Chapter 14

Receivers and Transmitters

In this chapter, William E. Sabin, WØIYH, discusses the "system design" of Amateur Radio receivers and transmitters. "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 14.1**, 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 Igen 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 Iout 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

$$V_{out} / V_{in} \tag{1}$$

In decibels (dB) it is

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

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 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 Ω). 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.

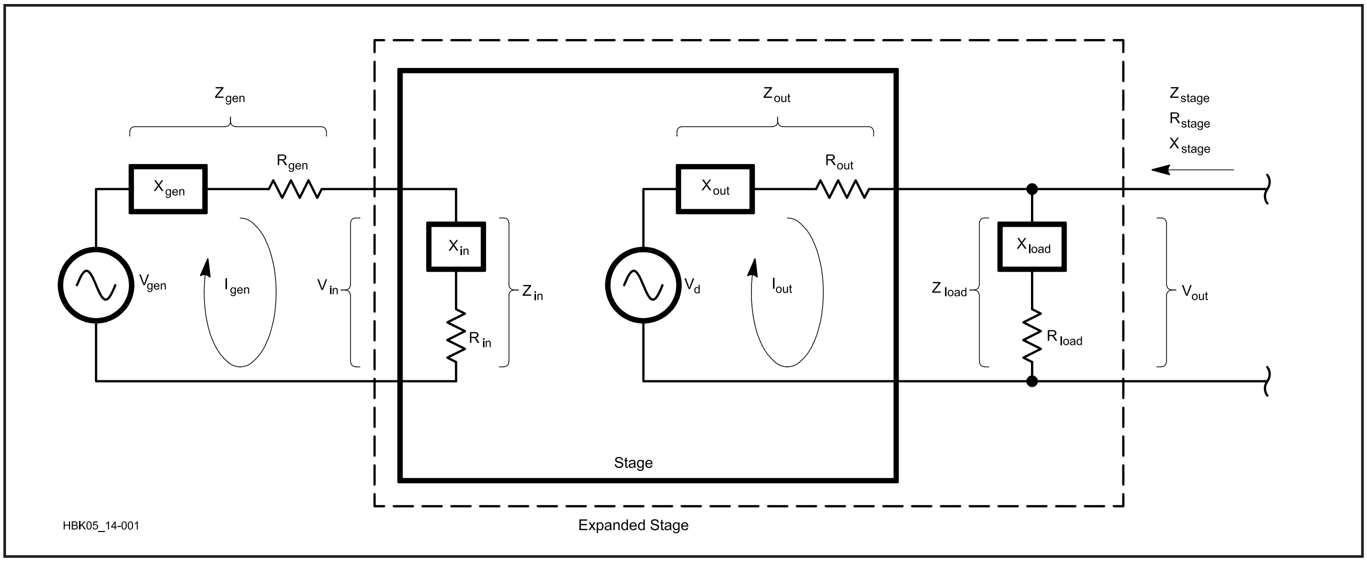

Fig 14.1—A single-stage building-block signal processor. The properties of this stage are discussed in the text.

Fig 14.2 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 14.1 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 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.

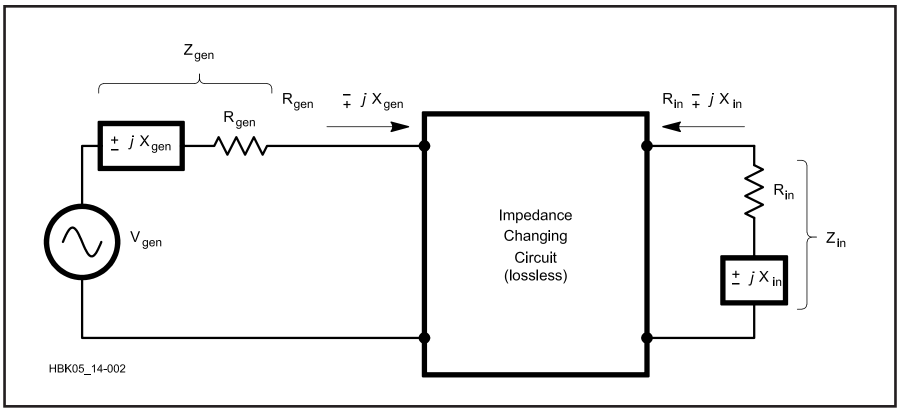

Fig 14.2—The conjugate impedance match of a generator to a stage input. The network input impedance is $R_{gen} \pm jX_{gen}$ and its output impedance is $R_{in} \pm jX_{in}$ (where either R term may represent a dynamic impedance). Therefore the generator and the stage input are both impedance matched for maximum power transfer.

Available Power Gain

Consider that in Fig 14.1 the stage and its output load Z_{load} constitute an "expanded" stage as defined by the dashed 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 14.1.

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 14.1, 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 14.1 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.

Feedback (Undesired)

One of the most important properties of the single-stage building block in Fig 14.1 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 14.1) 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 14.3**, a transistor amplifier, shown in skeleton form, with sharply tuned resonators at input and output.

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 14.3 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 14.4**.

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 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 "lossless feedback" circuits. These circuits use very efficient transformers instead of resistors or lossy networks that introduce thermal noise into a system.

Noise Factor and Noise Figure

The output resistance of the signal generator that drives a typical signal processing block such as shown in Fig 14.4 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

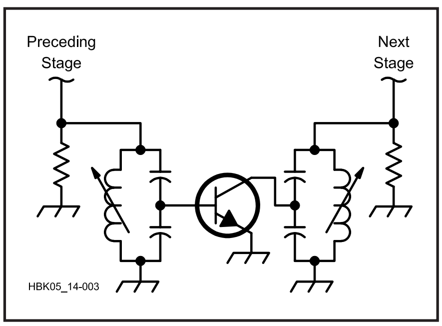

Fig 14.3—A double tuned transistor amplifier circuit that may oscillate due to excessive amplification and reverse coupling.

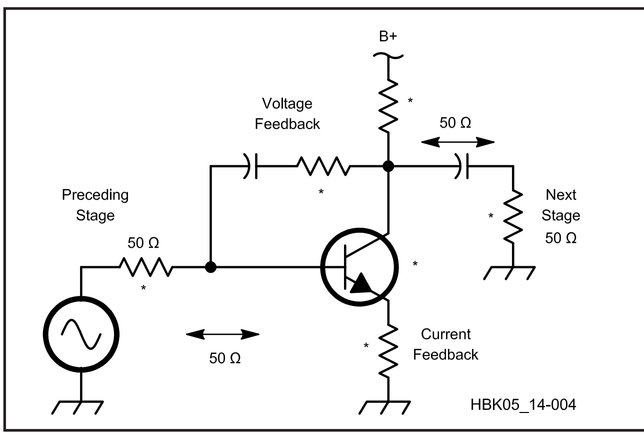

Fig 14.4—A cascadable amplifier using feedback. The feedback and therefore the amplifier performance depends on the load and driving-stage impedances.

Negative Feedback in RF Circuit Design

This sidebar shows how negative feedback can enhance the performance of the RF amplifier circuits that are used in homebrew receivers, transmitters or transceivers. The sidebar lists the advantages of negative feedback, and presents examples for the advantages. We will not consider automatic gain control (AGC) or automatic level control (ALC) circuits, often referred to as "envelope feedback" (see later sections of this chapter).

1. All amplifiers have nonlinearities that produce output frequencies (harmonics, intermodulation distortion and adjacent channel interference) of a certain amount that are not present in the input signal. Feedback can reduce the *percentage*.

2. Feedback can be used to increase or decrease the input impedance or output impedance of an amplifier stage. Actual L, C and R values are modified by feedback to "effective" values. The concept of a dynamic or lossless resistance that does not dissipate power or create thermal noise is introduced.

3. Feedback can improve the stability (freedom from a tendency to oscillate) of an amplifier. This includes the methods of feedback network compensation and also "neutralization" or "unilateralization" which means the reduction of internal feedback within the amplifier.

4. Feedback can make the frequency response, the input impedance and the output impedance of an amplifier more constant over a wide frequency band.

5. Feedback can reduce gain and frequency response changes due to temperature, supply voltage, value tolerances of inductors, capacitors and resistors and transistor parameter spreads. The performance variations over a large number of identical circuits are greatly reduced.

6. By controlling the performance of each stage in a multistage system, using feedback, the overall performance can be more accurately predicted and maintained with little or no "tweaking" of the individual stages.

Feedback in an amplifier stage nearly always reduces the gain (ratio of output to input voltage, current or power) to a lower value. For a certain overall gain requirement, more stages are required. This is in nearly all cases a mild penalty. Each extra stage is an additional source of the imperfections that were enumerated above. This means that the entire chain must be designed as a system in order to meet the system goals. Feedback can also be applied over two, or sometimes three cascaded stages, but is more difficult. In this brief overview we will focus mainly on single-stage feedback.

The Negative Feedback Concept

Fig A shows an amplifier with negative feedback. Prior to feedback, the amplifier has gain G and a phase reversal that we assume for simplicity is 180°. A portion of the output *voltage*, or perhaps a portion of the output *current*, goes through a feedback network to the input where it combines, possibly in *series* with or in *parallel* with the generator signal to produce a *modified* input signal that is smaller than it was with no feedback. This new input consists partly of generator signal, assumed to be perfect, and partly of an output signal, assumed to be imperfect. Note that the generator signal and the fed-back signal are in opposite phase. When this modified input signal is amplified (even though it is amplified imperfectly itself) the original imperfections that were

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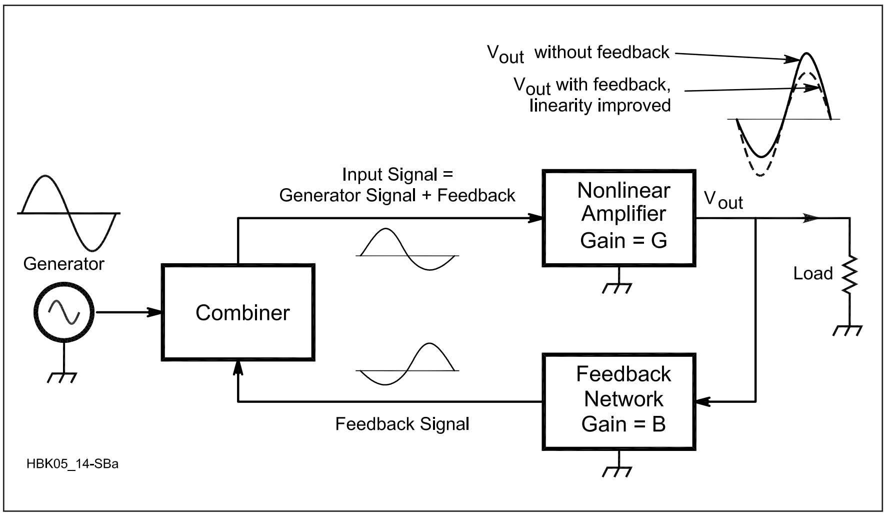


Fig A—Block diagram of a nonlinear amplifier with feedback to improve linearity.

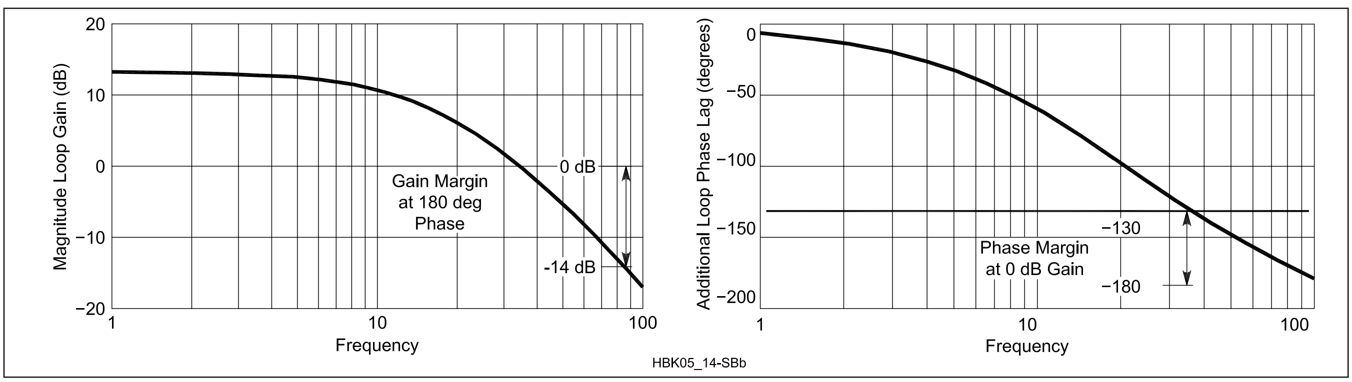


Fig B—Graph 1 shows the gain margin of the feedback loop when the loop phase shift is 180°. Graph 2 shows the phase margin of the feedback look when the loop gain is 0 dB.

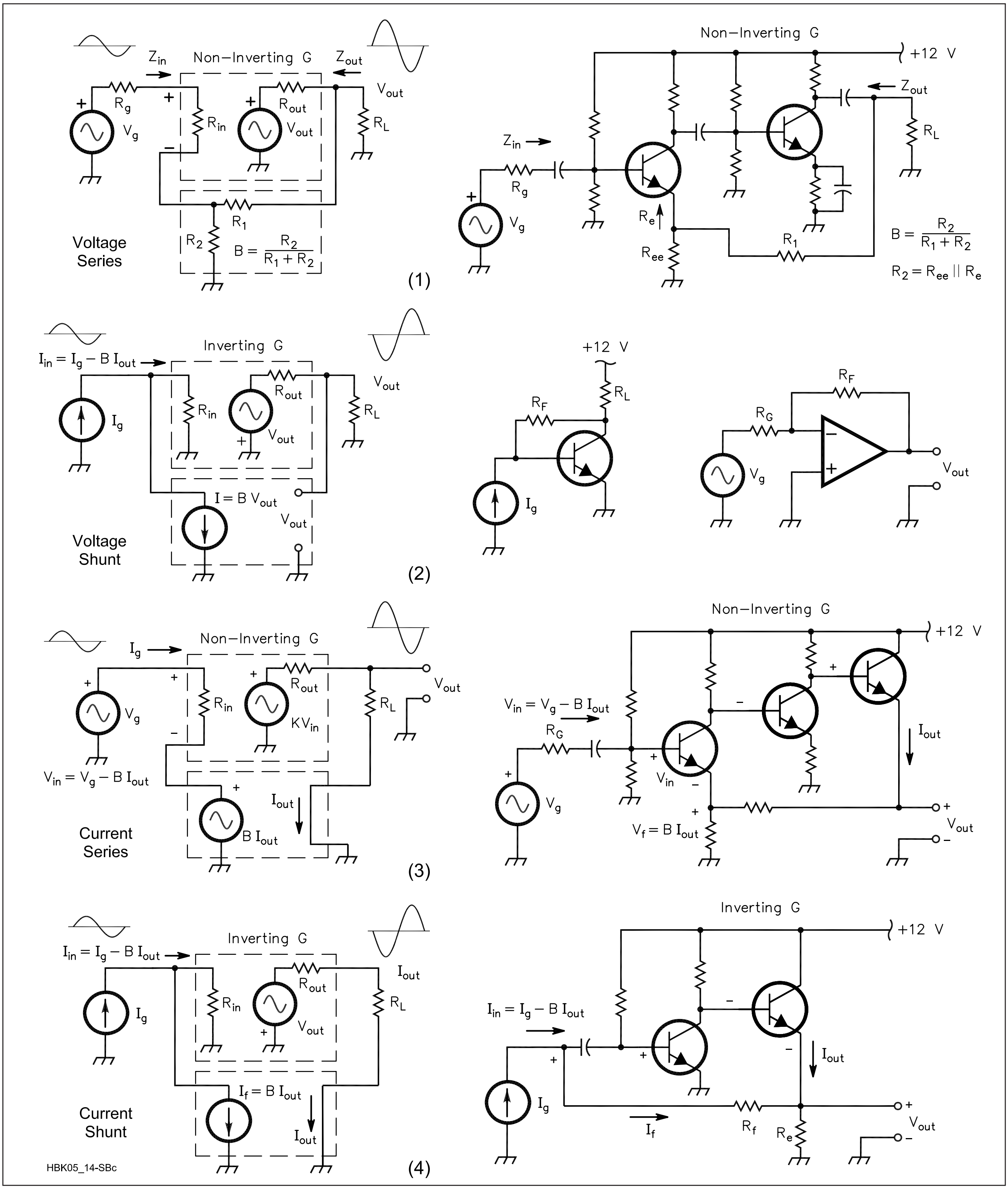


Fig C—Part 1 is an example of voltage-series feedback. A fraction, B, of the noninverted output voltage is in series with the input-signal voltage. Z_{out} is reduced and Z_{in} is increased. R_2 consists of R_{ee} in parallel with R_e (= 1/gm).
Part 2 is an example of voltage-shunt feedback. The signal generator is a constant-current source, I_g. Current B V_{out} is in parallel with I_g and in phase-opposition to I_g. The two sample circuits are a one-stage trans-resistance amplifier and an op-amp as an inverting voltage amplifier. Input impedance and output impedance are both reduced.
Part 3 is an example of current-series feedback, or a transconductance amplifier. The noninverted output current produces a feedback voltage of V_f = B I_{out}, which is in series opposition with the signal V_g. Z_{out} is increased and Z_{in} is also increased.
Part 4 is an example of current-shunt feedback. The inverted output current produces a feedback current If = B I_{out}, which is in shunt opposition with the signal I_g. Z_{out} is increased and Z_{in} is decreased.

created in the amplifier are reduced. That is, the fed-back imperfections are amplified and phase reversed, and at the output are in the correct phase that they *counteract* the amplifier's inherent imperfections. Also, we can now increase the generator level so that the output returns to the same value that it had with no feedback. The feedback amplifier is then less imperfect, but some residual imperfection always remains.

Gain and Stability

In Fig A as the product G B (known as the "loop gain") becomes large, the gain AFB with feedback is nearly 1/B where B is the feedback gain, usually of a linear, passive feedback network. In other words, the total gain AFB of the feedback amplifier is:

$$A_{FB} = \frac{G}{1 - GB} \approx \frac{1}{B} \text{ if } -GB >> +1.0 \qquad \text{(Eq A1)}$$

In an inverting amplifier (Fig A) G and AFB are negative numbers, due to the phase inversion in G. In a noninverting amplifier, G and AFB are positive numbers and the feedback network B must provide the phase inversion (the negative number). If G B in Eq A1 reaches the value +1.0, due to *additional* phase shift, the amplifier becomes unstable (the denominator becomes zero). The design of G and B must prevent this. We must make |G B|—the magnitude of G B—less than 1.0 (0 dB) at the frequency at which an additional phase shift of 180° occurs. This is called "gain margin." We must also make this additional phase shift of G B less than 180° at the frequency at which |G B| equals 1.0 (0 dB). This is called "phase margin." For small values of these margins, you can see erratic amplifier behavior and possibly oscillation on an oscilloscope and on a spectrum analyzer. In practice, a phase margin of 45° is usually a safe value that also provides a good gain margin. An important task is to perform a graphical plot, over a wide frequency range, of the magnitude and phase of the product G B, examine the margins and then modify the design as needed. **Fig B** shows typical loop gain and phase plots. The most desirable response is one that approaches unity (0 dB) loop gain at –6 dB per octave because this provides good gain and phase margins. As a follow-up, plot the frequency response of AFB (Eq A1) to verify the final result and see some of the benefits of the negative feedback. Equations very similar to Eq A1 apply to the other benefits of negative feedback that were previously mentioned. The "cost" of the feedback is that a larger generator signal is needed, or may require a pre-amplifier stage. This additional stage operates at a lower signal level and may not need as much feedback.

Four Negative Feedback Topologies

The advantages of negative feedback can be achieved in four basic ways, according to how the feedback affects the input impedance of the amplifier, which can be increased or decreased, and the output impedance of the amplifier, which can be increased or decreased. At the output, the feedback can be *derived* from the voltage across the load or the current through the load. At the input, the feedback can be *applied* in series with the generator or in parallel with the generator. These options provide the following four variations. **Fig C** shows the block diagrams for these options, and also a simple example of each type.

1. Voltage-Series. Reduces output impedance, increases input impedance.
2. Voltage-Shunt. Reduces output impedance, reduces input impedance.
3. Current-Series. Increases output impedance, increases input impedance.
4. Current-Shunt. Increases output impedance, reduces input impedance.

Feedback can be used to match impedances. For example, the 1000-Ω output resistance of an amplifier can be reduced by feedback to 50 Ω. It then correctly terminates a 50-Ω filter, transmission line or the next amplifier stage.—*William E. Sabin, WØIYH*

"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 14.4 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 14.5**. 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 14.5. 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 14.5 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

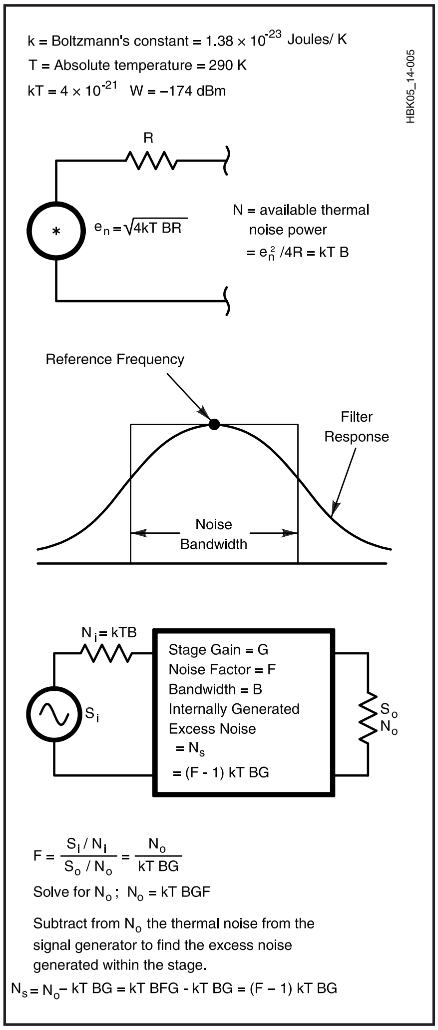


Fig 14.5—Diagram and equations that explain F, the noise factor of a single stage. The excess noise generated within the stage is also indicated. The definition of noise bandwidth is included.

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 14.5 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 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 14.5, 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 14.5. B is noise bandwidth. Using this information, the sensitivity is

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

In terms of the "open circuit voltage" from a 50-Ω 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 No is the concept of "noise temperature" which we discuss later under Microwave Receivers.

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 $Y = AX + BX^2 + CX^3 + \ldots$.... 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 14.6**).

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 14.6 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.

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 14.7** 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 14.7 illustrates this also.

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.

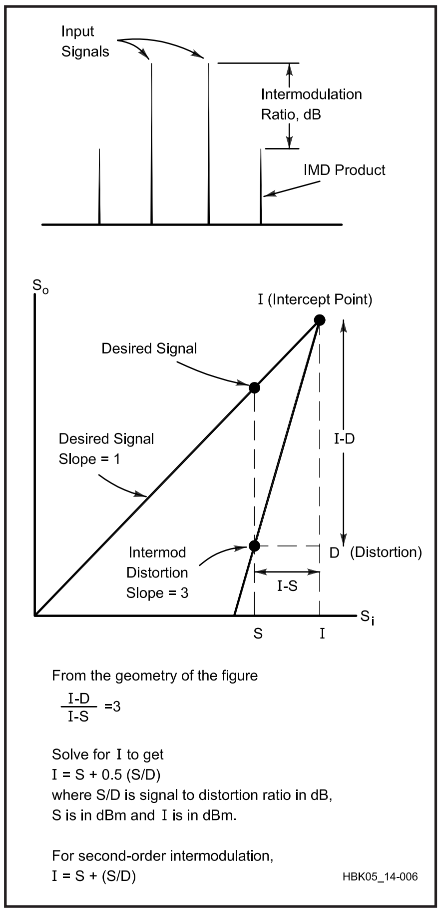


Fig 14.6—Top: IMD ratio (as displayed on a spectrum analyzer). Bottom: intercept point.

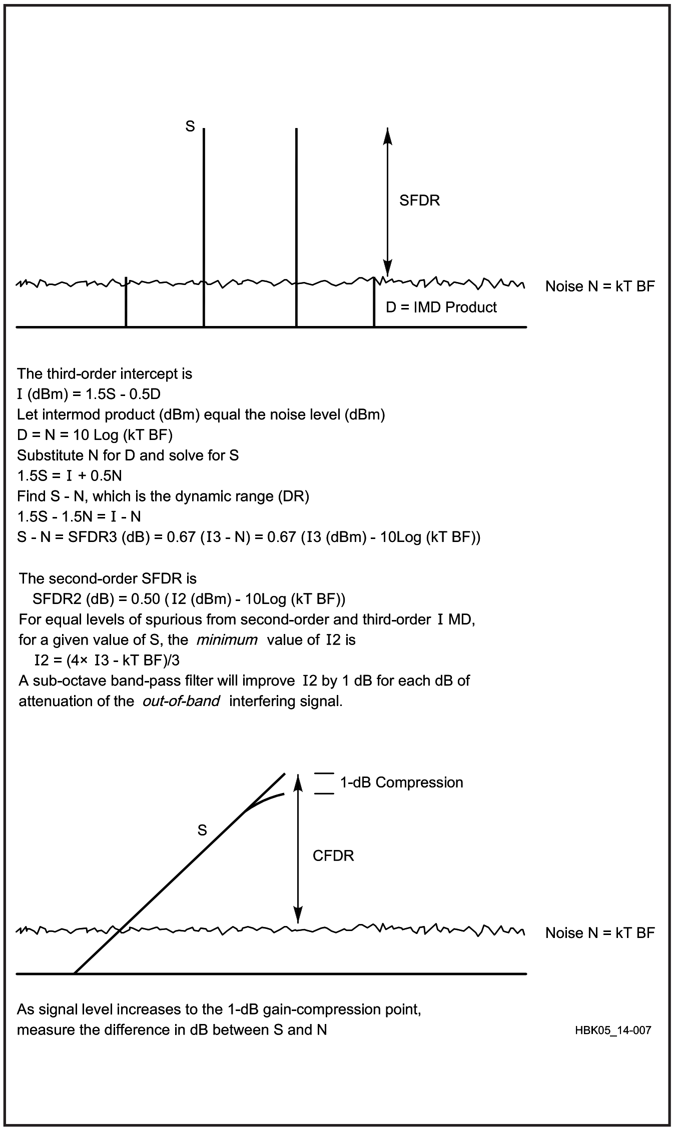


Fig 14.7—The definitions of spurious-free dynamic range (SFDR) and compression-free dynamic range (CFDR). The derivation yields a very useful equation for SFDR.

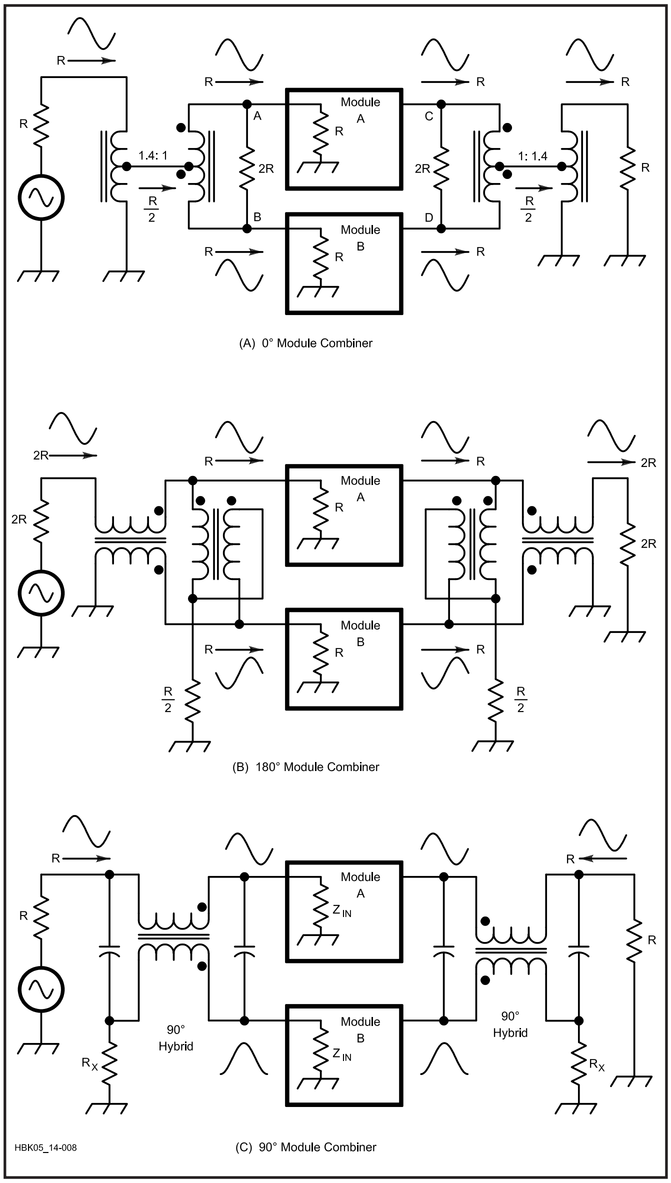


Fig 14.8—The three basic techniques for combining modules.

Fig 14.8 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.

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 (Ref 3). 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 14.9 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 L1, C1, C2 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 L2, C3, C4. 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.

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 14.9 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.

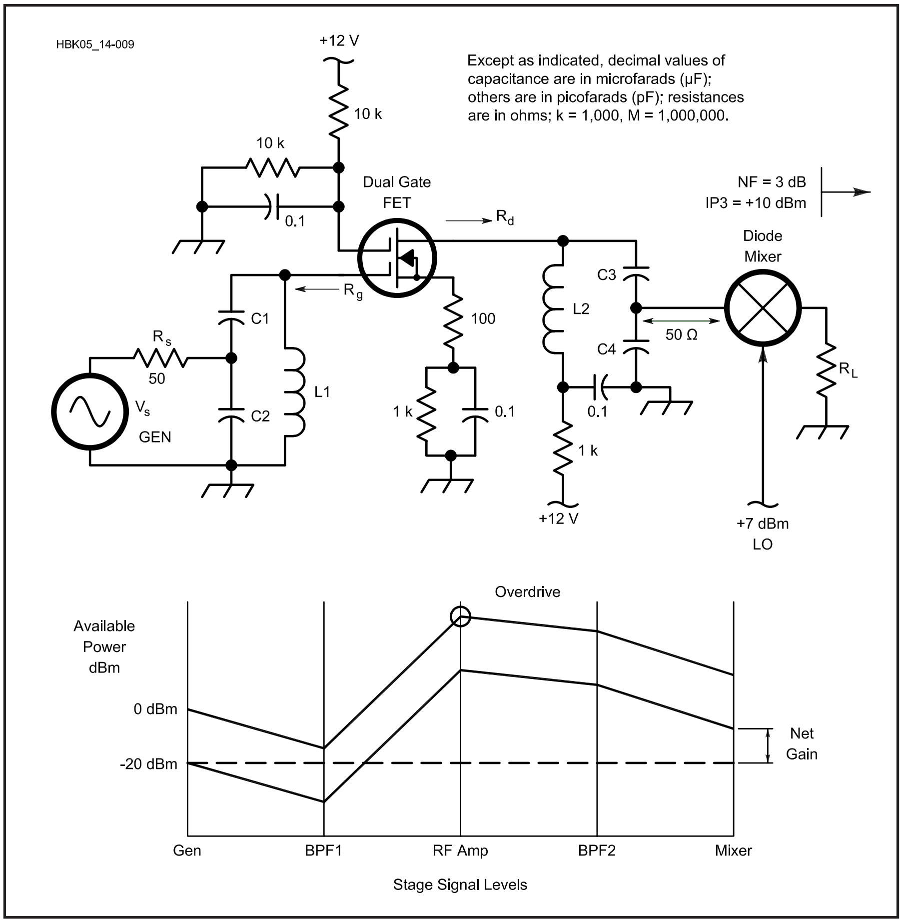


Fig 14.9—An example of cascaded stage design, a simple receiver front end.

Selectivity of Cascaded Stages

The simplified receiver example of Fig 14.9 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 14.10** 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.

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.

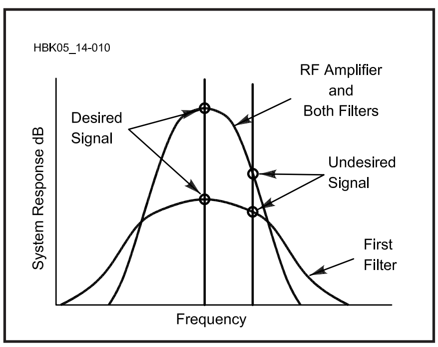


Fig 14.10—The gain and cumulative selectivity, between the generator and the mixer, of the example circuit in Fig 14.9.

Noise Factor of a Cascade

In the example of Fig 14.9, 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 14.11A**. 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 14.11A 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 multi-segment system in Fig 14.11B by applying it repetitively, first to stage N + 1 and N,

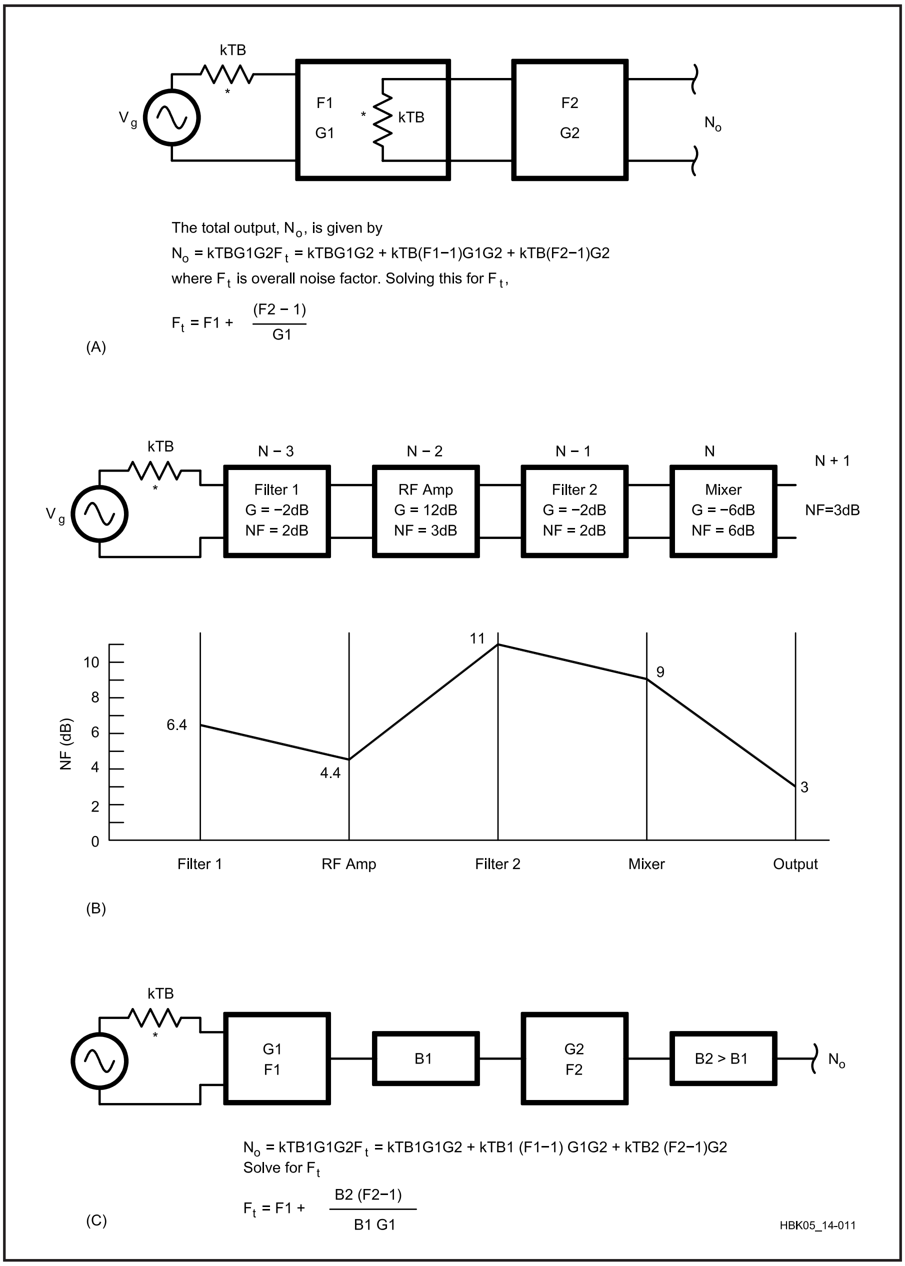


Fig 14.11—A: the noise factor of a two-stage network. B: cumulative noise factor for the example in Fig 14.10. C: noise factor when the bandwidth increases toward the output.

then to N and N – 1, 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 14.11B shows the cumulative noise figure (dB) at each point in the example of Fig 14.9. 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 14.11A 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 14.11C 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 (Ref 4).

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 14.12** 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 5).

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.

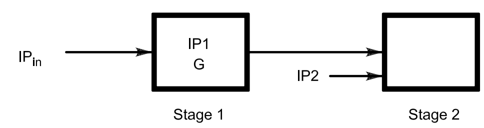
Let's briefly discuss each of these kinds of noise.

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 6).

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 7)). 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



Ip1 = Input intercept for stage 1 in watts, not dBm.
Ip2 = Input intercept looking into stage 2 in watts, not dBm.
Ip_{in} = Input intercept of the cascade in watts, not dBm.
G = Gain of stage 1 (not dB).
Convert IP_{in} from watts to dBm after calculation is finished.

For third-order intercept, $IP_{in} = \dfrac{1}{\dfrac{1}{IP1} + \dfrac{G}{IP2}}$

For second-order intercept, $IP_{in} = \dfrac{1}{\left(\dfrac{1}{\sqrt{IP1}} + \dfrac{G}{\sqrt{IP2}}\right)^2}$

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Fig 14.12—This is how IMD is calculated in cascade circuits.

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 6). 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 14.13A 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 of this topic). The general method to reduce wideband noise from the transmitter is to place the narrowband modulation 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 14.13B. 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 (Ref 8).

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 re-

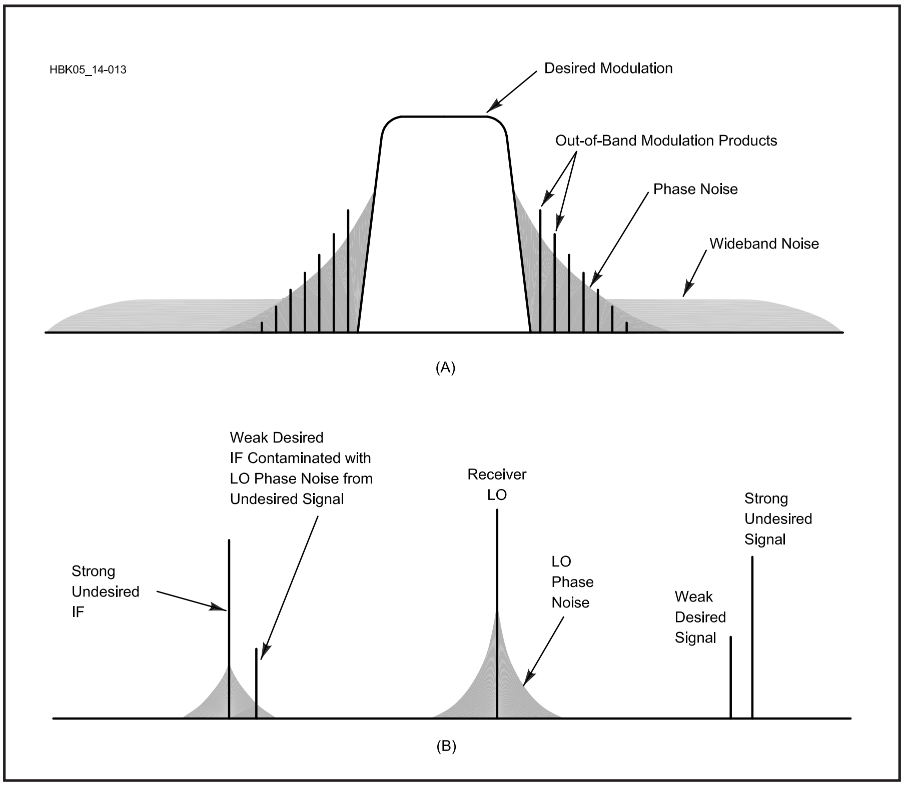


Fig 14.13—A: transmitter spectrum with discrete out-of-band products, phase noise and white noise. B: reciprocal mixing of LO phase noise onto an incoming signal.

quire 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 µV from a 50-Ω 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.

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 14.14 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

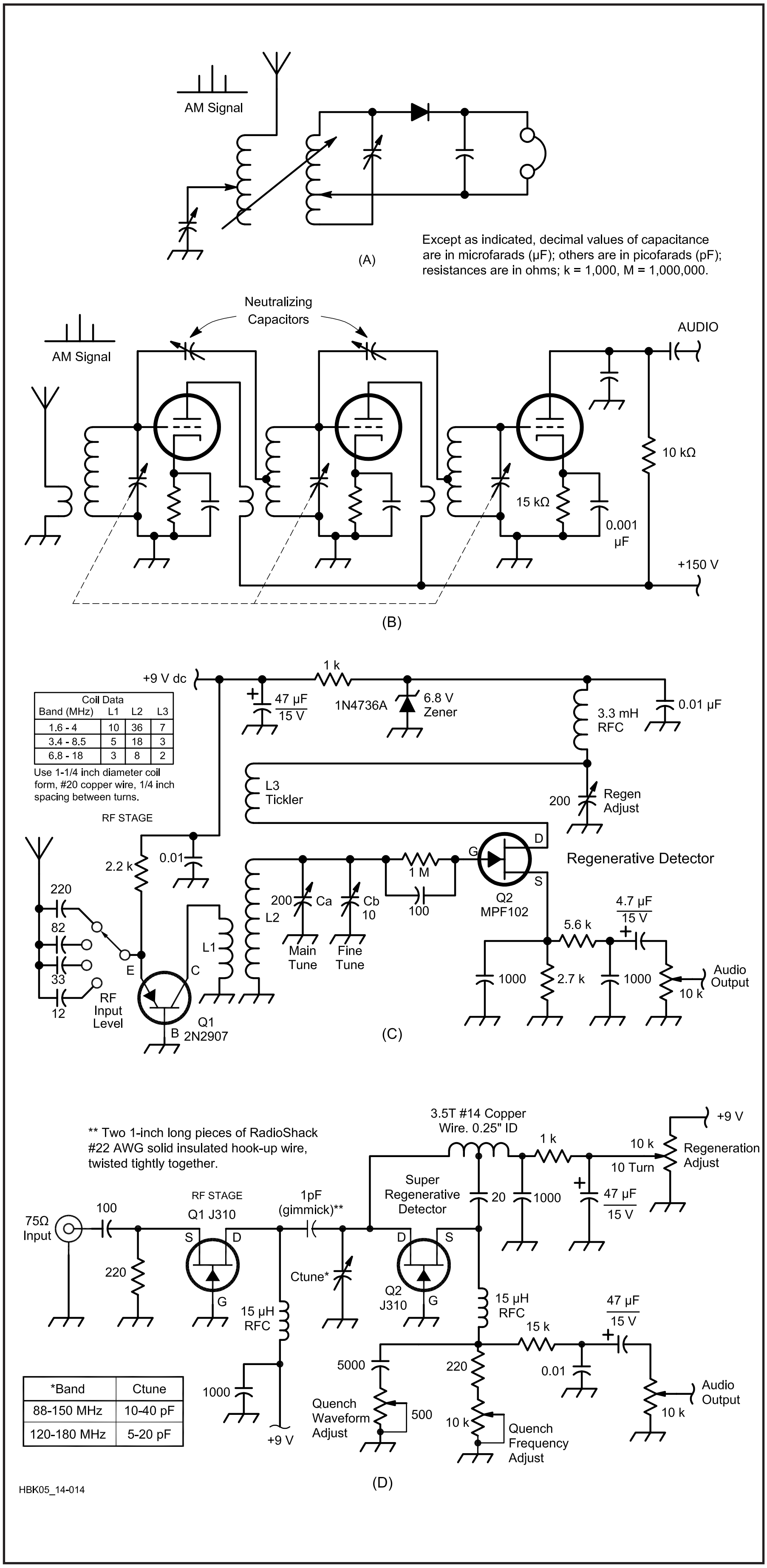
In Fig 14.14A the antenna circuit (capacitive at low frequencies) is series resonated by the primary coil to maximize the current through both, which also maximizes the voltage across the secondary. The semiconductor crystal rectifier then demodulates the AM signal. This demodulation process utilizes 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.

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 (non-coherently). Therefore the detector provides the same signal-to-noise ratio as a single sideband (SSB) signal (Ref 9).

The Tuned Radio Frequency (TRF) Receiver

As its name implies, the TRF receiver uses one or more tuned RF stages followed by a detector stage and audio amplifier. The variable capacitors (or sometimes variable coils) of each tuned stage track each other as the receiver is tuned. Each LC tuned circuit provides an additional band pass filter and so increases the overall selectivity of the receiver. One impor-



tant benefit of the TRF approach is very high quality audio when receiving AM signals. The selectivity varies with tuning however and is reduced at the high frequency end of the tuning range, where the capacitance is low and the L/C ratio is high. At the low end of the tuning range, selectivity can become too high and roll-off the modulation sidebands. This variation in selectivity (and gain) are the TRF's main drawbacks. A related type of receiver, the "Neutrodyne" was an early triode design that used neutralization capacitors to prevent its RF stages from oscillating. See Fig 14.14B. Multigrid tubes or dual-gate FETs usually do not need neutralization.

The Regenerative Receiver

Edwin Howard Armstrong invented the regenerative circuit around 1914. Fig 14.14C shows a modern version. Q2 is basically a modified JFET oscillator that uses positive feedback (termed "regeneration"), to greatly increase both gain and selectivity.

A portion of the detector's amplified RF output is fed back to its input, in phase, by tickler winding L3. The signal is then amplified over and over, providing a gain of about 20,000 (86 dB). To minimize drift, this circuit uses a regulated detector supply voltage and a "throttle" capacitor regeneration control. Regeneration also introduces negative resistance into the circuit, greatly increasing its selectivity. This avoids the necessity of using several tuned stages. As regeneration is increased above the oscillation point, both gain and selectivity go down.

The regenerative receiver is easy for beginners to "home-brew" and different coils (or capacitors) can simply be switched-in to provide a very wide tuning range. Like the TRF architecture, the regen also tends to provide hi-fi, low distortion audio. The tradeoff, however, is that regeneration must be user adjusted and this requires both practice and patience. When receiving AM signals, feedback is set to a point just below self-oscillation. For CW

Fig 14.14—A shows a simple crystal set receiver that is as much fun to build and use today as it was in the early days of radio. B is an example of a "Neutrodyne" receiver, which was a variation of the tuned radio-frequency (TRF) receiver. C is an example of a modern regenerative receiver circuit, using transistors as the active elements. D is a modern self-quenched superregenerative receiver circuit.

and SSB reception, the detector is adjusted so that it is operating above the oscillation threshold, thus providing a beat note.

An RF stage (Q1) is used to isolate the 1 mW detector from the antenna and for additional gain. An input attenuator allows the operator to reduce the RF input level, increasing selectivity and preventing the detector from "blocking" on strong CW or SSB transmissions. For more details see: Kitchin, "High Performance Regenerative Receiver Design," November/December 1998 *QEX*.

The Superregenerative Receiver

Invented by Edwin Howard Armstrong around 1922, the superregenerative circuit is an oscillating regenerative detector that is periodically shut down or "quenched" by a second oscillation, usually between 20 and 30 kHz. It is essentially an amplitude-modulated oscillator whose quenching oscillations allow the input signal to build-up to the oscillation threshold repeatedly, providing typical detector gains of one million (120 dB). Superregenerative detectors can employ a separate quench oscillator (separately quenched) or produce their own secondary relaxation oscillations (self quenched).

The "superregen" receiver can be used from the lower VHF range all the way up into the microwave region and provides a simple receiver of great sensitivity. Fig 14.14D shows a modern circuit design for Amateur Radio experimentation.

Although much easier for a home builder to construct than a VHF superhet, the superregenerative receiver's extremely high gain is often difficult to control, requiring careful design and construction. The superregen also has a high nonsignal background noise; however, a simple squelch circuit will cure this. If high selectivity is needed, several controls must be carefully adjusted while the receiver is tuned. But for wide-band AM and FM reception, the controls can be preset. A superregen is sometimes operated in the "linear mode" for low audio distortion, although standard operation is sufficient for communications quality audio.

The classic Amateur Radio article about superregenerative receivers is Ross Hull's July 1931 *QST* article, "Five Meter Receiver Progress." This article describes a successful superregnerative receiver for 56 to 60 MHz.

Nat Bradley, ZL3VN, has discovered a new use for the superregen. If a separate local oscillation is mixed with the received signal, so that the frequency difference between the two is equal to the quench frequency, the detector will provide direct narrow-band FM demodulation. For more details, see: Kitchin, "New Super Regenerative Circuits for Amateur VHF and UHF Experimentation," September/October 2000 *QEX*.

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.

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 14.15 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 14.16** 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.

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.

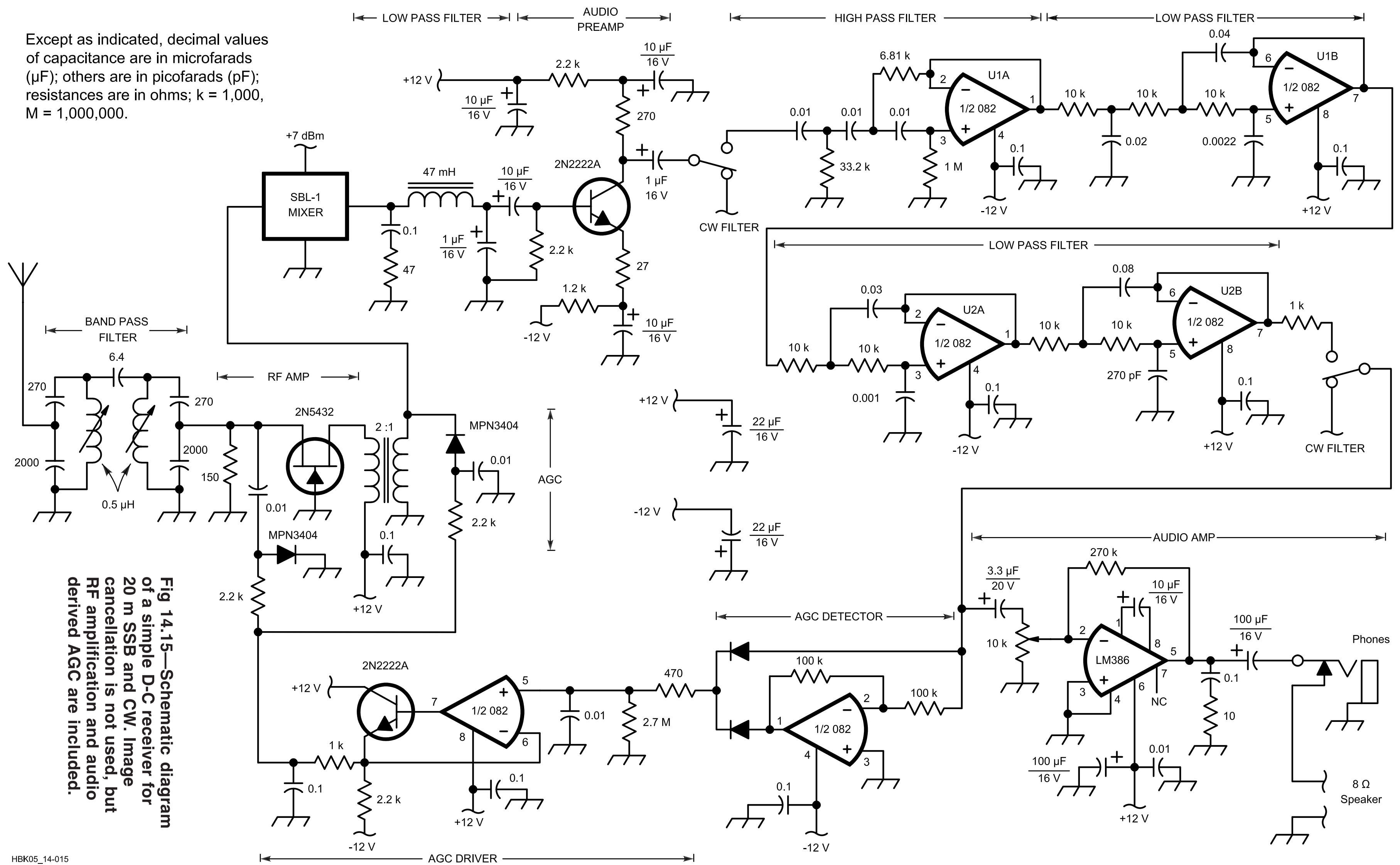


Fig 14.15—Schematic diagram of a simple D-C receiver for 20 m SSB and CW. Image cancellation is not used, but RF amplification and audio derived AGC are included.

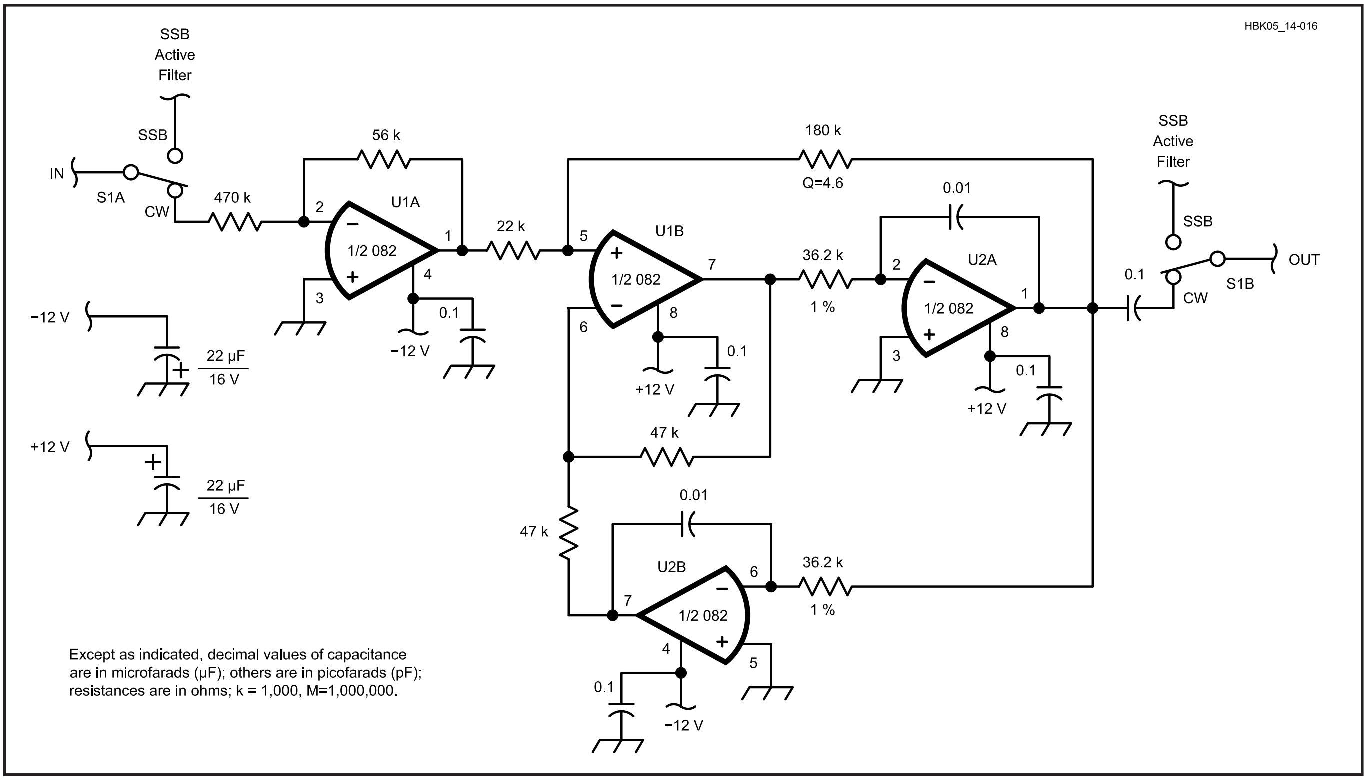


Fig 14.16—Schematic diagram of an active CW band-pass filter for the D-C receiver in Fig 14.15.

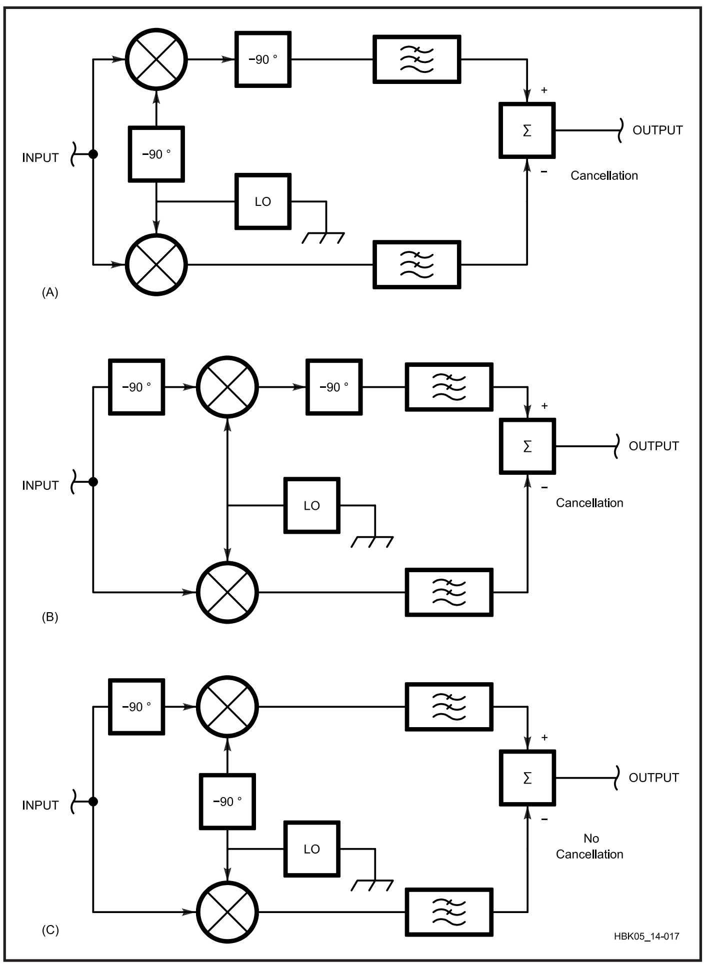


Fig 14.17—Both A and B are workable image cancelling mixer stages. The scheme at C will not cancel image signals.

The 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 14.15 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

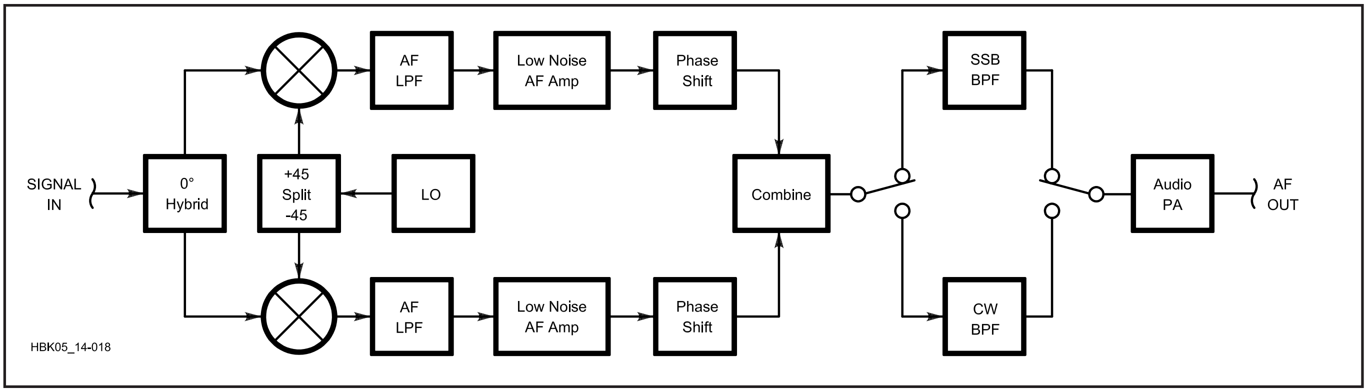


Fig 14.18—Typical block diagram of an image cancelling D-C receiver.

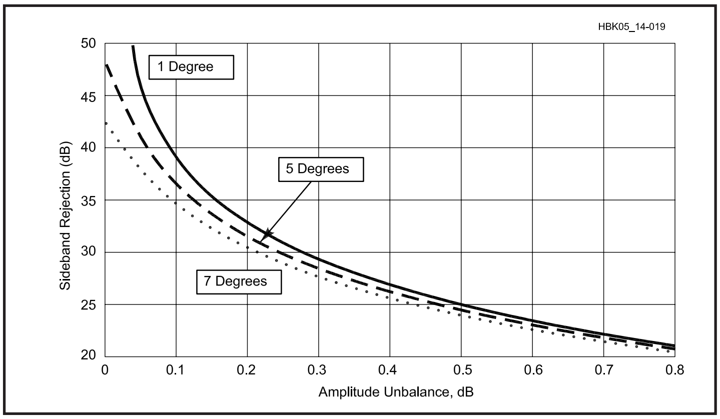


Fig 14.19—A plot of sideband rejection versus phase error and amplitude unbalance.

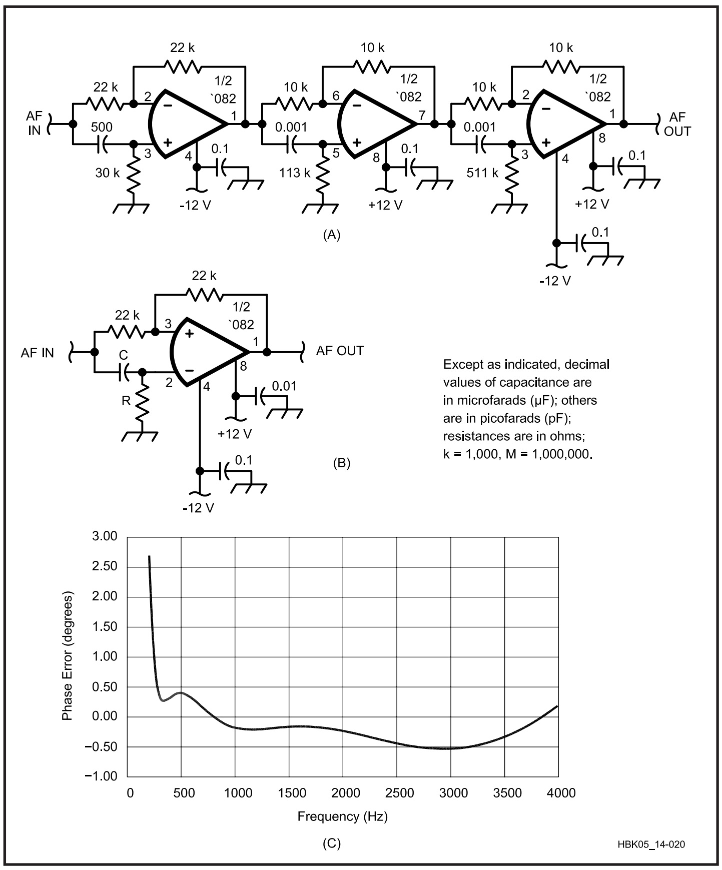


Fig 14.20—A: an example of an audio phase-shift circuit. B: a single all-pass stage. C: phase error vs audio frequency for a pair of circuits like that in A with appro-priate values of R and C.

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 11, 12). **Fig 14.17** 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 14.18** is a typical approach 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 14.19** 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.

AUDIO PHASE SHIFTERS

Fig 14.20A shows an example of an audio phase-shift network. The stage in Fig 14.20B 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/(2πRC). 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 13). Fig 14.20C shows the phase error for two circuits like the one shown in Fig 14.20A. 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 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 14.21A** 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 a perfect multiplier, as the equation in Fig 14.21 suggests, it is a linear mixer, these are the only output frequencies present and it has all the properties of any other linear circuit except for the change of frequency. If the mixer is a commutating (switching) mode mixer it is still a perfect mixer but additional frequencies of lesser amplitude are present. See the **Mixers, Modulators and Demodulators** chapter for a detailed discussion.

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

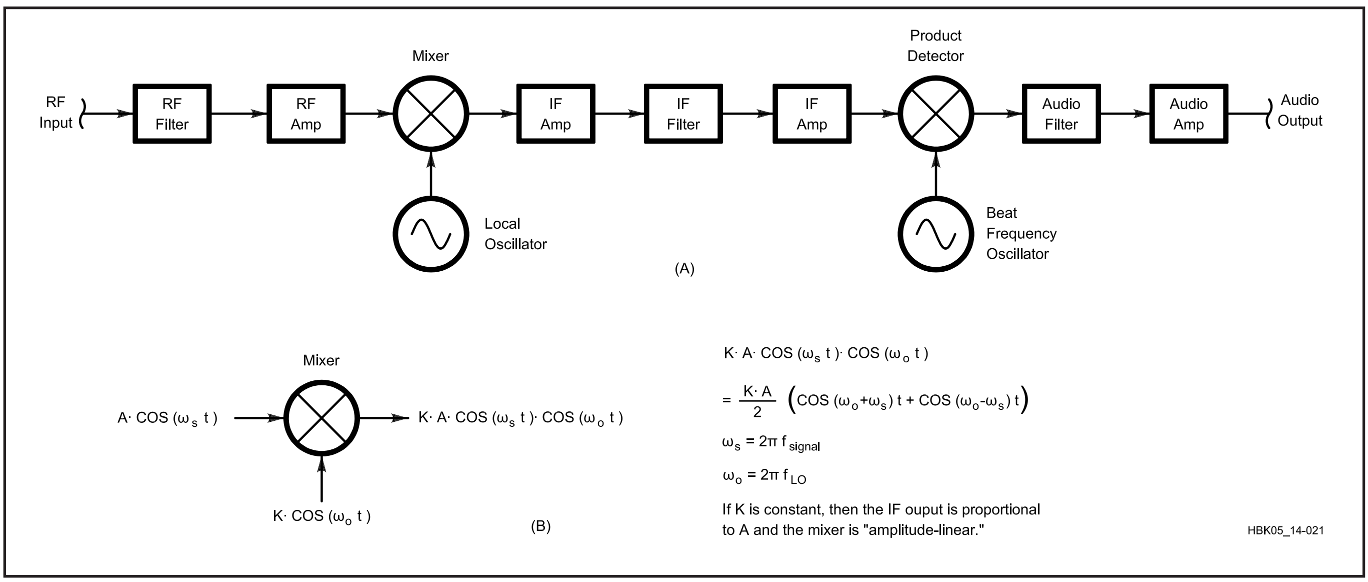


Fig 14.21—(A) Basic block diagram of a superhet receiver. (B) Showing how the input signal and a constant LO input produce a linear mixing action.

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.

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 14.22**. 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 14.22 is intentionally incomplete and meant for instructional purposes only; do not attempt to duplicate it as a project.

Block Diagram

The block labels of Fig 14.22 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 14.15 and Fig 14.16. 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 14).

Fig 14.23A 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 14.22B; 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 the response at 14.00 MHz.

Fig 14.23C 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 selectivity 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. 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 15). For amateur work the approach in Fig 14.22 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

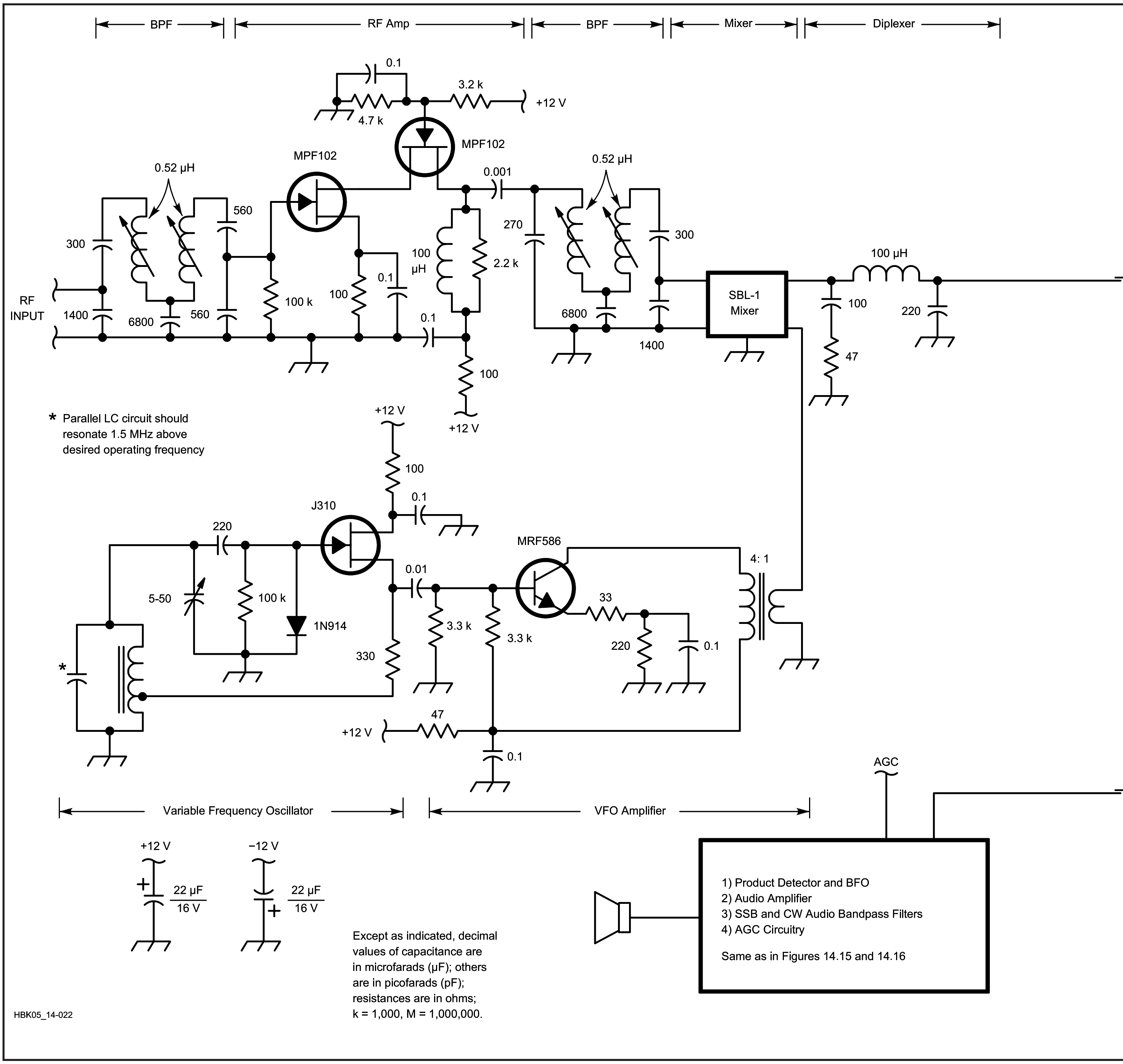
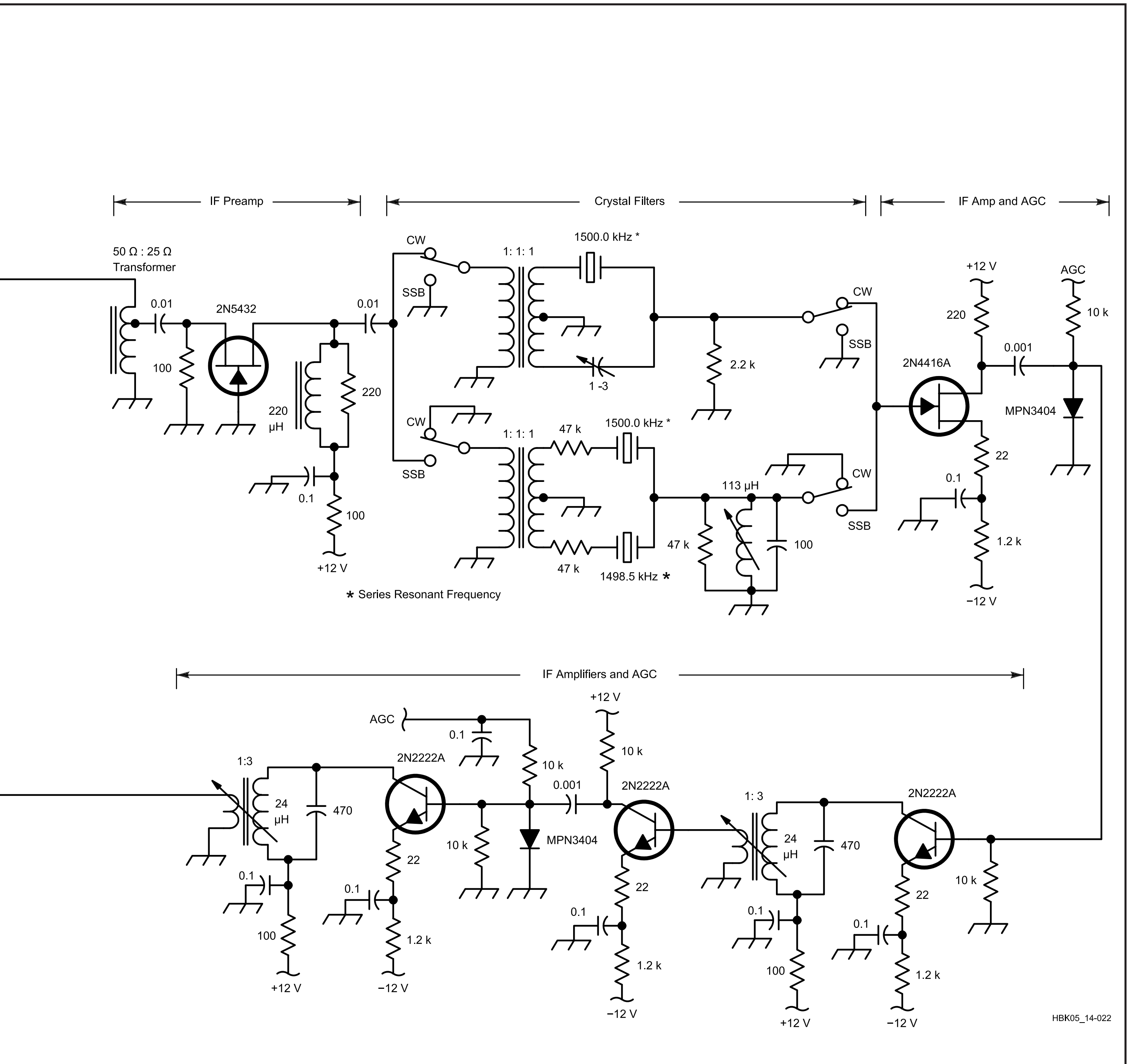


Fig 14.22—Specific example of a down-conversion superhet that is used to explain and analyze superhet behavior and design.

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 (Ref 16).

A matter of considerable interest concerns the way that IMD varies as the separation between the two tones increases. In Fig 14.22, 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



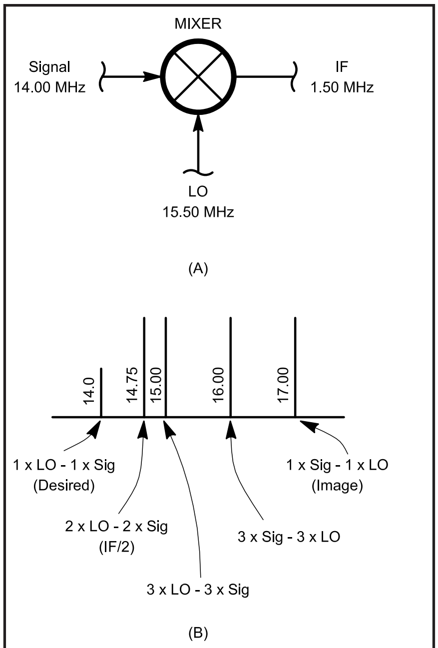
for ways to improve the overall performance.

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 14.22 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



RF Level = 0 dBm

RF Harmonic	LO Harmonic 1	2	3	4	5
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

HBK05_14-023 (C)

Fig 14.23—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.

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 14.22).
6. Terminate the mixer in such a way that its IMD is minimized. Fig 14.22 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 14.22) 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.

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 17).

The AGC Loop

Fig 14.24A 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 14.24B is achieved. The AGC action does not begin until a certain level, called the AGC threshold, is reached. The Threshold Volts input in Fig 14.24A 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 14.25 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.

AGC Time Constants

In Fig 14.24, 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 18). If we make R2 × C1 much longer, say 3 seconds or more, the AGC voltage remains almost constant until the R5, C2 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 lin-

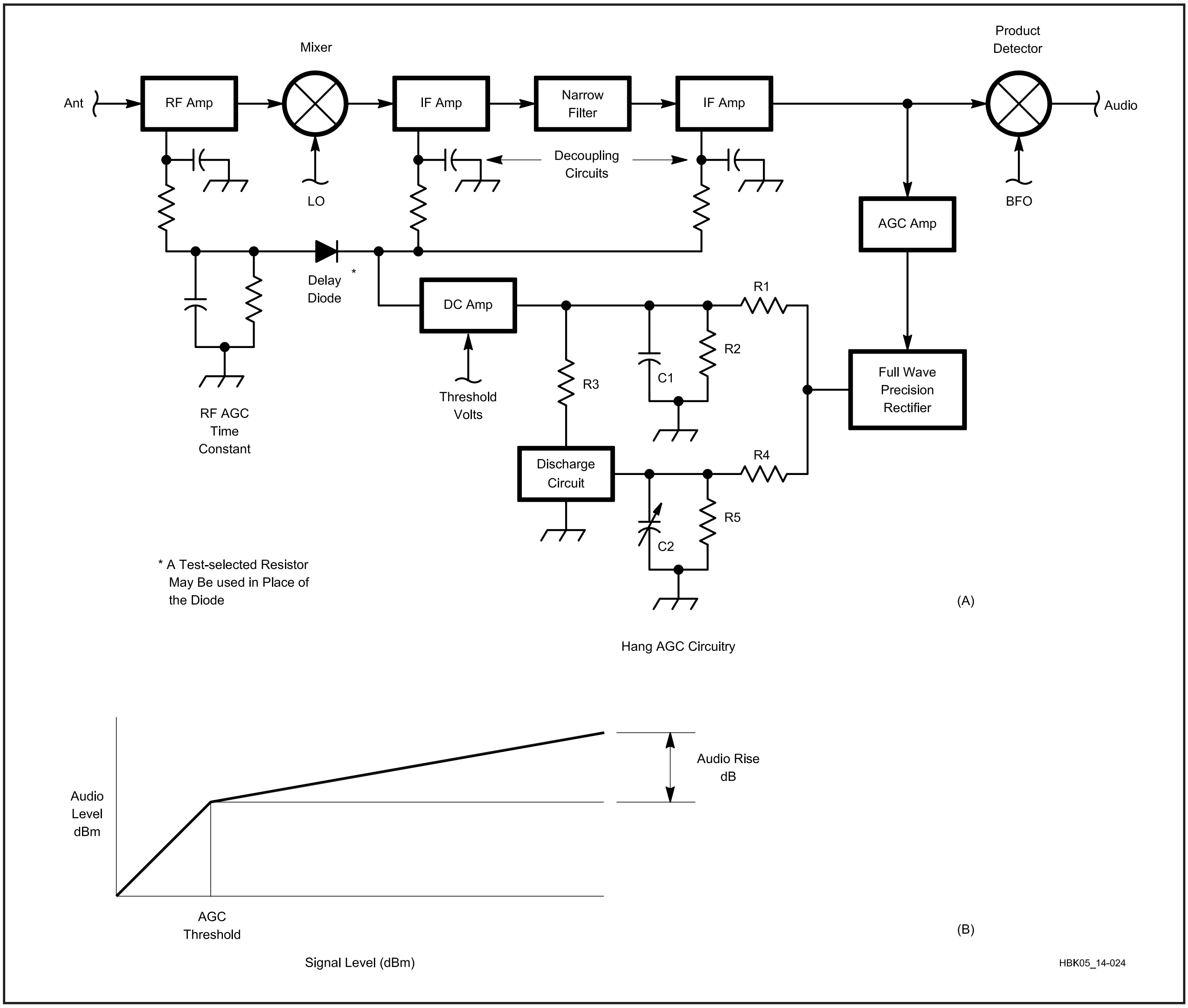


Fig 14.24—AGC principles. A: typical superhet receiver with AGC applied to multiple stages of RF and IF. B: audio output as a function of antenna signal level.

ear" 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 14.24A 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 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.

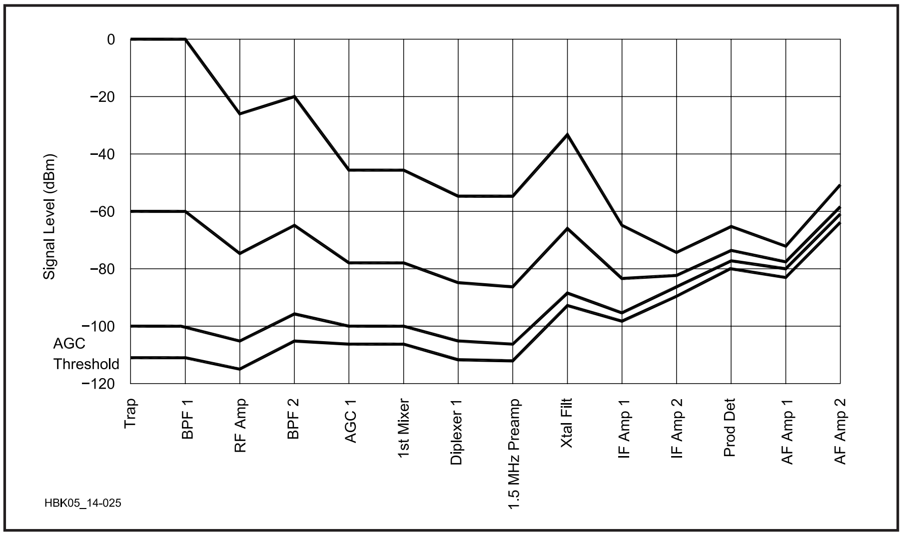


Fig 14.25—Gain control of a ham-band receiver using AGC. A manual gain control could produce the same result.

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 14.15 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 14.26 shows some gain controllable circuits. Fig 14.26A shows 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).

Fig 14.26B 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 14.26C 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 14.26D shows the high-end performance National Semiconductor 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 19). 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

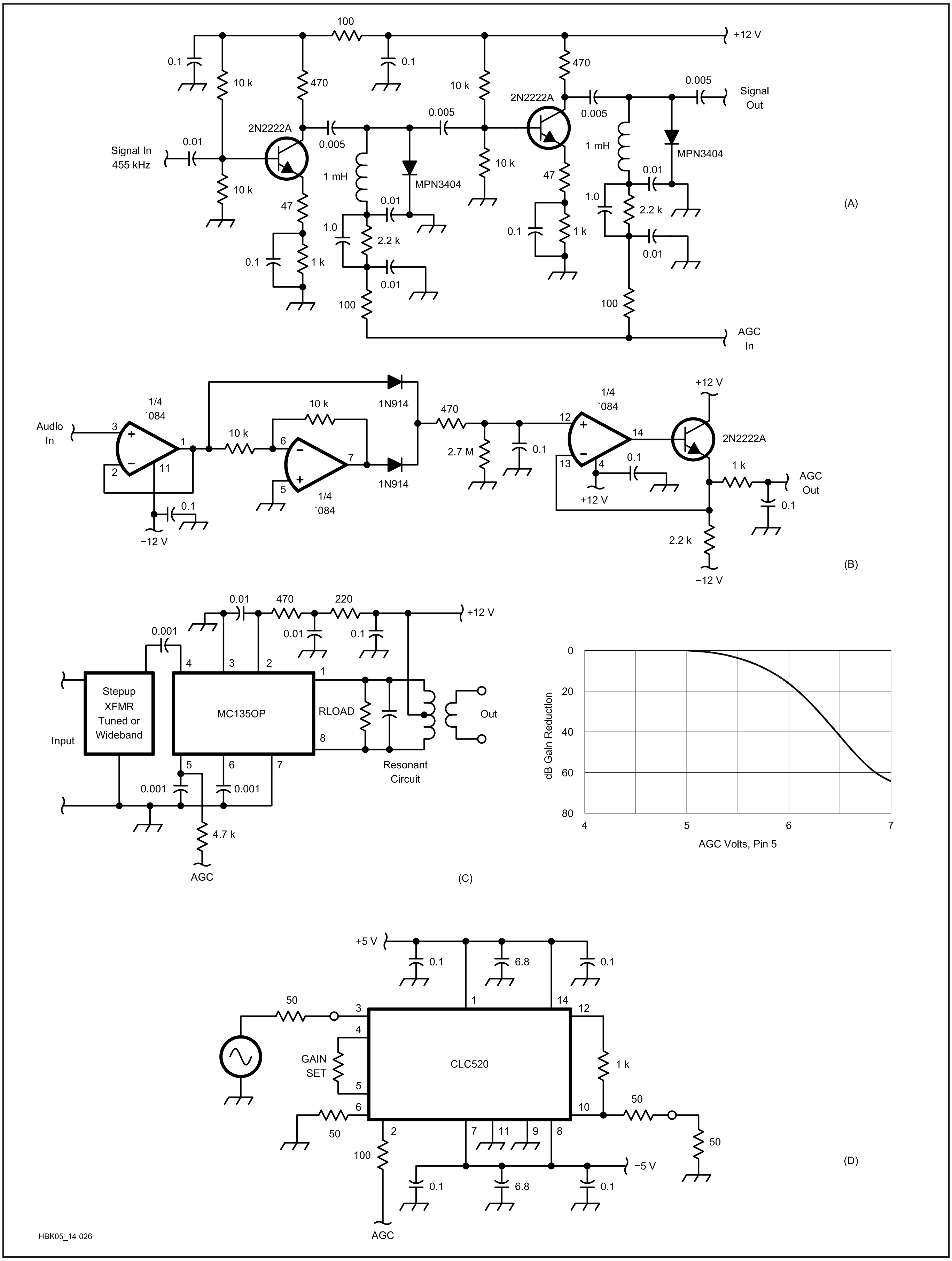


Fig 14.26—Some gain controllable amplifiers and a rectifier suitable for audio derived AGC.

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 14.15 and Fig 14.16) or digital (DSP) technology.

Some Simple Crystal Filters

Fig 14.27 and **Fig 14.28** present two crystal filters to consider for a simple down-conversion receiver with a 1.5-MHz IF (see Fig 14.23). The crystals are a set of three available from JAN Crystals. The filters are both driven from a low-impedance source (200 Ω, for example).

CW Filter

Fig 14.27A is a "semi-lattice" filter using a single crystal for CW work (Ref 19). 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 Fig 14.27B. If this filter is combined with an audio band-pass filter as in Fig 14.16, pretty good CW selectivity is possible. In Fig 14.27C, 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.

SSB Filter

Fig 14.28 is a "half-lattice" filter (Ref 19). 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 14.15 will improve the overall response considerably.

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 "Low Cost Series" of miniaturized torsional-mode filters for 455.0 kHz. They come in four styles with 3 dB/60 dB bandwidths of 0.3/0.5/1.5/2.0, 2.5/5.2 and 5.5/11 kHz. Used filters are sometimes available from various sources (Ref 15).

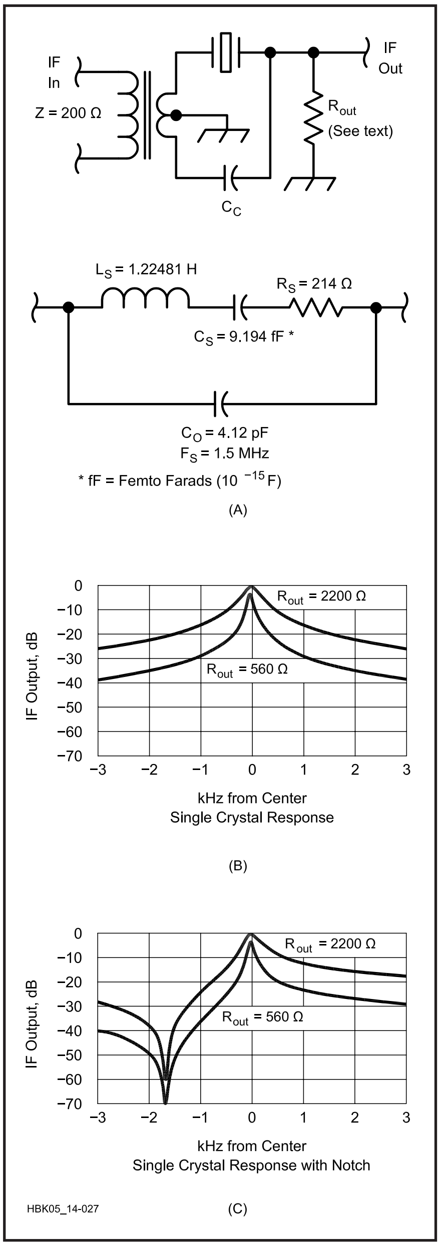


Fig 14.27—A single-crystal filter circuit for a simple CW receiver design. See also Fig 14.22.

Multielement Crystal Filters

A discussion of more complex crystal filters appears in the **RF and AF Filters** 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). **Fig 14.29** shows examples of relay and diode switching that work quite well. The relays can be inexpensive miniature RadioShack 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 15).

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

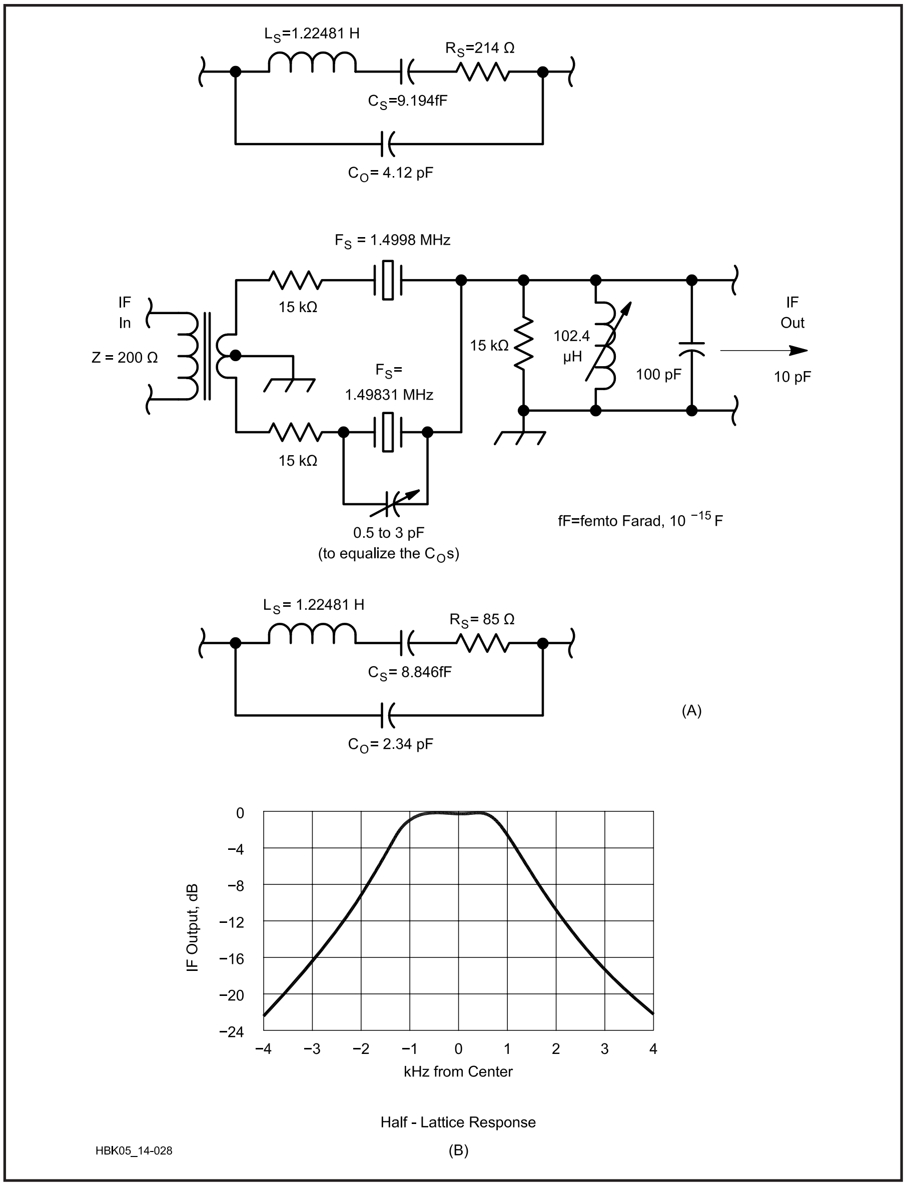


Fig 14.28—A two-crystal half-lattice filter for a simple SSB receiver. See also Fig 14.22.

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 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 (Ref 20).

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 (Ref 21).

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 14.30**. 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 band-switched. 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 14.30, 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.

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 22).

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 pre-

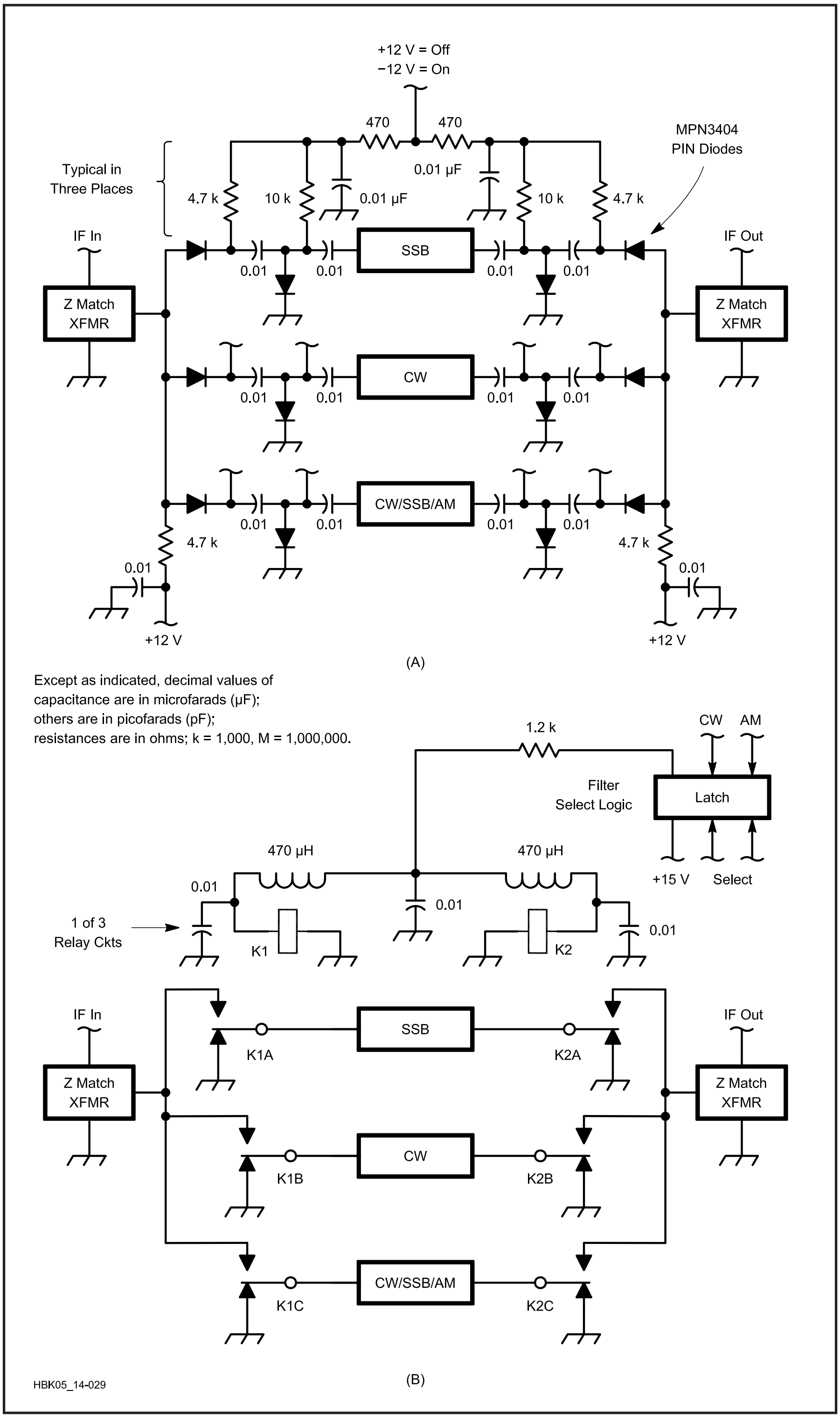


Fig 14.29—IF filter switching using PIN diodes (A) or relays (B).

ceding 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 23).

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.

An Up Converter Example

The block diagram in **Fig 14.31** 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 pre-selection 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.

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 speci-

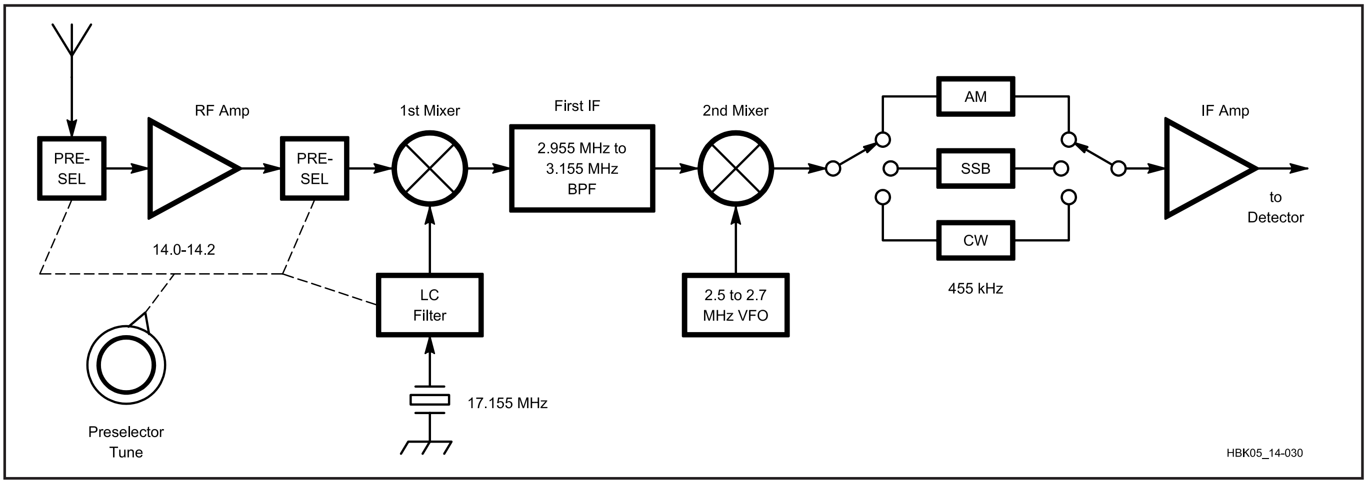


Fig 14.30—Block diagram of a double-conversion superhet that eliminates some of the tracking problems of the conventional superhet.

fications. 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.

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 14.31 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 14.31 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 feed-through, 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 14.31 shows.

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 14.31, 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.

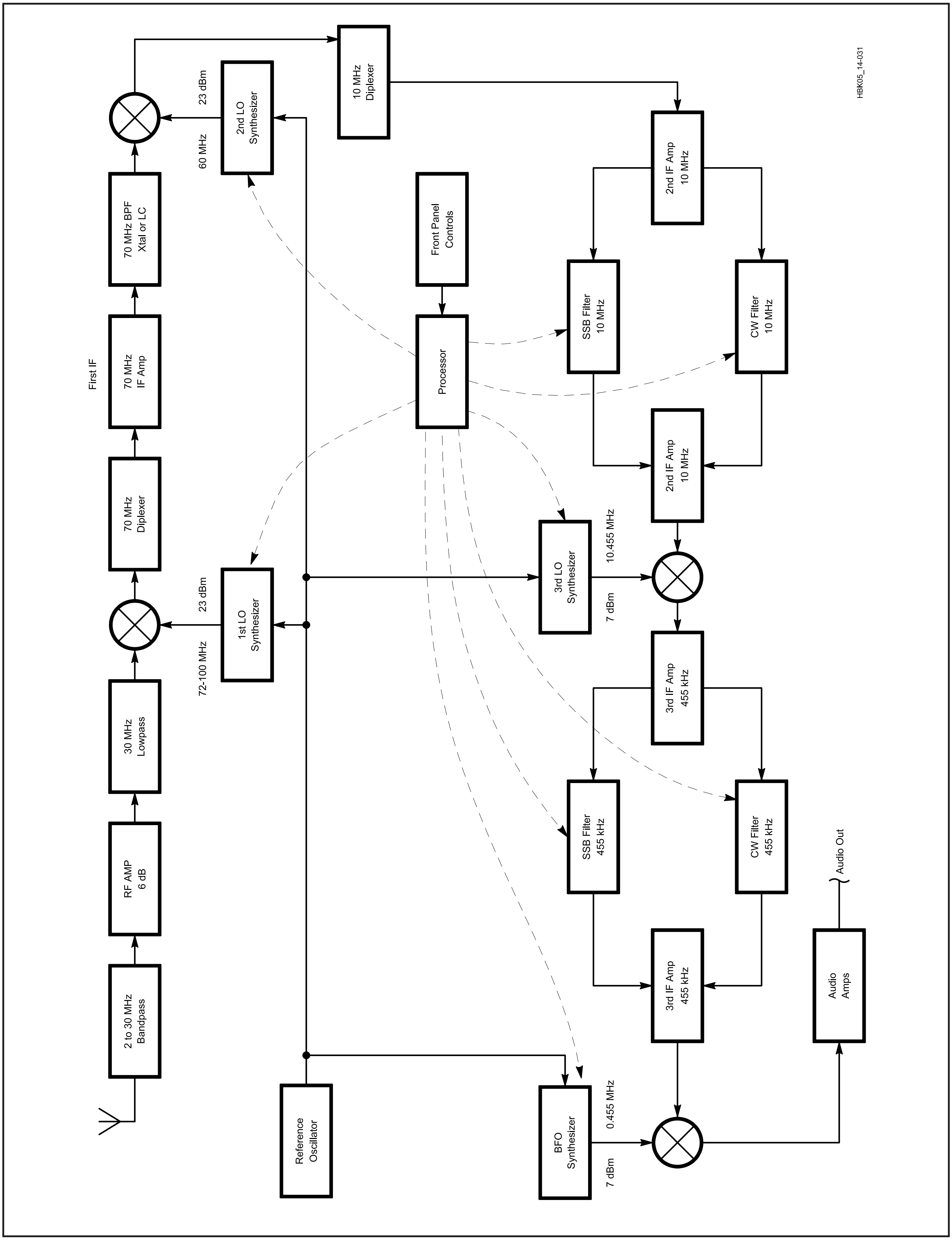


Fig 14.31—Block diagram of an HF up-conversion receiver for SSB and CW. Microprocessor control of receiver functions is included. LOs are from synthesizers.

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 14.31 is an acceptable compromise in this respect for a fairly high quality receiver.

Triple Conversion

The block diagram of Fig 14.31 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 narrowband 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 24).

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 14.31, 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.

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** 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** chapter describes how a noise blanker is typically implemented. (See also Ref 17.)

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. Yes, hams do use SSB for longer distance communications because of its better weak-signal performance. However, most voice communication in this range is done using FM. This section will focus on the differences between VHF/UHF and HF receivers.

NARROWBAND FM (NBFM) RECEIVERS

Fig 14.32A 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 14.32B 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 14.32A shows dual down-conversion.

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 pre-emphasis of the higher speech frequencies occurs at the transmitter and de-emphasis 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 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** 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

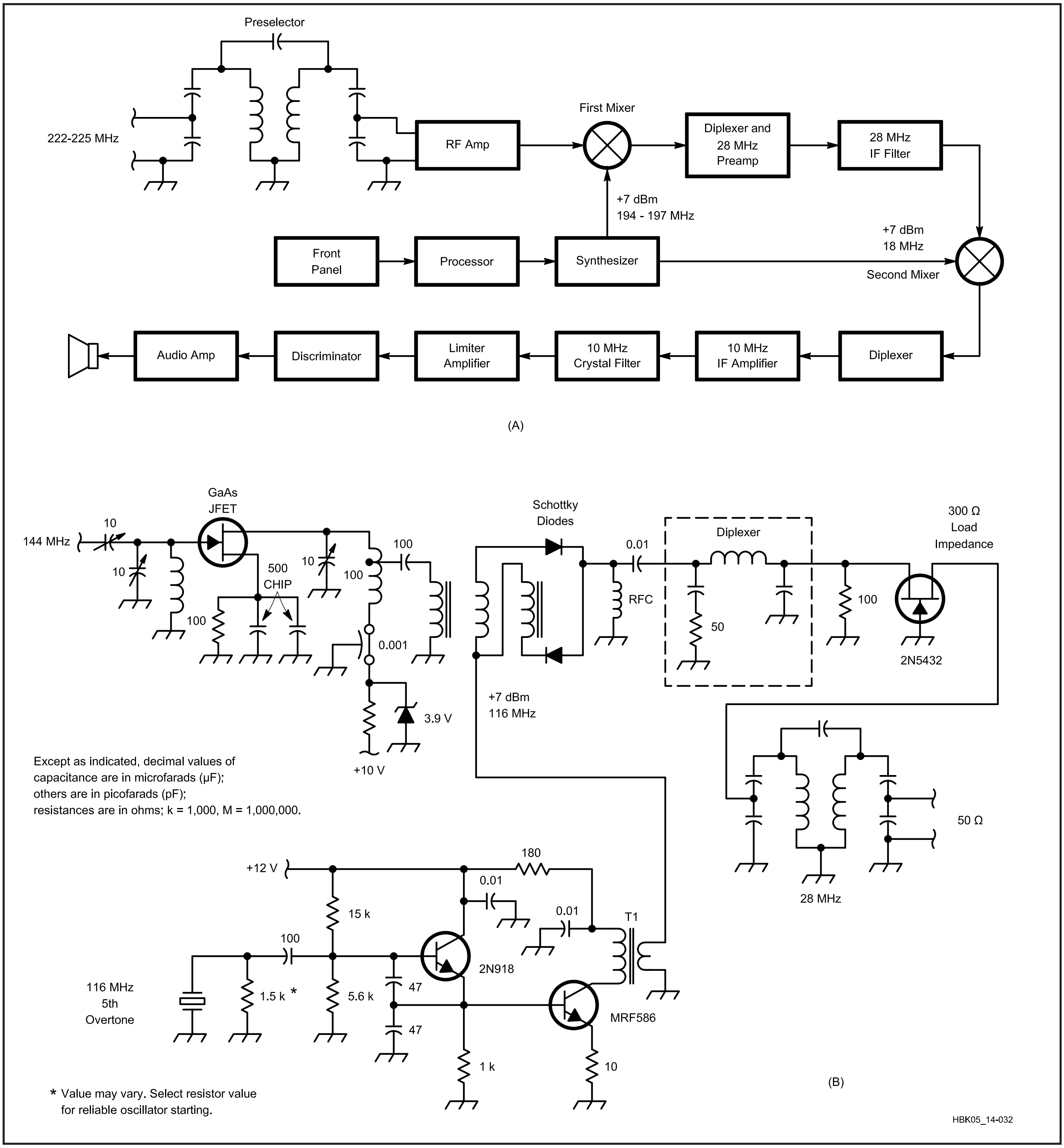


Fig 14.32—Some narrowband FM circuits. A: block diagram of a typical NBFM receiver. B: a front-end circuit with preselection and down conversion.

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 8). A similar ultimate SNR effect occurs in SSB receivers. On the other hand, a perfect AM receiver tends to suppress LO phase noise (Ref 9).

NBFM ICs

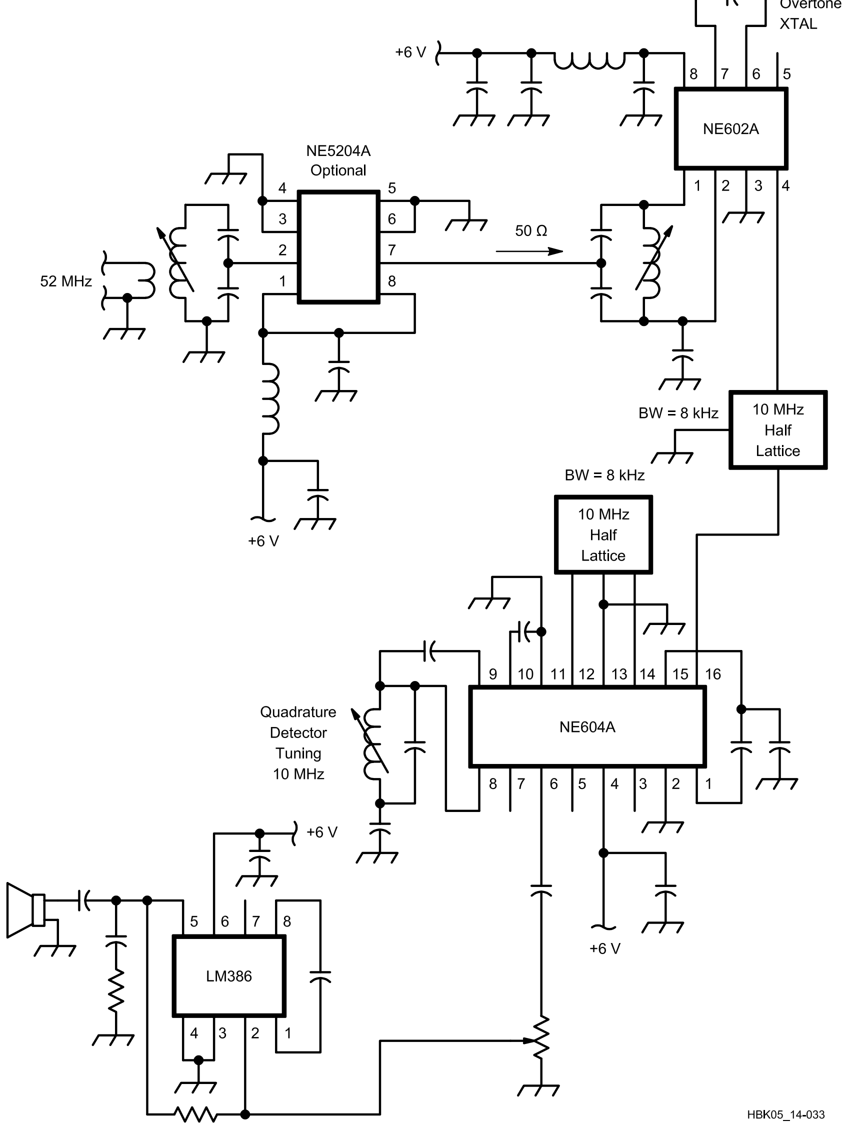
A wide variety of special ICs for NBFM receivers are available. Many of these were designed for "cordless" or cellular telephone applications and are widely used. **Fig 14.33** shows some popular versions for a 50-MHz NBFM receiver. One is an RF amplifier chip (NE/SA5204A) for 50-Ω input to 50-Ω output 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 25) 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. 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 down-conversion was mentioned previously in this chapter.

The diagram in Fig 14.33 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 designers should learn how to use these data sheets and other information.

Fig 14.33—The NE/SA5204A, NE/SA602A, NE/SA604A NBFM ICs and the LM386 audio amplifier in a typical amateur application for 50 MHz.



The best places to learn about data sheets are data books and application notes.

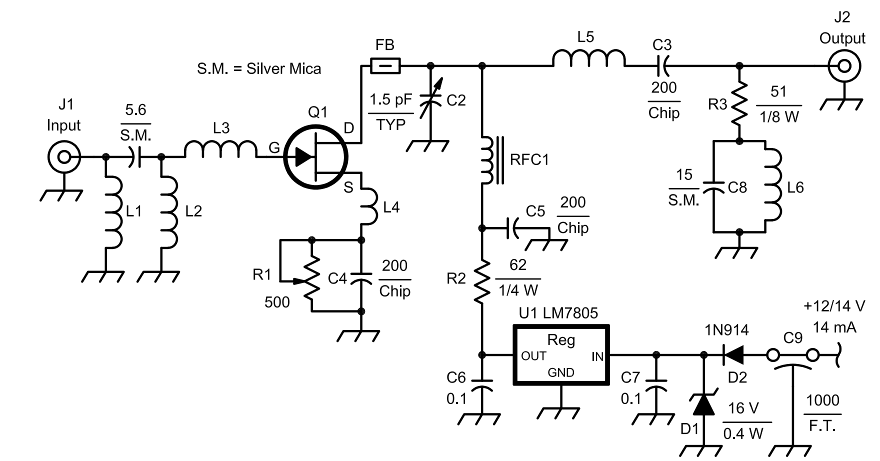
UHF TECHNIQUES

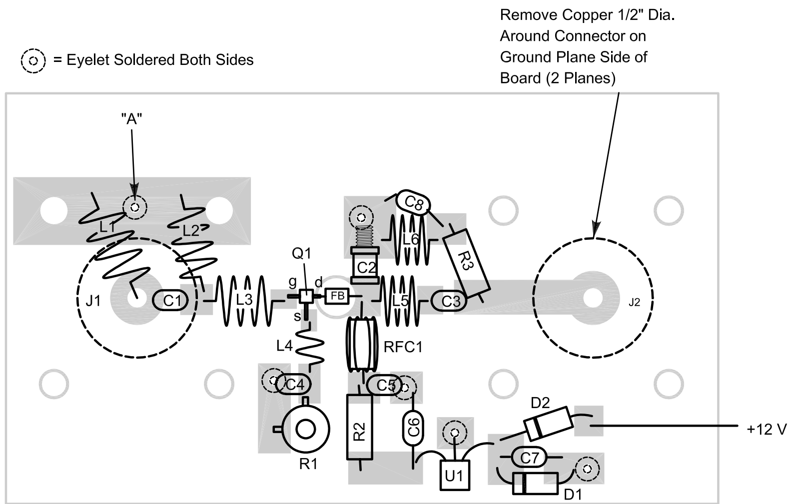
The Ultra High Frequency spectrum comprises the range from 300 MHz to 3 GHz. All of the basic principles of radio system design and circuit design that have been discussed so far apply as well in this range, but the higher frequencies require some special thinking about the methods of circuit design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

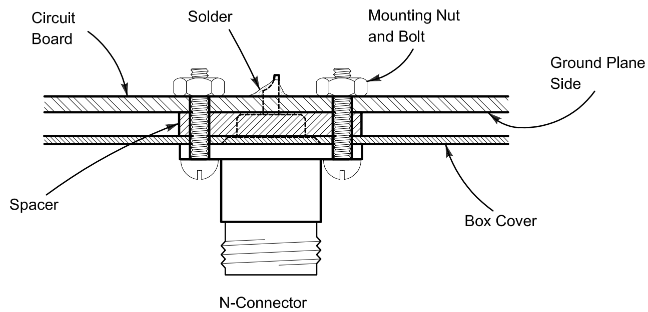
GAAS FET PREAMP FOR 430 MHZ

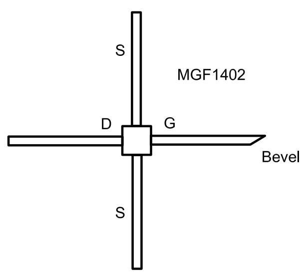
Fig 14.34 shows the schematic diagram and the physical construction of a typical RF circuit at 430 MHz. It is a GaAsFET preamplifier intended for low noise Earth-Moon-Earth or satellite reception. The construction uses chip capacitors, small helical inductors and a stripline surface-mount GaAsFET, all 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.

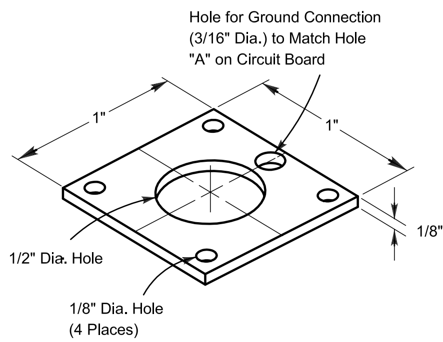
At higher frequencies, microstrip methods become more desirable in most cases because of their smaller dimensions. However, the advent of tiny chip capacitors and chip resistors have extended the frequency range of discrete components. For example, the literature shows methods of building LC filters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











Aluminum Spacer for N-Connector

HBK05_14-034

Fig 14.34—GaAsFET preamplifier schematic and construction details for 430 MHz. Illustrates circuit, parts layout and construction techniques suitable for 430-MHz frequency range.
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-µF 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).

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. 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 14.35A 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.

Cumulative Noise Figure

Fig 14.35B 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),

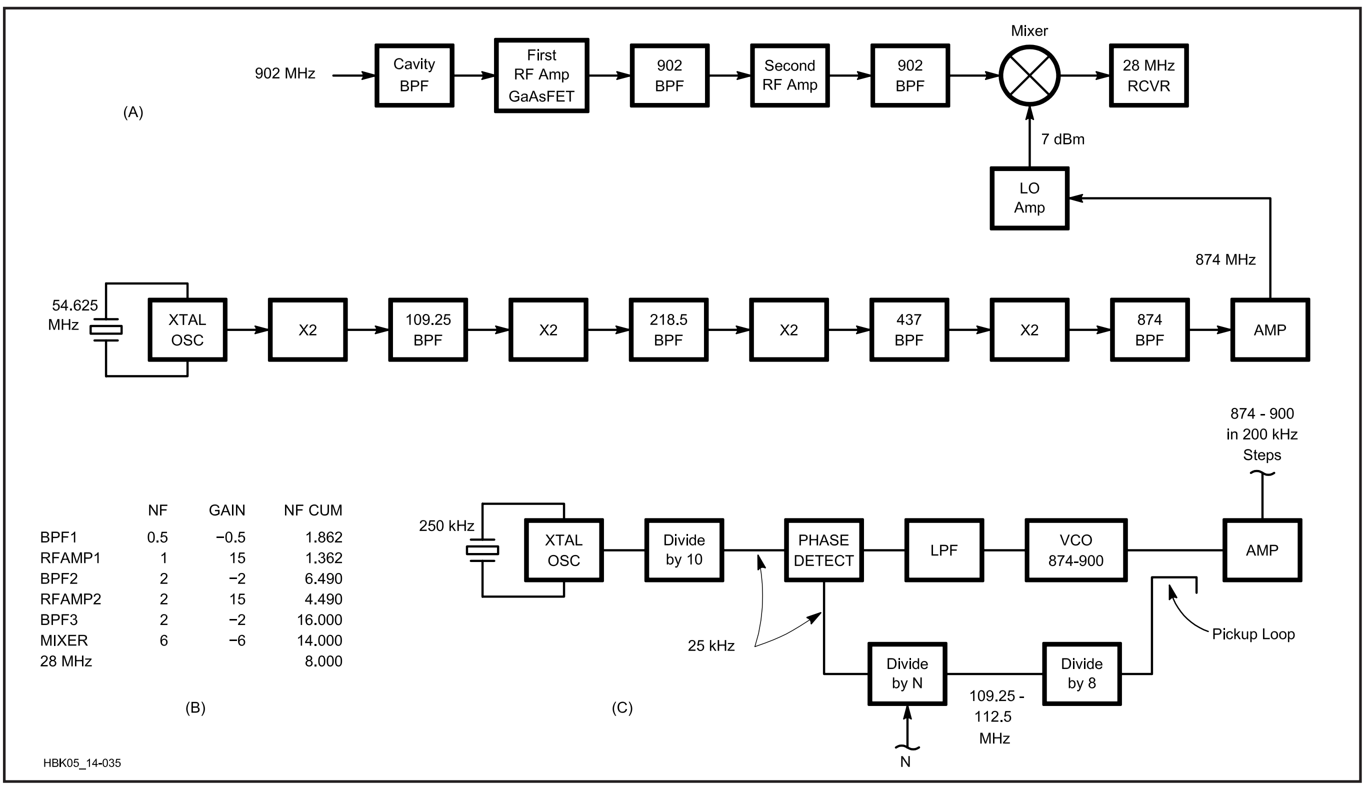


Fig 14.35—A down converter for the 902 to 928 MHz band. A: block diagram; B: cumulative noise figure of the signal path; C: alternative LO multiplier using a phase locked loop.

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 14.35C) 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 fre-

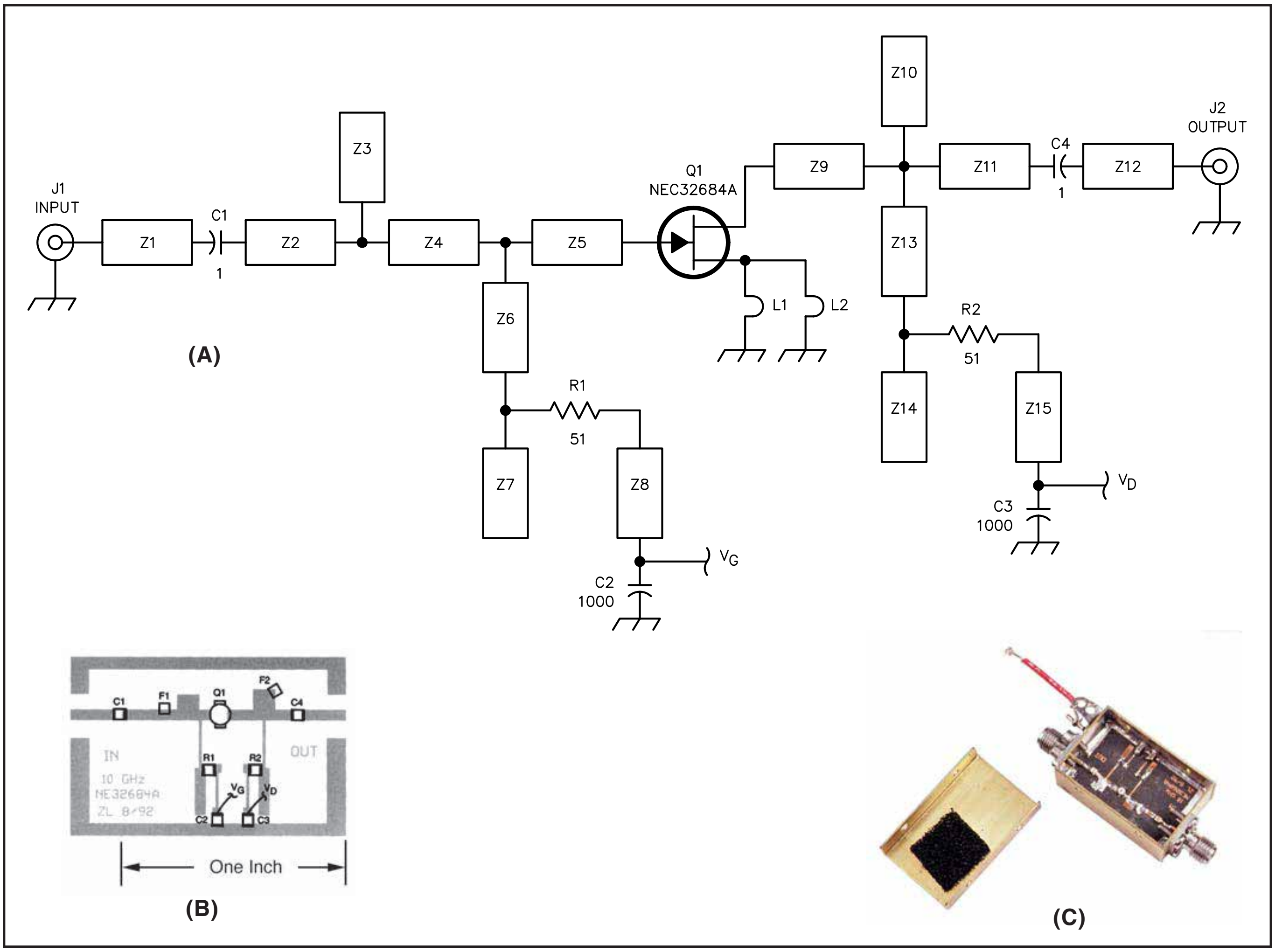


Fig 14.36—A low-noise preamplifier for 10 GHz, illustrating the methods used at microwaves. A: schematic. B: PC board layout. Use 15-mil 5880 Duroid, dielectric constant of 2.2 and a dissipation factor of 0.0011. For a negative of the board write, phone or e-mail the Technical Department Secretary at 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.

quencies 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 26, 27). 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 14.36A 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 28; 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. 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 the signal path terrain, atmosphere, transmitter and receivers systems. **Fig 14.37** shows a simplified example of how this works. This example is adapted from Dec 1980 *QST* (also see ARRL *UHF/Microwave Experimenter's Manual,* p 7-55). In the present context of receiver design we wish

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 feedline 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 14.37—Example of a 10-GHz system performance calculation. Noise temperature and noise factor of the receiver are considered in detail.

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 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 formula for cascaded noise factors.

$$TS = TG + TE1 + TE2/G1 + TE3/(G1G2) + TE4/(G1G2G3) + \ldots \tag{8}$$

where TS is the system noise temperature (including the generator, which may be an antenna) and TG 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 14.37 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:

$$N = k\ T_S\ B_N \tag{9}$$

where

$k = 1.38 \times 10^{-23}$ and

B_N= noise bandwidth

The system noise figure FS 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, FS, 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 29 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.

Block Diagram

Fig 14.38 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 14.37 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.
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 citcuit 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 to 106.5 MHz 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 29) is recommended reading for this very interesting project, which provides a fairly straightforward (but not extremely simple) way to get started on 10 GHz.

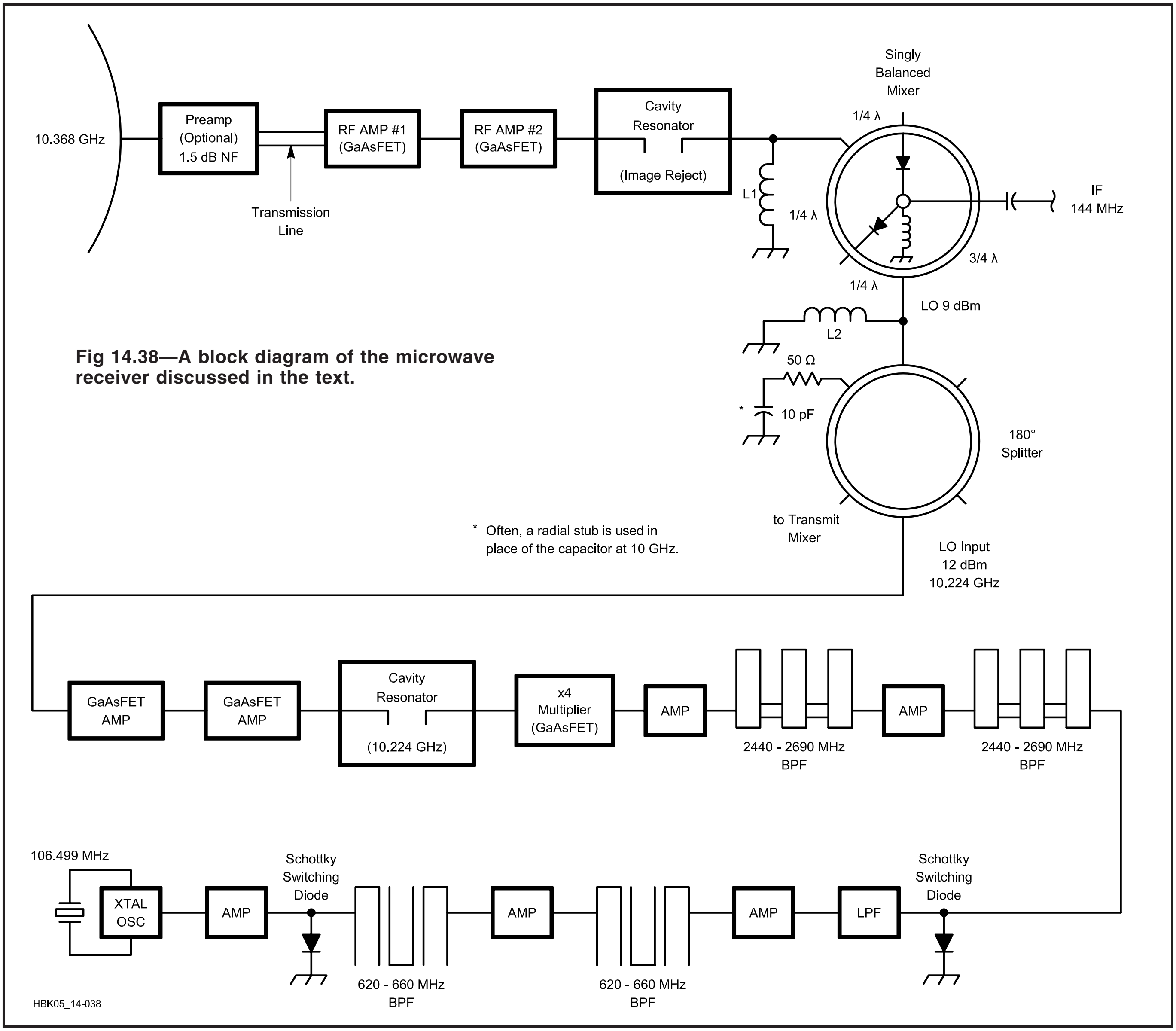


Fig 14.38—A block diagram of the microwave receiver discussed in the text.

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 14.39** is a block diagram of one approach. Let's discuss the various elements in detail,

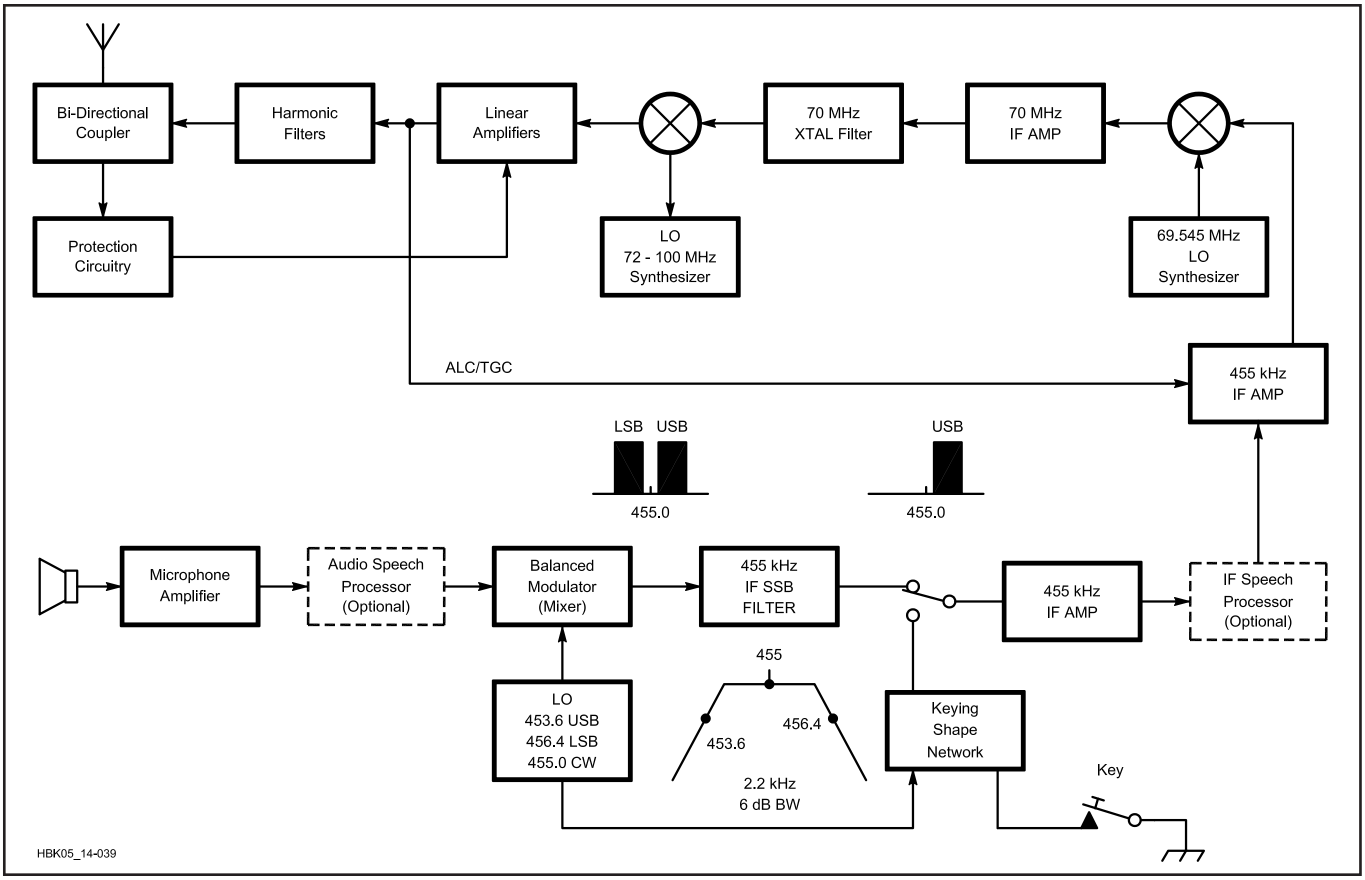


Fig 14.39—Block diagram of an up conversion SSB/CW transmitter. The various functions are discussed in the text.

starting at the microphone.

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. 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 14.40** uses a low-noise BiFET op amp. The 620-Ω resistor is selected 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.

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 14.41 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.

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 14.42A** works on this AGC principle. One section of a Signetics 571N IC is used. The other section can be connected as an 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 side-stepped 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 14.42B.

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 14.39) 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

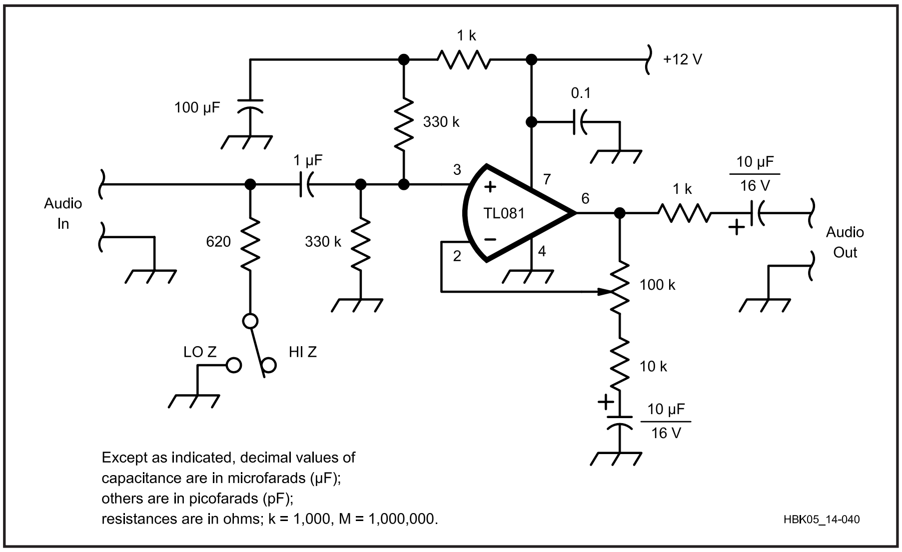


Fig 14.40—Schematic diagram of a simple op-amp microphone amplifier for low- and high-impedance microphones.

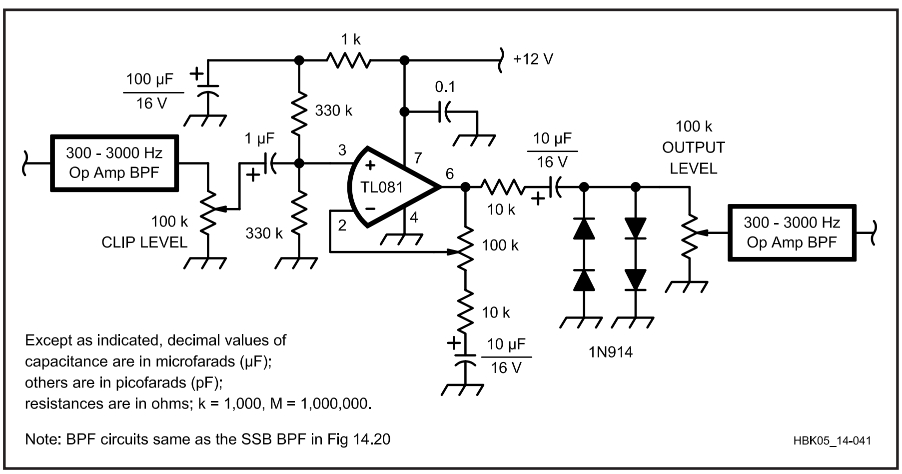


Fig 14.41—A simple audio speech clipper. The input signal is band pass filtered, amplified by 20 dB, clipped and band pass filtered again.

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 14.39 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 can be designed to transmit two independent sidebands (ISB, Ref 17).

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 20 years and is still one of the best and least expensive. **Fig 14.43** 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 14.44A**. 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

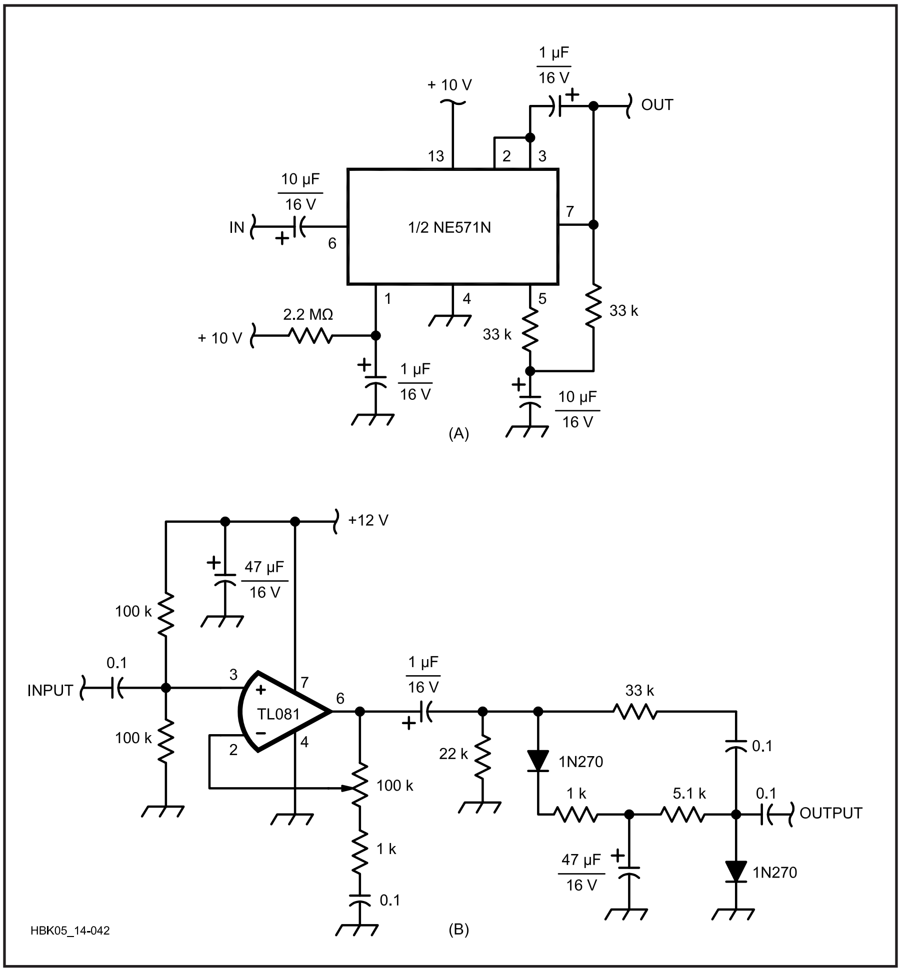


Fig 14.42—Typical solid-state compressor circuits. The circuit at A works on the AGC principle, while that at B is a forward-acting compressor.

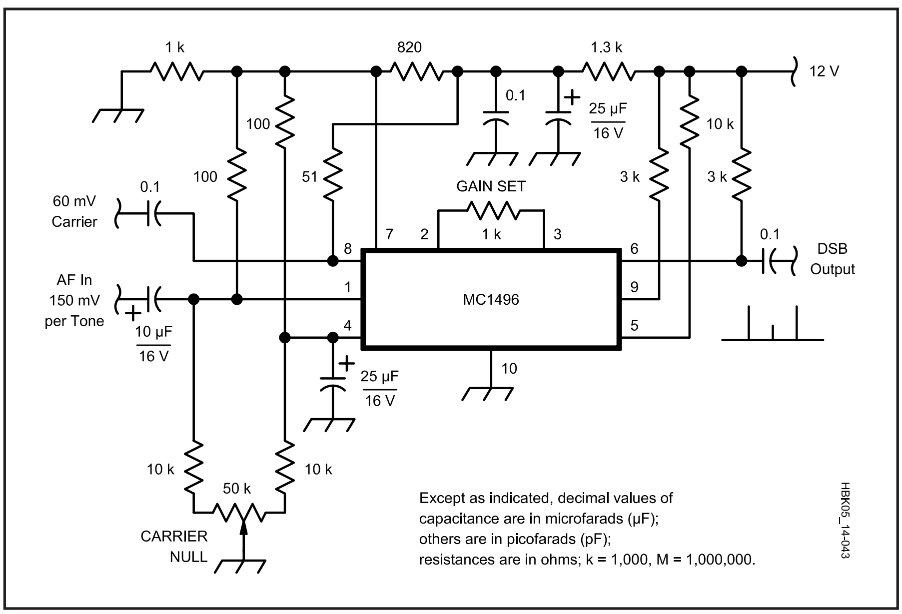


Fig 14.43—An IC balanced modulator circuit using the MC1496. The resistor from pin 2 to pin 3 sets the conversion gain.

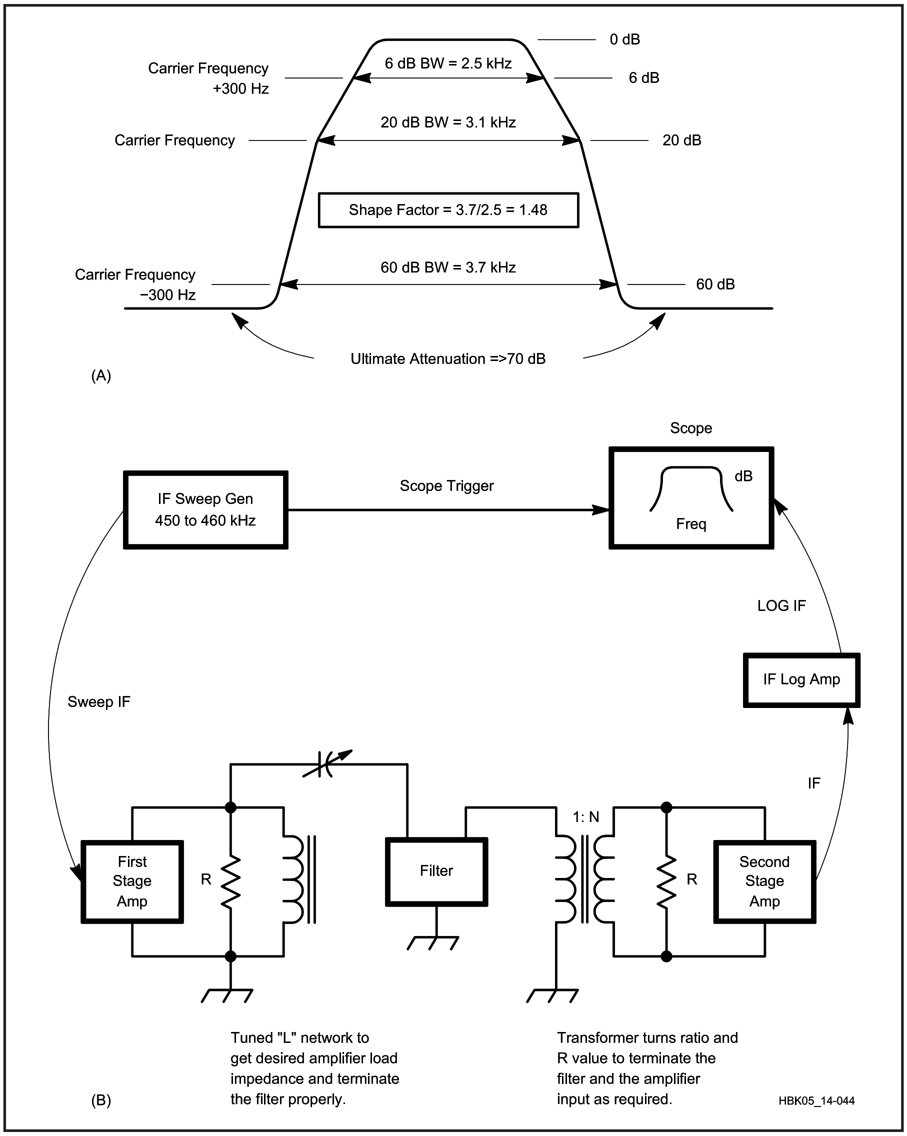


Fig 14.44—A: desired response of a SSB IF filter. B: one method of terminating a mechanical filter that allows easy and accurate tuning adjustment and also a possible test setup for performing the adjustments.

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 frequently 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 14.44B. There are three goals: The driver stage must see the desired load impedance: the stage after the filter must see the desired source (generator) impedance and the filter must be properly terminated at both ends. Fig 14.44B shows two typical approaches. 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 (Ref 17).

Fig 14.45A 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 freuqency. 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 14.46 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 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 14.45B 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 14.45A.

Another method, performed at audio, synthesizes mathematically the function of the IF clipper. This method is mentioned in Ref 17, and was an accssory 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 14.39 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 14.39, 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 14.31.

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 14.39 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

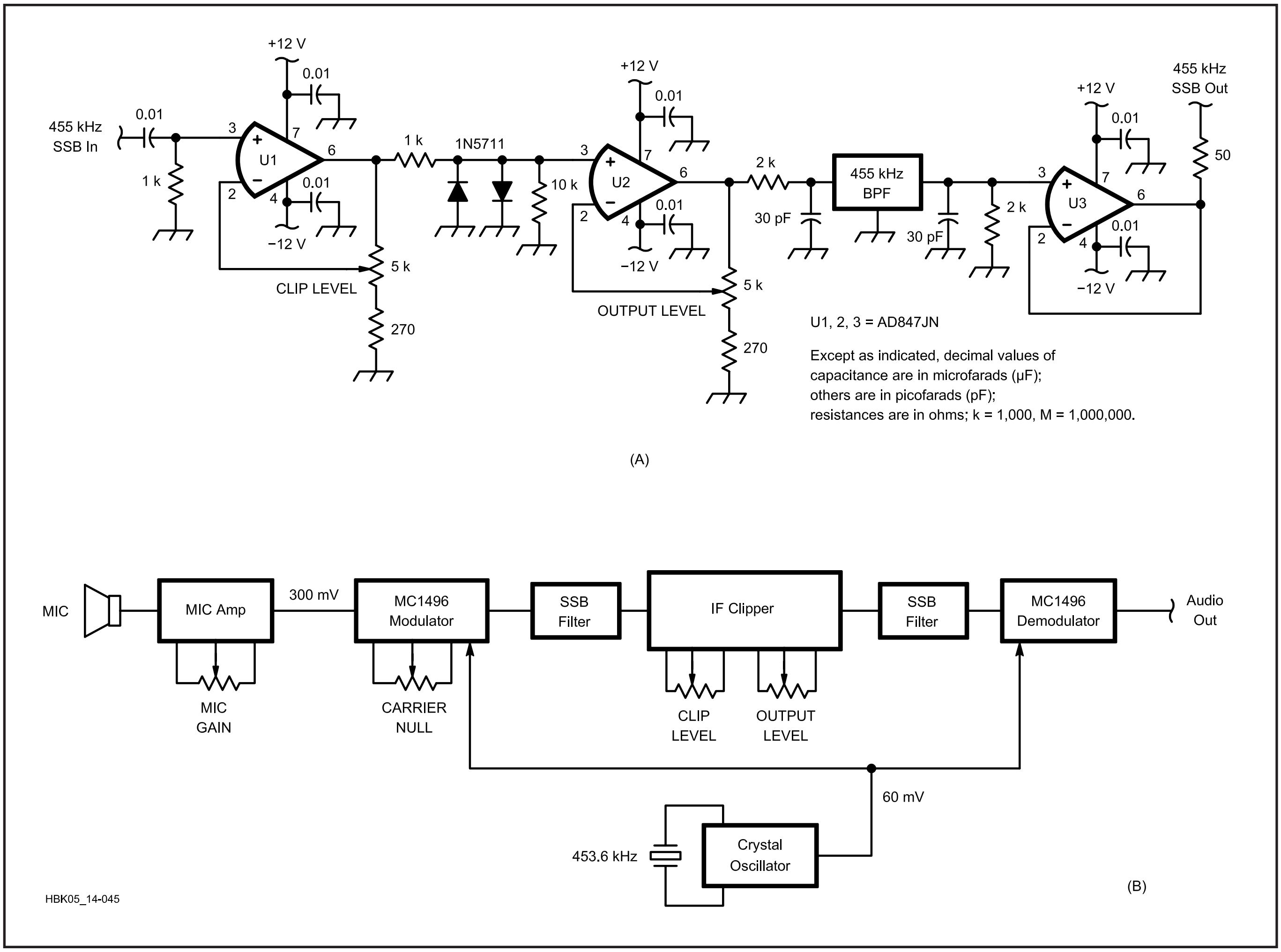


Fig 14.45—IF speech clipping. A: an IF clipper circuit approach. B: the audio signal is translated to 455 kHz, processed, and translated back to audio.

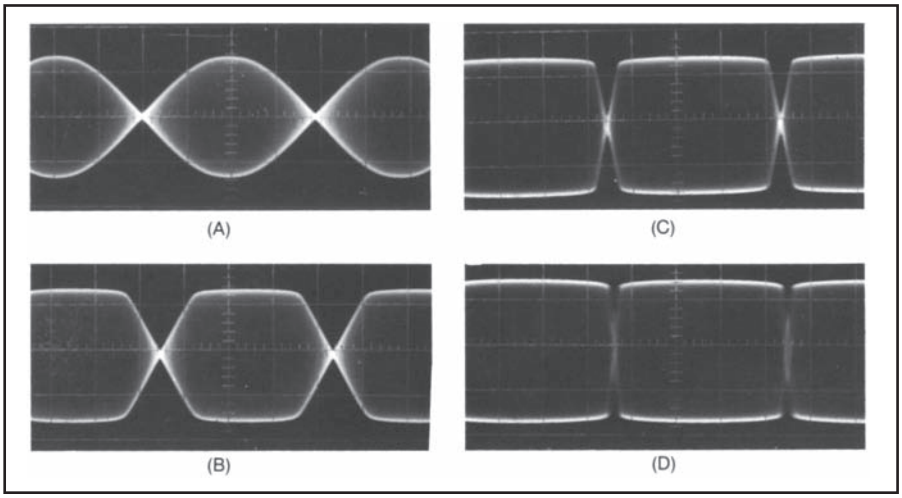


Fig 14.46—Two-tone envelope patterns with various degrees of RF clipping. All envelope patterns are formed using tones of 600 and 1000 Hz. At A, clipping threshold; B, 5 dB of clipping; C, 10 dB of clipping; D, 15 dB of clipping (from "RF Clippers for SSB," by W. Sabin, July 1967 *QST*, pp 13-18).

the LO can pass through. Each kind of balanced modulator circuit has its own method of doing this. The approach chosen in Fig 14.39 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 (CW) is a special kind of low-level amplitude modulation (a low-signal-level stage is turned on and off). It is special because the sideband power is subtracted from the carrier power, and not provided by a separate "modulator" circuit, as in high-level AM. It creates a spectrum around the carrier frequency whose amplitude and bandwidth are influenced by the rates of signal amplitude rise and fall and by the curvature of the keyed waveform. Refer to the graph of keying speed, rise and fall times, and bandwidth in the **Modes and Modulation Sources** chapter for some information about this spectrum. (That figure is repeated here as **Fig 14.47**.) The vertical axis is labeled Rise and Fall Times (ms). For a rise/fall time of 6 ms (between the 10% and 90% values) go horizontally to the line marked Bandwidth. A –20 dB bandwidth of roughly 120 Hz is indicated on the lower horizontal axis. Continuing to the K = 5 and K = 3 lines, the upper horizontal axis suggests code speeds of 30 wpm and 50 wpm respectively, as described in the text that accompanies the figure in the **Modes and Modulation Sources** chapter. These code speeds can be accommodated by the rise and fall times displayed on the vertical axis. For code speeds greater than these the Morse code characters become "soft" sounding and difficult to copy, especially under less-than-ideal propagation conditions.

For a narrow spectrum and freedom from adjacent channel interference, a further requirement is that the spectrum must fall off very rapidly beyond the –20 dB bandwidth indicated in Fig 14.47. A sensitive narrow-band CW receiver that is tuned to an adjacent channel that is only 1 or 2 kHz away can detect keying sidebands that are 80 to 100 dB below the key-down level of a strong CW signal. An additional consideration is that during key-up a residual signal, called "backwave" should not be noticeable in a nearby receiver. A backwave level at least 90 dB below the key-down carrier is a desirable goal.

Fig 14.48 is the schematic of one wave-shaping circuit that has been used successfully. A Sallen-Key third-order op amp low-pass filter (0.1 dB Chebyshev response) shapes the keying waveform, produces the rate of rise and fall and also softens the leading and trailing corners just the right amount. The key closure activates the CMOS switch, U1, which turns on the 455-kHz IF signal. At the key-up time, the input to the waveshaping filter is turned off, but the IF signal switch remains closed for an additional 12 ms.

The keying waveform is applied to the gain control pin of a CLC5523 amplifier IC (similar to the CLC520, shown in Fig 14.26D). This device, like nearly all gain-control amplifiers, has a *logarithmic* control of gain, therefore some experimental "tweaking" of the capacitor values was used to get the result shown in **Fig 14.49A**. The top trace shows the on/off operation of the IF switch, U1. The signal is turned on shortly before the rise of the keying pulse begins and remains on for about 12 ms after the keying pulse is turned off, so that the waveform falls smoothly to a very low value. The result is an excellent spectrum and an almost complete absence of backwave. The bottom trace shows the resulting keyed RF-output waveshape. It has an excellent spectrum, as verified by critical listening tests. The thumps and clicks that are found in some CW transmitters are virtually absent. The rise and fall intervals have a waveshape that is approximately a cosine. Spread-spectrum frequency-hop waveforms have used this approach to minimize wideband interference.

Fig 14.49B is an accurate SPICE simulation of the wave shaping circuit output before the signal is processed by the CLC5523 amplifier. To assist in adjusting the circuit, create a steady stream of 40 ms dots that can be seen on an RF oscilloscope that is looking at the final PA output envelope. It is important to make sure that the excellent waveshape is not degraded on its way to the transmitter output. Single-Sideband linear power amplifiers are well suited for a CW transmitter, but they must stay within their linear range, and the backwave problem must be resolved.

When evaluating the spectrum of an incoming CW signal during on-the-air operations, a poor receiver design can contribute problems caused by its vulnerability to a strong but clean adjacent channel signal. Clicks, thumps, front end overload, reciprocal mixing, etc can be created in the receiver. It is important to put the blame where it really belongs.

For additional information see "A 455-kHz IF Signal Processor for SSB/CW," William Sabin, WØIYH, *QEX*, March/April 2002, pp 11-16.

Wideband Noise

In the block diagram of Fig 14.39 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. Wideband 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 8).

The amplifiers after this mixer are also

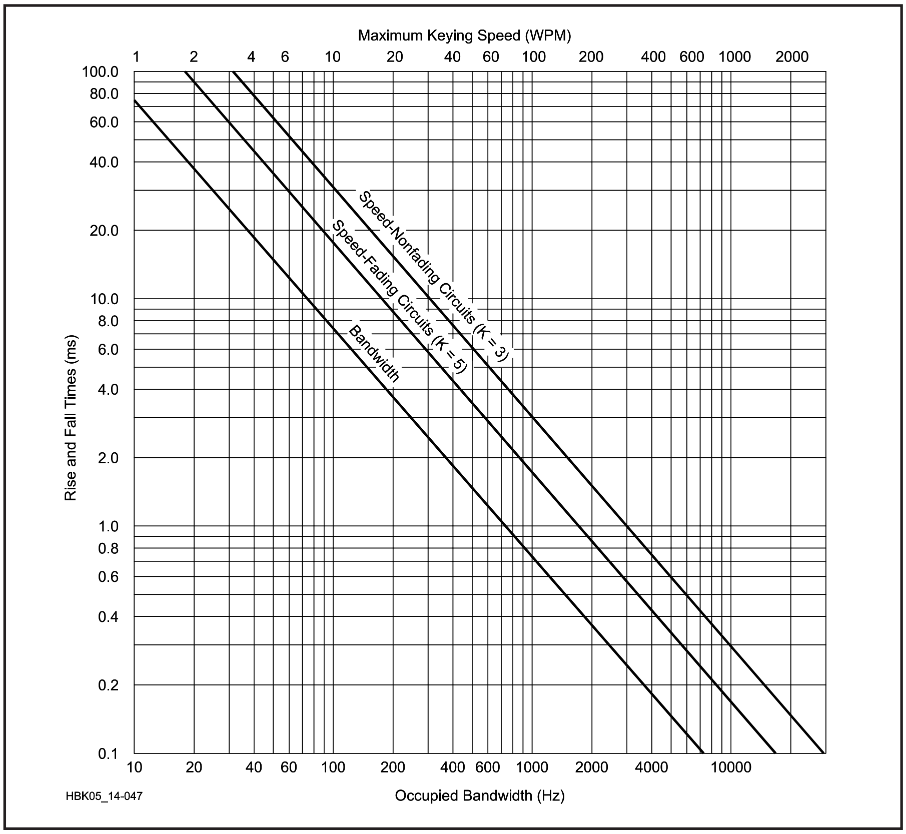


Fig 14.47—Keying speed vs rise and fall times vs bandwidth for fading and nonfading communications circuits. For example, for transmitter output waveform rise and fall times of approximately 6 ms, draw a horizontal line from 6.0 ms on the Rise and Fall Times scale to the Bandwidth line. Then draw a vertical line to the occupied bandwidth scale at the bottom of the graph. In this case the bandwidth is about 130 Hz. Also extend the 6.0 ms horizontal line to the K = 3 line for a nonfading circuit. Finally draw a vertical line from the K = 3 line to the wpm axis. The 6 ms rise and fall time should be suitable for keying speeds up to about 50 wpm in this example.

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 14.39 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."

Automatic Level Control (ALC)

The purpose of ALC is to prevent the various stages in the transmitter from being overdriven. Over-drive 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 14.39.

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 a while for the enhancement to take place, so weak components right after a strong peak are not enhanced very much. **Fig 14.50A** shows a complete ALC circuit that performs speech processing.

ALC in Solid-State Power Amplifiers

Fig 14.50B 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.

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 semiper-

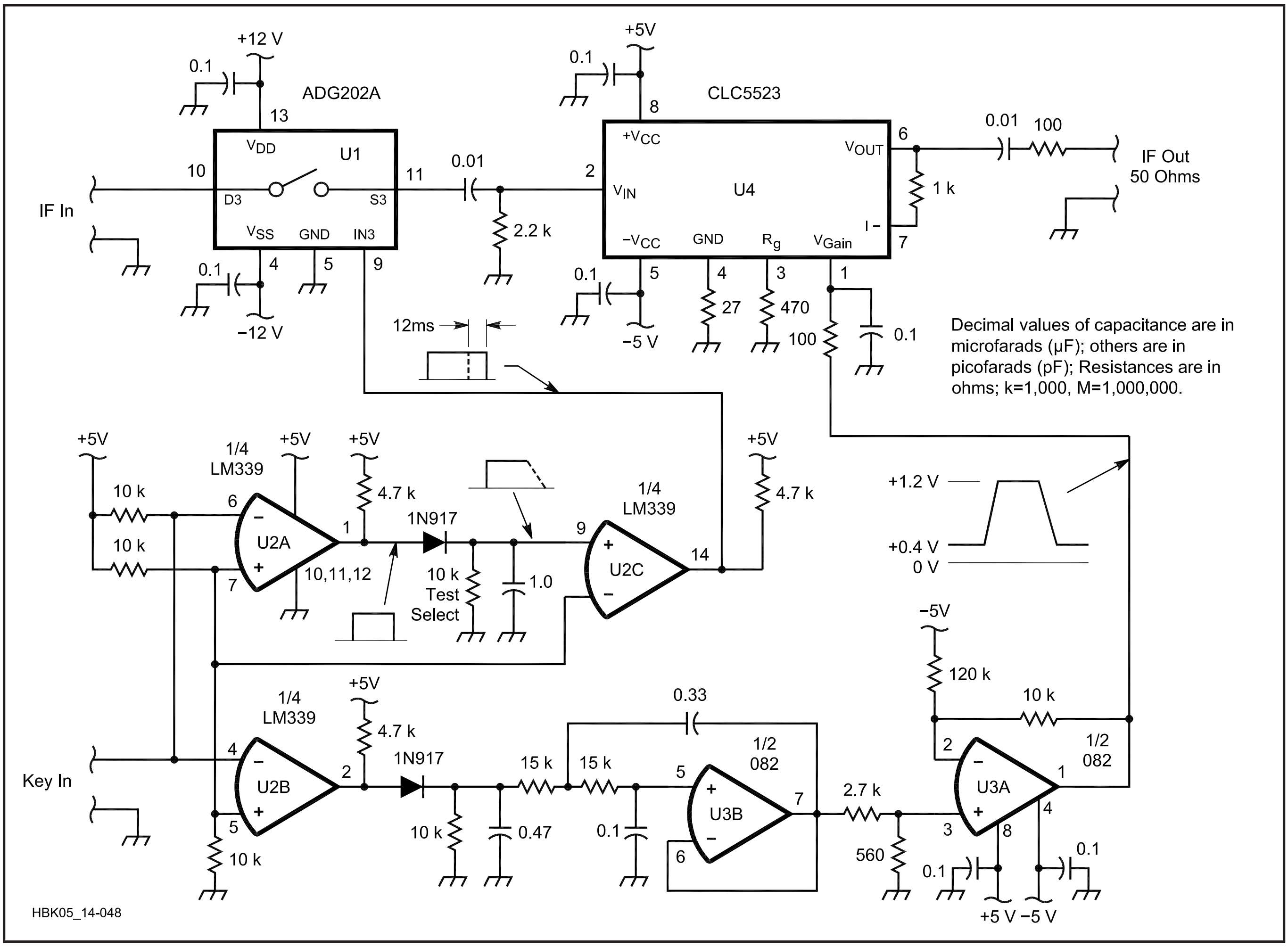
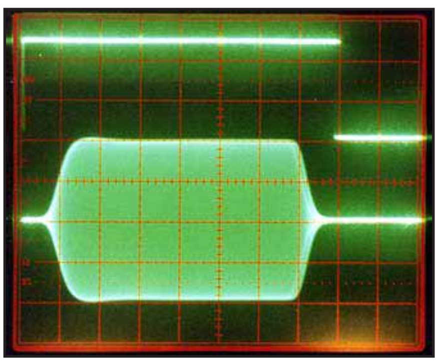
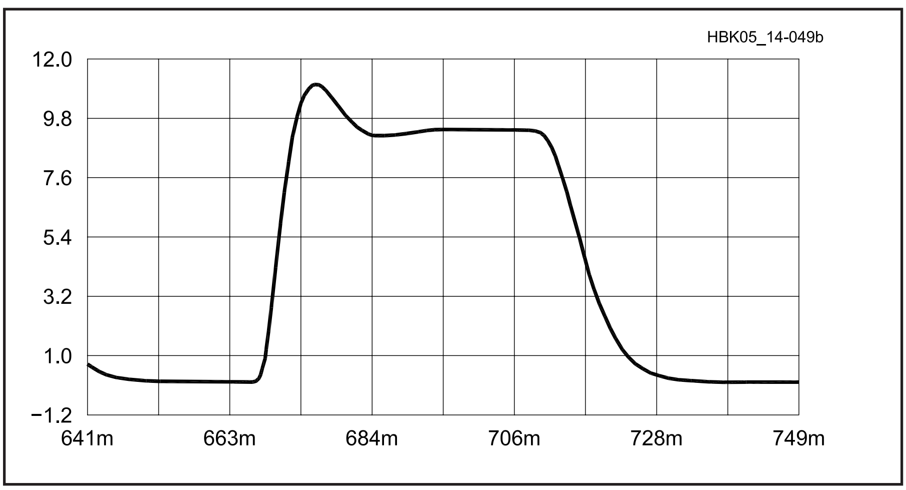


Fig 14.48—This schematic diagram shows a CW waveshaping and keying circuit suitable for use with an SSB/CW transmitter such as is shown in Fig 14.39.



(A)



(B)

Fig 14.49—Part A is the oscilloscope display of the CW waveshaping and keying circuit output. The top trace is the IF keying signal applied to S1 of Fig 14.48. The bottom trace is the transmitter output RF spectrum. Part B is a SPICE simulation of the waveshaping network. When this signal is applied to the logarithmic control characteristic of the CLC5523 amplifier, the RF envelope is modified slightly to the form shown in A.

manent. 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.

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.
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 that 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.

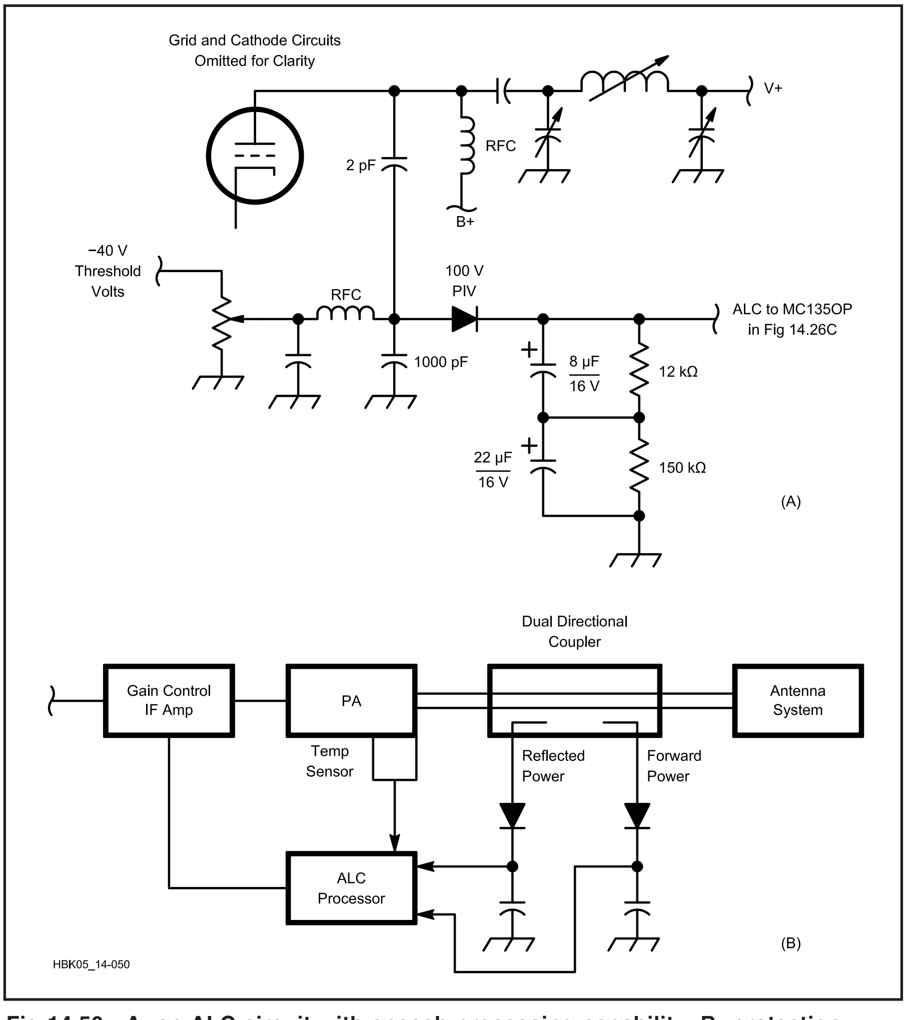


Fig 14.50—A: an ALC circuit with speech processing capability. B: protection method for a solid-state transmitter.

Frequency Multipliers

A passive multiplier using diodes is shown in **Fig 14.51A**. 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 14.51B. The FETs are biased in the square-law region and the 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 14.51C 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 14.51D. The input and output are both push-pull. The balance potentiometer

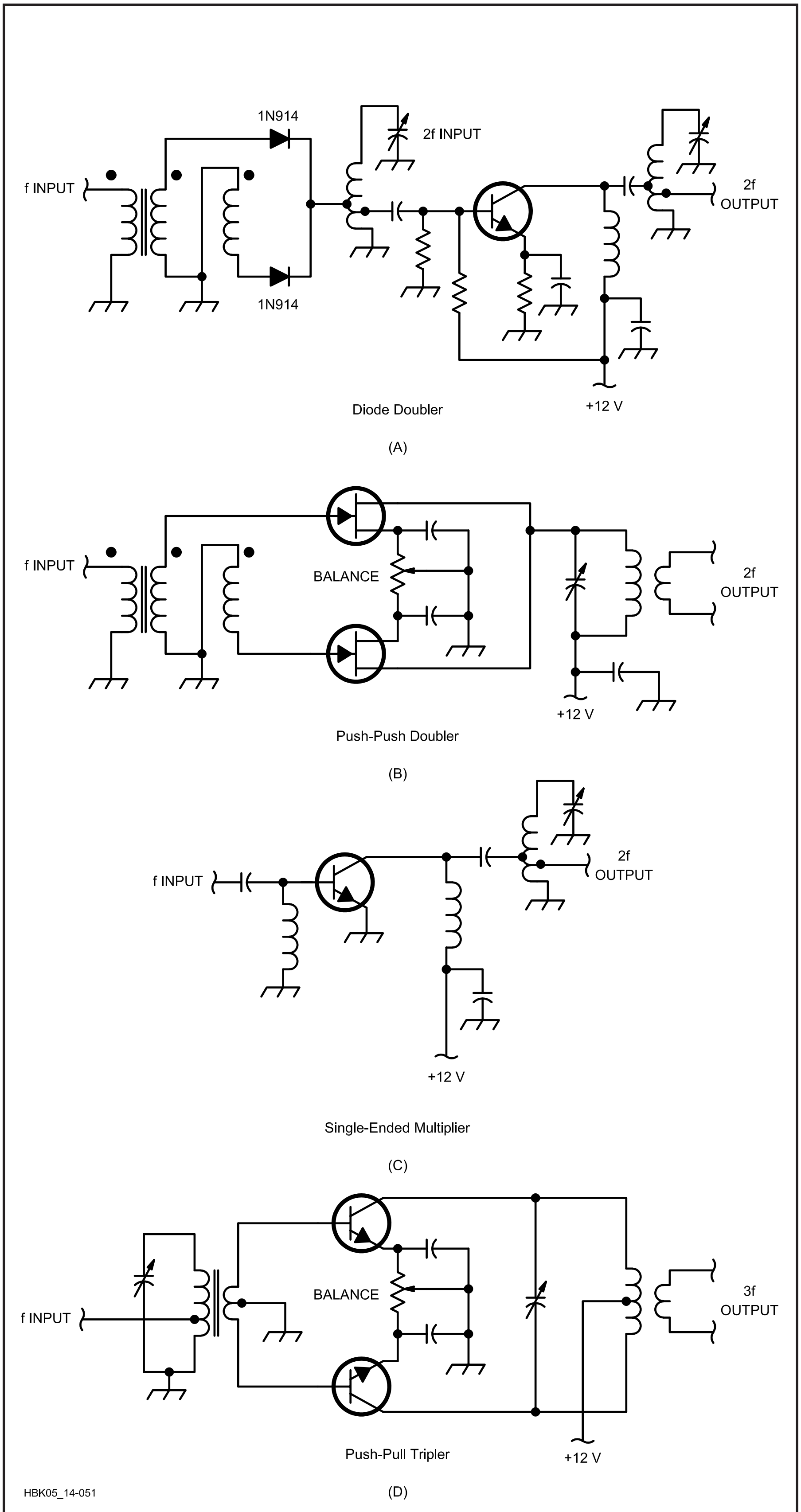


Fig 14.51—A: diode doubler. B: push-push doubler using JFETS. C: single-ended multiplier using a BJT. D: push-pull tripler using BJTs.

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. 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. 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 14.52A** is a typical schematic. For more information regarding design details there are two References: Hewlett-Packard application note AN-920 and Ref 30.

The varactor diode can also be used as a multiplier. Fig 14.52B 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.

IMPEDANCE TRANSFORMATION BETWEEN CASCADED CIRCUITS

One of the most common tasks the electronics designer encounters is to correctly interface between the output of one circuit and the input of an adjacent circuit, so that both are operating in the desired manner. We introduced this in Figs 14.1 and 14.2. In this segment we will present a unified overview of the general topic that should be helpful to the designer of RF circuits. This is a very large topic, so we must stick to basic ideas and give References. The networks we will consider are in two categories, broadband and narrowband. The modern trend in Amateur Radio is to employ the personal computer, using software such as *ARRL Radio Designer*, *SPICE*, *Mathcad*, the Smith Chart and associated design programs.

Impedance "Matching" and "Transformation"

Fig 14.53A shows a network connected between a generator with internal resistance R_{IN} and a load R_{OUT}. In this case the load and source are "matched" because each sees itself, looking into the network. As Figs 14.1

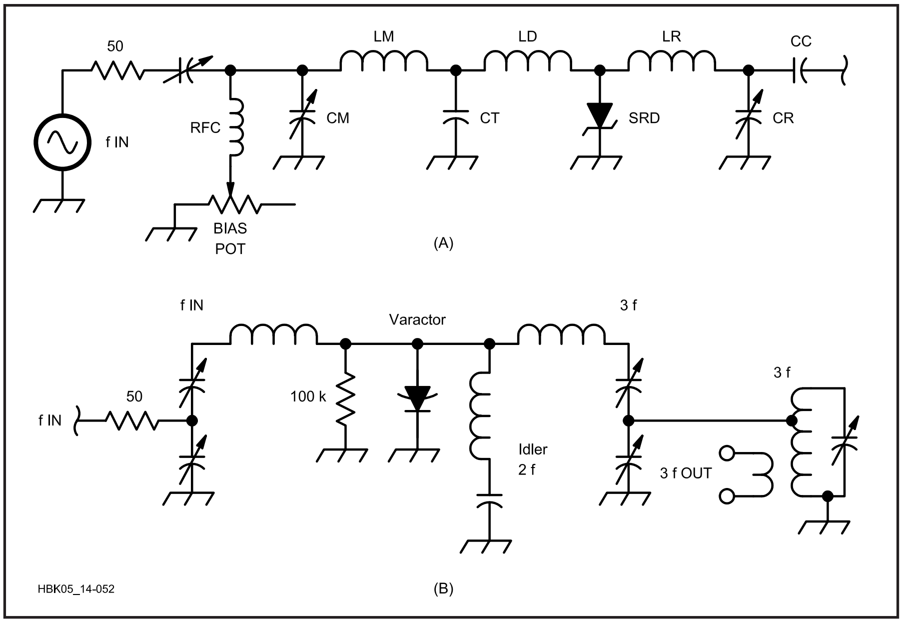


Fig 14.52—Diode frequency multipliers. A: step-recovery diode multiplier. B: varactor diode multiplier.

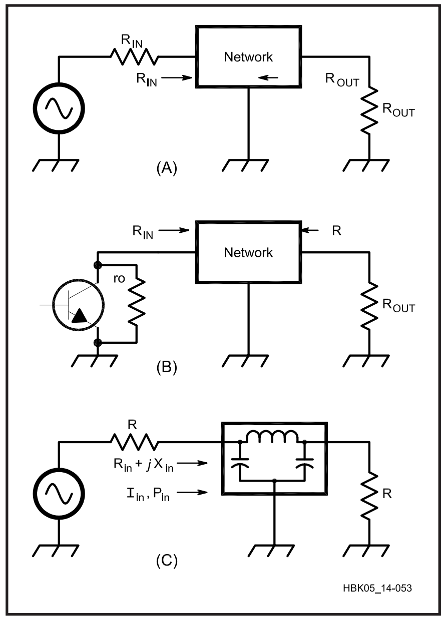


Fig 14.53—At A, matching network driven by generator. At B, matching network driven by transistor. At C, low-pass filter network.

and 14.2 explained, this idea extends to the idea of "conjugate match." There are many circuits that require this kind of impedance matching for textbook operation. In Fig 14.53B, showing a transistor amplifier, a different situation may exist. The transistor sees the R_{IN} that it needs for correct operation of the transistor.

That is, for a certain value of collector DC voltage and current and a certain allowed maximum value of collector RF voltage and current, a certain value of R_{IN} is required. However, the load R_{OUT}, looking back, may not see itself, but something much different, R, because the output resistance r_o of the transistor is not guaranteed to be the same as R_{IN} in Fig 14.53A. In this situation we could say that the load R_{OUT} is "transformed" to R_{IN}. This difference is important to understand in many applications. In some small-signal transistor circuits the dynamic output resistance r_o of the transistor and R_{OUT} are actually "matched" by a network as in Fig 14.53A. [Note: "dynamic" means that r_o is not a physical resistor but an internal, lossless *negative-feedback* property of the transistor]. In many other situations correct circuit performance requires that R_{OUT} see some specific value other than itself, looking into the output of the network.

A frequent practice for networks, including the *resonant* coupling networks (see Fig 14.59) is to combine a physical resistor with r_o so that the network is properly terminated at the input end. If the resistor is used, the transistor then sees the network and resistor combination; this combination should be the load that we want. For example, suppose r_o is 10 kΩ and an R_{IN} of 1 kΩ (including r_o) is required. A 2 kΩ network and a 2.5 kΩ resistor in parallel could be used and the network is then correctly double-terminated as in Fig 14.53A.

Of course, some signal power will be lost in the 2.5 kΩ resistor, but in low-level circuits we often accept that. An elegant alternative is to use additional negative feedback in the transistor circuit to reduce r_o to 1 kΩ and use a 1 kΩ network. In some circuits, especially RF power amplifiers, the dissipation in an extra resistor cannot be tolerated, and in this case the network internal impedance R, looking into the output terminals, is usually ignored and the main goal is to get the right R_{IN}.

The **RF Power Amplifiers** chapter shows how Pi and Pi-L networks for power amplifiers are calculated. Use simulation to verify that the circuit is behaving as you want and to get network component losses. Simulations also show that the loading by a resistor or tube/transistor dynamic r_o at the R_{IN} end changes the selectivity, and this can be monitored by actual measurement. The Q and LC values can be modified as needed.

Mismatch in Lossless Networks

Fig 14.53C shows a generator with resistance R and a lowpass filter (LPF) connected to an R load resistor. At very low frequencies the filter is "transparent" and the maximum power is delivered to the load. The inductances and capacitances are assumed to be lossless and the generator sees the R load. But at high frequencies the reactances cause the output power to be attenuated. The input impedance is $R_{IN} \pm j X_{IN}$. The power (real power) delivered to the filter input is $P_{IN} = I_{IN}^2 \times R_{IN}$.

Because L and C are lossless, this power P_{IN} must be identical to the power that is actually delivered to the R load. I_{IN}, X_{IN} and R_{IN} in the low-pass filter (LPF) are modified by the reactances in such a way that P_{IN}, therefore P_{LOAD}, is reduced. These changes are equivalent to an impedance mismatch between the generator and the filter. Or, we can say the generator no longer sees the correct load resistance R. This is the basic mechanism by which lossless networks control the frequency response.

BROADBAND TRANSFORMERS

Conventional Transformers

Fig 14.54A 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 14.54B 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 14.54C shows how these elements determine the frequency response, including a resonant 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 14.8, are often "conventional."

Fig 14.54D considers a typical application of a conventional transformer in a linear Class-A RF power amplifier. The load is 50 Ω and the maximum allowable transistor collector voltage and current excursions for linear operation are shown. The value of DC current and the sinewave limits are determined by studying the collector voltage-current curves (or constant-current curves) in the data manual to find the most linear region. To deliver this power to the 50 Ω load, the turns ratio is calculated from the equations. In an RF amplifier a ferrite or powdered-iron core

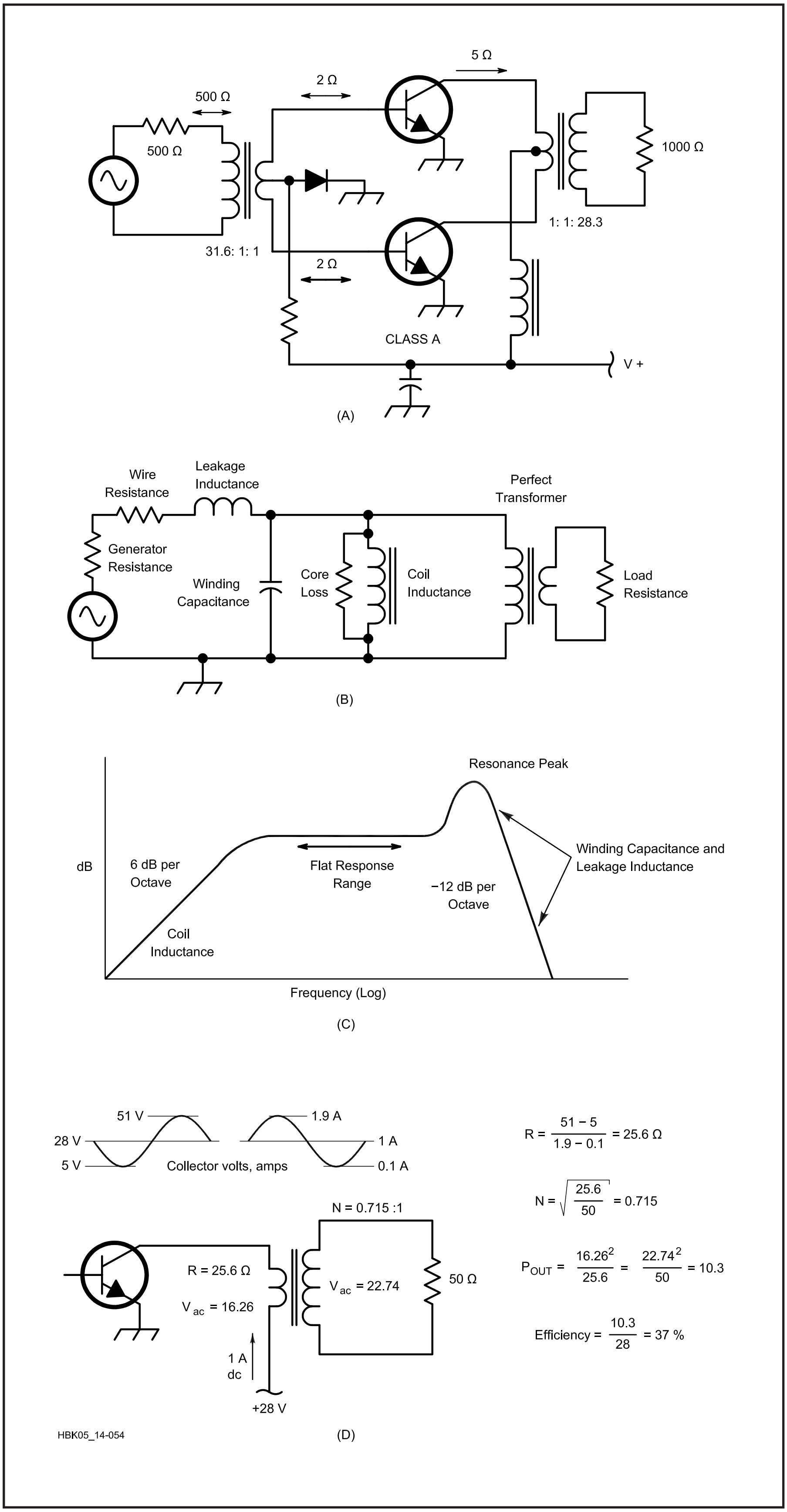


Fig 14.54—Conventional transformers in an RF power amplifier. Leakage reactances, stray capacitances and core magnetizations limit the bandwidth and linearity, and also create resonant peaks. D shows design example of transformer-coupled RF power amplifier.

would be used. The efficiency in this example is 38%.

Transmission Line Transformers

The basic transmission line transformer, from which other transformers are derived, is the 1:1 choke (or current) balun, shown in **Fig 14.55A**. We consider the following basic properties:

- A pair of close-spaced wires or a length of coax (ie, a transmission line) wraps around a ferrite rod or toroid or through a number of beads. For the 3.5 to 29.7 MHz band, type 43 ferrite ($\mu = 850$), or equivalent, is usually preferred. Other types such as 77 (at 1.8 MHz, $\mu = 2000$) or 61 (at VHF bands, $\mu = 120$) are used. The Z_0 of the line should equal R.
- Because of the ferrite, a large impedance exists between points A and C and a virtually identical impedance between B and D. This is true for parallel wires and it is also true for coax. The ferrite affects the A to C impedance of the coax inner conductor and the B to D impedance of the outer braid equally.
- The conductors (two wires or coax braid and center-wire) are tightly coupled by electromagnetic fields and therefore constitute a good conventional transformer with a turns ratio of 1:1. The voltage from A to C is equal to and in-phase with that from B to D. These are called the *common-mode voltages* (CM).
- A common-mode (CM) current is one that has the same value and direction in both wires (or braid and center wire). Because of the ferrite, the CM current encounters a high impedance that acts to reduce (choke) the current. The normal differential-mode (DM) signal does not encounter this CM impedance because the electromagnetic fields due to equal and opposite currents in the two conductors cancel each other at the ferrite, so the magnetic flux in the ferrite is virtually zero.
- The main idea of the transmission line transformer is that although the CM impedance may be very large, the DM signal is virtually unopposed, especially if the line length is a small fraction of a wavelength.
- A common experience is a CM current that flows on the outside of a coax braid due to some external field, such as a nearby antenna or noise source. The balun reduces (chokes) the CM current due to these sources. But it is very important to keep in mind that the common-mode voltage across the ferrite winding that is due to this current is efficiently coupled to the center wire by conventional transformer action, as mentioned before and easily verified. This equality of CM voltages, and also CM impedances, reduces the *conversion* of a CM signal to an *undesired* DM signal that can interfere with the *desired* DM signal in both transmitters and receivers.
- The CM current, multiplied by the CM impedance due to the ferrite, produces a CM voltage. The CM impedance has L and C reactance and also R. So L, C and R cause a broad parallel self-resonance at some frequency. The R component also produces some dissipation (heat) in the ferrite. This dissipation is an excellent way to dispose of a small amount of unwanted CM power.
- The main feature of the ferrite is that the choke is effective over a bandwidth of one, possibly two decades of frequency. In addition to the ferrite choke balun, straight or coiled lengths of coax

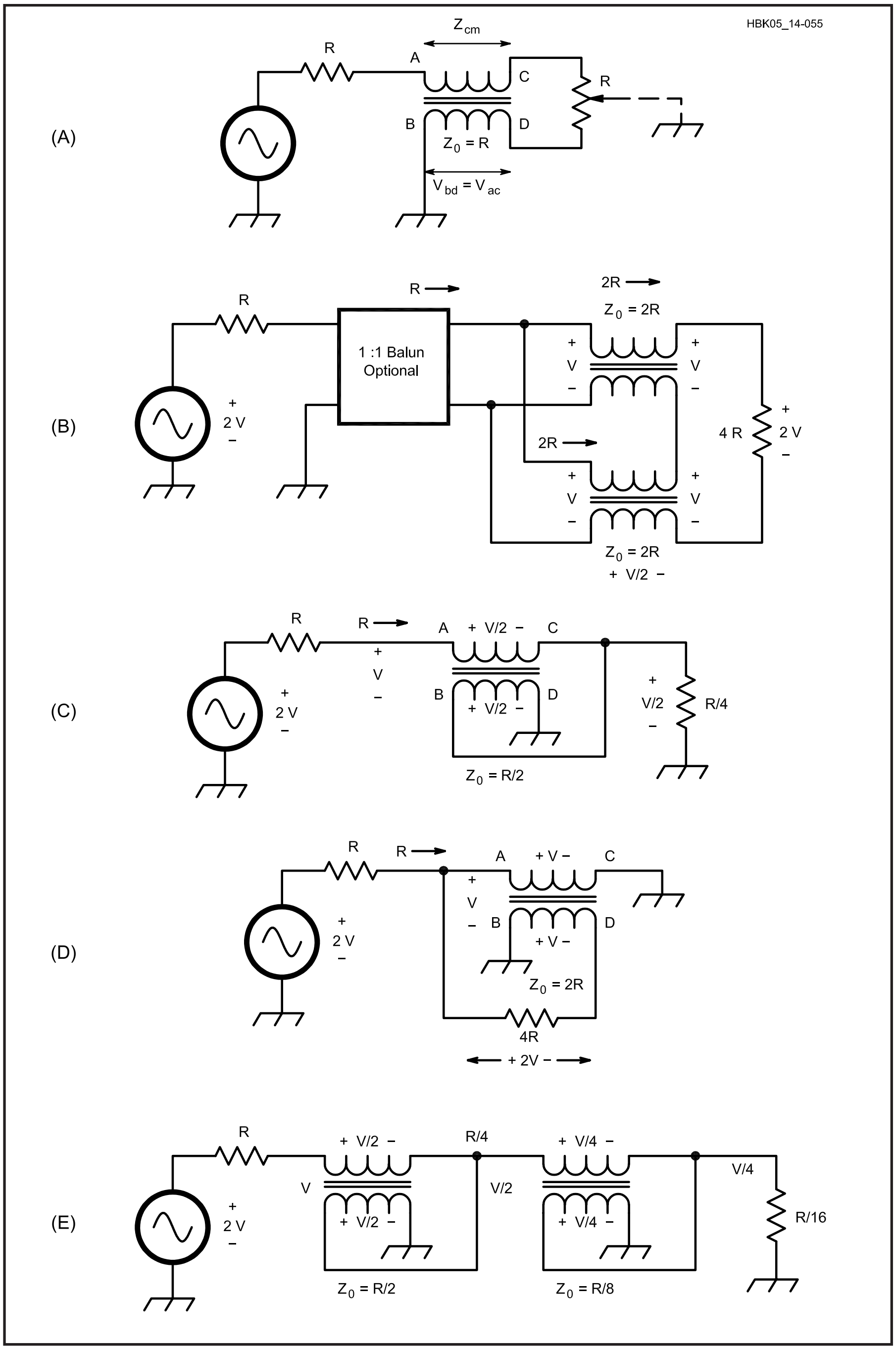


Fig 14.55—At A, basic balun. At B, 1:4 Guanella transformer. At C, Ruthroff transformer, 4:1 unbalanced. At D, Ruthroff 1:4 balanced transformer. At E, Ruthroff 16:1 unbalanced transformer.

(no core and almost no CM dissipation) are used within narrow frequency bands. A one-quarter-wave length of transmission line is a good choke balun at a single frequency or within a narrow band.

- The two output wires of the balun in Fig 14.55A have a high impedance with respect to, and are therefore "isolated" from, the generator. This feature is very useful because now any point of R at the output can be grounded. In a well-designed balun circuit *almost* all of the current in one conductor returns to the generator through the other conductor, despite this ground connection. Note also that the ground connection introduces some CM voltage across the balun cores and this has to be taken into account. This CM voltage is maximum if point C is grounded. If point D is grounded and if all "ground" connections are at the same potential, which they often are not, the CM voltage is zero and the balun may no longer be needed. In a coax balun the return current flows on the inside surface of the braid.

We now look briefly at a transmission line transformer that is based on the choke balun. Fig 14.55B shows two identical choke baluns whose inputs are in parallel and whose outputs are in series. The output voltage amplitude of each balun is identical to the common input, so the two outputs add in-phase (equal time delay) to produce twice the input voltage. It is the high CM impedance that makes this voltage addition possible. If the power remains constant the load current must be one-half the generator current, and the load resistor is 2V/0.5I = 4V/I = 4R.

The CM voltage in each balun is V/2, so there is some flux in the cores. The right side floats. This is named the *Guanella* transformer. If Z_0 of the lines equals 2R and if the load is pure resistance 4R then the input resistance R is independent of line length. If the lines are exactly one-quarter wavelength, then $Z_{IN} = (2R)^2 / Z_L$, an impedance inverter, where Z_{IN} and Z_L are complex. The quality of balance can often be improved by inserting a 1:1 balun (Fig 14.55A) at the left end so that both ends of the 1:4 transformer are floating and a ground is at the far left side as shown. The Guanella can also be operated from a grounded right end to a floating left end. The 1:1 balun at the left then allows a grounded far left end.

Fig 14.55C is a different kind, the *Ruthroff* transformer. The input voltage V is divided into two equal in-phase voltages AC and BD (they are tightly coupled), so the output is V/2. And because power is constant, $I_{OUT} = 2I_{IN}$ and the load is R/4. There is a CM voltage V/2 between A and C and between B and D, so in normal operation the core is not free of magnetic flux. The input and output both return to ground so it can also be operated from right to left for a 1:4 impedance stepup. The Ruthroff is often used as an amplifier interstage transformer, for example between 200 Ω and 50 Ω. To maintain low attenuation the line length should be much less than one-fourth wavelength at the highest frequency of operation, and its Z_0 should be R/2. A balanced version is shown in Fig 14.55D, where the CM voltage is V, not V/2, and transmission is from left-to-right only. Because of the greater flux in the cores, no different than a conventional transformer, this is not a preferred approach, although it could be used with air wound coils (for example in antenna tuner circuits) to couple 75 Ω unbalanced to 300 Ω balanced. The tuner circuit could then transform 75 Ω to 50 Ω.

Fig 14.56 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 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.

Tips on Toroids and Coils

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, also 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 toroidal 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, but how many of us are that lucky? Spot checks with an attenuator ahead of a receiver that is tunable to the harmonics are also very helpful.

There are many transformer schemes that use the basic ideas of Fig 14.55. Several of them, with their toroid winding instructions, are shown in **Fig 14.57**. Because of space limitations, for a comprehensive treatment we suggest Jerry Sevick's books *Transmission Line Transformers* and *Building and Using Baluns and Ununs*, both available from ARRL. For applications in solid-state RF power amplifiers, see Sabin and Schoenike, *HF Radio Systems and Circuits*, Chapter 12, also available from ARRL.

Tuned (Resonant) Networks

There is a large class of LC networks that utilize resonance at a single frequency to transform impedances over a narrow band. In many applications the circuitry that the network connects to has internal reactances, inductive or capacitive, combined with resistance. We want to absorb these reactances, if possible, to become an integral part of the network design. By looking at the various available network possibilities we can identify those that will do this at one or both ends of the network. Some networks must operate between two different values of resistance, others can also operate between equal resistances. As mentioned before, nearly all networks also allow a choice of selectivity, or Q, where Q is (approximately) the resonant frequency divided by the 3-dB bandwidth.

As a simple example that illustrates the method, consider the generator and load of **Fig 14.58A**. We want to absorb the 20 pF and the 0.1 µH into the network. We use the formulas to calculate L and C for a 500 Ω to 50 Ω L-network, then subtract 20 pF from C and 0.1 µH from L. As a second iteration we can improve the design by considering the resistance of the L that we just found. Suppose it is 2 Ω. We

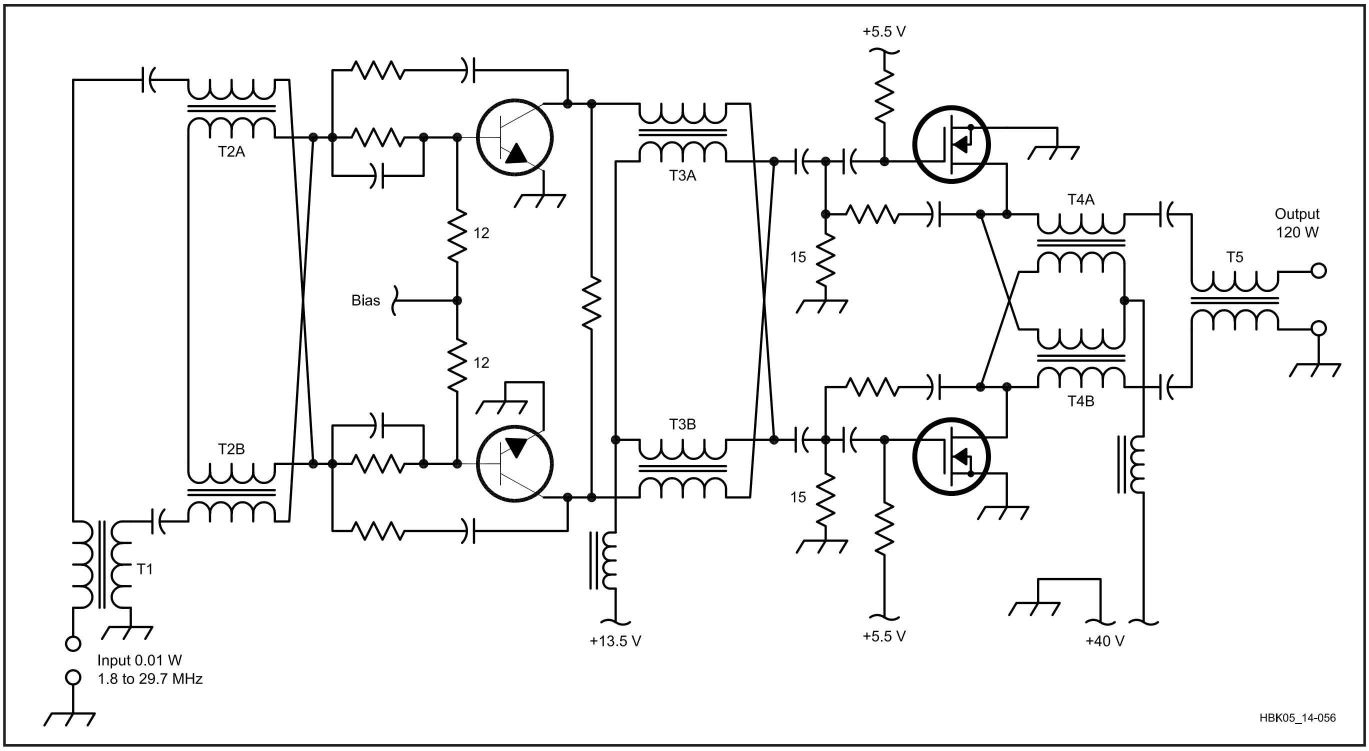


Fig 14.56—This illustrates how transmission-line transformers can be used in a push-pull power amplifier.

can recalculate new values L´ and C´ for a network from 500 Ω to 52 Ω, as shown in Fig 14.58B.

Further iterations are possible but usually trivial. More complicated networks and more difficult problems can use a computer to expedite absorbing process. Always try to absorb an inductance into a network L and a capacitance into a network C in order to minimize spurious LC resonances and undesired frequency responses. Inductors and capacitors can be combined in series or in parallel as shown in the example. Fig 14.58C shows useful formulas to convert series to parallel and vice versa to help with the designs.

A set of 14 simple resonant networks, and their equations, is presented in **Fig 14.59**. Note that in these diagrams RS is the low impedance side and RL is the high impedance side and that the X values are calculated in the top-down order given. The program *MATCH.EXE* can perform the calculations.

ARRL Radio Designer can also help a lot with special circuit-design problems and some approaches to resonant network design. It can graph the frequency response, compute insertion loss and also tune the capacitances and inductances across a frequency band. You may select the selectivity (Q) in such programs based on frequency-response requirements. The program can also be trimmed to help realize realistic or standard component values. A math program such as *Mathcad* can also make this a quick and easy process. You can find additional information for Pi and Pi-L networks in the **RF Power Amplifiers** chapter.

NBFM Transmitter Block Diagram

Fig 14.60 shows the phase-modulation method, also known as indirect FM. It is the most widely used approach to NBFM. Phase modulation is performed at 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

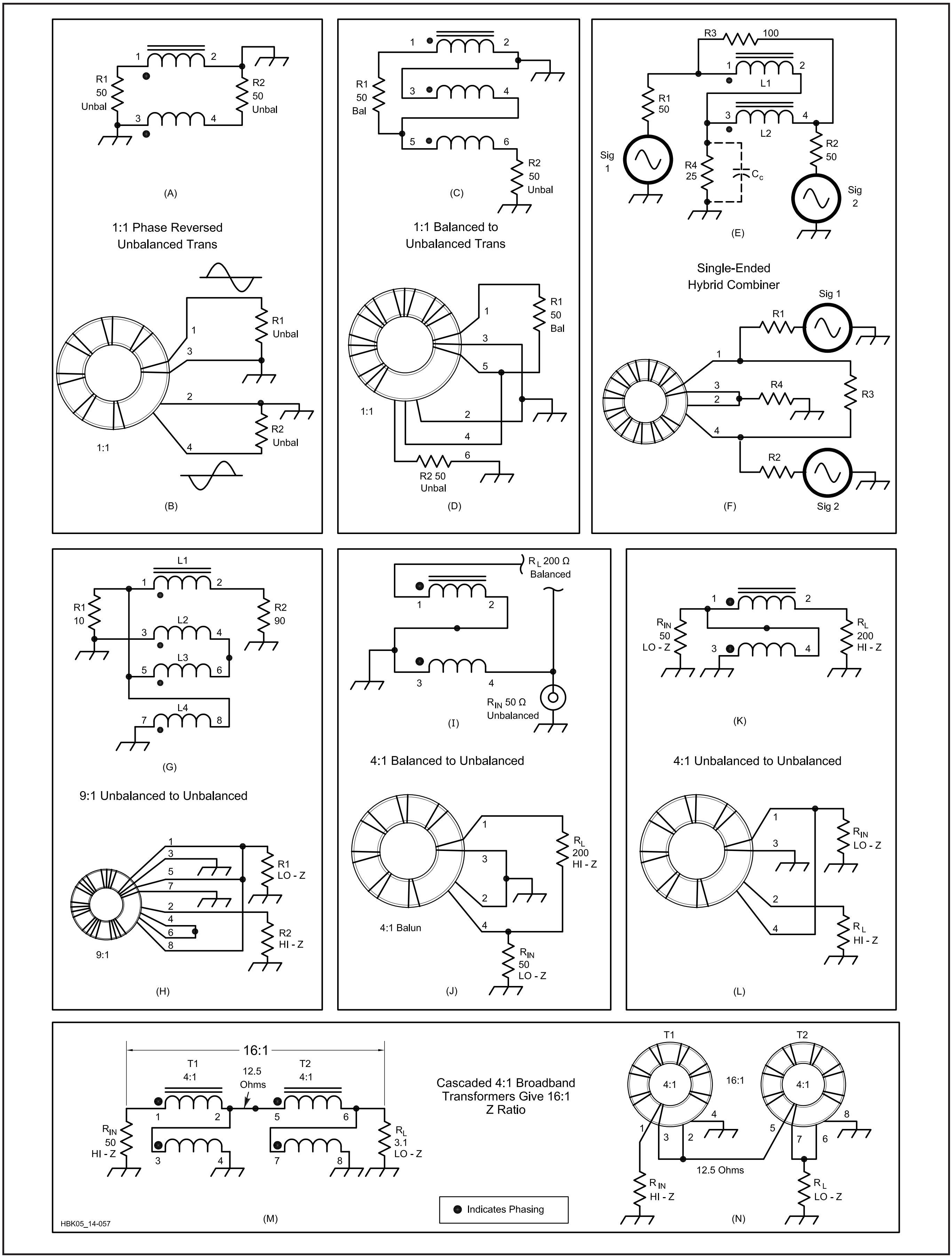


Fig 14.57—Assembly instructions for some transmission-line transformers. See text for typical magnetic materials used.

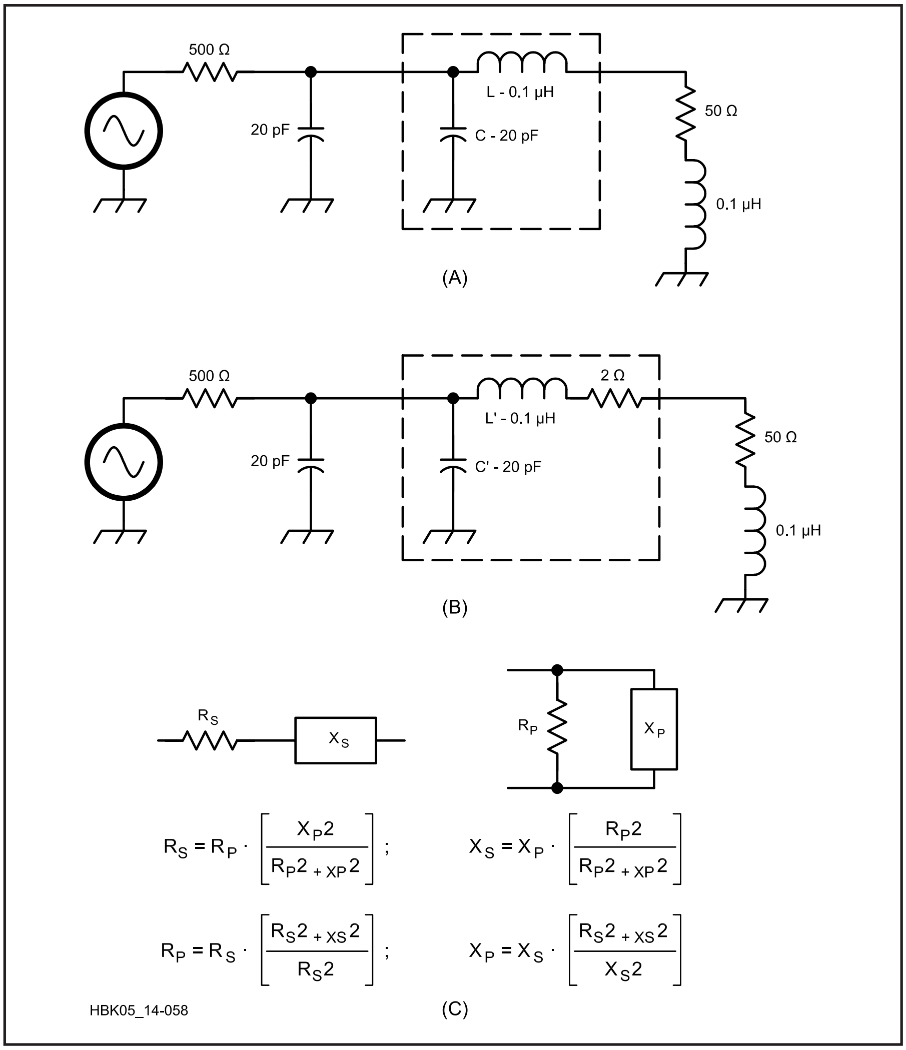


Fig 14.58—At A, impedance transformation, first iteration. At B, second iteration compensates L and C values for coil resistance. At C, series-parallel conversions.

op-amp integrator circuit in the audio amplifier accomplishes this. 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 31).

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.

"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** chapter describes such a system.

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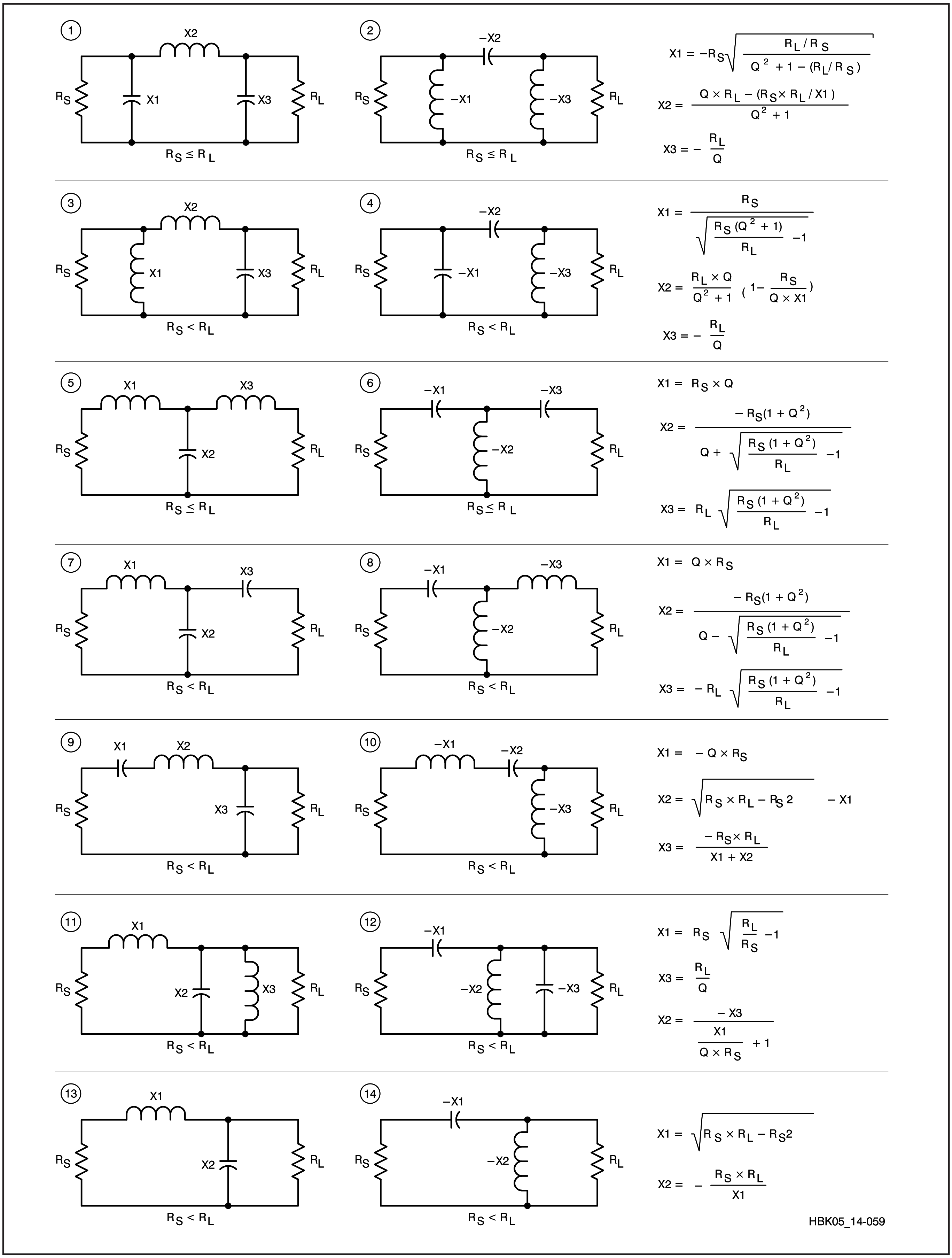

Networks 1 and 2:

$$X1 = -R_S\sqrt{\frac{R_L/R_S}{Q^2+1-(R_L/R_S)}}$$

$$X2 = \frac{Q \times R_L - (R_S \times R_L / X1)}{Q^2+1}$$

$$X3 = -\frac{R_L}{Q}$$

Networks 3 and 4:

$$X1 = \frac{R_S}{\sqrt{\frac{R_S(Q^2+1)}{R_L}-1}}$$

$$X2 = \frac{R_L \times Q}{Q^2+1}\left(1-\frac{R_S}{Q \times X1}\right)$$

$$X3 = -\frac{R_L}{Q}$$

Networks 5 and 6:

$$X1 = R_S \times Q$$

$$X2 = \frac{-R_S(1+Q^2)}{Q+\sqrt{\frac{R_S(1+Q^2)}{R_L}-1}}$$

$$X3 = R_L\sqrt{\frac{R_S(1+Q^2)}{R_L}-1}$$

Networks 7 and 8:

$$X1 = Q \times R_S$$

$$X2 = \frac{-R_S(1+Q^2)}{Q-\sqrt{\frac{R_S(1+Q^2)}{R_L}-1}}$$

$$X3 = -R_L\sqrt{\frac{R_S(1+Q^2)}{R_L}-1}$$

Networks 9 and 10:

$$X1 = -Q \times R_S$$

$$X2 = \sqrt{R_S \times R_L - R_S 2} - X1$$

$$X3 = \frac{-R_S \times R_L}{X1+X2}$$

Networks 11 and 12:

$$X1 = R_S\sqrt{\frac{R_L}{R_S}-1}$$

$$X3 = \frac{R_L}{Q}$$

$$X2 = \frac{-X3}{\frac{X1}{Q \times R_S}+1}$$

Networks 13 and 14:

$$X1 = \sqrt{R_S \times R_L - R_S 2}$$

$$X2 = -\frac{R_S \times R_L}{X1}$$

Fig 14.59—Fourteen impedance transforming networks with their design equations (for lossless components).

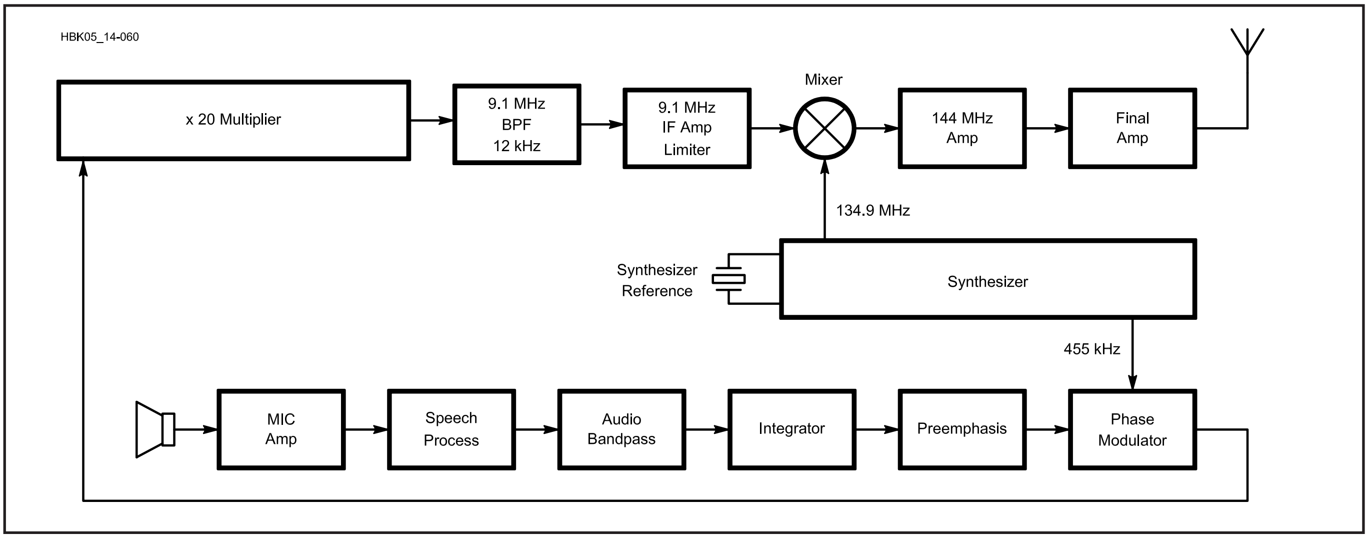

Fig 14.60—Block diagram of a VHF/UHF NBFM transmitter using the indirect FM (phase modulation) method.

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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 14.61**, 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, C13 and C14 are

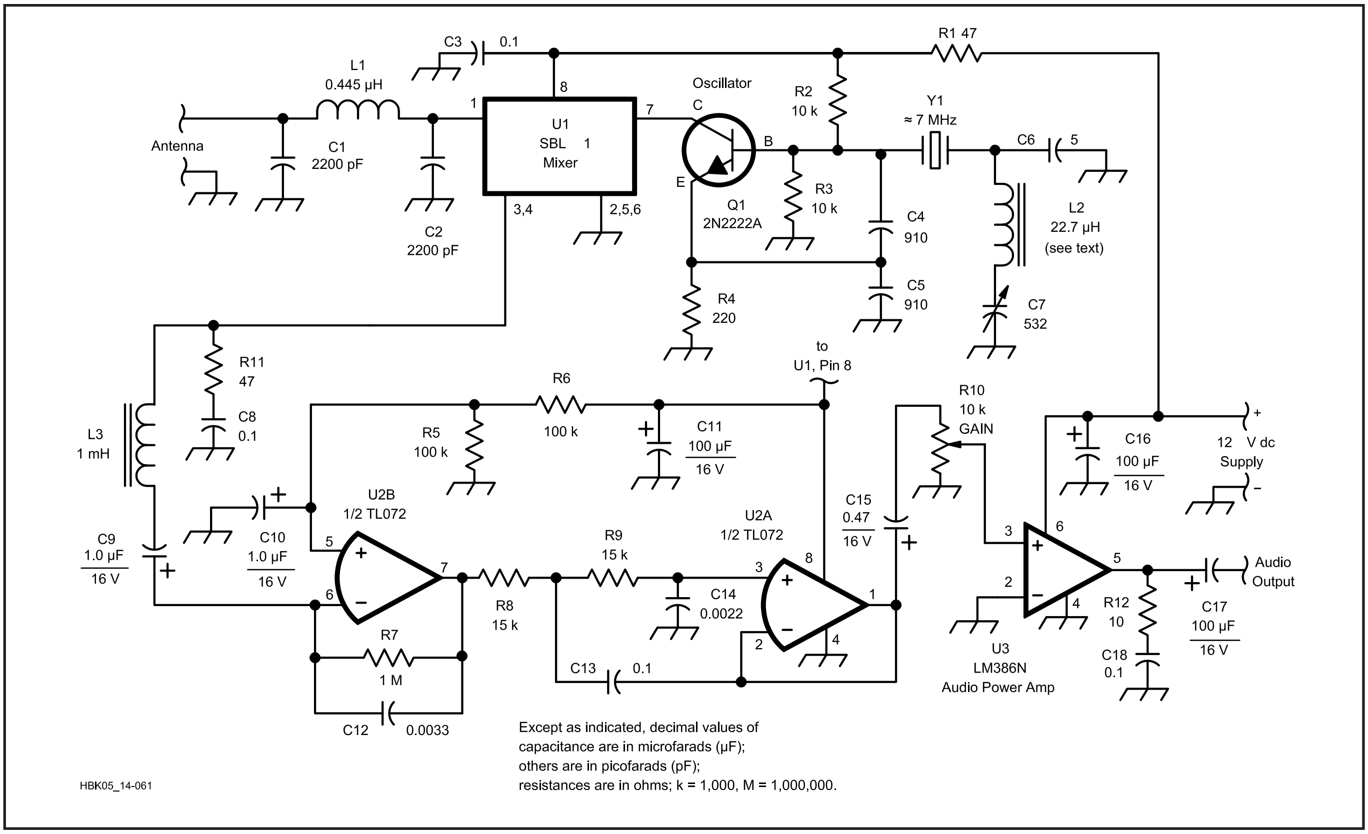


Fig 14.61—An SBL-1 mixer (U1, which contains two small RF transformers and a Schottky-diode quad), a TL072 dual op-amp IC (U2) and an LM386 low-voltage audio power amplifier IC (U3) do much of the Rock-Bending Receiver's magic. Q1, a variable crystal oscillator (VXO), generates a low-power radio signal 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 (800-346-6873, 817-483-4422). 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.
U2—TL072CN or TL082CN dual JFET op amp.
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 (RadioShack No. 271-215 or 271-1721 suitable).
U1—Mini-Circuits SBL-1 mixer.
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.

appropriate for listening to CW signals. If you want SSB stations to sound better, make the changes shown in the caption for Fig 14.61.

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.) See **Fig 14.62** for details on the physical layout of several important components used in the receiver. **Fig 14.63** shows photos of two different receivers using two different approaches to construction—one using a PC board and the other using "ugly" techniques.

If you use a homemade double-sided circuit board based on the PC pattern on the accompanying CD, 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 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

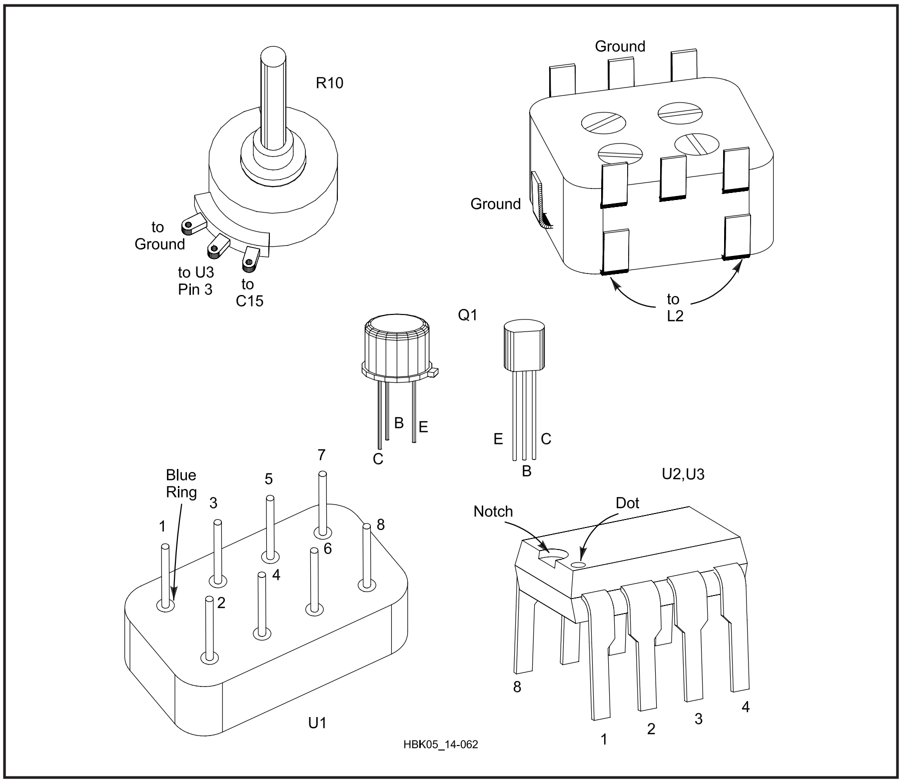
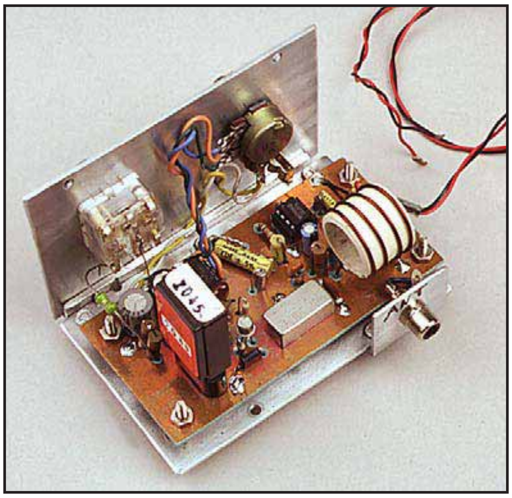


Fig 14.62—The Mouser Electronics part suggested for C7 has terminal connections as shown here. (You can use any variable capacitor with a maximum capacitance of 350 to 600 pF for C7, but its terminal configuration may differ from that shown here.) Two Q1-case styles are shown because plastic or metal 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.



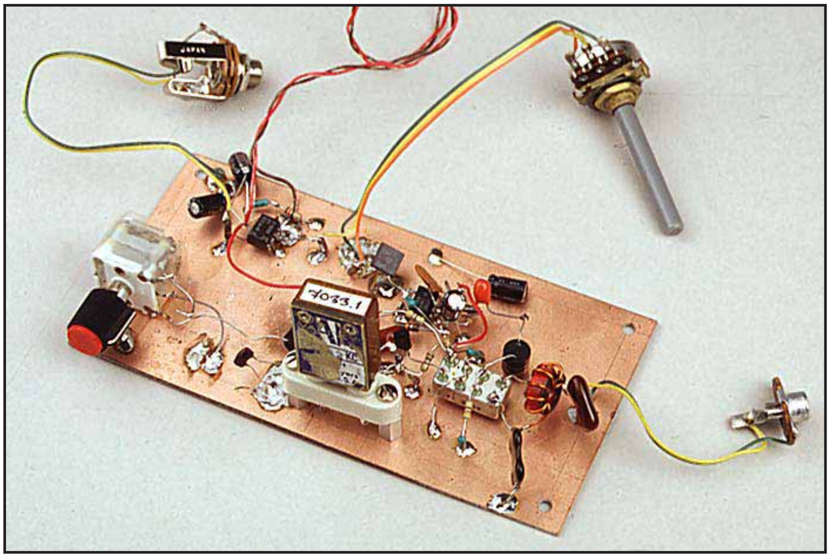


Fig 14.63—Ground-plane construction, PC-board construction—both of these approaches can produce the same good Rock Bending Receiver performance. (WI5W built the one that looks nice, and ex-W9VES—who wrote this caption—built the one that doesn't.)

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 14.64 is the schematic for the preamplifier. C1 and C2 are dc blocking capacitors. The device receives at the output lead, through RF choke L1 and limiting resistor R1. The only other components used are the bypass capacitors on the 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

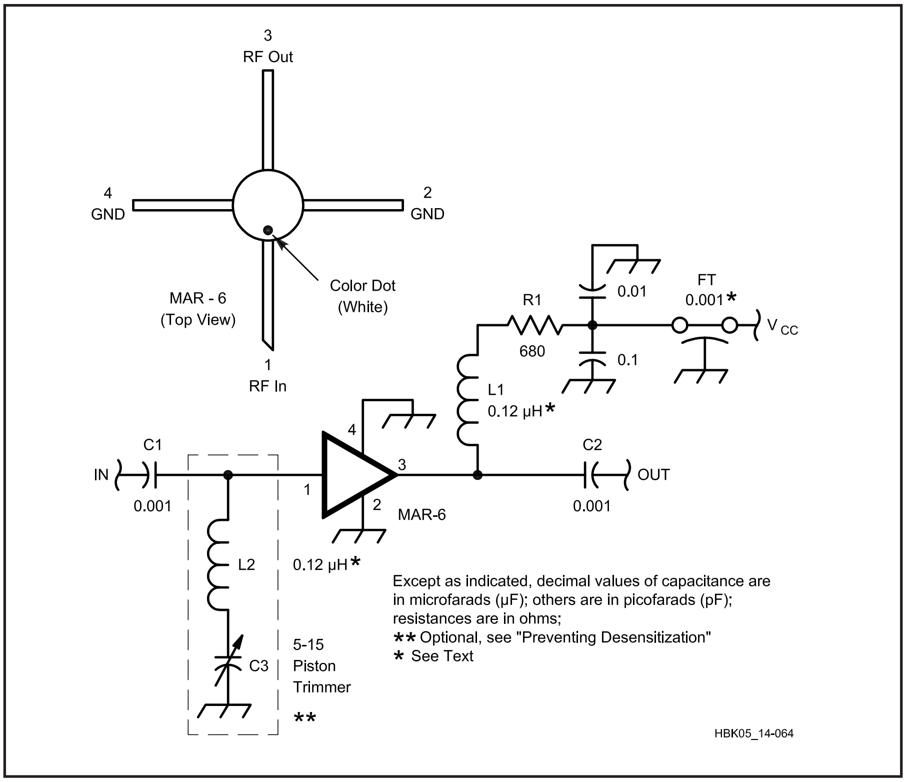


Fig 14.64—A schematic of the preamp circuit. Equivalent parts may be substituted.

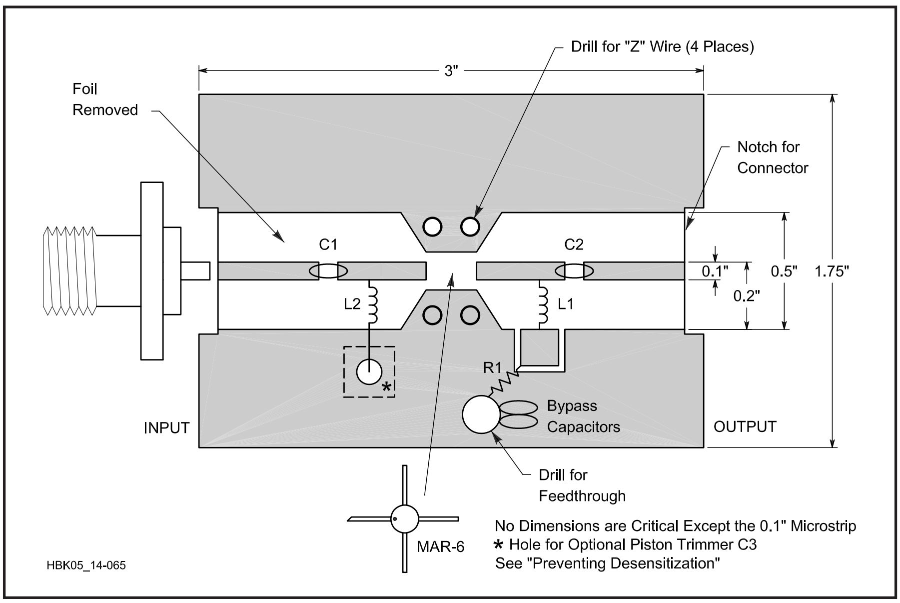


Fig 14.65—A part-placement diagram for KR8L's MMIC preamp. Dark areas are copper on the component (near) side. The reverse is a copper ground plane.

supply provided about 14.6 V, so a 680 Ω, 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 µH 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. 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 14.65 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 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 interference-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] 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. **Fig 14.66** is a photo of the front panel built by KK7B.

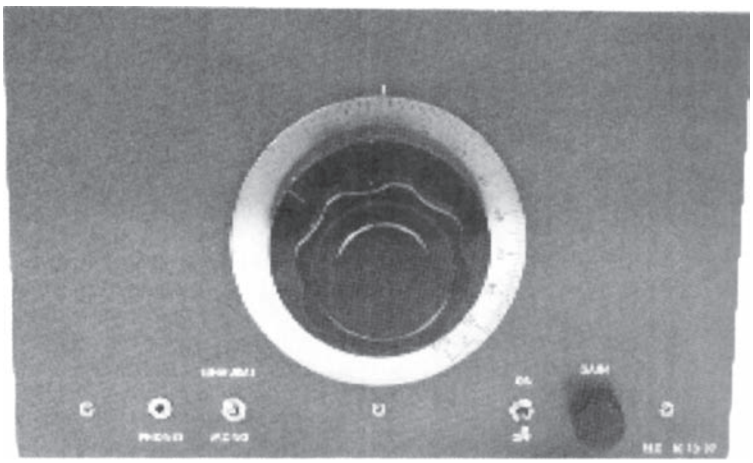


Fig 14.66—A receiver with presence . . . to fully appreciate this receiver, you've got to hear it! "Once my ears got used to the effect, they had to drag me away from this radio. This is one I gotta have!"—*Ed Hare, W1RFI, ARRL Lab Supervisor*

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 14.67**. The incoming signal is split into two paths, then mixed with a pair of local oscillators (LOs) with a relative 90º 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 multi-

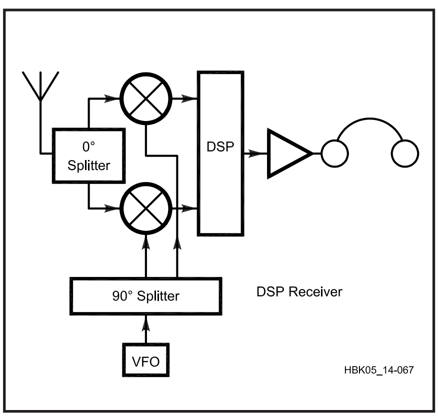


Fig 14.67—The simplified block diagram of a receiver using a DSP detector; see text.

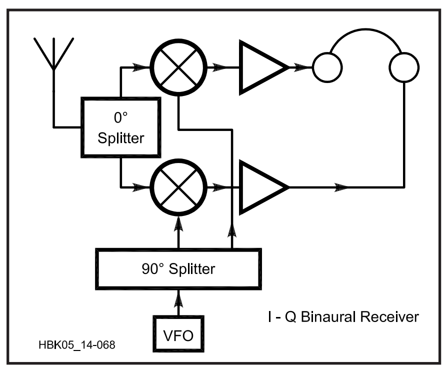


Fig 14.68—The block diagram of a binaural I-Q receiver that allows the ear/brain combination to process the detector output, resulting in stereo-like reception.

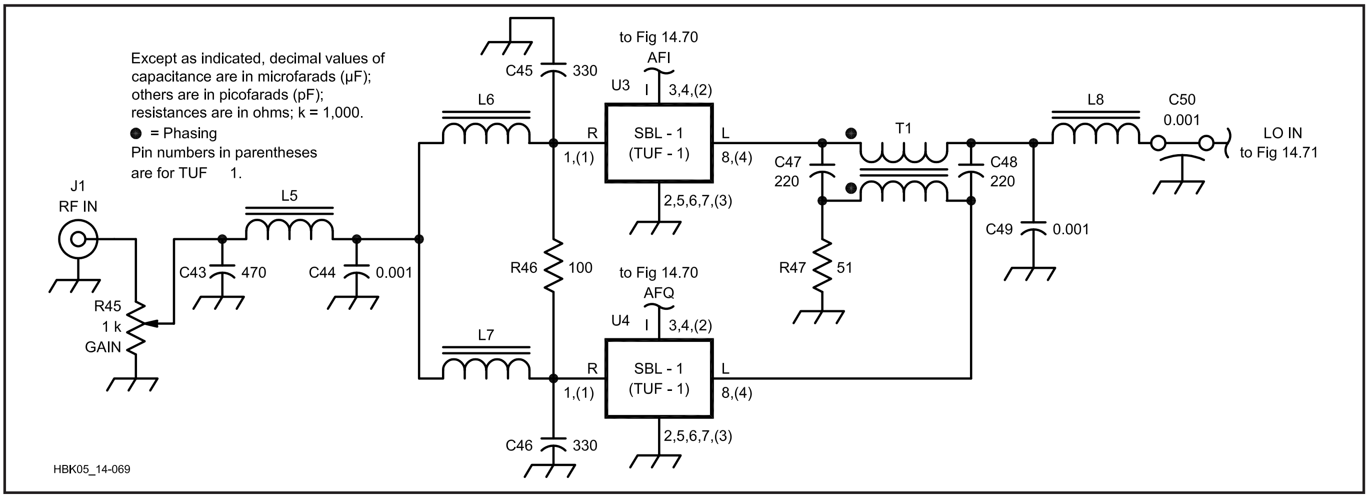


Fig 14.69—This diagram shows the front end and *I* and *Q* demodulators of the Binaural Weekender receiver. Unless otherwise specified, resistors are 1/4 W, 5% tolerance carbon-composition or film units. Equivalent parts can be substituted. Pin connections for the SBL-1 and TUF-1 mixers at U3 and U4 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.

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 µH, 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.

plication. 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 information 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 14.68**.

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 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 14.69, **14.70** and **14.71** show the complete receiver schematic. In Fig 14.69, 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 14.70 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 14.71 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

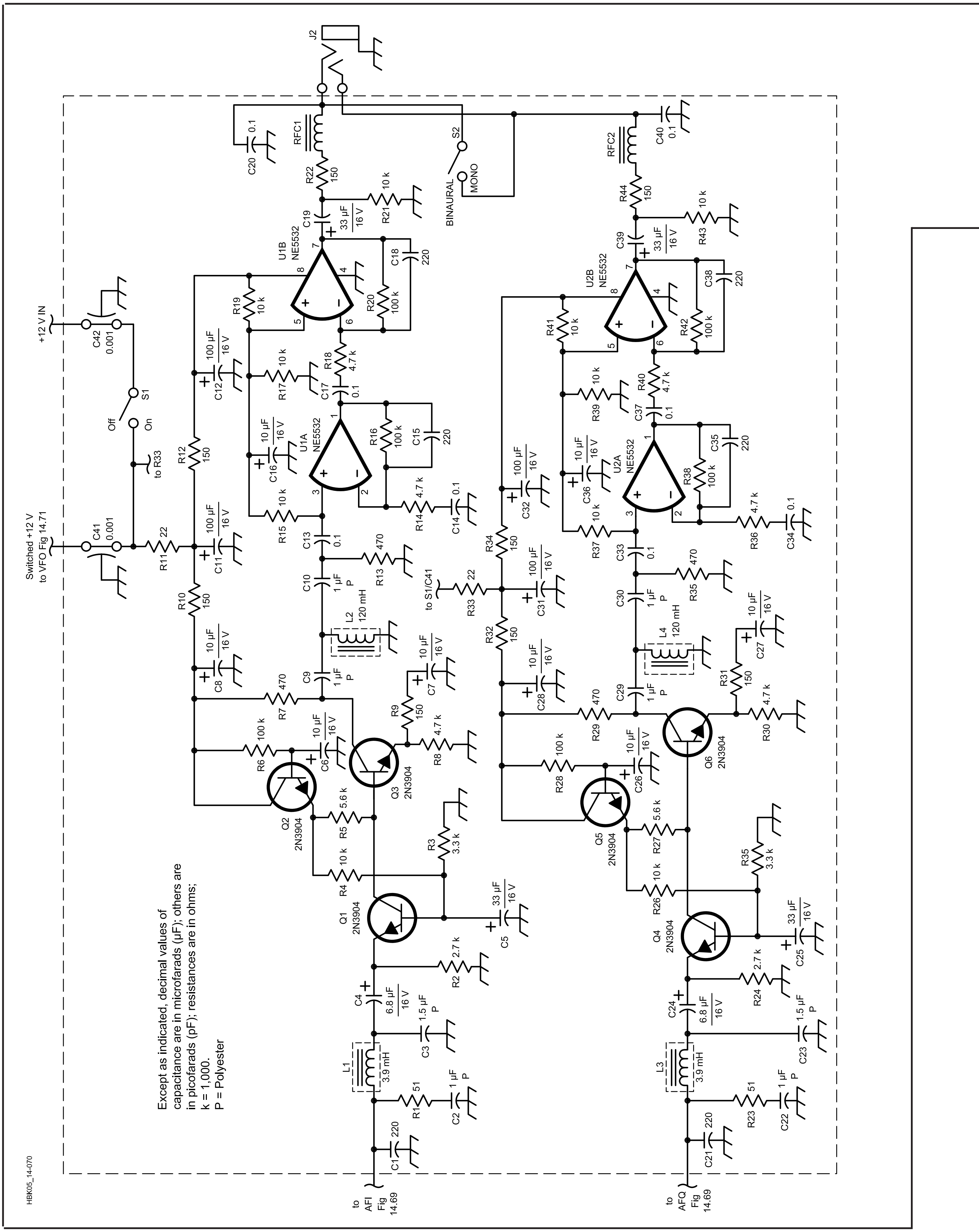

Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000. P = Polyester
HBK05_14-070
+12 V IN
Switched +12 V to VFO Fig 14.71
to S1/C41
to R33
Off
On
S1
S2
BINAURAL
MONO
J2
U1A NE5532
U1B NE5532
U2A NE5532
U2B NE5532
Q1 2N3904
Q2 2N3904
Q3 2N3904
Q4 2N3904
Q5 2N3904
Q6 2N3904
to AFI Fig 14.69
to AFQ Fig 14.69

Fig 14.70—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.

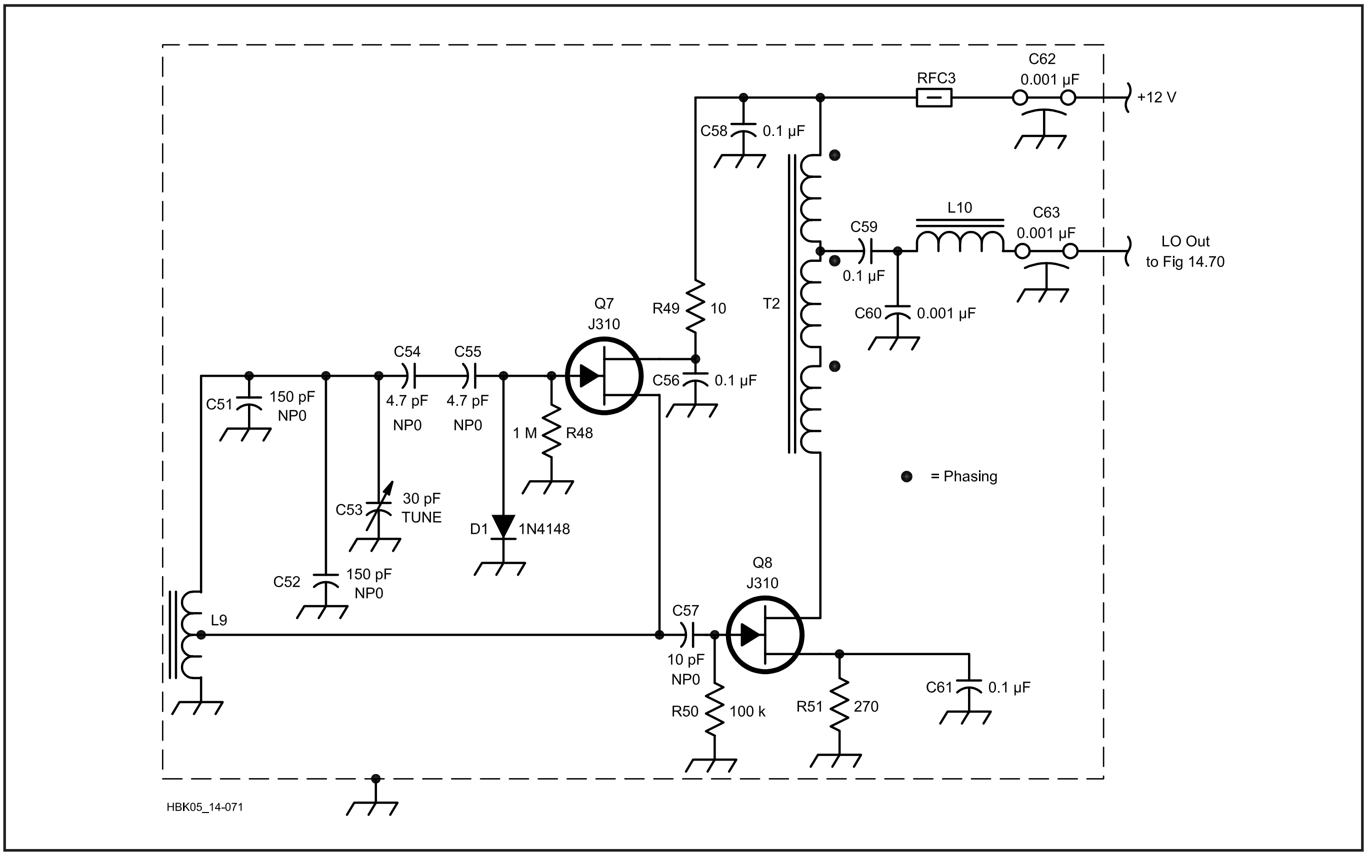


Fig 14.71—The diagram shows the prototype binaural receiver's VFO. The LO output is +10 dBm. This simple VFO works exceptionally well, but must be completely shielded for good D-C receiver performance. A receiver with an open PC-board VFO will work better if the variable oscillator is not running on the received frequency. As noted elsewhere, the kit version of the receiver uses a different VFO.

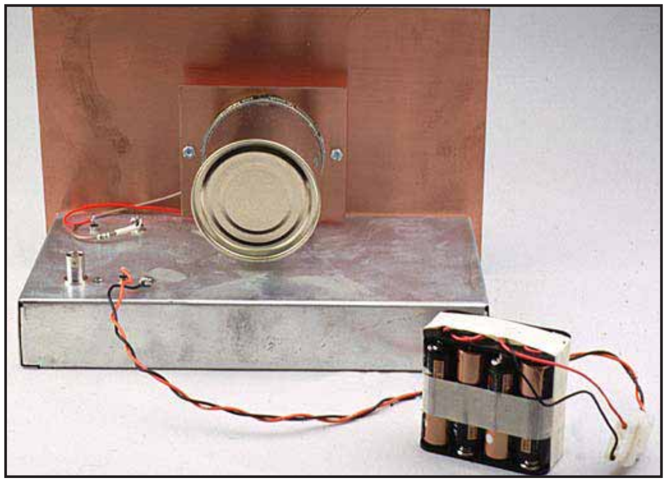
C51, C52—150 pF, NP0 disc ceramic.
C53—30 pF air-dielectric variable.
C54, C55—4.7 pF NP0 disc ceramic.
C56, C57, C59, C61—0.1 µF metal polyester (Panasonic V series).
C57—10 pF NP0 disc ceramic.
C60—0.001 pF metal polyester.
C62, C63—0.001 µF feedthrough capacitor.
D1—1N4148.
L9—1.5 µH, 22 turns #22 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).

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

Fig 14.66 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

(A)



(B)

Fig 14.72—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 can live again as a VFO shield in the prototype receiver.

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* much like the STEREO-MONO switch on an FM broadcast receiver—given the choice, it always ends up in the STEREO position!

The author 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 14.72** 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 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.

Notes

[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.

[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.

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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* V*HF Society,* Cedar Rapids, IA, July 1984, pp 27-33.

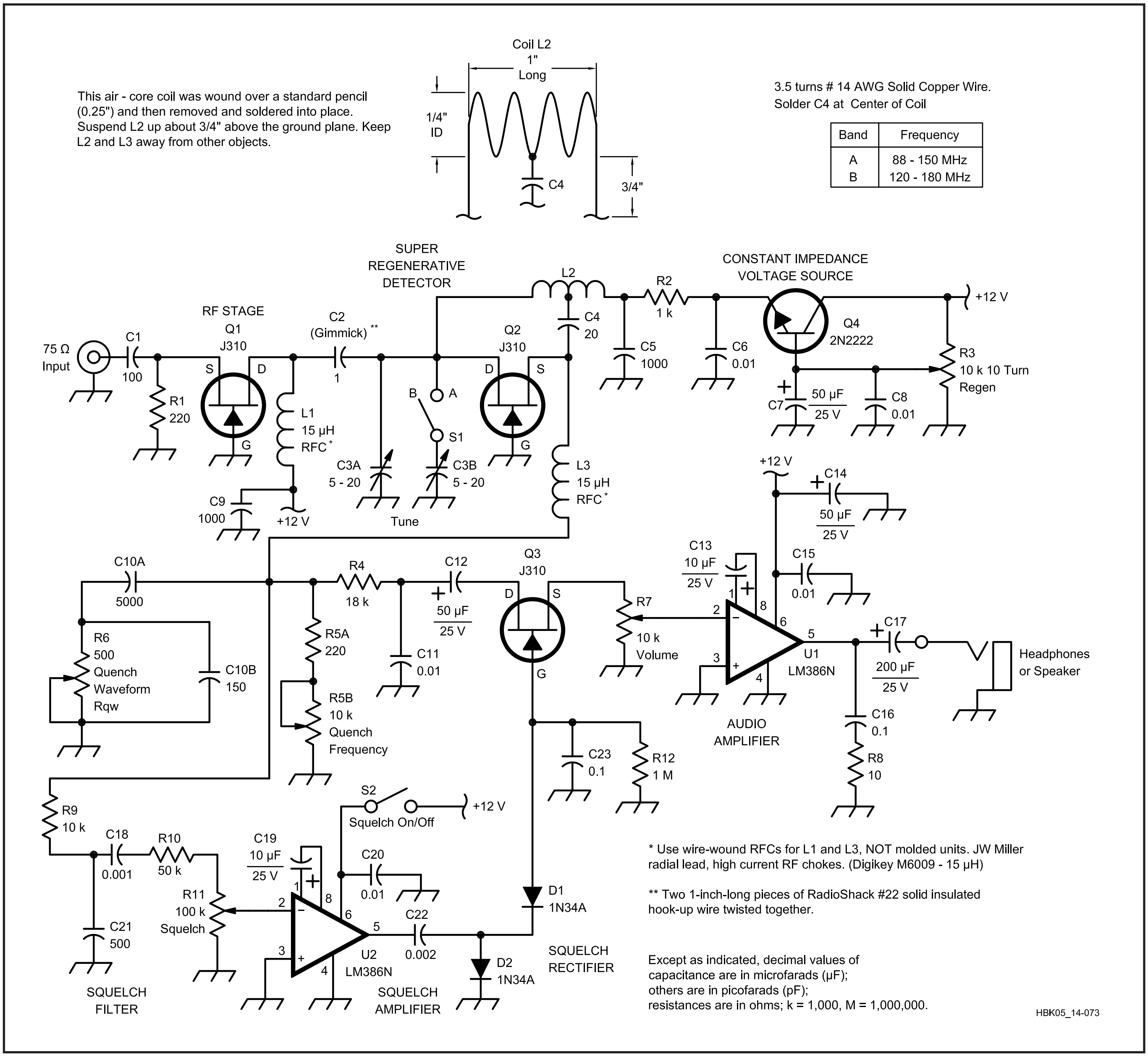


Fig 14.73—A Two-Band VHF receiver with squelch.

A SUPERREGENERATIVE VHF RECEIVER WITH SQUELCH

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 88 to 180 MHz in two bands. This covers the FM broadcast band, the aircraft band and the 2-meter ham bands. Above 2 meters, it also receives fire, police, marine and weather frequencies. The circuit draws about 20 mA.

Don’t expect to squeeze out the ultimate in selectivity or stability with this receiver, although it does provide a sensitivity of around 0.5 µV. It uses the principle of superregeneration for this high sensitivity with a low parts count.

For narrow bandwidth FM reception, this receiver does require careful adjustment of several controls as it is tuned. A low-cost PC board is available from FAR Circuits, which greatly simplifies construction and helps prevent any layout or wiring errors.

Circuit Description

Fig 14.73 shows the circuit. RF signals from the antenna enter via a 75Ω coaxial cable and are ac coupled to the source of JFET Q1. This untuned RF amplifier provides excellent stability, good output-to-input isolation, and modest voltage gain over a wide frequency range. It also has a low input impedance (to match the coax line) and a high output impedance, which minimizes loading on the detector. R1 provides protective dc bias for the JFET. L1 is an RF choke, which extracts the amplified RF signal from the JFET drain.

Q2 operates as a superregenerative detector in a modified Hartley oscillator configuration. Capacitor C3 and coil L2 set the tuning (and detector oscillation) frequency. L2 should be stretched so that it is about one inch in length. After the circuit has been built and the detector is operating correctly, the turns on L2 can be compressed or expanded to raise or lower the tuning range. A gimmick capacitor, C2, couples the signal from the RF stage to the detector. Its value is somewhere around 1 pF. The gimmick lightly couples signals into the detector. Any over-coupling would reduce selectivity and might create "dead spots" in the tuning range of the receiver.

Band switching is accomplished simply and easily, by using S1 to switch-in either one or two gangs of tuning capacitor C3. S1 was wired directly between the two "hot" terminals of C3 using two *very* short lengths of #14 AWG copper wire. These support the switch quite well. With this arrangement, it is necessary to build the receiver with an open top, so you can reach in to change bands.

The value of C3 is not critical. A two or three-gang capacitor salvaged from an old FM radio will work nicely. Other small variable capacitors may be substituted, as long as their maximum capacity is not too great. Small mica capacitors may be wired in series or parallel with C3, or turns may be added or subtracted from L2, to change the tuning range. Capacitors C4 and C5 should be mica (or NP0 ceramic), as they need to be low-drift, high-Q devices.

The RC components on the Q2 source set up a secondary relaxation oscillation that self-quenches the receiver at a super audio rate. Resistor R6 slows the build-up and decay of this oscillation and modifies its waveform into a sine wave for high selectivity. Q4 reduces any variations in quench waveform shape as the regeneration control is varied. The audio output from Q2 is low pass filtered by R4 and C11. This prevents the quenching oscillations from reaching the audio stage and also improves the audio quality. The filtered audio signal is ac coupled to Q3 and then on to the volume control and LM386 audio amplifier.

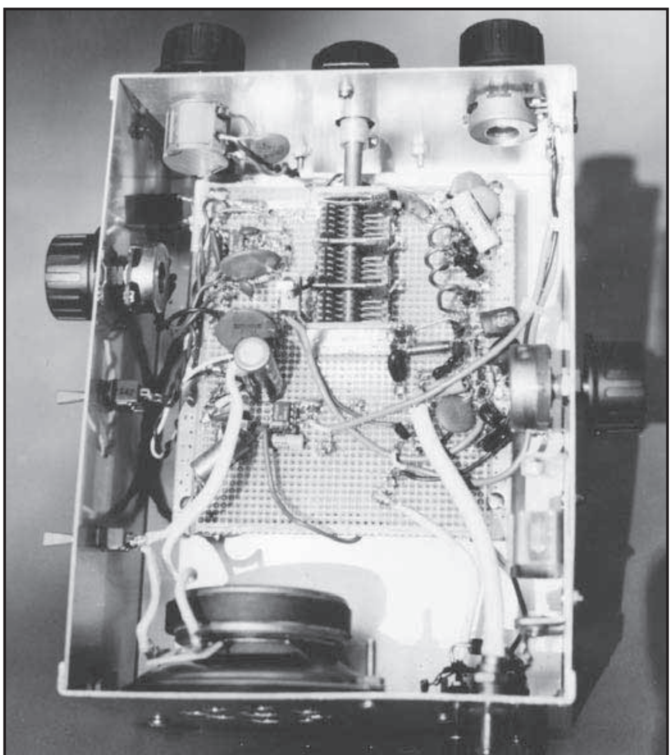
A second filter circuit consisting of R9, C21 and C18, R10 passes mainly the upper audio frequencies to amplifier U2. This drives diodes D1 and D2, and the negative voltage output then squelches the receiver, by turning off Q3 during no-signal periods.

Construction Guidelines

Be sure to solder all ground leads of the FAR Circuits' board to both the top and bottom of the board, because these are not plated-through holes. When using a hand-wired board, it is essential that the drains of Q1 and Q2 be about one-quarter inch apart, with the gimmick capacitor connected between them. All wiring should use the shortest leads possible. The use of a rigid metal enclosure greatly improves the receiver's frequency stability. A vernier dial and a 10-turn potentiometer are recommended for easy TUNING and REGEN control. The metal ground plane of the PC board is connected to the metal enclosure using a single short wire. Use shielded wiring for all controls. Mount C3 directly on the PC board ground plane, then pass its shaft through an oversized hole in the metal front panel, without touching the panel. Use plastic knobs for the controls.

Testing and Operation

Test the audio stage first; then check the detector and RF stage. Check the audio by placing your finger on the wiper of R7. To test the detector, leave out C2. With the SQUELCH turned-off, set R6 to mid-position, turn the REGEN level up high and adjust R5B until the receiver oscillates. You should hear a loud "rushing" noise. Check that oscillation occurs over the entire tuning range adjusting R5B and R6 as needed. If there are any dead spots, move L2 and L3 farther away from all other objects. Then insert C2 and twist the



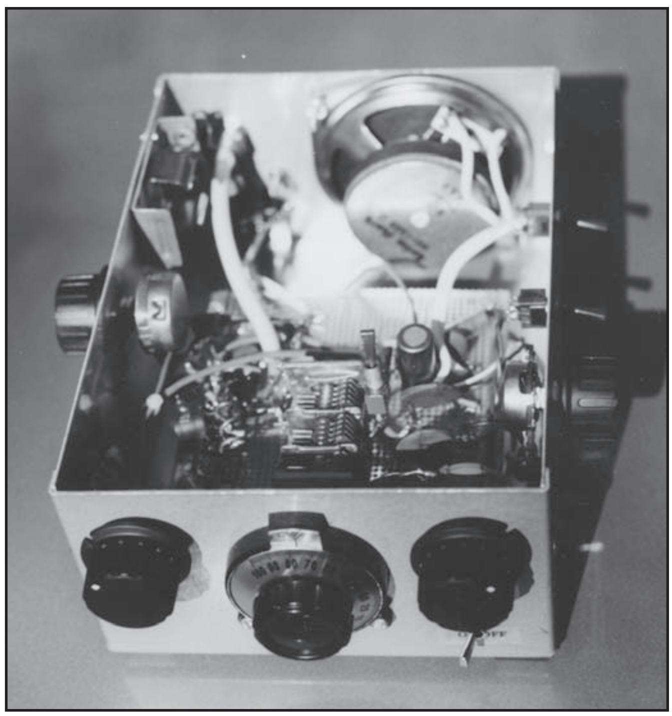


Fig 14.74—Photos of the completed Two-Band VHF receiver.

turns together as much as possible without affecting detector operation.

Turn the REGEN up high on the 88-108 MHz FM broadcast band (wide-band FM mode). With QUENCH WAVEFORM (R6) set to 0 Ω; adjust the REGEN control for minimum distortion. Use the same settings for the 118-136 MHz aircraft band (AM mode) and set the SQUELCH so that the no-signal background noise is just muted. For operation on the 2-meter band (NBFM mode), increase R6 to about midrange and set REGEN fairly high. Tune-in a station, reduce the REGEN level until the audio level increases, and then carefully adjust TUNING and REGEN for maximum volume. Higher R6 resistance will be needed when receiving very strong signals. When receiving NBFM, headphone level is more than adequate but speaker volume will be low. Additional audio gain can be added between R7 and U1, if desired.

This receiver can be modified to cover the 6-meter band by changing a few components. Use a 50-pF air variable capacitor for C3, with a 40-pF mica capacitor in parallel. Increase L1 and L3 to 33 µH, L2 to 7 turns, C10A to 7000 pF, C4 to 50 pF; C5 to 2000 pF, omit C10B, twist C2 more tightly to approximately 2 pF.

For further reading, see Kitchin "New Super Regenerative Circuits for Amateur VHF and UHF Experimentation," September/October 2000 *QEX*. **Fig 14.74** shows photos of a finished receiver.

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 on the accompanying CD.

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_{DSon} (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 oscil-

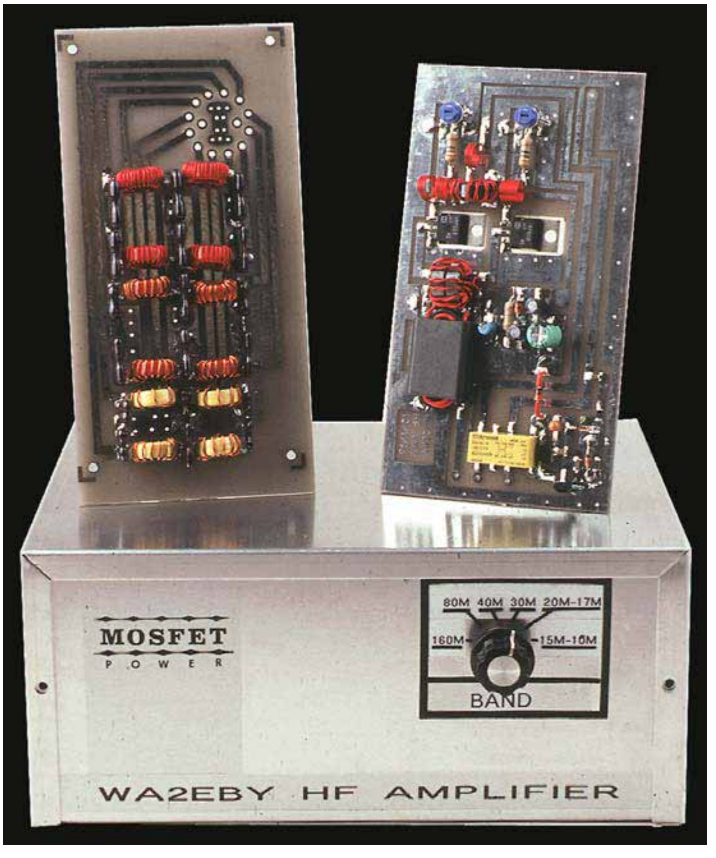


Fig 14.75—The WA2EBY MOSFET HF Amplifier produces over 40 W output with 1 W of input power.

lation 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.

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. The author recommends 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. A photo of the WA2EBY amplifier is shown in **Fig 14.75**.

Considerable experimentation with several designs resulted in the circuit shown in **Fig 14.76**. This amplifier consists of two power MOSFETs operating in push-pull, and employs an RF-sensed TR relay.

During receive, TR relay K1 is de-energized. 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 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.

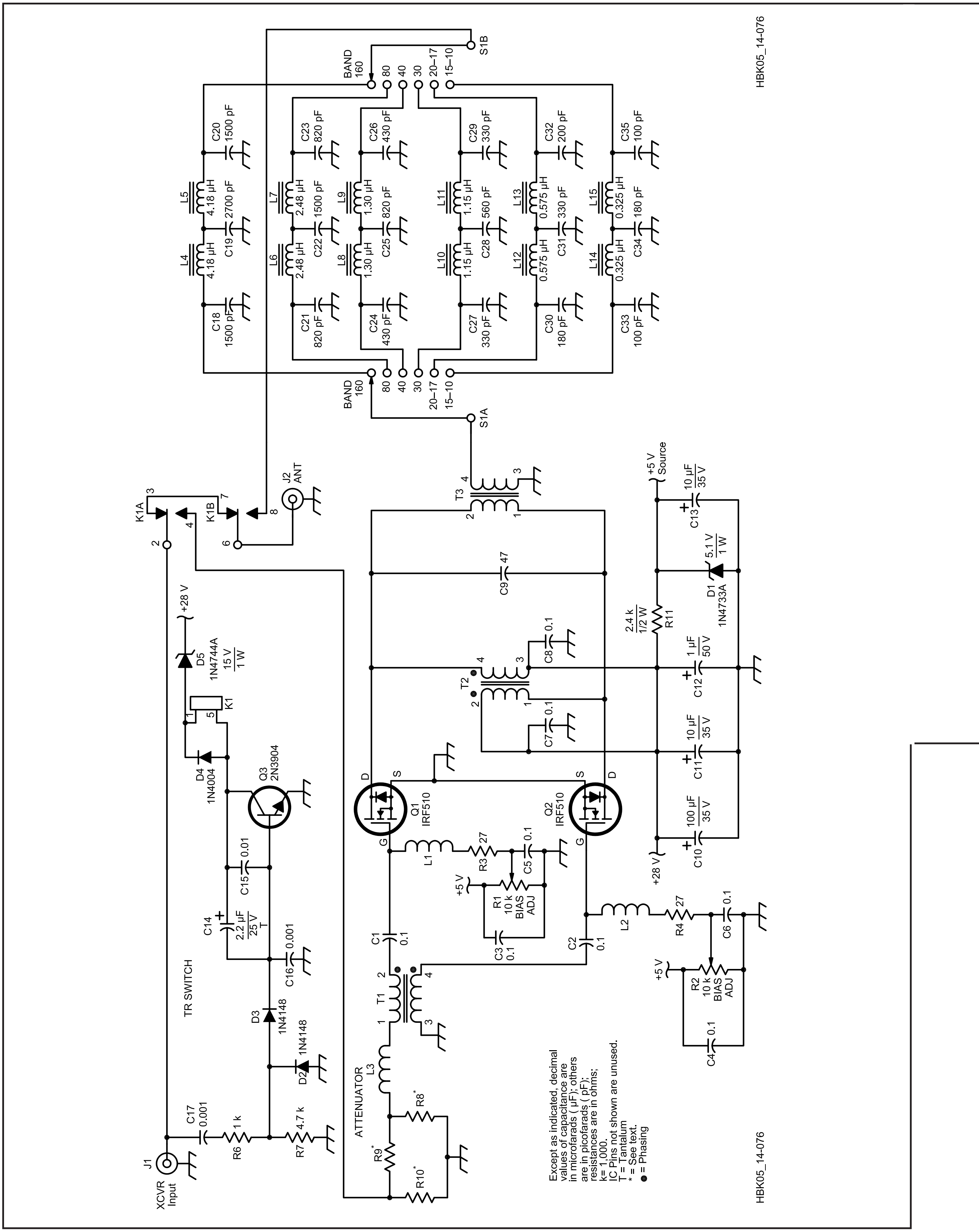
HARMONIC FILTERING

Although biased for class AB linear operation, this amplifier (like others of its type) exhibits some degree of non-linearity, 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 14.76.

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. 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



HBK05_14-076
S1B
BAND
160
80
40
30
20–17
15–10
C20 1500 pF
C23 820 pF
C26 430 pF
C29 330 pF
C32 200 pF
C35 100 pF
L5 4.18 µH
C19 2700 pF
L4 4.18 µH
L7 2.48 µH
C22 1500 pF
L6 2.48 µH
L9 1.30 µH
C25 820 pF
L8 1.30 µH
L11 1.15 µH
C28 560 pF
L10 1.15 µH
L13 0.575 µH
C31 330 pF
L12 0.575 µH
L15 0.325 µH
C34 180 pF
L14 0.325 µH
C18 1500 pF
C21 820 pF
C24 430 pF
C27 330 pF
C30 180 pF
C33 100 pF
S1A
K1A
K1B
J2 ANT
T3
C9 47
+5 V Source
C13 10 µF 35 V
D1 1N4733A 5.1 V 1 W
R11 2.4 k 1/2 W
+28 V
D5 1N4744A 15 V 1 W
K1
T2
C8 0.1
C12 1 µF 50 V
C7 0.1
C11 10 µF 35 V
C10 100 µF 35 V
D4 1N4004
Q3 2N3904
Q1 IRF510
Q2 IRF510
L1
R3 27
C5 0.1
R1 10 k BIAS ADJ
C3 0.1
C15 0.01
C14 2.2 µF 25 V T
C16 0.001
C1 0.1
C2 0.1
L2
R4 27
C6 0.1
R2 10 k BIAS ADJ
C4 0.1
TR SWITCH
D3 1N4148
D2 1N4148
T1
ATTENUATOR
L3
R8*
R9*
R10*
C17 0.001
R6 1 k
R7 4.7 k
J1
XCVR Input
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k= 1,000.
IC Pins not shown are unused.
T = Tantalum
* = See text.
● = Phasing
HBK05_14-076

Fig 14.76—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 Ω 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.

Fig 14.77—This rear-panel view of the amplifier shows the heat sink. The filter board mounts on the back side of the front panel.

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 **Fig 14.77**. 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 #4-40 mounting 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 14.1**. 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 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.

Table 14.1
Low-Pass Filter Inductor Winding Information

(Refer to Fig 14.76)

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.

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 in-

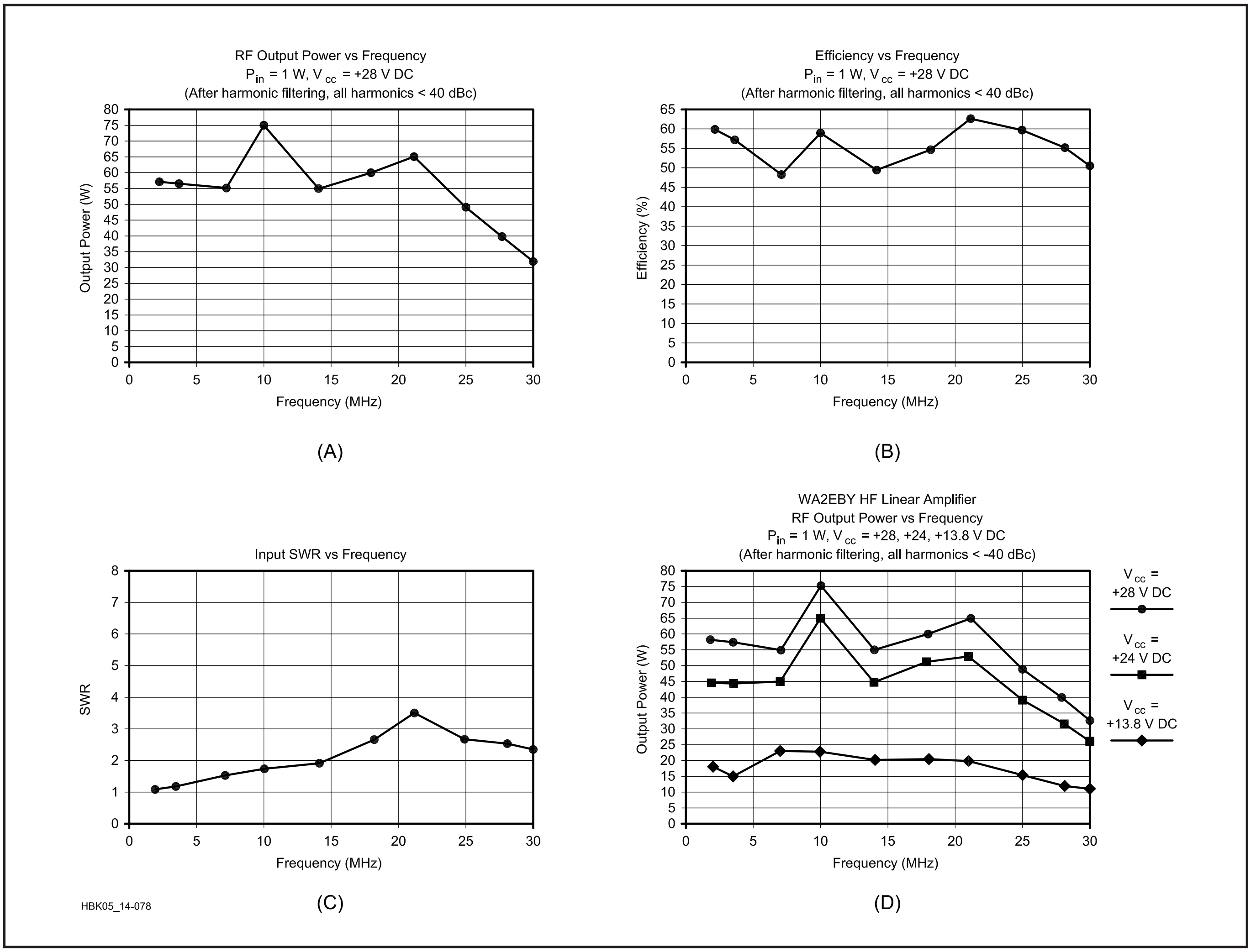


Fig 14.78—A shows the amplifier RF output power. B shows the amplifier efficiency. C shows the amplifier input SWR. D shows the amplifier RF output power versus supply voltage.

ductors. 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 14.78A**.

As shown in Fig 14.78B, this amplifier achieves an efficiency of better than 50% over its frequency range, except at 7 MHz where the efficiency drops to 48%.

Fig 14.78C 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 14.76) at the input and supplying 2 W to the pad input. This keeps the amplifier drive at 1 W.

Fig 14.78D 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.

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

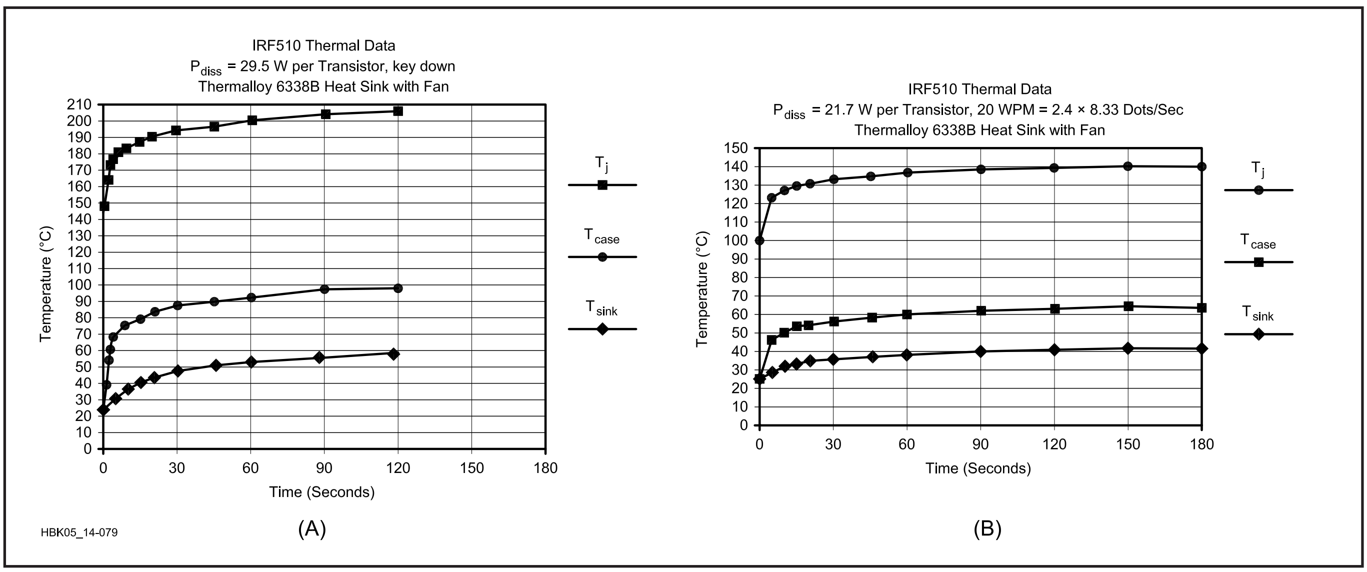


Fig 14.79—A is a graph of the amplifier thermal performance during key-down conditions. B is a graph of the amplifier thermal performance during simulated CW conditions.

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 14.79A**. Data was taken under dc operating conditions with power-dissipation levels set equal to conditions under RF operation. A Radio Shack 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 14.79B). 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 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 2W-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, and 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 14.76.) 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. Amidon, Inc provides parts kits for this project (see Note 9).

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 Experi-mental 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]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.

[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.

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 14.80** 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] This 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.

VFO CIRCUIT DESCRIPTION

The VFO is a tapped-coil Hartley oscillator that is optimized for low phase noise (see Fig 14.80B). 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.

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.

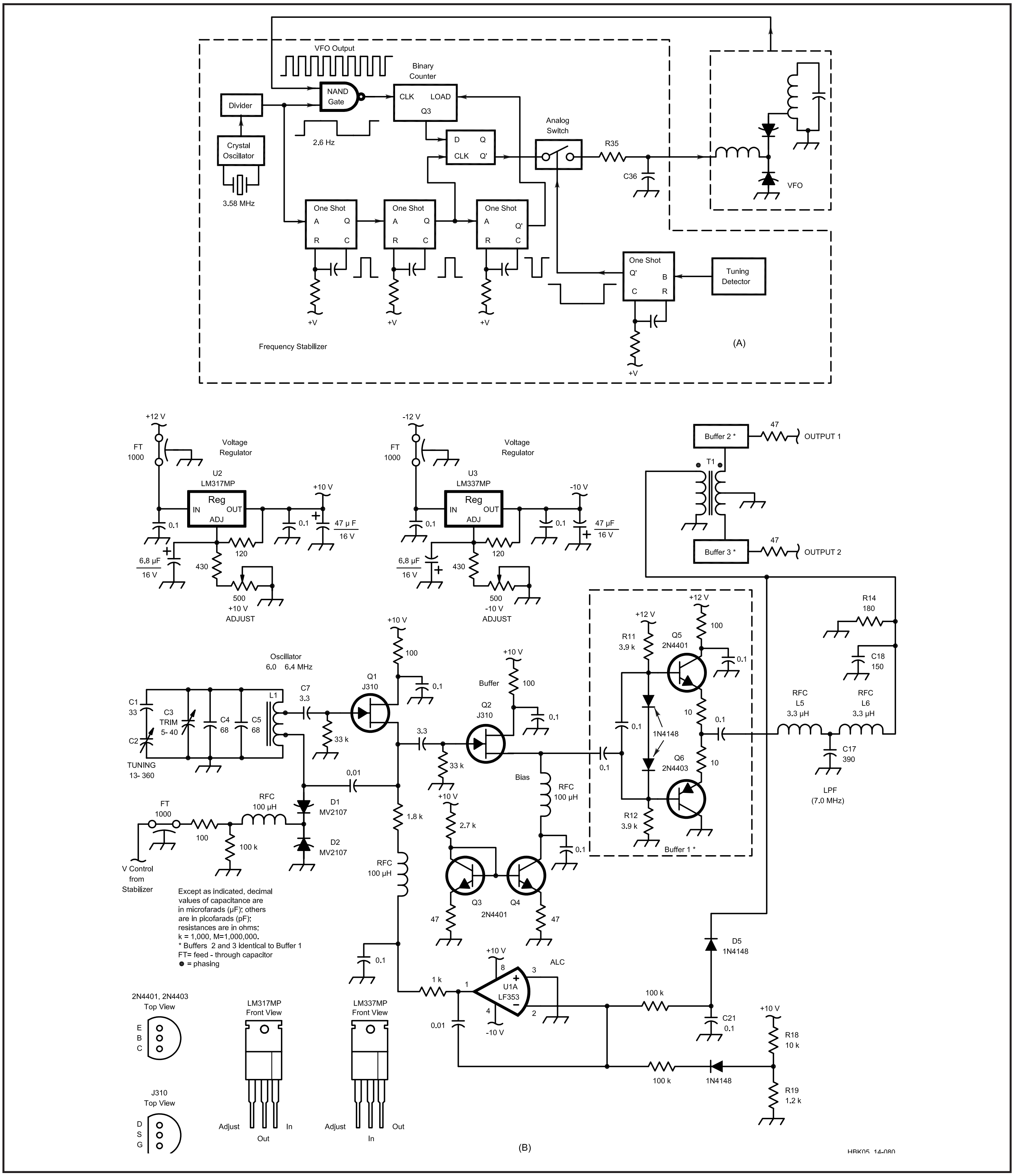


Fig 14.80—At A—Block diagram of the frequency stabilizer and VFO connections. At B—VFO schematic. Buffers 2 and 3 are identical to Buffer 1. Most of the parts are available from Mouser Electronics, Digi-Key Corporation or Allied Electronics. The cores for L1 and T1 are from Amidon Associates. Use 1/4-W, 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).

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 14.81**) 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 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 14.82**) 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 retriggerable—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.

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 µF 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.

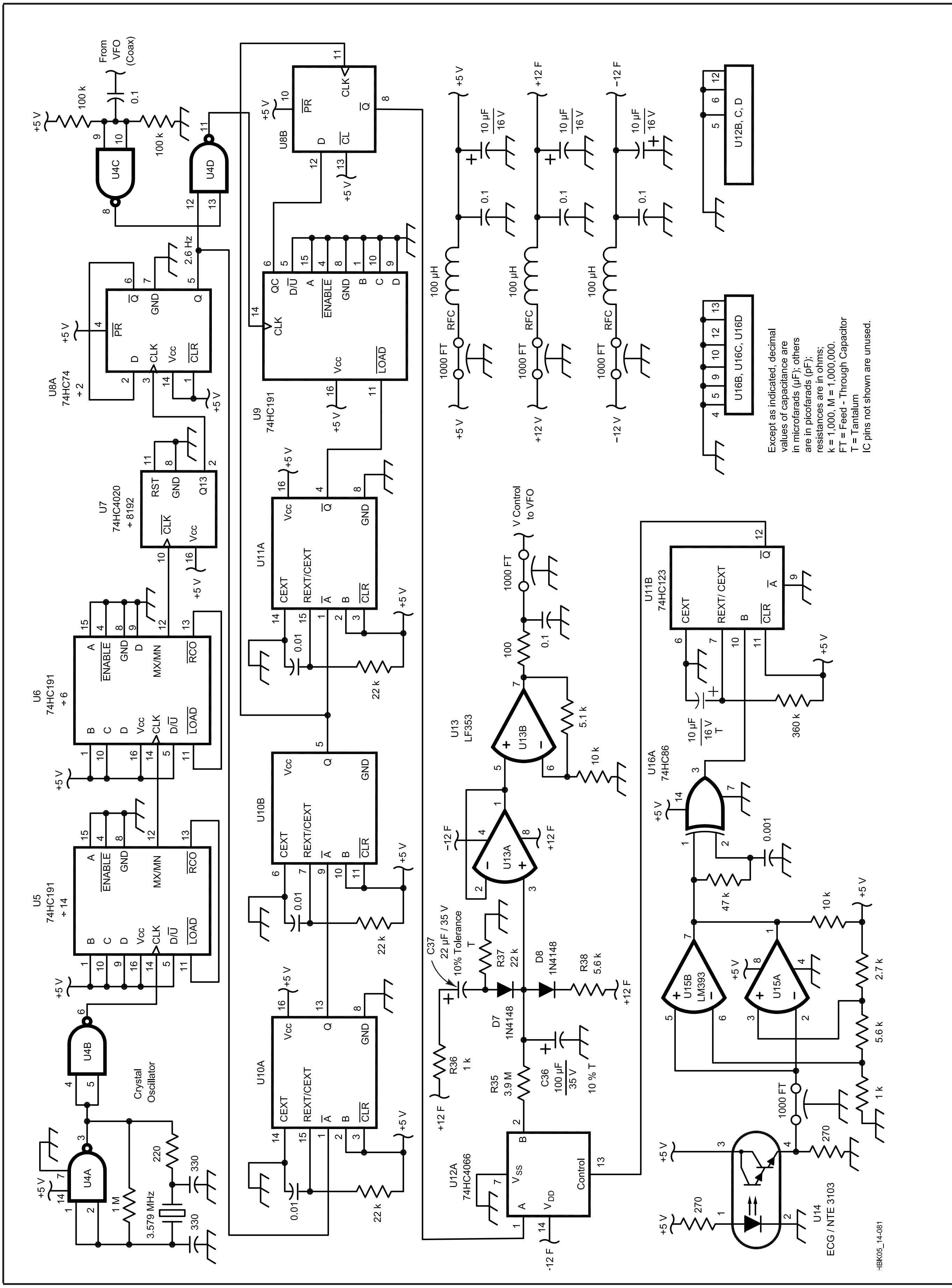

From VFO (Coax)
Crystal Oscillator
3.579 MHz
U4A
U4B
U4C
U4D
U5 74HC191 ÷ 14
U6 74HC191 ÷ 6
U7 74HC4020 ÷ 8192
U8A 74HC74 ÷ 2
U8B
U9 74HC191
U10A
U10B
U11A
U11B 74HC123
U12A 74HC4066
U13 LF353
U13A
U13B
U14 ECG / NTE 3103
U15A
U15B LM393
U16A 74HC86
U12B, C, D
U16B, U16C, U16D
2.6 Hz
V Control to VFO
C37 22 µF / 35 V 10% Tolerance T
R37 22 k
D8 1N4148
R38 5.6 k
D7 1N4148
R36 1 k
R35 3.9 M
C36 100 µF 35 V 10 % T
RFC 100 µH
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.
HBK05_14-081

Figure 14.81—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).

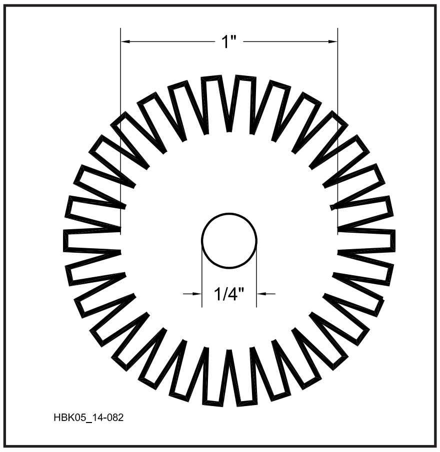


Figure 14.82—Mechanical details of interrupter wheel. Use a good reduction drive and make one tooth for every 20 to 40 Hz of frequency change. Use any rigid, opaque material.

A SIMPLE REGENERATIVE HF RECEIVER FOR BEGINNERS

This project was designed by Charles Kitchin, N1TEV, and originally published in the September 2000 issue of *QST*. It features a low cost, portable shortwave receiver that's ideal for a Scout radio merit badge or for learning basic radio theory. This project is fun to build and easy to get working. This little radio is a fun way to discover ham-band QSOs, news, music and all the other things the shortwave bands have to offer. With it, you can receive dozens of international shortwave broadcast stations at night. Although this little receiver is quite sensitive, it naturally won't match the performance of a commercial HF rig. If you have not used a regenerative receiver before, you'll have to practice adjusting the controls, but that's part of the adventure. Many of today's experienced "homebrewers" got their start by building simple circuits just like this one. You'll gain experience in winding a coil and following a schematic. As your interest in radio communication develops, you can build a more complex receiver later.

This little set requires only a single hand-wound coil and consumes just 5 mA from a 9-V battery. At that rate, an alkaline battery can provide approximately 40 hours of operation. The sound quality of this receiver is excellent when using Walkman headphones. The radio can also drive a small speaker. To simplify construction, a low-cost PC board is available from FAR Circuits.[1] You can house the receiver in a readily available RadioShack plastic project box.

CIRCUIT DESCRIPTION

Fig 14.83 shows the schematic. L1 and C1 tune the input signal from the antenna. Regenerative RF amplifier Q1 operates as grounded-base Hartley oscillator. Its positive feedback provides a signal amplification of around 100,000. The very low operating power of this stage, only 30 μW, makes this receiver very portable and prevents interference to other sets in the area. R2 controls the amount of positive feedback (regeneration). D1 and C4 comprise a floating detector that provides high sensitivity with little loading on Q1. The relatively low back-resistance of the 1N34 germanium diode (don't use a silicon diode here!) provides the necessary dc return path for the detector.

VOLUME control R5 sets the level of detected audio driving U1, an LM386 audio amplifier. C5 provides low-pass filtering that keeps RF out of the audio amplifier. R4 isolates the low-pass filter from the detector circuit when the volume control is at the top of its range. The bottom of the VOLUME control, R5, and pin 3 of the LM386 float above ground, so that both inputs of the IC are coupled. This allows the use of a 100-kΩ VOLUME control; this high resistance value prevents excessive loading of the detector. D5 protects the receiver from an incorrectly connected battery. L1 is wound on a coil form using a standard 35-mm plastic film can or 1-inch-diameter pill bottle. C1 can be any air-dielectric variable capacitor with a maximum capacitance of 100 to 365 pF. A two ganged capacitor from an old AM radio will work well. Total frequency coverage varies with the capacitance value used, but any capacitor in that range should cover the 40-meter ham band and several international broadcast bands. If you use a capacitor with a max. capacity larger than 150 pF, the receiver will cover a very wide frequency range but it will be more difficult to tune-in an individual station. In that event, the optional fine-tuning control (see the inset of Fig 14.83) is recommended. **Figs 14.84** and **14.85** show the construction details.

D6 functions as a poor man's Varactor (voltage-variable capacitor). As the volt-

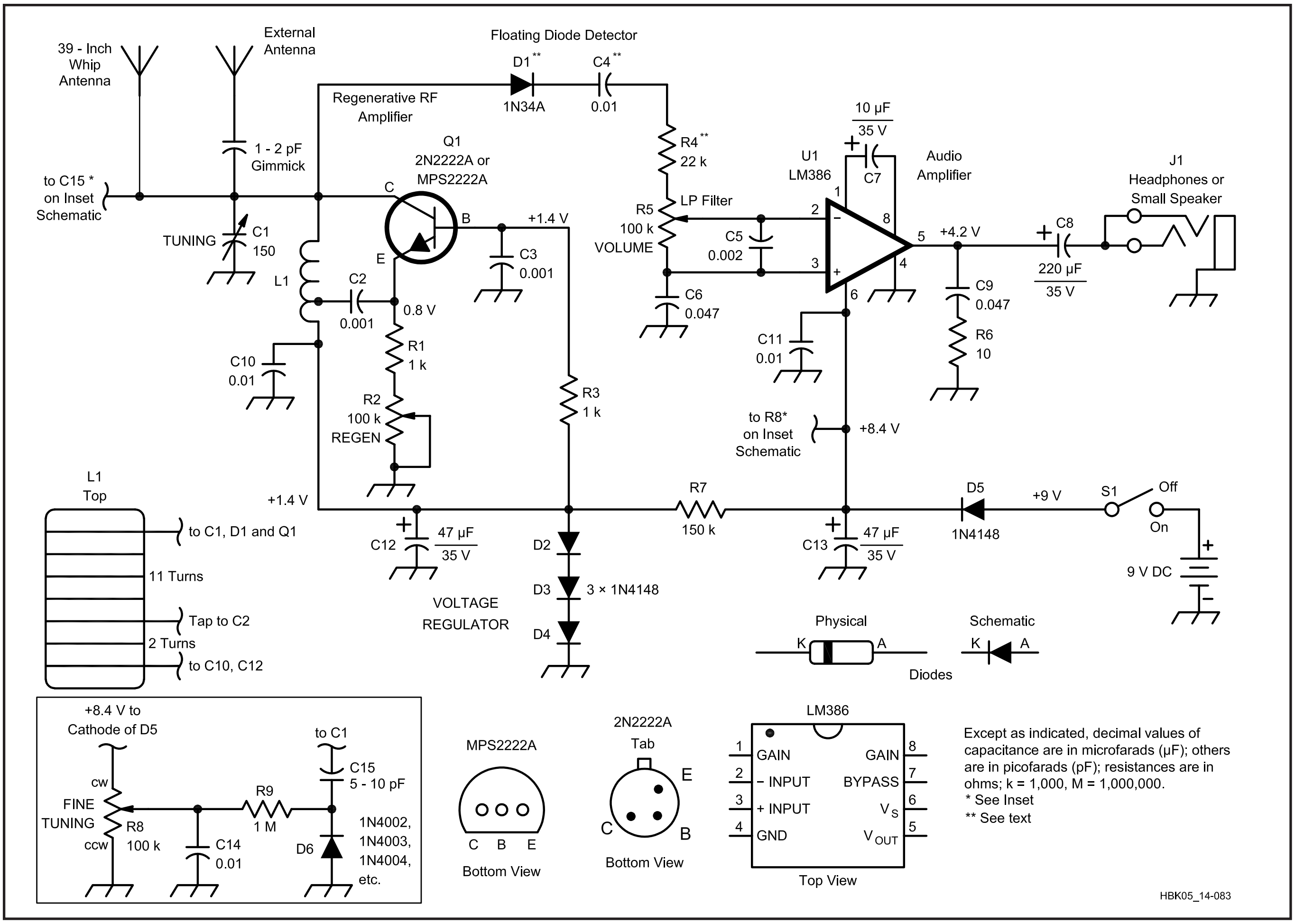


Fig 14.83—Schematic of the simple regen receiver. Unless otherwise specified, resistors are 1/4-W, 5%-tolerance carbon-composition or metal-film units. Part numbers in parentheses are RadioShack. Equivalent parts can be substituted; n.c. indicates no connection.

C1—150 pF (maximum value) air-dielectric variable capacitor; see text.
C2, C3—0.001 µF, 50 V (or more) disc ceramic (RS 272-126).
C4, C10, C11, C14—0.01 µF, 50 V (or more) disc ceramic (RS 272-131).
C5—0.002 µF, 50 V (or More) disc ceramic (use two RS 272-126 connected in parallel).
C6, C9—0.047 µF, 50 V disc ceramic (RS 272-134).
C7—10 µF, 35 V electrolytic (RS 272-1025).
C8—220 µF, 35 V electrolytic (RS 272-1017).
C12, C13—47 µF, 35 V electrolytic (RS 272-1027).
C15—5 to 10 pF, 50 V (or more) mica (RS 272-120).
D1—1N34A *germanium* diode (RS 276-1123); don't use a silicon diode here.
D2-D5—1N4148 or any similar diode (RS 276-1122).
D6—1N4003 silicon diode (RS 276-1102).
J1—1/8-inch, three-circuit jack (RS 274-246).
L1—See text.
Q1—2N2222A NPN transistor (RSU11328507) or MPS2222A (RS 276-2009).
R1, R3—1 kΩ (RS 271-1321).
R2, R5—100 kΩ potentiometer, linear taper (RS 272-092).
R4—22 kΩ (RS 271-1339).
R6—10 Ω (RS 271-1301).
R7—150 kΩ (RSU11345287) or use series-connected 100 kΩ (RS 271-1347) and 47 kΩ (RS 271-1342) resistors.
R8—100 kΩ audio-taper pot (RS 271-1722); connect so that clockwise rotation increases the voltage at the junction of the pot arm, R9 and C14.
R9—1 MΩ (RS 271-1356).
S1—SPST miniature toggle (RS 275-612).
U1—LM386N-1 audio amplifier (RS 276-1731).
Misc—PC board (see Note 1); 8-pin DIP socket for U1 (RS 276-1995A); 9-V battery clip (RS 270-325); three knobs (RS 274-402A); project box (RS 270-1806); #6-32 screws and nuts, rubber feet; 9-V battery, RadioShack 22-gauge solid, insulated hook-up wire.

age from FINE-TUNING control R8 is increased, the diode is reverse biased and its capacitance decreases. For two-band operation, use a 150-pF capacitor for C1 and install a miniature toggle switch, right on the capacitor itself, to add an additional 250-pF mica capacitor in parallel with C1. With the capacitor switched-in, the receiver will now tune the 80-meter band.

COIL WINDING

For the coil winding, use 22-gauge solid-conductor insulated hook-up wire. Drill a mounting hole in the bottom of the coil form. Then, drill two small holes in the side of the coil form, near the top. Wind the coil starting from the top of the form going to the bottom, keeping the turns well above the PC board. Feed one end of the wire through the first hole to the inside of the form, then out through the second. Tie

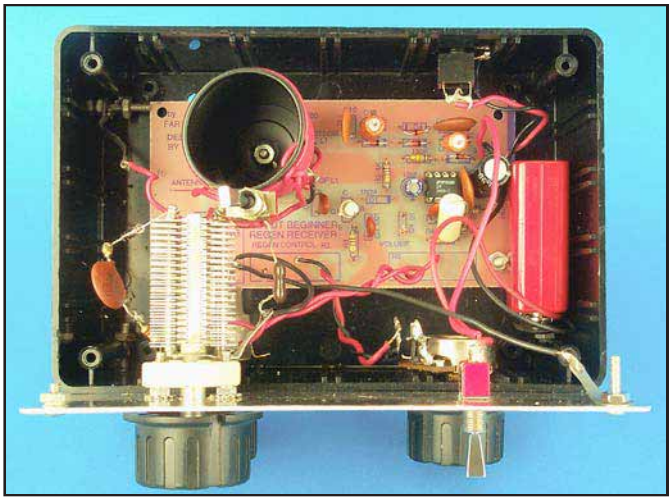


Fig 14.84—In this view of the receiver, the TUNING capacitor is at the left. Immediately behind it is the coil, L1. Just to the left of the TUNING capacitor, you can see D1, C4 and R4, as discussed in the text.

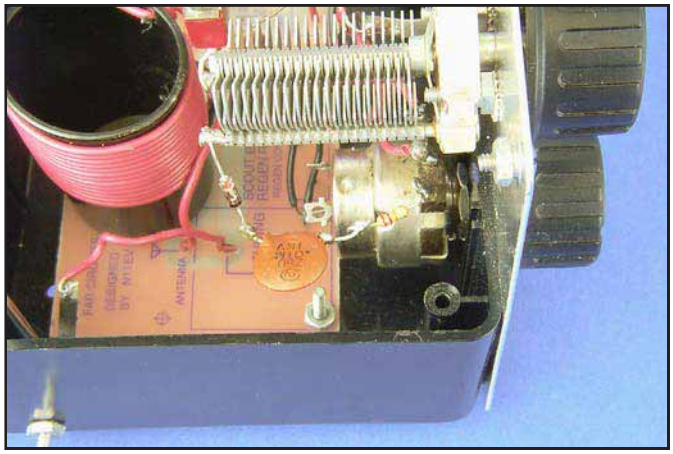


Fig 14.85—This close-up shows the interconnection of series-connected D1, C4 and R4 between the TUNING capacitor and VOLUME control.

a knot in the wire where it enters the form—this keeps the wire from loosening-up later on. Be sure to leave a two to three inch length of wire at each end of the coil so you can make connections to the PC board. You can wind the coil in either direction, clockwise or counterclockwise. Tightly wind the the wire onto the form, counting the turns as you go. Keep the turns close together and don't let the wire loosen as you wind. To make the coil tap, wind 11 turns on the coil form. While holding the wire with your thumb and index finger, mark the tap point and remove the insulation at that point. Solder a two to three-inch piece of wire to the tap. Continue winding turns until the coil is finished (13 turns total). Keep the free end of the wire in place using a piece of tape and drill two more holes in the coil form where the winding ends. Feed the wire end in and out of the coil as before and tie a knot at the end to hold the winding in place. When the coil is finished, remove the tape then carefully solder the three wires from the coil (bottom, tap and top) to their points on the PC board keeping the wire lengths as short as possible. Mount the coil away from any metal objects.

CONSTRUCTION DETAILS

The prototype receiver used a Radio Shack plastic project box, with a metal front panel. For best performance, the floating detector must be wired using short, direct connections. Therefore, these components are not mounted on the PC board. Mount the VOLUME control, R5, close to the TUNING capacitor, C1. Connect D1, C4 and R4 in series between the "hot" side of C1 (the stator) and the top of the VOLUME control. See Fig 14.85. Mount the TUNING capacitor and the regeneration (REGEN) control on opposite sides of the front panel. The VOLUME and REGEN controls are best mounted near the bottom of the front panel to keep their connecting wires to the PC board as short as possible. Use very short twisted wires or shielded wires for the VOLUME and REGEN control connections. Be sure to connect a wire between the metal front panel and the PC board ground. If you use the RadioShack jack specified for J1 (RS 274-276), connect pins 2 and 5 together and attach that common lead to C8. Ground pin 1 of the jack. If you intend to use a small speaker, connect it between pins 1 and 3. Then, when headphones are plugged in, the speaker will be disconnected automatically.

This receiver will pick up lots of stations using a 39-inch whip antenna connected directly to the "hot" side of C1 (the junction of C1, L1 and the collector of Q1). The set will be easily detuned by hand and body movements, however. For best performance, use an external 20 foot, or longer, length of insulated wire run outside to a tree. But be sure not to connect this wire directly to the set, as this will load down Q1 and ruin selectivity. Instead, use a small "gimmick" capacitor, approximately 1 to 2 pF, between the external antenna and C1. This is easy to do: simply twist together two pieces of solid insulated hook-up wire, each about two inches long. Solder one wire to C1, the other to a banana jack on the back of the set. Twist the wire's other (insulated) ends together approximately three times, depending on antenna length. Twist the wires enough to bring in lots of stations but not enough to load down Q1.

TESTING AND OPERATING THE RECEIVER

Set the VOLUME and REGEN controls to midrange, plug in the headphones, attach the battery and turn on the receiver. Check to see that the audio stage is working by placing a finger on the wiper of the VOLUME control and listening for a buzz. Then use just three feet of hook-up wire connected directly to C1, as a temporary antenna. Adjust the REGEN control until the set produces a "live" sound, indicating that Q1 is oscillating. If not, then carefully recheck the wiring and measure the voltages labeled on the schematic, using a high-impedance DVM or multimeter. Common problems are Q1 being wired backwards (emitter and collector connections reversed) and the wires from coil L1 connected to the wrong places on the PC board.

Use two hands when operating the receiver: one for tuning, the other for controlling regeneration. For international broadcast stations or AM phone operation, carefully, adjust the REGEN control so that Q1 operates just below oscillation. For CW and SSB, increase the REGEN level so that the set *just* oscillates, providing the required local oscillation for these modes. If you operate this receiver close to another radio, the regen's 30-µW oscillator might cause interference. For details on building a higher-performance regen receiver for serious CW and SSB reception see: Kitchin, "High Performance Regenerative Receiver Design," Nov/Dec 1998 *QEX*.

[1]A PC board for this radio is available from FAR Circuits. Price: $5 each plus $1.50 shipping for up to three boards.