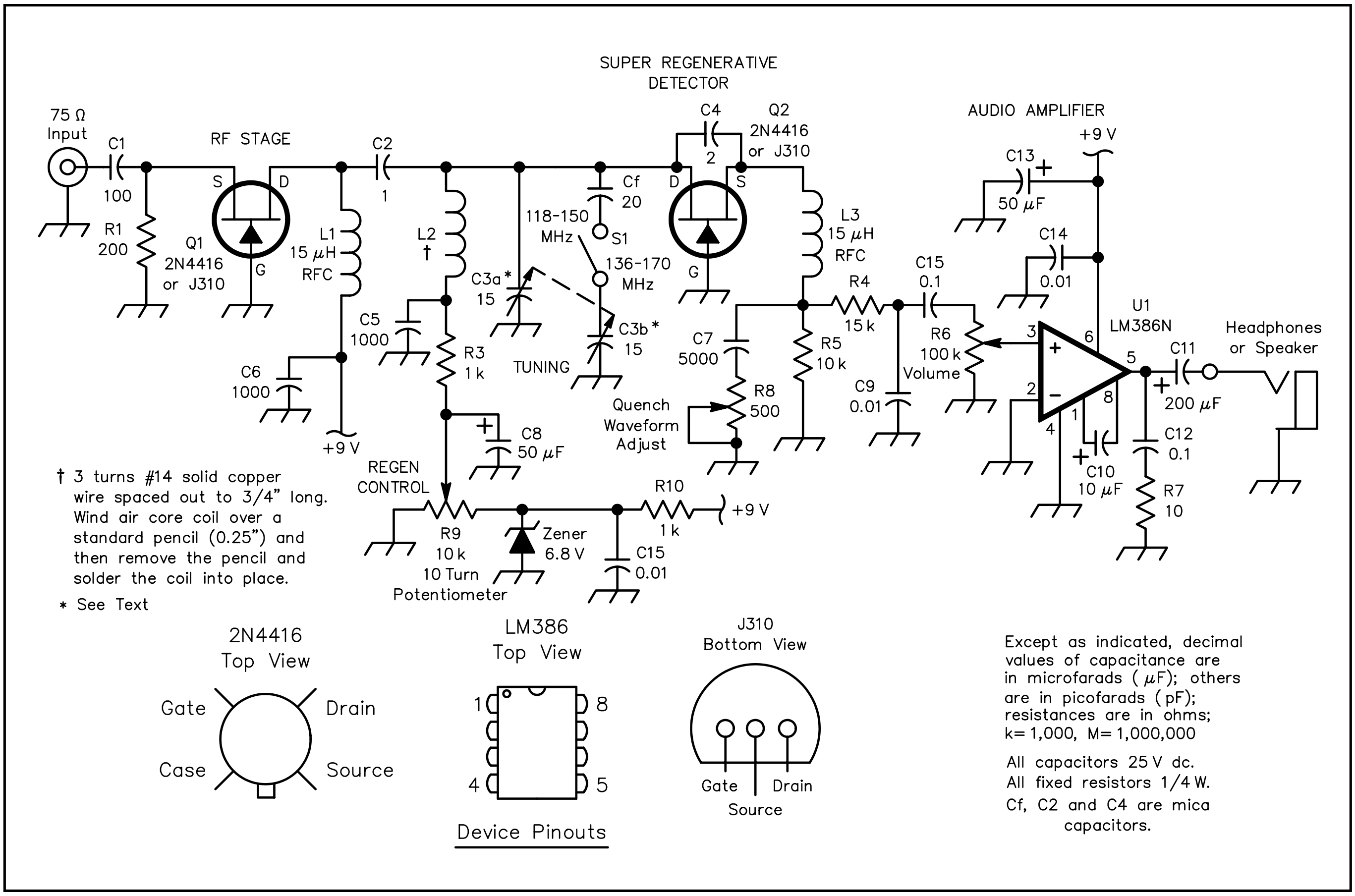


Fig 17.86—Schematic diagram of the receiver. C3 is a dual section, air-variable capacitor. C2 can be a fixed commercial capacitor, or a pair of twisted, insulated wires (see text). Cf can be increased to lower the low-frequency tuning range or decreased to increase this range. Compressing L2 will lower both tuning ranges.

multiple ground paths can prevent the detector from oscillating. It is important that the superregenerative detector's tuning coil be physically located away from other objects—particularly chassis ground, the bottom and sides of the equipment box, if it is metal, and any shielding that may be present.

Avoid printed circuit boards: with all their components (especially the main tuning coil) mounted very close to the ground plane, the detector will be loaded down and may fail to oscillate properly, if it all. Instead, just use a piece of copper clad board and some standoffs. Suspend the components above the board on the standoffs or use the parts that have leads grounded to hold the other components above the board (this is known as the *dead bug* or *ugly* construction method). The parts are readily available from most small parts suppliers listed in the **References** chapter.

Put the copper board inside a small box or use a block of wood and a piece of metal for the front panel. If you plan on placing the entire receiver inside a closed metal box, first build the circuit outside the box and be sure it oscillates before placing it inside.

Mount the tuning capacitor onto the copper board and pass its shaft through an oversized hole in the front panel. Mount all the other controls (except S1) to the front panel and connect them to the board using the shortest leads possible.

Receiver circuits are usually built backwards. Start with the audio stage. Wire it up as far back as the volume control and then test the stage by turning the control up halfway and placing your finger on the wiper (listen for a buzz). If it's not working, recheck the wiring or use a voltmeter to troubleshoot the problem. Be sure the supply voltage is present and that the voltage on pin 5 of the LM386 is at half supply.

After the audio stage is working, wire the detector and RF stage, but don't connect the two stages together (leave-out capacitor C2 in the schematic). Be sure to locate Q1 very close to L1 since C2 needs to be connected with very short leads. Now, with no load on the detector, set R8 to mid position and turn up the regeneration control, R9, until oscillation occurs. You should hear a loud rushing noise. This indicates the detector is superregenerating.

L3 and C4 are the only components with critical values. Since your layout and construction will be different from the author's, some experimentation may be needed to get the detector to oscillate. Start with the values shown in the figure. Try changing the value of L3 first, followed by C4. Be sure the detector oscillates over the entire tuning range. R8 and R9 also affect oscillations. If you have any dead spots, you may want to move L2 and L3 farther away from ground. Once the values are set, no further changes should be needed. The C2 value shown, 1 pF, will work fine, but if you want to optimize the performance you will want to vary this. You also can use a 5-pF capacitor and tap down on the coil for best performance. Alternately, you can try using a pair of twisted, insulated wires for C2. Start with 4 turns, and increase the number until it sounds as though you are affecting the detector's oscillations.

Tuning the Receivers

For optimum sensitivity from these receivers, use "fresh" 9-V batteries. The frequency range can be adjusted by adding 10 or 15 pF in parallel with C3a (to move both the high range and low range down), in parallel with C3b (to move only the high range down) or reducing Cf to move the high range up. The turns on the main tuning coil, L1b, can be compressed or expanded to change both bands.

In these self-quenched circuits, the regeneration control also varies the quench frequency rate. For AM and wide-band FM reception this is a handy feature, since the operator only needs to adjust a single control. The setting for optimum sensitivity can usually be found by simply advancing the control past the detector's oscillation threshold and then to a point just below where the background (mush) noise suddenly begins to increase rapidly.

But for narrow-band FM reception, the second control is needed. Set R8 at midscale, then adjust the regeneration control for strong oscillation (high sensitivity). Next tune in the carrier of the desired station. After tuning to the center of the carrier, lower the regeneration level until the audio level increases sharply. If you turn the control too far down, the detector will squeal. The use of R8 creates a narrow-band "widow" on the regeneration control between the point where the detector first begins to oscillate and the point where (narrow-band) audio begins to drop off rapidly. Increasing R8's resistance widens this region but lowers detector sensitivity. Because of their interaction, the regeneration control and the quench waveform control both need to be adjusted for narrow-band FM reception.

Fig 17.87 shows a simple antenna for the receiver. Mount the dipole vertically, since most transmissions in this frequency range are vertically polarized.

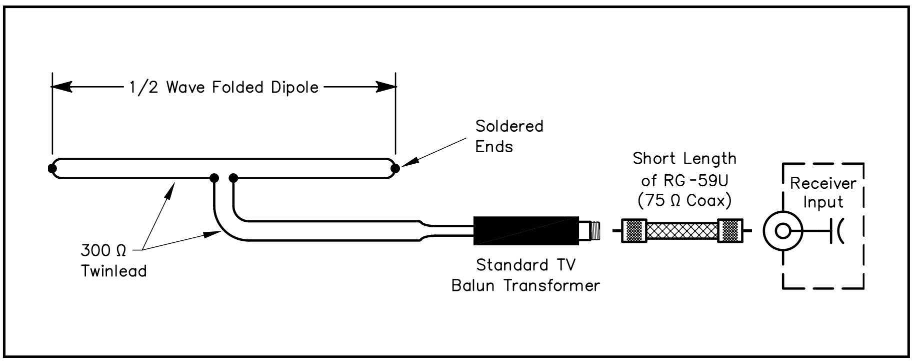


Fig 17.87—This simple dipole, made from 300-Ω TV line, will work fine. The TV line is attached to a coax feed line by a standard TV balun transformer. Make the dipole 39 inches long for the 2-meter band and a bit longer, 44.5 inches, for the aircraft band.

Tuning Around

Yes, the construction technique is ugly. Yes, you will have to adjust R8 and R9 constantly for best reception. Yes, you will have to practice finding stations and calibrating the dial. But, you can build it in one or two evenings, hear hundreds of stations, AM, WBFM or NBFM, that transmit in this frequency range, and get surprisingly good performance from two FETs and one integrated circuit.

You can modify this receiver to cover the 6-m band by changing a few components. Connect both sections of C3 together, and increase L1 and L3 to 30 mH. Use 8 turns of #14 wire for L2. C2 becomes 2 pF and C4 5 pF. Compress or expand the turns of L2 to cover the band.

A 30/40 W SSB/CW 20-M TRANSCEIVER

This project is a unique *Handbook* offering. Unlike many projects, this one was not created to be duplicated by the home constructor. This project is for the confident and experienced builder who is looking for a challenge. The transceiver was created in the ARRL Lab by Zack Lau, KH6CP/1, to meet an unusual set of design goals.

Some old-timers may remember Project Goodwill. The project purchased and distributed QRP 20-m CW transceiver kits to prospective hams in many foreign countries with small ham populations. Those transceivers ran out a few years ago, and ARRL began development of a new transceiver.

The goal was to supply an inexpensive, reliable, ready-built transceiver that provided moderate (not QRP) RF output in both the CW and SSB modes. **Fig 17.88** shows the transceiver block diagram. This transceiver was intended for inexpensive commercial manufacture. One of the important design goals was to eliminate as much tuning as possible, much like the microwave no-tune transverters. Of course, this conflicted with the next goal, which was to make the design as cheap as possible without sacrificing too much performance. Another goal was to accommodate a wide-tolerance PC-board process (single-sided with thick traces—50-mils minimum). Finally, the board was to use controls and connectors that are board-mounted.

In an attempt to simultaneously reduce the amount of tuning and keep costs down, a decision was made to use bilateral filters, mixers, and amplifiers where possible. It could be argued that the SBL-1s are too costly for an inexpensive transceiver project, but they offer several advantages. Being inherently bilateral, one needs only two mixers instead of four, reducing the number of parts needed. See the schematic in **Fig 17.89**. Also, they have pretty good carrier suppression, which eliminates a trimmer that would require adjustment, a very important consideration for this project. Finally, they are low impedance devices, which makes broadband matching much easier. As a bonus, they offer good dynamic range performance, sometimes outperforming NE602-based designs by 15 to 20 dB.

The no-tune requirement caused real problems in the power amplifier. The push-pull amplifier uses a pair of MRF477s that could probably be replaced by a single narrow-band MRF477 amplifier. A narrow-band amplifier would require adjustment, but one would get a significant savings. Additionally, a narrow-band amplifier would give more gain, allowing the driver stage to be run with a bit more feedback, increasing stability. The final amplifier shown is only conditionally stable, it will oscillate with severe mismatches. Fortunately, these transistors are pretty rugged, though you may wish to add some sort of SWR foldback circuitry for added insurance. To keep the costs down, such circuitry wasn't developed. A more complex bias circuit that independently biased the transistors might be better, but the simple circuit shown seems to work adequately with matched pairs of RF transistors.

The biasing circuit shows an interesting change in technology—it's probably more cost effective to use an integrated circuit and a variable resistor than to use the old alternative—a hand-selected 2-W resistor. Of course, if you just happen to have a junk box full of 2-W resistors, you're in good shape.

An 8-MHz IF was chosen, which allows the use of cheap microprocessor crystals and a 6-MHz VFO. Matching transformers are used because the mixer and amplifiers present close to 50-Ω terminations. Two separate four-pole filters are used to get a total of eight poles of filtering.

The cost cutting is most apparent in the simple AGC circuit. Initially, a hang-AGC circuit was used with good results, but it was removed to trim the cost as much as possible. An inferior audio-derived AGC circuit was used instead. One advantage of the audio-derived circuit is that it's less susceptible to BFO leakage, something that can easily be a problem when putting almost all the circuitry for an SSB transceiver on a single, single-sided PC board.

The audio amplifier is an old standby LM380 with an appropriate RC stabilization network on the output. While this IC does not have as much gain, or output, as newer chips, like the Signetics TDA 1015, it does seem easier to use. That's an important consideration when filling a big board with circuits that

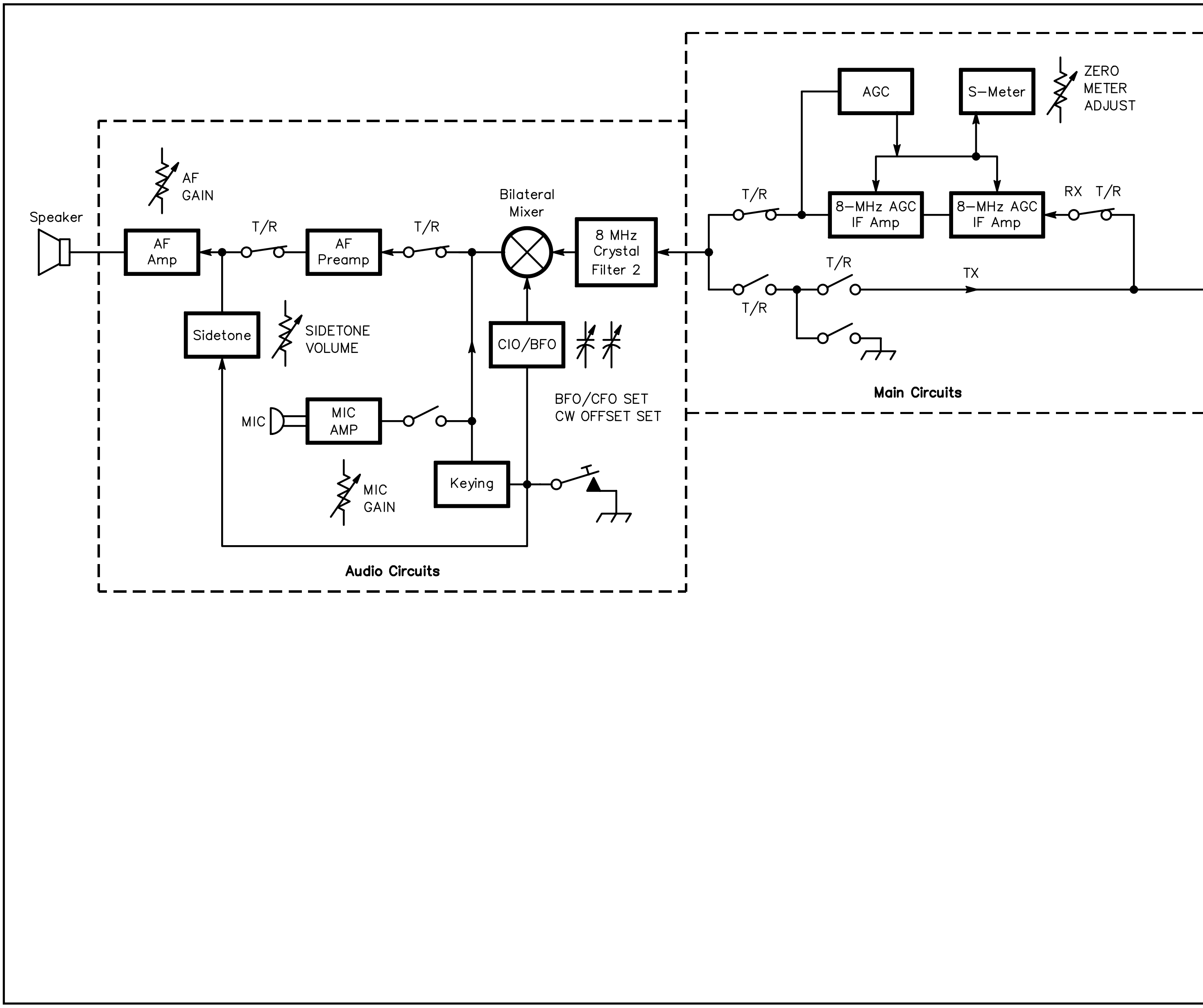


Fig 17.88—Block diagram of the transceiver.

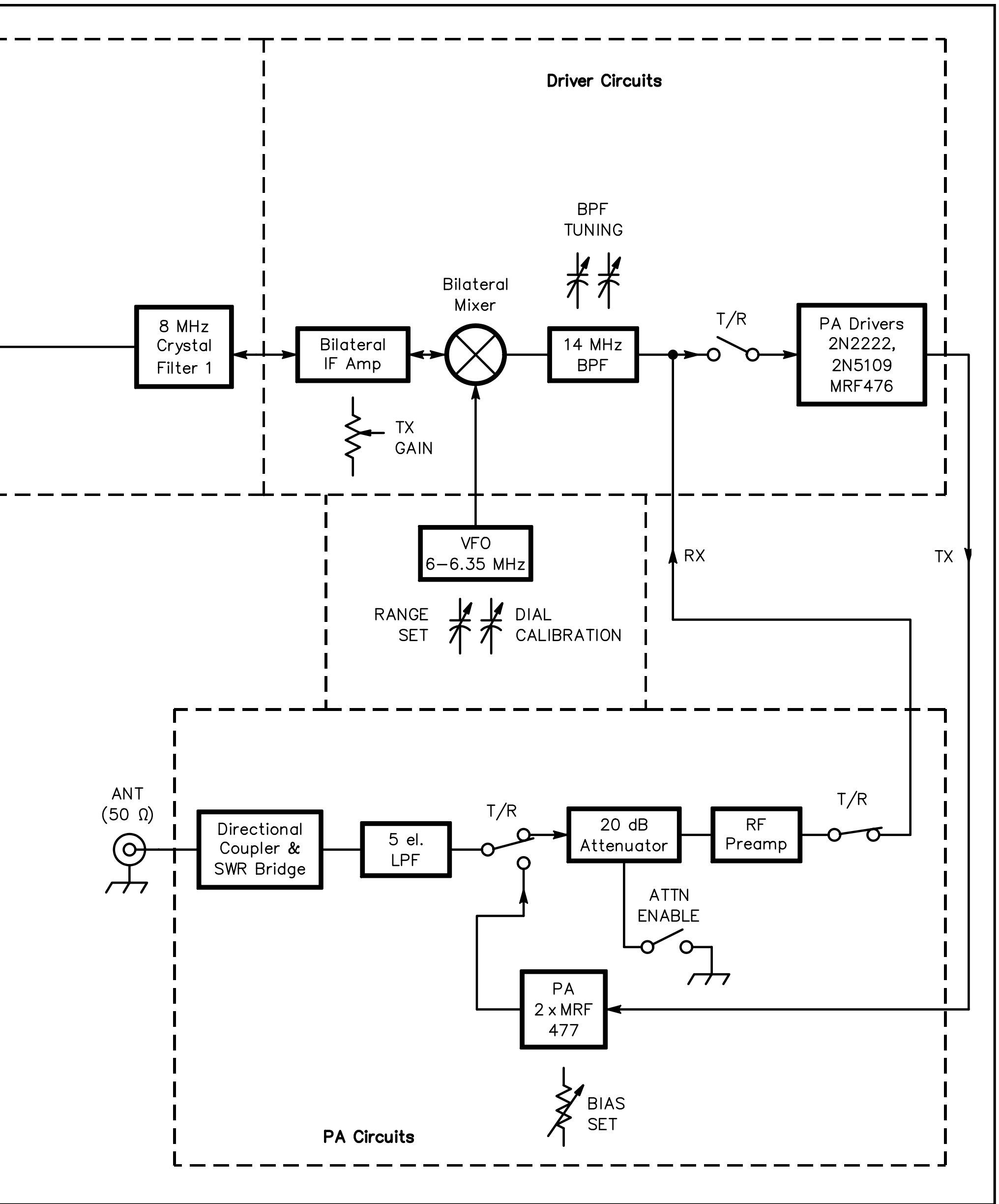
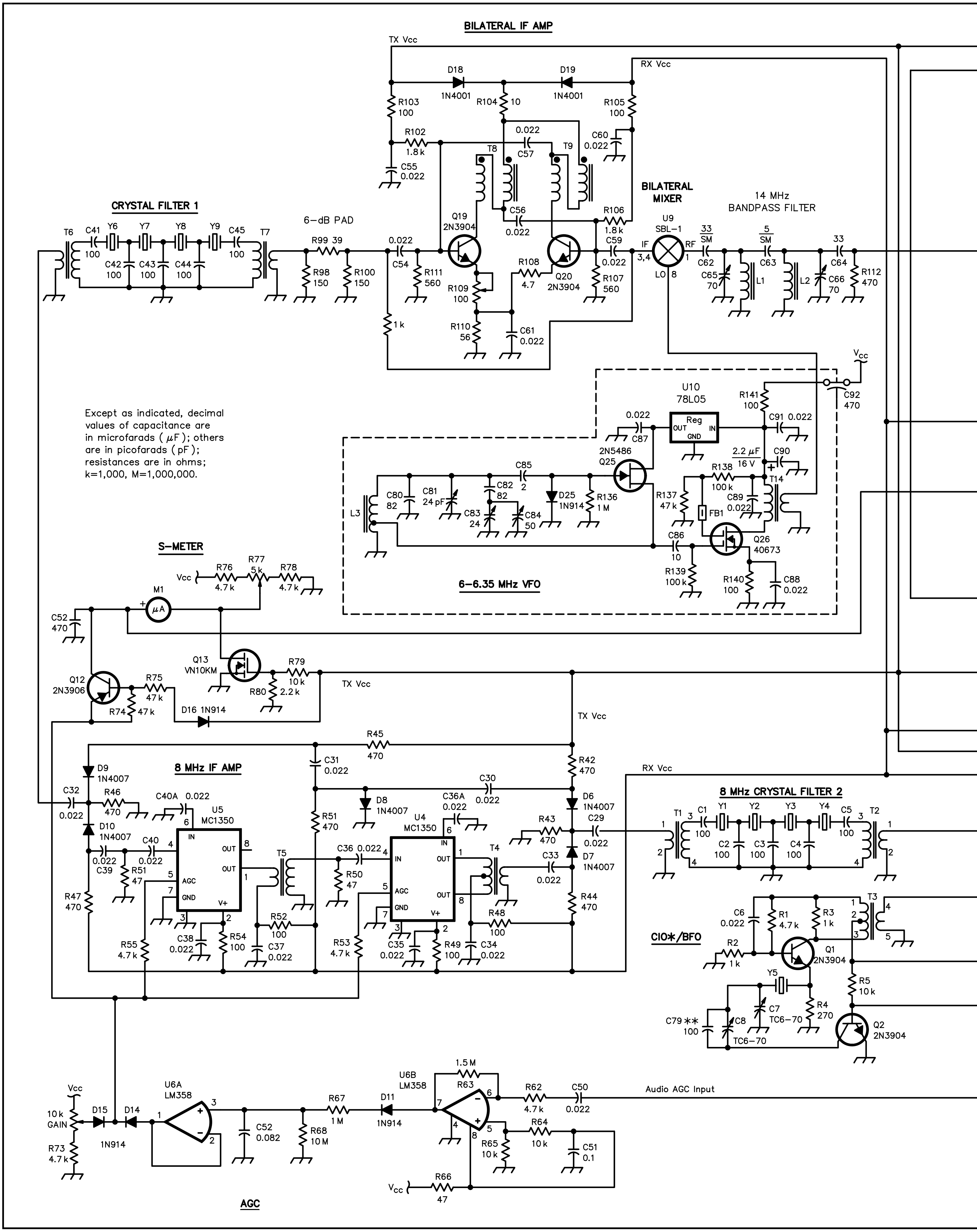
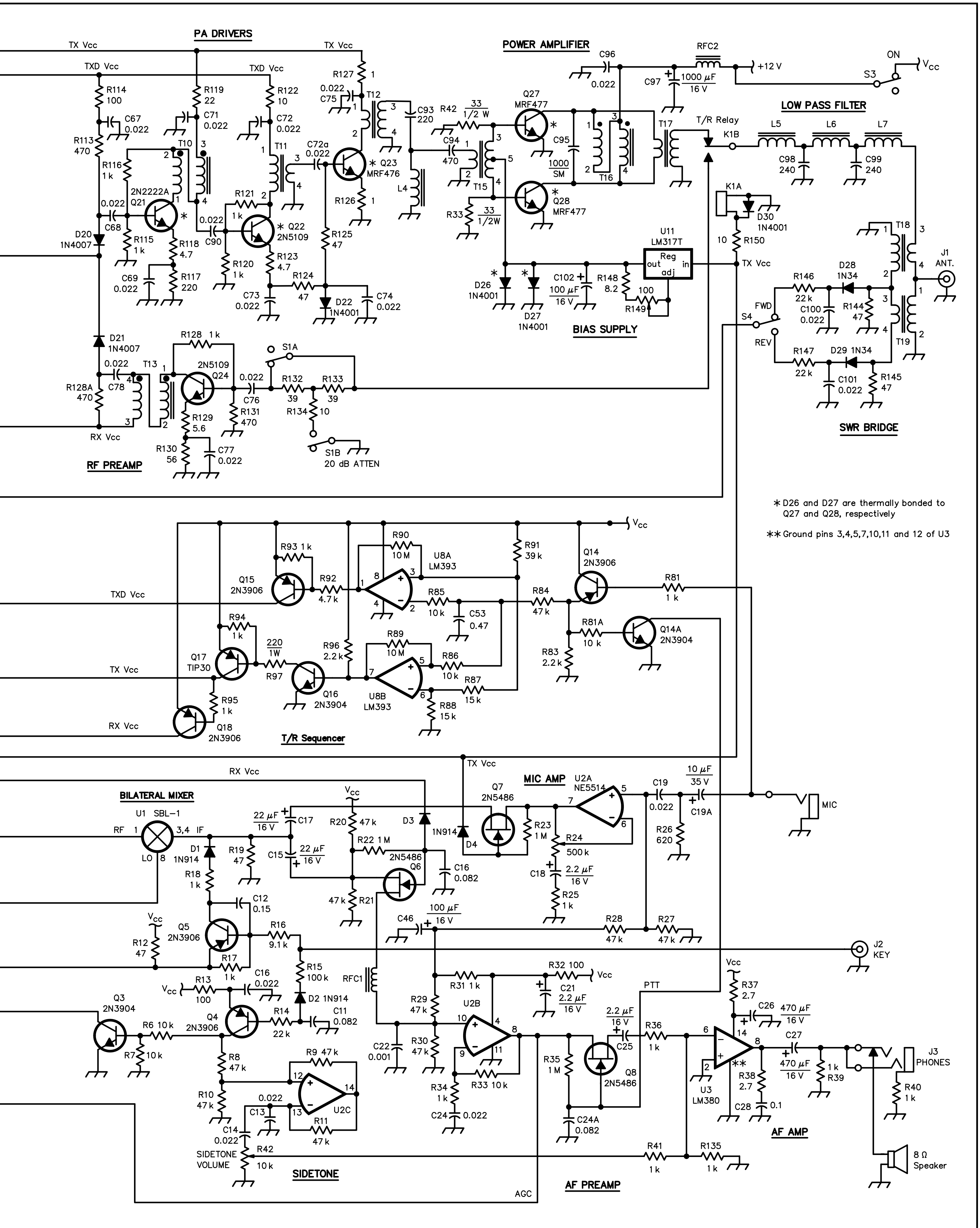

Driver Circuits
BPF
TUNING
Bilateral
Mixer
8 MHz
Crystal
Filter 1
Bilateral
IF Amp
14 MHz
BPF
T/R
PA Drivers
2N2222,
2N5109
MRF476
TX
GAIN
VFO
6–6.35 MHz
RX
TX
RANGE
SET
DIAL
CALIBRATION
ANT
(50 Ω)
Directional
Coupler &
SWR Bridge
5 el.
LPF
T/R
20 dB
Attenuator
RF
Preamp
T/R
ATTN
ENABLE
PA
2 x MRF
477
BIAS
SET
PA Circuits

Fig 17.89—Schematic diagram of the 20 m SSB/CW transceiver.

C1-5, C41-45—100-pF silver mica.
C7, C8, C65, C66—70-pF maximum ceramic or film trimmer capacitors.
C83—24-pF air trimmer (Johnson 189-509 used).
C84—50-pF tuning capacitor.
D6, D7, D8, D9, D10, D20, D21—1N4007 rectifier diodes used as PIN diodes.
D26, 27—1N4001 rectifier diodes thermally coupled to the nearest MRF477.
D28, D29—1N34 germanium diodes.
K1—Omron G2U-112-US SPDT relay or equiv.
L1, L2—17 t #22 enameled wire on a T-50-6 core.
L3—35 t #28 enameled wire on a T-50-6 core. Tap 8 turns up from ground end.
L4—13 t #28 enameled wire on a T-37-6 core.
L5, L7—12 t #22 enameled wire on a T-44-2 core.
L6—15 t #22 enameled wire on T-68-6 toroid core.
Q1, Q2, Q3, Q14, Q16, Q19, Q20—2N3904 NPN BJT.
Q4, Q5, Q12, Q15, Q18—2N3906 PNP BJT.
Q6, Q7, Q8, Q25—2N5486 N-channel JFET.
Q13—VN10KM VMOS FET, or equiv.
Q17—TIP 30 PNP BJT.
Q21—2N2222A metal cased NPN BJT.
Q22, Q24—2N5109 NPN BJT with heat sinks.
Q23—MRF476 NPN BJT with heat sinks.
Q26—40673 dual-gate MOSFET, or equiv.
Q27, Q28—MRF477 NPN BJT with appropriate heat sinks.
RFC1—1 mH toroidal RF choke. FT-37-72 toroid filled with one layer of #28 enameled wire.
RFC2—8 t #18 enameled wire on an FT-50-43 core.
T1, T2, T6, T7—12 t #28 enameled wire primary feeds crystal filter. 5 t #28 enameled wire secondary on an FT-37-43 core.
T3—20 t #28 enameled wire on an FT-37-43 core. Tap 7 t from collector end. Secondary is 4 t of #28 enameled wire over the primary.
T4, T5—18 t #28 enameled wire center-tapped primary on an FT-37-43 toroid. Secondary, 2 t #28 enameled wire. (One way of reducing the IF gain is to use only one half of the primary.)
T8, T9, T10, T13—5 t bifilar wound #28 enameled on an FT-37-43 toroid.
T11—5 t #24 enameled wire primary on an FT-37-43. Secondary, 3 t #22 enameled wire (to MRF 476).
T12—3 t #23 enameled wire primary on an FT-37-43 toroid. 5 t #22 enameled wire secondary (output).
T14—18 t #28 enameled wire primary. 5 t #28 secondary on an FT-37-43 core.
T15—Input is 3 t #24 enameled wire on a BLN-202-73 balun core. Output to transistors is 1 t center-tapped #24 enameled wire.
T16—8 t #20 bifilar wound on an FT-50-43 ferrite toroid.
T17—Output transformer. $^{3}/_{16}$-inch brass tubing and unetched circuit-board stock (double-sided) primary through 4 FB-63-43 ferrite beads. 4 t #20 enameled wire secondary through tubing.
T18, T19—31 t #24 enameled wire on a T-50-3 toroid. 1 t #24 enameled wire secondary.
U1, U9—SBL-1 double-balanced mixer.
U2—NE5514 quad op amp.
U3—LM380 audio amplifier.
U4, U5—MC1350 IF amplifier.
U6—LM358 dual op amp.
U10—78L05 regulator.
U8—LM393 dual comparator.
U11—LM317T adjustable regulator.
Y1, Y2, Y3, Y4, Y6, Y7, Y8, Y9—Matched 8-MHz microprocessor crystals (HC-18/U case). 250-Hz matching should be adequate.
Y5—8-MHz microprocessor crystal.


BILATERAL IF AMP
CRYSTAL FILTER 1
6-dB PAD
BILATERAL MIXER
14 MHz BANDPASS FILTER
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.
S-METER
6–6.35 MHz VFO
8 MHz IF AMP
8 MHz CRYSTAL FILTER 2
CIO*/BFO
AGC
Audio AGC Input


PA DRIVERS
POWER AMPLIFIER
LOW PASS FILTER
T/R Relay
BIAS SUPPLY
RF PREAMP
SWR BRIDGE
20 dB ATTEN
∗D26 and D27 are thermally bonded to Q27 and Q28, respectively
∗∗Ground pins 3,4,5,7,10,11 and 12 of U3
T/R Sequencer
BILATERAL MIXER
MIC AMP
SIDETONE
SIDETONE VOLUME
AF PREAMP
AF AMP
ANT.
KEY
MIC
PHONES
8 Ω Speaker
TX Vcc
TXD Vcc
RX Vcc
PTT
AGC
+12 V
ON

are bound to interact if you give them a chance. 2N5486 JFETs are used for audio switching. While it's certainly possible to reduce the parts count by using quad switching chips instead of transistors, ICs are more difficult to lay out properly in a complicated circuit. The sidetone is a triangle wave.

The VFO uses a JFET and a MOSFET buffer amplifier. Unfortunately MOSFETs are apparently being phased out, but you can still get them from a variety of surplus sources.

Zack built an “ugly” prototype on several unetched PC boards, and that circuit went through many changes as he worked with the manufacturer. What you see here is the circuit as it was when development stopped. It is presented in the *Handbook* because there has been a consistent demand for an SSB transceiver with more that a few watts of output. Consider this a work in progress. If you build it, drop the *Handbook* Editor a note telling your experience and ideas. With a little more development, this could be a useful and widely popular project.

Construction

If the circuits are all working properly, alignment should be pretty easy. While there are two interactive adjustments for the band-pass filter, the BFO/carrier insertion oscillator and the VFO alignment, they aren’t terribly difficult. Other adjustments are the S-meter zero, the sidetone volume, transmit gain, MRF477 bias setting, and mic-gain adjustment.

The bias setting for the finals is a trade-off, as one does get more gain and better IMD performance with a higher bias current, but the amount of heat generated is higher. A setting that results in a total current of 1 A during transmit seems to work well.

The mic gain should be adjusted for 30 W on voice peaks for good linearity, while 40 W is available on CW. One of the prototypes has a noise floor of –129 dBm.

1996 Updates

Lance, WS2B, tells us that Motorola recently discontinued the MRF 476 (Q23). NTE Cross Reference software gives the NTE236 as an alternative device.

Dan's Small Parts may have the 40673 and the VN10KM.

THE NORCAL SIERRA: AN 80-15 M CW TRANSCEIVER

Most home-built QRP transceivers cover a single band, for good reason: complexity of the circuit and physical layout can increase dramatically when two or more bands are covered. This holds for most approaches to multiband design, including the use of multipole switches, transverters and various forms of electronic switching.[1]

If the designer is willing to give up instant band switching, then plug-in band modules can be used. Band modules are especially appropriate for a transceiver that will be used for extended portable operation, for example: back-packing. The reduced circuit complexity improves reliability, and the extra time it takes to change bands usually isn't a problem. Also, the operator need take only the modules needed for a particular outing.

The Sierra transceiver shown in **Fig 17.90** uses this technique, providing coverage of all bands from 80 through 15 m with good performance and relative simplicity.[2] The name Sierra was inspired by the mountain range of the same name—a common hiking destination for West Coast QRPers. The transceiver was designed and built by Wayne Burdick, N6KR, and field tested by members of NorCal, the Northern California QRP Club.[3]

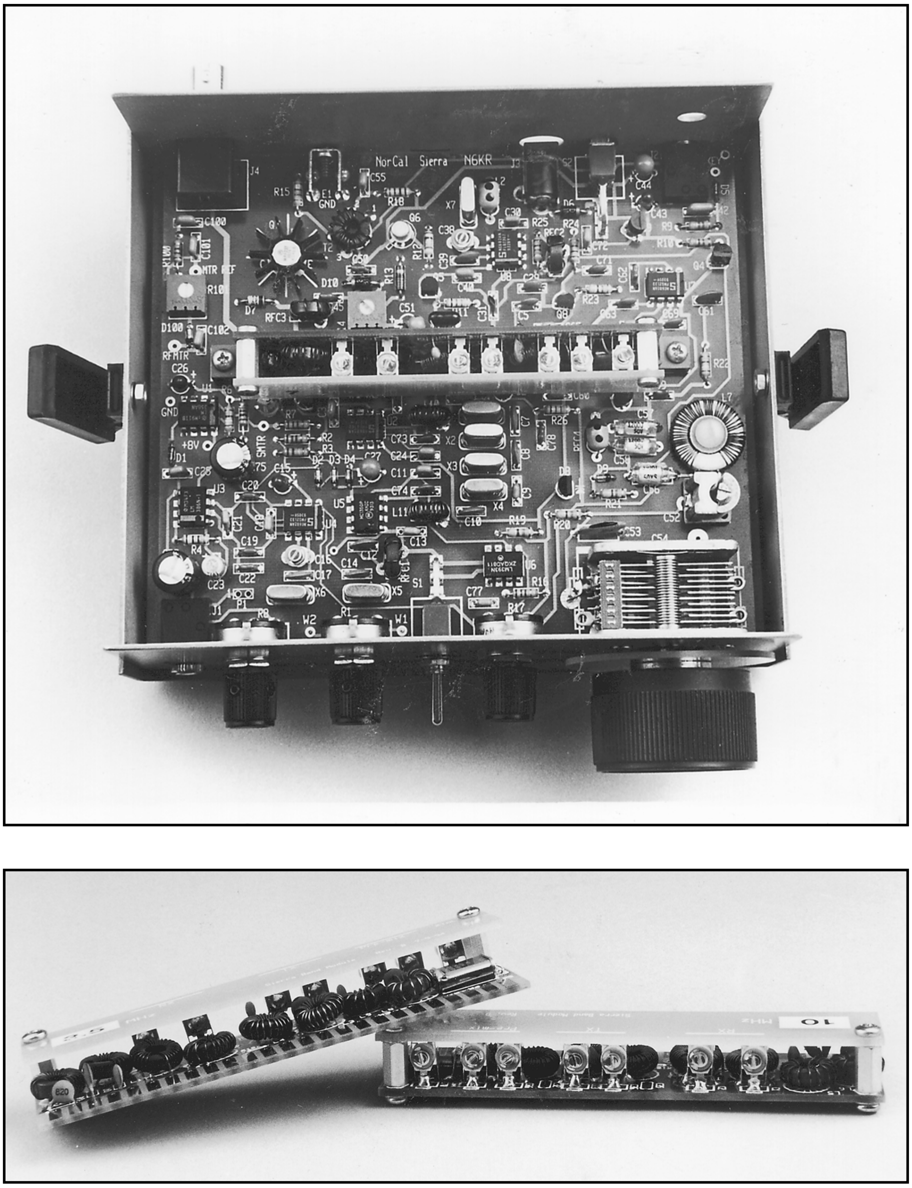


Fig 17.90—The Sierra transceiver. One band module is plugged into the center of the main PC board; the remaining boards are shown to the left of the rig. Quick-release latches on the top cover of the enclosure make it easy to change bands.

1 One of N6KR's previous designs, the Safari-4, is a good example of how complex a band-switched rig can get. See "The Safari-4...." Oct through Dec 1990 *QEX*.

2 Band modules for 160, 12 and 10 m have also been built. Construction details for these bands are provided in the Sierra information packet available from the ARRL. Go to the ARRL Web page: **http://www.arrl.org/notes/** to download a template package or write to the ARRL Technical Department Secretary and request the '96 *Handbook* Sierra template package.

3 For information about NorCal, write to Jim Cates, WA6GER, who is in the **References** Address List. Please include an SASE.

Features

One of the most important features of the Sierra for the portable QRP operator is its low current drain. Because it has no relays, switching diodes or other active band-switching circuitry, the Sierra draws only 30 mA on receive.[4] Another asset for field operation is the Sierra's low-frequency VFO and premixing scheme, which provides 150 kHz of coverage and good frequency stability on all bands.

The receiver is a single-conversion superhet with audio-derived AGC and RIT. It has excellent sensitivity and selectivity, and will comfortably drive a speaker. Transmit features include full break-in keying, shaped keying and power output averaging 2 W, with direct monitoring of the transmitted signal in lieu of sidetone. Optional circuitry allows monitoring of relative power output and received signal strength.

Physically, the Sierra is quite compact—the enclosure is 2.7×6.2×5.3 inches (HWD)—yet there is a large amount of unused space both inside and on the front and rear panels. This results from the use of PC board-mounted controls and connectors. The top cover is secured by quick-release plastic latches, which provide easy access to the inside of the enclosure. Band changes take only a few seconds.

Circuit Description

Fig 17.91 is a block diagram of the Sierra. The diagram shows specific signal frequencies for operation on 40 m. **Table 17.4** provides a summary of crystal oscillator and premix frequencies

4 Most multiband rigs draw from 150 to 500 mA on receive, necessitating the use of a larger battery. A discussion of battery life considerations can be found in "A Solar-Powered Field Day," May 1995 *QST*.

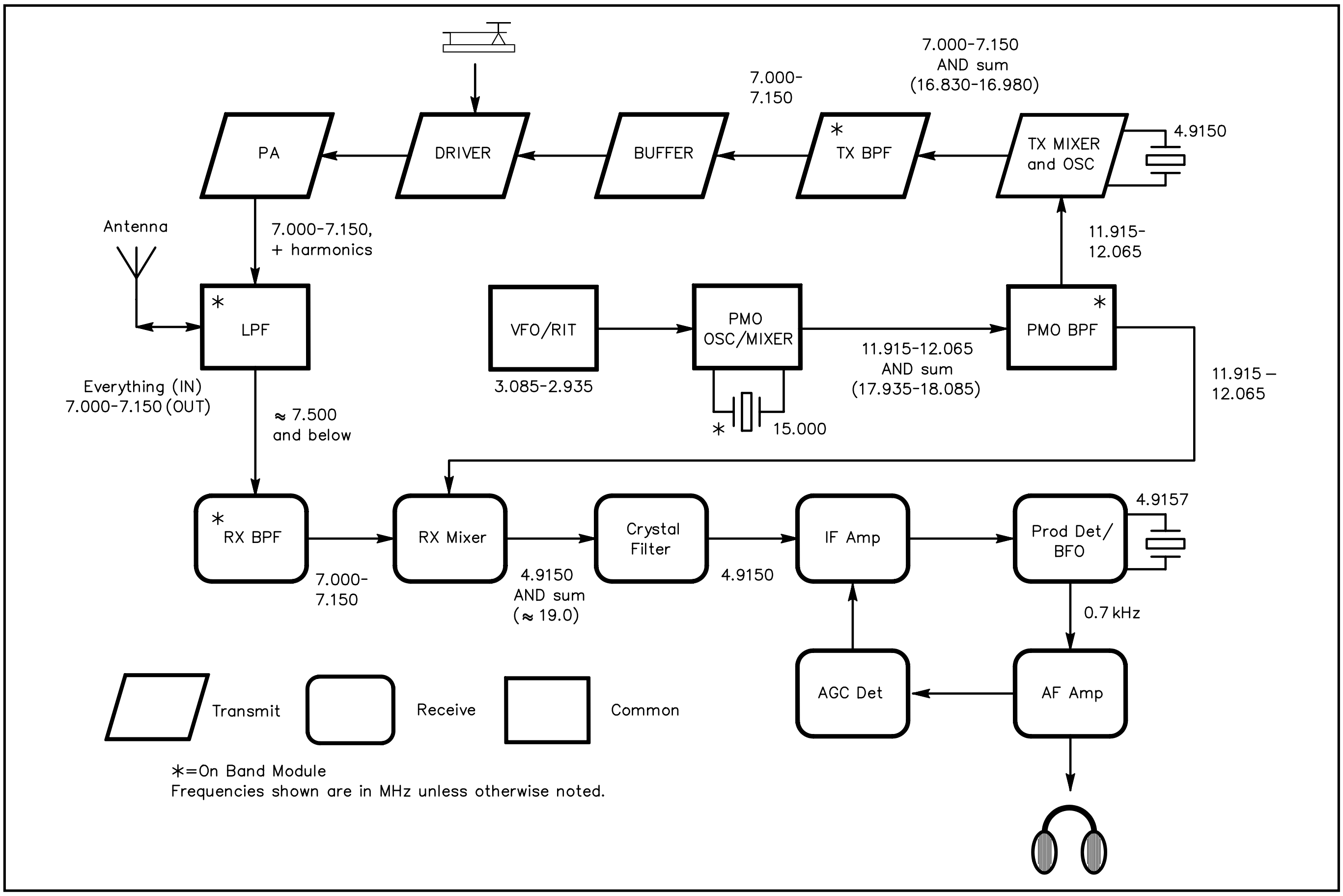


Fig 17.91—Block diagram of the Sierra transceiver. Three different-shaped symbols are used to show transmit, receive and common blocks. Those blocks with an asterisk (*) are part of the band module. Signal frequencies shown are for 40 m; see Table 17.4 for a list of crystal oscillator and premix frequencies for all bands.

Table 17.4
Crystal Oscillator and Premix (PMO) Frequencies in MHz

The premixer (U7) subtracts the VFO (2.935 to 3.085 MHz) from the crystal oscillator to obtain the PMO range shown. The receive mixer (U2) subtracts the RF input from the PMO signal, yielding 4.915 MHz. The transmit mixer (U8) subtracts 4.915 MHz from the PMO signal to produce an output in the RF range.

RF Range	*Crystal Oscillator*	*PMO Range*
3.500-3.650	11.500	8.415-8.565
7.000-7.150	15.000	11.915-12.065
10.000-10.150	18.000	14.915-15.065
14.000-14.150	22.000	18.915-19.065
18.000-18.150	26.000	22.915-23.065
21.000-21.150	29.000	25.915-26.065

for all bands. The schematic is shown in **Fig 17.92**. See **Table 17.5** for band-module component values.

On all bands, the VFO range is 2.935 MHz to 3.085 MHz. The VFO tunes "backwards": At the low end of each band, the VFO frequency is 3.085 MHz.

U7 is the premixer and crystal oscillator, while Q8 buffers the premix signal prior to injection into the receive mixer (U2) and transmit mixer (U8).

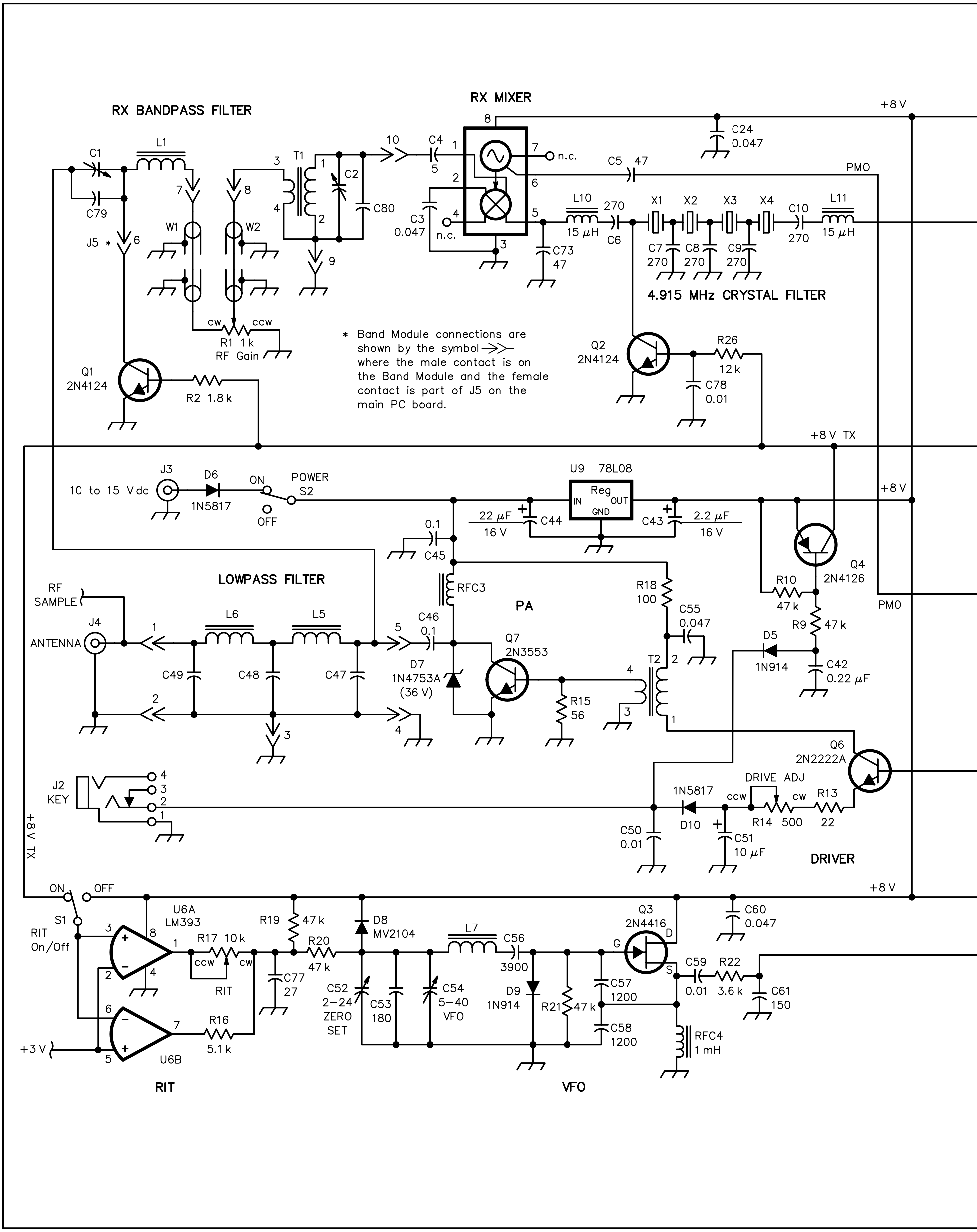
A low-pass filter, three band-pass filters and a premix crystal make up each band module. To make the schematic easier to follow, this circuitry is integrated into Fig 17.92, rather than drawn separately. J5 is the band module connector (see the note on the schematic).

The receive mixer is an NE602, which draws only 2.5 mA and requires only about 0.6 V (P-P) of oscillator injection at pin 6. An L network is used to match the receive mixer to the first crystal filter (X1-X4). This filter has a bandwidth of less than 400 Hz. The single-crystal second filter (X5) removes some of the noise generated by the IF amplifier (U7), a technique W7ZOI described.[5] This second filter also introduces enough loss to prevent the IF amplifier from overdriving the product detector (U4).

[5] *Solid-State Design*, p 87.

Fig 17.92—Schematic of the Sierra transceiver. Parts that change for each band are shown in Table 17.5.

C1, C2, C33, C36, C64, C66, C70—9-50 pF right-angle-mount ceramic trimmer (same for all band modules, Mouser 24AA084)
C16, C38—Ceramic trimmer, 8-50 pF (Mouser 24AA024).
C52—Air variable, 2-24 pF (Mouser 530-189-0509-5).
C53—Disc, 180 pF, 5%, NP0.
C54—5-40 pF air variable with 8:1 vernier drive.
C56—Polystyrene, 3900 pF, 5%.
C57, C58—Polystyrene, 1200 pF, 5%.
D6, D10—1N5817, 1N5819 or similar .
D7—36 V, 1 W Zener diode (Mouser 333-1N4753A).
D8—MV2104 varactor diode, or equivalent.
J1, J2—PC-mount 3.5-mm stereo jack with switch (Mouser 161-3500).
J3—2.1-mm dc power jack (Mouser 16PJ031).
J4—PC-mount BNC jack (Mouser 177-3138).
J5—50 PIN, dual-row edgeboard connector with 0.156-inch spacing (Digi-Key S5253-ND).
L10, L11—18 µH; 18 t #28 enameled wire on an FT-37-61 toroid.
L2—Miniature RFC, 15 µH (Mouser, 43LS185).
L7—19 µH; 58 t #28 enameled wire on a T-68-7 toroid.
Q5—U310, J310, 2N4416 or other high-transconductance device.
R1, R8—PC-mount 1-kΩ pot (Mouser 31CW301).
R14, R101—500 Ω trimmer (Mouser 323-4295P-500).
R17—PC-mount 10-kΩ pot (Mouser 31CW401).
RFC1—3.5 µH; 8 t #26 enameled wire on an FT-37-61 toroid.
RFC2—7 µH; 4 t #26 enameled wire on an FT-37-43 toroid.
RFC3—34 µH; 9 t #26 enameled wire on an FT-37-43 toroid.
RFC4—Miniature RFC, 1 mH (Mouser 43LS103).
S1, S2—SPDT, PC mount, right angle toggle switch with threaded bushing. C&K 7101SDAV2QE is used in the kit; Digi-Key CKN1059-ND will work but does not have a threaded bushing.
T2—Primary: 12 t #26 enameled wire; secondary: 3 t on an FT-37-43 toroid.
U1—LM358N dual op-amp IC.
U2, U4, U7, U8—NE602AN mixer-oscillator IC.
U3—LM386N-1 audio amplifier IC.
U5—MC1350P IF amplifier IC.
U6—LM393N dual comparator IC.
U9—8 V regulator, TO-92 package (Digi-Key AN78L08-ND).
W1, W2—RG-174 coaxial jumper, about 3 inches long (see text).
X1-X7—4.915 MHz, HC-49 (Digi-Key CTX050). X1 through X5 should be matched (their series-resonant frequencies within 50 Hz).


RX BANDPASS FILTER
RX MIXER
+8 V
C24
0.047
C5 47
PMO
C1
L1
C79
J5 *
W1
W2
T1
C2
C80
10
C4
5
C3
0.047
n.c.
L10
15 μH
270
C6
X1
X2
X3
X4
C10
270
L11
15 μH
C7
270
C8
270
C9
270
C73
47
4.915 MHz CRYSTAL FILTER
R1 1k
RF Gain
cw
ccw
* Band Module connections are shown by the symbol where the male contact is on the Band Module and the female contact is part of J5 on the main PC board.
Q1
2N4124
R2 1.8 k
Q2
2N4124
R26
12 k
C78
0.01
+8 V TX
J3
D6
1N5817
ON
OFF
POWER
S2
10 to 15 Vdc
U9 78L08
Reg
IN
OUT
GND
22 μF
16 V
C44
C43
2.2 μF
16 V
0.1
C45
Q4
2N4126
R10
47 k
R9
47 k
PMO
LOWPASS FILTER
RF SAMPLE
J4
ANTENNA
L6
L5
C49
C48
C47
RFC3
PA
C46
0.1
D7
1N4753A
(36 V)
Q7
2N3553
R15
56
R18
100
C55
0.047
T2
D5
1N914
C42
0.22 μF
Q6
2N2222A
J2
KEY
+8 V TX
DRIVE ADJ
1N5817
D10
C50
0.01
C51
10 μF
R14 500
R13
22
DRIVER
ON
OFF
S1
RIT On/Off
U6A
LM393
R17 10 k
ccw
cw
RIT
R19
47 k
R20
47 k
C77
27
R16
5.1 k
U6B
+3 V
D8
MV2104
C52
2–24
ZERO SET
C53
180
C54
5–40
VFO
L7
C56
3900
D9
1N914
R21
47 k
C57
1200
C58
1200
Q3
2N4416
G
D
S
C59
0.01
R22
3.6 k
C61
150
C60
0.047
RFC4
1 mH
RIT
VFO

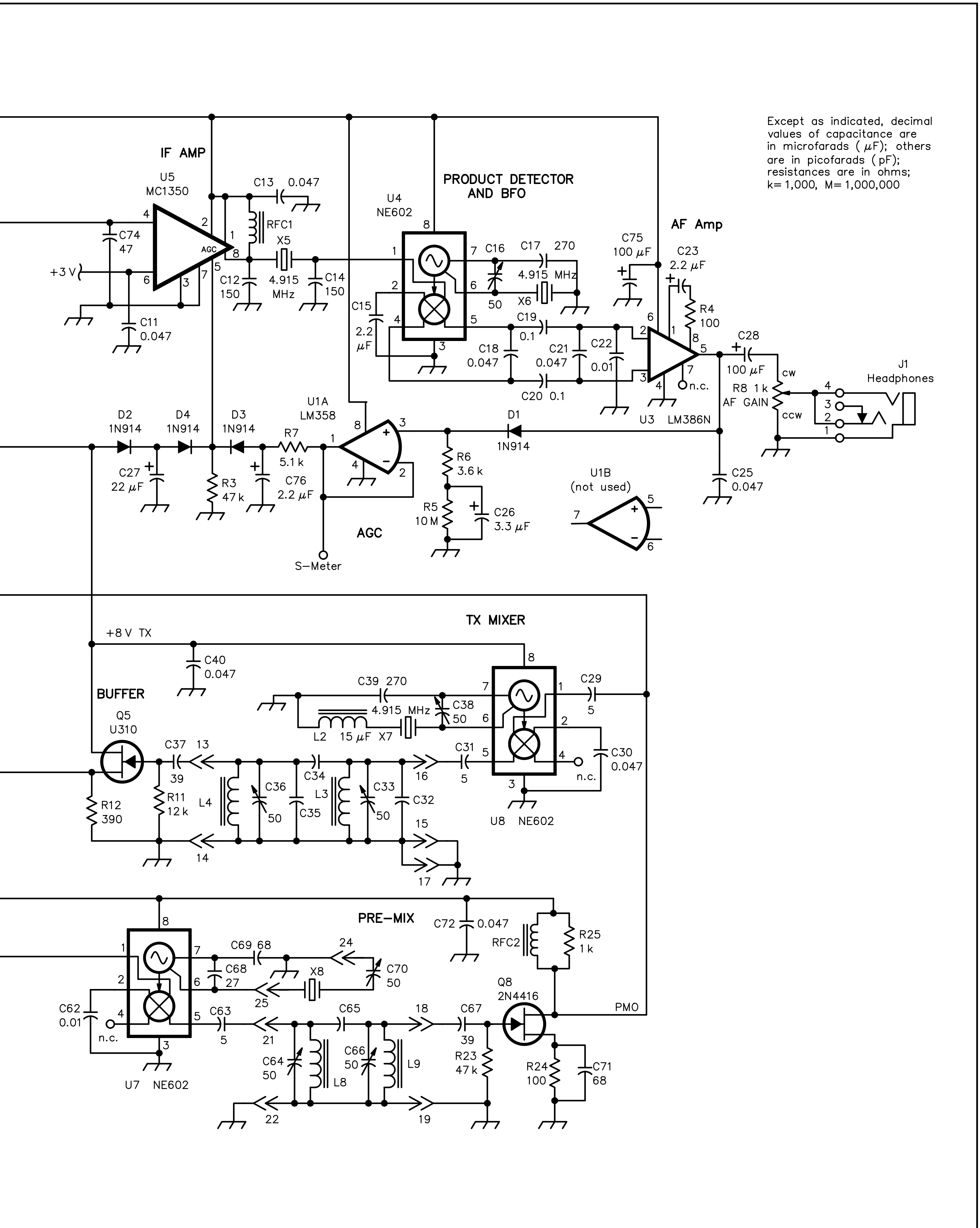

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k= 1,000, M= 1,000,000
IF AMP
U5
MC1350
AGC
C74
47
+3 V
C11
0.047
C13 0.047
RFC1
X5
4.915
MHz
C12
150
C14
150
PRODUCT DETECTOR
AND BFO
U4
NE602
C16
50
C17 270
4.915 MHz
X6
C15
2.2
µF
C18
0.047
C19
0.1
C21
0.047
C22
0.01
C20 0.1
AF Amp
C75
100 µF
C23
2.2 µF
R4
100
C28
100 µF
U3 LM386N
n.c.
R8 1 k
AF GAIN
cw
ccw
J1
Headphones
D2
1N914
D4
1N914
D3
1N914
R7
5.1 k
U1A
LM358
C27
22 µF
R3
47 k
C76
2.2 µF
R6
3.6 k
R5
10 M
C26
3.3 µF
D1
1N914
C25
0.047
AGC
S–Meter
U1B
(not used)
TX MIXER
+8 V TX
C40
0.047
BUFFER
Q5
U310
C37
39
R11
12 k
R12
390
L4
C36
50
C34
C35
L3
C33
50
C32
C39 270
4.915 MHz
C38
50
L2 15 µF X7
C31
5
C29
5
C30
0.047
n.c.
U8 NE602
PRE–MIX
C72 0.047
RFC2
R25
1 k
C69 68
C68
27
X8
C70
50
C62
0.01
n.c.
C63
5
C65
C67
39
Q8
2N4416
PMO
C64
50
L8
C66
50
L9
R23
47 k
R24
100
C71
68
U7 NE602

Table 17.5

Band Module Components

All crystals are fundamental, 15-pF load capacitance, 0.005% frequency tolerance, in HC-49 holders. Fixed capacitors over 5 pF are 5% tolerance. All coils are wound with enameled wire.

				Band		
Part	*80 m*	*40 m*	*30 m*	*20 m*	*17 m*	*15 m*
C32, C35	33 pF, 5%	47 pF, 5%	not used	not used	not used	not used
C34	5 pF, 5%	5 pF, 5%	2 pF, 5%	2 pF, 5%	2 pF, 5%	2 pF, 5%
C47, C49	820 pF, 5%	330 pF, 5%	330 pF, 5%	220 pF, 5%	150 pF, 5%	150 pF, 5%
C48	1800 pF, 5%	820 pF, 5%	560 pF, 5%	470 pF, 5%	330 pF, 5%	330 pF, 5%
C65	5 pF, 5%	5 pF, 5%	2 pF, 5%	1 pF, 5%	1 pF, 5%	1 pF, 5%
L1	50 µH, 30 t #28 on FT-37-61	14 µH, 16 t #26 on FT-37-61	5.2 µH, 36 t #28 on T-37-2	2.9 µH, 27 t #28 on T-37-2	1.7 µH, 24 t #28 on T-37-6	1.9 µH, 25 t #28 on T-37-6
L3, L4	32 µH, 24 t #26 on FT-37-61	5.2 µH, 36 t #28 on T-37-2	4.4 µH, 33 t #28 on T-37-2	2.9 µH, 27 t #28 on T-37-2	1.7 µH, 24 t #28 on T-37-6	1.9 µH, 25 t #28 on T-37-6
L5, L6	2.1 µH, 23 t #26 on T-37-2	1.3 µH, 18 t #26 on T-37-2	1.0 µH, 16 t #26 on T-37-2	0.58 µH, 12 t #26 on T-37-2	0.43 µH, 12 t #26 on T-37-6	0.36 µH, 11 t #26 on T-37-6
L8, L9	8.0 µH, 12 t #26 on FT-37-61	2.5 µH, 25 t #28 on T-37-2	1.6 µH, 20 t #28 on T-37-2	1.3 µH, 18 t #26 on T-37-2	0.97 µH, 18 t #26 on T-37-6	0.87 µH, 17 t #28 on T-37-6
T1 (Sec same as L1)	Pri: 2 t #26 on FT-37-61	Pri: 1 t #26 on FT-37-61	Pri: 3 t #26 on T-37-2	Pri: 2 t #26 on T-37-2	Pri: 2 t #26 on T-37-6	Pri: 2 t #26 on T-37-6
X8	11.500 MHz (ICM 434162)	15.000 MHz (ICM 434162)	18.000 MHz (ICM 434162)	22.000 MHz (ICM 435162)	26.000 MHz (ICM 436162)	29.000 MHz (ICM 436162)

The output of the AF amplifier (U3) is dc-coupled to the AGC detector. U3's output floats at $V_{cc}/2$, about 4 V, which happens to be the appropriate no-signal AGC voltage for the IF amplifier when it is operated at 8 V. C26, R5, R6, C76 and R7 provide AGC loop filtering. Like all audio-derived AGC schemes, this circuit suffers from pops or clicks at times.

Transmit signal monitoring is achieved by means of a separate 4.915 MHz oscillator for the transmitter; the difference between this oscillator and the BFO determines the AF pitch. Keying is exponentially shaped, with the rise time set by the turn-on delay of transmit mixer U8 and the fall time determined by C51, in the emitter of driver Q6.

CONSTRUCTION

The Sierra's physical layout and packaging make it relatively easy to build and align, although this isn't a project for the first-time builder. The boards and custom enclosure described here are included as part of an available kit.[6] Alternative construction methods are discussed below.

With the exception of the components on the band module, all of the circuitry for the Sierra is mounted on a single 5×6 inch PC board. This board contains not only the components, but all of the controls and connectors as well. The board is double-sided with plated-through holes, which permits flexible arrangement of the circuitry while eliminating nearly all hand-wiring. The only two jumpers on the board, W1 and W2, are short coaxial cables between the RF GAIN control and the receiver input filters.

A dual-row edge connector (J5) provides the interface between

[6] Full and partial kits are available. The full kit comes with all components, controls, connectors, and a detailed assembly manual. Complete band modules kits are available for 80, 40, 30, 20, 17 and 15 m. For information, write to Wilderness Radio (see Address List in the **References** chapter).

the main board and the band module. The 50 pins of J5 are used in pairs, so there are actually only 25 circuits (over half of which are ground connections).

The band module boards are 1.25 × 4 inches (HW). They, too, are double-sided, maximizing the amount of ground plane. Because the band modules might be inserted and removed hundreds of times over the life of the rig, the etched fingers that mate with J5 are gold-plated. Each etched finger on the front is connected to the corresponding finger on the back by a plated through hole, which greatly improves reliability over that of a single finger contact.

Each band module requires eight toroids: two for the low-pass filter, and two each for the receive, transmit and premix band-pass filters. The builder can secure the toroids to the band module with silicone adhesive or Q-dope. Right-angle-mount trimmer capacitors allow alignment from above the module. Each band module has a top cover made of PC board material. The cover protects the components during insertion, removal and storage.

The VFO capacitor is a 5-40 pF unit with a built-in 8:1 vernier drive. The operating frequency is read from a custom dial fabricated from 0.060-inch Lexan. The dial mounts on a hub that comes with the capacitor.

The Sierra's custom 0.060-inch aluminum enclosure offers several benefits in both construction and operation. Its top and bottom covers are identical U-shaped pieces. The bottom is secured to the main board by two 0.375-inch standoffs, while the top is secured to the bottom by two long-life, quick-release plastic latches. As a result, the builder can easily remove both covers to make "live" adjustments or signal measurements without removing any controls, connectors or wires. The front and rear panels attach directly to the controls and connectors on the main board. This keeps the panels rigid and properly oriented.

As can be seen in the photograph, the interior of the rig is uncluttered. NorCal QRP Club members have taken advantage of this, building in keyers, frequency counters and other accessories—and even storing up to four band modules in the top cover. One popular addition is an S/RF meter, the circuit shown in **Fig 17.93**.

The construction techniques described above represent only one way to build the Sierra; other physical layouts may better suit your needs. For example: If no built-ins are needed, the rig could be built in a smaller enclosure. You could replace the VFO capacitor with a small 10-turn pot and a varactor diode. If necessary, eliminate RIT and metering.

If a different physical layout is required, determine the orientation and mounts for the band module connector first, then arrange the various circuit blocks around it. Use short leads and good ground-plane techniques to avoid instability, especially on the band modules. Point-to-point or "dead-bug" construction are possible, but in some cases shields and additional decoupling may be required. Use a reliable connector if band modules will be repeatedly inserted and removed.

Alignment

The minimum recommended equipment for aligning the rig is a DMM with homemade RF

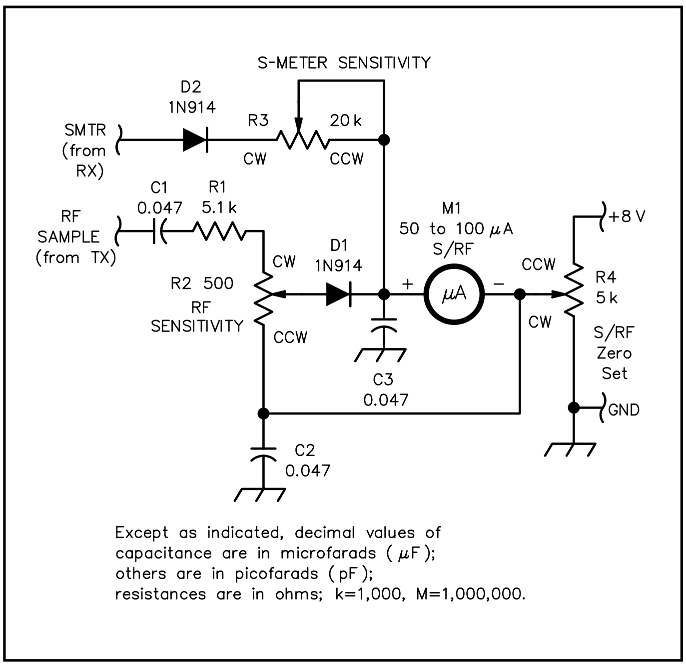


Fig 17.93—Metering circuitry for the Sierra transceiver.

probe and a ham-band transceiver. Better still is a general-coverage receiver or frequency counter.[7] Start with a 40- or 20-m module; these are usually the easiest to align.

First, set the VFO to the desired band edge by adjusting C52. If exactly 150 kHz of range is desired, squeeze or spread the windings of L7 and readjust C52 iteratively until this range is obtained. RIT operation can also be checked at this time. Reduce the value of R19 if more RIT range is desired.

Prepare each band module for alignment by setting all of its trim caps to midrange. (The final settings will be close to midpoint in most cases.)

Receiver alignment is straightforward. Set BFO trimmer C16 to midrange, RF GAIN (R1) to maximum and AF GAIN (R8) so that noise can be heard on the phones or speaker. On the band module, peak the premix trimmers (C64 and C66) for maximum signal level measured at Q8's drain. Set the fine frequency adjustment (C70) by lightly coupling a frequency counter to U7, pin 7. Next, connect an antenna to J4 and adjust the receiver filter trimmers (C1 and C2) for maximum signal. The AGC circuitry normally requires no adjustment, but the no-signal gain of the IF amplifier can be increased by decreasing the value of R3.

Before beginning transmitter alignment, set the drive-level control, R14, to minimum. Key the rig while monitoring the transmitted signal on a separate receiver and peak the transmit band-pass filter using C33 and C36. Then, with a dummy load or well-matched antenna connected to J4, set R14 to about 90% of maximum and check the output power level. It may be necessary to stagger-tune C33 and C36 on the lower bands in order to obtain constant output power across the desired tuning range. On 80 m the –3 dB transmit bandwidth will probably be less than 150 kHz.

Typically, output on 80, 40 and 20 m is 2.0-2.5 W, and on the higher bands 1.0-2.0 W. Some builders have obtained higher outputs on all bands by modifying the band-pass filters. However, filter modification may compromise spectral purity of the output, so the results should be checked with a spectrum analyzer. Also, note that the Sierra was designed to be a 2-W rig: additional RF shielding and decoupling may be required if the rig is operated at higher power levels.

PERFORMANCE

The Sierra design uses a carefully selected set of compromises to keep complexity low and battery life long. An example is the use of NE602 mixers, which affects both receive and transmit performance. On receive, the RF gain will occasionally need reduction when strong signals overload the receive mixer. On transmit, ARRL Lab tests show that the rig complies with FCC regulations for its power and frequency ranges.

Aside from the weak receive mixer, receiver performance is very good. There are no spurious signals (birdies) audible on any band. ARRL Lab tests show that the Sierra's receiver has a typical MDS of about –139 dBm, blocking dynamic range of up to 112 dB and two-tone dynamic range of up to 90 dB. AGC range is about 70 dB.

The Sierra's transmitter offers smooth break-in keying, along with direct transmit signal monitoring. There are two benefits to direct monitoring:

- the clean sinusoidal tone is easier on the ears than most sidetone oscillators and
- the pitch of the monitor tone is the correct receive-signal pitch to listen for when calling other stations.

The TR mute delay capacitor, C27, can be reduced to as low as 4.7 µF to provide faster break-in keying if needed.

The prototype Sierra survived its christening at Field Day, 1994, where members of the Zuni Loop Expeditionary Force used it on 80, 40, 20 and 15 m. There, Sierra compared favorably to the Heath HW-9 and several older Ten-Tec rigs, having as good or better sensitivity and selectivity—and in most cases better-sounding sidetone and break-in keying. While the other rigs had higher output power, they couldn't touch the Sierra's small

[7] The alignment procedure given here is necessarily brief. More complete instructions are provided with the ARRL template package and the kit. See note 2 to obtain a template.

size, light weight and low power consumption. The Sierra has consistently received high marks from stations worked too, with reports of excellent keying and stability.

CONCLUSION

At the time this article was written, over 100 Sierras had been built. Many have been used extensively in the field, where the rig's unique features are an asset. For some builders, the Sierra has become the primary home station rig.

The success of the Sierra is due, in large part, to the energy and enthusiasm of the members of NorCal, who helped test and refine early prototypes, procured parts for the field-test units and suggested future modifications.[8] This project should serve as a model for other clubs who see a need for an entirely new kind of equipment, perhaps something that is not available commercially.

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[8] The author would like to acknowledge the contributions of several NorCal members: Doug Hendricks, KI6DS; Jim Cates, WA6GER; Bob Dyer, KD6VIO; Dave Meacham, W6EMD; Eric Swartz, WA6HHQ, Bob Warmke, W6CYX; Stan Cooper, K4DRD; Vic Black, AB6SO; and Bob Korte, KD6KYT.

A BROADBAND HF AMPLIFIER USING LOW-COST POWER MOSFETs

Many articles have been written encouraging experimenters to use power MOSFETs to build HF RF amplifiers.[1-8] That's because power MOSFETs—popular in the design of switching power supplies—cost as little as $1 each, whereas RF MOSFET prices start at about $35 each!

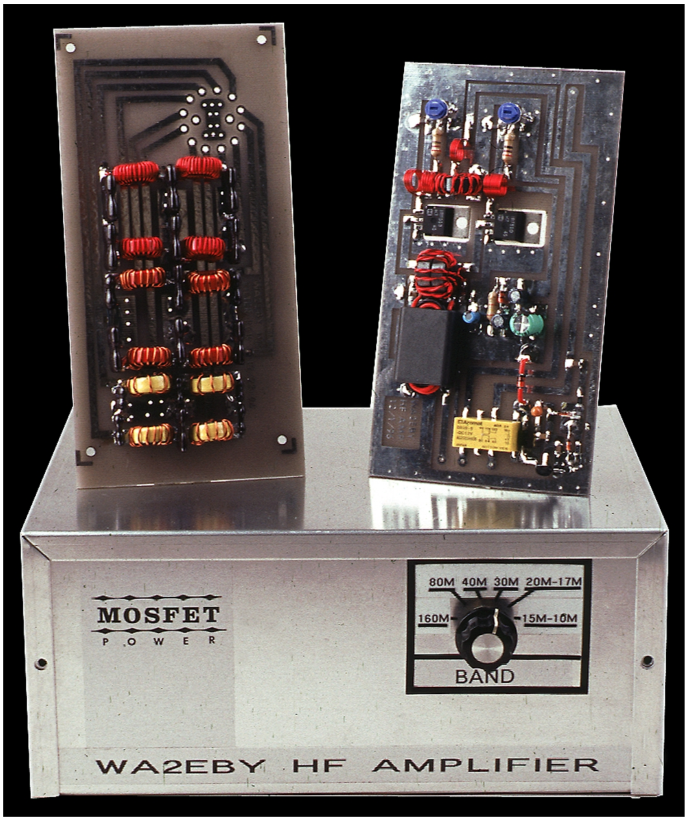
Mike Kossor, WA2EBY, designed and built this amplifier after hundreds of hours experimenting with power MOSFETs. The construction projects described in Notes 1 to 8, provide useful information about MOSFETs and general guidelines for working with them, including biasing, parasitic-oscillation suppression, broadband impedance-matching techniques and typical amplifier performance data.

With the design described here, 1 W of input power produces over 40 W of output (after harmonic filtering) from 160 through 10 meters. In addition to the basic amplifier, there is an RF-sensed TR relay and a set of low-pass filters designed to suppress harmonic output and comply with FCC requirements. The amplifier is built on double-sided PC board and requires *no tuning*. Another PC board contains the low-pass filters. Power-supply requirements are 28 V dc at 5 A, although the amplifier performs well at 13.8 V dc.

There are no indications of instability, no CW key clicks and no distortion on SSB has been reported by stations contacted while using the amplifier. To make it easy for you to duplicate this project, PC boards and parts kits are available, all at a cost of about $100![9] Etching patterns and parts-placement diagrams are included in the **References** chapter.

AN OVERVIEW OF MOSFETs

MOSFETs operate very differently from bipolar transistors. MOSFETs are voltage-controlled devices and exhibit a very high input impedance at dc, whereas bipolar transistors are current-controlled devices and have a relatively low input impedance. Biasing a MOSFET for linear operation only requires applying a fixed voltage to its gate via a resistor. With MOSFETs, no special bias or feedback circuitry is required to maintain the bias point over temperature as is required with bipolar transistors to prevent thermal runaway.[10] With MOSFETs, the gate-threshold voltage increases with increased drain current. This works to turn off the device, especially at elevated temperatures as transconductance decreases and $R_{DS(on)}$ (static drain-to-source *on* resistance) increases. These built-in self-regulating actions prevent MOSFETs from being affected by thermal runaway. MOSFETs do not require negative feedback to suppress low-frequency gain as is often required with bipolar RF transistors. Bipolar transistor gain increases as frequency decreases. Very high gain at dc and low frequencies can cause unwanted, low-frequency oscillation to occur in bipolar transistor RF amplifiers unless negative feedback is employed to prevent it. Low-frequency oscillation can damage bipolar transistors by causing excess power dissipation, leading to thermal runaway.



This rear-panel view of the amplifier shows the heat sink. The filter board mounts on the front panel.

MOSFET LIMITATIONS

Of course, MOSFETs do have their limitations. The high gate impedance and the device structure make them susceptible to electrostatic discharge (ESD) damage. Some easily applied precautions prevent this: Use a soldering iron with grounded tip; use a wrist strap connected to ground through a 1 MΩ resistor to bleed off excess body charge while handling MOSFETs and do all work on an anti-static mat connected to ground via a 1 MΩ resistor.

The sensitivity of a MOSFET's gate to static and high-voltage spikes also makes it vulnerable to damage resulting from parasitic oscillation. This undesired self-oscillation could result in excessive gate-to-source voltage that permanently damages the MOSFET's gate insulation. Another MOSFET limitation is gate capacitance. This parameter limits the frequency at which a MOSFET can operate effectively as an RF amplifier. I recommend reviewing the referents of Notes 1 to 3 if you are interested in more detailed information about MOSFETs.

POWER MOSFET RF AMPLIFIERS

The author built several power MOSFET amplifiers to check their performance. His experiments underscore the need to observe *exact* construction techniques and physical layout if similar performance is to be expected. Although he used PC board construction, his results differed significantly in several of the experiments because the circuit layout was not the same as the original layout.

Considerable experimentation (and I do mean considerable!) with several designes resulted in the circuit shown in **Fig 17.94**. This amplifier consists of two power MOSFETs operating in push-pull, and employs an RF-sensed TR relay.

During receive, TR relay K1 is deenergized. Signals from the antenna are connected to J2 and routed through K1 to a transceiver connected to J1. (This path loss is less than 0.3 dB from 1.8 MHz through 30 MHz.) In transmit, RF voltage from the transceiver is sampled by C17 and divided by R6 and R7. D2 and D3 rectify the RF voltage and charge C16. Q3 begins to conduct when the detected RF voltage across C16 reaches approximately 0.7 V. This energizes K1, which then routes the transmitted RF signal from J1 to the amplifier input and switches the amplifier output to the low-pass filter block and then to the antenna at J2. RF-sensed relay response is very fast. No noticeable clipping of the first CW character has been reported.

An RF attenuator (consisting of R8, R9 and R10) allows you to adjust the amplifier input power to 1 W. (The parts list contains resistor values to reduce the output of 2 or 5 W drivers to 1 W.) The 1 W signal is then applied to the primary of T1 via an input impedance-matching network consisting of L3. T1 is a 1:1 balun that splits the RF signal into two outputs 180º out of phase. One of these signals is applied by C1 to the gate of Q1. The other signal is routed via C2 to the gate of Q2. The drains of Q1 and Q2 are connected to the primary of output transformer T3, where the two signals are recombined in phase to produce a single output. T3 also provides impedance transformation from the low output impedance of the MOSFETs to the 50 Ω antenna port. DC power is provided to the drains of Q1 and Q2 by a phase-reversal choke, T2. This is a very effective method to provide power to Q1 and Q2 while presenting a high impedance to the RF signal over a broad range of frequencies. The drain chokes for Q1 and Q2 are wound on the same core, and the phase of one of the chokes (see the phasing-dot markings on T2) is reversed. C9 increases the bandwidth of the impedance transformation provided by T3, especially at 21 MHz.

The 5 V bias supply voltage is derived from 28 V by Zener diode D1 and current-limiting resistor R11. Bypass capacitors C3, C4, C5, C6 and C13 remove RF voltages from the bias supply voltage. Gate bias for Q1 and Q2 is controlled independently. R1 adjusts the gate-bias voltage to Q1 via R3 and L1. R2 works similarly for Q2 via R4 and L2.

At low frequencies, the amplifier input impedance is essentially equal to the series value of R3 and

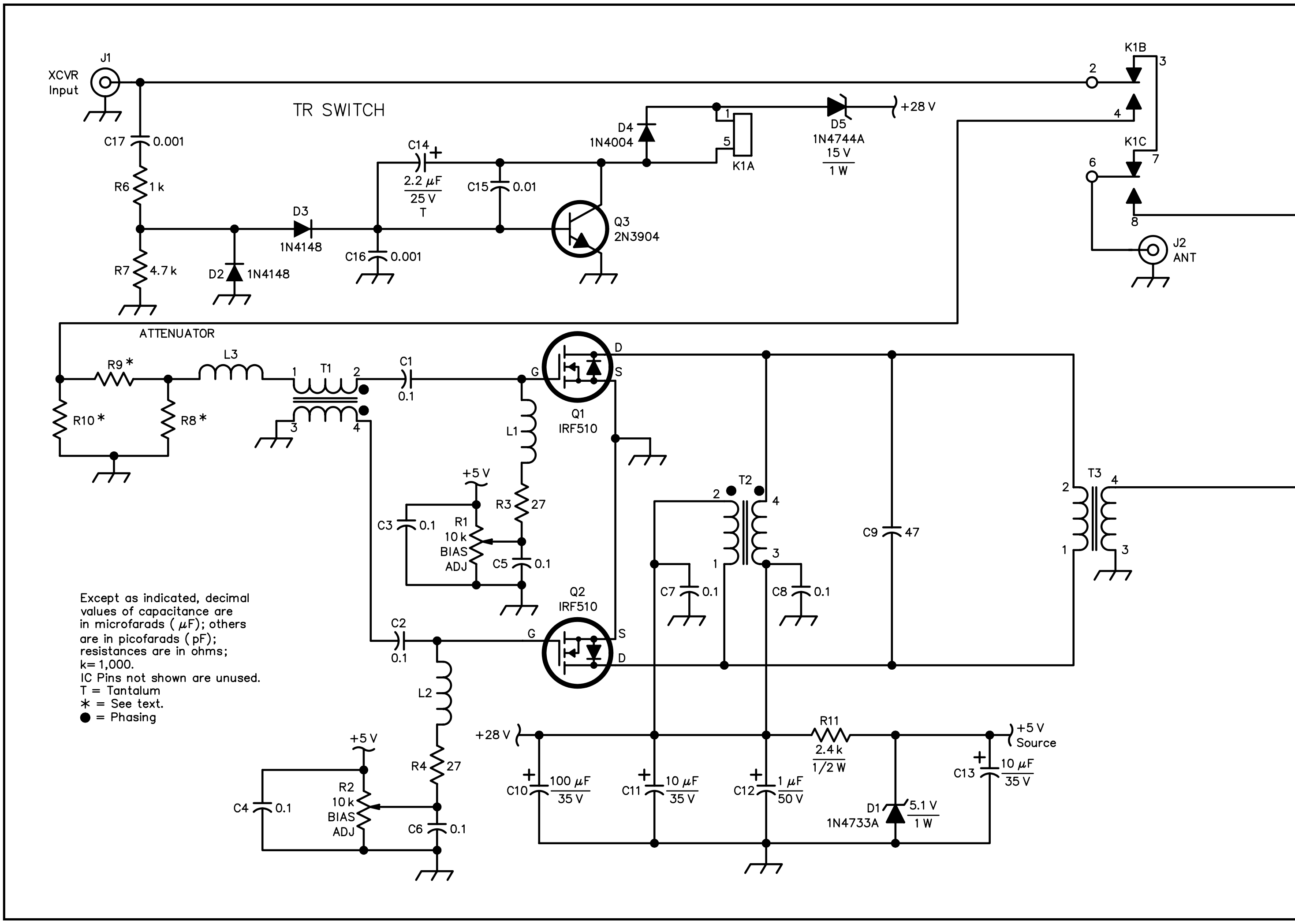
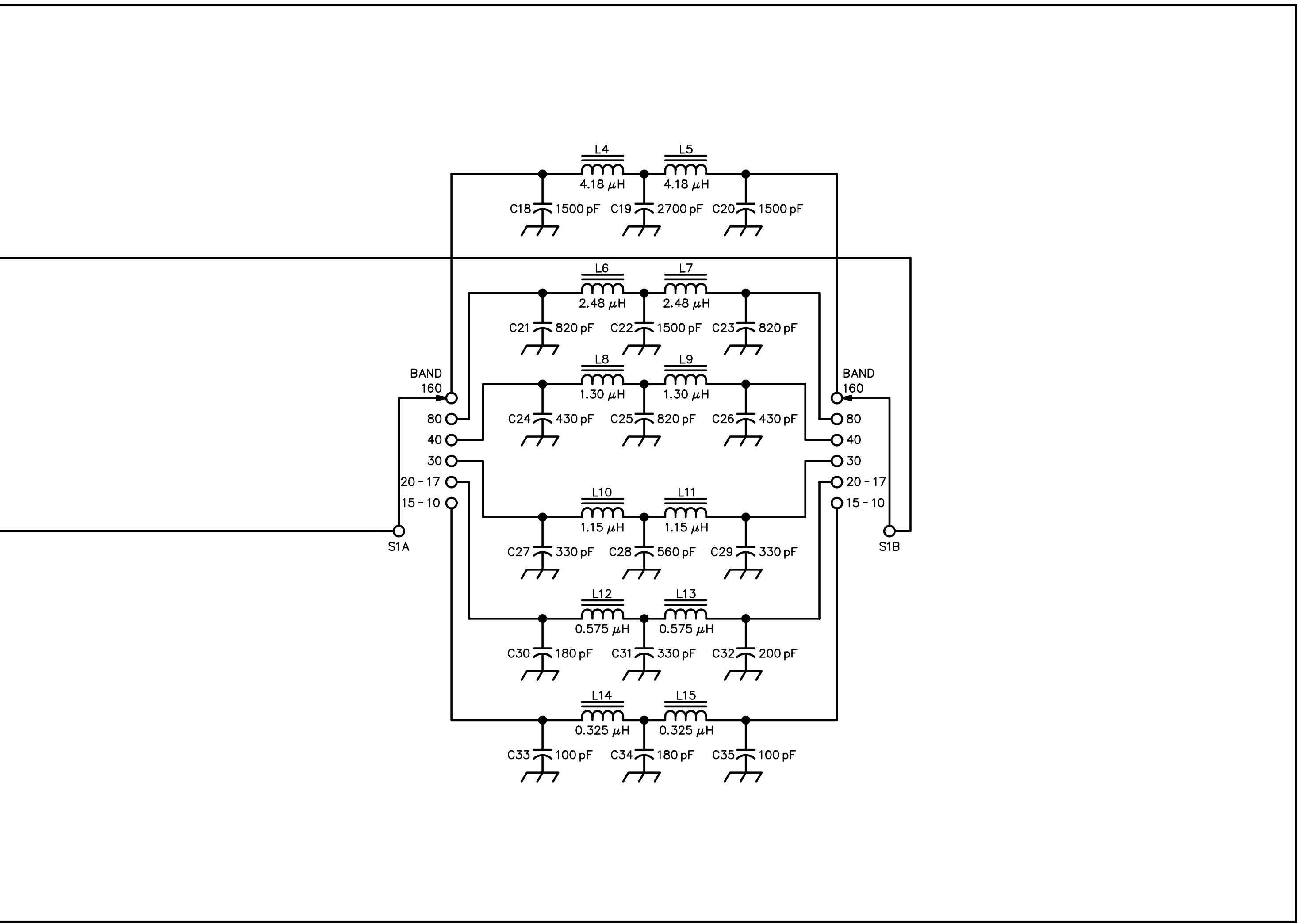


Fig 17.94—Schematic of the MOSFET all-band HF amplifier. Unless otherwise specified, resistors are $^1/_4$ W, 5% tolerance carbon-composition or film units. The low-pass filter section shows some filter component values that differ from the calculated values of a standard 50 Ω-input filter. Such differences improve the impedance matching between the amplifier and the load. Capacitors in the filter section are all dipped mica units. Equivalent parts can be substituted. Part numbers in parentheses are Mouser; see Note 9 and the References chapter for contact information.

C1-C8—0.1 µF chip (140-CC502Z104M)
C9—47 pF chip (140-CC502N470J)
C10—100 µF, 35 V (140-HTRL35V100)
C11, C13—15 µF, 35 V (140MLR35V10)
C12—1 µF, 50 V (140-MLRL50V1.0)
C14—2.2 µF, 35 V tantalum (581-2.2M35V)
C15—0.01 µF chip (140-CC502B103K)
C16, C17—0.001 µF chip (140-CC502B102K)
C18, C20, C22—1500 pF (5982-19-500V1500)
C19—2700 pF (5982-19-500V2700)
C21, C23, C25—820 pF (5982-19-500V820)
C24, C26—430 pF (5982-15-500V430)
C27, C29, C31—330 pF (5982-19-500V330)
C28—560 pF (5982-19-500V560)
C30, C34—180 pF (5982-15-500V180)
C32—200 pF (5982-15-500V200)
C33, C35—100 pF (5982-10-500V100)
D1—1N4733A, 5.1 V, 1 W Zener diode (583-1N4733A)
D4—1N4004A (583-1N4004A)
D2, D3—1N4148 (583-1N4148)
D5—1N4744A, 15 V, 1 W Zener diode (583-1N4744A)
J1, J2—SO-239 UHF connector (523-81-120) or BNC connector (523-31-10)
K1—12 V DPDT, 960 W coil, 12.5 mA (431-OVR-SH-212L)
L1, L2—9$^1/_2$ turns #24 enameled wire, closely wound 0.25-in. ID
L3—3$^1/_2$ turns #24 enameled wire, closely wound 0.190-in. ID
Q1, Q2—IRF510 power MOSFET (570-IRF510)
Q3—2N3904 (610-2N3904)
R1, R2—10 kΩ trim pot (323-5000-10K)
R3, R4—27 Ω, $^1/_2$ W (293-27)
R6—1 kΩ chip (263-1K)
R7— 4.7 kΩ chip (263-4.7K)
R8—130 Ω, 1 W (281-130); for 7 dB pad (5 W in, 1 W out)



R9—43 Ω, 2 W (282-43); for 7 dB pad (5 W in, 1 W out)
R10—130 Ω, 3 W (283-130); for 7 dB pad (5 W in, 1 W out)
R8, R10—300 Ω, $^1/_2$ W (273-300); for 3 dB pad (2 W in, 1 W out)
R9—18 Ω, 1 W (281-18); for 3 dB pad (2 W in, 1 W out)
R11—2.4 kΩ, $^1/_2$ W (293-2.4K)
S1—2 pole, 6 position rotary (10YX026)
T1—10 bifilar turns #24 enameled wire on an FT-50-43 core.
T2—10 bifilar turns #22 enameled wire on two stacked FT-50-43 cores.
T3—Pri 2 turns, sec 3 turns #20 Teflon-covered wire on BN-43-3312 balun core.
Misc: Aluminum enclosure 3.5 × 8 × 6 inches (HWD) (537-TF-783), two TO-220 mounting kits (534-4724), heat-sink compound (577-1977), amplifier and low-pass filter PC boards (see Note 9), heat sink (AAVID [Mouser 532-244609B02]; see text), about two feet of RG-58 coax, #24 enameled wire and #20 Teflon-insulated wire.

R4. L1 and L2 improve the input-impedance match at higher frequencies. The low value of series resistance provided by R3 and R4 also reduces the Q of impedance-matching inductors L1 and L2, which improves stability. DC blocking capacitors C1 and C2 prevent loading the gate bias-supply voltage.

C14 keeps transistor Q3 conducting and K1 energized between SSB voice syllables or CW elements. Without C14, K1 would chatter in response to the SSB modulation envelope and fast keying. Increasing the value of C14 increases the time K1 remains energized during transmit. The reverse voltage generated by K1 when the relay is deenergized is clamped to a safe level by D4. D5 drops the 28 V supply to 13 V to power 12 V relay K1. D5 can be replaced with a jumper if K1 has a 28 V dc coil or if you intend to operate the amplifier with a 13.8 V dc supply.

HARMONIC FILTERING

Although biased for class AB linear operation, this amplifier (like others of its type) exhibits some degree of nonlinearity, resulting in the generation of harmonics. This push-pull amplifier design cancels even-order harmonics (2f, 4f, 6f, etc) in the output transformer, T3. Odd-order harmonics are not canceled. Second-order harmonics generated by the amplifier are typically less than –30 dBc (30 dB below the carrier) whereas third-order harmonics are typically only –10 dBc. FCC regulations require all HF RF-amplifier harmonic output power to be at least –40 dBc at power levels between 50 to 500 W. To meet this requirement, it is common practice for HF amplifiers to use low-pass filters. Separate low-pass filters are needed for the 160, 80, 40 and 30 meter bands. The 20 and 17 meter bands can share the same low-pass filter. So, too, the 15, 12 and 10 meter bands can share a common low-pass filter; see Fig 17.94.

Switching between the six filters can be a messy wiring problem, especially on the higher-frequency bands where lead lengths should be kept short for optimum performance. This problem is solved by mounting all six low-pass filters on a PC board. A two-pole, six-position rotary switch (S1) mounted directly on the same PC board manages all filter interconnections. One pole of S1 connects the amplifier output to one of the six filter inputs, while the other pole of S1 simultaneously connects the corresponding filter output to the TR relay, K1. Only two coaxial-cable connections are required between the RF amplifier and the low-pass filter board.

AMPLIFIER CONSTRUCTION

The amplifier is constructed on a double-sided PC board with plated through holes to provide top-side ground connections. Chip resistors and capacitors were used to simplify construction, but leaded capacitors may work if lead lengths are kept short. First, assemble all chip capacitors and resistors on the PC board. Tweezers help to handle chip components (or see the vacuum handler for chip components described in the **Station Setup and Accessory Projects** chapter. Work with only one component value at a time. (Chip caps and resistors can be very difficult to identify!) Chip capacitor and resistor mounting is simplified by tinning one side of the PC board trace with solder before positioning the capacitor or resistor. Touch the soldering iron tip to the capacitor or resistor to tack it in place. Finish mounting by soldering the opposite side of the component. *Don't apply too much heat to chip capacitors*. The metalized contacts on the capacitor can be damaged or completely removed if too much heat is applied. Use a 15 to 20 W soldering iron and limit soldering time to five seconds.

Mount axial-leaded resistors, diodes and remaining capacitors next. To avoid damaging them, mount inductors and transformers last. L1 and L2 are wound on a $^1/_4$ inch drill-bit shaft. By wrapping the wire around the shaft 10 times, you'll get $9^1/_2$ turns. The last turn arcs only a half-turn before entering the PC board. L3 is wound on a 0.190-inch diameter drill bit with $3^1/_2$ turns wound the same way as L1 and L2. Mounting K1 is simplified by first bending all its leads 90° outward so it lies flat on the PC board. Be sure to follow the anti-static procedures mentioned at the beginning of this project while handling MOSFETs. The gate input can be damaged by electrostatic discharge!

When winding T3, wind the primary first and add the secondary winding over the primary. Be sure to use Teflon-insulated wire for these windings; the high operating temperatures encountered will likely melt standard hook-up wire insulation.

Heat Sinking

Together, Q1 and Q2 dissipate up to 59 W. A suitable heat sink is required to prevent the transistors from overheating and damage. I used an AAVID 244609B02 heat sink originally designed for dc-to-dc power converters. The amplifier PC board and heat sink are attached to an aluminum enclosure by two #4-40 screws drilled through the PC board, enclosure and heat sink at diagonally opposite corners. See **Photo A**. A rectangular cutout in the enclosure allows Q1 and Q2 direct access to the heat sink. This is essential because of the large thermal impedance associated with the TO-220 package (more on this topic later). Mark the locations of the transistor-tab mounting-hole location in the center of the heat sink, between the cooling fins. Disassemble the heat sink to drill 0.115 inch holes for #4-40 mounting screws, or tap #4-40 mounting holes in the center of the heat-sink fins.

Use mica insulators and grommets when mounting Q1 and Q2 to prevent the #4-40mounting screws from shorting the TO-220 package drain connections (tabs) to ground. Coat both sides of the mica insulator with a *thin* layer of thermal compound to improve the thermal conduction between the transistor tab and the heat sink. Be sure to install the mica insulator on the heat sink *before* assembling the amplifier PC board to the enclosure and heat sink. The mica insulators are larger than the cut outs in the PC board, making it impossible to install them after the PC board is mounted.

LOW-PASS FILTER CONSTRUCTION

Inductor winding information for the low-pass filters is provided in **Table 17.6**.

Single Band

A PC-board trace is available on the amplifier PC board next to amplifier output (J3) to allow the installation of a single-band low-pass filter between the terminals of J3 and the J4 input to K1. This is handy if you intend to use the amplifier on one band only. The input inductor of the low-pass

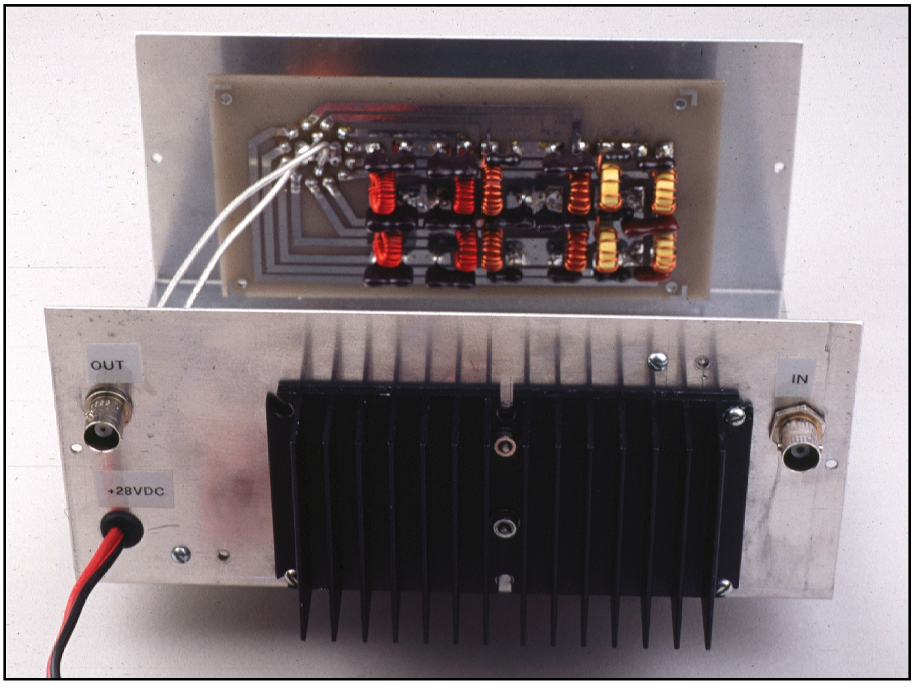


Photo A

Table 17.6
Low-Pass Filter Inductor Winding Information
(Refer to Fig 17.94)

Inductor Number	*No. of Turns*	*Core*
L4, L5	30	T-50-2
L6, L7	22	T-50-2
L8, L9	16	T-50-2
L10, L11	14	T-50-2
L12, L13	11	T-50-6
L14, L15	8	T-50-6

Note: All inductors are wound with #22 enameled wire except for L4 to L7, which are wound with #24 enameled wire.

filter connects from J3 to the single PC trace adjacent to J3. The output inductor connects in series between the single PC trace to J4. The three filter capacitors connect from J3, J4 and the PC-board trace near J3 to ground. *This single trace is not used when multiple filters are required*. Remember to remove the single trace adjacent to J3 on the amplifier PC board before attaching the amplifier board between the RF connectors on the rear panel of the enclosure.

Multiple-Band Filters

Using the amplifier on more than one band requires a different approach. A set of six low-pass filters is built on a double-sided PC board with plated through holes to provide top-side ground connections. A PC-board mount, two-pole, six-position rotary switch does all low-pass filter selection. Silver-mica, leaded capacitors are used in all the filters. On 160 through 30 meters, T-50-2 toroids are used in the inductors. T-50-6 toroids are used for inductors on 20 through 10 meters. The number of turns wound on a toroid core are counted on the toroid's OD as the wire passes through the core center. (The **Circuit Construction** chapter provides complete details for winding toroids.) Assemble one filter section at a time starting with the 160, 80, 40 and 30-meter filters. With the switch mounting position at your upper left, the 160-m filter input (C18) is near the top edge of the board and the filter output (C20) is near the bottom edge. *The last two filters are out of sequence;* the 15-10 meter filter comes *before* the 20-17 meter filter) and the inputs/outputs are reversed to simplify the PC-board layout. The input capacitors, C30 and C33, are mounted on the board *bottom edge*, and output capacitors, C32 and C35, are on the *top edge*.

Use care when assembling the rotary switch. All 14 terminals must fit through the PC board without damaging or bending the pins. Make sure there are no bent pins before you attempt assembly. Insert the rotary switch into the PC board. Do *not* press the rotary switch all the way into the PC-board holes flush with the ground plane! If you do, the top flange of the signal pins may short to the ground plane.

BIAS ADJUSTMENT

The biasing procedure is straightforward and requires only a multimeter to complete. First, set R1 and R2 fully counterclockwise, (0 V on the gates of Q1 and Q2). Terminate the RF input and output with 50 Ω loads. Next, connect the 28 V supply to the amplifier in series with a multimeter set to the 0-200 mA current range. Measure and record the idling current drawn by the 5 V bias supply. The value should be approximately 9.5 mA (28 – 5.1 V) / 2.4 kΩ = 9.5 mA). Set the Q1 drain current to 10 mA by adjusting R1 until the 28 V supply current increases by 10 mA above the idling current (9.5 + 10 = 19.5 mA). Next, adjust R2 for a Q2 drain current of 10 mA. This is accomplished by adjusting R2 until the 28 V supply current increases by an additional 10 mA (to 29.5 mA).

AMPLIFIER PERFORMANCE

With a 28 V power supply and 1 W of drive, the RF output power of this amplifier exceeds 40 W from 1.8 MHz through 28 MHz. Peak performance occurs at 10 MHz, providing about 75 W after filtering! A performance graph for this amplifier is shown in **Fig 17.95A**.

As shown in Fig 17.95B, this amplifier achieves an efficiency of better than 50% over its frequency range, except at 7 MHz where the efficiency drops to 48%.

Fig 17.95C shows the input SWR of the amplifier. It exceeds 2:1 above 14 MHz. The input SWR can be improved to better than 2:1 on all bands by adding a 3 dB pad (R8-R10 of Fig 17.94) at the input and supplying 2 W to the pad input. This keeps the amplifier drive at 1 W.

Fig 17.95D graphs the amplifier RF output power as a function of drain supply voltage. During this test, the amplifier RF drive level was kept constant at 1 W. As you can see, even when using a 13.8 V dc supply, the amplifier provides over 10 W output (a gain of more than 10 dB) from 1.8 to 30 MHz.

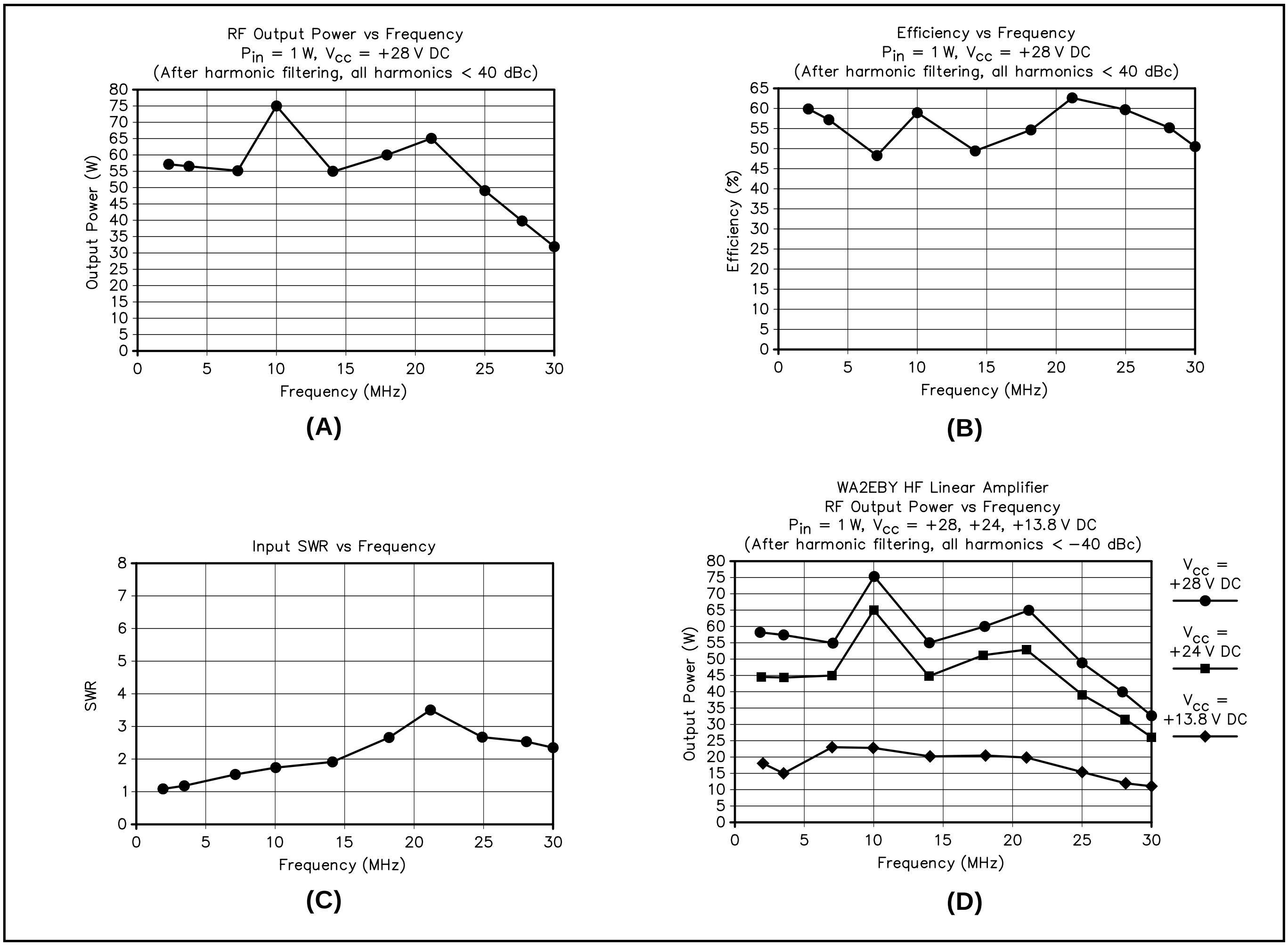


Fig 17.95—A shows the amplifier RF output power. B shows the amplifier efficiency. C shows the amplifier input SWR. D shows the amplifier RF output power versus supply voltage.

OPERATION

The amplifier requires no tuning while operating on any HF amateur band. You must, however, *be sure to select the proper low-pass filter prior to transmitting*. If the wrong low-pass filter is selected, damage to the MOSFETs may result. Damage will likely result if you attempt to operate the amplifier on a band with the low-pass filter selected for a lower frequency. For example, driving the amplifier with a 21 MHz signal while the 1.8 MHz low-pass filter is selected will likely destroy Q1 and/or Q2.

The amplifier can also be damaged by overheating. This limitation is imposed by the TO-220 packages in which Q1 and Q2 are housed. The thermal resistance from junction to case is a whopping 3.5°C/W. This huge value makes it virtually impossible to keep the junction temperature from exceeding the +150°C target for good reliability. Consider the following conditions: key down, 1 W input, 53 W output on 7 MHz (worst-case band for efficiency). The amplifier consumes 28 V × 4 A = 112 W, of which 53 W are sent to the antenna, so 59 W (112 W – 53 W = 59 W) are dissipated in Q1 and Q2. Assuming equal current sharing between Q1 and Q2, each transistor dissipates 29.5 W. To keep the transistor junction temperature below +150°C requires preventing the transistor case temperature from exceeding 46.8°C (150 – [3.5 × 29.5]) while dissipating 29.5 W. Also, there is a temperature rise across the mica insulator between the transistor case and heat sink of 0.5°C/W. That makes the maximum allowable heat-sink temperature limited to 46.8 – (0.5 × 29.5) = 32°C. In other words, the heat sink must dissipate

59 W (29.5 from each transistor) with only a 7°C rise above room temperature (25°C). Even if the junction temperatures were allowed to reach the absolute maximum of 175°C, the heat sink temperature must not exceed 57°C. Accomplishing this requires a heat sink with a thermal resistance of (57 – 25) / 59 = 0.54°C/W. This is far less than the 1.9°C/W rating of the AAVID 244609B02 heat sink I used. The situation may seem bleak, but all is not lost. These calculations make it clear that the amplifier should not be used for AM, FM or any other continuous-carrier operation. The amplifier should be used only for CW and SSB operation where the duty cycle is significantly reduced.

Thermal performance of the amplifier is illustrated in **Fig 17.96A**. Data was taken under dc operating conditions with power-dissipation levels set equal to conditions under RF operation. A RadioShack brushless 12 V dc fan (RS 273-243A) blows air across the heat sink. Key down, the maximum rated junction temperature is reached in as little as five seconds. Prolonged key-down transmissions should be avoided for this reason.

Under intermittent CW conditions, the situation is very different. Transistor-case temperatures reached 66°C after operating four minutes under simulated CW conditions at 20 WPM (60 ms on, 60 ms off). The corresponding junction temperature is +141°C (based on an equivalent RMS power dissipation of 21.7 W per transistor). This keeps the junction temperature under the 150°C target (see Fig 17.96B). One simple way to reduce power dissipation is to reduce the power-supply voltage to 24 V. RF output power will decrease about 10 W from the maximum levels achieved with a 28 V supply.

From a thermal standpoint, the IRF510 power MOSFET is a poor choice for this RF amplifier application. Although I must say I am impressed with the robustness of these devices considering the times I spent testing them key down, five minutes at a time, without failure. Q1 and/or Q2 may need to be replaced after a year or so of operation because of the compromise in reliability. Considering their low cost, that is not a bad trade-off.

STABILITY

High gain, broad bandwidth and close input/output signal routing (within the TR relay) all work against stability. With a good load (< 2:1 SWR) the amplifier is stable from 1.8 MHz through 39 MHz. Oscillation was observed when the transmitter frequency was increased to 40 MHz. The output load match also affects stability. I spent a great deal of time trying to make this design unconditionally stable

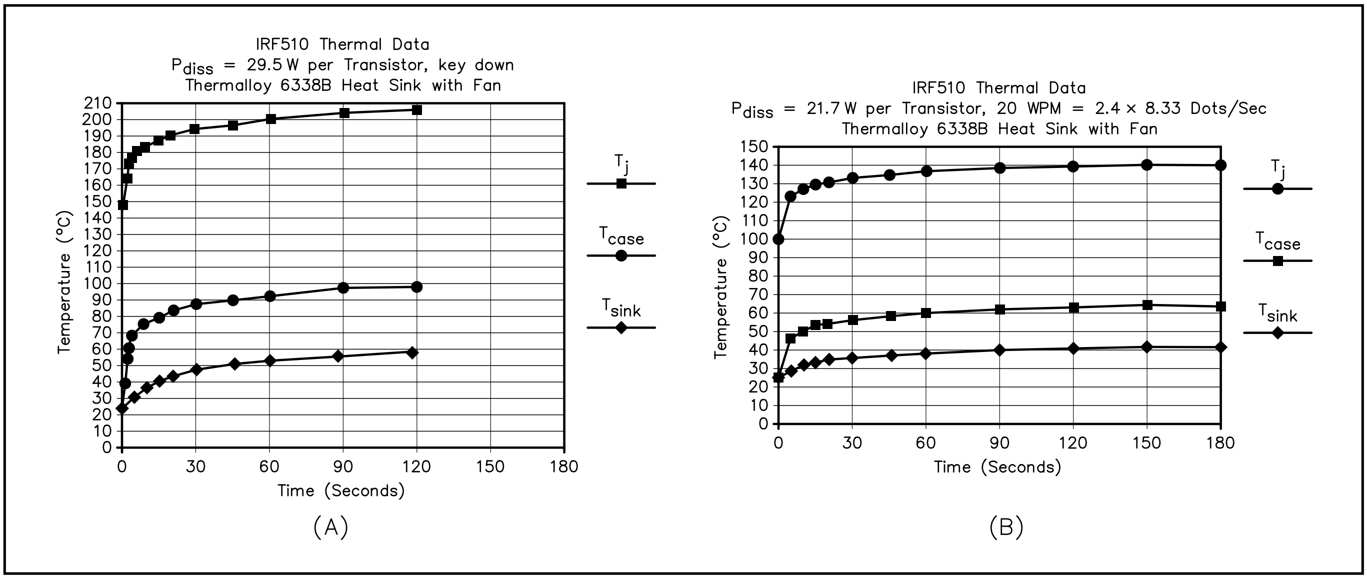


Fig 17.96—A is a graph of the amplifier thermal performance during key-down conditions. B is a graph of the amplifier thermal performance during simulated CW conditions.

even with loads exceeding 3:1 SWR without sacrificing output power (gain) at 28 MHz without success. I did identify some reasonable compromises.

One of the easiest ways to improve stability and the input SWR seen by the RF source is to add an RF attenuator (pad) at the amplifier input. An attenuator is absolutely required if the transmitter (driver) provides more than 1 W to the amplifier. R8, R9 and R10 form an RF attenuator that attenuates the transmitter drive level, but does not attenuate received signals because it is only in the circuit when K1 is energized. To drive this amplifier with a 2-W-output transmitter requires use of a 3-dB pad. The pad improves the amplifier input SWR and the isolation between the amplifier's input and output. The drawback is that 1 W is wasted in the pad. Likewise, a 5-W driver requires use of a 7-dB pad, but 4 W are wasted in the pad. (Values for R8, R9 and R10 to make a 3-dB pad and a 7-dB pad are given in the parts list of the caption for Fig 17.94.) Installing a pad requires cutting the PC-board trace *under R9*, otherwise R9 would be shorted out by the trace. Make a small cut (0.1 inch wide) in the trace under R9 before soldering R9 in position. R8 and R10 have the same values, but may have different power ratings. Connect R10 between the RF input side of R9 and ground. Install R8 between the amplifier side of R9 and ground.

An impedance mismatch between the output of a 1-W driver and the amplifier input can be a source of instability. (Obviously, if the driving transmitter output power is only 1 W, you can't use a pad as described earlier.) If you encounter stability problems, try these remedies: Place a resistor in parallel with L1 and L2 to decrease the Q of the amplifier matching network (try values between 50 and 220 Ω). Try reducing the value of L3 or eliminating L3 entirely. Both of these modifications improve stability, but reduce the amplifier output power above 21 MHz.

SUMMARY

This project demonstrates how inexpensive power MOSFETs can be used to build an all-band linear HF power amplifier. Frequency of operation is extended beyond the limits of previous designs using the IRF510 and improved input-impedance matching. Long-term reliability is recognized as a compromise because of the poor thermal performance of the low-cost TO-220 package.

If you have been thinking about adding an amplifier to your QRP station, this project is a good way to experiment with amplifier design and is an excellent way to become familiar with surface-mount "chip" components. Mouser Electronics and Amidon, Inc provide parts kits for this project (see Note 9). These parts kits make it very easy to get started and more economical to "homebrew" this project.

ACKNOWLEDGMENTS

The author thanks the following individuals associated with this project: Harry Randel, WD2AID, for his untiring support in capturing the schematic diagram and parts layout of this project; Al Roehm, W2OBJ, for his continued support and encouragement in developing, testing, editing and publishing this project; Larry Guttadore, WB2SPF, for building, testing and photographing the project; Dick Jansson, WD4FAB, for thermal-design suggestions; Adam O'Donnell, N3RCS, for his assistance building prototypes; and his wife, Laura, N2TDL, for her encouragement and support throughout the project.

Notes

[1] Doug DeMaw, W1FB, "Power-FET Switches as RF Amplifiers," *QST*, Apr 1989, pp 30-33. See also Feedback, *QST*, May 1989, p 51.

[2] Wes Hayward, W7ZOI, and Jeff Damm, WA7MLH, "Stable HEXFET RF Power Amplifiers," Technical Correspondence, *QST*, Nov 1989, pp 38-40; also see Feedback, *QST*, Mar 1990, p 41.

[3] Jim Wyckoff, AA3X, "1 Watt In, 30 Watts Out with Power MOSFETs at 80 Meters," Hints and Kinks, *QST*, Jan 1993, pp 50-51.

[4] Doug DeMaw, W1FB, "Go Class B or C with Power MOSFETs," *QST*, March 1983, pp 25-29.

[5] Doug DeMaw, W1FB, "An Experimental VMOS Transmitter", *QST*, May 1979, pp 18-22.
[6] Wes Hayward, W7ZOI, "A VMOS FET Transmitter for 10-Meter CW," *QST*, May 1979, pp 27-30.
[7] Ed Oxner, KB6QJ (ex-W9PRZ), "Build a Broadband Ultralinear VMOS Amplifier," *QST*, May 1979, pp 23-26.
[8] Gary Breed, K9AY, "An Easy-to-Build 25-Watt MF/HF Amplifier," *QST*, Feb 1994, pp 31-34.
[9] Parts for this project are available in five modular kits. The following three kits are available from Mouser Electronics: Amplifier components (Mouser P/N 371-HFAMP1) consisting of the amplifier PC board and all PC-board-mounted components (except for the ferrite cores). Price: $35, plus shipping. Amplifier hardware kit (Mouser P/N 371-HFAMP2) consisting of the aluminum enclosure, two UHF connectors, two TO-220 mounting kits, AAVID heat sink and one container of heat sink compound. Price: $30 plus shipping. Low-pass filter kit (Mouser P/N 371-HFAMP3) consisting of the low-pass filter PC board, rotary switch and all PC-board-mounted capacitors (inductor cores are *not* included). Price: $35, plus shipping. Part-placement diagrams accompany the PC boards. See the **References** chapter for contact information.

PC boards only are available from Mouser Electronics: HF amplifier board (#371-AMPPWB-2); filter PC board (#371-LPPWB-2). Price $15 each, plus shipping.

The following two kits are available from Amidon Inc: Amplifier ferrite kit (Amidon P/N HFAFC) containing the ferrite cores, balun core and magnet and Teflon wire to wind the transformers for the HF amplifier. Price: $3.50 plus shipping. Low-pass filter cores kit (Amidon P/N HFFLT) containing all iron cores and wire for the low-pass filters. Price: $4.50 plus shipping. See the **References** chapter for contact information.

[10] Motorola Application Reports Q1/95, HB215, *Application Report AR346*.

Thermal runaway is a condition that occurs with bipolar transistors because bipolar transistors conduct more as temperature increases, the increased conduction causes an increase in temperature, which further increases conduction, etc. The cycle repeats until the bipolar transistor overheats and is permanently damaged.

AN EXPERIMENTAL $^1/_2$-W CW TRANSMITTER

Using the transmitter shown in **Fig 17.97**—a 74HC240 octal inverting buffer IC operating as a crystal oscillator and power amplifier—Lew Smith, N7KSB, has worked all continents and over 30 countries using just a roof-mounted ground-plane antenna. It is a simple, inexpensive means of having fun on 10, 15 or 20 m.

It's important to operate the circuit at a supply voltage from 7.8 to 8 V as a compromise between maximum power output and device safety.

The logic chips have built-in input and output buffers. The extra gain provided by the extra stage makes it harder to get rid of key clicks. The $^1/_2$-W rig's key-click filter therefore uses an unusually large (33-ms) time constant.

The output stages in 74HC*xxx* devices are designed to have equal value pull-up and pull-down transistors. This minimizes even-order harmonics, simplifying the $^1/_2$-W rig's output filtering.

The 74HC240 can directly drive a power MOSFET amplifier, as exemplified by the 74HC240-IRF510 transmitter described in "An Easy-to-Build 15-W Transmitter" (*Hambrew* magazine, Spring 1994).

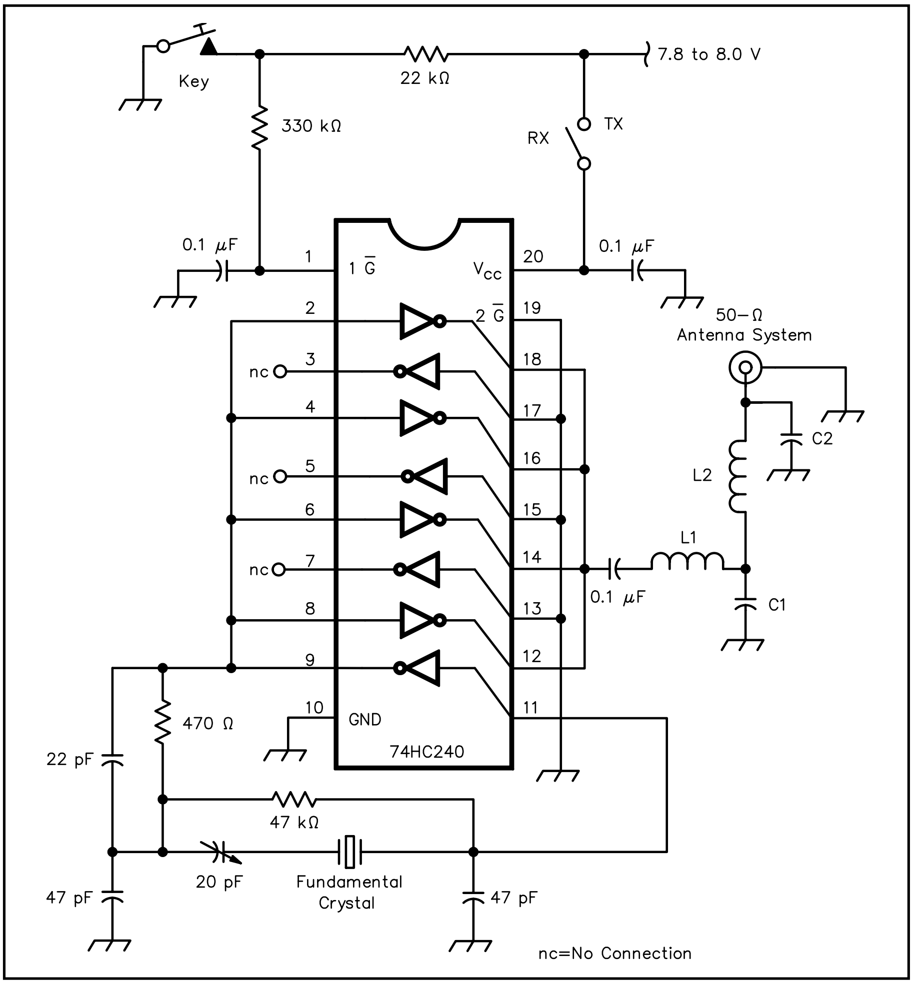


Fig 17.97—N7KSB's experimental $^1/_2$-W CW transmitter uses a 74HC240 high-speed CMOS octal buffer: One section serves as a crystal oscillator; four sections amplify the signal; three sections are unused. (See Table 17.7 for output filter values.) Because the IC dissipates $^1/_2$ W on 20 m and 0.9 W on 10 m, it needs a heat sink. (Epoxy the IC to the ground plane, dead-bug style.) L1 is sensitive to lead length, so plug-in filters (Lew uses phono plugs and jacks) may require removal of one or two turns to compensate for the extra lead length. (ARRL Lab tests indicate that this transmitter meets current FCC spectral-purity specifi-cations for equipment in its power-output class and frequency range. The test version produced 0.51 W at 14 and 21 MHz, and 0.47 W at 28 MHz, with a 7.83-V supply.)

Table 17.7

Output Filter Component Values for the 74HC240 Transmitter

Band (m)	*10*	*15*	*20*
C1 (pF)	330	470	680
C2 (pF)	100	150	220
L1 (turns, length)	3 t, $^5/_8$″	4 t, $^5/_8$″	5.5 t, $^5/_8$″
L2	7 t, $^5/_8$″	10 t, 1″	12 t, 1″

C1 and C2 are mica or ceramic; L1 and L2 are #14 wire with air cores ($^3/_8$″ diam form).

A DRIFT-FREE VFO

By following several design guidelines, Jacob Makhinson, N6NWP, built a low-cost, easy-to-construct LC VFO with a very low level of phase noise. The article originally appeared in December 1996 *QST*.

The method shown makes the oscillator essentially drift-free, with very little phase noise, VFOs built with these techniques are viable in applications where low overall noise level and wide dynamic range is of great importance.

The technique can also spare VFO designers the drudgery of more conventional drift-compensating techniques.

Many VFO designs have appeared in the Amateur Radio literature, and the quest for a low-drift VFO hasn't ceased. If the frequency-stability requirements are stringent, the thermal-drift compensation can be very tedious. Wes Hayward's *QST* article[1] devoted to VFO drift compensation is an excellent example of this difficult pursuit.

DESIGN CRITERIA

To avoid degradation of the receiver's front end, several requirements should be imposed on the phase noise level of the VFO. An excessively high level of close-in phase noise (within the bandwidth of the SSB signal) may reduce the receiver's ability to separate closely spaced signals. As an example, a 14-pole crystal filter described in Note 2 provides adjacent-signal rejection of 103 dB at a 2-kHz offset. This requires the use of a VFO with –139 dBc/Hz phase noise at a 2-kHz offset.

$$Pn = P - 10 \log(BW) = -103 - 10 \log(4000) = -139 \text{ dBc/Hz}$$

where

Pn = VFO phase-noise spectral density, in decibels relative to the carrier output power, in a 1-Hz bandwidth (dBc/Hz)

P = VFO power level (dBc) in a given bandwidth (BW)

BW = test bandwidth, in Hertz

In addition, excessive close-in phase noise may lead to reciprocal mixing, where the noise sidebands of a VFO mix with strong off-channel signals to produce unwanted IF signals.

Excessive far-out phase noise may degrade the receiver dynamic range. In a properly designed receiver, the phase-noise-governed dynamic range (PNDR) should be equal to or better than the spurious-free dynamic range (SFDR). We can calculate the PNDR:[3]

$$PNDR = -Pn - 10 \log(BW)$$

Assuming the PNDR equals the SFDR at 112 dB in a 2.5-kHz IF noise bandwidth, the required far-out phase noise level is –146 dBc/Hz:

$$Pn = -SFDR - 10 \log(BW) = -112 - 34 = -146 \text{ dBc/Hz}$$

Another form of VFO instability—frequency drift—has always been a nuisance and a great concern to the amateur community. The objective of this project was to keep the long-term frequency drift (seconds, minutes, hours) under 20 Hz. This includes thermal drift from both internal heating and environmental changes.

BLOCK DIAGRAM

The block diagram of **Fig 17.98A** shows the LC VFO and the frequency stabilizer. The stabilizer monitors the VFO frequency and forms an error signal that is applied to the VFO to compensate for frequency drift. This technique, which is capable of stabilizing a VFO to within a few hertz, was devised by Klaas Spaargaren, PA0KSB, and first described in *RadCom* magazine in 1973.[4] My project builds upon Spaargaren's idea and presents a few refinements.

The stabilizer converts a free-running VFO into an oscillator that can be tuned in the usual fashion, but then locks to the nearest of a series of small frequency steps. Unlike traditional PLL frequency synthesizers, the stabilizer has no effect on the phase-noise performance of the VFO; it only compensates for thermal drift.

The timing signal (2.6 Hz) is derived from a crystal oscillator via a frequency divider. The timing signal drives a NAND gate to provide a crystal-controlled time window, during which the binary counter counts the VFO output. When the gate closes, the final digit of the count remains in the counter. For counts 0 to 3, the Q3 output of the counter is a logic 0; for counts of 4 to 7, a logic 1.

The result is stored in a D flip-flop memory cell: When the 2.6-Hz timing signal goes low, the first of three one-shots triggers. The second follows and clocks the binary counter Q3 output into the memory cell. The negative-going pulse from the third resets the counter for the next counting sequence.

The output of the memory cell is applied to an RC integrating circuit with a time constant of several minutes. This slowly changing dc voltage controls the VFO frequency via a couple of Varicaps connected to a tap on the VFO coil.

If the counter output is 0, the memory-cell output is 1, which charges C and increases the VFO frequency. A counter output of 1 discharges C and decreases the VFO frequency. The stabilizer constantly searches for equilibrium, so the VFO frequency slowly swings a few hertz around the lock frequency. The circuit limits the frequency swing to a maximum of ±2 Hz, typically ±1 Hz.

A difficulty arises when the operator changes frequency because the control voltage is disturbed. If the memory-cell output connects directly to the RC integrator, the frequency correction that occurs immediately after tuning results in a frequency hop. To overcome this problem, an analog switch disconnects the integrator from the memory during tuning. The tuning detector—an infrared interrupter switch and a one-shot—controls the analog switch.

CIRCUIT DESCRIPTION

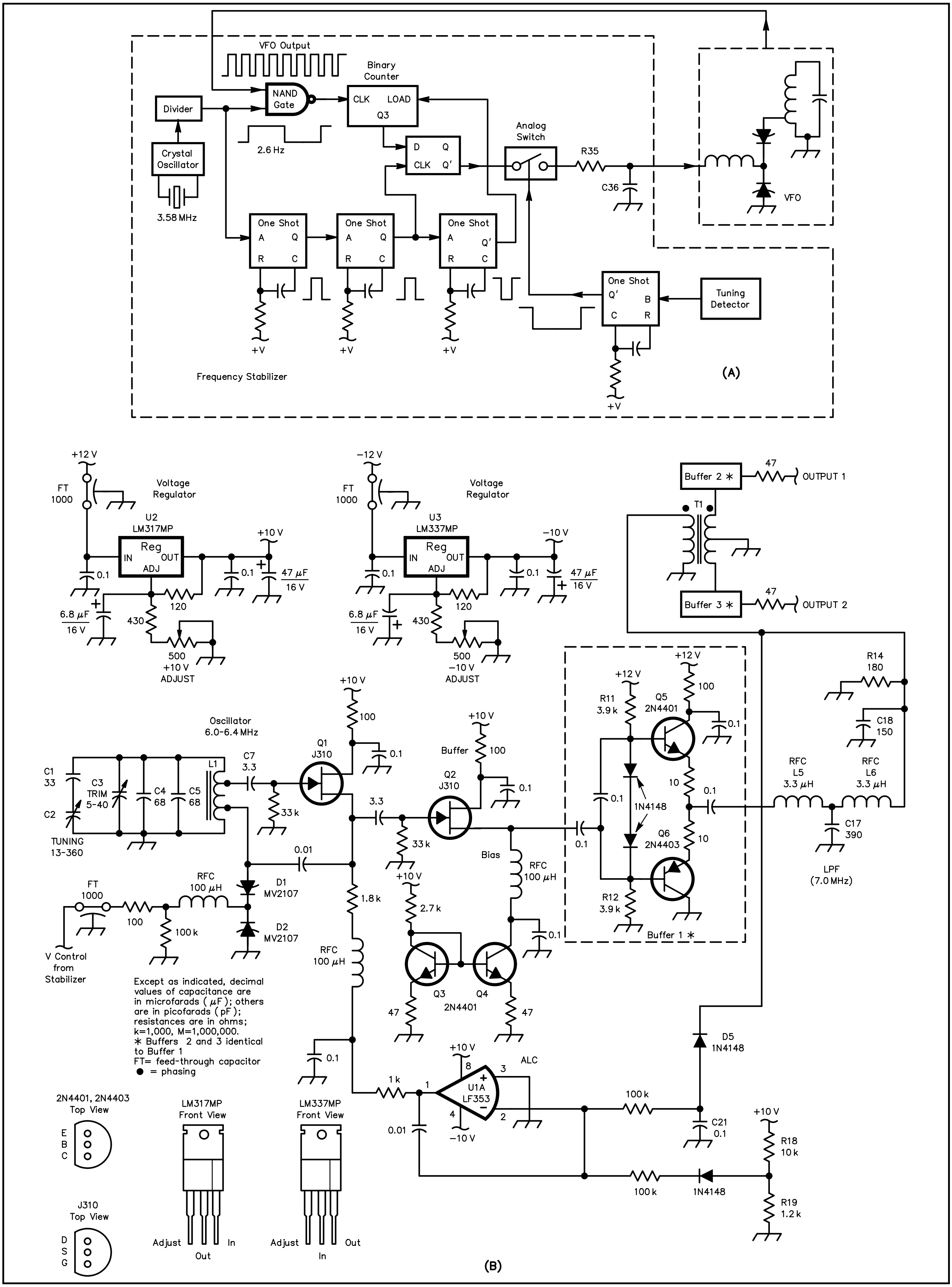
VFO

The VFO is a tapped-coil Hartley oscillator that is optimized for low phase noise (see **Fig 17.98B**). It follows the design rules compiled by Ulrich Rohde, intended to minimize the phase noise in oscillators.[5]

The tank coil, L1, has an iron-powder toroidal core; coil Q exceeds 300. C1, C4, C5 and C7 are NP0 (C0G) ceramic capacitors (5% or 10% tolerance). C2 is the main tuning capacitor, and C3 is a small ceramic trimmer capacitor.

Fig 17.98—At A—Block diagram of the frequency stabilizer and VFO connections. At B—VFO schematic. Buffers 2 and 3 are identical to Buffer 1. Most of the parts are available from Mouser Electronics, Digi-Key Corporation or Allied Electronics. The cores for L1 and T1 are from Amidon Associates.[6] Use 1/4-W, 5%-tolerance carbon-composition or film resistors and ceramic, 20%-tolerance capacitors unless otherwise indicated. RF chokes or encapsulated inductors may be used for those labeled "RFC."

Q1, Q2—J310, N-channel JFET (Allied)
D1, D2—MV2107 or ECG/NTE613 tuning diode (Varicap, Allied)
L1—29 turns of #18 AWG enameled copper wire on a T-80-6 iron-powder toroidal core tapped at 4 turns and 20 turns from the cold end (Amidon)
T1—#32 AWG enameled copper wire on a BN-43-2402 two-hole ferrite balun core (Amidon) primary: 5 turns; secondary: 16 turns, center tapped
Vector part #8007 circuit board (Digi-Key)
Vector part #T44 terminals (Digi-Key)


VFO Output
Binary Counter
NAND Gate
CLK
LOAD
Q3
Divider
2.6 Hz
Crystal Oscillator
3.58 MHz
D Q
CLK Q′
Analog Switch
R35
C36
VFO
One Shot
A Q
R C
One Shot
A Q′
R C
+V
One Shot
Q′ B
C R
Tuning Detector
Frequency Stabilizer
(A)
+12 V
FT 1000
Voltage Regulator
U2 LM317MP
Reg
IN OUT
ADJ
+10 V
0.1
47 μF 16 V
6.8 μF 16 V
120
430
500 +10 V ADJUST
−12 V
U3 LM337MP
−10 V
500 −10 V ADJUST
Buffer 2 ∗
47
OUTPUT 1
T1
Buffer 3 ∗
OUTPUT 2
+12 V
100
R11 3.9 k
Q5 2N4401
R14 180
C18 150
10
1N4148
Q6 2N4403
R12 3.9 k
Buffer 1 ∗
RFC L5 3.3 μH
RFC L6 3.3 μH
C17 390
LPF (7.0 MHz)
+10 V
Oscillator 6.0-6.4 MHz
Q1 J310
C7 3.3
C1 33
C2
C3 TRIM 5-40
C4 68
C5 68
L1
TUNING 13-360
33 k
Buffer
Q2 J310
Bias
RFC 100 μH
0.01
D1 MV2107
D2 MV2107
FT 1000
RFC 100 μH
100
100 k
V Control from Stabilizer
1.8 k
2.7 k
Q3 Q4 2N4401
47
D5 1N4148
ALC
U1A LF353
8 3 1 4 2
1 k
100 k
C21 0.1
R18 10 k
R19 1.2 k
1N4148
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.
∗ Buffers 2 and 3 identical to Buffer 1
FT= feed-through capacitor
● = phasing
2N4401, 2N4403 Top View
E B C
J310 Top View
D S G
LM317MP Front View
Adjust In Out
LM337MP Front View
Adjust Out In
(B)

The VFO frequency range is set from 6.0 MHz to 6.4 MHz (to accommodate a 20-meter receiver with an 8-MHz IF). The loaded Q of the resonator is kept high by using a tapped coil and loose coupling to the gate of the FET through C7 (more than 8 kΩ at 6 MHz). The RF voltage swing across the resonator exceeds 50 V, P-P. Varicaps D1 and D2, which compensate for thermal drift, are connected across the coil's lower tap (less than 14% of the total turns) and have a negligible effect on overall phase-noise. J310 is the TO-92 version of U310—a very low-noise FET in HF applications.

An ALC loop limits the voltage swing. The signal is sampled at the primary of T1, rectified by the D5-C21 network and fed to the inverting input of an integrator, U1A, where it is compared against the reference voltage at the junction of R18 and R19. The dc voltage at the integrator output sets Q1's drain current so that the signal swing at T1's primary is always 2.5 V, P-P. The ALC loop also makes VFO performance independent of Q1's pinch-off voltage. The signal at Q1's source is a 6.5-V, P-P, sinusoid with almost no distortion.

Q2 is a high-impedance buffer that is loosely coupled to Q1. Q2's drain current is set to 3.4 mA (by the constant-current source, Q3-Q4) regardless of Q2's pinch-off voltage.

Buffer 1 is a push-pull stage biased into slight conduction by resistors R11 and R12. It has excellent linearity and a very low output impedance, which is required to drive an LC filter. The filter (L5, L6, C17, C18 and R14) is a four-pole, 0.1-dB Chebyshev low-pass filter with a ripple frequency of 7 MHz. All harmonics at the VFO output are at least 45 dB below the fundamental.

T1 provides the two complementary outputs required for a commutation mixer and raises the voltage swing at the VFO output.

Buffers 2 and 3 are electrically identical to buffer 1. They further decouple the VFO from its load and serve as low-distortion 50-Ω drivers. The signal level at each output is 4 V, P-P, when driving a high-impedance load (eg, a CMOS gate), +10 dBm when driving a 50-Ω load.

FREQUENCY STABILIZER

NAND gates U4A and B (see Fig 17.99) comprise a Pierce crystal oscillator. The timing signal (2.6 Hz) appears at the output of the frequency divider (U5, U6, U7 and U8A). The exact frequency of the crystal and the timing signal is unimportant, but the stabilizer has been optimized for 2.3 to 2.7 Hz.

There are two requirements for the crystal oscillator: No harmonics should fall in the IF passband, and the crystal should have a low temperature coefficient. Crystal-oscillator thermal drift should not exceed 10 Hz within the temperature operating range. Crystals in HC-33 cases with frequencies between 2.0 and 3.58 MHz worked best for me. The frequency divider is sufficiently flexible to provide the desired timing-signal frequency.

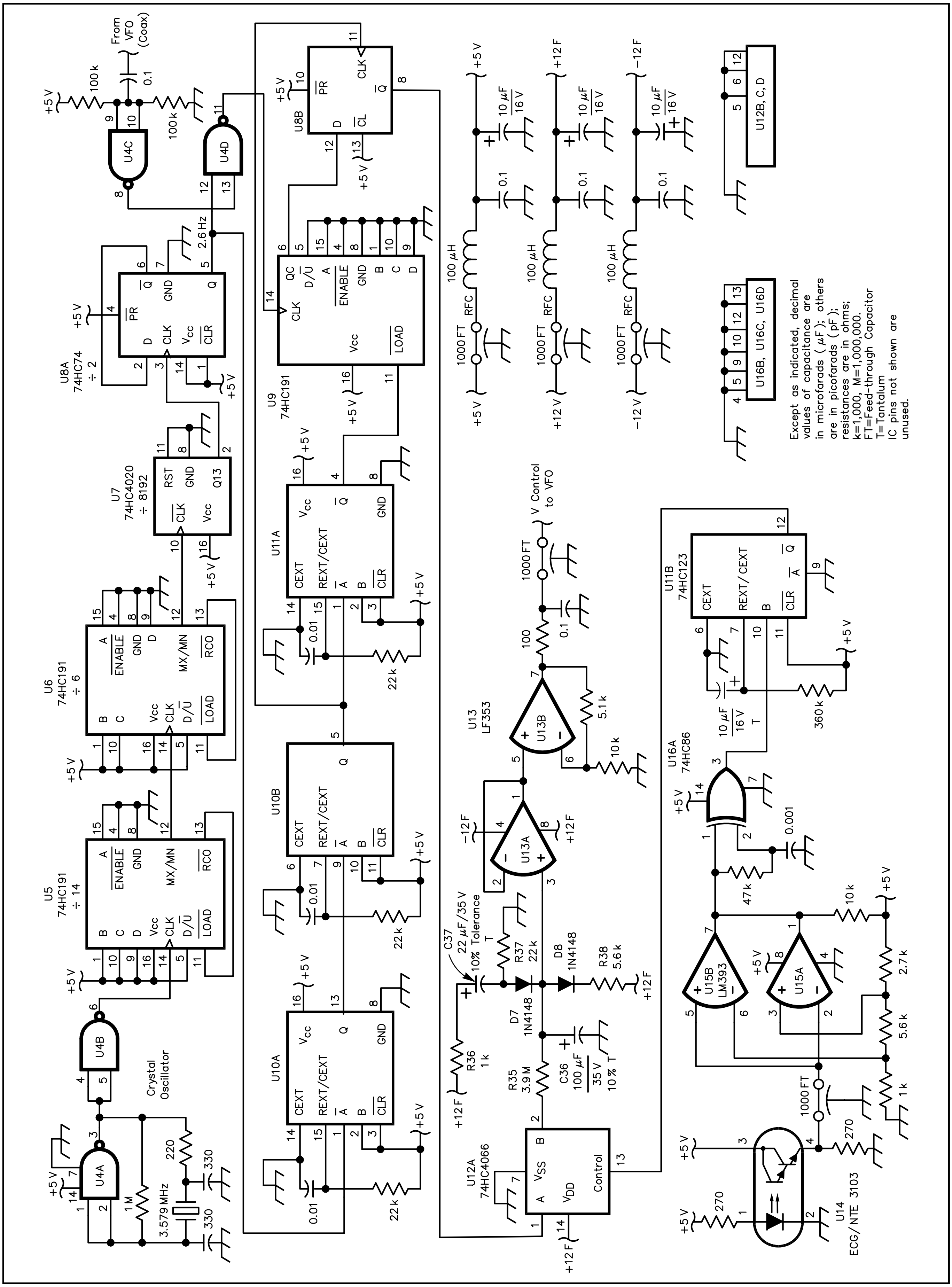
U4C, biased into a linear range, converts the sinusoidal signal from one of the two VFO outputs into a square wave. U4D gates the VFO signal bursts into the clock input of the binary counter, U9. At the end of every burst, the final digit is held by the counter.

The falling edge of the timing signal triggers U10A, the first of three cascaded one-shots. The pulse at the output of U10B clocks the data from the counter into U8B. The pulse at the output of U11A resets U9.

If the number of pulses in each successive burst is equal (no VFO drift), U9 constantly counts the same

Fig 17.99—Stabilizer schematic. Use 1/4-W, 5%-tolerance carbon-composition or film resistors and ceramic, 20%-tolerance capacitors unless otherwise indicated.

U4—74HC00 Quad NAND gate
U5, U6, U9—74HC191 presettable 4-bit binary counter
U7—74HC4020 14-bit binary ripple counter
U8—74HC74 dual D flip-flop
U10, U11—Dual 74HC123 one-shot
U12—4066 quad analog switch
U14—ECG/NTE3103 optical interrupter (Darlington output, Allied, see Note 6)



From VFO (Coax)
Crystal Oscillator
3.579 MHz
U4A
U4B
U4C
U4D
U5 74HC191 ÷ 14
U6 74HC191 ÷ 6
U7 74HC4020 ÷ 8192
U8A 74HC74 ÷ 2
U8B
U9 74HC191
U10A
U10B
U11A
U11B 74HC123
U12A 74HC4066
U12B, C, D
U13 LF353
U13A
U13B
U14 ECG/NTE 3103
U15A
U15B LM393
U16A 74HC86
U16B, U16C, U16D
2.6 Hz
V Control to VFO
C37 22 μF/35 V 10% Tolerance T
C36 100 μF 35 V 10% T
R35 3.9 M
R36 1 k
R37 22 k
R38 5.6 k
D7 1N4148
D8 1N4148
100 μH RFC
1000 FT
Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000. FT=Feed-through Capacitor T=Tantalum IC pins not shown are unused.

number, and the output of U8B never changes. In practice, however, U8B constantly toggles between two states. The integrating circuit, R35-C36 (time constant = 6.5 minutes), converts the toggling into a slowly changing voltage. Varicaps D1 and D2 transform a few millivolts of change into ±1 or 2 Hz change of VFO frequency.

U13A, a high-input-impedance buffer, prevents the discharge of C36. U13B, a noninverting amplifier with a gain of 1.5, ensures compliance between the control-voltage range and the capacitance-per-volt ratio of the Varicaps (1 to 6 V for best performance). Network R36, R37, C37 and D7 establishes the initial dc voltage applied to the varicaps; the value is set by the C37-C36 voltage divider.

An infrared interrupter switch, U14, serves as sensor in the tuning-detector circuit. The slotted interrupter detects the movements of a serrated disc (see **Fig 17.100**) on the VFO reduction-drive shaft. U15A and B, a two-level limit comparator, converts the signal at its input into pulses. U16A produces trigger pulses for the one-shot, U11B, by detecting both leading and falling edges of the signal at its input. U11B is retrig-gerable—its Q' output stays low during manual tuning and for 3.6 seconds after tuning stops. Analog switch U12A disconnects C36 from the flip-flop during tuning, thus preserving the capacitor charge. This system does not provide for an RIT control.

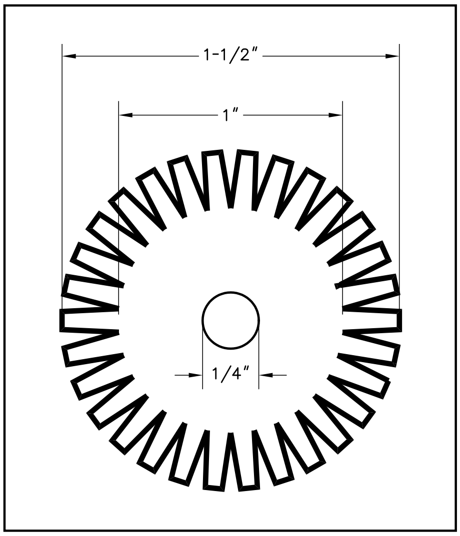


Fig 17.100—Mechanical details of interrupter wheel. Use a good reduction drive and make one tooth for every 20 to 40 Hz of frequency change. Use any rigid, opaque material.

CONSTRUCTION

The VFO and the stabilizer are in separate boxes. Mount components within the enclosures on the perf board's foil side. Make ground connections to the foil plane. Use Vector pins as terminal posts for the input and output signals.

The VFO box is a die-cast aluminum enclosure ($4^{11}/_{16} \times 3^{11}/_{16} \times 2^{1}/_{16}$ inches) to ensure mechanical rigidity. The two RF outputs exit the box via BNC connectors and coax. DC enters via feedthrough capacitors. Rigidly attach C2 to the enclosure wall. Cover L1 with a low-loss polystyrene Q dope and place it as far as possible from the ground plane and enclosure walls. The layout is not critical, but observe standard RF building methods: use short leads, dress them for minimum coupling and solder bypass capacitors directly to the ground plane close to the terminal they bypass.

The stabilizer is in a $5^{1}/_{2} \times 3 \times 1^{1}/_{4}$-inch LMB aluminum enclosure. Component placement and layout is not critical, but keep component leads short around the crystal oscillator. Use a BNC connector for the signal from the VFO module. Solder the ground pins of all ICs directly to the ground plane, and decouple each power-supply pin of ICs U4 through U9 to ground via a 0.1-mF capacitor. Route all dc voltages to the module via feedthrough capacitors to avoid RF leakage. Mount U14 so that the serrated disc is in the middle of the slot.

Mount C37 in a socket in case you need to adjust its value: For unknown or varying VFO drift direction, use a 22-µF capacitor to place the initial varicap control voltage at midrange ($V_c \approx 2.9$ V). Use 10-µF if VFO drift is predominantly negative ($V_c \approx 1.5$ V), and 33-µF if it's predominantly positive ($V_c \approx 4.0$ V).

MEASUREMENTS

The VFO thermal drift without the stabilizer was under 800 Hz at room temperature (after 90 minutes) and under 1500 Hz when the ambient temperature was raised 20°C. There was no attempt to compensate for thermal drift.

With the frequency stabilizer connected, the thermal drift did not exceed 10 Hz at room temperature, 20 Hz when raised 20°C. In one of the experiments, power was on for several days, and drift was under 10 Hz at room temperature. Frequency lock is attained in less than 10 seconds after the power is switched on.

With the components shown in the schematic, the stabilizer can compensate for a maximum 1800-Hz drift with a 25°C temperature rise. To compensate for a greater frequency drift, select varicaps with higher diode capacitances; the frequency swing will increase from ±1 or 2 Hz to a higher value.

NOTES

[1] Wes Hayward, W7ZOI, "Measuring and Compensating Oscillator Frequency Drift," *QST*, Dec 1993, pp 37-41.

[2] Jacob Makhinson, N6NWP, "Designing and Building High-Performance Crystal Ladder Filters," *QEX*, Jan 1995, pp 3-17.

[3] Peter Chadwick, G3RZP, "Phase Noise Intermodulation and Dynamic Range," *Frequency Dividers and Synthesizers IC Handbook*, Plessey Semiconductors, 1988, p 151.

[4] Klaas Spaargaren, PA0KSB, "Technical Topics: Crystal-Stabilized VFO" *RadCom*, Jul 1973, pp 472-473. Comments followed in later "Technical Topics" columns. Also see, "Frequency Stabilization of L-C Oscillators," *QEX*, February 1996, pp 19-23.

[5] Ulrich Rohde, KA2WEU/DJ2LR, *Digital PLL Frequency Synthesizers* (Englewood Cliffs: Prentice-Hall, 1983) p 78.

[6] For contact information for Mouser Electronics, Digi-Key Corp, Allied Electronics and Amidon Inc, see the Address List in the **References** chapter.