output variation with tuning, this output is constant for the total tuning range. The oscillator frequency is not directly related to the transmitter output frequency, so there are few problems relating to stray power amplifier energy in the oscillator circuitry. Finally, the output can be turned off and on by controlling a digital reset line in the divider. As such, there is a perfect method for keying without every changing the oscillator operating frequency. The oscillator runs continuously and does not change frequency during a transmit interval. The usual mechanisms for generating chirp are absent.

The oscillator, divider, and filter portion of the 20-m transmitter is shown in **Fig 5.49**. A crystal at 9.373 MHz (HC-49, 20-pF load) was chosen to provide about 10 kHz of tuning around the desired output frequency of 14.06 MHz. The range is obtained without any crystal series inductance. However, the builder may wish to add inductance to extend the tuning range. The output from oscillator Q1 drives Q2, conditioning the signal for logic compatibility. This then drives a 74HC74 divide-by-2 chip. During normal key-up conditions, pin 1 is held low by Q3. This "reset" prevents any output from appearing from the IC. When the key or spot switch are pressed, pin 1 goes high, and the divider generates the desired 4.687-MHz output. The 5 V bias for U1 is obtained from U2, a low power regulator.

A 2-kΩ pull up resistor on U1's Q output helps to ensure that the output goes all the way to 5 V during operation, establishing the logic level, and hence, the RF output level. The combination of the resistor and the chip circuitry generate a load of approximately 1 kΩ to provide filter loading at the input end, establishing the values for C8 and C9. The filter is designed for a 50-Ω output load. The available power at the third harmonic is about 0 dBm. This filter is designed for a bandwidth of 400 kHz at 14 MHz. With the inductors used, the filter insertion loss is about 3 dB.

A buffer amplifier, Q5, increases the output from the filter to a comfortable +11 dBm. Q5 is only powered on key-down intervals, controlled by a delayed switch, Q6, which also provides the needed control signal "A" for U1. A 4.7-µF capacitor keeps this switch "on" for a short interval after key down. The 1-kΩ resistor in series with the 4.7-µF capacitor allows the "A" signal to immediately change with the initial application of the key while the transmitter output is still shaped with the circuitry around Q9. This creates a "time sequence" keying scheme, similar to one applied to vacuum tube transmitters of the 1950's era.

The buffer output is applied to a 100-Ω pot functioning as a Drive control, and then to a keyed driver, Q7. This stage and the output power amplifier are shown in **Fig 5.50**. These components are on a separate board from the earlier circuitry, further isolating the circuits. The driver, a medium power bipolar feedback amplifier, is capable of an output of up to 300 mW. The keying is done with Q9, a shaping integrator-switch.

The output amplifier uses an inexpensive HEX FET. Some regulated 5-V energy is stolen from the other board and applied to a pot that generates bias for the FET PA. The bias is adjusted by monitoring the FET drain current with a sensitive meter and is set for a current of close to 1 mA. This amplifier will run in Class B, off during key up conditions, allowing the use of electronic T/R switching. However, forward FET bias enhances both gain and stability. The FET output is matched with a modified LCC type T-network consisting of L5 and a pair of mica compression trimmer capacitors. This is followed by additional low pass filtering. The output is set to 4 W by adjusting the drive and tun-

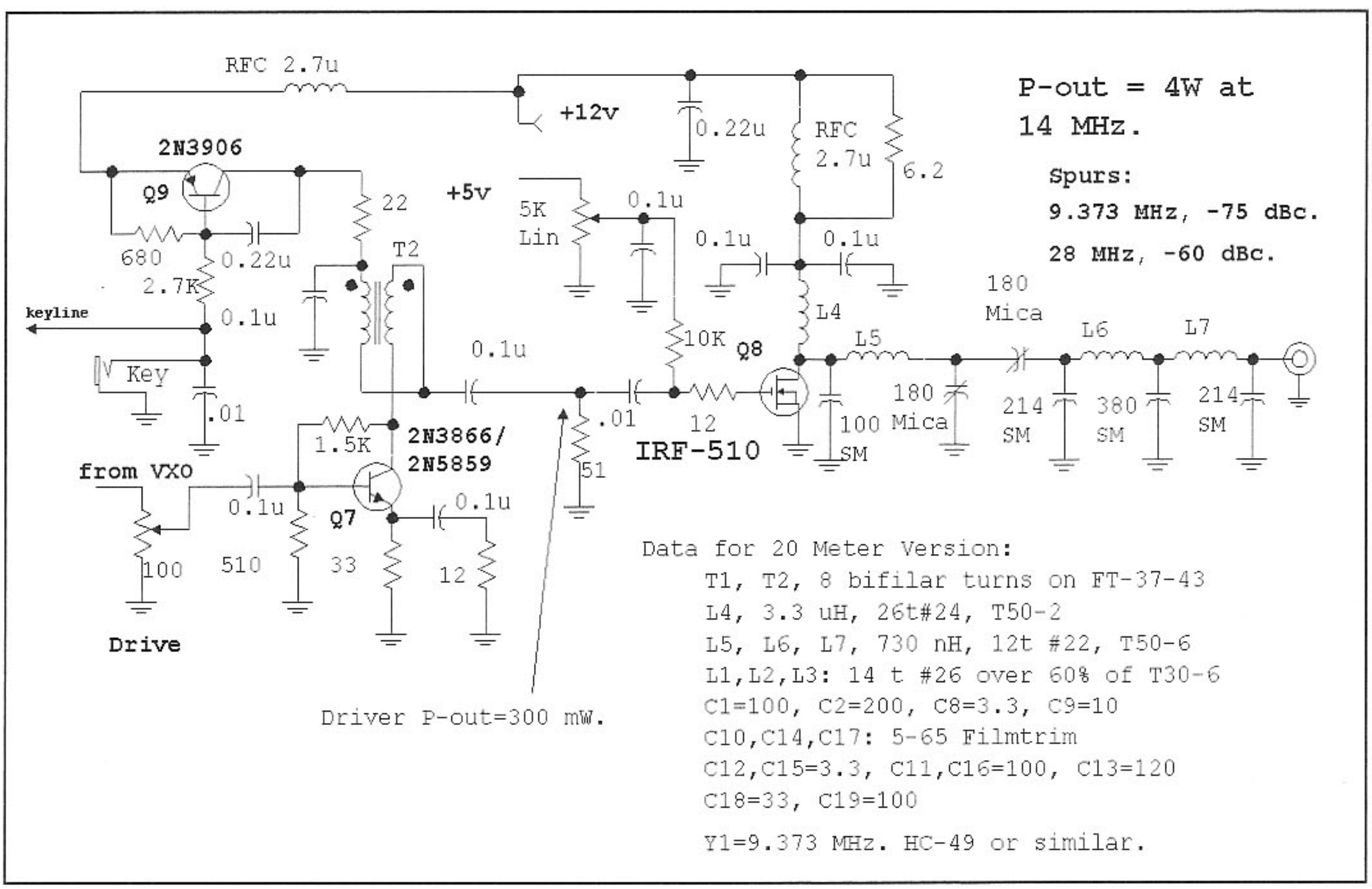


Fig 5.50—Keyed driver and power amplifier for the transmitter.

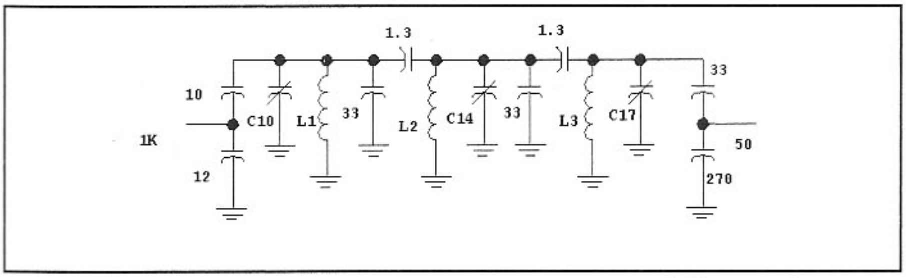
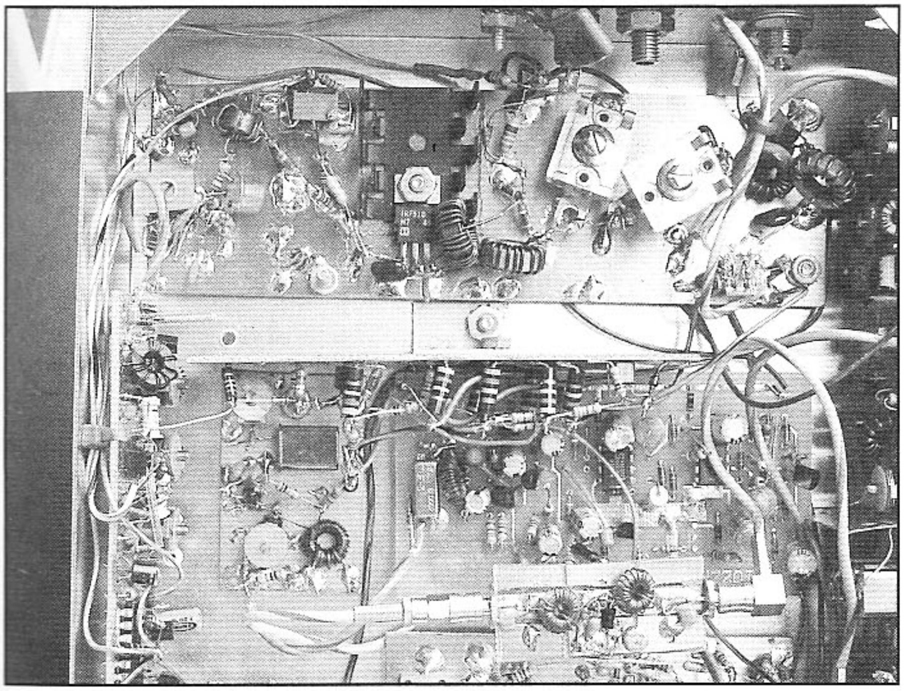


Fig 5.51—A 21-MHz bandpass filter. The inductors and the variable capacitors are identical to those used in the 14-MHz design.



The 4-W output power amplifier is shown at the top of the photo. The board includes the keyed driver, drive control pot, and bias pot. The box housing this rig also includes a 20-meter receiver (The "Easy 90-14") described in Chapter 6.

ing the T-network capacitors for maximum output.

A subtle instability was noted during the transmitter turn-on process. In an effort to make the transmitter as clean as possible, an extra 2.7-µH RFC had been included in the drain line. But a low level oscillation was noted in the PA. An oscilloscope examination revealed a frequency of 300 kHz. This turned out to be the result of a resonance between the 2.7 µH and the bypass capacitors. A 6.2-Ω resistor was paralleled across the RFC and the oscillation was eliminated. This illustrates the subtlety of wideband bypassing of power stages in a transmitter. See the information on decoupling in Chapter 2.

The only spurious responses noted in the output were at the crystal oscillator frequency and at the transmitter 2nd harmonic, but they were below the desired output by 75 and 60 dB, respectively. Yet the transmitter is built with no internal shielding or other complexities.

A 21-MHz version of this design would be especially practical, for it could use an existing 14-MHz crystal. A 21-MHz bandpass filter is shown in **Fig 5.51** to aid the designer/builder in realizing a rig for that band.

Although the digital divider was originally implemented for use with simple low power transmitters, it lends itself well to general-purpose applications with LC oscillators as well as crystal-based designs.

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CHAPTER 6

Transmitters and Receivers

6.0 SIGNALS AND THE SYSTEMS THAT PROCESS THEM

The basic building blocks of amplifiers, filters, oscillators, mixers, and frequency multipliers have been discussed. We now begin to combine these components to build the equipment that provides communications. We begin the chapter with a look at CW, AM, DSB, SSB, and FM signals. Block diagrams are then shown for the equipment we build to deal with these signals. Later sections will present detailed design methods and examples.

Signals are presented as equations. We then show graphs in the time and frequency domains, the results we would observe with either an oscilloscope or spectrum analyzer. This discussion is not intended to be complete, but is merely a sketch of signal forms. A complete treatment is found in communications texts.[1]

The first signal we consider is the audio, or *baseband* representation. This might represent the output of a receiver or a voice signal that we apply to a transmitter microphone input. A receiver output from a CW signal is generally a rather pure sine wave, perhaps at a frequency of 1000 Hz. Mathematically this is

$$v(t) = \sin(2\pi f t) \qquad \textbf{Eq 6.1}$$

where v(t) indicates that the voltage is a function of time, f is the frequency in Hz, and t is time in seconds. Graphed in the time domain, the tone is the familiar sine wave, **Fig 6.1**. The energy is confined to a single frequency, so the spectrum, or frequency domain representation is a single line, **Fig 6.2**. The 1-V amplitude has a spectrum with a height of 1 V. It is more common within the radio frequency design arena to see spectra calibrated in terms of power.

The human voice is not a sine wave, but a combination of tones forming complicated patterns in both time and frequency. The actual signals are difficult to handle with simple equations and are different for every voice. So, we approximate a voice signal with several sine waves. The baseband example we use (**Figs 6.3** and **6.4**) has three tones of $f_1 = 1000$, $f_2 = 2500$, and $f_3 = 400$ Hz with respective amplitudes of 0.6, 1, and 0.5 V. The total baseband signal is

$$v_b(t) = 0.5 \sin(2\pi f_3 t) + 0.6 \sin(2\pi f_1 t) + 1 \sin(2\pi f_2 t) \qquad \textbf{Eq 6.2}$$

Traditional amplitude modulation is familiar as an AM broadcast signal. This is generated by changing—or modulating at an audio rate—the amplitude of a *carrier*. The carrier is merely a single sinusoid. A frequency of 100 kHz is used in our

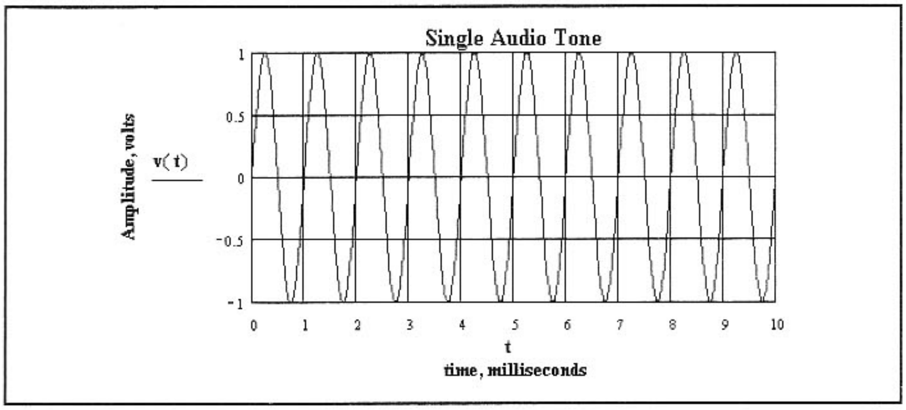


Fig 6.1—A single audio tone as a function of time.

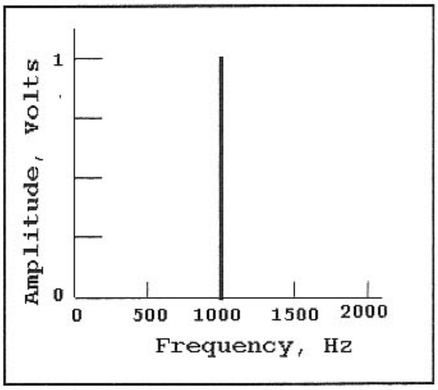


Fig 6.2—The 1000 Hz audio tone in the frequency domain.

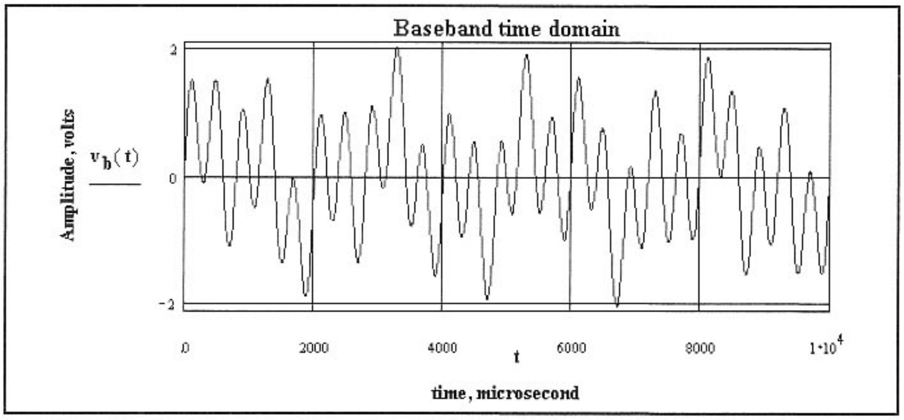


Fig 6.3—The time-domain graph of the three audio tones.

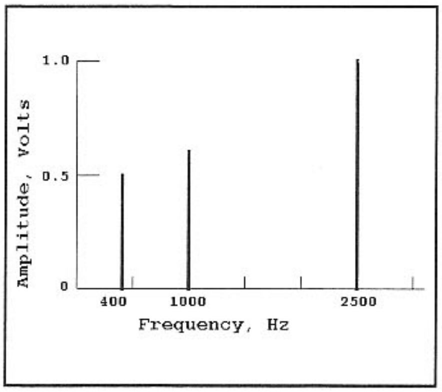


Fig 6.4—The frequency-domain graph of the three audio tones.

examples. The graphs and equations are the same as the earlier single-tone audio signal, except that the frequency is higher.

The carrier amplitude is modulated to generate the AM signal of **Eq 6.3**.

$$v_a(t) = \left(1 + 0.3 \sin\left(2 \pi f_{aud}\ t\right)\right) \times \sin\left(2 \pi f_c\ t\right)$$

Eq 6.3

where f_c is the carrier frequency of 100 kHz and f_{aud} is the audio frequency of 1 kHz. The 0.3 factor is a modulation index and indicates 30% modulation. The time domain signal is shown in **Fig 6.5** with a spectrum in **Fig 6.6**. The two curves are related through appropriate mathematics, which follow from the trig identity shown in the *Trig Identities for Signal Analysis* sidebar. A detailed mathematical analysis will always tie the two domains together. Modulations that are simple in one domain are often complicated and messy in the other.

The time domain waveform shows that the amplitude of the RF sine wave varies, exceeding the original carrier amplitude for part of the cycle. The frequency domain graphs show that extra energy to be contained in the frequency domain sidebands while the carrier remains constant with no audio variation. This is easily confirmed by observation with a spectrum analyzer or receiver that will resolve the carrier from the sidebands.

A 100-kHz carrier modulated by the three-tone baseband signal is shown in **Fig 6.7** and **Fig 6.8**.

The multi-tone amplitude modulation is described by

$$v_{am}(t) = \left(1 + 0.3\, v_b(t)\right) \sin\left(2 \pi f_c\ t\right)$$

Eq 6.4

where the sine term represents the carrier and $v_b(t)$ is the baseband signal from **Eq 6.2**. The first set of parentheses on the right side of the equal sign in **Eq 6.4** contains the unity term, which leads to the carrier in the final result, and the complex audio signal $v_b(t)$ that generates the sidebands.

A double sideband signal results when audio is applied to a *balanced* modulator driven by a local oscillator. The resulting output for a single modulating audio tone is

$$v_d(t) = \sin\left(2 \pi f_{aud}\ t\right) \sin\left(2 \pi f_c\ t\right)$$

Eq 6.5

where the first term is the audio while the second is the carrier. The term with unity in **Eq 6.4** is missing from **Eq 6.5**, indicating that the carrier is no longer present. The waveforms are shown in **Fig 6.9** and **Fig 6.10**.

The result of a double-sideband generator driven with the multiple-tone audio is then

$$v_{dsb}(t) = \sin\left(2 \pi f_{U2}\ t\right) + \sin\left(2 \pi f_{L2}\ t\right) + 0.6 \sin\left(2 \pi f_{U1}\ t\right) + 0.6 \sin\left(2 \pi f_{L1}\ t\right) + 0.5 \sin\left(2 \pi f_{U3}\ t\right) + 0.5 \sin\left(2 \pi f_{L3}\ t\right)$$

Eq 6.6

where the frequencies shown represent the

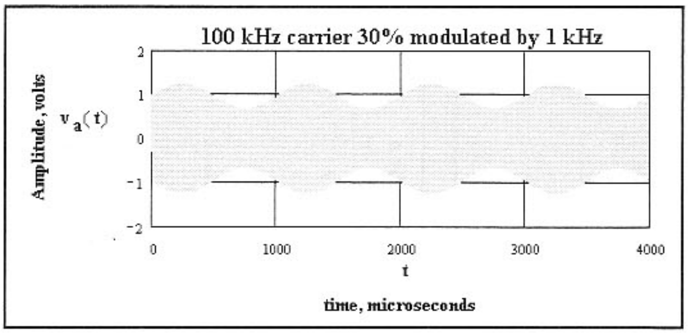


Fig 6.5—The carrier amplitude here is 1 V. Modulation causes the amplitude to depart from this value. The energy appears in the figure to be a solid mass of energy, but if we zoom in, plotting only a small fraction of the curve shown, we will see the details of the RF oscillation. This could be done experimentally with an oscilloscope triggered from the RF waveform.

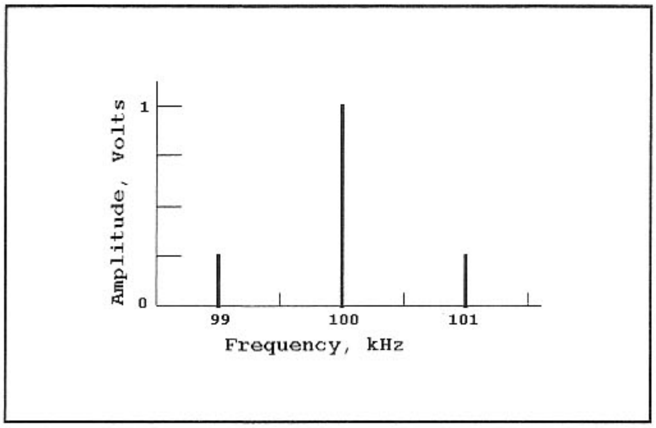


Fig 6.6—Frequency-domain representation of an AM signal. The carrier at 100 kHz is modulated at 1 kHz to generate two sidebands below and above the carrier.

Trig Identities for Signal Analysis

In high school trigonometry class you may have learned some useful identities. One of them relates the product of two sine functions:

$$\sin(a)\sin(b) = \frac{1}{2}\cos(a-b) - \frac{1}{2}\cos(a+b)$$

Our analysis of amplitude modulation started with a carrier of amplitude A:

$$A\sin(\omega_c\, t)$$

where $\omega_c = 2\pi f_c$ is a carrier frequency expressed in radians/sec, with f_c in Hz. The amplitude is allowed to vary about a base value.

$$A = A_0\left(1 + m\sin(\omega_a\, t)\right)$$

where ω_a is an audio frequency in radians/sec and m is a modulation index. The modulated wave becomes:

$$v(t) = A_0\left(1 + m\sin(\omega_a\, t)\right)\sin(\omega_c\, t)$$ which expands to:

$$v(t) = A_0\sin(\omega_c\, t) + A_0\sin(\omega_c\, t)\, m\sin(\omega_a\, t)$$

The first term is the carrier, which varies only with time at the carrier rate, ω_c. The second term is the product of audio and RF carrier sine waves. Expansion with the identity yields:

$$A_0\, m\sin(\omega_a\, t)\sin(\omega_c\, t) = A_0\, m\left[\frac{1}{2}\cos\left[(\omega_c - \omega_a)t\right] - \frac{1}{2}\cos\left[(\omega_c + \omega_a)t\right]\right]$$

and then:

$$A_0\, m\sin(\omega_a\, t)\sin(\omega_c\, t) = A_0\, m\left[\frac{1}{2}\cos\left[2\pi(f_c - f_a)t\right] - \frac{1}{2}\cos\left[2\pi(f_c + f_a)t\right]\right]$$

The two cosine waves on the right are the lower and upper sidebands of the AM signal.

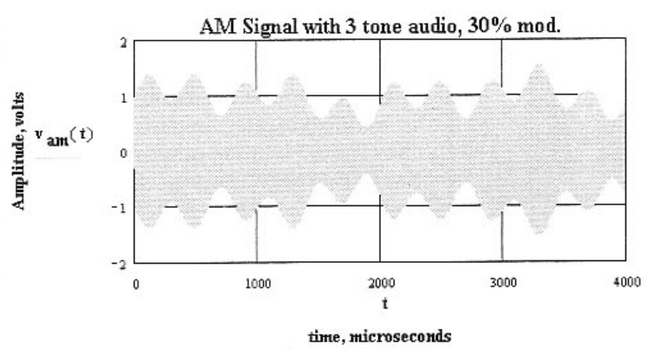


Fig 6.7—A three-tone baseband signal modulates a 100-kHz audio tone.

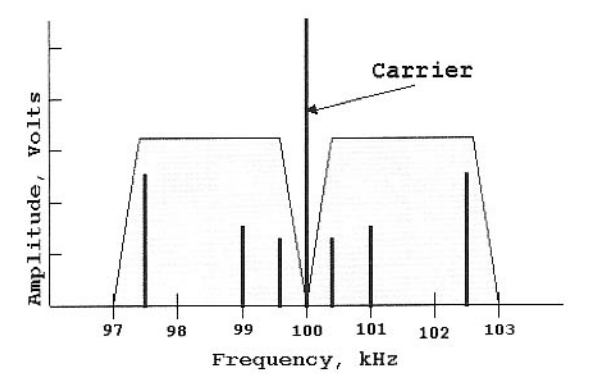


Fig 6.8—Frequency-domain view of amplitude modulation with a three-tone baseband signal. The two sideband regions are now shaded.

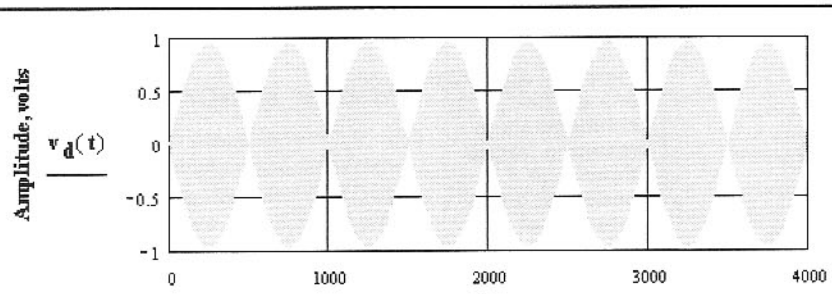


Fig 6.9—100-kHz double-sideband output with 1-kHz audio.

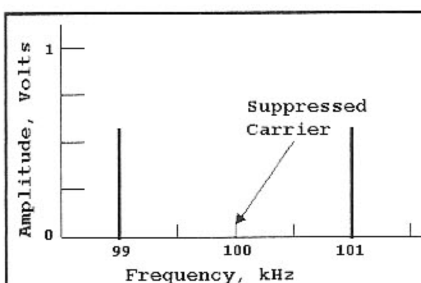


Fig 6.10—Frequency-domain view of a DSB signal with a single audio tone. Two output frequencies are created.

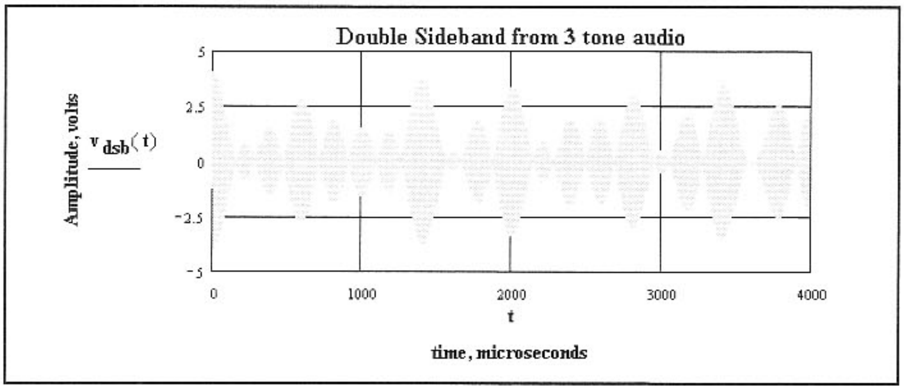


Fig 6.11—Double sideband with a multi-tone audio, time domain.

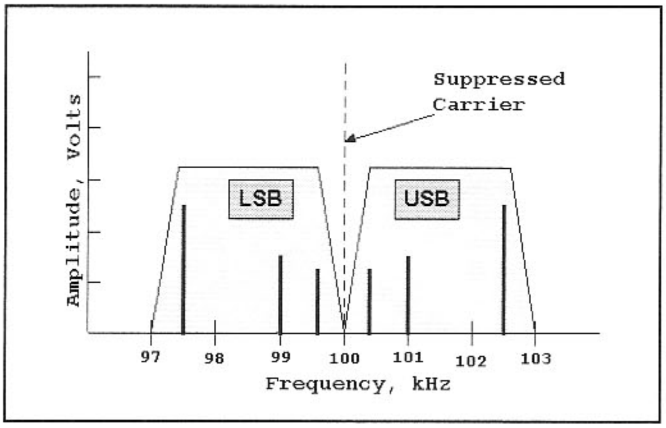


Fig 6.12—Frequency-domain representation of DSB with multiple-tone audio. The upper and lower sideband parts of the spectrum are highlighted.

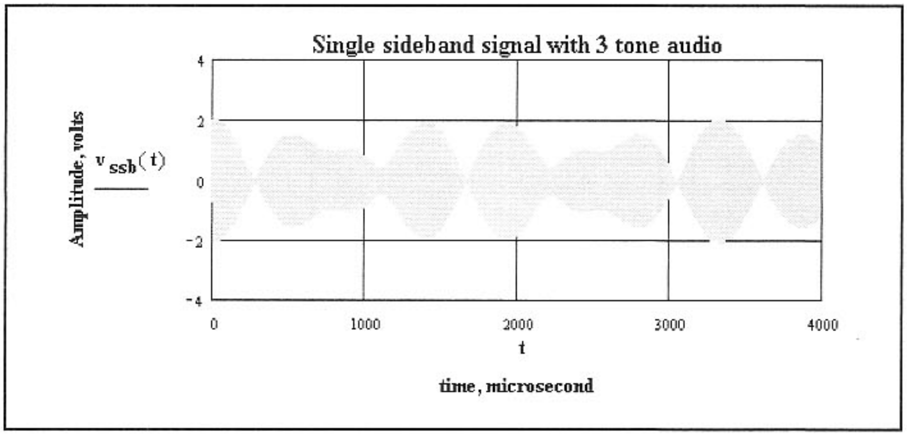


Fig 6.13—Single-sideband signal from a three-tone baseband input.

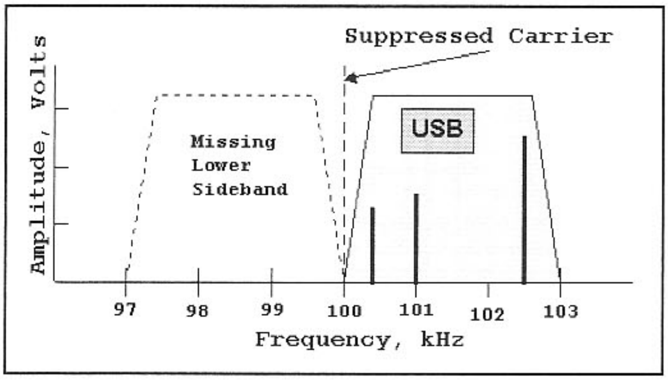


Fig 6.14—Spectrum of a single-sideband signal resulting from a three-tone baseband audio input.

upper and lower sideband components resulting from audio components at f_1, f_2, and f_3. The DSB signals are shown in **Fig 6.11** and **Fig 6.12**.

A single sideband (SSB) signal is described by eliminating one of the sidebands. For this example, we retain the upper sideband, resulting in

$$v_{ssb}(t) = \sin(2\pi f_{U2}\, t) + 0.6 \sin(2\pi f_{U1}\, t) + 0.5 \sin(2\pi f_{U3}\, t) \qquad \textbf{Eq 6.7}$$

The corresponding graphs are **Fig 6.13** and **Fig 6.14**.

The SSB signal, when viewed in the frequency domain, is really nothing more than an exact replica of the original baseband signal, except that it is now translated linearly to a higher frequency. If a lower sideband signal had been generated, it would have been a replica of the original with an inversion. That is, what had started as a high audio frequency of 2500 Hz now appears as the lowest frequency.

A frequency-modulated signal is described by

$$v_{fm}(t) = \sin\left[2\pi f_c \left(1 + m \sin(2\pi f_a\, t)\right) t\right] \qquad \textbf{Eq 6.8}$$

If we pick a 10-kHz carrier and modulate it with a 1-kHz audio signal, we see the time domain signal of **Fig 6.16**. The amplitude is constant, but the frequency varies.

Extracting the spectrum for this signal is mathematically much more difficult than it was with the other signals, for the audio sine wave is now *inside* the argument for the basic signal before modulation, as seen in **Eq 6.8**. Signals appear about the carrier, spaced by the audio frequency. However, several sets appear. A 1 kHz audio tone produces signals at +/–1, +/–2 kHz and so on, as shown in **Fig 6.17**. The strength of the sidebands *and the carrier* depend on m, still a modulation index, and are described by Bessel functions.[2] No FM equipment is described in this book, but the equations are included for completeness.

Block Diagrams

We now examine basic transmitters and receivers, beginning with simple CW gear. A CW transmitter generates a carrier at a single frequency with no modulation other than the off-on keying that imposes the familiar encoding. A simple CW transmitter is shown in **Fig 6.18**. The circuit begins with an oscillator operating at the final output frequency. Typical oscillators are usually fol-

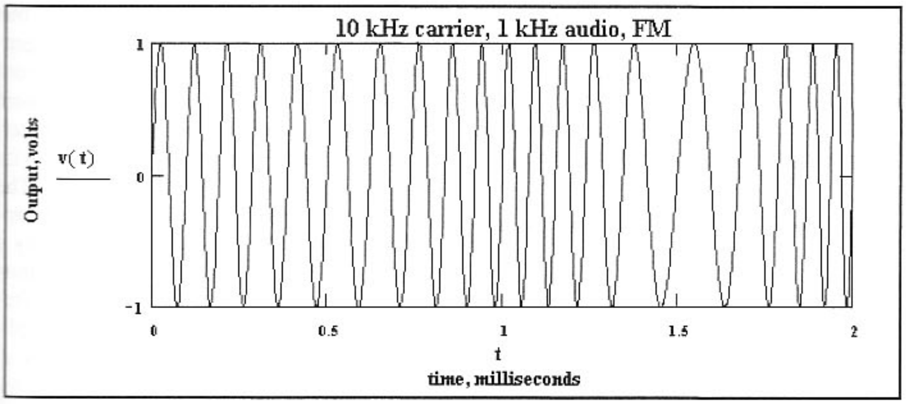


Fig 6.16—Time domain representation of an FM signal.

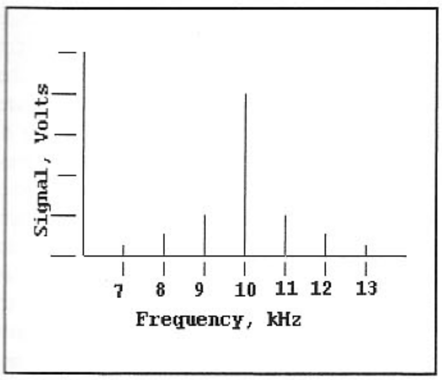


Fig 6.17—Spectrum of an FM signal, 10-kHz carrier with 1-kHz audio. This graph represents what we might observe with a typical spectrum analyzer. We often see plots like this with some components below the frequency axis, indicating a sign change when frequency is modulated rather than amplitude.

lowed by amplifiers (perhaps several) to increase output power. The final block is a low-pass filter to remove harmonics.

The amplifiers serve the additional function of buffering the oscillator. Buffers may have low gain, but have much more gain in the normal forward direction than in the reverse one. A typical 20-dB gain design might have a gain of –30 dB in the reverse direction. This serves to prevent large transmitter output signals from reaching the oscillator. Common-base (gate) amplifiers usually feature excellent reverse isolation.

A crystal or an LC resonator determines the oscillator frequency. The oscillator should be shielded from the rest of the transmitter to prevent transmitter output components from reaching it. An oscillator is most sensitive to signals at frequencies within the loaded bandwidth of the resonator controlling the oscillator. Hence, shielding is especially important for the simple transmitter of Fig 6.18. Poor shielding or inadequate buffering allows the frequency to change (pulling) when the amplifiers are keyed on. The output frequency then differs from that when the amplifier is off.

The modified circuit of Fig 6.18B uses a frequency multiplier between the oscillator and the power amplifiers. The buffering action of a frequency multiplier is profound. Signals travelling from the output backward in a buffer remain at the output frequency. The buffer input, including the oscillator, is not usually sensitive to this. With the transmitter output at a multiple of the oscillator frequency, it no longer has components within the bandwidth of the oscillator tank, so is not susceptible to the pulling mentioned. Indeed, it is often practical to build transmitters with no inter-stage shielding whatsoever if multipliers are used. A bandpass filter is used at the multiplier output to suppress direct feed-through from the oscillator and harmonics— other than the desired one— that are often present. The filter can often be as simple as a single resonator if the multiplier is just a balanced frequency doubler. More often, we use double or triple tuning at the output of multipliers.

A mixer is often used within a CW transmitter with a bandpass filter to select the desired frequency, shown in **Fig 6.19**. This example has a 2-MHz variable-frequency oscillator, a 5-MHz crystal-controlled oscillator, and an output at 7 MHz. The VFO tunes a 150-kHz range to cover the CW portion of the 7-MHz band. The bandpass filter must be wide enough to pass the entire range, but should not be a lot wider, for spurious mixer products must also be suppressed by the filter. The 5-MHz component will be suppressed by balance

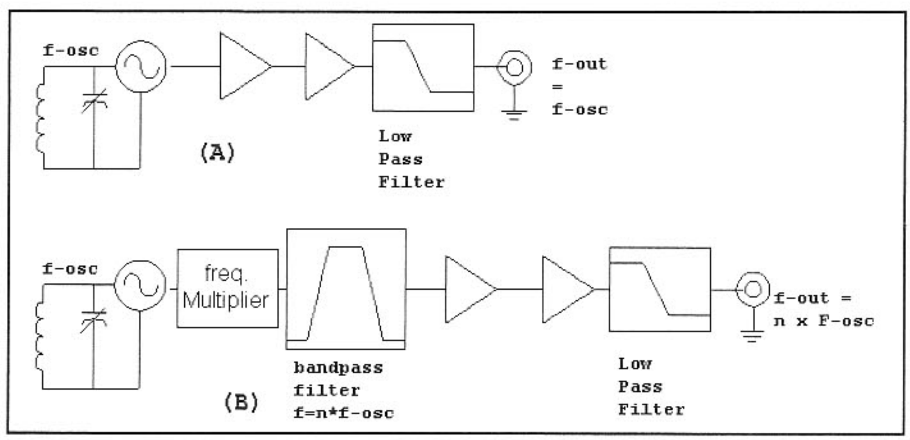


Fig 6.18—Simple CW transmitters with a *master oscillator* and a *power amplifier* are traditionally called a MOPA design. Design "A" has the oscillator and amplifier operating at the same frequency while that at "B" uses frequency multiplication.

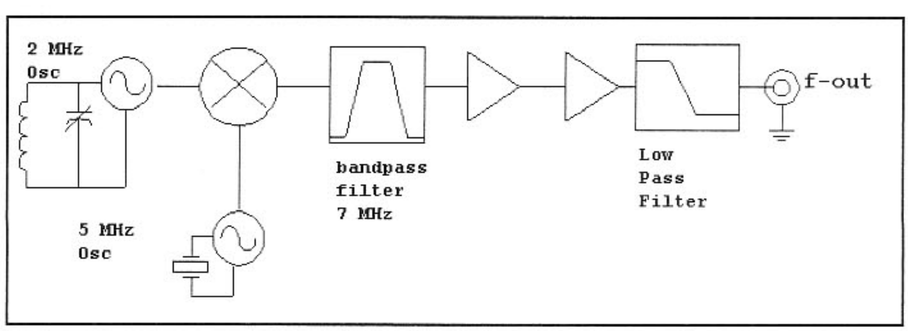


Fig 6.19—A CW transmitter using a mixer. Frequency stability is improved owing to use of a lower frequency for the variable-frequency oscillator. Careful bandpass filtering is required at the mixer output to preserve spectral purity.

in the mixer, but may often need to be further attenuated by the bandpass filter. A typical circuit would often use a triple-tuned filter if intended to meet modern standards.

These methods are not restricted to simple CW transmitters. Heterodyne methods are also useful when building local oscillator systems for SSB or similar equipment.

A CW signal is received by heterodyning the radio frequency energy down to baseband where it can be heard. This may occur in one step in a direct-conversion (including regenerative) receiver or in several steps in a conventional superheterodyne. The key element in a direct-conversion receiver is the mixer, or as it is usually called in applications with an audio output, the product detector. The input signal, usually relatively weak, is applied to the RF port of a mixer driven by a strong local oscillator. Two mixer outputs will appear, but only the audio difference frequency is used. The signal is usually amplified further and is applied to headphones. A block diagram is shown in **Fig 6.20**. The input preselector filter protects the receiver from strong signals at frequencies far removed from those being received. The low-pass filter routes audio to the amplifiers while preventing other mixer products or mixer feed-through components from reaching the amplifier. Direct conversion receivers are covered in much greater detail in Chapter 8.

An instructive experiment tunes the frequency of a signal generator attached to a direct conversion receiver. One will then hear an audio beat note, the difference frequency between the generator and the receiver local oscillator. The output frequency is shown in **Fig 6.21** as a function of generator frequency. Tuning the receiver with a fixed generator produces an identical result. The response is double sided; for every tuning of a simple direct conversion receiver, there are two different input frequencies that can produce the same output signal. One response is called the audio image of the other. This makes it challenging to use such a receiver in severely congested bands. But the simplicity and other good qualities of a direct conversion receiver will often compensate for this problem.

The traditional solution to the audio image problem is the single-signal superheterodyne receiver shown in the block diagram of **Fig 6.22.** The incoming signal is processed in a preselector filter and then applied to a mixer. The output is still at a radio frequency, but one that is different from the incoming signal, an intermediate frequency, or IF. This 7-MHz receiver uses a 1-MHz IF with an LO in the 6-MHz region. The 1-MHz signal from the mixer is filtered with a narrow bandwidth circuit. It is further amplified and applied to a second mixer, now functioning as a product detector to produce an audio output. After some audio gain, headphones are driven. The LO for the product detector is called a beat frequency oscillator, or BFO.

Assume that the 1-MHz IF filter has a bandwidth of 500 Hz, centered exactly at 1 MHz. The receiver LO will be tuned to 6.040 MHz. This means that the incoming signals that will produce an output are cen-

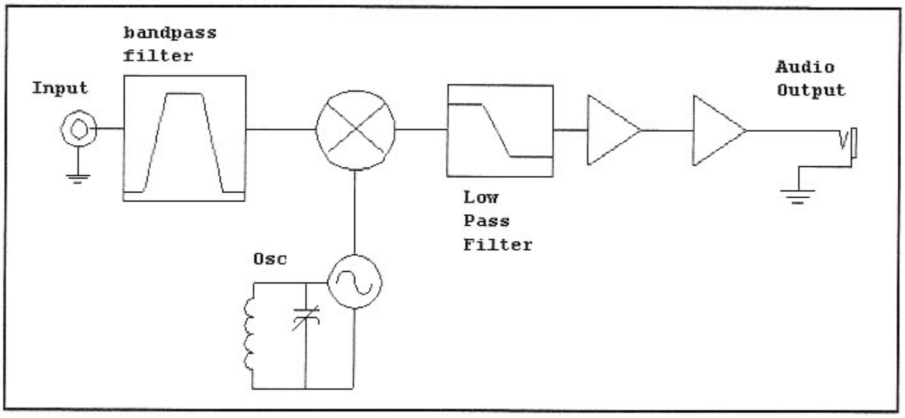


Fig 6.20—Direct-conversion receiver. The incoming signal is applied to a mixer where it is converted directly to audio without intermediate processing.

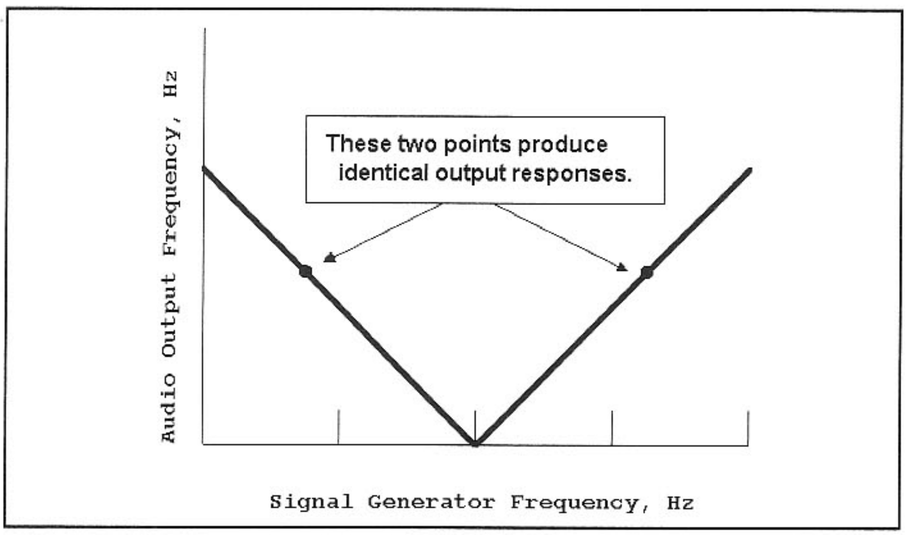


Fig 6.21—Tuning response of a fixed-tuned DC receiver while varying a signal generator applied to the input. A 1000-Hz beat note is available from the generator at two different generator frequencies. One response is the audio image of the other.

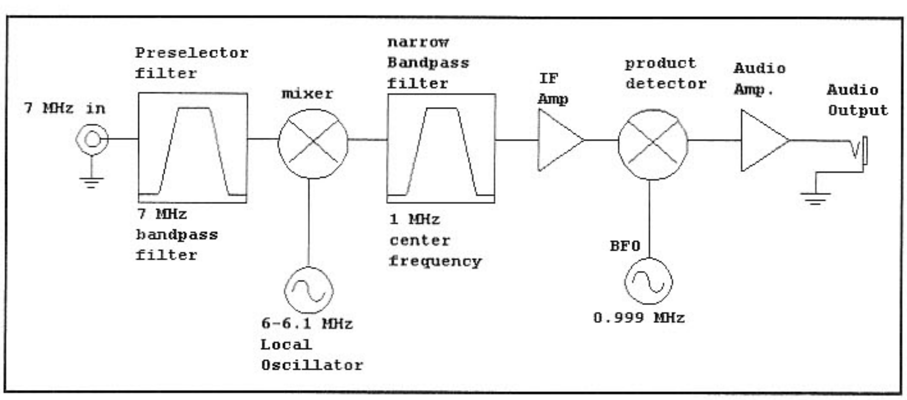


Fig 6.22—A simple single-conversion superheterodyne receiver featuring a "single-signal response." A narrow filter, usually using a quartz crystal, follows the mixer.

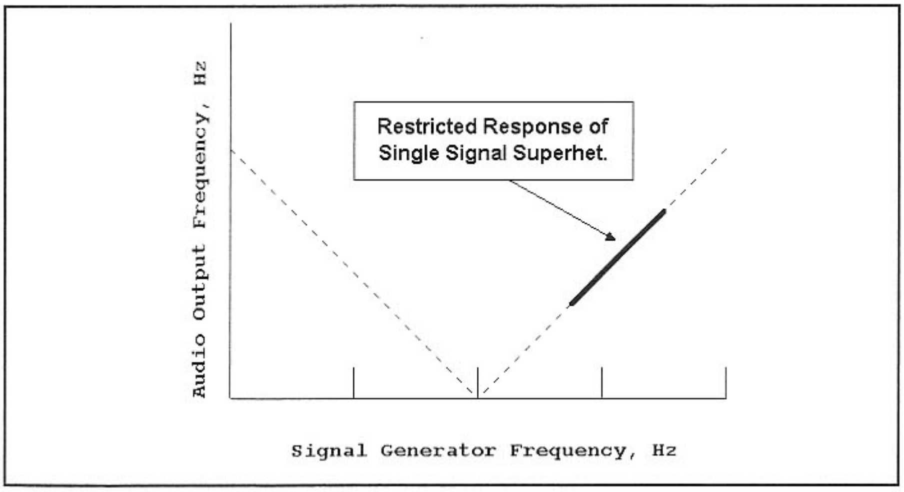


Fig 6.23—Tuning response to the single-signal superhet. The output from a single source occurs in a single area on the dial.

tered at 7.04 MHz and occur in a 500-Hz band, 250 Hz on either side of 7.04 MHz. Signals within that band are the only ones that will produce an IF output. Set the BFO to 0.999 MHz, 1 kHz away from the IF center. An IF signal at 1 MHz will then produce a 1-kHz beat note. But the only beat notes that are possible for this BFO setting are in a 500-Hz wide span from 750 to 1250 Hz. Repeating the earlier experiment performed with the direct conversion receiver yields the result of **Fig 6.23**.

A single-signal response can also be obtained with phasing methods, and related schemes. These are covered in detail in Chapter 9.

DSB

Let's return to the transmitter problem, but now consider the generation of a double sideband signal. (Double-sideband, full-carrier amplitude modulation is of great historic interest, especially to collectors, but is not the most-used method of voice communications today. We won't treat the method in this book.) The key element needed to generate DSB is a balanced mixer. It will be driven with a suitable RF local oscillator and low level audio from an amplified microphone. The output, shown earlier in Fig 6.10, contains the two sidebands symmetrically spaced about a suppressed carrier. Further amplification and low-pass filtering completes the transmitter. A simple DSB transmitter is shown in **Fig 6.24**. A typical simple DSB transmitter will have a carrier that is suppressed by 30 to 40 dB with respect to either sideband. Although simple and compatible with existing SSB equipment, DSB transmitters are rarely used today, largely due to the excess spectrum used.

Audio intelligence is impressed on the signal in DSB and SSB transmitters with a block traditionally shown as a balanced modulator. The modulator is really just a mixer with a particular application. It is usually a balanced circuit, for that is the mechanism used to suppress the carrier output. See balance in Chapter 5.

The direct-conversion receiver shown earlier (Fig 6.20) will allow DSB signals to be received. Each of the two sidebands will be heterodyned down to baseband where they will add to produce an audio output. It is *vital* that the BFO be *exactly*

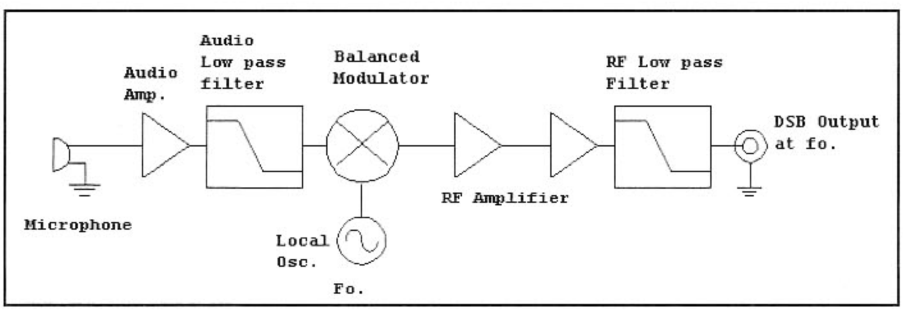


Fig 6.24—A double sideband transmitter.

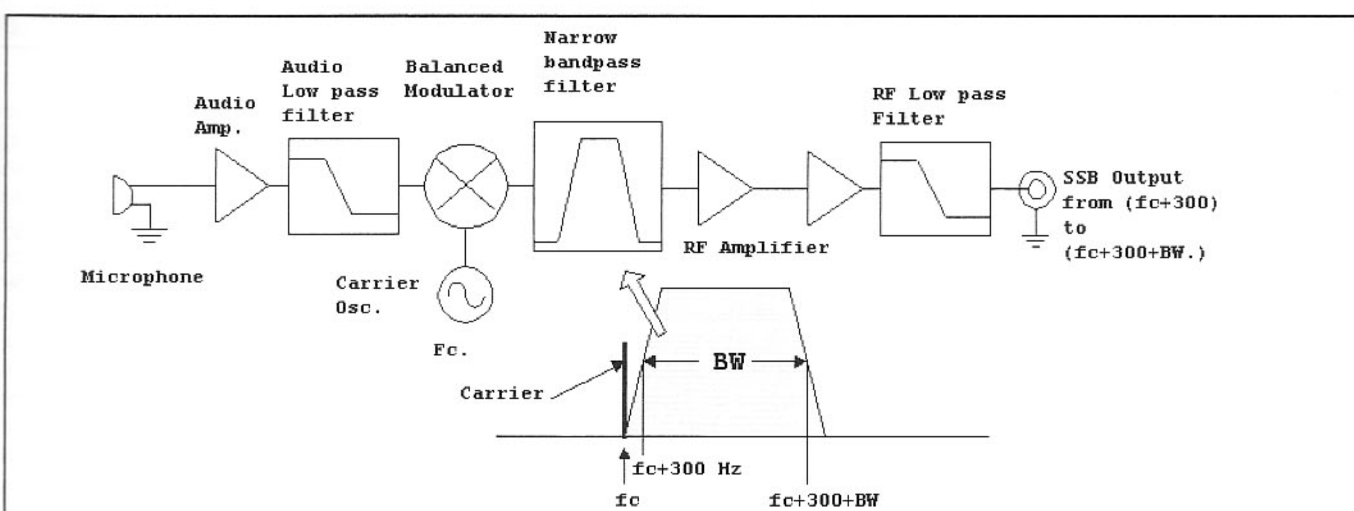


Fig 6.25—A traditional SSB transmitter using the filter method. A narrow filter follows a balanced modulator to remove one of two sidebands present on the DSB output of the modulator.

on the frequency of the suppressed carrier. This is so difficult in practice that a DC receiver is normally not suitable for DSB applications.

The most popular method used to generate SSB is shown in **Fig 6.25**. This is traditionally called the *filter method*, for a narrow bandpass filter is used to select one of two sidebands generated by a balanced modulator. See Figs 6.12 and 6.14. The other dominant way to get SSB is the phasing method, treated in great detail in Chapter 9. The phasing method is based upon mathematics following from the *Trig Identities for Signal Analysis* sidebar earlier in this chapter where multiplication of two sine waves is performed with a doubly balanced mixer.

The SSB transmitter shown in Fig 6.25 has a severe difficulty—it operates at only a single frequency, that of the filter used to generate the sideband. A practical filter-type SSB transmitter topology is presented in **Fig 6.26** where an SSB signal is generated at an intermediate frequency. The resulting SSB is then heterodyned to a desired output frequency where it is bandpass filtered, amplified, low-pass filtered, and applied to an antenna.

Assume the narrow filter used to create the SSB signal at IF is configured to create an upper sideband. For example, let the carrier frequency be 9.000 MHz with a filter extending from 9.0003 to 9.003 MHz, a bandwidth of 2.7 kHz. Set the LO to 37.4 MHz and design the LC bandpass filter to cover 28 to 29 MHz. The resulting signal is then at 28.4 MHz. The transmit mixer has both sum and difference frequency outputs and the LC bandpass has selected the difference, producing a carrier output of $(F_{LO} - F_C)$ for the suppressed carrier. The sideband frequency within the IF will be $F_C+\delta$ where δ is a small positive difference frequency. This value is greater than the carrier, so this is an upper sideband. Because the LC bandpass is configured for a difference output, the signal output will be $(F_{LO} - (F_C+\delta))$, which expands to $(F_{LO} - F_C - \delta)$. This is less than the suppressed and translated carrier at $(F_{LO} - F_C)$, so we now have a lower sideband signal. A designer must always be aware of such inversions. They can be useful for the designer, for crystal filters without ideal symmetry (lower sideband ladder of Chapter 3) are easily built.

The simple direct-conversion receiver in Fig 6.20 is effective in receiving an SSB signal. The difficulty that we encountered with DSB is no longer present, for there is no coherent information in the spectrum formerly occupied by the suppressed sideband to be heterodyned to baseband, eliminating the need for extreme stability. If the BFO is in error by 100 Hz, the received voice may sound unusual, but will still be intelligible.

Even though there is negligible opposite-sideband energy transmitted by a properly designed and adjusted SSB transmitter, that does not mean that the spectrum where that opposite sideband would have been is not used. That spectrum is usually occupied by another SSB station. If a direct-conversion receiver was tuned to a desired signal, the undesired signal would produce completely garbled audio, making simple direct-conversion receivers unsuitable in a densely populated band.

A superheterodyne receiver like that in **Fig 6.27** is usually used to receive SSB. The incoming signal is filtered in a preselector, heterodyned to an IF, and is passed through a bandpass filter. The bandwidth of that filter, usually built with quartz crystals, is wide enough to pass all of the speech spectrum that is transmitted, but little more. A typical SSB receiver will have a bandwidth from 2 to 3 kHz. The filter shape is fairly flat over the passband, but then has steep skirts so that energy in an adjacent "channel" will not interfere with the signal being received. The nar-

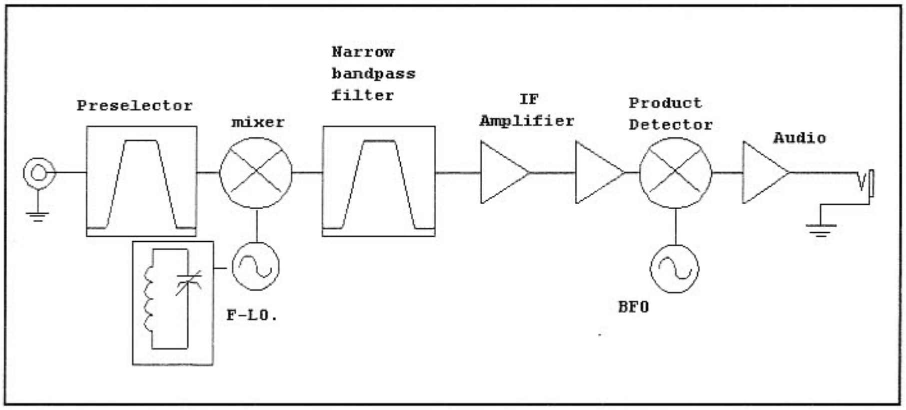


Fig 6.27—A traditional superhet SSB receiver. The response from only one sideband is allowed owing to the narrow-bandwidth crystal filter and the relationship of the BFO frequency to that filter.

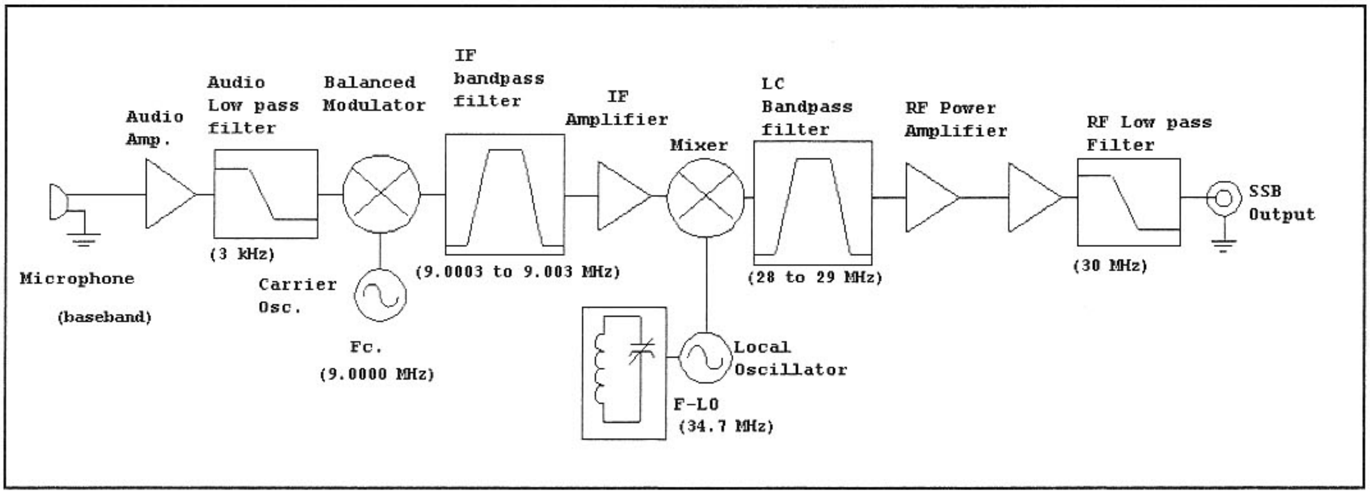


Fig 6.26—A practical filter type SSB transmitter where a mixer translates the output of a fixed-frequency SSB generator to a variety of outputs.

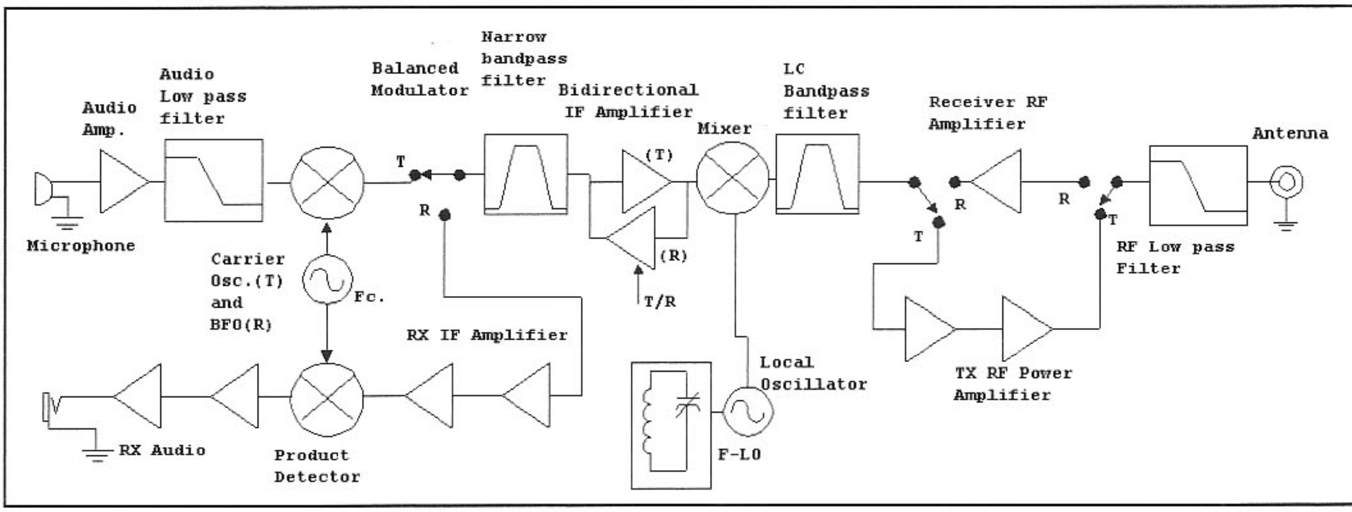


Fig 6.28—An SSB transceiver, a system for both receiving and transmitting an SSB signal. Economy and operating convenience are gained by sharing elements between functions. It is most common to share oscillators and a crystal filter, which is done here. This circuit also shares a mixer between the receiver and transmitter, and uses a bidirectional IF amplifier, a circuit that, with dc switching, will amplify signals moving in either direction. The amplifier circuits are presented later in the text.

row bandpass filter in the SSB receiver is followed by IF amplifiers, a product detector with BFO, and an audio amplifier.

The BFO must be carefully set in the SSB receiver. It should be fixed so that one edge of the filter (a –6 dB point) corresponds to an audio note of about 300 Hz. The other edge will be determined by the filter bandwidth. Typically the BFO is at a point on the filter response that is 20 or 30 dB below the nominal, flat response. The same constraints are used in setting up the carrier oscillator in the filter method transmitter.

The SSB receiver can produce sideband inversion just as we illustrated in the transmitter. The builder/designer should go through the numbers to confirm the behavior. Using popular vernacular, "You do the math."

The SSB receiver, although designed to receive SSB, is also well suited to CW. So long as the filter has good stopband attenuation, the response will also be single signal, as can be confirmed by repeating the experiment we have done with both the direct conversion and the CW superheterodyne. Readjustment of the BFO can compromise the single signal characteristic. An SSB filter is often considered too wide for optimum CW performance, especially in a heavily used band.

The SSB receiver is also well suited for reception of DSB signals. The filter in the receiver rejects one of the sidebands present at the receiver antenna terminal.

Finally, we see that combining Figs 6.26 and 6.27 will result in a transceiver where many circuit elements can be shared between transmit and receive functions. Most transceivers share all oscillators and the crystal filter between the two functions. **Fig 6.28** shows a typical block diagram, here with a design that also shares a mixer between functions, and uses a bidirectional amplifier. No matter what schemes the designer may elect to use, he or she should take care to preserve performance in both transmit and receive functions.

6.1 RECEIVER FUNDAMENTALS

A receiver is characterized by numerous parameters. It must have considerable gain, for the signals we wish to hear are weak. The receiver must also be selective, allowing signals with only slightly differing frequencies to be isolated, received, with useful information processed. The receiver must also include detection in one form or another, producing an output frequency that we can hear. The detection may consist of a rectifier that extracts information about amplitude variations of the radio frequency signal, a discriminator that evaluates signal frequency, or a mixer excited by an LO with a frequency at or very close to the incoming one.

All functions must be executed in a way that does not compromise the information from an original signal. Hence, local oscillators must be stable with respect to the stability of the signals being processed. Filters that provide selectivity must be wide enough to pass the desired information related to the received signals. The gain must be generated without adding excessive noise. Receiver performance specifications generally relate to how well the various required jobs are done.

We begin our receiver investigation with a primitive experiment, an examination of headphones, the generally preferred transducer for converting an electrical signal into sound. (Although we all tend to assume that headphones are optimum, some will argue that a speaker is preferred for weak signals. Individual experiments are required.) The experiment uses a 50-Ω audio-signal source with known output power. See Chapter 7.

A large collection of monaural and stereophonic headphones were examined, old and new. The two ear-pieces were usually operated in series. The typical phones were low (4 Ω) to medium impedance (20 to 35 Ω per side), often representing a reasonable impedance match to the 50-Ω generator. The signal source was adjusted with each headphone set until a signal was just detectable in a quiet room.

The most sensitive headphones were obsolete, inexpensive types consisting of little more than 2-inch diameter speakers mounted next to each ear. Two pair from our collection were capable of producing a detectable output with an available input of –85 dBm. That is, the applied signal was 85 dB below one milliwatt from a 50-Ω audio source.

Several of the phones were nearly as sensitive including some newer Koss TD/65 (90 Ω per side) used for routine communications. The Koss sensitivity was –80 dBm, with better clarity than provided by many others. Several lightweight inexpensive phones (Sony Walkman class) had sensitivity from –60 to –70 dBm. Very old high impedance phones had similar sensitivity, but only after being impedance matched.

A typical listening level will be significantly higher than our *threshold*, but still well below a milliwatt. From these experiments, we will assume that a minimum receiver must be capable of producing an output of –50 dBm for the weakest signal to be encountered. The weakest signals that we normally encounter in HF CW communications are –130 to –140 dBm, indicating a needed gain of around 90 dB. Although this is a subjective result, it represents a design beginning.

Our first simple receiver is shown in **Fig 6.29**. A high-gain audio amplifier with low input and output impedance was built with a gain of 87 dB. The amplifier is combined with an external diode ring mixer, 7-MHz local oscillator and input 7.5-MHz low-pass filter to form a complete direct-conversion receiver. An antenna was connected, producing numerous signals in the 40-m band. The receiver had the usual bright response that we expect from direct-conversion designs. (DC receivers are discussed in much greater detail in Chapter 8.)

The amplifier did more than make the signals louder. It generated noise, apparent when power was first applied. While the noise was not so loud as to be objectionable, it would obscure some weak signals we expected to hear. When a signal generator was attached and adjusted, the best we could hear was about –130 dBm, well away from the –140 dBm expected with many simple direct-conversion receivers.

Why is this receiver so noisy? Little noise is generated in the first element in the system, the diode ring mixer, a passive element without gain. Rather, the noise in this design is generated in the amplifier that follows the mixer.

This noise is not the result of a poor op-amp choice, but a poor design with respect to noise. Negative feedback in an amplifier reduces input impedance. The impedance looking into the inverting amplifier input of a 5532, *with a 5.6-kΩ feedback resistor*, is about 1 Ω. We modify this with an added series 56-Ω resistor to generate a 57-Ω impedance to approximately match the mixer, a requirement for low mixer distortion. The available signals from the mixer are all absorbed, but only the fraction of the power delivered to the 1-Ω input is amplified. The remaining power is merely converted to heat. All of the available noise current from the input resistor flows in the op-amp input. The result is poor noise figure, a degradation in the input signal-to-noise ratio in the process of amplification. This amplifier is contrasted with the popular design where the first audio amplifier is a common-base bipolar transistor. In that design, almost all of the available power is presented to the active device.

The fundamental receiver parameter used to characterize the noise that limits sensitivity is noise figure (NF), introduced in section 2.6. NF is a measure of the degradation of signal-to-noise ratio by a processing element, be it a complete receiver or a single stage.

Let's assume that we wish to infer receiver noise figure by driving the receiver with a signal generator. The input *signal* power is established by the *available* power from the generator. (This may differ from the actual power delivered to the source.)

Input available noise power is that available from whatever resistor might be attached to the input, given by

$$P_n = k\,T\,B \qquad \textbf{Eq 6.9}$$

where k is Boltzmann's constant, T is temperature in kelvins, and B is the bandwidth in Hz in which the noise is observed. The standard temperature used for noise determinations is 290 K, close to a normal room temperature. This noise power is independent of the resistance. The noise power is distributed uniformly over all frequencies. If receiver bandwidth is increased, the noise power increases accordingly.

Attaching a room temperature resistor to the input of a receiver provides a source of noise. The signal generator, with its output resistance, will also serve this function. If the generator level is changed by attenuation, output resistance seen by the receiver remains constant to maintain a constant available noise power.

The output signal and noise are measured by attaching a load (usually a speaker or earphones) monitored by an ac voltmeter, ideally one that provides a true rms response. Noise output can be monitored alone by momentarily turning the generator off. When the signal is again applied, along with the input noise, the

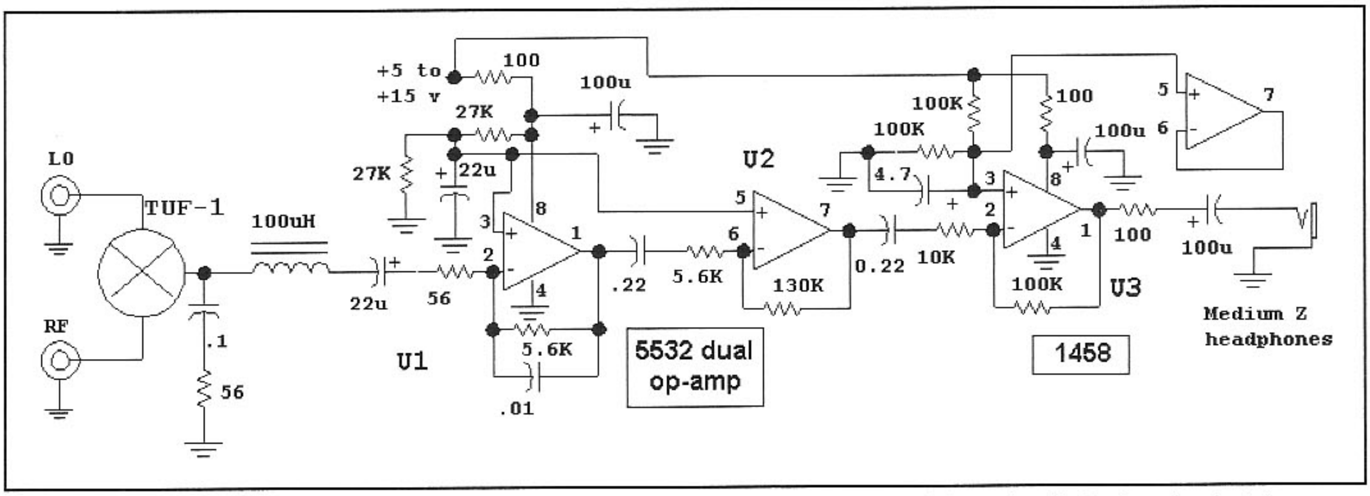


Fig 6.29—A basic direct-conversion receiver. An audio amplifier with a gain of 87 dB follows the diode ring. See text for discussion.

output will be an output signal + noise power. An output signal-to-noise ratio can then be calculated. Noise figure can then be calculated.

Noise figure is usually measured with a noise source of known power, usually well above the noise power available from a 290-K resistor. See Section 2.6 and noise measurements in Chapter 7.

The greatest virtue of noise figure as a receiver parameter is that it is bandwidth invariant. If we increase the bandwidth during a NF measurement, we will process more noise in the receiver. But the output will also increase in proportion, leaving the noise gain, the ratio of output noise to input noise, a constant.

Another measure of receiver sensitivity is *minimum discernable signal*, or MDS. This is the available input signal from a generator that will cause the output power to increase by 3 dB over what is present without the applied signal. In this condition the signal and the noise have equal output powers.

MDS is directly related to room temperature NF by

$$\text{MDS (dBm)} = -174 \text{ dBm} + \text{NF(dB)} + 10 \log(B) \qquad \textbf{Eq 6.10}$$

We measured the noise figure of one of our receivers to be 7 dB with a nominal bandwidth of 500 Hz. **Eq 6.10** then predicts MDS of –140 dBm. A direct measurement of MDS where we look for a 3-dB increase in output above the noise floor as we apply signal produced an almost identical result of –141 dBm.

It is interesting to listen to this receiver with the signal generator attached. We find that we can hear the MDS, but not much further into the noise.

We now increase the receiver bandwidth to 2.4 kHz by switching in a new crystal filter, increasing the bandwidth factor in Eq 6.10 to 33.8 dB. MDS becomes –133.2 dBm with a 7-dB noise figure. A measurement will usually confirm this number. Noise measurement in a wider bandwidth is generally easier than it is with narrow band systems owing to less fluctuation in the meter movement. But major errors can and often do occur as a result of slight gain variations with frequency in either the IF or the receiver audio circuitry— errors that generate a narrower noise bandwidth than expected. A direct NF measurement is generally preferred over one of MDS, where only a ratio of two noise powers must be determined.

An ideal receiver with measured MDS commensurate with the filter BW will often let a listener hear signals that are much weaker than indicated by the MDS. Why? The human ear and brain are a vital part of the communications system and they are capable of acting like a filter of considerably narrower bandwidth than the voice bandwidth of the receiver. This effect is observed with both wide bandwidth superheterodyne designs and direct conversion receivers. Indeed, many seasoned weak-signal VHF enthusiasts including moonbounce specialists normally use wider SSB-bandwidth filters.

Many argue that noise figure is rarely a significant receiver parameter, especially for HF reception. An NF of 10 or 12 dB at 28 MHz, with much higher numbers at lower frequencies will usually provide as much sensitivity as one can use. A practical receiver test is very simple: While listening to background noise on a band, disconnect the antenna. If the noise drops significantly, the receiver NF is as good as it needs to be.

NF is much more important as a design parameter. The essence of modern receiver design is a quest for dynamic range, and NF specifies the lower end of such a range.

Equation 6.10 relates NF to MDS, suggesting that little is to be gained with extremely low noise figures. Consider, for example, a receiver with a 200-Hz bandwidth and 3-dB NF. Equation 6.10 predicts MDS of –148 dBm. Dropping noise figure to a spectacular 0.5 dB results in only a 2.5 dB sensitivity improvement to –150.5 dBm. This is what a careful MDS measurement would demonstrate. But in reality, the practical improvement could be much more than this. The dilemma comes about when we pick a noise temperature of 290 K for our standard. This choice defined the "input" noise in Eq 6.9. But if the input noise resulted not from the 290 K resistor related to our measurement, but from an antenna pointed at a quiet part of the sky, the input noise might well relate to a resistor with a temperature as low as 20 K. A more refined calculation would show that MDS could be as low as –158 dBm for this example. A related concept of *noise temperature* was used to obtain this result.[3]

The noise factor of a two-stage cascade is

$$F = F_1 + \frac{(F_2 - 1)}{G_1} \qquad \textbf{Eq 6.11}$$

where F is the net noise factor, F_1 and F_2 are the noise factors for the first and second stage, and G_1 is the available power gain for the first stage. All numbers are power ratios and not dB values.

Consider an example shown in **Fig 6.30**. The first amplifier has a gain of 12 dB and a 3-dB NF while the second stage has an 8-dB NF. Related power ratios are $F_1 = 2$, $F_2 = 6.3$, and $G_1 = 15.8$, yielding $F = 2.34$, or $NF_{NET} = 3.7$ dB. The first stage noise performance dominates in this example. Once we know how to evaluate a cascade of two stages, we can apply the process in steps to evaluate an arbitrary cascade, including an entire receiver front end.

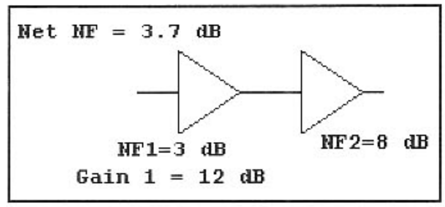


Fig 6.30—Example calculation for noise figure of a cascade of two stages.

Many of the circuit blocks that we used in receivers and transmitters are room temperature passive parts with no gain elements. These include not only the popular passive switching-mode mixers, but attenuators and filters. Generally, the NF of a passive circuit equals the insertion loss of that circuit. Hence, a diode ring mixer with a 6 dB conversion loss (gain = –6 dB) will have a 6-dB NF. A bandpass filter with an insertion loss of 2 dB will, similarly, have NF = 2 dB and Gain = –2 dB.

Fig 6.31 illustrates a receiver front end where several elements contribute to the noise figure. This circuit will include an RF amplifier, for we are interested in relatively low noise figure. Two bandpass filters are used. The first is a single resonator ahead of the RF amplifier while the second is a double tuned circuit. A diode-ring mixer is followed by a feedback amplifier that uses a bipolar transistor with high dc emitter current. The overall cascade has net gain of 15 dB and a net noise figure of 7.1 dB.

Front-end bandpass filters usually do not impact overall noise figure. In the receiver example just presented the system bandwidth is determined by a crystal filter that follows the attenuator. This filter is usually narrow (3 kHz or less) and the two L/C bandpass filters shown as the first and third elements in the cascade are wide (a few hundred kHz). The crystal filter then sets the overall response. The bandpass filters in the cascade have no more impact on noise figure than an attenuator would. The situation would be considerably different if the crystal filter was replaced with a wide L/C filter with equal or wider bandwidth than those in the front end.[4]

Some RF Amplifiers and Attenuators

Many modern HF receivers use no RF amplifier, for adequate noise figure can be obtained without it. Most commercial gear has a NF of 10 to 12 dB at and below 30 MHz. A practical sensitivity test was outlined above. There are some situations where an RF amplifier can be useful, even at HF. This is especially true at 21 and 28 MHz during periods of marginal propagation. It is then useful to switch a low noise amplifier into the signal path. Such an amplifier is not normally needed and should not be used merely to make signals louder. We will illustrate a few circuits that we have built, used, and measured.

A favorite RF amplifier is a common gate JFET circuit. A J310 is used for HF applications, while a U310 is preferred for VHF and UHF. (The surface mounted version of the J310 should be excellent for both!) The basic amplifier is shown in **Fig 6.32**. The FET is biased for a current of 12 to 14 mA, determined by FET I_{DSS} and source resistor. The gain is only about 2 dB with this amplifier if the drain load resistor, R, is set at 680 Ω. In spite of the low gain, the amplifier is still very useful. It has a good input and output impedance match, so offers a good interface to filters and mixers. It is most useful for the excellent reverse isolation. The reverse gain (S12) was measured as –43 dB. This is an excellent amplifier for use with direct conversion receivers when attempting to reduce *tunable hum*, discussed in Chapter 8. The circuit is turned on with $V_{CONTROL}$ = +5 or so. The gain is reduced by 40 dB when turned off.

Gain goes up to 6.5 dB in this circuit when the drain load resistor is eliminated. In that configuration, the third order output intercept was +28 dBm, measured at 14 MHz with fairly flat gain up to 50 MHz. (Intercepts were introduced in section 2.6.) Lower frequency performance is improved with a larger inductance RF choke.

Higher gain is available if the output is tuned, shown in **Fig 6.33**. The output drain resistance for this amplifier is close to 10 kΩ, allowing it to form one termination of a bandpass filter. The variation shown with a single tuned output circuit has a typical gain of 12 to 13 dB with a 50-Ω load. The 50-Ω input match is a 15-dB return loss. Noise figure was 5.0 dB at 21 MHz.

This amplifier has no tuning at the input, for C1 and L1 are both large. Lower noise figure is often obtained with a suitable input network, one that usually degrades input impedance match. The designer can generally design an input network that will present a needed impedance to the input if the value for optimum NF is known. We didn't have that data for the J310, but were able to find hints. Specifically, Chip Angle, N6CA, has built amplifiers with the U310 for several VHF bands. The U310 is the same chip, but is packaged in a metal can with the gate attached to the can. We were able to analyze his circuits and scale his input networks to lower frequency. The result was an amplifier with a measured 1.5-dB NF, but with a poor input match and gain of only 12 dB. This occurred at 21 MHz with L1 = 1.26 µH and C1 = 39 pF. The noise match point that we inferred was Γ_{OPT} = 0.89 at 7°.[5]

A common source JFET should be capable of low noise performance. The practical difficulty in building such a circuit is often stability. Cascode connected JFETs should be considered. Neutralization is also practical, although rarely used.

The humble source follower should not be discounted as a low-noise amplifier. A suitable circuit is shown in **Fig 6.34**. A link-coupled input drives the gate through

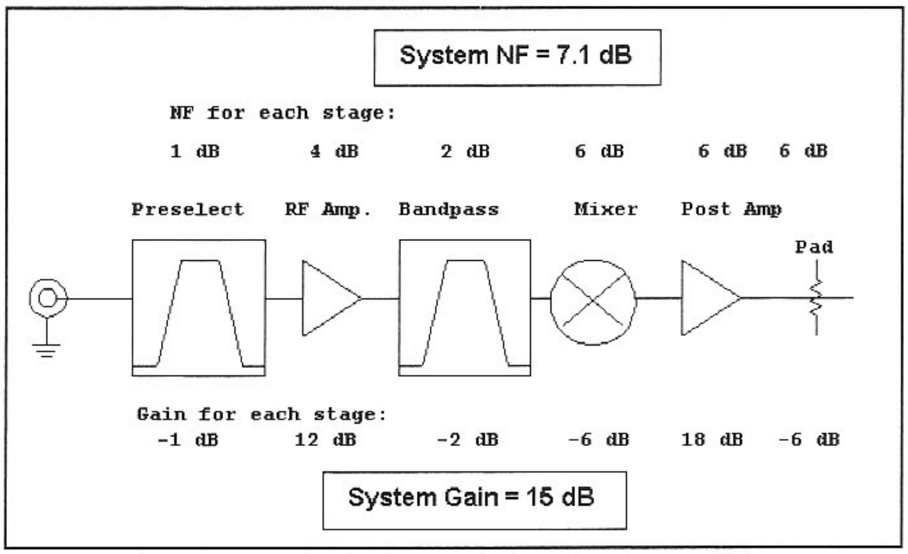


Fig 6.31—A six-stage cascade showing a typical receiver front end. The stages consist of a wide filter, an RF amplifier, a steeper skirted bandpass filter, a diode ring mixer, a post-mixer amplifier, and finally, a 6-dB attenuator.

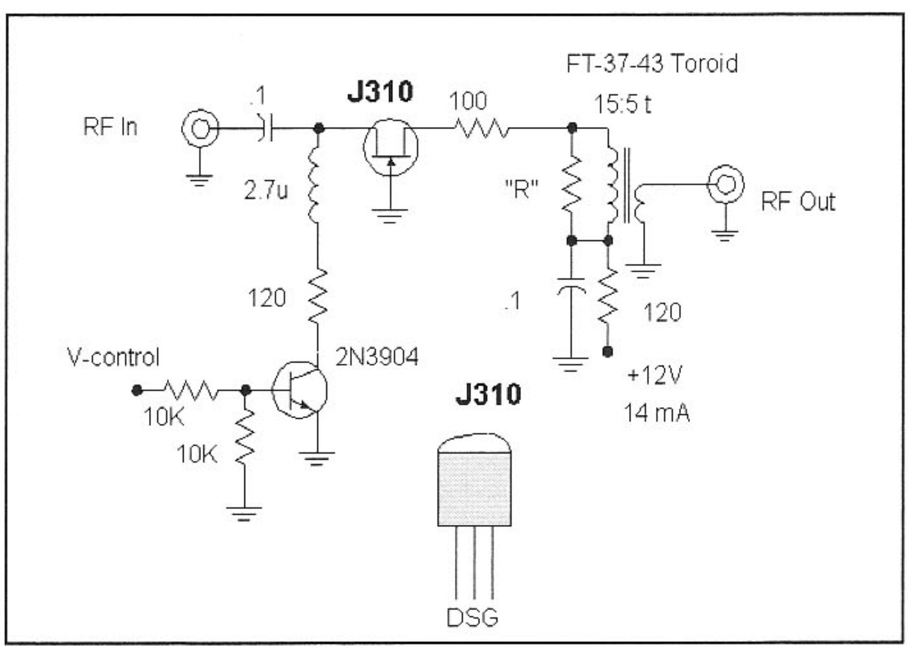
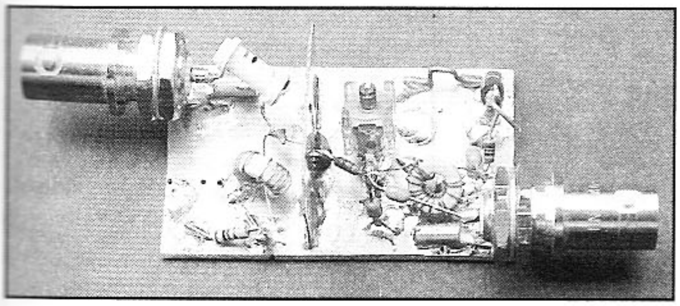


Fig 6.32—A common-gate amplifier using a JFET. The 100-Ω resistor at the drain suppresses UHF oscillations. See text regarding the drain load resistor, "R."



Close up of common-gate low-noise amplifier using a J310.

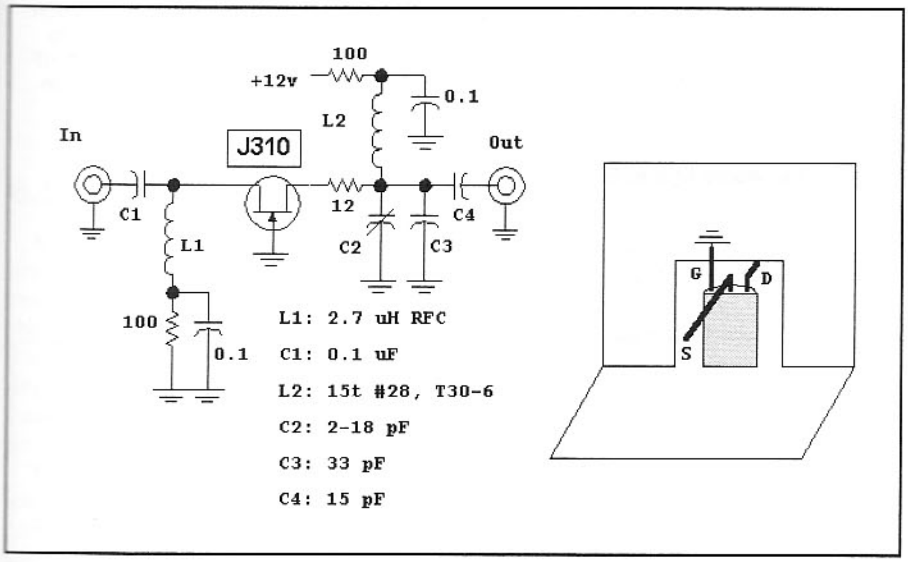


Fig 6.33—A 21-MHz RF amplifier. This circuit, with the values shown, provides a gain of 14 dB with a 5-dB noise figure. Redesign of the input network produced a NF of 1. 5 dB, but with reduced gain of 12 dB. A shield between the source input circuit and the output drain circuit is advised, especially if high-Q solenoid coils are used. It is generally not required when using toroids, although the gate should be grounded with short lead length.

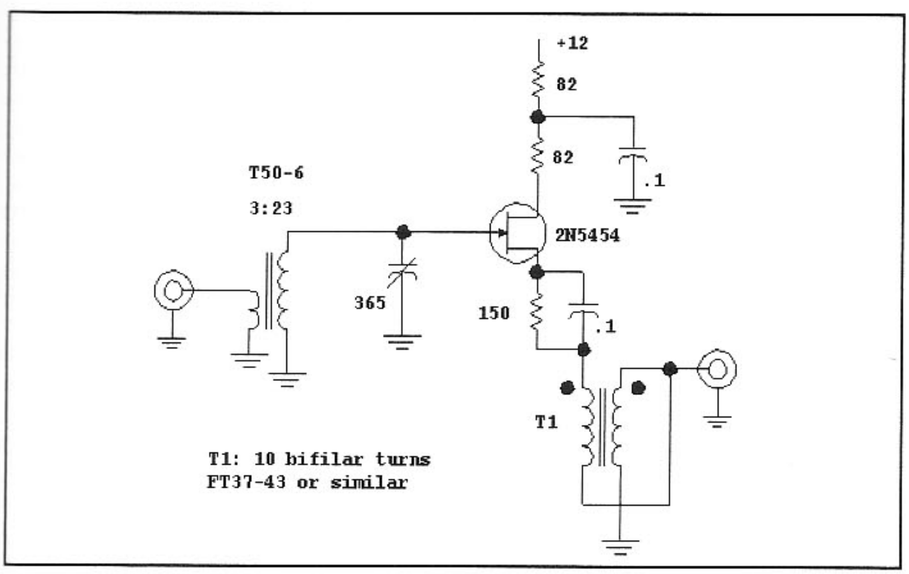


Fig 6.34—Source follower functioning as a low-noise amplifier. The drain resistor serves to suppress UHF parasitic oscillations. The components shown will tune from 6 to 22 MHz.

a tuned circuit with a sizable impedance transformation. The output is then extracted from the source with a ferrite transformer. An example amplifier measured gain of 11 dB with NF = 1.9 dB. No stability problems were noted. The output match was good, although the input is severely mismatched.

Dual gate MOSFETs make excellent RF amplifiers as shown in **Fig 6.35**. This circuit was tuned for both the 21 and the 14 MHz bands with similar results obtain with each. The 14-MHz circuit is shown. A pi-network transforms the 50-Ω source to "look like" an impedance of 2000 Ω at gate-1 of the FET. The network was designed for a Q of 10 and used an existing 2.7-μH RFC. The drain is matched with a ferrite transformer followed by a 6-dB pad. This amplifier provide a gain of 16.5 dB (including the loss of the pad) with a 3.6-dB noise figure. The circuit had an output intercept of +12.5 dBm.

The gain is often excessive with dual-gate MOSFETs. Better overall receiver dynamic range is afforded by reduced gain. The pad helps, but it compromises the amplifier intercept performance, for the amplifier must have a 6 dB higher intercept to get the quoted value. Even the 1200-Ω drain load resistor compromises IMD performance. Source degeneration provides an alternative, achieved by disconnecting the source bypass capacitor. Gain dropped to 9 dB for the circuit shown (with pad), and the noise figure increased slightly to 4.1 dB with OIP3 = +14 dBm.

The low-Q inductor used in the input pi-network compromises the noise figure. Replacing it with a toroid dropped the 3.6-dB NF to 2.5 dB. Even lower values are available if a higher impedance is chosen for the pi network. The input match is very poor with all variations of this amplifier.

Many of the feedback amplifiers described throughout this text are suitable for RF amplifier application. The noise figures can be in the 3 dB area with some transistors. For example, we have measured a 3-dB NF with a 2SC1252 operating with 20-mA emitter current.

The modern trend in amateur receivers is to include an RF amplifier that can be switched into the circuit if needed. That switching is best done with relays, although PIN diodes can also be used if done with extreme care to avoid second-order intermodulation. It is also common to include one or two attenuators that can be switched ahead of a receiver. An attenuator equally decreases the strength of all signals reaching the front end. Often the signals we are trying to copy are strong enough that an attenuation of 10 dB will not cause a sensitivity problem. The real

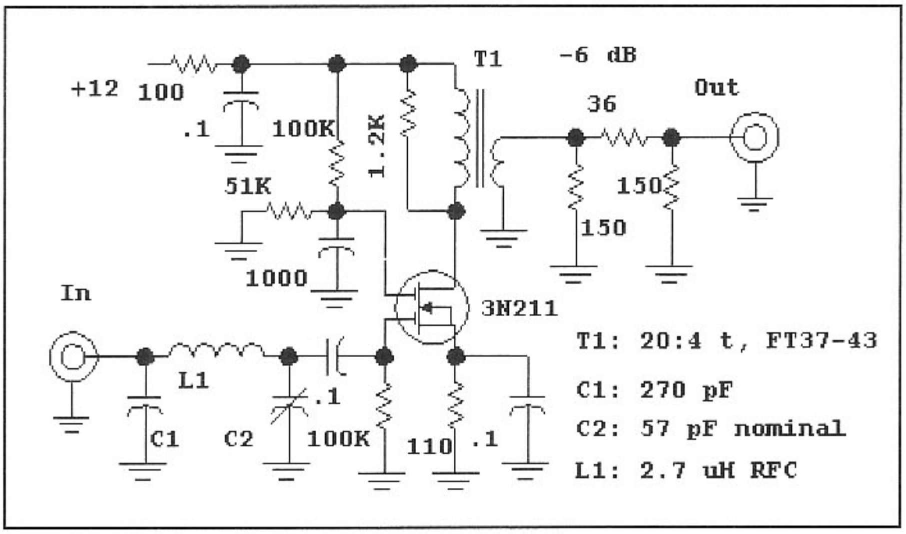


Fig 6.35—Dual-gate MOSFET RF amplifier. This version used an RF choke at L1 with Qu = 50. A higher Q inductor will drop the amplifier noise figure. See text.

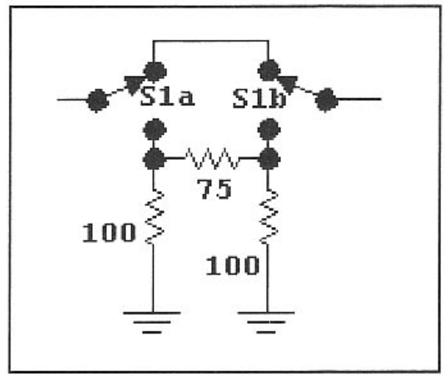


Fig 6.36—A 50-Ω, 10-dB pad using standard resistors and a toggle switch. Short lead lengths should be used to provide good performance over the HF region. Relay switching could also be used.

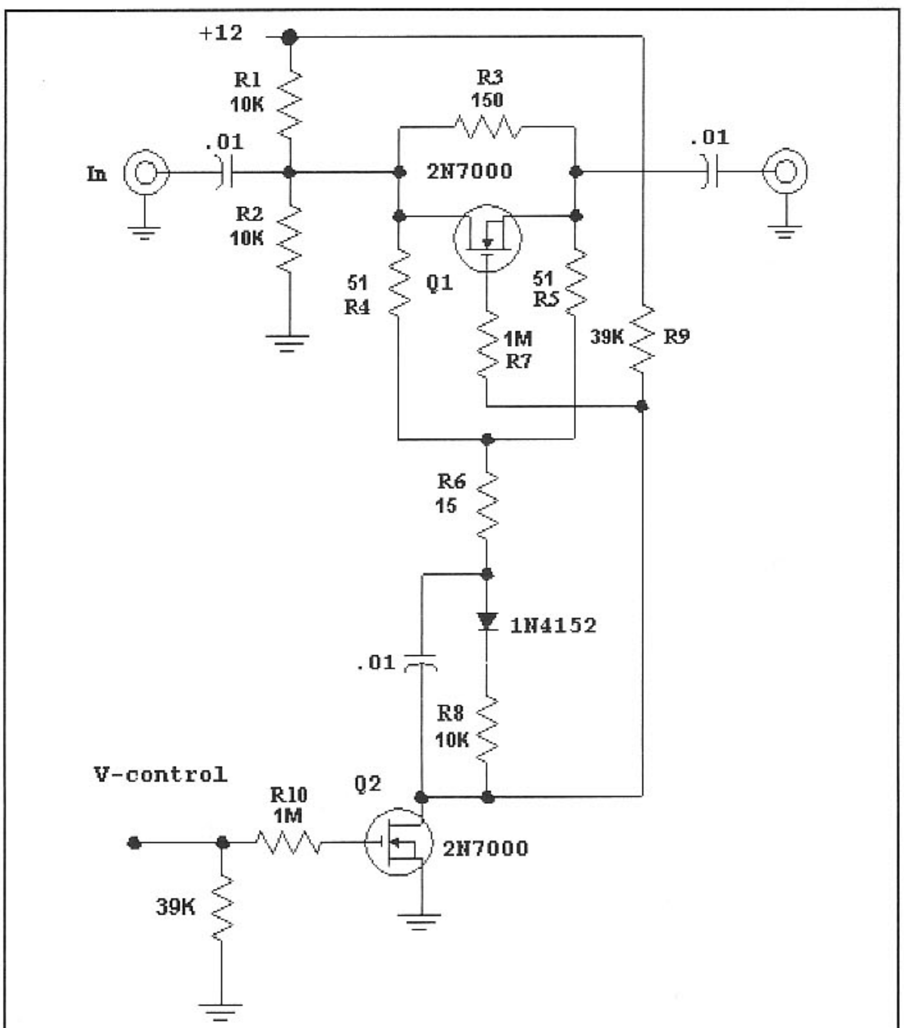


Fig 6.37—A 10-dB pad using electronic switching. A bridged-Tee pad (R3, 4, 5, 6) is switched with low-cost MOSFETs. During thru operation, Q1 is on while Q2 is off. Q2 comes on during attenuated operation. Current consumption is about 1 mA.

Dual-Gate MOSFET Availability

The dual gate MOSFET was a very popular consumer device from 1970 to 1980 and was readily available from a number of sources. The part provides low noise, moderate to high amplifier intercepts, and reasonable power consumption. They also offer good AGC performance. They are now more difficult to obtain than they were in the past.

But Dual-Gate MOSFETs are still available. Several suppliers in Japan continue to manufacture a variety of components. The NEC 3SK131 is an excellent part, but it is available only in a surface-mount form.

Phillips manufactures a large variety of dual-gate devices. These are often listed in some US catalogs. Again, these devices appear predominantly in SMT format.

Generally, it is quite straightforward to substitute one MOSFET in a circuit designed for another. There may be a few different biasing details, but these can be extracted from data sheets, which are generally available on the World Wide Web. Experiments may be required if data is not available.

Finally, most circuits using dual-gate MOSFETs can be built with N-channel JFETs in a cascode configuration. This is illustrated in the IF amplifier part of this chapter.

utility of an attenuator is that most distortions drop faster with signal strength than the signals themselves. Hence, if strong signals within a band are causing gain compression or intermodulation distortion, a small decrease in the strength of the offending signals can completely eliminate the problems.

A passive attenuator is shown in **Fig 6.36.** The typical miniature toggle switch works well for pads of this sort with 10 to 20 dB attenuation.

A scheme is shown in **Fig 6.37** where 2N7000 MOSFETs replace a mechanical switch. The FETs are both RF and dc switches in this application. A pair of resistors, R1 and R2, create a 6-V supply. R9 will bias Q1 into conduction in the low attenuation position with the Q2 gate low. The Q1 channel is then held at 6 V. But when Q2 is turned on, R6 is switched to RF ground. The dc potentials also change to turn Q1 off. We measured an insertion loss of 0.38 dB with this circuit, with a 10 dB gain step. The 14-MHz IIP3 exceeded +35 dBm during low attenuation, and was +26.5 dBm in the attenuation position.

6.2 IF AMPLIFIERS AND AGC

A superheterodyne receiver uses an intermediate frequency between an initial mixer and detector, primarily as a means for obtaining selectivity. It is this selectivity that selects the sideband received, or provides single-signal CW reception. The IF is the usual place for adding and controlling receiver gain through voltage control.

Voltage-controlled gain is usually realized with integrated circuits. But the most popular parts are slowly, but surely disappearing as the consumer markets evolve toward larger scales of integration. Accordingly, this section contains two goals. First, we hope to illustrate some IF amplifier methods that can be applied before the semiconductors disappear. And of greater import, we hope to illustrate some methods that others can use to develop their own IF circuits.

Early superhets used tuned IF amplifiers, providing selectivity throughout the amplifier while modern designs usually use local filtering. Signals exit a mixer, pass through a filter (usually built from quartz crystals) to reach the IF amplifier. As such, the IF amplifiers are protected from strong out of band signals, the sources of performance-compromising distortions. Reasonable linearity is still useful to preserve low *in-band* distortion.

The importance of IF noise figure is illustrated in **Fig 6.38** where we calculate receiver noise figure for a system with the front end treated earlier. The front end had a 7.1-dB NF with total gain of 15 dB. We start with a lossy crystal filter with 10-dB insertion loss and find that overall system noise figure is always above 10 dB, even if the IF NF is as low as 3 dB. A more realistic filter loss of 3 dB provides an overall NF in the 8 to 9 dB region, even with fairly noisy IF amplifiers. IF Amplifier noise figure, including the loss of any filter ahead of it, can have a major impact on system performance!

The distortion properties of IF amplifiers will become more important in emerging receiver topologies. These receivers, largely based upon digital signal processing, use wide IF filters followed by an IF amplifier driving an analog-to-digital converter. The receiver is then completed through digital calculations. Distortion within the IF amplifier and the A-to-D converter become vital.

In the following pages we will consider a number of IF amplifier circuits. We will examine them for noise figure, gain, gain variation, and IMD. Some complete IF systems will be shown.

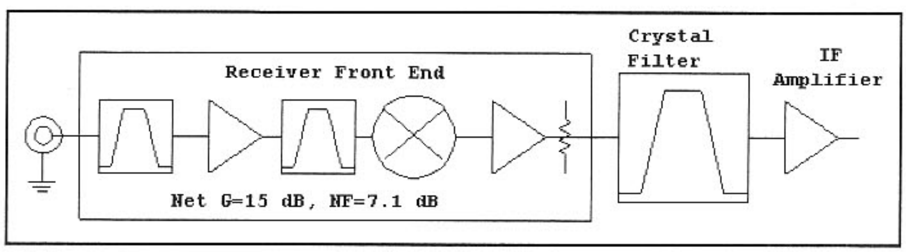


Fig 6.38—The front end presented earlier in Fig 6.31 is combined with a crystal filter of known insertion loss, followed by an IF amplifier. If the filter has a 10-dB IL, a 7-dB IF noise figure will produce a system NF of 10.6 dB.

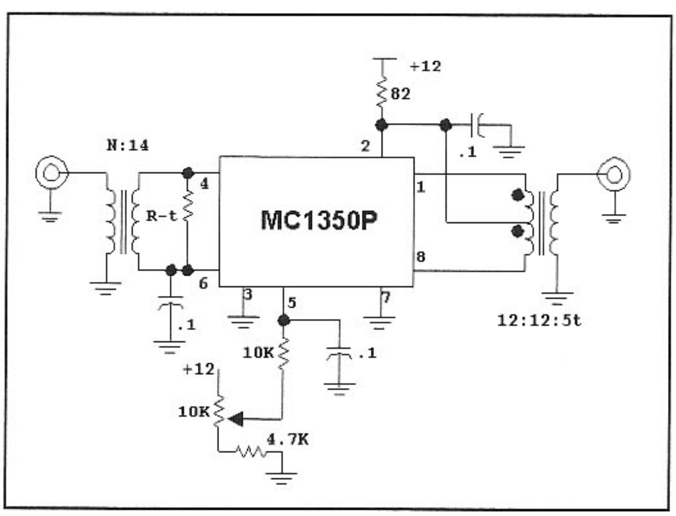


Fig 6.39—Amplifier for examination of the MC1350P. Gain is reduced by over 60 dB by increasing the dc current into pin 5.

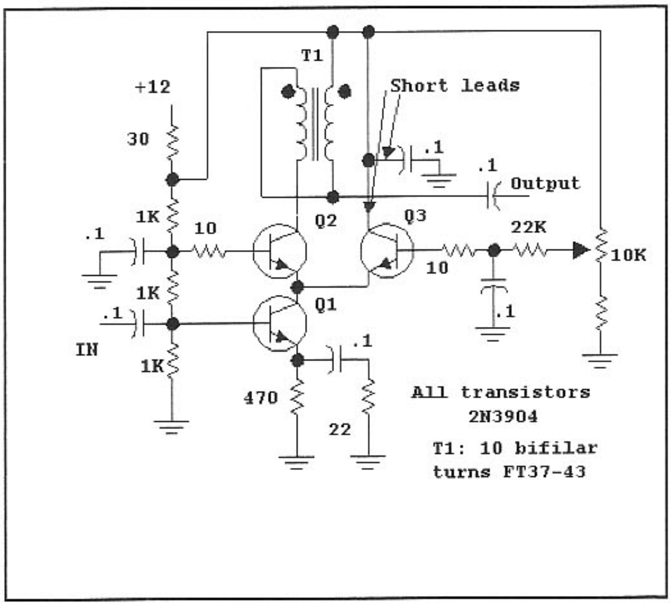


Fig 6.40—Bipolar transistor discrete IF amplifier with gain reduction using the same mechanism as used in the MC1350P. Control range was 70 dB, experimentally controlled with a 10-kΩ manual IF gain.

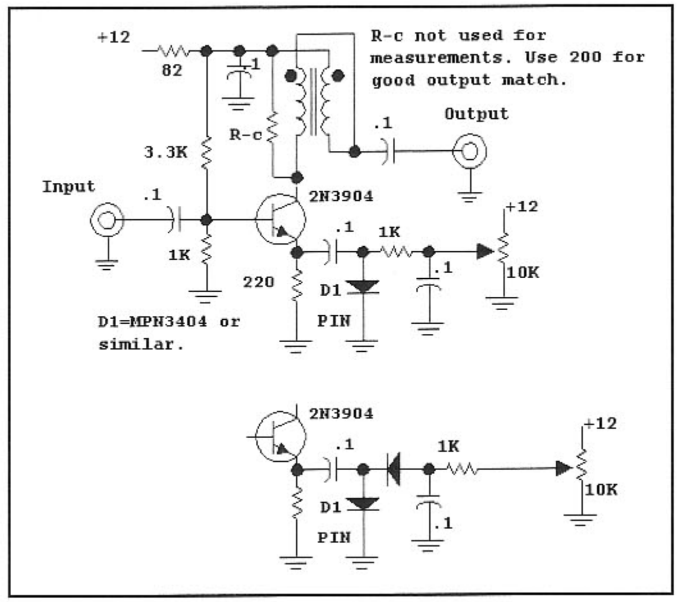


Fig 6.41—Simple gain-controlled amplifier. The inset shows the use of two PIN diodes to increase the control range slightly with the same control current. Many diode types work with this circuit; see text. The 10-kΩ pot establishes manual IF gain.

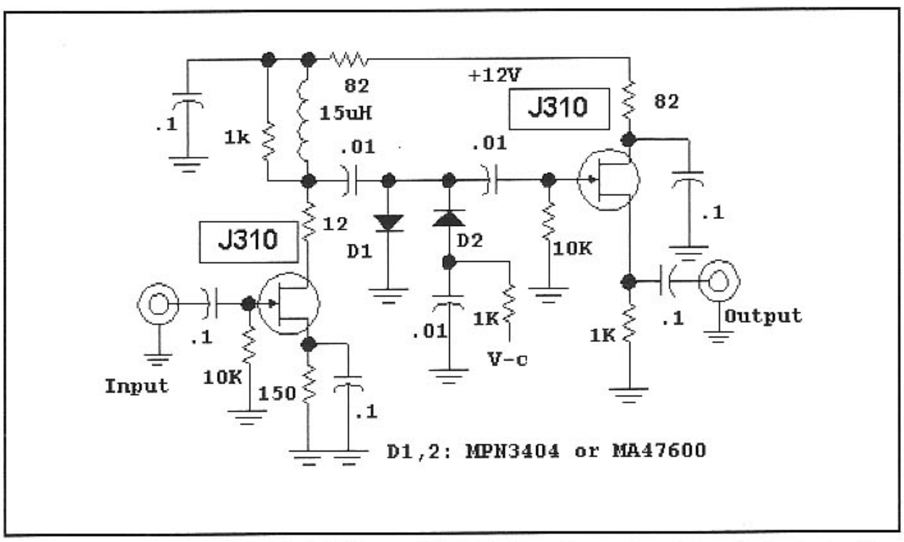


Fig 6.42—AGC amplifier with FETs and PIN diodes. Manual gain is controlled with the 10-kΩ pot.

The first amplifier presented uses the popular Motorola MC1350P. Although this device is, at this writing, slated to be discontinued, it will probably be available for a while from distributors, or from surplus. The methods used in the 1350 can also be realized with discrete components. The MC1350P test circuit is shown in **Fig 6.39**.

The input between pins 4 and 6 (the input differential pair) looks like a 2700-Ω resistance paralleled by 8 pF at 10 MHz. This was approximately matched with a 2:14 turn ferrite transformer with no R_T used. The output, consisting of open collectors of a differential transistor pair, was terminated with a ferrite transformer, producing a 10-MHz gain of 47 dB. The gain-control range was over 65 dB. The noise figure was 5.1 dB, but degraded to 10.3 dB when the gain was reduced by 10 dB.

The relatively high input impedance is rarely suitable for termination of crystal filters. Extra resistance, R_T, is often paralleled with the input to achieve a needed impedance. R_T = 620 Ω produced a net impedance near 500 Ω, a common value needed to terminate crystal filters. This was matched to 50 Ω with a 4:14 turn ratio ferrite transformer. Gain dropped to 39 dB, as expected. Full gain noise figure was 6.6 dB, increasing to 14.1 dB with 10-dB gain reduction. Changing R_T to 220 Ω with a new matching transformer produced further degradation.

Fig 6.40 shows a breadboard circuit with internal workings similar to the '1350, although the IC has additional differential input and output buffering. The Q1 collector current passes through Q2 that operates as a common base amplifier. Gain is reduced by increasing the base bias on Q3 so that emitter current and signal current are both *robbed* from Q2. This circuit provided measured gain of 16.5 dB, 70-dB gain-control range, and good IMD performance. Noise figure was 7 dB at maximum gain, but degrading to 19 dB with 10-dB gain reduction. We noted a noise peak when Q2 and Q3 conducted equal currents. Careful examination revealed the same effect with the MC1350.

A bipolar transistor circuit using PIN-diode emitter degeneration is shown in **Fig 6.41**. Although simple, this circuit offers promise. Gain at 10 MHz was measured at 30 dB with a MPN3404 PIN diode. Gain control range was also 30 dB. A builder may wish to load the collector with a resistor to produce slightly less gain per stage with a better output impedance match. Noise figure was 5.2 dB and hardly changed with a 10-dB gain reduction. Several diode types were evaluated in this circuit. Power rectifiers such as the 1N4006 or 1N647 worked well with low distortion, although large diode capacitance reduced gain control range. While a 1N4152 worked, IMD was severe at some currents.

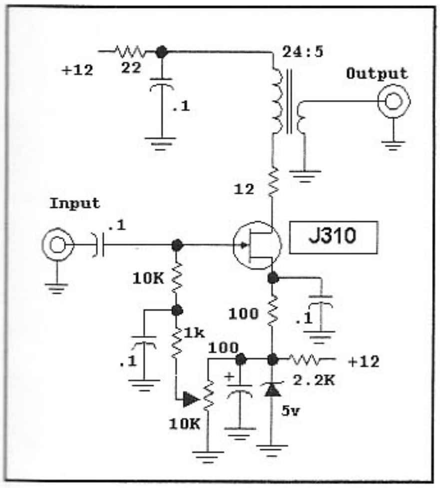


Fig 6.43—A single JFET is biased toward pinchoff with the reverse bias developed across the Zener diode. This amplifier offers 13.5 dB gain and a 37-dB gain range. The transformer, wound on an FT37-43, was available on the bench at the time of testing. The 10-kΩ pot sets gain.

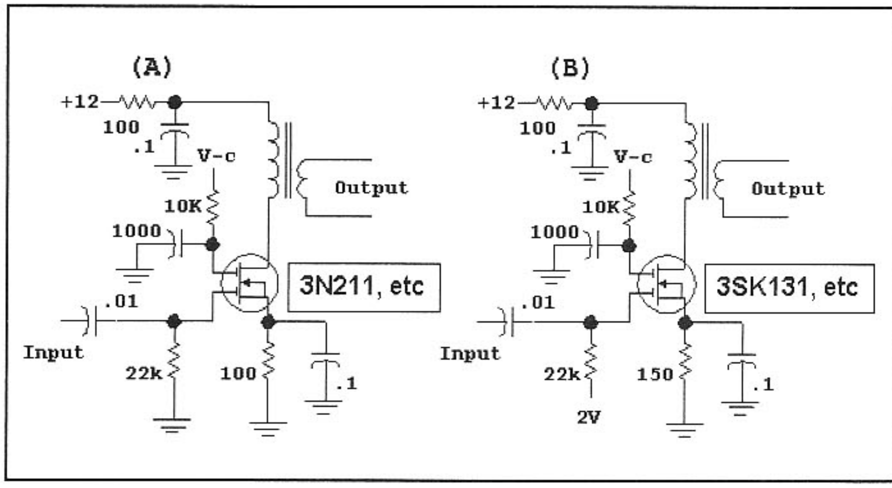


Fig 6.44—Two variations of a basic dual-gate MOSFET amplifier with variable gain. The circuit at (B) has the larger gain variation. The labeling of FETs is arbitrary, for these circuits are intended to be generic. The 3SK131, an SMT device from NEC is popular and is recommended.

PIN diodes can be combined with FETs for interesting IF amplifiers. **Fig 6.42** shows an amplifier where a FET serves as a common-source amplifier, followed by shunt PIN diodes driving a source-follower output. Output could also be obtained from the first FET drain through a transformer. This topology has many possibilities. Gain was 13 dB with a 60-dB gain range when the FET was driven from 50 Ω. NF was poor in this topology, but became very good when the first FET was driven from a higher impedance via an L-network. Gain also increased.

The performance of this amplifier is critically dependent on diode type. IMD was very low with MA47600 diodes from Microwave Associates. Experiments with devices from HP are recommended using the 5082-3080, or HSMP-3814. We observed some gain compression in this circuit with the MPN3404.

A very simple JFET IF amplifier is shown in **Fig 6.43** where gain is reduced as gate bias moves toward pinchoff. This circuit is configured (with a Zener diode) for a single power supply, although a negative supply for the biasing would be preferred. The circuit shown barely has adequate power supply voltage, but basic performance is excellent. Initial gain is 13.5 dB (at 10 MHz) with a smooth control range of 37 dB. Noise figure at maximum gain was 4.6 dB, increasing to 7.6 dB with 10 dB of gain reduction. Input intercept was +10 dBm at maximum gain, dropping eventually to –7 dBm as gain drops. However, intercept degrades

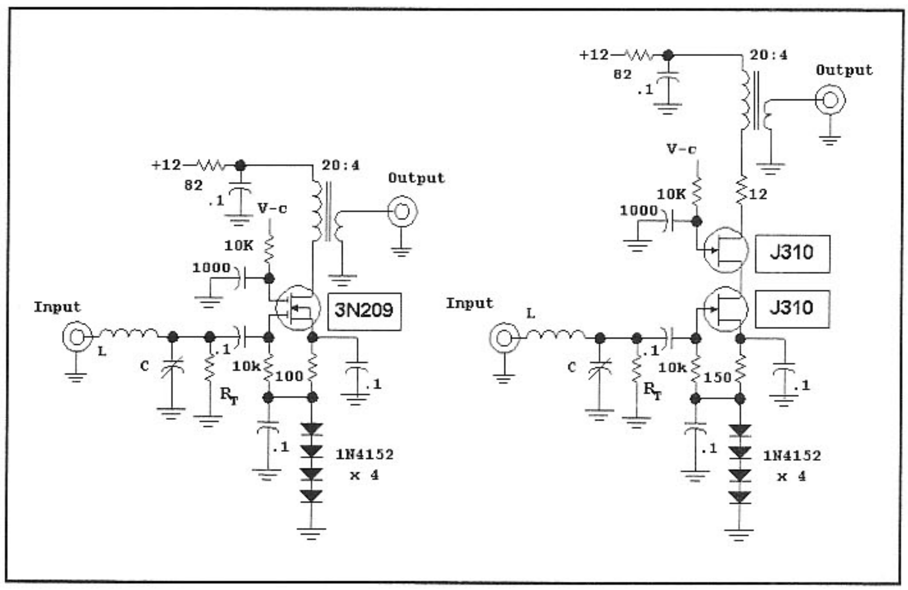
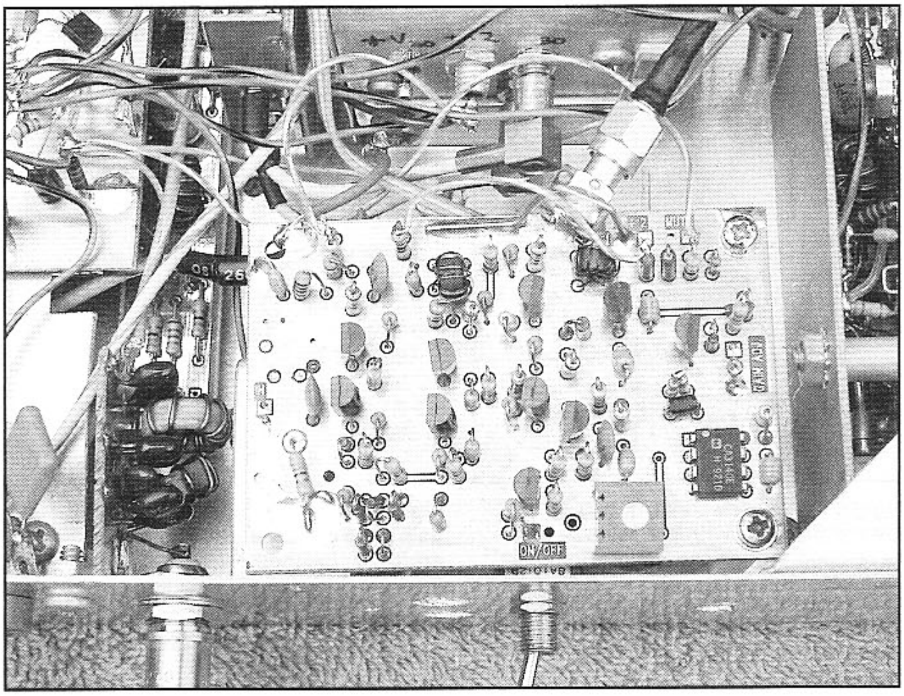


Fig 6.45—An IF amplifier using either a dual-gate MOSFET or a cascode connection of JFETs. These amplifiers use diode strings in series with the FETs for biasing, allowing substantial gain reduction with reduced control voltage. Transformers use #28 wire on an FT-37-43 ferrite toroid. Measurements were done at 10 or 14 MHz.

slower than gain, so IMD products are always decreasing with gain reduction. The measurements were done with 50-Ω input drive. An input network presenting a higher impedance to the gate will increase gain and drop noise figure.

A popular IF device is the dual-gate MOSFET. See the earlier sidebar regarding part availability. With two basic configurations in **Fig 6.44**, that at (A) is the more fundamental. The FET is self-biased with a source resistor while gate 1 is at dc ground. Gate 2 is normally biased at about 1/3 of V_{dd} to produce maximum gain. Moving the voltage on gate 2 in *either* direction will reduce gain. This topology has a limited gain reduction (less than 10 dB) available unless gate 2 is



IF amplifier using a cascode JFET pair.

extended to negative voltages.

Fig 6.44B shows a popular variation used in many imported transceivers. Here, gate 1 is positively biased to about 2 V. With this biasing on gate 1, stage gain variation exceeds 30 dB with positive gate 2 voltages.

Fig 6.45 shows additional variations we examined. One uses a 3N209. The biasing is similar to that in the previous figure, part A, but uses a string of diodes in the source lead with gate 1 biased at the top of the diodes. With the 3N209 circuit shown and without R_T, maximum gain was 28 dB and gain variation was nearly 60 dB. The noise figure was 2.5 dB with the L network designed to present an impedance of 2.3 kΩ to gate 1. Inserting a 3-kΩ resistor for R_T generates a proper termination for the L-network, causing gain to drop to 20 dB and NF to increase to 6.6 dB, but now with a well-matched input. Noise figure degrades only slightly with gain reduction. Very careful gate-2 bypassing is required with all circuits using dual-gate MOSFETs to prevent UHF oscillation. The bypass capacitor should have fairly small C (1000 pF) so AGC dynamics are not altered, and capacitor lead length should be short. A drain resistor (10 to 100 Ω) will also help stability.

IMD performance was modest with a typical IIP3 being –11 dBm. However, intercepts improved as gain was reduced. This means that distortion products always drop faster with gain reduction than the desired signals.

The circuit on the right side of Fig 6.45 uses a cascode connection of J310 JFETs. A slightly larger source resistor was used to obtain similar stage current, typically 8 mA at full gain. This amplifier produced a maximum gain of 28 dB with a 34-dB gain variation (R_T absent.) The 3-dB NF degraded little with 10 dB gain reduction. A typical input intercept was –3 dBm with IMD products dropping faster than the desired output signals.

IF Systems

As we begin to assemble a complete IF system, the first question we ask is "How much gain is needed?" Often, the required gain is very small. In such a case, one can still realize AGC in the IF with a voltage-controlled attenuator. Such a circuit is shown in **Fig 6.46** where PIN diodes are arranged in a ladder of series and shunt elements. Diode current is controlled with a bipolar differential pair. Q2 is completely "on" at maximum gain, conducting all the current offered by Q3. This current flows through series elements with no current flowing in the shunt parts. Some of the current is shifted from Q2 to Q1 as gain is reduced, increasing I in shunt elements and removing it from the series ones. This circuit has a gain range of about 50 dB. Performance is better (lower insertion loss at max. gain) with premium PIN diodes, but is surprisingly good with 1N4006 rectifier diodes. Rectifiers often use a PIN structure to secure high breakdown voltage, but may still have high capacitance when compared with "RF parts."

The total IF gain needed in a traditional AM receiver can be relatively high, for the usual AM detector requires high drive for reasonable fidelity. The product detectors used in CW and SSB receivers are linear to low levels. IF gain is then picked for good sensitivity with the weakest signals and is

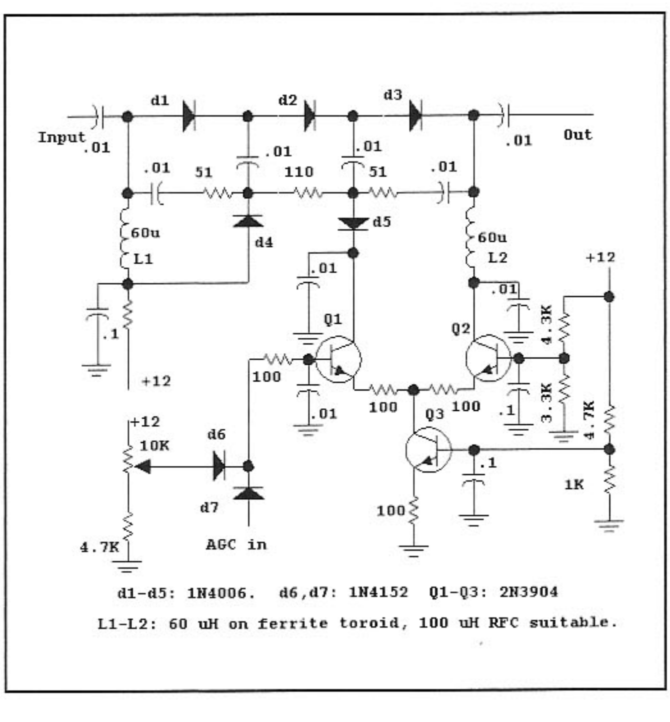


Fig 6.46—IF attenuator circuit offering a 50-dB gain-control range with an insertion loss of about 2 dB at 5 MHz. The 10-kΩ pot is a manual gain control.

reduced as signals get larger. The IF in a digital receiver (one where an IF signal is applied to an A-to-D converter) may have more severe requirements related to matching the input signal requirements of the A-to-D.

The usual IF system provides two outputs. One drives the signal detector while the other is applied to an AGC detector, a circuit providing dc output in proportion to the RF input voltage. Some AGC detectors are shown in **Fig 6.47**. The two outputs must be well isolated. It is especially important that BFO energy from the product detector not reach the AGC detector where it can be detected to reduce IF gain. Noise on the BFO (see the oscillator chapter discussion of noise) that reaches the IF can also inter-modulate with signals to compromise performance.

A dc signal emerges from the AGC detector. It is usually amplified and processed with op-amps for application to the controlled stages.The detector may have a threshold with no output until a minimum input signal is applied. This dc threshold must be exceeded before any gain reduction occurs, resulting in a threshold for RF detection. Once the signals are strong enough to exceed the detector threshold, the AGC holds the output nearly constant with only a slight increase with louder applied signals. **Fig 6.48** shows a plot for one of our receivers, showing output signal Vs available input power. The threshold was adjustable and was set to occur with an input signal of –97 dBm. MDS for this CW receiver was under –140 dBm, so there is a moderate range of signals available before any AGC action occurs. This is an "ear-saver" design, one that protects the user from loud signals, but produces a receiver sound not compromised by AGC. Most commercial transceivers use AGC systems designed to make all signals sound nearly the same. This is clearly an open area for the individual designer/builder.

Diodes are often used to combine two control signals applied to an IF amplifier, shown in **Fig 6.49**. The two signals can come from a manual gain control and an AGC detector, or they may originate from two parts of an AGC system. Similar methods are used to mute receive IF amplifiers during transmit periods.

Fig 6.50 shows a system with two stages of gain with cascode connected J310s followed by a fixed gain differential amplifier. A 1:1 turns ratio ferrite transformer couples the signal from the cascode to the dif-pair. IF output is extracted from one collector of the pair while the AGC detector is driven by the other isolated output.

The experimental development of this circuit started with the first stage, Q1 and Q2. The gain control range was only 30 dB with three diodes in the chain, but increased with 5 diodes. Single stage current was 10 mA at maximum gain, but dropped to about 1 mA at minimum gain. A second stage, Q3 and Q4, was added, sharing the

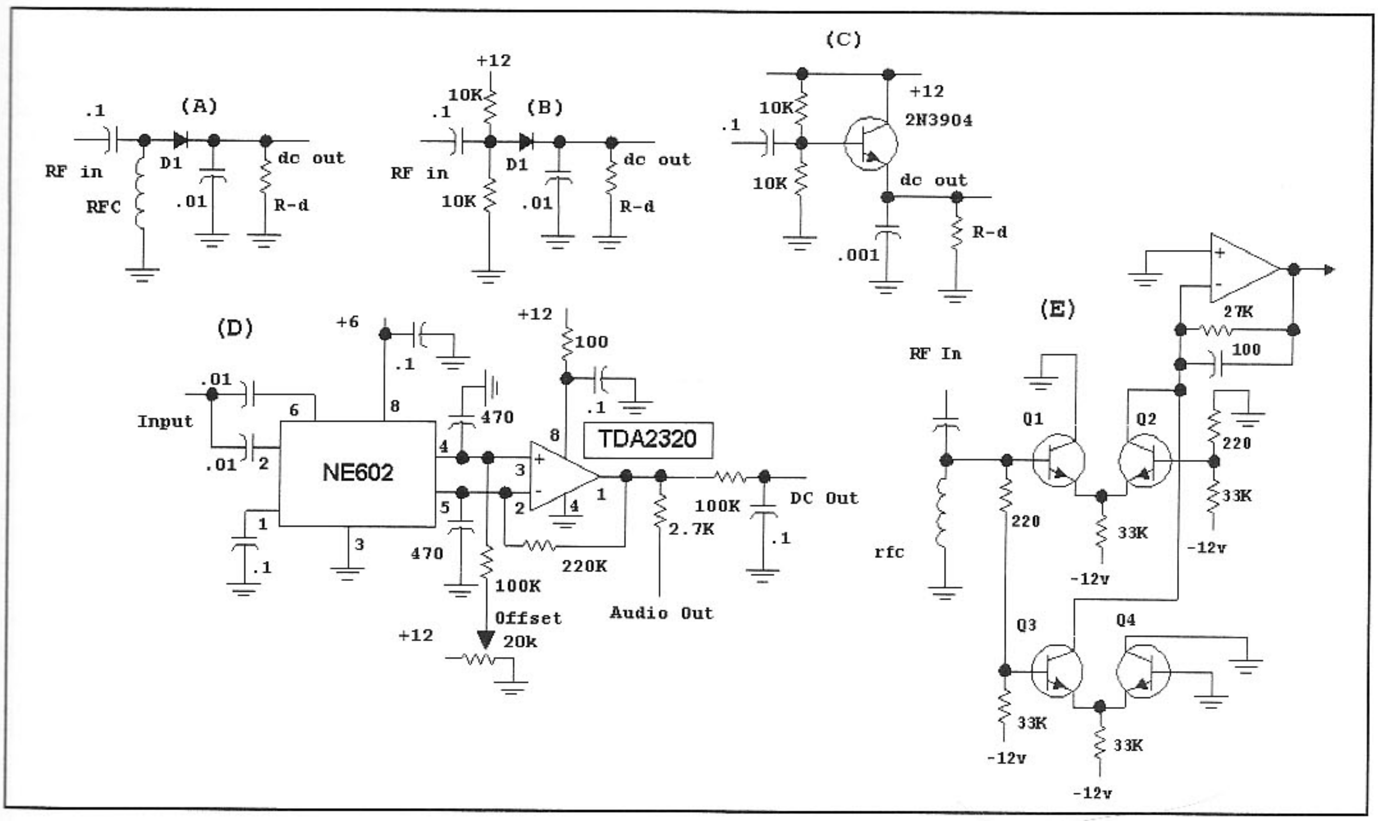


Fig 6.47— Several RF detectors suitable for examining the output of an IF amplifier. (A) shows a traditional diode detector with fast signal diodes. (B) is similar although the diode anode is now biased for a small direct current. (C) shows an emitter follower functioning as a detector. As the input voltage becomes more positive, causing the normal rectification in the e-b diode, collector current flows to charge the capacitor. (D) shows a sensitive detector, suitable for AM demodulation as well as level detection. The Gilbert cell mixer now functions as a multiplier, for both input ports are driven by the same signal. A 10-mV input yields several volts of dc output. If that input is 40% modulated, the audio output will be several volts peak-to-peak. This circuit was designed by W7AAZ. Many op-amps are suitable including the TL074 and NE5532. (E) uses a pair of differential amplifiers, each with an 80-mV input offset, causing each to operate as a detector. Cross coupling of the outputs cancels ac in the output through balance, producing a current input to an op-amp. A dual supply is usually required for this circuit. This detector was used by Carver (W7AAZ) in his high-performance IF system.[6]

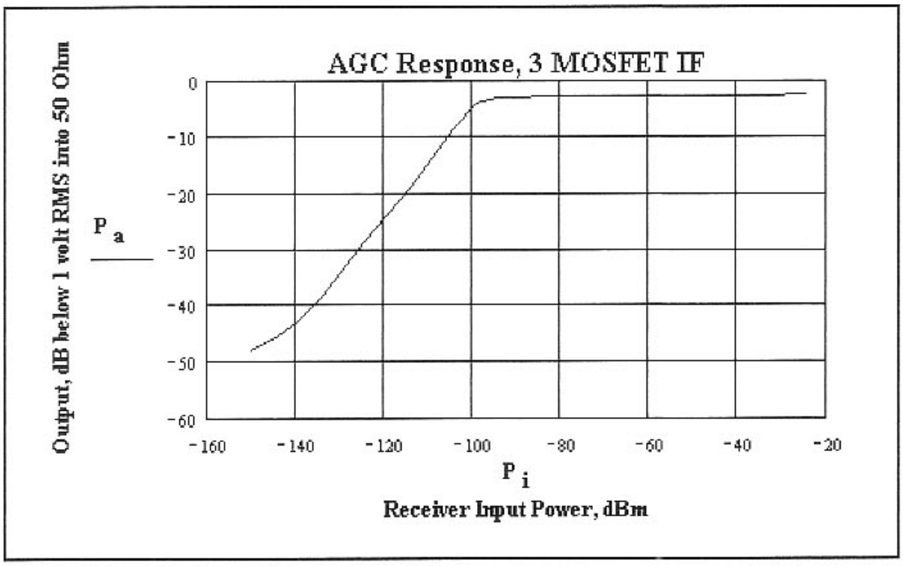


Fig 6.48—Receiver output vs input for a CW receiver. The threshold was specifically set in accord with operator preferences. The IF amplifier is shown later in Fig 6.56.

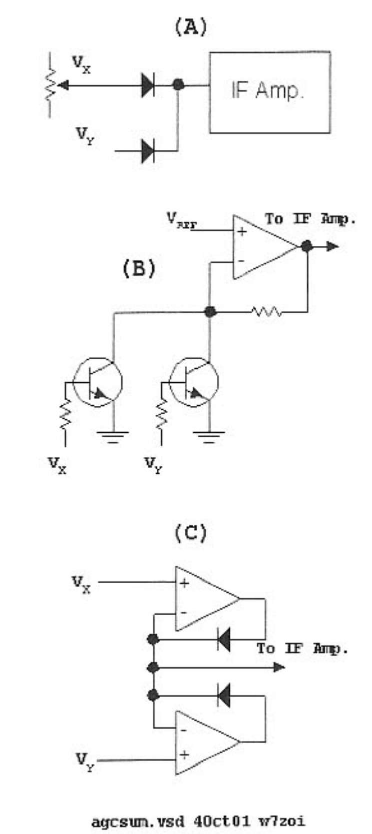


Fig 6.49—Diodes combine signals applied to an AGC amplifier. At A, an AGC signal and one from a manual gain control are selected with the more positive one setting the voltage applied to the amplifier. In B, two signals applied to transistor bases establish currents that are summed in an op-amp. Both inputs contribute in this case. The version in C uses diodes within feedback loops of op-amps to form "perfect rectifiers," which establish a very sharp transition between active inputs. This scheme was elegantly used in Carver's IF amplifier.[7]

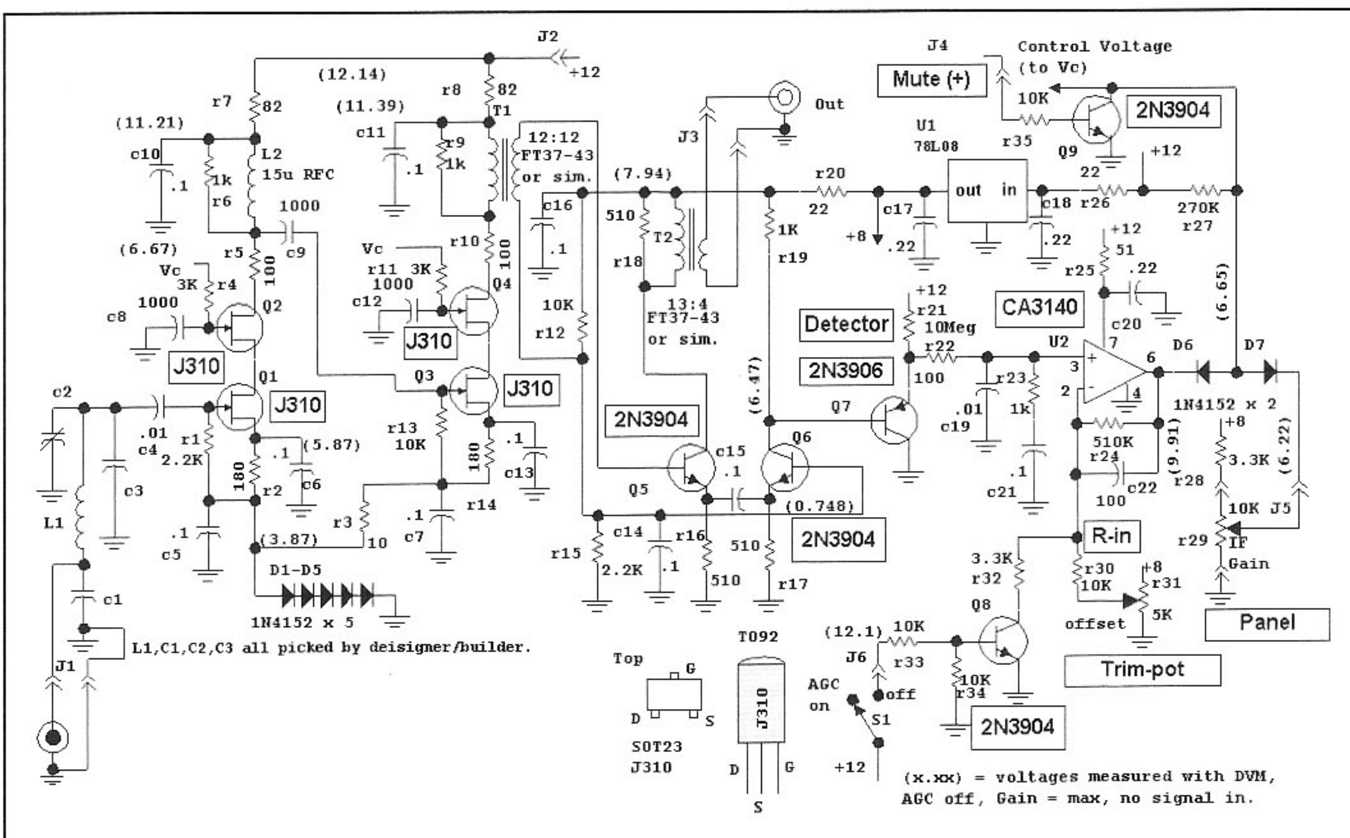


Fig 6.50—A general-purpose IF Amplifier module using cascode J310 JFETs. See text for details.

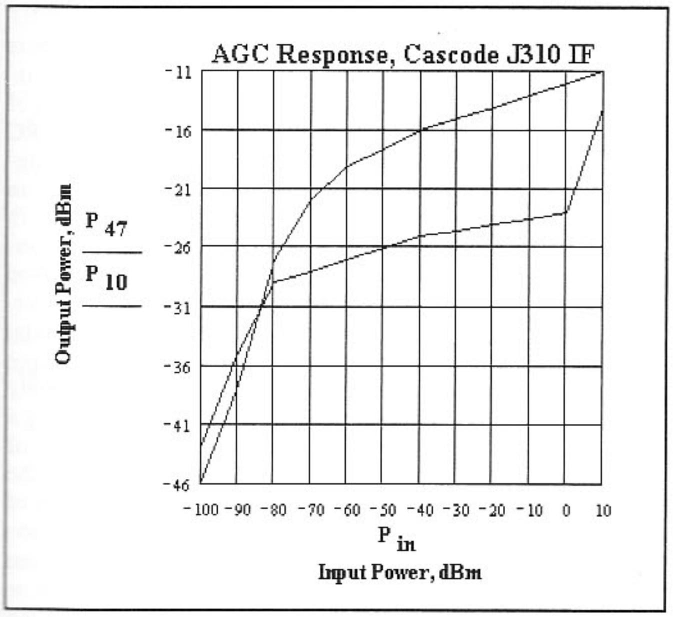


Fig 6.51—IF system output vs input for the IF system using two cascode-connected J310 stages. The two curves are for two different values of "input resistor" in the op-amp, which alters system dc gain. See text for details.

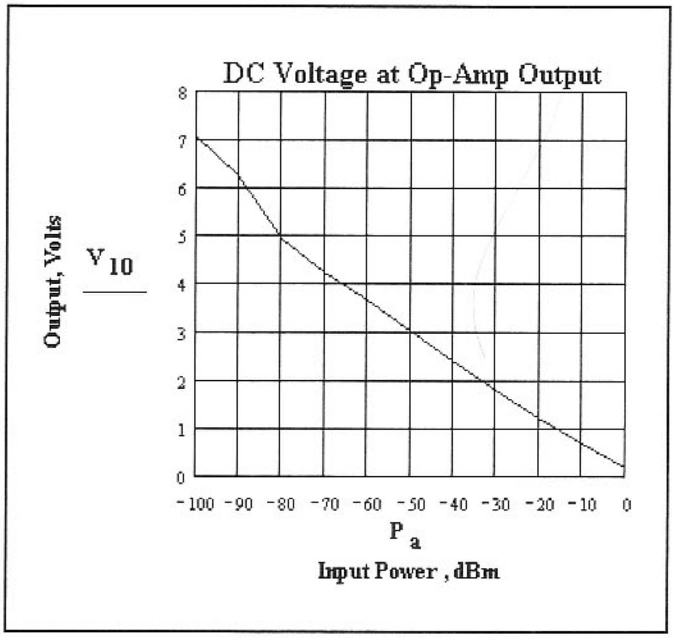


Fig 6.52—DC level at the op-amp output. This voltage may be used directly to drive an "S meter," driven with an op-amp dc follower.

diode chain with the first pair. A J310 source follower was temporarily added to provide an output. The gain variation was now 93 dB at 10 MHz, increasing to 108 dB at 5 MHz. There was a high pass gain characteristic, a result of the 15 μH RFC. Larger values should be used at lower frequency. The gain control voltage should be between 0 and 6 V. Values above 6 V produced a slight gain *decrease*, so that region should not be used.

The 9-MHz gain was 28 dB with no input network other than a blocking capacitor. NF was then 7 dB with R1 at 10 kΩ. A 9-MHz pi network was then added to present a 2-kΩ impedance to the first gate, causing gain to jump to 44 dB while NF dropped to an impressive 1 dB. The NF was maintained with 10-dB gain reduction. We then replaced R1 with a 2.2-kΩ resistor, so the network now causes a good 50-Ω impedance match to appear at the input. NF was now up to 5 dB, increasing to 6 dB with a 20-dB gain reduction. The designer/builder needs to design his or her own networks to apply this circuit to the filters used.

The rest of the circuit was now built, initially using 47 kΩ for R30, R_{IN} at U2. The no-signal dc voltage at the detector output (emitter of Q7) was 6.8 V, so the arm of "offset" pot R31 was set initially to this value. The Op-amp, U2, buffers the control voltage appearing across the timing capacitors, C19 and C21. The loop is closed, generating AGC action, when the op-amp output is connected to the controlled stages through diode D6. The response is shown in the upper curve of **Fig 6.51**. Although the loop is well behaved, it is not very *tight*, allowing considerable output variation between threshold and the upper input-signal limit. Input resistor R_{IN} was dropped to 10 kΩ (increasing loop gain) to produce the preferred response in the lower curve. But the

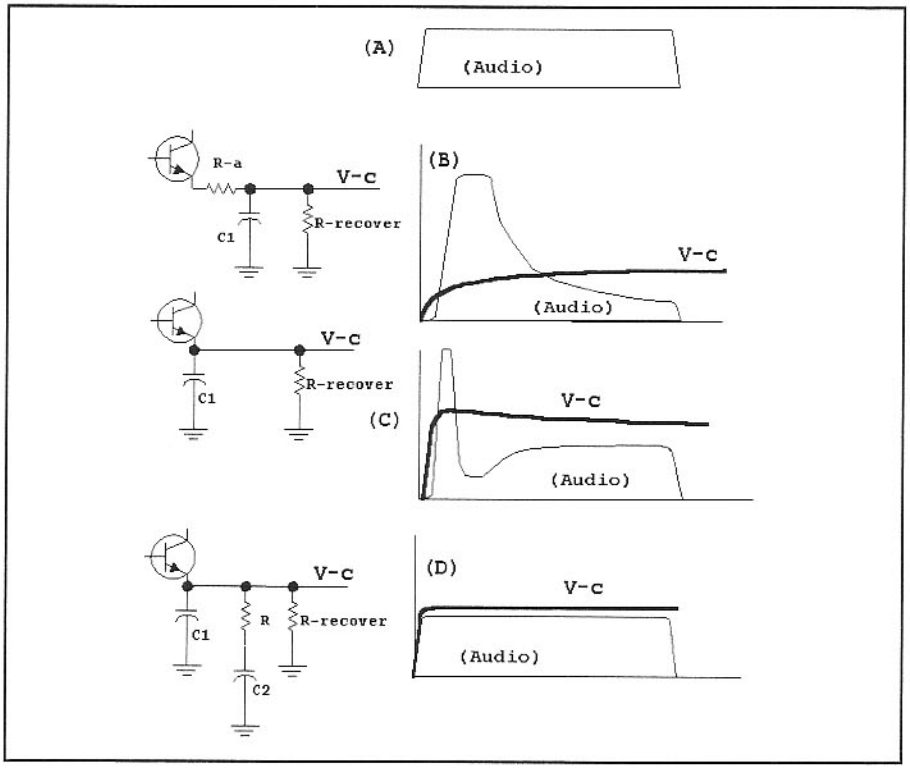


Fig 6.53—Audio envelopes and timing capacitor values vs time. See text for details.

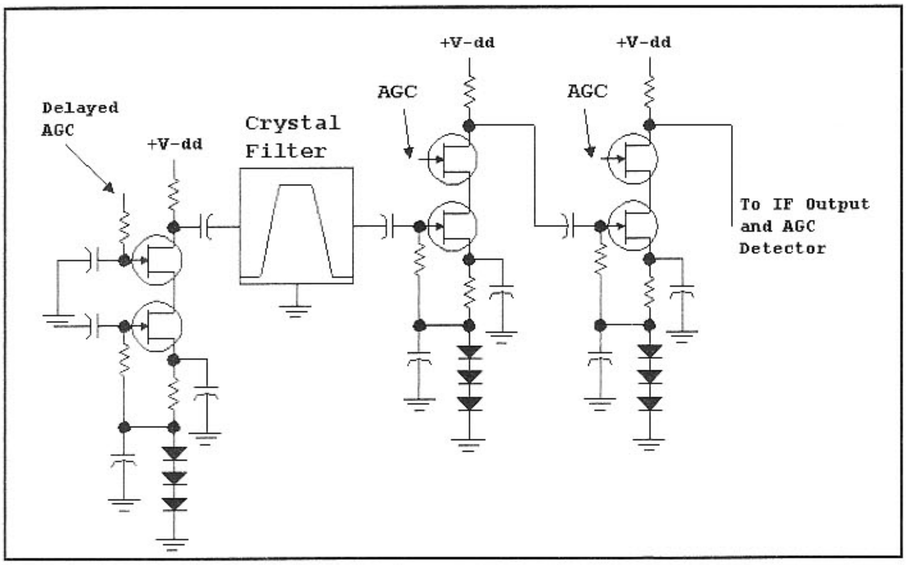


Fig 6.54—System with a crystal filter within the AGC loop. See text for discussion.

system is now ineffective at input levels above 0 dBm. The reason for this becomes clear if we examine the curve of **Fig 6.52** showing dc voltage at the U2 output. The dc voltage has reached 0 by the time the input gets to 0 dBm, so no further gain reduction is possible. Adjustment of the offset pot, R31, will probably fix this anomaly, if it becomes a problem. Such levels would rarely be encountered in most receivers.

The relatively clean dc variation in Fig 6.52 suggests that a signal-strength meter could be driven directly by the op-amp. If this is done, additional circuitry should be added for any "calibration" that might be desired with the S meter. The offset pot is not intended for this purpose, but only to set AGC threshold.

The attack time in the circuit of Fig 6.50 is determined by the detector (Q7) output impedance, by timing capacitors C19 and C21, and by R23. R21 and the capacitors establish recovery characteristics. The values shown were approximate and may require later changes.

A PNP detector was used in the previous circuit. Consider a more general case with an NPN (or a diode) detector charging memory capacitors. **Fig 6.53.** shows some audio envelopes and related capacitor values, V_C. The input to the receiver (or IF system) is a chain of Morse dots (dits.) Even if the receiver is to be used only for SSB, this represents a good test method. Set the strength of the dits to be low, AGC to "off," and the manual gain control to drop the IF gain to produce the response shown in Fig 6.53A. This is an ideal audio envelope with a well-defined rise and fall time.

Having observed the ideal system without AGC, we now increase the strength of the dit chain and activate AGC. Generally we wish to have a near instantaneous fast attack, with a slow decay, yielding the same audio response we saw with the ideal case. But that does not always occur.

Fig 6.53B shows a single timing capacitor, C1, with a modest detector output impedance, R_A. The resulting slow attack allows the audio to climb to high levels, and then drop over the course of the dit as the capacitor voltage stabilizes. Decreasing the attack time, realized by reducing RA, reduces this distorting behavior. But in the extreme this generates the behavior shown in Fig 6.53C where the timing capacitor charges very fast before the gain is reduced. The audio drops to a level below the proper one, but grows to the right value after the loop "catches up." In the extreme, there is no audio for a period until the timing capacitor discharges enough to allow the IF gain to increase to a value that produces a stable result. This is the well known "pop" occurring with some AGC systems.

A solution is found with two (or more) timing capacitors, C1 and C2. C1 is smaller than before and can be charged quickly with the detector output impedance. This may reduce the gain, but for only a short time. Much of the charge on C1 discharges through R to be deposited on C2, increasing that voltage and the resulting V_C value. The process repeats with each cycle of the IF system. This behavior, closer to the ideal, is presented in Fig6.53D.

The process is more complicated than the simple picture we have painted, for there are delays within all IF amplifiers. For example, the control gates of the JFET cascode circuits are connected to bypass capacitors with series decoupling resistors. The bypassing is a necessary part of the cascode connection. The related RC forms a low pass filter that causes the signal at the controlled gate to arrive *after* an input is applied. The delay is short with the values we used, but can be much larger. Signals arriving at the IF input are delayed through a narrow bandwidth filter, generating an output that grows at a finite rate, allowing a fast AGC system to keep up.

In some applications we wish to apply AGC to an RF or IF amplifier preceding a narrow filter, shown in the example of **Fig 6.54**. The filter delay is now within the loop. That is, we detect after the delay of the filter, allowing the signal to grow too large to avoid overloading early stages. The delay can cause severe overshoot or popping if gain reduction is applied directly to the first stage. The preferred solution is to purposefully *delay* the control signal applied to the early stage with a long time constant. Good system dynamics result only when the controlled elements after the narrow filter have enough range and speed to reduce the gain far enough to restrict the output for a short while, only to recover, allowing the delayed stage time to assume part of the overall gain reduction.

We used a string of Morse code dots as a means for evaluation and adjustment of an AGC system. This is not a mere illustration, but a useful experimental method. A simple PIN diode modulator called the *Ditter* is presented in the measurement chapter for just this purpose. The dits are created with a 555 timer IC, but could be generated with a function generator, now offering adjustment ability. The Ditter includes an output to drive the external triggering input of a dual trace oscilloscope. One 'scope channel then shows the control voltage while the other monitors audio or IF output. Ideally, an AGC loop needs to be tested over a wide range of signals, for stability can vary with level.[8]

Audio Derived AGC

Simple equipment sometimes uses audio derived AGC where a detector samples the audio signal to charge a timing capacitor. That voltage is processed and applied to IF amplifiers for AGC. The attraction of this is that audio amplitudes are large, for most of the receiver gain has been realized. Little more gain is required to complete the AGC system. But there is a major difficulty with audio derived AGC. This relates to the sampling nature of the detection process. The detectors we

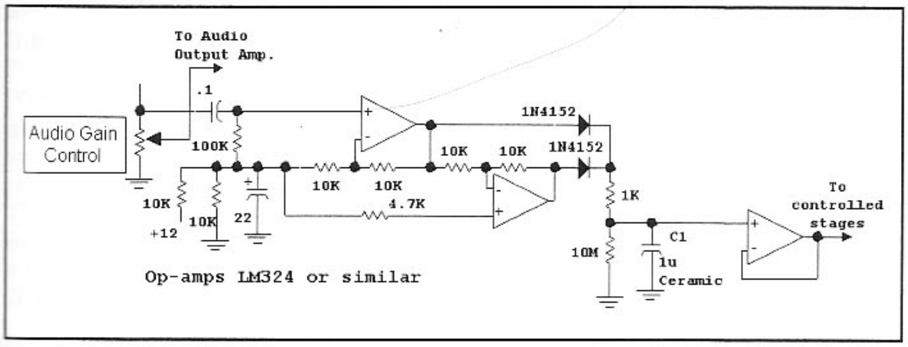


Fig 6.55—Full wave audio detector for use in simple AGC systems.

have examined obtain one sample for each peak of the waveform being detected. Audio waveforms have fewer peaks, especially if the signal is a low-pitched CW carrier. This allows the receiver to be overwhelmed in the period between peaks.

A partial solution to the low frequency difficulty lies in audio filtering. A high-pass filter (with several elements) ahead of both the AGC detector and audio output will prevent very low beat notes from reaching either. A cutoff of around 300 Hz is suggested.

A typical full wave detector for use in an audio derived AGC is shown in **Fig 6.55** with both positive and negative audio peaks contributing to the output. A slow recovery is set by the 10-MΩ resistor across C1, which can be made faster with a smaller resistor. Shorting C1 will turn the AGC off. The system shown is suitable for IF amplifiers like the MC1350P. Level shifting or inversion may be required for other controlled circuits.

Mention was made earlier of difficulties with filters within an AGC loop. This problem can be especially severe when audio filters are included within a loop. Audio filtering is better applied after detection for the AGC loop.

Although audio derived systems present major design challenges, good performance is still possible. This becomes evident when high-end professional-level audio-recording equipment is studied.

Practical FET IF System Examples

The Cascode JFET amplifier presented earlier was developed as a complete, practical module, Fig 6.50, for use in a Monoband SSB/CW Transceiver. This circuit can be built with other FET types, with appropriate circuit changes. The JFETs should be roughly matched for I_{DSS} (+/–10%) and should all be of the same type.

The initial adjustment of the IF amplifier starts by removing one end of R30 from the board. The AGC is turned on with no signals present and the voltage on pin 6 of U2 is measured and recorded in the notebook. The voltage on the arm of R31 is then set for the same value. R30 is again installed in the circuit. R31 can be re-adjusted later to alter AGC threshold.

A similar MOSFET IF amplifier is shown in **Fig 6.56**. This circuit uses three gain stages using 3N209 MOSFETs, a type available in our junkbox. Those wishing to duplicate this circuit should consider the 3SK131 or similar available SMT parts. After three gain stages, the signal is applied to a differential PNP amplifier. One side is terminated in a 510-Ω resistor, providing a properly matched drive for a "tail end" crystal filter. This filter serves to eliminate noise generated within the IF amplifier at frequencies other than that of the main filter. It also distributes the selectivity improving the stopband attenuation of the overall system. The noise filter is terminated in a resistor and a FET follower output stage feeding a product detector.

The main IF input selectivity is provided by a 10th order filter with a 500-Hz bandwidth, designed for a Gaussian-to-12-dB response. (This filter, a KVG XL-10M, is regrettably no longer available. They are sometimes found on the surplus market, but few were manufactured.) The IF system was breadboarded without printed boards in a multiple-section surplus milling. One section contains the main filter input while another has the output and the first IF amplifier. Another houses the 2nd and 3rd IF stages while yet another holds the differential amplifier and an NPN detector. Feedthrough capacitors route the signal through the milling where the dc parts of the AGC loop reside.

The input circuitry is critical to the components used. A ferrite transformer matches the 50-Ω drive to the main crystal filter impedance of about 300 Ω. The filter output is then transformed up to 2200 Ω with a low Q pi-network where a 2.2-kΩ input resistor at Q1 terminates the filter. This topology guarantees a reasonable noise figure with a proper impedance match for the crystal filter, vital in preserving the specified performance. The pi-network used an existing RF choke, although a toroid with higher Q_U would be preferred.

This IF has a bandwidth just under 500 Hz with a measured system sideband suppression in excess of 120 dB. The dc AGC response was presented earlier in Fig 6.48. The threshold may be adjusted with R-th (2.5 kΩ) shown in the schematic.

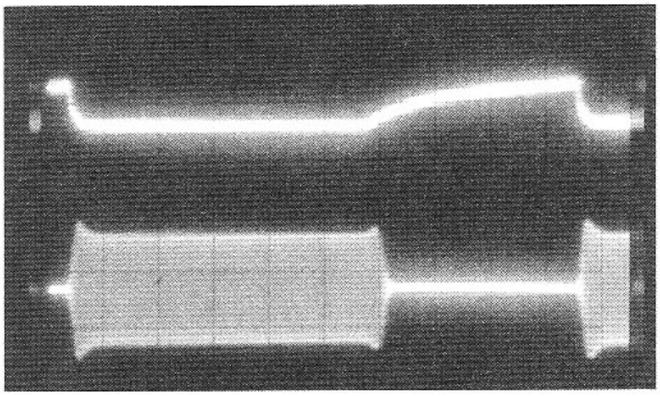
The attack and recovery are determined by the components in the *Timing* section of the circuit. An NPN detector, Q6, charges a feedthrough capacitor that feeds a signal out of the milled enclosure to a CA3140 op-amp that then drives inverter Q8. The Q8 collector then drives the timing capacitors. The primary one is a .01 µF, which is tied to a 0.1µF/10 kΩ combination paralleled by a 1µF/100 kΩ pair. These values were established with the ditter mentioned earlier. The voltage on the timing capacitor and the audio signal are shown in a photo.

A Hang AGC System

Fig 6.57 shows an AGC system with the unusual characteristic of using two timing systems. One is driven by IF signals, so it has the advantages of quick attack. The other comes from the audio.

During receiver operation, signals within the IF cause C2 to charge, which reduces receiver gain. If the signal is a short lived one or even a noise burst, C2 will quickly discharge through Q10, a low pinchoff JFET switch. However, if the signal is present for a reasonable period (around a hundred milliseconds), audio will have been amplified by Q11 to charge C1 negatively. This drives Q10 into pinchoff, which disconnects it from C2. The only discharge path for C2 is now a 22-MΩ resistor, so recovery is slow, causing the gain to *hang* at a nearly constant level. But if the audio disappears for a short period, C1 discharges, Q10 is no longer pinched off, and C2 quickly discharges, returning the receiver to full gain.[9] The audio detector is an "open-loop" process that modifies the basic closed-loop IF AGC system, so does not alter system dynamics.

The hang scheme can be adapted to our FET IF systems with relative ease, as shown in **Fig 6.58**. The partial circuit in



Timing signals for the MOSFET IF Amplifier during AGC testing.

(A) is set up for NPN detectors while (B) accommodates PNP detectors. The builder must provide many design details.

Evolving Designs

Clearly, many of the methods can be combined. For example, W7AAZ built an IF amplifier using a dual-gate MOSFET input stage followed by an MC1350P. By applying AGC to both FET gates, he was able to obtain a very wide AGC range in a relatively simple design. PIN diodes can also be added to existing systems to stretch the range of FET or bipolar amplifiers, integrated or not.

Perhaps the most impressive IF design we have seen is that presented by Bill Carver, W7AAZ, in *QST*, May, 1996.[10] This circuit is based upon the AD600 series of integrated circuits from Analog Devices. Although expensive, these parts offer outstanding performance. They feature a wide AGC range that is extremely *dB linear* (the gain in dB is directly proportional to the control voltage). Bill's complete paper is included on the CD included with this book.

Carver's IF amplifier included a number of outstanding features not found in other circuits. His circuit used three amplifier blocks where gain reduction occurred, just as one of the previous circuits shown (Fig 6.56) used three stages. Our simple circuit had gain reduction applied to all stages at once. But Carver's IF used a sequential gain reduction. The last stage had gain reduced by 40 dB before any other reduction occurred. Further reduction was applied then to the

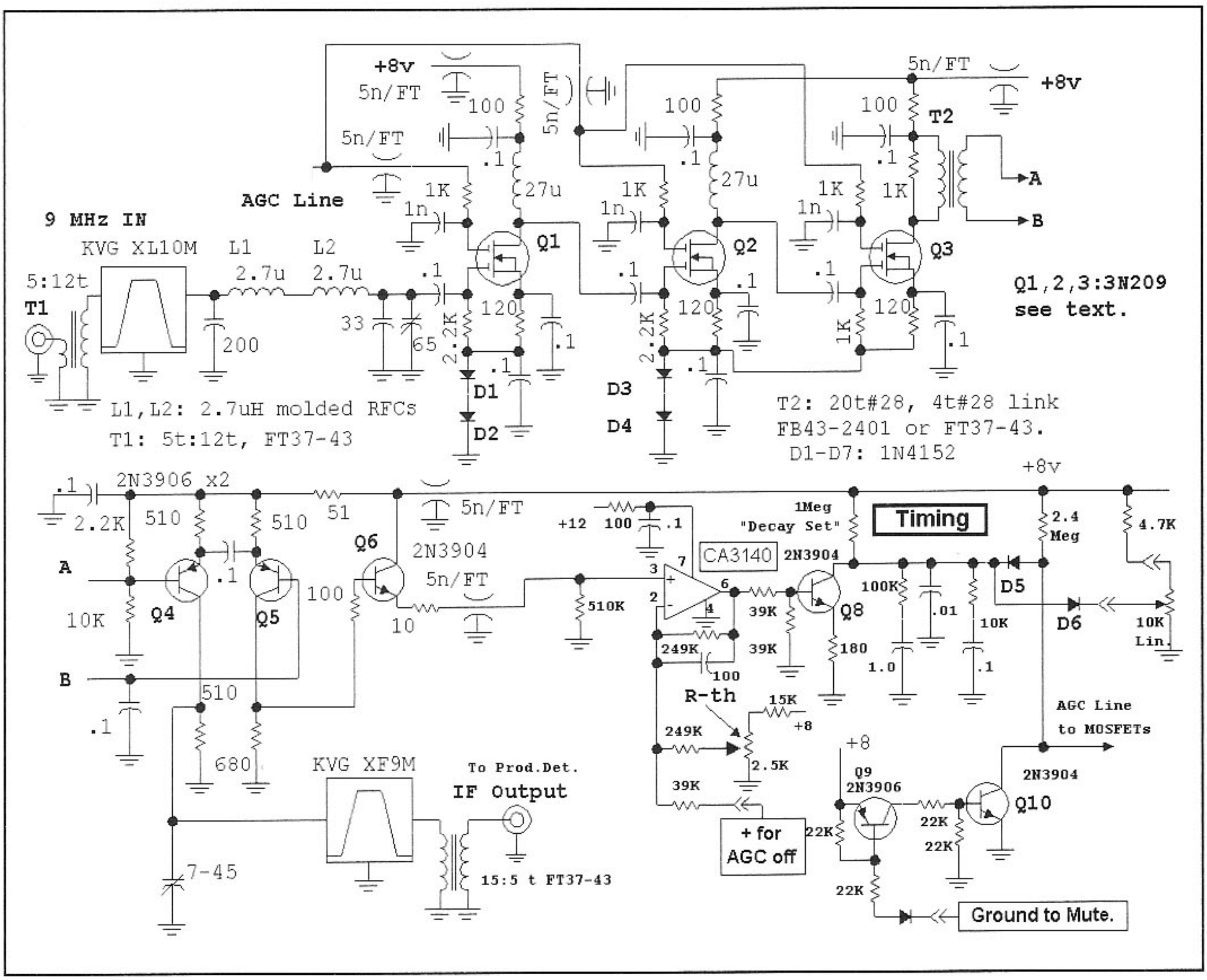


Fig 6.56—IF amplifier using three gain reduction stages with dual-gate MOSFETs. See text for discussion.

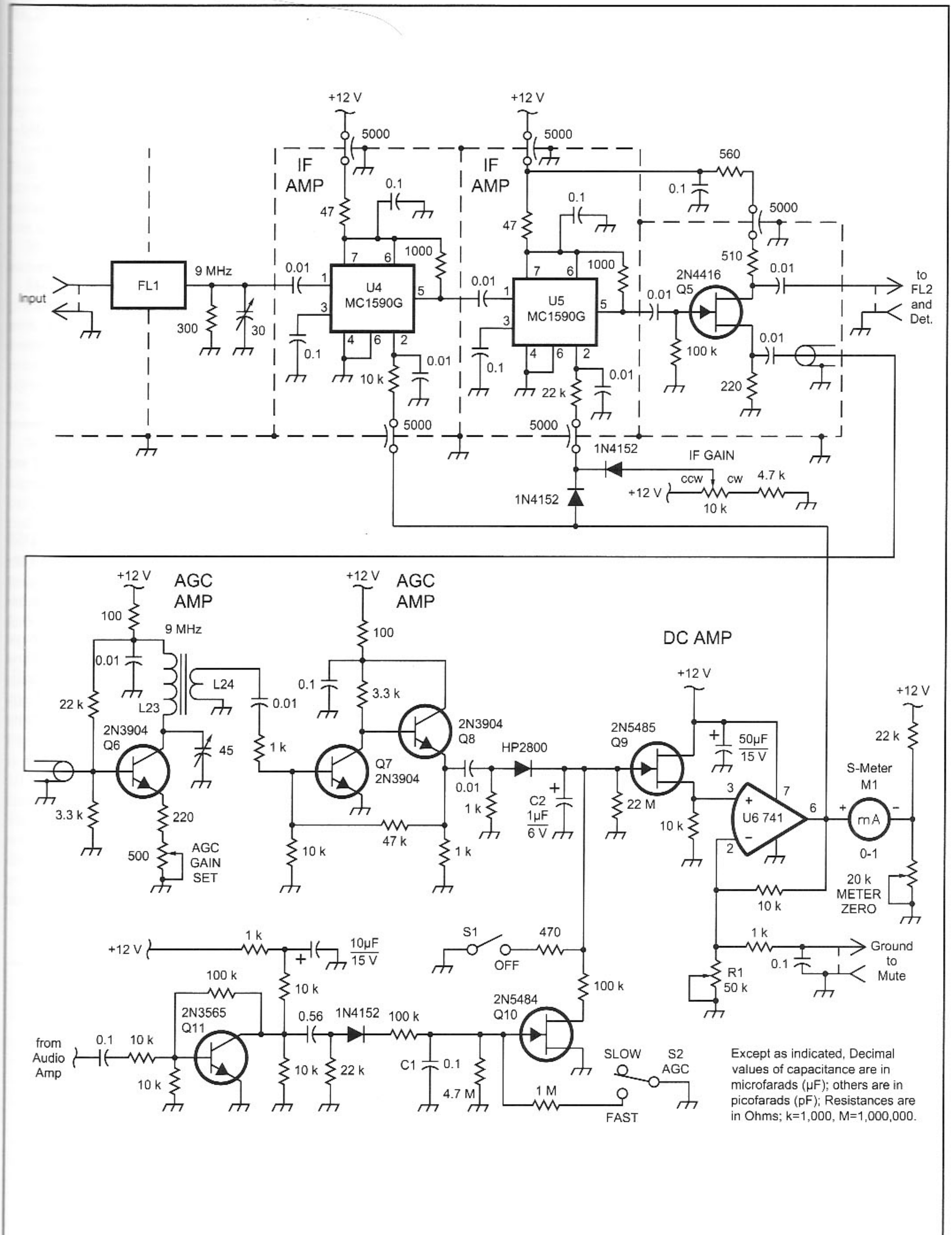


Fig 6.57—A full hang-type AGC system with two timing systems. The IF-derived AGC offers quick attack while "hang time" is established by the audio.

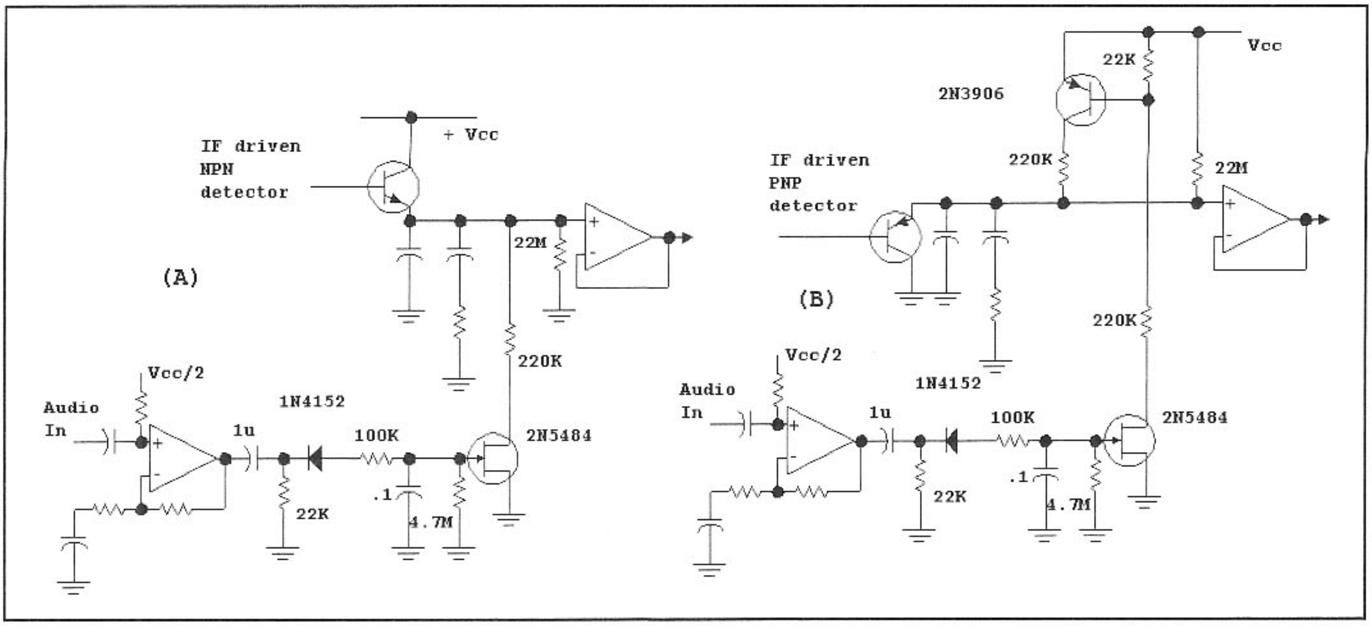


Fig 6.58—Adapting a hang AGC to IF amplifiers with NPN or PNP detectors. See text discussion.

middle stage, and after a total of 80-dB reduction, to the input stage. This was possible because of the buffering used within the AD600 and the use of "perfect rectifiers" in the control circuits. The Carver system also used a second gain reduction loop with a bandpass filter between stages, optimizing dynamic behavior while keeping noise low.

The Carver paper included another unusual feature that will become more common with emerging receivers: He used a feed-forward scheme where the AGC detector not only controlled the gain of stages ahead of the detector, but altered the gain in stages following detection. In principle, one could carry these methods to the extreme where an accurate detector establishes gain in later stages without a need for negative feedback. This could be realized with hardware (a log amplifier and detector with variable gain IF amplifiers and stepped gain audio amplifiers) or software with a DSP system. Delay in filters or amplifiers presented a problem with traditional negative feedback systems, but now becomes an asset, providing time for calculations in a DSP based system. These DSP methods have already, at this writing, been used for a few years in some very- high-performance military equipment from Rohde and Schwarz, and will be described for use in DSP transceivers described in this book.[11]

6.3 LARGE SIGNALS IN RECEIVERS AND FRONT END DESIGN

The range of signals available to our receivers can be very wide indeed. The weakest signals we can hear are limited by noise, and drop to typical levels of –140 dBm or less in a CW bandwidth. These are rare at HF, but common at VHF. But signals can also be very strong. The strongest sky-wave propagated signals we encounter will depend on our antenna, but can sometimes be as strong as a microwatt (–30 dBm,) or even more with high gain antennas.

Most of our concern for large signal performance relates to the receiver *front end,* the part of a receiver between the antenna connector and the place where receiver bandwidth determining selectivity is obtained, usually the first crystal filter. The front end usually consists of much more than the "first stage."

We have two concerns when dealing with the large signals. First, "How loud can the signals be that we try to copy with our receivers?" This problem relates to both front ends and to gain control. Second, "What is the range of signals that can be present within the receiver front end without causing problems when we attempt to receive average or weak signals?" This is the more complicated and subtle problem with the more interesting challenge.

Fig 6.59 shows a partial receiver block diagram for a 14-MHz single-conversion superhet with a 2-MHz IF. The calculated front-end filter response is shown in **Fig 6.60**. The center frequency response is normalized to 0 dB, so the response at 10 MHz can be used to evaluate worst-case image rejection, 76 dB for this example. The front-end bandwidth, over 400 kHz, is wide enough to not require any adjustment during receiver use. These filters contribute little to the receiver signal selectivity and do *not* impact noise figure and sensitivity. But they are vital in protecting the receiver from other responses.

The narrow crystal filter in the IF determines the receiver selectivity. The response of two crystal filters are shown in **Fig 6.61**. Both filters were designed for a bandwidth of 2500 Hz, but one filter uses four crystals while the more selective one uses eight. The beat frequency oscillator (BFO) is normally placed 300 Hz below the lower passband edge for an upper sideband response. The voice frequencies then recovered by this 2500-Hz bandwidth filter extend from 300 to 2800 Hz. Opposite sideband response is then well defined. Owing to the filter skirt shape, sideband suppression is critically dependent on position within the passband. For the four-element filter, sideband suppression extends from only 14 dB at the low audio end to 43 dB at the high end. The 8-element filter offers much better sideband suppression, but is still only 27 dB at the low audio end. It grows to 87 dB at the high audio extreme. Similar response can be expected in a filter method SSB transmitter. The improved response of the phasing method is dramatic for sideband suppression *at low audio frequency*. This suggests that combinations of a superhet and the phasing method may offer spectacular performance, an old, but still viable option.

Several undesired phenomena occur in

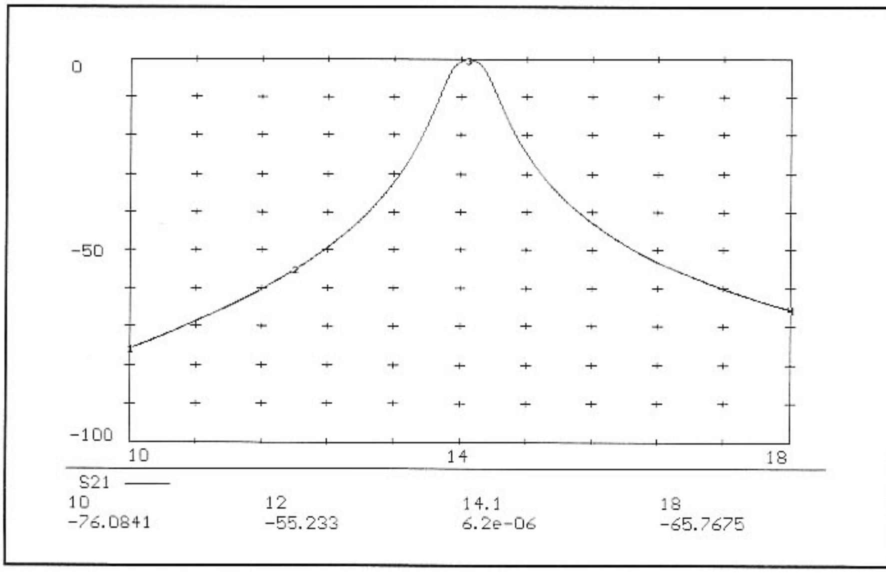


Fig 6.60—The response of the front end from 10 to 18 MHz. The image rejection at 10 MHz is 76 dB. This is a computer generated ideal plot. The 3-dB bandwidth is 0.41 MHz, centered at 14.1 MHz. This response results from a single- tuned circuit at the antenna and a double-tuned circuit between the RF amplifier and the mixer.

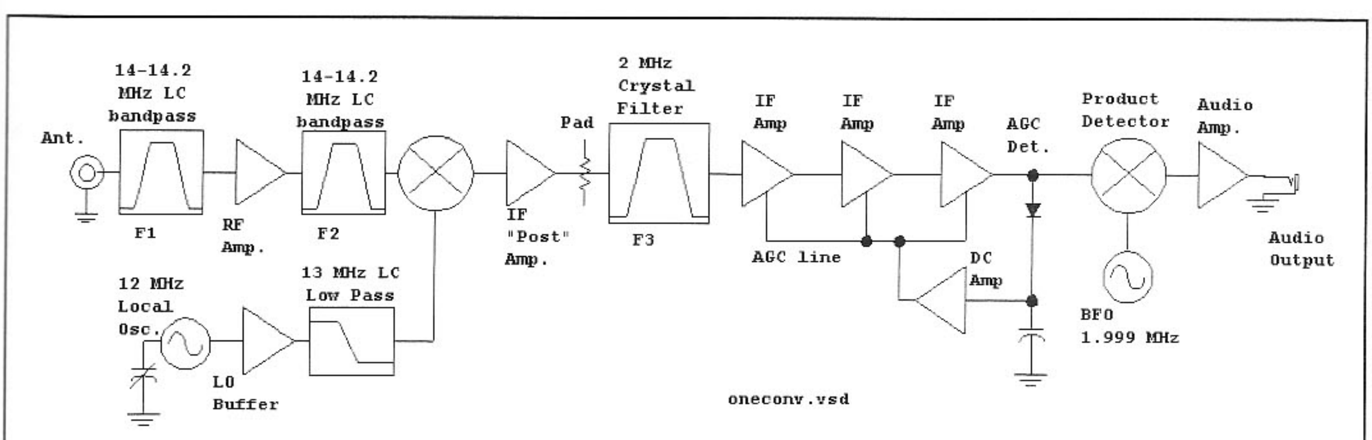


Fig 6.59—14-MHz receiver with a 2-MHz IF. The LO tunes from 12 to 12.2 MHz, so the image extends from 9.8 to 10 MHz.

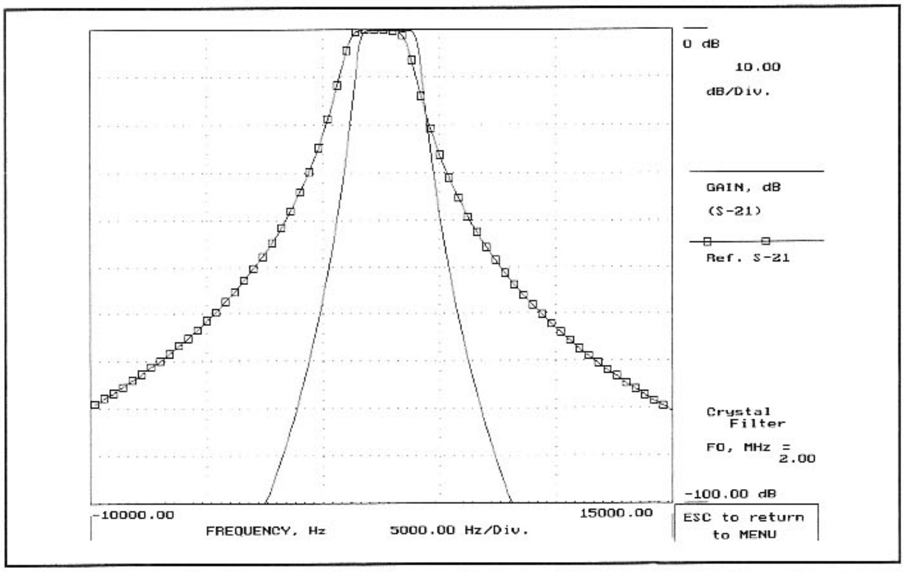


Fig 6.61—Response of two crystal filters. While both have a bandwidth of 2.5 kHz, one uses only 4 crystals (trace marked with small squares) while the other uses 8. Both were designed for a Butterworth response. Steeper skirts are afforded by a Chebyshev response. See text for discussion.

a receiver front end to compromise performance. These include:

- Gain compression: If we examine the front end as a module and measure gain, we find a constant value for most signals. However, as the signals grow, we eventually find a level where the gain is reduced over the small signal value. We usually specify the 1 dB compression point, that available input power in dBm where gain is reduced from the small signal value by 1 dB. A simple way to measure gain compression uses two signals or "tones." One is of weak to average strength and is the one tuned by the receiver during the test. The other is much stronger and is placed within the front-end bandwidth, but well outside the receiver bandwidth. A typical spacing for an SSB receiver might be 20 to 50 kHz. The strong signal is increased until the weaker one drops by 1 dB. This can be a difficult measurement to perform. The IF filter must have enough stopband attenuation to keep the strong signal from creeping into the IF where undesired AGC detection might occur. Further, the measurement is often compromised by reciprocal mixing, or noise blocking, which is described below. Gain compression is easily defined, but rarely a great problem.
- Cross modulation: This was a common specification when AM was the dominant modulation mode. It is measured with two input signals. The first is an *average* strength carrier with no modulation of it's own. The second is a much stronger modulated carrier spaced away from the weak carrier by several receiver bandwidths. It is often 30% amplitude modulated by an audio sine wave. We increase the strength of the modulated carrier while the receiver is tuned to the weaker one, waiting until the modulation of the stronger appears on the weaker one.
- Phase noise blocking, or reciprocal mixing: This problem was described in the oscillator chapter. Phase noise blocking occurs when a strong signal is applied to the receiver at a frequency slightly away from the receiver's tuned frequency. Noise sidebands on the receiver LO will mix with the incoming signal to produce an IF response. The offending energy is a noise rather than a carrier, so the response is proportional to receiver bandwidth. For this reason, the response, when measured, is usually normalized to a 1-Hz bandwidth. Measurement is complicated by noise on a generator that might be used to measure it. It is difficult to differentiate between the two, justifying the term *reciprocal mixing*. Noise blocking shows up as a problem on the air when a strong local signal appears. If the offending signal is on CW, the noise shows up as a keyed hiss that become stronger as the receiver is tuned toward the signal. It is a fundamental problem that is "fixed" only with careful LO design. Reciprocal mixing is a major problem with frequency synthesized radios and offers *the single most fundamental challenge* to the design of advanced communications equipment. An integral part of this challenge is that of eliminating spurious responses in frequency synthesizers, sometimes quite significant when DDS is used.
- Second-order intermodulation distortion: Generally, intermodulation distortion (IMD) occurs when two or more signals are applied to the input of a receiver, creating distortion products at frequencies other than the input. Second-order IMD produces sum and difference frequencies. The sample receiver of Fig 6.59 used a 2-MHz IF, so two inputs that were separated by 2 MHz could generate an output at the IF. Inputs at, for example, 13 and 15 MHz could generate the distortion products. However, this is unlikely, for our receiver is preceded with considerable filtering. Signals at these frequencies are attenuated before reaching the later parts of the front end. Second-order IMD is characterized by an intercept, as outlined in Chapter 2.
- Harmonic distortion: This is a distortion created within the receiver where the output is a harmonic of an input. For example, second-order harmonic distortion would occur if a strong 1 MHz signal was applied to the front end. A second-harmonic signal would then be created within the receiver and produce a signal in the 2-MHz IF. A more common distortion might be generated from a strong 7-MHz signal. The 14-MHz second harmonic created in the receiver front end is available for subsequent conversion. But the example front end filtering is extreme enough that little 1 or 7 MHz energy would ever reach the front end. Direct harmonic distortion is rarely a problem in a well pre-selected receiver, one with good input filters. But most commercial receivers, today, are not well pre-selected.
- Third-order intermodulation distortion: Like second-order IMD discussed above, this distortion is the result of two input tones. This product is perhaps the most difficult distortion to eliminate, for it occurs close to a pair of incoming frequencies. It is a *third-order* product because there are essentially three frequencies that create the product. If two input frequencies, f_1 and f_2, are applied to a receiver, the distortion occurs at $(2 f_1-f_2)$ and $(2 f_2-f_1)$. In the first example, f_1 is used twice, so the 3 inputs are f_1, f_1, and f_2. (Note that *order* can also be related to the exponent on a dominant term in a power series description of the distorting device, but that relationship is often ambiguous.[12]) Consider two example inputs of 14.04 and 14.05 MHz, directly within our input filters. The distortion products now appear at 14.03 and 14.06 MHz. The front

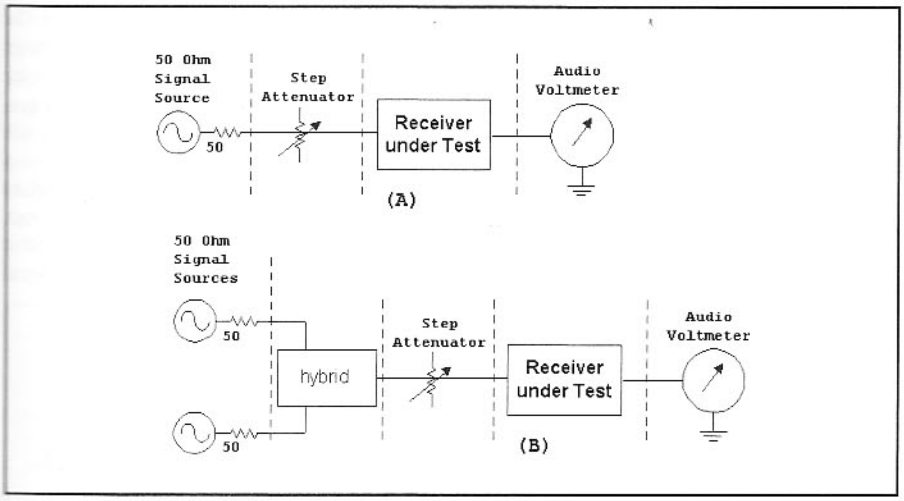


Fig 6.62—Setup for measurement of receiver dynamic range. See text for discussion.

end filtering does nothing to attenuate the original signals that cause the distortion, nor does it attenuate the products once they have been generated. First impressions suggest that this distortion would ruin all communications, but things are not that severe. The detail that saves our receivers is the characteristic that a third-order distortion product will increase or decrease in proportion to the *cube* of the input signals. So, if input signals become 1 dB weaker, the resulting distortion decreases by 3 dB. Third-order IMD in a receiver is characterized by a third-order input intercept. Although third-order IMD is an insidious problem, it is easy to measure. Generally, anything we do to a front-end design to improve IMD will also improve gain compression and second-order IMD. For these reasons, the third-order input intercept becomes a central design consideration for receivers.

Dynamic Range and Intercepts

We often hear folks talking about *dynamic range* of an amplifier or receiver, but the term is often ill defined. When asked about it, the person will say it is the difference in dB between the largest signal that a circuit can handle and the smallest. But what is the weakest signal and what defines it? How large can the largest be and how do we define that?

We use the following *receiver* definition: Two-tone dynamic range is the dB difference between two signal levels; The weakest signal that a receiver can deal with is the *minimum discernable signal*, or MDS while the strongest signal is one of two signals of equal strength that produce a third-order distortion product with a response equal to that of the MDS.

MDS was defined earlier and is the available power from a room temperature signal source that will cause the output to increase by 3 dB above the background noise. MDS is related to receiver noise figure and bandwidth by

$$\text{MDS (dBm)} = -174 \text{ dBm} + 10\log(\text{BW}) + \text{NF} \qquad \textbf{Eq 6.12}$$

where BW is the receiver *noise* bandwidth in Hz and NF is the noise figure in dB. Noise bandwidth is usually close to signal bandwidth at the –6 dB points.[13] For example, a receiver with a 2.5-kHz bandwidth and a 10-dB noise Figure has a –130-dBm MDS. The test setup used to measure MDS is shown in **Fig 6.62A**. The signal in dBm available to the receiver is the generator output less the attenuation value in dB.

After measuring MDS, a second signal source is added to the test set, as shown in Fig 6.62B. The sources are adjusted to have equal outputs. The *hybrid* in that figure is a circuit element that combines the outputs of two 50-Ω generators to form one 50-Ω source while isolating the two generators from each other. (See Chapter 7 under Return Loss Bridge.) The combined output is adjusted as needed in the step attenuator. The level available to the receiver input is adjusted until the response on the meter is exactly the same 3-dB-above-the-noise response that we saw when measuring MDS.

Consider an example. First, turn AGC off for all DR and intercept measurements. With no input signals, the audio output from our receiver is 5 mV, RMS. This is the result of receiver noise. We now inject a 14.010-MHz signal from a generator and adjust the level and receiver tuning until the audio output is 7.1 mV, 3 dB above the noise level. This happened with a generator output of –130 dBm, which becomes the MDS. Next, we set up the signal generators at 14.03 and 14.05 MHz, leaving the receiver tuned to 14.01 MHz. We increase the level of the two tones until we get the same output that we saw with the MDS measurement. This occurs with a signal at the input of –44 dBm per tone. Each tone is 86 dB above MDS, so our two-tone dynamic range is 86 dB.

We can measure the receiver input third-order intercept directly with the same equipment. (See Chapter 2, section 6, to

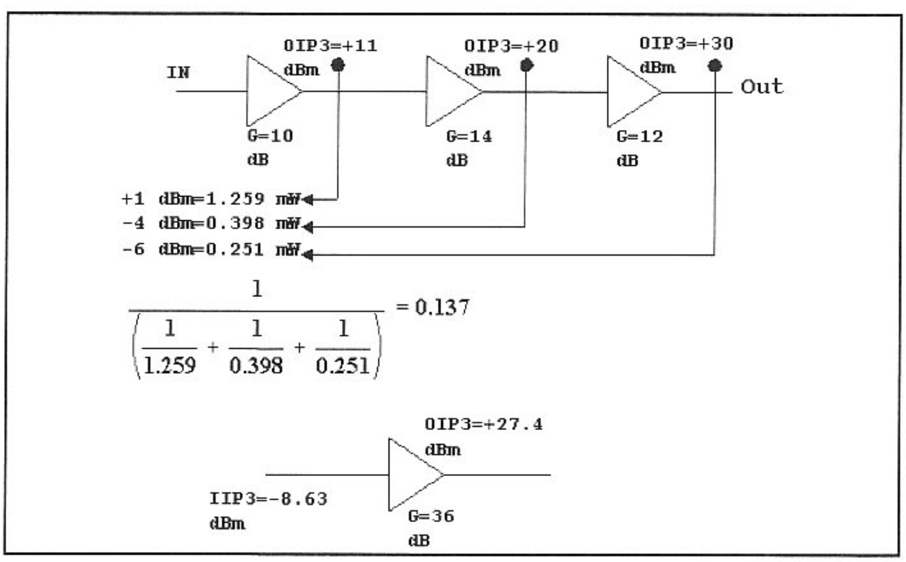


Fig 6.63—Three amplifier stages are cascaded. The intercept for the cascade is calculated by normalizing the intercepts to one plane in the system, converting values from dBm to mW, combining values in the way that resistors in parallel are combined, and then converting back to dBm. See text for details.

see how intercept is defined and measured.) Set the attenuator output for a larger output per tone than was used in the direct DR measurement. Tune the receiver to 14.01 MHz and note an output of 100 mV in the audio voltmeter. We note that the available signals at 14.03 MHz and 14.05 MHz is –31 dBm per tone or per signal. We now tune the receiver to 14.03 MHz where we encounter a very loud signal. The attenuator is increased until the output level is again at 100 mV, finding that this happened when we had added 60 dB of attenuation. Hence, the distortion products are 60 dB below the desired response. This is the IMD Ratio, or IMDR. Rewriting an equation from section 2.6

$$IP_{3in} = P_{in} + \frac{IMDR}{2} \qquad \textbf{Eq 6.13}$$

allowing us to calculate the input intercept for the receiver as –1 dBm. While doing this measurement, it is instructive to change the input from –31 to –29 dBm, or a similar small amount. With 2-dB-larger input signals, we see IMD products that are 6 dB stronger. The IMDR becomes 56 dB, still leaving an input intercept of –1 dBm. If IP_{3in} remains a constant, the front end is said to be *well behaved.*

Two formats are used to indicate intercepts. The one we have used for an input intercept is IP_{3in}. The IP_3 part indicates that it is a third-order intercept while *in* signifies an input rather than output intercept. An equally valid designation is *IIP3* where the first *I* denotes input. The second format relates to the *output* intercept, symbolized by IP_{3out} or *OIP3*. Avoid associating the term *intercept point* with a number, for it is only confusing when the plane of definition is not specified. Strictly speaking, intercept point is the intersection of two curves.

Intercepts are not mere esoteric curiosities or receiver figures-of-merit. Rather, they are tools, useful parameters available to the designer. Intercepts offer two major capabilities:

- If the input intercept of a receiver (or any system) is known, the intermodulation distortion is well defined for all input levels.
- If the intercepts and gains for all stages in a system are known, they can be combined to calculate the intercept for the complete system. Input and output intercepts for a single stage differ by the small-signal stage gain.

Equation 6.13 lets us calculate distortion for any input level.

The intercept of a cascade was treated earlier and is illustrated here with an example: a three-stage amplifier shown in **Fig 6.63**. This cascade might be part of a wideband amplifier to be used in an SSB transmitter. The output intercepts of the three stages are known: +11, +20, and +30 dBm. The respective gains are 10, 14 and 12 dB. Recall that the input intercept of an amplifier is related to the output intercept through the stage gain. This difference is not restricted to a single stage. The output intercepts for each stage can be normalized, or "moved" to the input of the overall system, becoming +1, –4, and –6 dBm. The individual intercepts are merely adjusted by the gains in the movement process. The normalized values are converted from dBm to power in milliwatts. The values are then combined in the same way that *resistors-in-parallel* are combined, producing a net input intercept of 0.137 mW, or –8.6 dBm. The parallel resistor analogy has no significance other than being an easily remembered formula.

This can also be presented in a generalized equation

$$IP_3 = -10 \log\left(\sum_{i=1}^{N} 10^{-\frac{IP_i}{10}} \right) \quad \textbf{(General case)}$$

$$= -10 \log\left(10^{-\frac{IP_1}{10}} + 10^{-\frac{IP_2}{10}} + 10^{-\frac{IP_3}{10}} \right) \quad \textbf{(N-3, 3-stage example)}$$

Eq 6.14

where IP_3 now represents the intercept of the cascade and IP_i is the intercept of the i-th stage with all intercepts being normalized to a single plane in the amplifier. In our example, we normalized all intercepts to the system input. However, we could have picked the output, or any interface between stages. (The equation is derived in *Introduction to Radio Frequency Design.*) This method is a worst-case analysis where the intermodulation *voltages* from each stage add in phase. Our measurements indicate that this analysis works well in practical systems, so long as the individual stages are well-behaved, as defined earlier.

Receiver dynamic range is related to intercept and MDS by a simple equation. MDS is further related to bandwidth and noise figure, offering a more general equation.

$$DR(dB) = \left(\frac{2}{3}\right)(IIP3 - MDS)$$

$$= \left(\frac{2}{3}\right)(IIP3 + 174 - NF - 10 \log(BW))$$

Eq 6.15

where IIP3 is the input third order intercept, NF is system noise Figure, BW is the system bandwidth. Recall that kT = –174 dBm at 290 K, explaining that term in the equation.

Some Front-End Design Examples

We are now in a position to evaluate some receiver front-end designs. A few examples will be presented using data obtained from measurements we have performed.

The first example is a popular one among the QRP clan, a receiver front end based upon the Phillips NE602 or NE612. Our evaluation data was presented in Chapter 5. A front-end block diagram, **Fig 6.64**, includes gains, intercepts, and noise figures for the stages. The result of applying the dynamic range analysis is also included. This is a simple design with only one active block, the mixer. The dynamic range is modest at 83 dB, although sensitivity is quite good. The noise figure

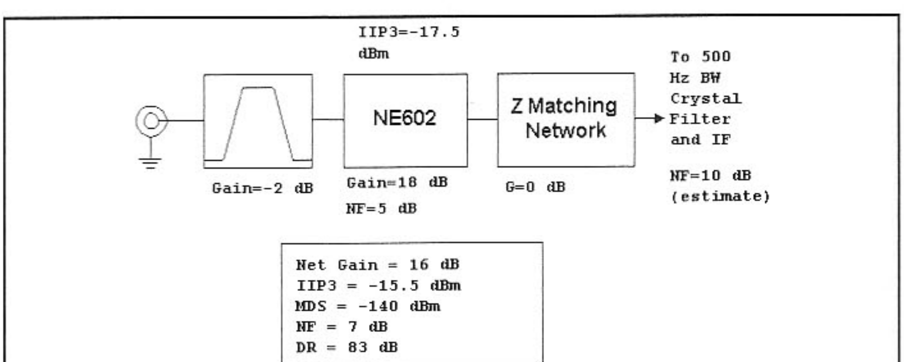


Fig 6.64—A simple receiver front end using the NE602. The IF system is estimated to have a noise figure of 10 dB.

is essentially that of the IC plus the insertion loss of the bandpass filter preceding it. Care must be exercised in implementing this design if this DR is to be realized. For example, chip intercept could be altered if output is extracted only from one output terminal. On the other hand, careful mismatch at the input may decrease gain to actually increase input intercept with only a modest noise figure change. Some builders claim a 90-dB dynamic range with NE602 front ends with this bandwidth. Clearly, careful measurements are always worthwhile.

In spite of the good MDS obtained from the NE602, some builders are tempted to add an RF amplifier. In other situations, an NE602 is used as a second mixer in a receiver, having been preceded with gain. The trade-off is illustrated in **Fig 6.65**. A bandpass filter with a 1-dB loss is followed by a low gain RF amplifier. The signal then passes through the original 2-dB-loss filter before arriving at the mixer. This design offers a 2-dB improvement in sensitivity, but at the price of a 5-dB decrease in dynamic range.

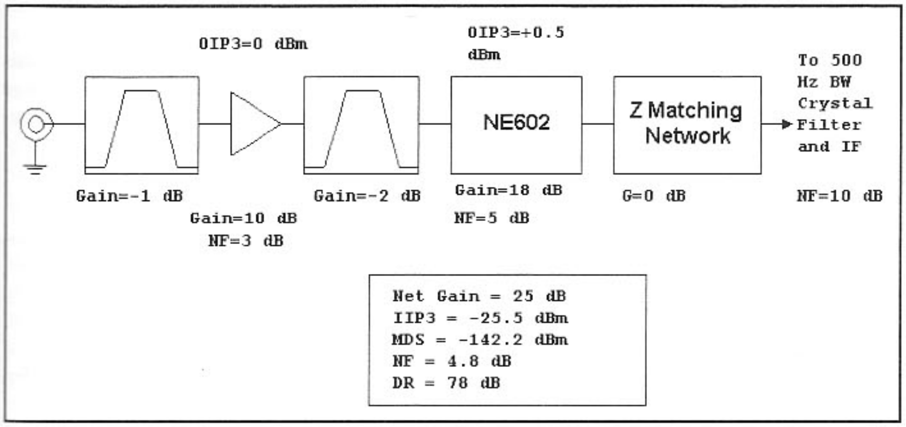


Fig 6.65—An RF amplifier is added to the previous design, offering slightly improved MDS at the cost of degraded dynamic range.

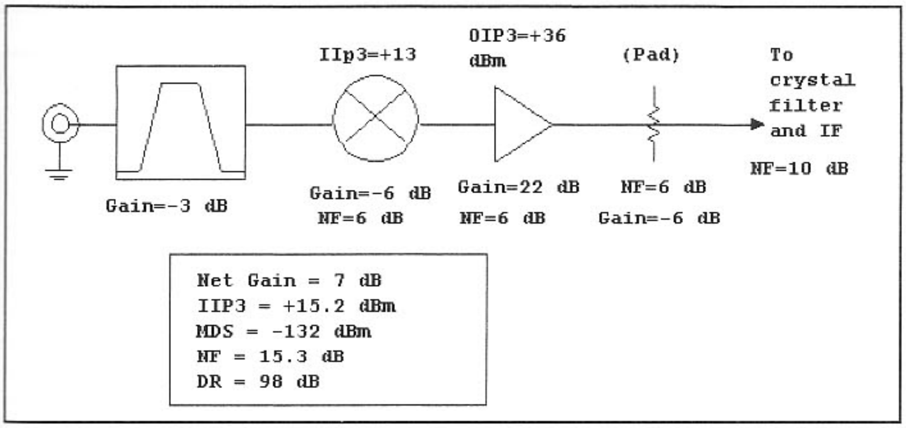


Fig 6.66—Basic front end with a diode-ring mixer followed by a high-current bipolar feedback amplifier.

The next sample front end, **Fig 6.66**, is the opposite extreme. Here we use a diode ring mixer as the first element, followed by a post mixer amplifier with high current. This is the sort of front end we recommend for the 160, 80 or 40-m amateur bands where low noise figure is rarely needed. Although MDS is 8 to 10 dB higher than the previous designs, dynamic range is 98 dB. The mixer in this design is a +7-dBm-type ring such as the Mini-Circuits SBL-1, TUF-1 or TUF-3. If an even stronger TUF-1H was substituted, DR over 100 dB is easily within reach in a simple design. The post mixer feedback amplifier would ideally use a part specified just for this application, such as the 2N5109 with 40 or 50 mA. However, a parallel pair of 2N3904s will do a surprisingly good job, again with 40 mA of total current.

Many builders question the use of a passive mixer with no gain. But it is exactly this lack of gain that leads to the low noise. The passive nature of the circuit eliminates the noise-generating elements that compromise some other mixers. There is no substitution for actual design.

The high noise figure of the bare-ring-mixer front end is usually not suitable for the higher bands. The designer will often want to add an RF amplifier to obtain lower NF. This modification is illustrated in **Fig 6.67**. The modest RF amplifier improves sensitivity by several dB while only reducing dynamic range by 2 dB. Too much RF gain could severely compromise performance.

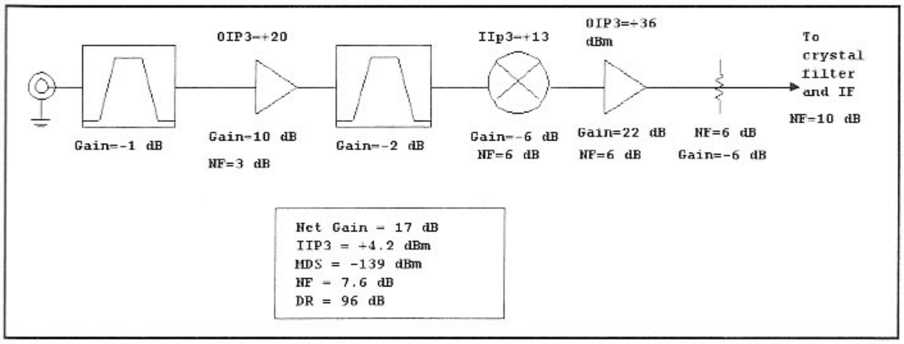


Fig 6.67—An RF amplifier is added to the basic diode-ring front end, significantly improving noise figure while compromising DR by only 2 dB.

The Receiver Factor

The two-tone dynamic range presented above has a major disadvantage as a receiver figure-of-merit: DR is a strong function of bandwidth. This is a direct result of MDS used in the DR equation. A CW receiver with a 500 Hz bandwidth will produce a higher DR than a SSB design with much wider bandwidth. Measurements of MDS are difficult, often complicated by un-planned filtering in the receiver audio section. While this filtering may or may not have much impact on the way a receiver sounds, the measured results are altered.

Both input intercept and noise figure for a receiver are generally bandwidth invariant parameters. The first is a measure of strong signal performance while the other defines weak signal behavior. They can be combined by taking the difference. We call this the receiver factor, **R = IIP3-NF**. The receiver using a diode ring front end without RF amplifier, Fig 6.66, had R=0 dBm while the NE602 receiver with an RF amplifier, Fig 6.65, provided R = –30.3 dBm. While both sample receivers used a CW bandwidth, the R-values would be the same if they were built with SSB filters. Later in this chapter we will describe a receiver with an astounding R = +35 dBm!

The noise figure, and hence, the receiver factor may change slightly with bandwidth with some receivers. This is usually the result of differing filter insertion loss as bandwidth is switched.[14]

A General Purpose Monoband Receiver Front End

Although there are numerous routes to the construction of a high performance front end, a dependable robust topology consists of the following cascade:

- A simple bandpass filter;
- A low-gain RF amplifier;
- A bandpass filter with two or more resonators;
- A diode-ring mixer;
- A post-mixer amplifier using a low-noise bipolar transistor with negative feedback;
- An attenuator that creates a stable impedance at both the output and, through the behavior of the feedback circuitry, the input of the post mixer amplifier;
- A crystal filter;
- And finally, an IF amplifier.

Generally, receivers designed with this front end have produced dynamic range within a couple of dB of the values pre-

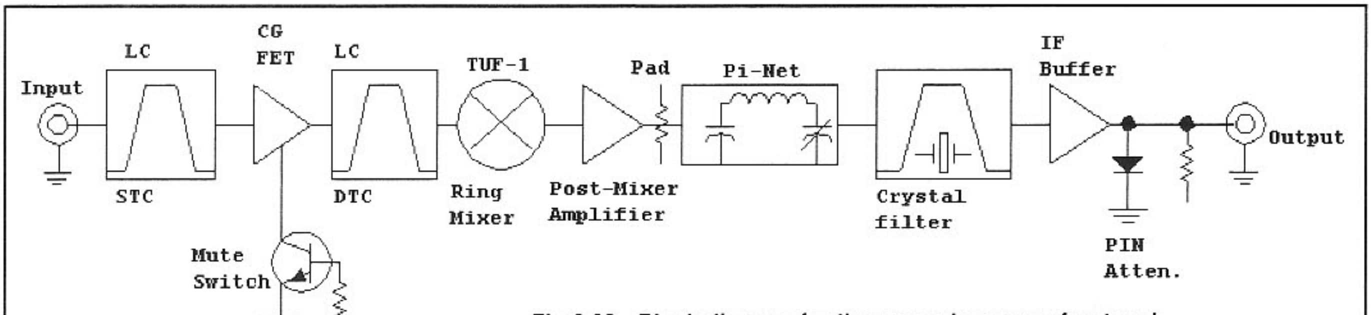


Fig 6.68—Block diagram for the general-purpose front end.

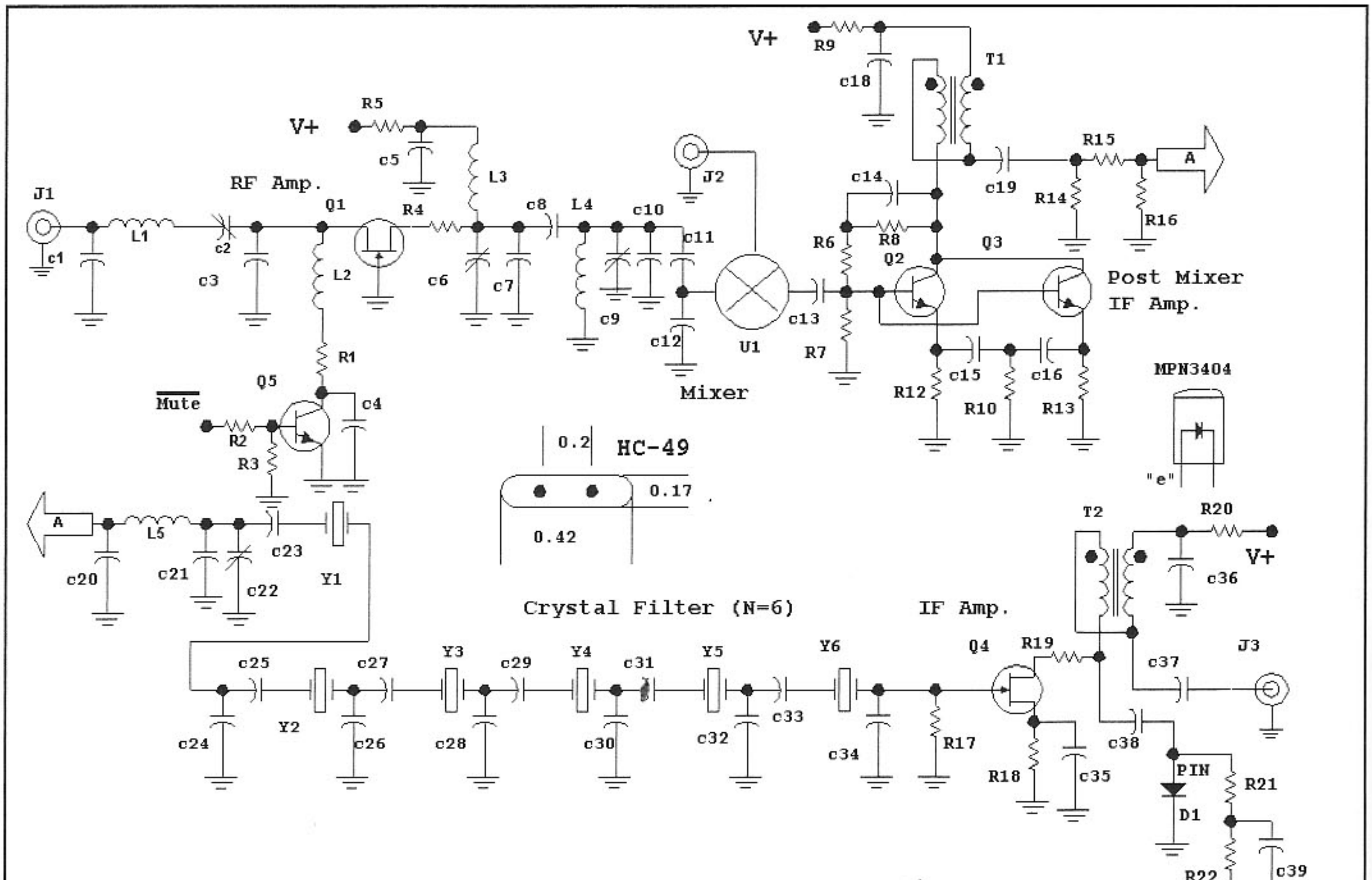


Fig 6.69—Schematic for the general-purpose front end. See text for details.

dicted by the analysis presented when using measured data for the individual stages. The block diagram for this front end is shown in **Fig 6.68**.

A small circuit board was designed and fabricated for this front end and includes a crystal filter of up to 6 crystals. The 50-Ω impedance of the pad is increased with a pi-network to whatever value needed by the filter. The other end of the crystal ladder is terminated in the proper resistor and a common source JFET amplifier. A PIN diode attenuator is also included in the IF amplifier output for those applications where no other IF gain control is available. A muting switch for the RF amplifier is also included. The complete schematic is given in **Fig 6.69**.

The input pre-selector filter is a single tuned circuit. It begins as a 3-element low-pass filter, but the usual inductor is replaced with a series tuned circuit. This simple topology degenerates into a low pass filter in the VHF stopband, a useful attribute when trying to avoid spurious responses related to stray VHF signals.

The second bandpass filter, a double-tuned circuit, appears after the RF amplifier where noise figure has been established. Insertion loss is not as critical as it might be without the amplifier. This means that the filter bandwidth can be narrow enough to ensure very good image rejection. It also allows us to use small toroid cores, if desired.

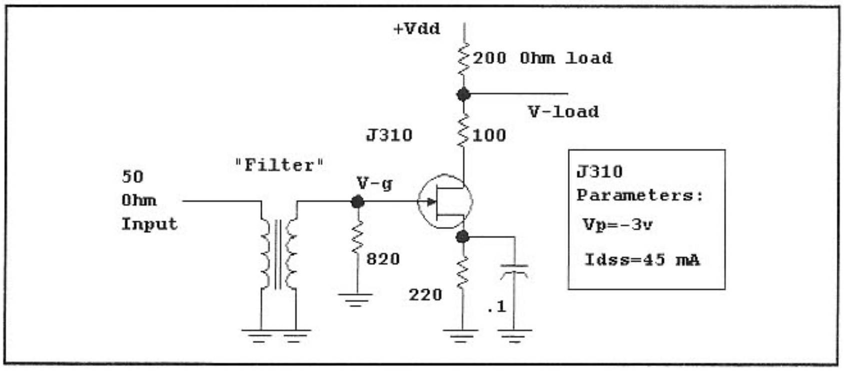
Two bandpass filters should be used in designs that include an RF amplifier. An RF amplifier that is not preceded by a filter is subject to overload from local signals, particularly the strong VHF broadcasts that most of us experience. A filter should also appear after the RF amplifier, immediately preceding the mixer. This circuit, often termed the image-stripping filter, establishes image rejection. If it was only present ahead of the RF amplifier, it would not suppress noise at the image frequency that is created by the RF amplifier.

The RF amplifier we chose is a common-gate JFET design. It is capable of very low noise figure while offering good intermodulation distortion and high power output when needed. It also can have very good reverse isolation, serving to suppress signals at the mixer that would otherwise find their way to the antenna terminal. But it can also be challenging, for the common gate FET amplifier can tend to oscillate. The spurious oscillations, which usually occur at a few hundred MHz, occur when the layout is poor or leads are too long. Generally, too much fuss is propagated in much of the electronics literature regarding long leads in solid-state circuitry, but this is a place where it really does matter, particularly with the FET gate lead.

A cure for the instability is resistance in series with the drain. This is not a mere experimental band-aid, but a circuit detail justified with analytic evaluation. Greater resistance generates even better stability. We have used 100 Ω in this application, for it provides margin without altering the low frequency (HF and low VHF) gain. The resistor should be placed as close to

Gain of a JFET Amplifier

The IF amplifier used in the output of the general-purpose receiver front end is a common-source configuration with a transformer output presenting a 200-Ω load to the FET drain. Amplifier gain depends on the impedance presented to the input.



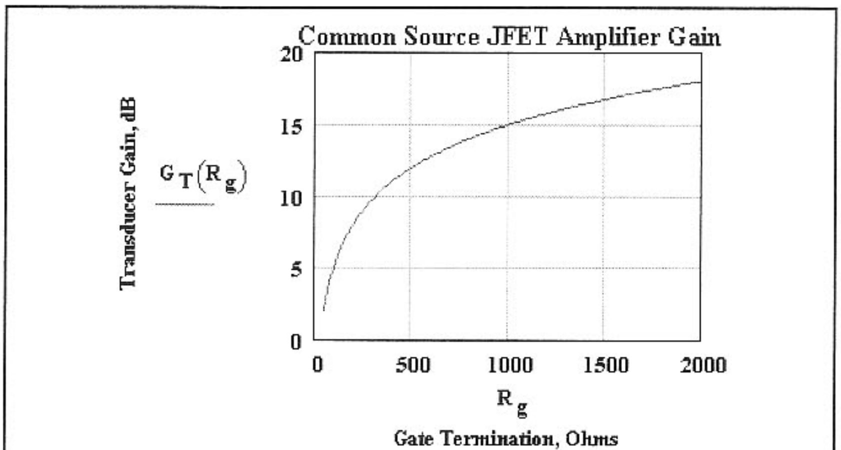
The "filter" is the combination of an impedance-transforming network and a crystal filter in this instance. The 50-Ω source is transformed to match a higher resistance, 820 Ω in the schematic above, with the composite "filter."

If 1 mV is presented to the input, the voltage at the gate will be increased by the square root of the impedance ratio, here a factor of 4.05. So, Vg = 4.05 mV. The FET bias current is 7.92 mA in this instance, so the transconductance is gm = 0.0126 S, using equations presented in Chapter 2. The drain signal current is then

GM•VG = 0.051 milliampere.

This current develops an output voltage across the 200 Ω load, Vout = 10.21 mV. (The 100-Ω resistor is significant only in reducing the effective supply voltage. It is included to suppress parasitic oscillations.)

Output power is $V^2/200 = 5.21 \times 10^{-7}$ W. But the available input power is 1 mV across 50 Ω, or 2×10^{-8} W, so transducer power gain is 26, or 14.2 dB. The important detail here is that power gain is a strong function of the impedance terminating the filter, shown in the curve below.



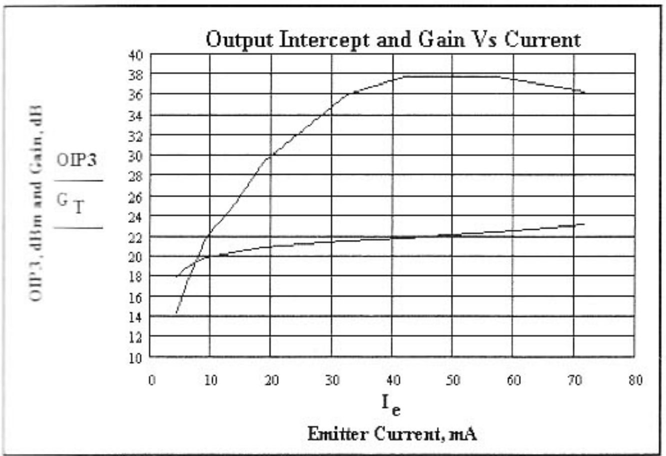


Fig 6.70—Gain (lower curve) and output intercept for one or two 2N3904s in parallel. Two devices should be used for currents above 20 mA, while total current over 40 mA is not recommended except as an experiment.
T1 = 10 bifilar turns on an FT37-43.
R9 = 47, R8 = 1kΩ, R6 = 1.5 kΩ, R7 = 680 Ω, R10 = 6.8 kΩ,
R12 = R13, which are picked to set the dc emitter current.
R12 = R13 = 100 Ω for 30 mA total current.
C13,14,15,16,18,19 = 0.1µF.

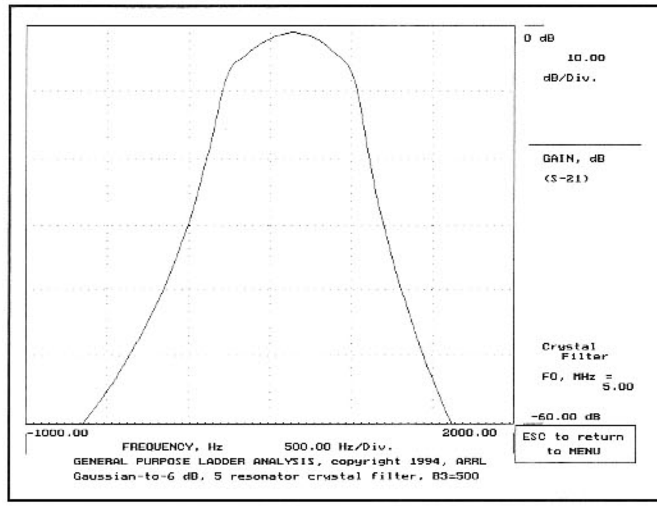


Fig 6.71—Calculated response for the Gaussian-to-6-dB crystal filter. The shape is Gaussian for the top 6 dB, but then reverts to a Chebyshev-like skirt response. The k and q data for this filter were obtained from Zverev's *Handbook of Filter Synthesis*, Wiley, 1967.

the FET as the board layout or breadboard allows. A simple shielding method for a J310 RF amplifier was shown earlier in this chapter. The shield was not needed on this circuit board.

The RF amplifier output resistance is around 10,000 Ω. That value was used while designing the input termination for the double tuned circuit while the output is set for a 50-Ω termination.

The RF amplifier FET is biased on when the NPN switch is saturated. The builder should design control circuitry to apply a positive voltage to the control input during receive intervals.

This module uses mixers in the TUF family from Mini-Circuits. Either the TUF-1 or TUF-3 should work well with +7 dBm of LO power. A high level mixer (TUF-1H or TUF-3H with +17-dBm LO power) will also fit in the board and will provide even higher dynamic range, but only when followed by an adequately strong post-mixer amplifier. The mixer is generally the DR defining element within the system.

The post-mixer amplifier is a critical element. Enough current should be used to guarantee the desired dynamic range. However, too much current can also be wasteful, especially in applications where batteries are used. The layout used in the general-purpose board is for two paralleled 2N3904s, shown in Fig 6.69. Resistors R12 and R13 determine the total current, which should be equal. Only one transistor is required if total current is 20 mA or less. Gain and output intercept are presented vs total amplifier current in **Fig 6.70**. A home station design where power is abundant might use 30 or 40 mA while 10 mA may be enough for a portable application. No heat sink has been needed for a pair of 2N3904s at 40 mA total current. Larger transistors with higher power dissipation ratings can, of course, be used.

The designer/builder must design the crystal filter for the desired bandwidth. While the board will accommodate up to 6 crystals, fewer may suffice. In one application using a 5-crystal CW bandwidth filter, we found that stopband attenuation was less than indicated by calculations. Two measures restored performance: First, all crystal metal cases were grounded to a wire bus. Second, a shield was soldered to the ground foil between the crystal filter and the post mixer amplifier.

The builder/designer has considerable flexibility available when choosing the terminating resistance for a crystal filter. This choice impacts the design of the IF amplifier. The design procedure is summarized in the *Gain of a JFET Amplifier* sidebar. Higher gain is available with the higher impedance values.

The PIN diode will provide up to 30-dB attenuation. This is especially handy for applications where no additional IF gain is used.

The Easy-90 Receiver

The general-purpose front end was used to build a simple receiver for the 20-m CW band, dubbed the *EZ90-14C*. The 90 indicates a two-tone dynamic range in excess of 90 dB, which is achieved with ease with this receiver. The receiver architecture is one without an IF/AGC amplifier. Front-end parts are tabulated in the following list.

The 5-element 5-MHz crystal filter for this receiver was designed for a 3-dB bandwidth of 500 Hz and a Gaussian-to-6-dB shape. This shape has the virtue of a good time-domain characteristic, keeping ringing to a minimum in a narrow filter. The stopband attenuation is still reasonable. An added virtue of transitional filters, including this Gaussian-to-6-dB, is a relative insensitivity to exact component value, allowing a minor degree of "slop" when being constructed. On the down side, this filter lacks the familiar circuit symmetry of Butterworth and Chebyshev designs. We built this 5-MHz filter with available crystals that had good Q, often over 200,000. Crystal frequencies were matched to within 10 Hz. Design details are presented in Chapter 3. A calculated response for this crystal filter is shown in **Fig 6.71**.

Several different filter designs were tried in this receiver. While a Cohn design worked, it used a terminating resistance under 200 Ω. This severely impacted the IF amplifier gain as outlined in the IF sidebar. (A Cohn type crystal filter is, of course, possible with a higher terminating impedance, but the simple design method presented in Chapter 3 is then invalid.) A Gaussian-to-6 dB filter with a 250-Hz bandwidth and 500-Ω terminations worked well, but was too narrow for the intended application

The front-end board output is routed directly to the product detector, shown in the detector-audio board in **Fig 6.72**. This

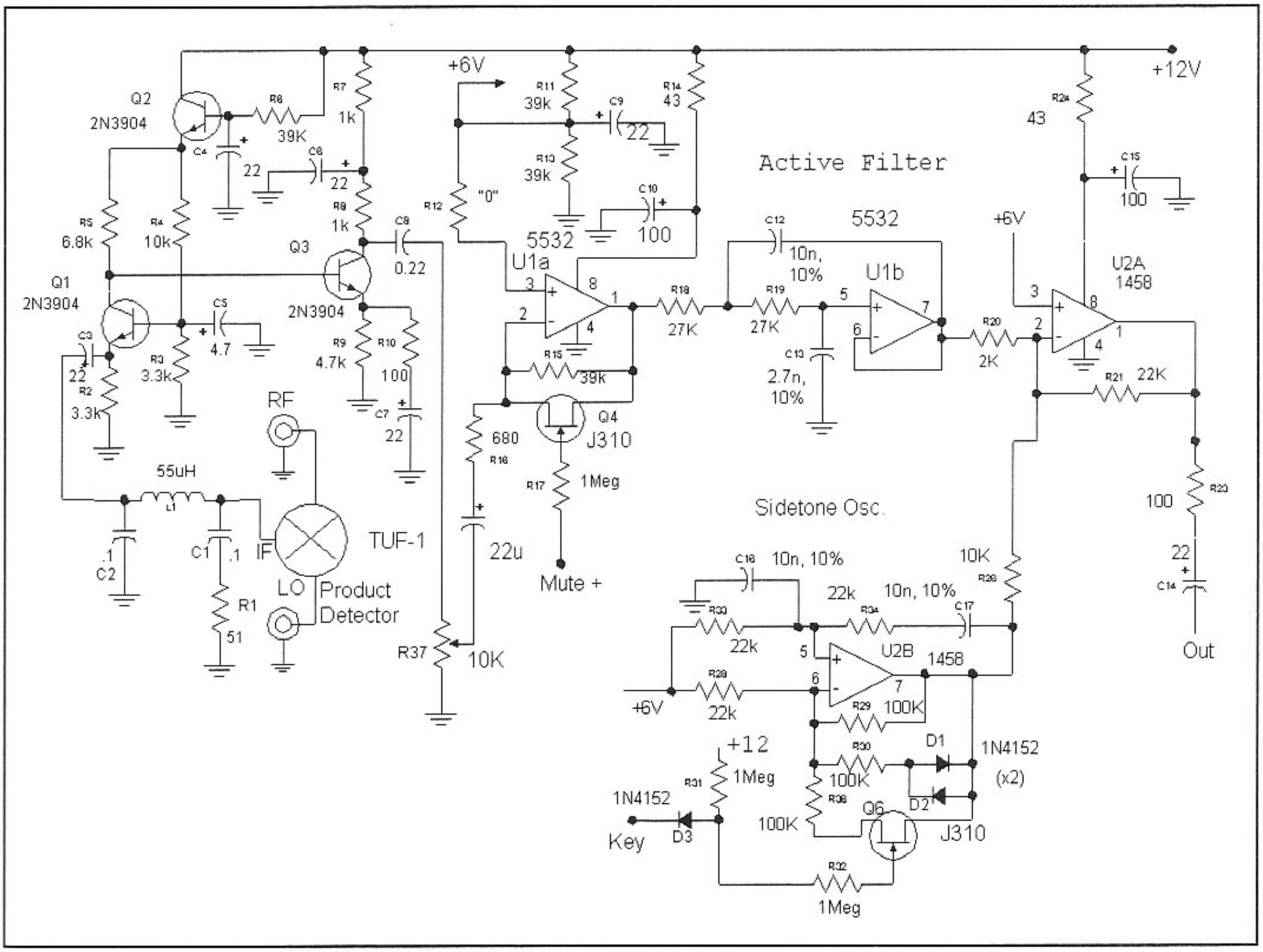


Fig 6.72—Audio amplifiers, product detector, and sidetone oscillator for the EZ-90C receiver.

EZ90-14C

Parts List for the 20-Meter "Easy 90" Receiver

C1,C3: 470 pF SM or NP0 ceramic
C2,C6,C9,C22: 65 pF, 10 mm air variable (Sprague Goodman GYC65000)
C4,5,13,14,15,16,18,19,35,36,37,39: 0.1 µF
C7:82 pF
C8: 2.2 pF
C10: 56 pF
C11: 22 pF
C12: 200 pF
C20: 820 pF
C21: 220 pF
C23: 470 pF
C24: 68 pF
C25: short circuit
C26: 100 pF
C27: 150 pF
C28: 100 pF
C29: 100 pF
C30: 150 pF
C31: 100 pF
C32: 82 pF
C33: short circuit
C34: not used
Q1, Q4: J310
Q2, Q3, Q5: 2N3904
D1: MPN3404 or similar PIN diode
L1: 27t #28 on T30-6
L2, L5: 4.7 µH molded RFC, Q>=50
L3, L4: 1.04 µH, 16 t #28, T30-6
T1:T2 10 bifilar turns #28, FT37-43
R1: 180
R2, R3: 10 kΩ
R4: 100
R5: 47
R6: 1.5 kΩ
R7: 680
R8: 1 kΩ
R9: 47
R10: 6.8
R12, R13: 100
R14, R16: 150
R15: 36
R17: 820
R18: 220
R19:100
R20: 47
R21: 1 kΩ
R22: 680
U1: TUF-1 or TUF-2 or TUF-3
Y1, 2, 3, 4, 5: HC49 crystals, 5 MHz, Lm=98 mH, C0=3 pF (see text)
Y6: not used; add short circuit

module design has been used in several projects. A TUF-1 provides the detector function. Bipolar audio amplifiers drive an audio gain control, followed by an op-amp providing gain and an RC active low pass filter with a peak at 700 Hz. The Q is kept low in this version. The audio is muted with a shunt FET switch.

The BFO for the product detector is shown in **Fig 6.73**. This is breadboarded on a small scrap of circuit board material.

Fig 6.74 shows a 9-MHz VFO for the EZ90-14C. The oscillator is a voltage-tuned Colpitts circuit purposefully configured for low inductance. The high fixed-tank capacitance is desirable for low phase noise. This LO produces a narrow tuning range of about 20 kHz with the available tuning diode. This receiver is used with a transmitter with restricted tuning range, so the narrow range is acceptable. The builder/designer may wish to use a combination of varactor tuning and a traditional variable capacitor to achieve a wider tuning range. Alternatively, higher L could be used to cover the entire CW band with a varactor diode.

The VCO output is extracted from a FET follower that then drives a power amplifier to provide the +7 dBm LO power needed by the ring mixer. Power amplifier degeneration is adjusted to set output level. An 8-V regulator supplies the VCO. It also provides a stable bias for the tune pot and a stable 4 V for an op-amp reference. The gain and offset in the op-amp are set up to supply a 5 to 10 V swing on the varactor diode.

A receiver noise figure measurement produced NF = 6.6 dB. If a noise bandwidth of 800 Hz is used with this, MDS of –138 dBm is suggested. However, a direct measurement of MDS produced –141 dBm. The difference is attributed to the narrow audio filter that restricts overall noise bandwidth. DR measurement produced a value of 95 dB, for IIP3 = –1.5 dBm. Using this value for IIP3, receiver factor is R = –8.1 dBm.

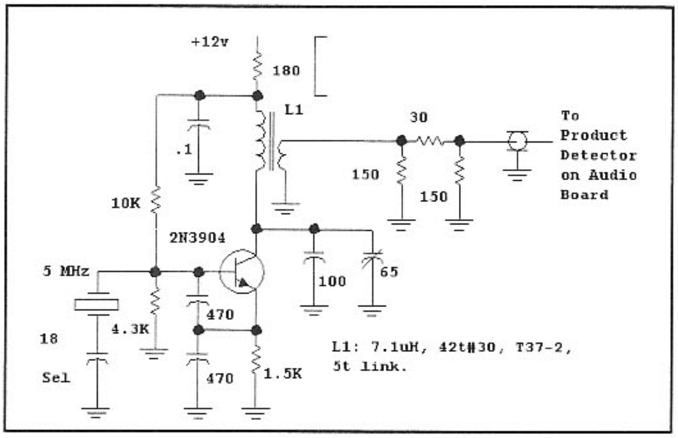
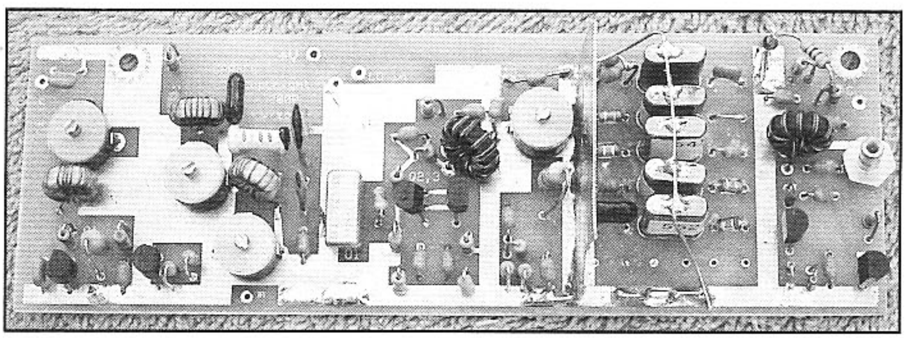


Fig 6.73—BFO for the EZ90-14C. A variable capacitor can be used in series with the crystal for final adjustment. It was replaced with a fixed capacitor in our receiver.



General-purpose receiver front end board used in the EZ90-14.

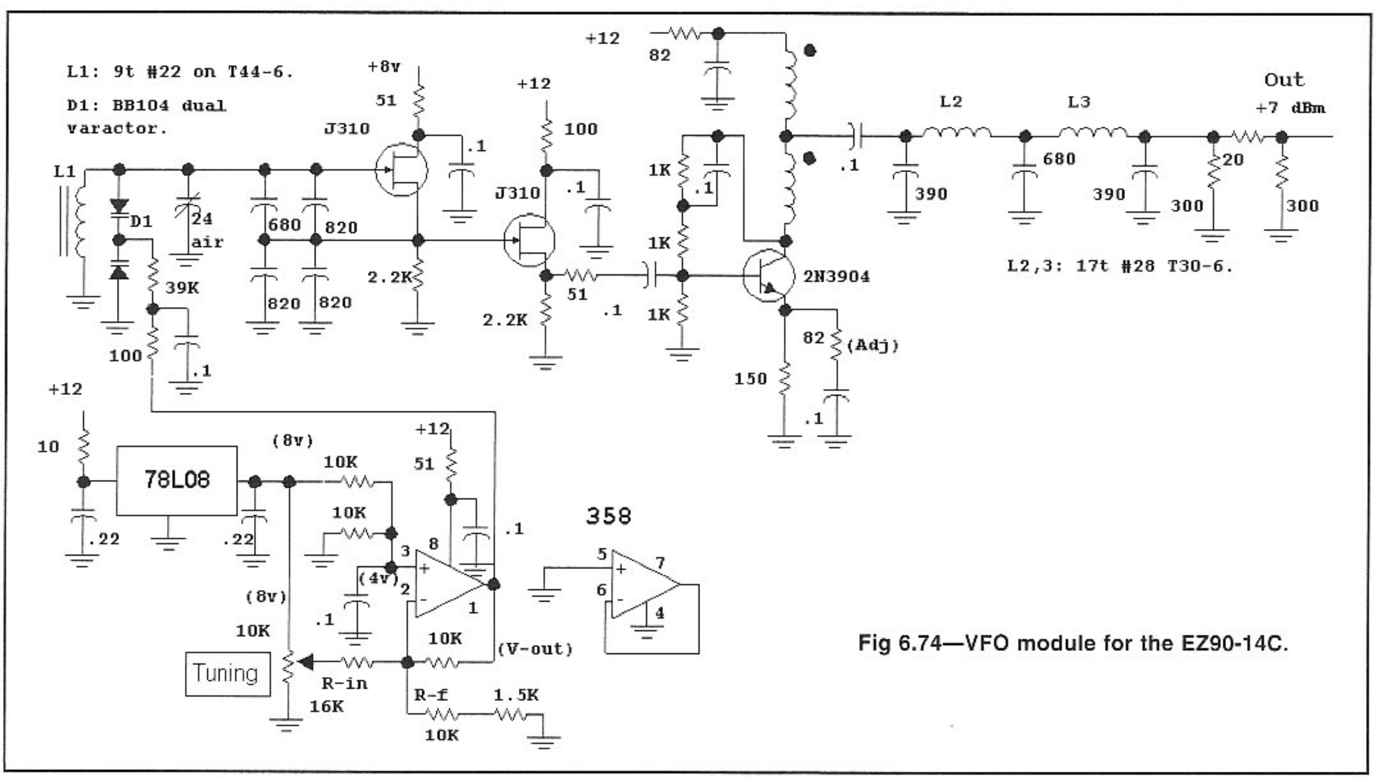


Fig 6.74—VFO module for the EZ90-14C.

The receiver is packaged with a 14 MHz VXO transmitter described in Chapter 5. The narrow receiver tuning range eliminates most birdies from being a problem. In spite of this, one was encountered in the form of a feedthrough of 15-MHz WWV energy. This signal got into the enclosure on the antenna connector where it then found it's way onto the grounds that reached the product detector. There, the normal third harmonic response of the diode ring allowed the 15-MHz component to be directly converted, to produce baseband audio. The problem was eliminated with a 5-MHz low-pass filter inserted in the line between the front end and the detector audio board. The problem would never have occurred if the receiver had not been built with completely unshielded boards.

Generally this receiver will hold up well in a contest environment, although we find it in need of some AGC for those moments when a really strong signal is encountered. Limiting in the audio output op-amp produces a clipped response when the strong signals appear, saving the operator's ears. The very "hot" receiver (low MDS) was designed for portable situations where noise levels are much lower than we find in a home environment.

A 14-MHz Receiver

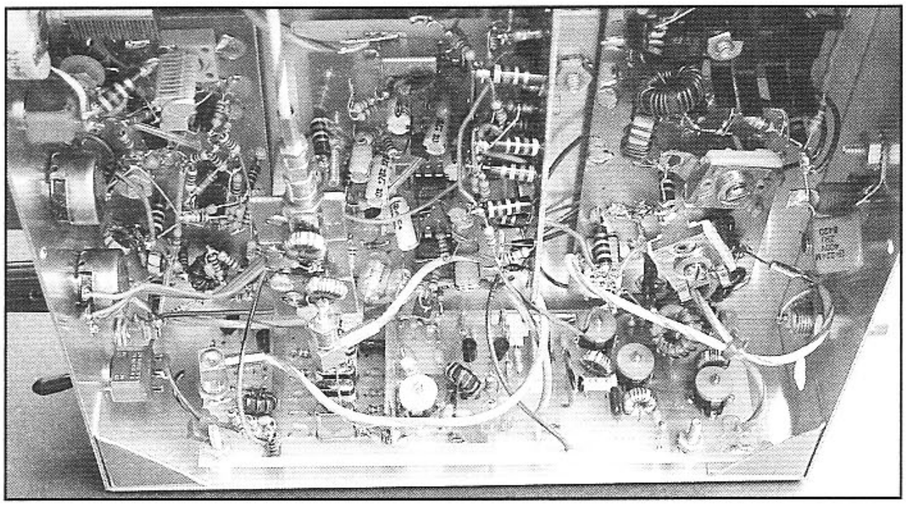
This receiver is an updated version of two earlier designs.[15] The changes include repackaging (smaller size) with improved shielding, a new frequency counter with lower power requirements, and a reduced noise IF system. This receiver is similar to the EZ90, but features the shielding needed for high dynamic range.

The receiver is a CW only design using filters with reasonable time domain characteristics. While these filters are no longer available, it should be possible for the aggressive builder to build viable substitutes. The 9-MHz IF system was described earlier in detail in Fig 6.56. The design features three stages of gain using dual-gate MOSFETs and crystal filters at both the IF input and output. The IF circuitry is built with breadboards into a multiple section milled aluminum enclosure.

The front end (**Fig 6.75**) begins with a bipolar RF amplifier biased to $I_e = 12$ mA, which produces low noise figure while maintaining an intercept that is high enough to not degrade overall receiver IIP3. The amplifier is preceded by a single resonator preselector and followed by a double tuned image-stripping filter.

The mixer uses a TUF-1 with +7 dBm LO drive. A higher LO level is applied to a 3 dB hybrid that splits the signal into two isolated components. One drives the mixer while the other is attenuated and available for transceive applications. The mixer has two inputs, selected by a small relay. One is the normal 14 MHz signal from the double tuned circuit while the other comes from other equipment at either 4 or 14 MHz. The mixer output is applied to the familiar feedback amplifier and pad combination. The front end is housed in a 4 × 4 × 1 inch milled aluminum box.

The BFO and Product Detector, shown



General-purpose receiver front end board installed in the EZ90-14 Receiver.

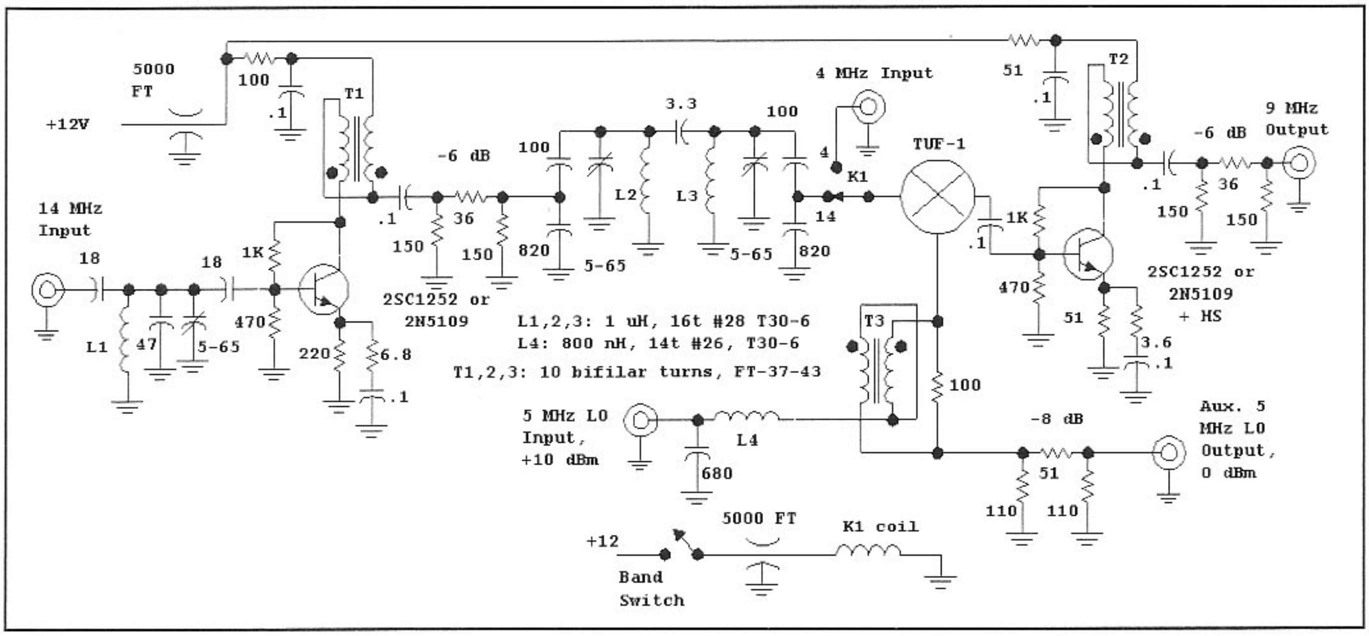
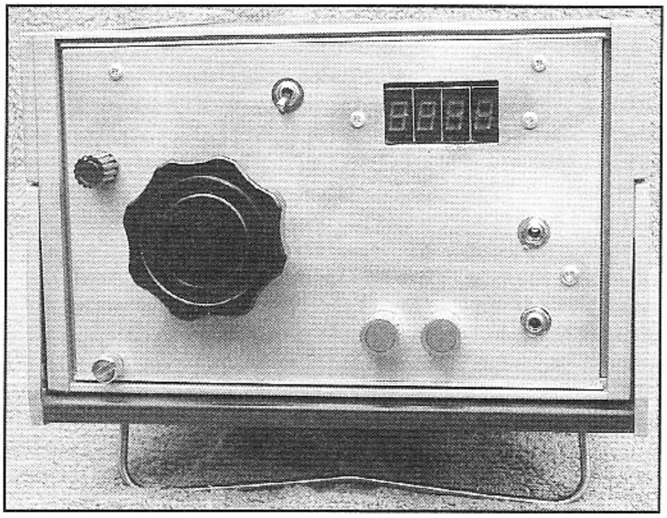
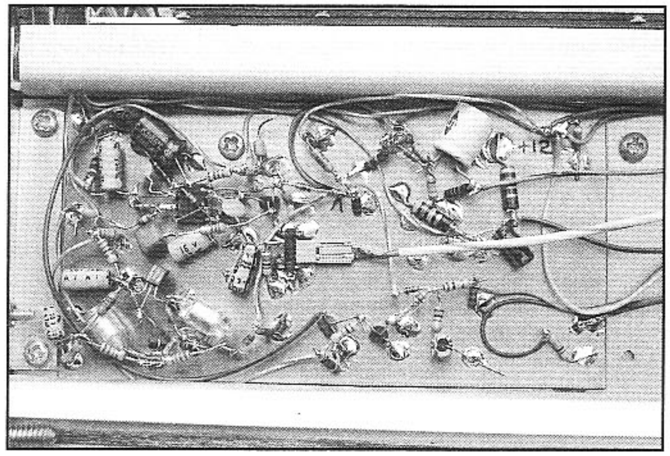


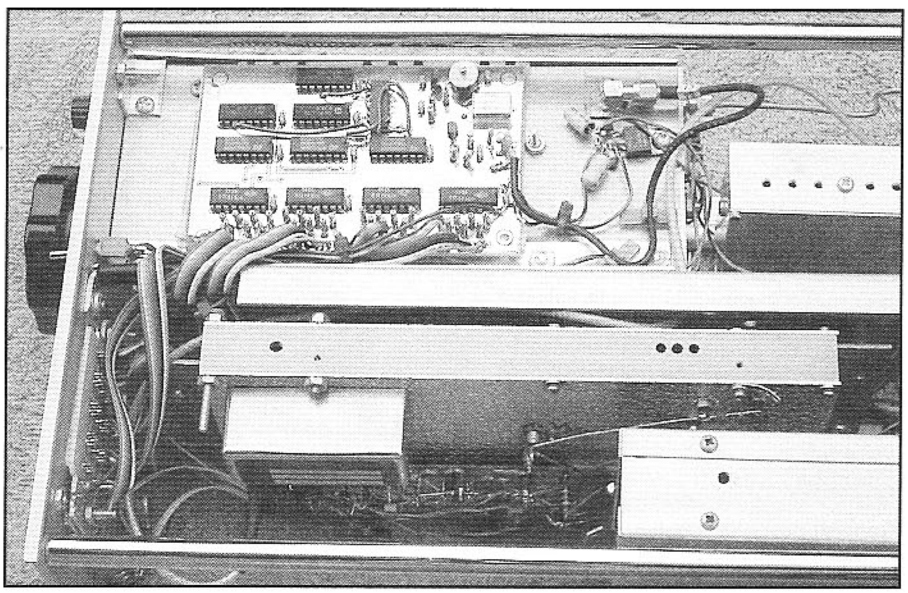
Fig 6.75—Front end for the 14-MHz receiver. The circuit is built largely with breadboarding methods.



Front panel view of receiver.



Close up view of audio amplifiers.



Inside of 14-MHz receiver. Upper left is the frequency counter, upper right is the front end, middle is IF chain, and lower right is product detector/BFO.

in **Fig 6.76**, is traditional. A diode ring moves the 9-MHz IF signal to base band while a bipolar transistor serves the BFO function.

The 5-MHz local oscillator is shown in **Fig 6.77**. The design uses a Colpitts VFO with a JFET. A JFET buffer drives a feedback amplifier output stage. The output power is large enough to drive the hybrid splitter and mixer in the front-end module. Varactor diode tuning will eventually be added to provide an RIT function. The related CMOS frequency counter was described in Chapter 4.

The receiver audio system is shown in **Fig 6.78**. U1 provides audio gain, muting, and a convenient place to inject a sidetone signal. This drives an audio gain control and the output stage, U2 and Q2. The output operates as a class A amplifier with a standing current of about 90 mA. This will drive a small speaker or headphones of virtually any impedance. The high current is not a prob-

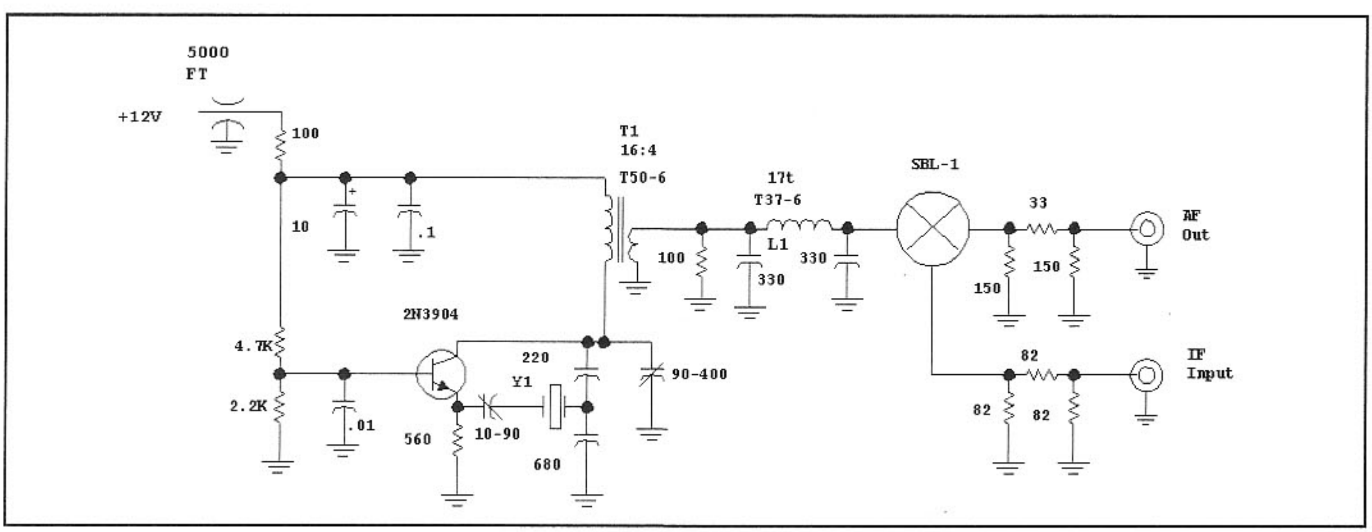


Fig 6.76—BFO and Detector for the 14-MHz receiver.

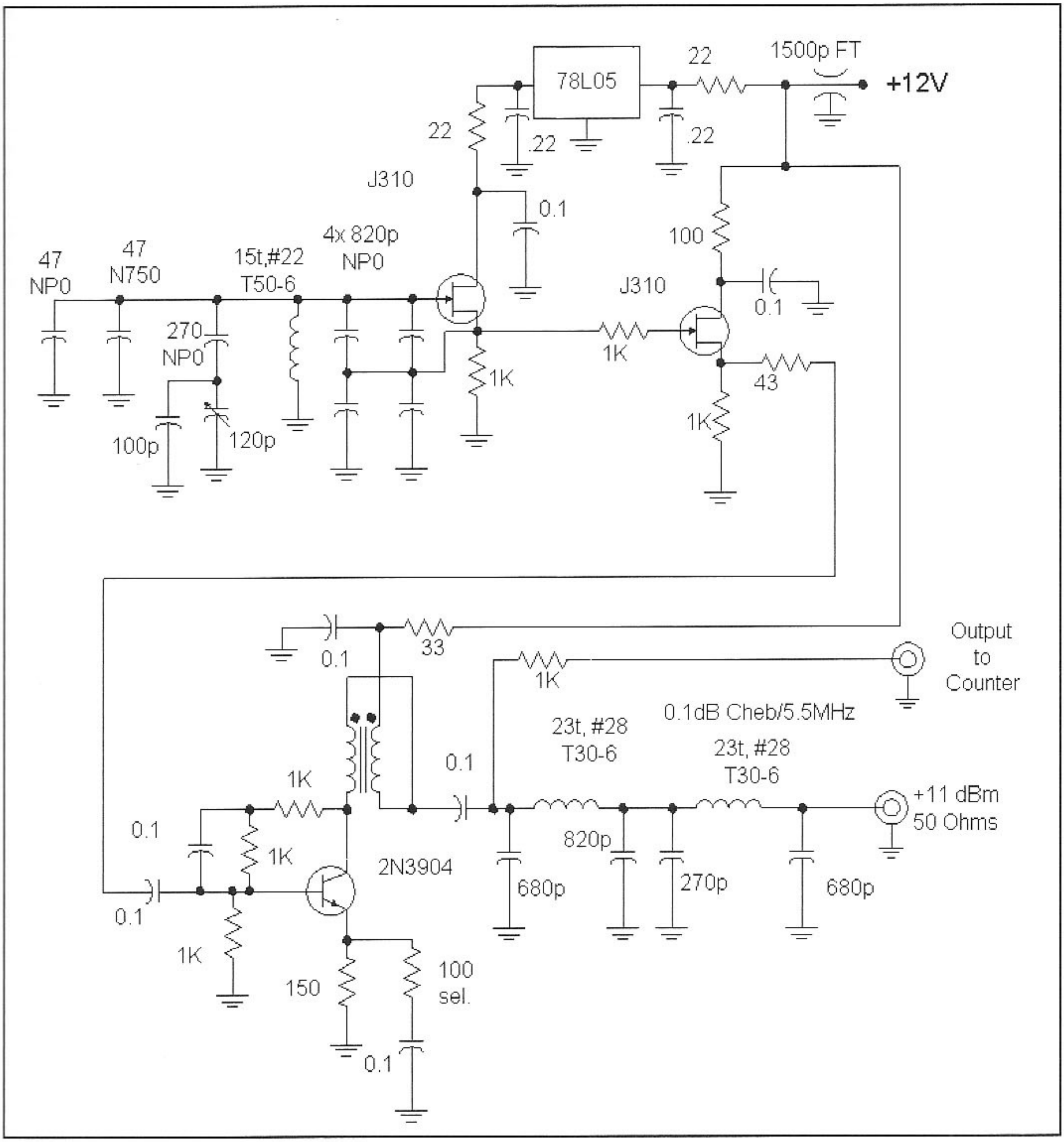


Fig 6.77—LO system for the 14-MHz receiver. The N750 capacitor provides temperature compensation as measured with a small homebuilt thermal chamber. All other capacitors in the oscillator have an NP0 temperature coefficient.

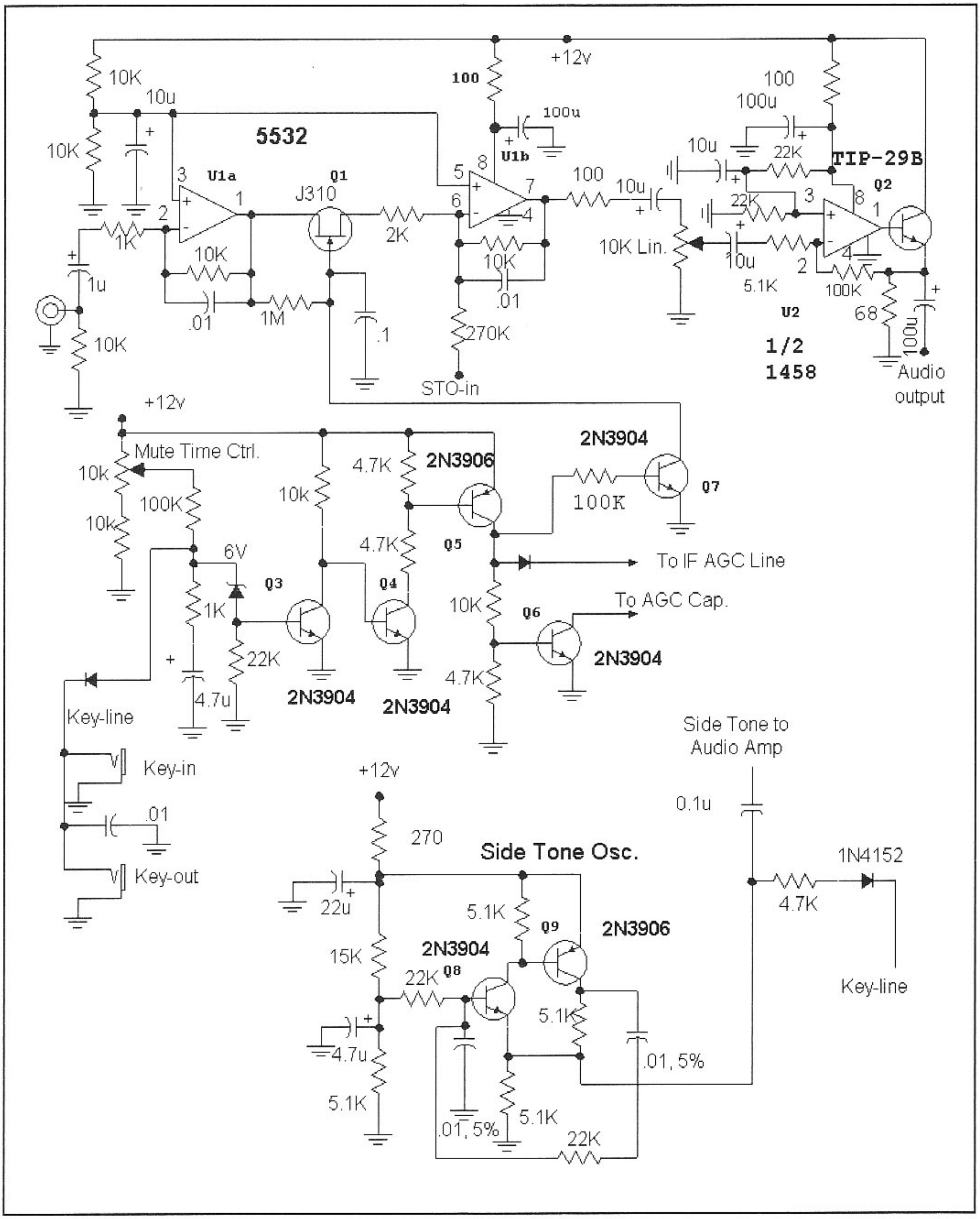


Fig 6.78—Audio and control system for the receiver. See text for details.

lem, for the receiver is used only in a home environment. Q3 and related components generate a time delay, establishing the time the receiver is muted following a key closure. Placing the function in the receiver allows use with many transmitters that may not include interface circuits. The key line loops in and out of the receiver.

Q8 and Q9 form an unusual Weinbridge sidetone oscillator. In key-up conditions the two transistors and the two 5.1-kΩ emitter resistors form an amplifier with a non-inverting gain of two. This is not high enough to support oscillation. But when the key is pressed, the 4.7-kΩ resistor causes the voltage gain to exceed 3, allowing oscillation to begin. The frequency is determined by the 5% capacitors and 22-kΩ resistors. Oscillator output is obtained from the emitter of Q8. This point does not change dc value as the circuit is keyed, preventing a keyed voltage spike in the audio.

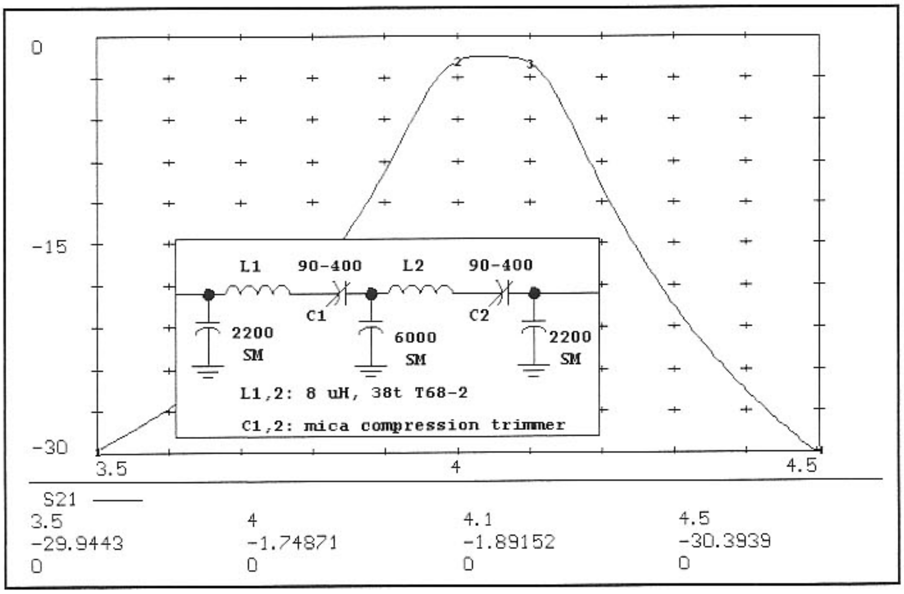


Fig 6.79—Filter for use at the output of crystal controlled converters to be used with the 4-MHz input in the 14-MHz receiver.

Overall Results

This receiver is a design that has evolved for several years, so the performance is fairly stable. Prior to a major rebuild in 1998, the receiver used an IF based upon MC-1350P integrated circuits. While adequate, the noise performance was marginal. Receiver noise figure is now maintained as IF gain is reduced, producing a receiver that continues to sound "bright," when used for weak or strong signals.

Noise figure was measured as 7 dB. The measured MDS was around −141 dBm while IIP3 was +1.5 dBm for DR of 95 dB. The LO system, although difficult to evaluate, seems to a have phase noise less than −140 dBc/Hz at a 5 kHz carrier offset. Thermal stability is excellent, although this occurred only after a minor struggle. Examination showed that an RF choke in the oscillator FET source had poor temperature characteristics. Removal of that component and further compensation produced a stable oscillator, illustrating the virtue of careful testing and response to test results. The LO, although lacking the control features of a synthesized system, is completely free of spurious responses. The receiver is just as much fun to use as the original was in 1974.

Converters

The receiver has been used with crystal-controlled converters for numerous bands. Although a traditional dual conversion system does not offer the dynamic range of a single conversion design, it can be close if converter gain is kept low. The typical converters consist of a preselector filter, a diode ring mixer with crystal controlled oscillator, a post mixer amplifier, and pad. An RF amplifier is used for the higher bands. Some sort of 4-MHz bandpass filter is then required to guard against any second conversion images. One filter we have used is shown in **Fig 6.79** with calculated response. The filter may reside with the converter or with the basic receiver. All of our converters use a crystal 4 MHz above the incoming band, preserving the frequency counter accuracy.

6.4 LOCAL OSCILLATOR SYSTEMS

Fig 6.80 shows a number of traditional LO configurations found in receivers and transceivers. Not shown are the common synthesized schemes found in "modern" commercial equipment. Frequency synthesis was discussed in Chapter 4. Many considerations presented here apply to synthesizers as well as simpler systems.

The simplest system is that of Fig 6.80A. A free running LC oscillator operates at the desired output frequency. It is buffered, sometimes with more than one amplifier if higher power is required. Low pass or bandpass filtering is included to remove harmonics. The signal will eventually drive a mixer, with many types requiring LO drive that is free of even-order harmonics. Odd harmonics are

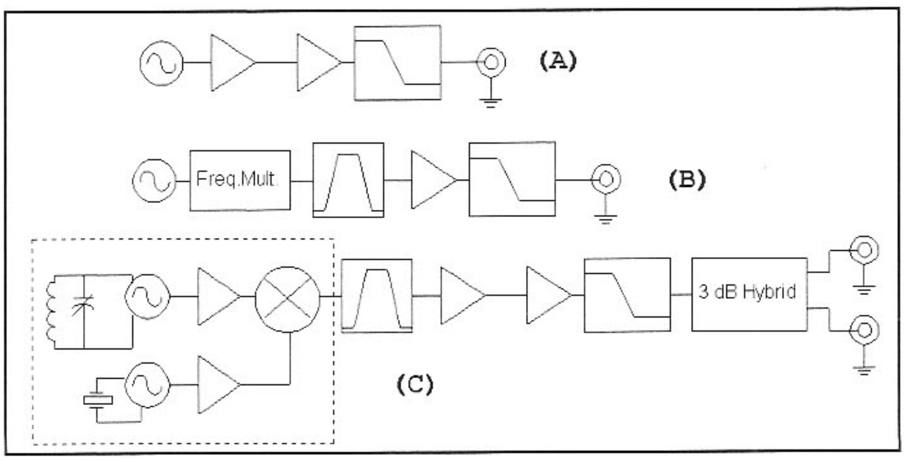


Fig 6.80—Local-oscillator systems for use with communications systems. See text for details.

allowed with the familiar diode rings, for they produce a symmetrical signal, a square wave in the extreme. Even-order harmonics can upset the balance needed for good port-to-port isolation. Details are discussed further in Chapter 8.

Frequency multiplication is often used, Fig 6.80B, for the buffering offered is excellent. In some cases a multiplier is needed to increase the frequency of a fundamental- mode VXO to the VHF region. While crystal-controlled oscillators may be possible at the needed frequency, overtone modes are usually used at VHF, which cannot be pulled with the ease of a fundamental mode oscillator. A bandpass filter follows the frequency multiplier. This is needed to selected the desired harmonic while suppressing all other components. Balanced frequency multipliers are recommended when possible, for they ease the level of filtering and shielding required. The frequency multiplication process is often a lossy one, so more amplifiers may be required. More than one gain stage may be required. Finally, a low pass filter reduces the harmonics generated by the amplifiers.

A frequency multiplier system like that of Fig 6.80B need not alter stability. Any drift in the oscillator will be multiplied with the carrier signal. So a 1-kHz drift in an oscillator that is frequency tripled will produce a 3-kHz shift in the output, leaving the fractional change constant. This drift is still low with multiplied crystal oscillators.

The premix scheme of Fig 6.80C is popular, using a mixer to produce an output resulting from two oscillators. One input is usually from a free running LC circuit while the other is crystal controlled. For example, a 25-MHz transceiver with an IF of 6 MHz might use a 31-MHz LO system. This could be realized with a 4.5-MHz free running VFO and a 26.5-MHz crystal-controlled oscillator. The frequency drift is dominated by the LC circuit, which can be fairly stable owing to the low frequency.

Assume this example system is to tune a 300 kHz range from 30.9 to 31.2 MHz. The VFO will then tune from 4.4 to 4.7 MHz. Before construction begins, or a crystal is ordered, a spur analysis should be performed. This was discussed in the mixer chapter. There are no severe problems with the frequencies used in this example.

Spurious responses, when present, can be reduced with careful attention devoted to LO mixer drive levels. A normal diode ring should be driven with a LO signal of +7 dBm, the 26.5-MHz signal in our example. The "RF" input should be confined to a maximum level of –10 dBm. The "specifications" for the mixer list a much higher level, around 0 dBm. This is the level allowed without damage to the mixer. But spurious responses grow dramatically with drive level. It is important to actually measure levels. An available RF power of –10 dBm should be established with a suitable substitutional measurement with a power meter or 50 Ω terminated oscilloscope, discussed further in Chapter 7.

The example mixer will have –17 dBm outputs at 22 and 31 MHz. A bandpass filter will select the higher. Either a double or triple tuned circuit is suitable. This application requires at least a 300-kHz bandwidth. A wider filter may be preferred, for a 1% bandwidth LC filter is lossy with typical toroid coils. But a 1-MHz bandwidth at a 31-MHz center would be an easy filter to design, build, and tune. A typical filter insertion loss might be 3 dB, resulting in a filter output of –20 dBm. If the eventual system output must be +10 dBm, a net gain of 30 dB is required. This is difficult with one gain stage, but easily realized with two. Feedback amplifiers with general-purpose transistors such as the 2N3904 or MPSH10 are suggested. Again, measurements are required. Avoid input overdrive as a means of obtaining the desired mixer output.

Layout can be critical with the mixer system. The filtered mixer output is low at –20 dBm. Yet there are two very strong signals present: an *RF* input (the VFO) at 4.4 to 4.7 MHz, and a crystal generated *LO* at a robust +7 dBm at 26.5 MHz. Spurious mixer outputs should be at least 50 or 60 dB below the desired level of –20 dBm, or at –80 dBm. The crystal oscillator output reaches +7 dBm. It is reasonable to obtain 50 to 60 dB of suppression between points on a circuit board. But 87-dB suppression presents a greater challenge.

Fig 6.81 shows one way we might build this LO system. The block diagram is in part A while part B shows a typical single board layout. This might be either a breadboard or a printed circuit board, either using a nearly solid metal top foil. When this layout is built and measured, we see the spurious outputs mentioned earlier. The crystal oscillator signal (26.5 MHz) is present in the output, as is a weaker VFO component at 4.5 MHz. But spurious outputs may not just indicate an inadequate bandpass filter. Even when that filter is improved, the spurs may persist, a result of poor layout.

A number of problems are present with this layout. Large RF currents flow in the oscillators, often larger than indicated by the output levels. Those currents flow in the ground plane. If a solid ground plane is used, attenuated oscillator current will be found in the ground foil around and beyond the bandpass filter, now free to feed into the output. The amplifier after the filter has a wide bandwidth and increases the spurious level.

Radiated oscillator signals reach the output coaxial connector. The center wire and the ground connection between the box wall and the circuit board foil form an open loop. That loop is now free to intercept some of the radiated energy. A better connection to the outside world would extend coaxial cable on a bulkhead connection until the board is reached. A twisted-wire pair also works well.

Single-point grounds for each stage are common in audio systems and are appropriate for RF designs. Similar regional grounding can confine oscillator ground current to a small part of the overall board. This would also prevent coupling between the individual oscillators.

The scheme that produces much better performance is shown in part C of Fig 6.81. The board ends with the mixer, situated very close to an output connector. The loop area related to the output connection is kept small. A coaxial environment is maintained through the bandpass filter with the following amplifiers then built on an open board. Examples are shown later in photographs. A 5-element low-pass filter follows, attenuating harmonics created in the amplifiers. The final element in most systems is a splitter-combiner, allowing two 50-Ω loads to be driven. This circuit usually has a 25-Ω input impedance, provided by a modification to a 50-Ω low-pass filter.

Active mixers with lower LO power requirements may be preferred for premixed LO applications. While the NE602 is suitable, higher-level Gilbert Cells like the MC1496 or the Texas Instruments Japan SN16913P are preferred. The later part is soon due to be discontinued with no similar replacement on the horizon. The AD-831 or AD-8343 from Analog Devices should be investigated.

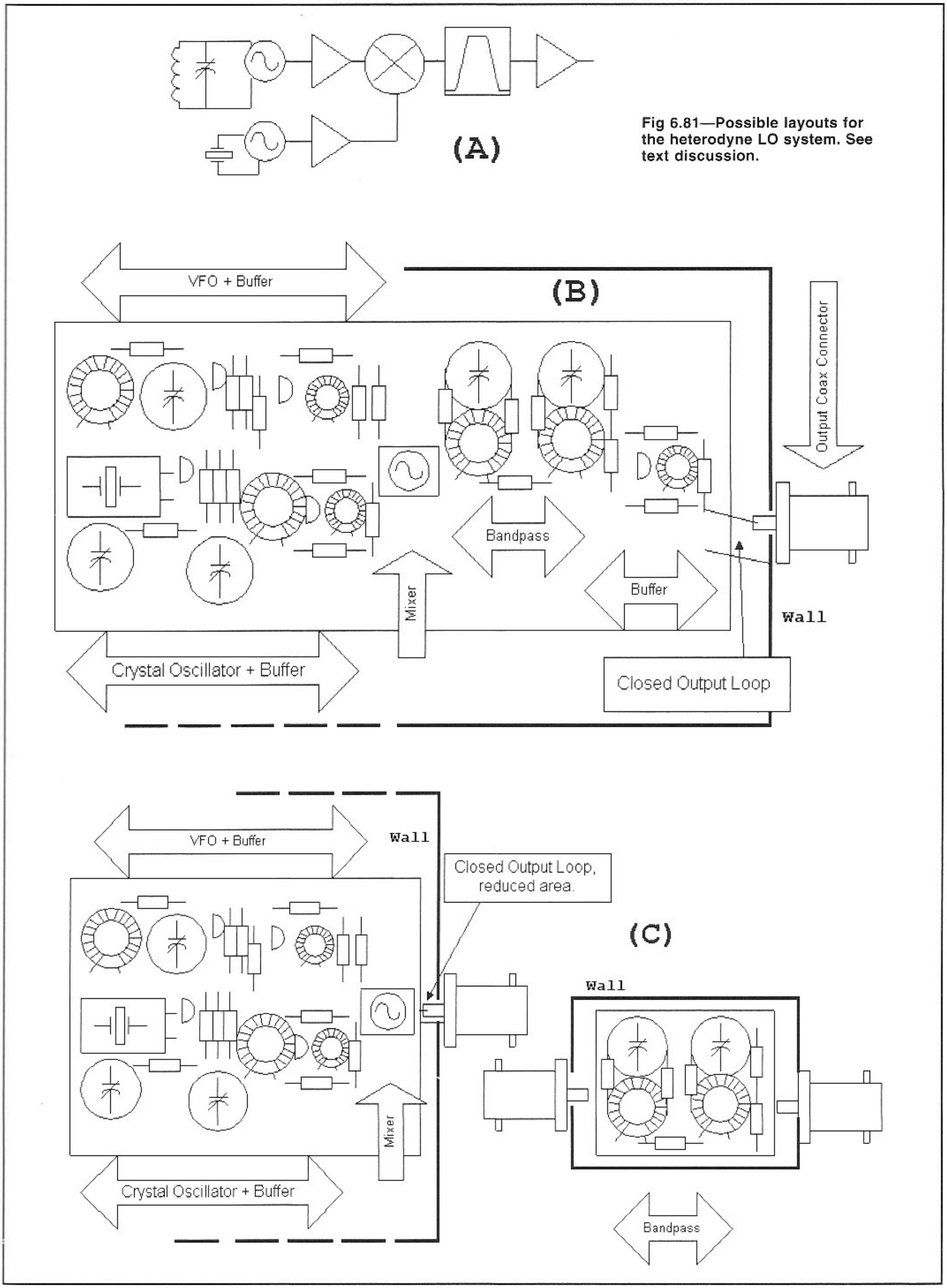


Fig 6.81—Possible layouts for the heterodyne LO system. See text discussion.

6.5 RECEIVERS WITH ENHANCED DYNAMIC RANGE

All of the elements within the front end must be enhanced when striving for high dynamic range (DR). It is usually the mixer (or mixers) that are the critical elements, the parts to be upgraded. However, as soon as we improve a mixer in a typical receiver, the amplifiers become stressed. It is mandatory that we examine all components up to and including the selective filters.

Intermodulation intercept and noise figure are both vital elements in a wide DR receiver. Any NF improvement will allow reduced gain in critical areas, thus relaxing intercept requirements.

A major change in receiver architecture can sometimes make a large difference. We will show a front end later that eliminates *all* gain ahead of the initial selectivity, thus achieving stellar intercept performance while maintaining an adequately low noise figure.

In the last chapter we saw that the input intercept (IIP3) for a +7 dBm LO type diode ring mixer could be +11 to +16 dBm. This is the value that we might measure with a 50-Ω, wideband termination. A high level mixer with +17 dBm LO drive will show IIP3 values 10 dB higher, with typical values in the vicinity +24 dBm.

Fig 6.82 illustrates these design concepts with a front-end block diagram. The first element is a single tuned circuit preselector filter. The wide bandwidth of 1.5 MHz keeps the insertion loss (IL) below 0.5 dB so long as inductor Q_u exceeds 250. Decreasing bandwidth to 350 kHz would cause IL to increase to 1.6 dB, again using inductor with Q_u = 250.

The next element is an RF amplifier. A bipolar feedback amplifier with a pad is used here, shown in **Fig 6.83**, which includes the input preselector schematic.

The next system element, **Fig 6.84**, is the main preselector filter, the one that establishes image rejection and protects the mixer from spurious responses. The circuit begins with a 9-element low-pass filter, followed by a 3-resonator bandpass with a bandwidth of 300 kHz. The receiver using this filter was a committed CW design that tuned only the bottom 150 kHz of the band, so the narrow preselector was not a limitation. The circuit ends in a 3-dB pad that establishes filter termination and helps preserve mixer performance. The low-pass filter guarantees stellar suppression of VHF signals, a problem in a metropolitan environment.

One might argue that this preselector is more extensive than needed. Our goal was to realize a "100 dB" receiver. That meant not only that the two-tone dynamic range should exceed 100 dB, but that all spurious responses should be suppressed by the same amount. One such spur occurred with 16-MHz input signals that reached the IF via a 5:1 spur. The image and spur rejection plus the 9-MHz IF feedthrough rejection could only be guaranteed with an extensive preselector. Such filters have high insertion loss, 6.5 dB here when the pad is included. It is this high loss that made the RF amplifier necessary.

Two preselector networks are required whenever an RF amplifier is used. Some initial selectivity protects the system from out of band energy. A single network at the input is generally insufficient, for it would allow image noise generated in the RF amplifier to be converted to the mixer IF.

The next element is the mixer, an SRA-1H using +17-dBm LO injection. The mixer is driven from a 5-MHz LO system. A design with fewer spurious responses would move the LO to 23 MHz. A heterodyne approach shown earlier (Fig 6.81A)

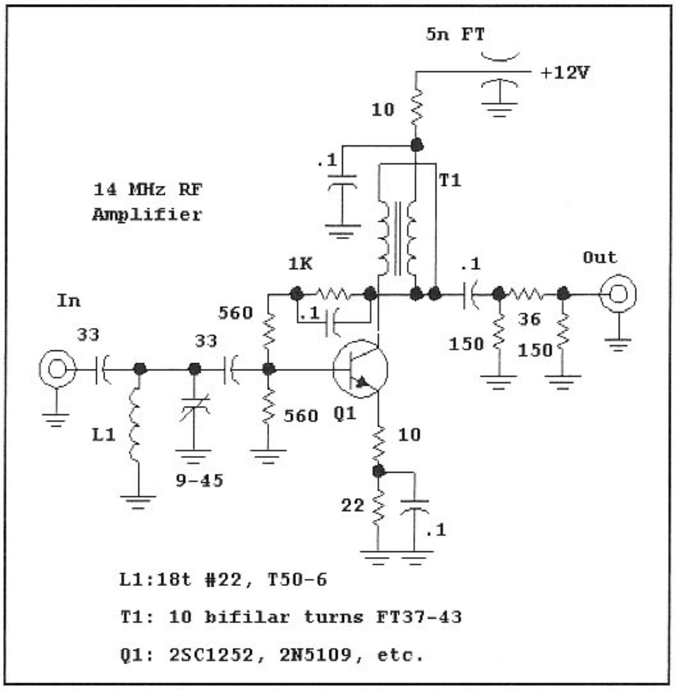


Fig 6.83—RF amplifier with preselector network. This amplifier used parallel feedback from the output tap. Feedback directly from the collector is preferred.

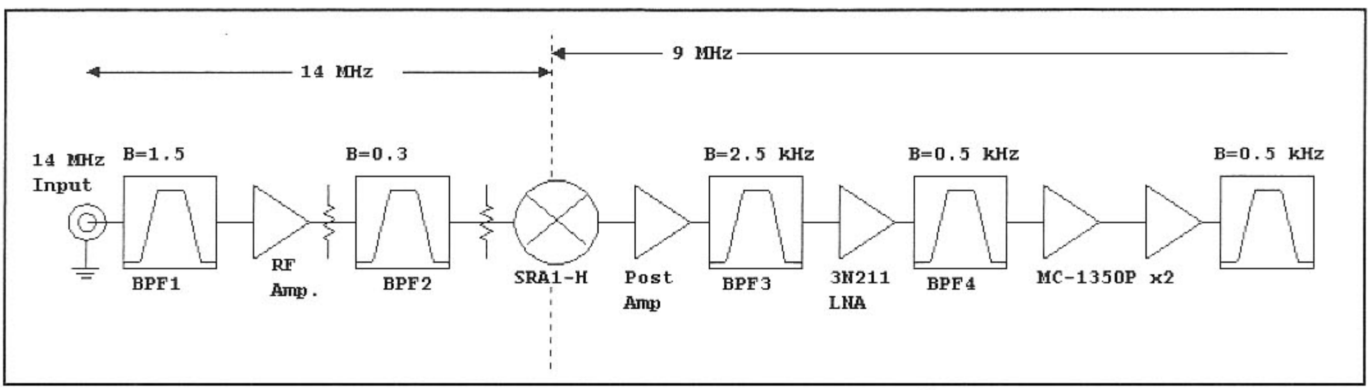


Fig 6.82—Block diagram of an early high-dynamic-range receiver. The various elements are shown in schematics. See text for stage-by-stage discussion.

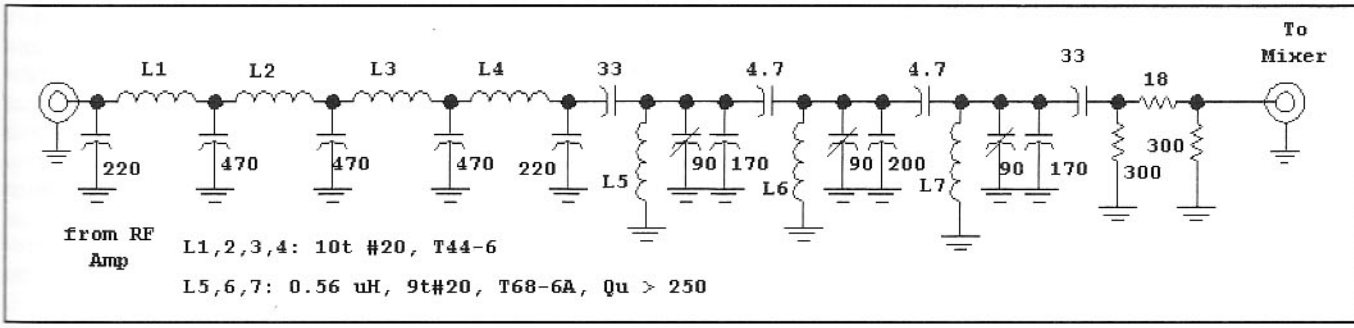


Fig 6.84—Image-stripping preselector filter used with the receiver. This filter provides over 100-dB suppression of images and other spurious responses.

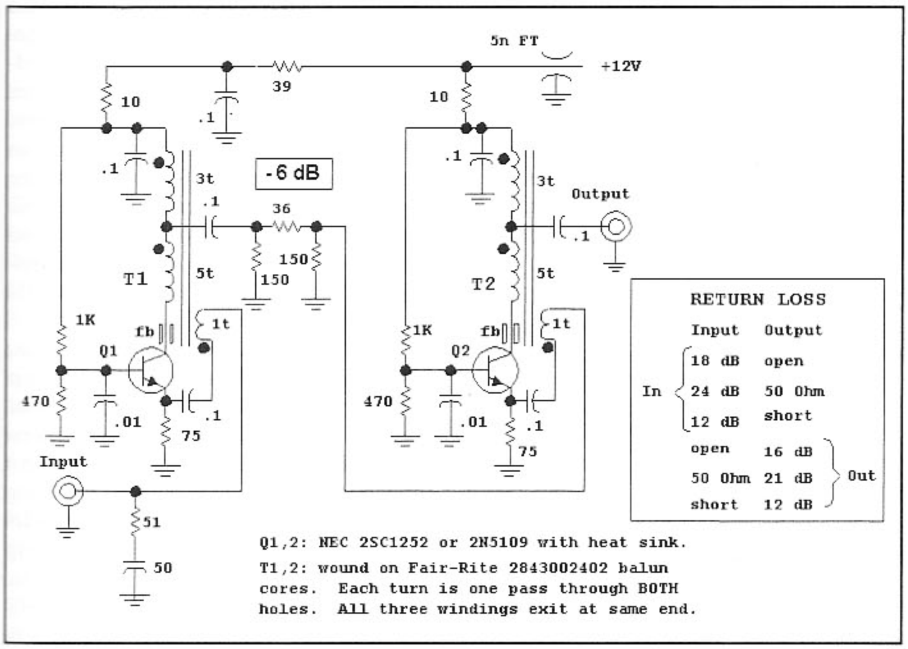
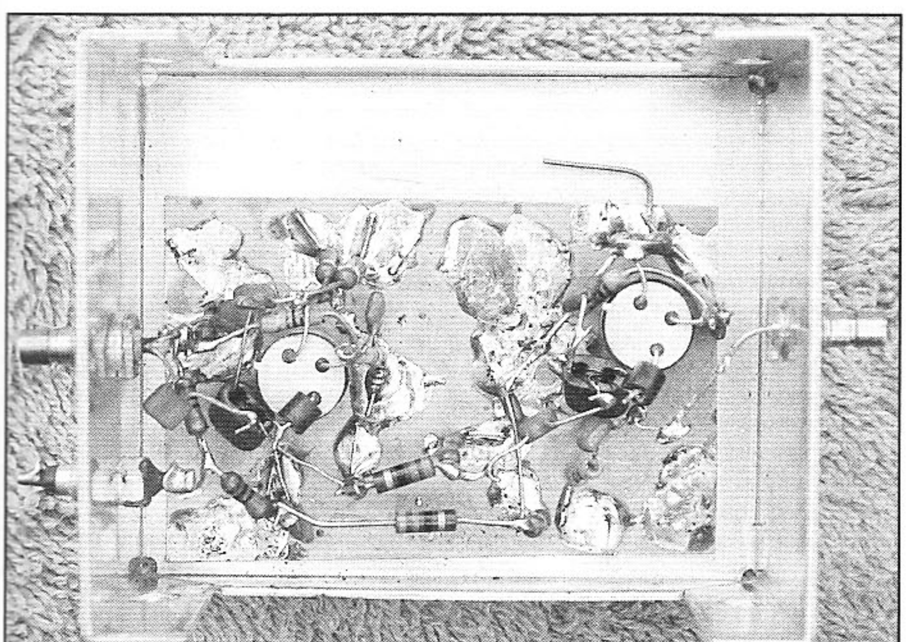


Fig 6.85—Two-stage Norton amplifier used in the CW receiver.



Two-stage Norton Amplifier.

is suitable. This would allow a wider preselector bandwidth with reduced loss, allowing less gain to be used in the RF amplifier, extending dynamic range.

The next front-end element is a post mixer amplifier, shown in **Fig 6.85**. This circuit uses the transformer feedback Norton amplifier topology presented in Chapter 2. That circuit has good noise figure and low IMD, but poor port-to-port isolation. Moreover, the terminal impedances are strongly dependent on the load at the opposite ports. This means that the strongly varying crystal filter input impedances would appear at the mixer output, degrading IMD performance. Placing a pad between two Norton amplifier stages solved the problem here. Overall amplifier gain was 11.5 dB with OIP3 = +42 dBm and NF = 5.7 dB. The individual stages had a 4.1-dB NF. The figure includes measured return loss for the input when terminated in a variety of outputs, and similar results for the output.

Overall front-end gain is low in this receiver. The main crystal filter that this design used was a 10-element circuit with 500-Hz bandwidth, which had a 10-dB insertion loss. The high IL was an acceptable price for the spectacular performance. But receiver NF would be compromised if the IF was driven from the low gain front end. So, a "roofing filter" was used to follow the front end. This lower loss filter with a 2.5-kHz bandwidth was followed by a fairly low noise amplifier that then drove the narrow CW filter. This topology compromises dynamic range for very close tone spacing, but is an otherwise useful technique.

Evaluation of this receiver produced an 8-dB noise figure (MDS = –139 dBm) with IIP3 = +13 dBm for dynamic range = 102 dB and Receiver factor R = +5 dBm. The receiver served as a self-test vehicle during development. The IF system was built and used with an earlier receiver. It then provided the narrow bandwidth needed for IMD measurements. This allowed direct evaluation of mixers, amplifiers, and filters. A key to the development was the ability to insert attenuators between stages. This then allows the designer/builder to pinpoint the distortion source.

Some interesting details emerged from this investigation. Our first attempts to use the 2.5-kHz roofing filter were frustrated by IMD in the filter, confirmed with the insertion of pads in the system. A new filter from a different manufacture eliminated this difficulty, leaving the mixer as the critical element. The mixer was not *well behaved*, showing better IIP3 when operated at higher levels than it did when IMD products were close to the receiver MDS. Lower level data is quoted.

The receiver was built with the front end segmented into several modules, each in a shielded box and interconnected with coaxial cable. The shielding continues through the IF, BFO, and Product Detector. Power is supplied to the modules via feedthrough capacitors. The 50-Ω interface allows easy measurement of individual modules and quick changes in gain distribution. It also prevents the sorts of interactions and instabilities that can (and usually do) arise when such systems are built in the open. Finally, it provides shielding against radiated and conducted energy from digital circuitry that might be used in other parts of the receiver. Shielding "by the stage" is generally much more important and useful than shielding afforded by one metal box around equipment. This is an old design and duplication is not encouraged.

Fast Forward—Modern Receivers

A more up-to-date front end is shown in **Fig 6.86**, where the incoming signal is converted to a VHF first IF. The design shown is not an example we have built, but one that should be possible with existing technology. It has features not found in earlier designs, but also introduces problems. Up-conversion is typical with most modern gear.

The first IF in this example is 70 MHz with the LO running above the IF. These up-converted designs are usually general coverage receivers, tuning from 50 kHz to 30 MHz. The example receiver uses a 70 to 100-MHz LO injection, generated by frequency synthesis. The input low-pass filter has a cutoff at 30 MHz and establishes image rejection. The image for this example is at the sum of the LO and the IF, 140 to 170 MHz. Images are no longer an issue so long as the low pass filter works as designed.

A bandpass preselector filter is still used in the front end of Fig 6.86. If none were used, the receiver would be subject to overload by signals far removed from the input. On the other hand, it is now practical to keep the preselector bandwidth wide enough that IL is low, which helps to maintain a low noise figure. Common practice uses half octave filters with two bandpass filters for each frequency doubling. This is often approximated with filters of around 5-MHz bandwidth. Narrower filter bandwidth could be useful.

Gains, noise figures, and intercepts are given with critical stages in Fig 6.86. The passive high-level (+17-dBm LO) mixer resembles that of the last receiver with 6-dB NF and conversion loss with an input intercept of +25 dBm. The post mixer amplifier has 12 dB gain, a low noise figure of 2 dB, and IIP3 of +25 dBm (OIP3 = +37 dBm.) Note that this amplifier is actually weaker (lower intercepts) than the post-amp used in the earlier receiver. This is practical, for signals are smaller, a result of using no RF amplifier. (Also, the post-amp in the previous receiver was stronger than necessary!)

Some design rules emerges from these studies: If the output intercept of one stage equals the input intercept of the following stage, each will contribute equally. If one of the two stages is to be dominant, it should have an intercept at the common plane that is 6 dB above the other. Note that these are not "rules-of-thumb," but results of analysis.

Data is included in the figure for crystal filter IL and IF noise figure. The result for this receiver is an overall noise figure of 10.5 dB with IIP3 of +26 dBm and R = +15.5 dBm. In a 500 Hz bandwidth this would generate a two tone DR of 108 dB. The mixer is the critical, performance-determining element defining system IMD.

Although the numbers appear good in this design, there are a couple of details that can severely degrade them. The first is the 70-MHz crystal filter. This element has a bandwidth of 20 kHz, easily realized with today's technology. But with such a wide bandwidth, a tone separation of 50 to 100 kHz would be required to achieve the calculated intercept. The same measurements done at 10 or 20-kHz separation would produce lower IIP3 values.

A second major problem relates to the bandpass filters used in the design. They are typically switched with PIN diodes at the filter *input* and output. Diodes at the input are not protected by the bandpass filters and are then subjected to a wide frequency spectrum. Both second and third-order intermodulation distortion can then generate products that severely compromise performance. Performance can sometimes be improved by increasing the bias current for conducting diodes. The better solution is substitution of improved diodes. The HP-8052-3081 is recommended.[16] The Siemens BAR17 or M1204 are also recommended.[17]

A vital diode parameter is *carrier lifetime*, which should be greater than 2 ms in this application. (Carrier lifetime is a measure of the life of carriers within the diode when reverse biased after a period of conduction.) Some high voltage rectifiers display long enough lifetimes, but tend to be lossy. PIN diodes built specifically for RF switching display lower loss, but only some have the long lifetimes needed for switching at HF and especially MF. The popular MPN3404 and similar devices used in this text are not generally suitable for high DR applications. Diodes need to be measured and characterized for RF performance so they can be included in a system analysis.

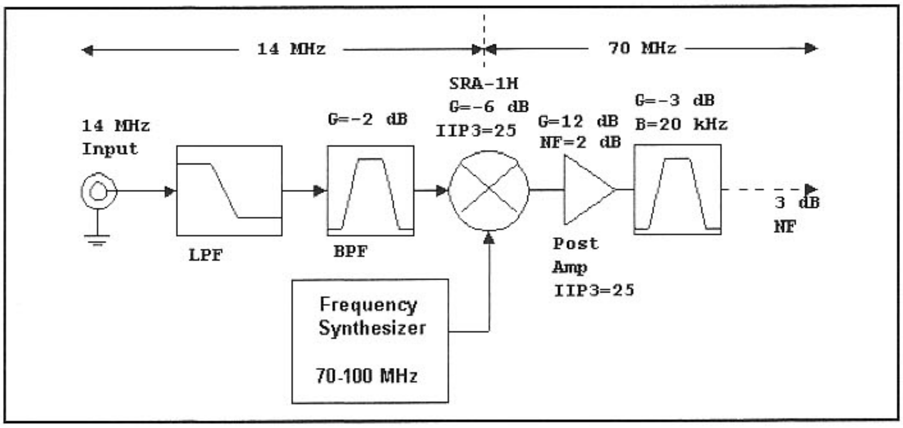


Fig 6.86—Front end typical of modern equipment, although this example is designed for performance beyond the norm. The bandpass filter, shown for 14 MHz center frequency, will have a bandwidth of several MHz and will be switched with relays or PIN diodes. See text.

Another flaw with the up conversion block diagram arises with the VHF crystal filter. IMD in these filters is often worse than seen with lower frequency filters. It should be characterized and considered in system analysis. The filter should have enough selectivity to allow the VHF IF signal to be converted down to a lower frequency IF where additional processing occurs. The conversion should be relatively spur and image free. It is common in current designs to amplify and heterodyne the signal to a low enough frequency that it can be applied to an analog to digital converter (ADC), producing a digital data stream suitable for digital signal processing (DSP.)

Additional distortion sources are found in the low-pass and, more often, in the bandpass filters ahead of the mixer. Filter intercepts depend primarily on the magnetic properties of the inductors used in the filters. They will also depend on the peak energy stored in the component during operation. Running 1 mW of power through a low-pass filter usually results in relatively low current flowing in the inductors used in that filter, so small cores are suitable. But the same 0 dBm applied to a narrow bandpass filter may produce much higher inductor current, producing intermodulation distortion. For example, we have observed in-band IIP3 of approximately +30 dBm for a three-resonator 10-MHz filter with 300-kHz bandwidth. Changing from T37-6 to larger T50-6 cores increased IIP3 to about +50 dBm. We have also observed severe IMD with inexpensive slug tuned coils. As with all things related to high DR equipment, meticulous measurements should replace lore.

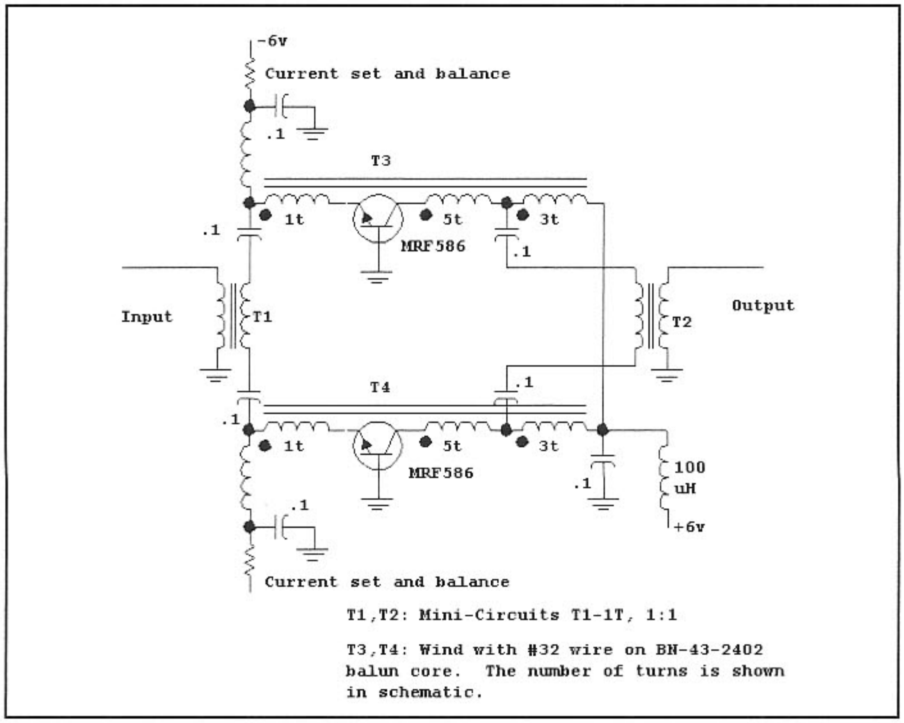


Fig 6.87—High-performance post-mixer amplifier. The transistors were biased to 40 mA each for OIP3 = +48 dBm. Dual power supplies are used for amplifier bias. This amplifier represents very good performance that we have not duplicated.

Moving toward higher Dynamic Range

The front ends described can be extended to provide even better performance by substitution of improved circuit elements. Primarily, the high-level mixer can be improved. Higher-level diode rings are available, some using up to 1/2 W (+27 dBm) LO power. With another 10 dB of LO comes a similar increase in IIP3.

Perhaps the more appealing mixers are those using FETs. They are capable of very high intercepts, have IL similar to the highest-level diode mixers, but require little LO power. This does not imply though that LO drive can be treated with casual abandon. Passive FET mixers usually have LO signals applied to the gates. They must be driven hard to ensure fast switching; symmetry is critical to preserve balance. The FET ring popularized by Oxner is capable of IIP3 up to +40 dBm or a bit higher with conversion loss around 8 dB. Makhinson, N6NWP, reported 7 dB loss with square wave LO drive.[18] The mixer of greater interest is the H-mode mixer generated by Horrabin, G3SBI.[19] IIP3 of +55 dBm was reported with a conversion loss from 8 to 9 dB when using the same FETs as applied with the Oxner mixer. A simplified version will be described later featuring IIP3 > +40 dBm with loss at 5 dB.

One of these high intercept mixers may well have OIP3 of +35 dBm or higher. To be dominant, post mixer amplifiers should have IIP3 of +40 dBm or higher. This moves the output intercepts into the +48 to +52 dBm range. Such amplifiers are possible with very high currents, or with modest currents and careful design. **Fig 6.87** shows the amplifier used by Makhinson in his receiver. Two Norton-type transformer-feedback amplifiers are used in push-pull to achieve a gain of 8 dB with OIP3 of +48 dBm and NF = 2.5 dB.

Colin Horrabin built a version of this amplifier with improved performance. He shifted to a single ended power supply, but increased current to 60 mA per transistor. He changed transistor type to the MRF-580A and added ferrite beads to the collectors for stability considerations. Transformers were hand wound on balun cores and the transistors were heat sunk to a copper substrate. He obtained the spectacular results of OIP3 = +56 dBm with Gain = 8.8 dB and noise figure under 1 dB! The amplifier was at the limits of his NF measurement capability, and IMD determination was also stressed. He also reported that transformers had to be selected for lowest IMD. Nothing is casual at this performance level.[20]

In a later variation of his earlier amplifiers, Makhinson used a push-pull pair of Norton feedback amplifiers that drove a differential pair of common base amplifiers. The second stage common-base circuit provided good reverse isolation while the input transformer feedback design afforded low noise. The lower second-stage reverse isolation generated an input impedance independent of output termination for the two-stage design.[21]

Another approach to balanced amplifier design is that of Engelbrecht, shown in **Fig 6.88** (see Chapter 3.)[22,23] The incoming signal is split in a 3-dB quadrature coupler. The two hybrid outputs are then 90° out of phase with each other as they are applied to the amplifier inputs. If the impedance match at amplifier #1 is less than perfect, there will be a power reflection. The action at the input to amplifier #2 will be identical, for the amplifiers are identical. Each reflected component undergoes another 90° reflection as it progressed back to the input. The two reflected components are 180° out of phase with each

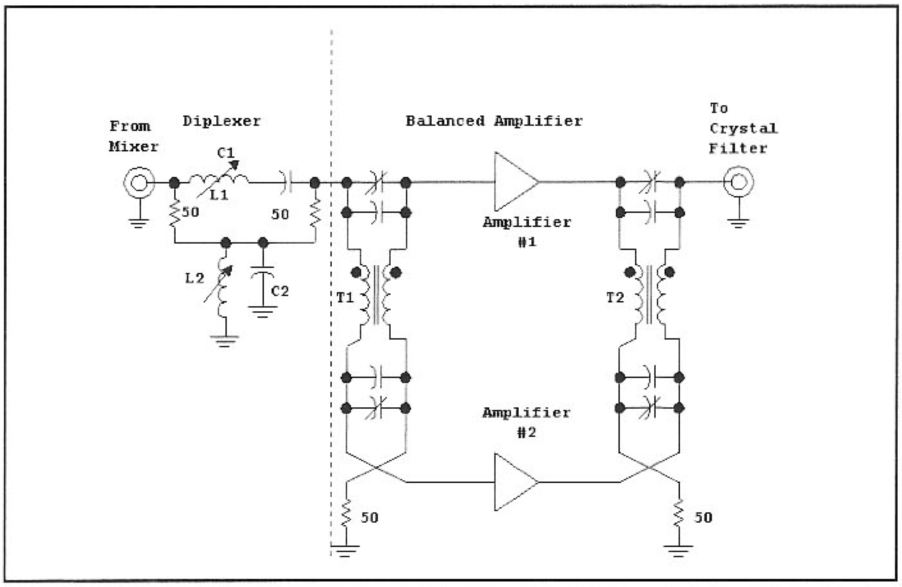


Fig 6.88—Balanced amplifier method of Engelbrecht. See text for discussion.

other by the time they reach the input, so the input impedance is always 50 Ω.

The coupler of Fig 6.88 generates a 90° phase shift at all frequencies, but equal output amplitudes at only one crossover point. A bandpass/bandstop diplexer provides a termination at all frequencies far from the design center. Depending on the nature of the crystal filter, a diplexer may be useful at the output port as well.

Front Ends Without Early Amplifiers—The Triad Receiver

The up-conversion system of Fig 6.86 is a child of compromise, illustrating the tradeoffs often taken to achieve general coverage. The ability to tune the entire HF spectrum was once considered a performance virtue. It is now, since the advent of WARC bands, merely an economic ploy. The aggressive designer/builder need not adhere to such guidelines. He or she can configure a system that will offer high performance on a few selected bands. The IF can be at HF where crystal filters can be narrow without severe loss and with low IMD. Preselector filters with only modest loss can be used with the best available mixers.

The problems with post mixer amplifiers remain. The ideal solution is to merely eliminate them. This can be done with a switching-mode mixer if a crystal filter with constant, frequency flat input impedance can be applied. Such a block diagram is shown in **Fig 6.89**. The circuit is the result of several years of collaborative effort on the part of Bill Carver, W7AAZ, Harold Johnson, W4ZCB, and Colin Horrabin, G3SBI—collectively referred to here as the *Triad*.[24]

The Mixer

The key element in this receiver is the H-mode mixer shown in **Fig 6.90**. The basic mixer was presented in Chapter 5. This example uses a readily available and inexpensive quad-MOSFET-Bus Switch, the Fairchild FST3125M. The device is also available from other vendors. (This part was suggested to the Triad by Giancarlo Moda, I7SWX.) The H-mode mixer is one with RF applied to a transformer, T1, which generates a balanced source of RF. The two resulting signals are then applied to the center taps of transformers T2 and T3. Four FETs connect windings to ground in pairs. Two IF outputs are generated on the secondary windings of T2 and T3.

The FST3125M uses a 5-V bias, required by the quad logic inverters included in the IC. The FETs and related transformers are biased at half this supply with a resistive divider. Symmetry is emphasized in the construction method shown in the photographs. A sandwich of two circuit boards contains the mixer, diplexer and following crystal filter described below. The mixer chip is on the lower board while the diplexer and filter are on the upper one. The three transformers actually reside between the two boards, serving as the routes from one to the other and back.

The digital portion of the mixer circuitry dealing with the LO is shown in **Fig 6.91**. A signal of +10 dBm is applied to the mixer board at *twice* the desired LO frequency. It is converted to a digital form with two NAND gates (74AC00) and is then routed to a divide-by-two circuit using a 74AC109 J-K flip-flop. The flip-flop contains an inhibit input, which is driven by the remaining NAND gates, providing a convenient means for turning the LO off. This may be used during receiver mute periods or as a noise blanking input. This method of blanking is especially effective, for it is out of the main signal path and has few of the distorting effects usually related to the blanking function, other than that intrinsic to modulation.

An alternative logic section is presented in **Fig 6.92**. This scheme uses a VHF local oscillator that is then divided by any even integer from 4 to 18. This method is used in the W7AAZ version of the receiver.

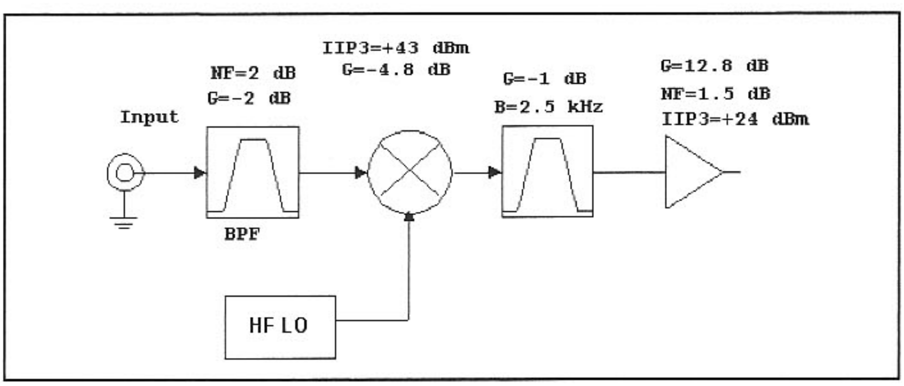


Fig 6.89—Receiver front end using no amplifiers before initial selectivity is obtained. This is the basis of the W7AAZ/W4ZCB/G3SBI receiver described below.

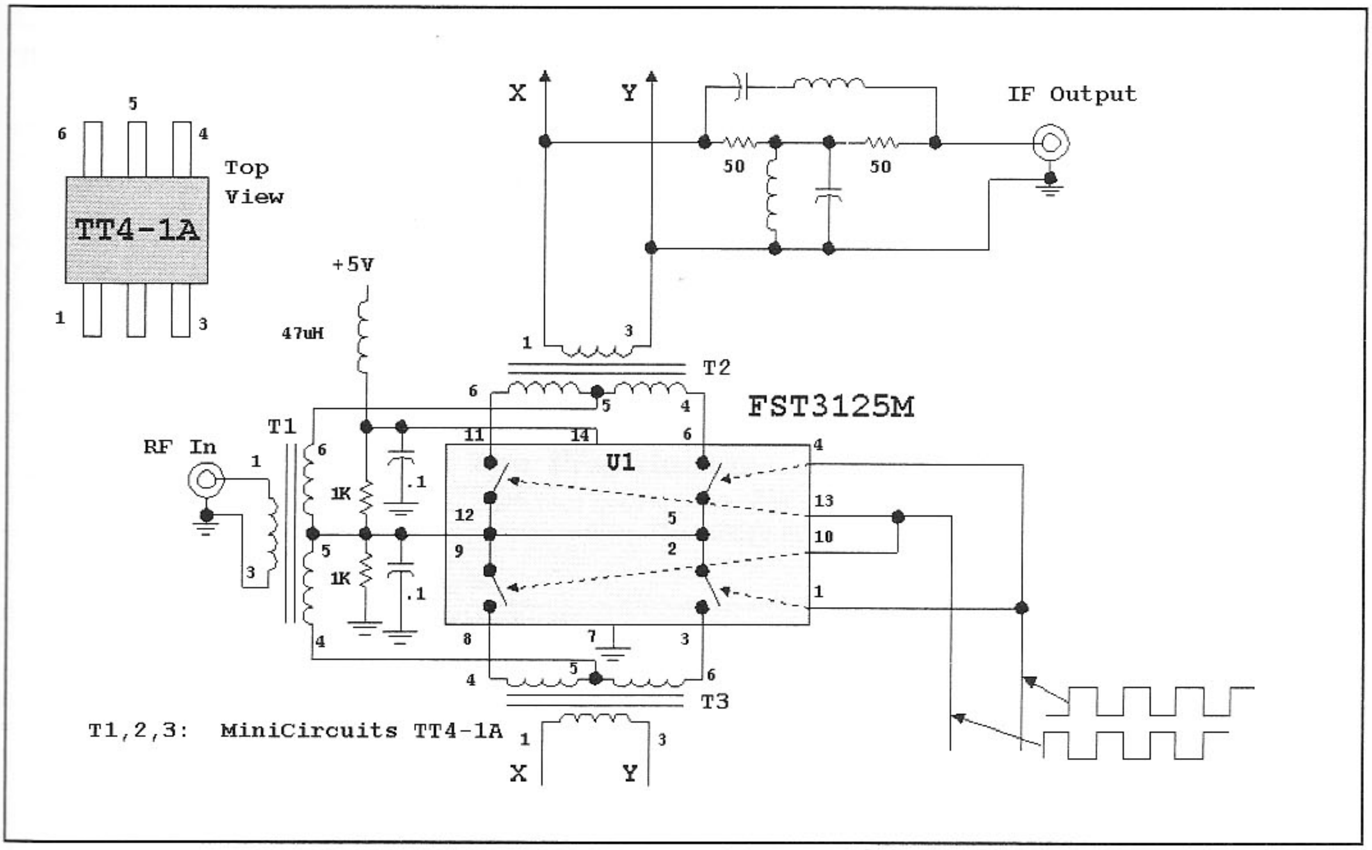


Fig 6.90—Mixer portion of the high-level front end. Commercially available transformers are used in this design. U1 consists of four MOSFET switches controlled by lines 1, 4, 10 and 13, linked with the dotted lines in the figure. See Chapter 3 for design of the Q = 1 diplexer at the IF port for compatibility with the chosen IF.

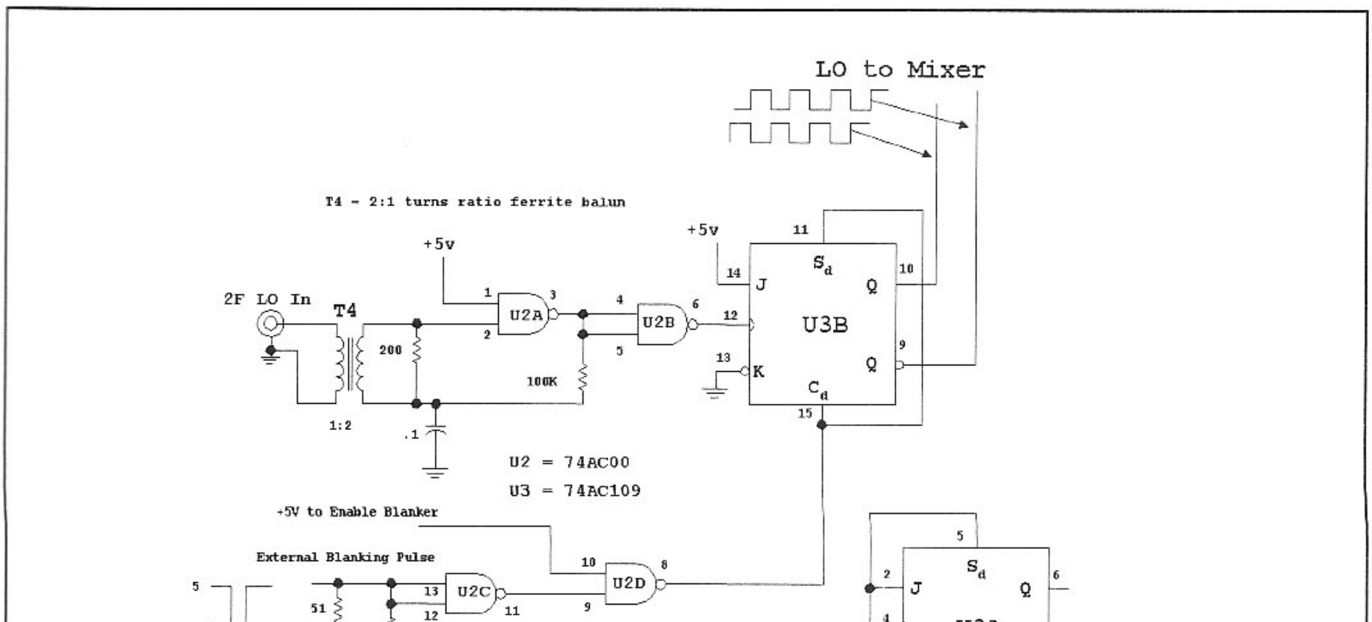


Fig 6.91—Logic circuits provide high-frequency LO drive for the H-mode mixer. Input is at twice the needed LO frequency. The designer/builder must add power supply connections to the ICs.

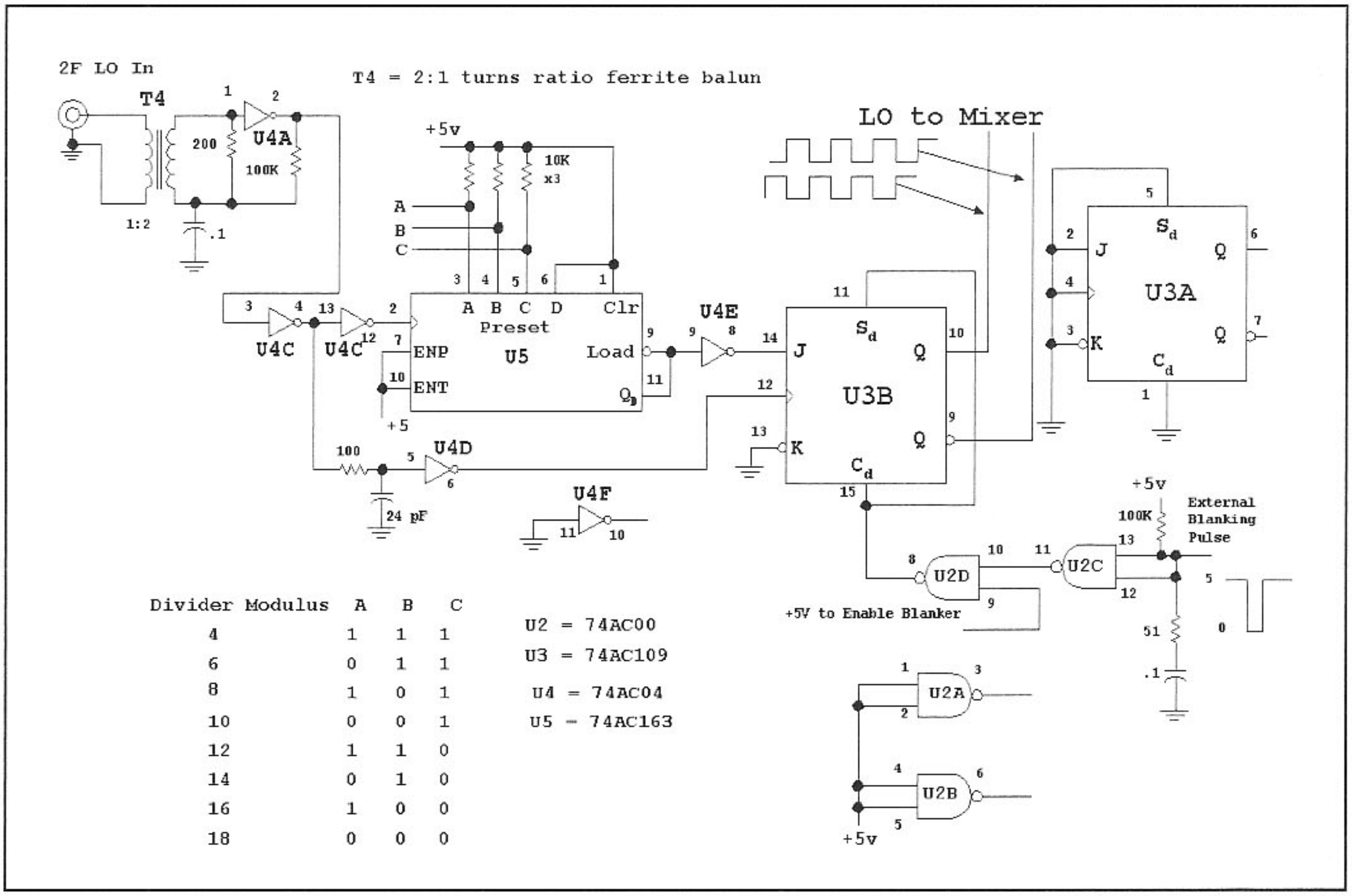


Fig 6.92—Logic circuits accept an input from a VHF synthesizer. The output is then divided by an even number between 4 and 18 before reaching the high-level mixer. The designer/builder must add power supply connections to the ICs.

The Roofing Crystal Filter

A poor mixer termination will severely degrade IIP3. A filter with a 50-Ω input impedance at all frequencies, inside and outside the passband, is shown in **Fig 6.93**.

The crystal filter is a critical element in the overall front end and requires careful design and adjustment by the designer/builder. The crystal frequencies are picked to produce a passband that overlaps that of the dominant filter in the receiver IF system, measured before this filter is built. The crystals will then be ordered from a reliable supplier. High crystal Q should be sought, for it will directly impact filter IL. The builders saw their best filters with loss under 1 dB with others under 2 dB. Even if the receiver is to be used mainly on CW, a wider design filter bandwidth is used in the interest of low loss.

Careful measurements are required to adjust this filter. A spectrum analyzer with a tracking generator is ideal, but should have stability commensurate with narrow crystal filters. Sweeps measuring input and output impedance match should, however, extended from near dc to VHF.

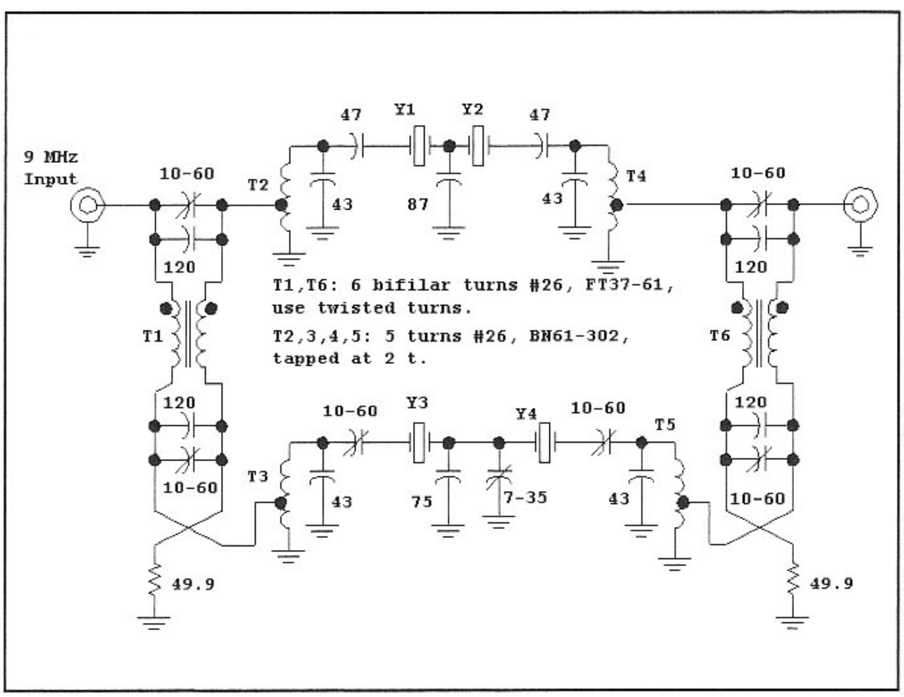


Fig 6.93—Crystal filter serving a "roofing" function. This circuit operates at 9 MHz, but can be redesigned for other frequencies within the HF spectrum. The variable capacitors with Y3 and Y4 are adjusted to match the one filter to the one using Y1 and Y2. The quadrature hybrids are adjusted for optimum impedance match at both ports. See text.

An Amplifier to follow the Roofing Filter

Fig 6.94 shows the amplifier that follows the mixer. This circuit must have reasonable performance, although not as stellar as would be needed without the filter. With only two crystals per side, the roofing crystal filter has limited skirt selectivity, allowing some large signals to appear beyond the filter.

The amplifier is a feedback circuit with four parallel JFETs. The total current is high at 85 to 100 mA, so the circuit has good distortion performance. The circuit began conceptually as a transformer matched common-gate amplifier; a topology with a well-defined, low input impedance.[25] A winding is added to the transformer to apply some signal to the gate. The result is a circuit that has neither terminal as common, yet has a well-defined 50-Ω input impedance while featuring low noise figure. This circuit has a typical NF of 1.5 dB with some versions measuring 1.2 dB. The output is transformer coupled with a drain load resistor to ensure a good output match.

Bill Carver, W7AAZ, modified the bifilar output auto-transformer with another winding that drives an adjustable capacitor, C-N, to couple energy back to the gate. This capacitor is adjusted for low reverse coupling. The result is a neutralized amplifier featuring low noise, high IIP3, excellent input and output impedance match, and good reverse isolation.

This circuit can be adjusted for an input return loss greater than 30 dB in the 3 to 30-MHz region. Typical gain is 12.8 dB with IIP3 = +24 dBm. A heat sink is built for the four FETs by drilling four holes in a piece of 1/8-inch-thick aluminum. The FETs are pushed into the holes, which are then filled with epoxy. Carver has also built similar amplifiers with six FETs, but the same 100-mA total current. These circuits require no heatsink.

The Preselector

The final element in the front end is the preselector filter. The basic form is shown in **Fig 6.95**, a top coupled set of parallel resonators. Reed relays are used at each end for band switching. Extensive decoupling (not shown) is used with the relays. The filters were designed to have a maximum insertion loss of 2 dB. A 5-resonator filter was used for 160 m while 3 or 4 were sufficient for the other bands. Toroids were used for all inductors with emphasis on larger sizes for high unloaded Q and low IMD. A 6 mix was used for the lower bands with 10 for the upper ones. Most capacitors were 1% silver mica types. The only variable capacitors were some trimmers used for coupling on the highest bands. Components were carefully measured prior to installation and inductor turns were spread or compressed slightly for fine-tuning. This was sufficient for the

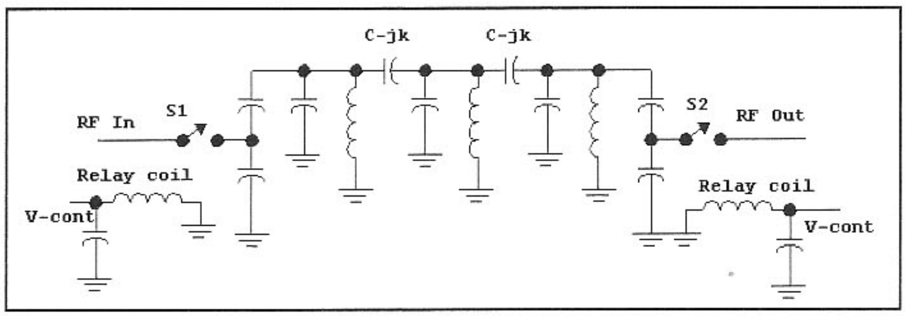


Fig 6.95—General form of preselector filters used for the high-performance receiver. While a 3-element filter is shown, some bands used up to 5 resonators.

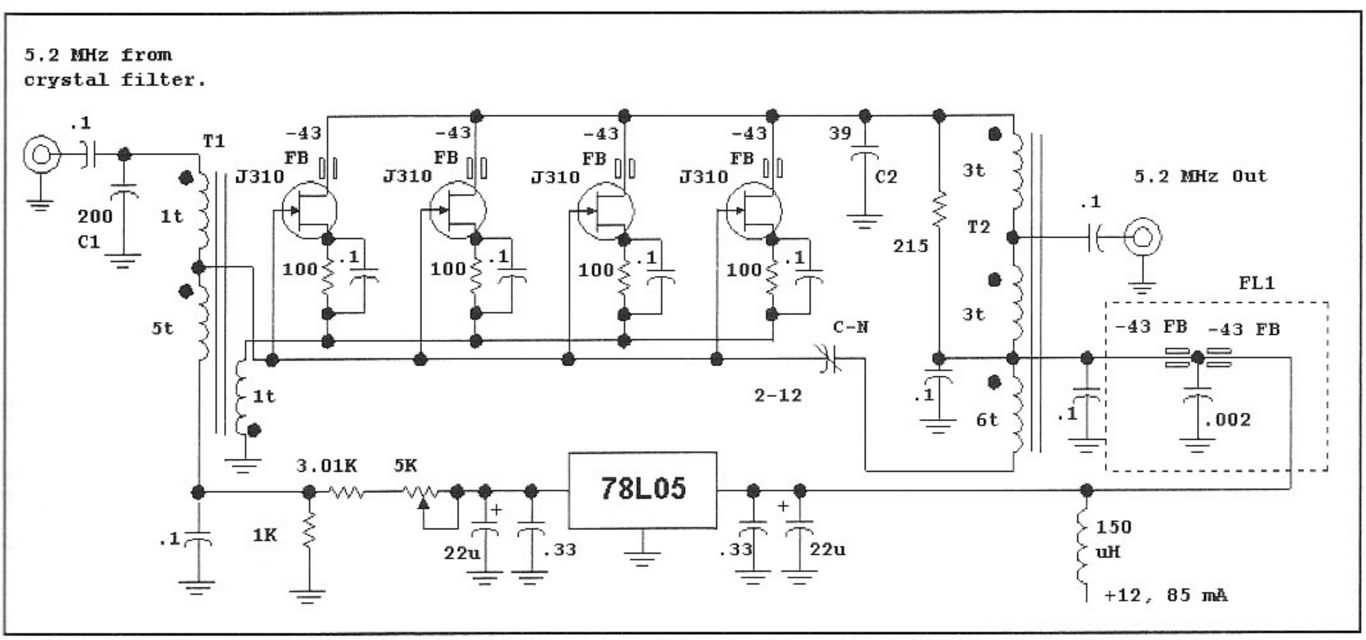
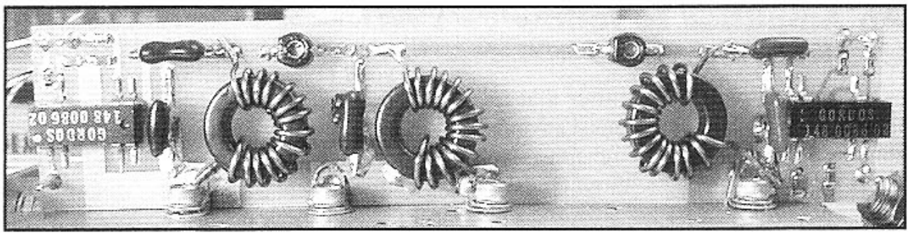


Fig 6.94—Amplifier that follows the roofing crystal filter. This particular version operates at 5.2 MHz, but can be optimized for any frequency in the HF spectrum. T1 is wound on a BN61-202 two-hole balun (binocular) core. The primary (grounded winding) is made from small copper or brass tubing through the balun holes. Alternatively, braid from RG174 coaxial cable may be used. The 5-turn and 1-turn windings are then wound with #28 or smaller wire. T2 consists of a pair of bifilar windings on a BN43-202 two-hole balun core. One bifilar winding forms the two 3-turn windings while the other bifilar pair is connected to form the 6-turn winding. Remember that one turn on a two-hole balun core is a pass through both holes. C1 and C2 are approximately resonant with transformers T1 and T2. FL-1 is a three wire monolithic element, but can be built with discrete components. C-N is adjusted for best reverse isolation (lowest S12.) All resistors are 1% metal film, 1/4 W.

lower bands while the Dishal method was applied for the upper frequencies.[26,27]

The design goal for the preselector filters was a stopband attenuation of 90 dB or more. This was realized, but it required considerably more effort that anticipated. The filters were all built on boards with components in a long narrow line for best input to output isolation. The stopband performance was only realized after the on-board grounds were isolated. Each resonator was grounded directly to the large metal plate that supported the boards. It was also important to carefully place the various filters in the stack. A situation to avoid was an adjacent filter that operated at an image. For example, if the receiver used a 5-MHz IF with LO at 9 MHz, the 4-MHz image is 14, so the 80 and 20-m filters should not be next to each other. Details of construction are shown in the photograph. This is yet another place where detailed measurements are required.



21 MHz bandpass filter used in W7AAZ version of the Triad Receiver. No variable tuning capacitors are used. The trimmers adjust coupling.

An Oscillator

A voltage-controlled oscillator developed by Harold Johnson, W4ZCB, is presented in **Fig 6.96.** It has been applied in a number of ways including acting as the controlled oscillator in experimental synthesizers and one-on-one phase-lock loops. The circuit operates in the 80 to 110 MHz region and is then divided from VHF in the circuit shown earlier in Fig 6.92.

The heart of the VCO is a helical resonator. This element offers an unloaded Q of 700, performance difficult to obtain at HF. A metal lathe is needed for the construction. The resonator is housed in a section of 1.5-inch-diameter copper tubing with copper-pipe ends. The helix consists of 9 turns of #12 wire wound on a 0.75-inch diameter tubular form that was machined from RF grade polystyrene rod. After the rod was machined to 0.8 inch outside diameter, threads were cut at an 8-turn-per-inch pitch. The inside of the rod was then removed with a large drill bit, leaving a wall thickness of approximately 1/8 inch. Material was retained at one end for mounting. The #12 wire was wound and approximately spaced before being threaded onto the form.

The helix has two taps. Output is extracted from one 1/4 turn up from ground while the drain is attached at 1/2 turn from ground. The outputs are buffered with a quad buffer. One output drives the mixer while the other is for synthesizer use. Detailed information regarding tap placement and resonator construction is given in a note from W4ZCB included on the CD that accompanies this book.

Two different methods were used for phase noise measurement. In one, the VCO under test was phase locked to an HP-8640B signal generator. The baseband output was filtered, amplified, and analyzed with an HP-312 selective voltmeter. The other system uses the HP8640 as a local oscillator with a high level mixer. The output is applied to a narrow crystal filter. The signal is amplified and further filtered, and is then detected. The two systems offer good agreement.

This oscillator, after division by 8, provided phase noise of –155 dBc/Hz at a 20 kHz spacing. The noise dropped to –163 dBc/Hz at 50 to 75 kHz; at 100 kHz it was beyond the range of the measurement equipment. A one-on-one PLL will provide some close in clean-up. Thermal stability was good enough to allow direct use without any stabilization, although this is not common and should not be expected with similar designs.

The Overall Triad Receiver

We have described the receiver front end, the portion that generates the wide dynamic range. The four-FET amplifier (Fig 6.94) is normally followed by the major crystal filter used in the receiver. The bandwidth and performance vary with the members of the Triad. The main IF system is the design offered by Carver in *QST* for May, 1996, a circuit based upon the Analog Devices AD600. The rest of the receiver is standard, although DSP enhancements are planned. The plans also

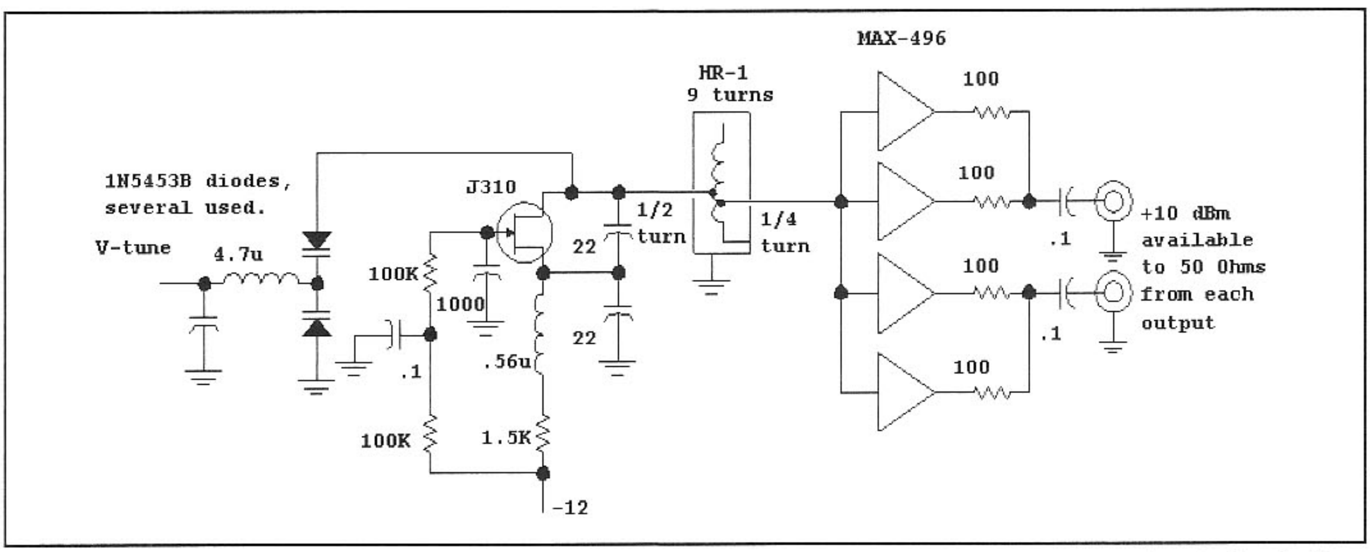
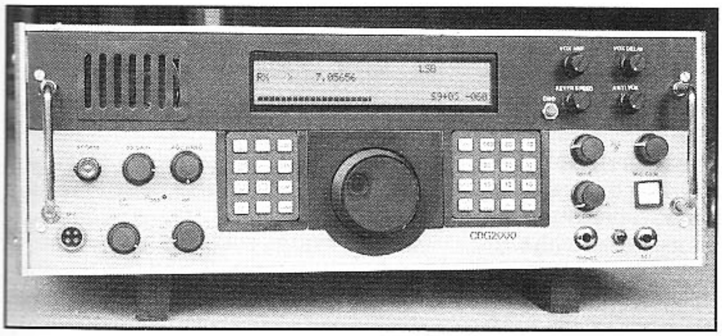


Fig 6.96—VHF helical-resonator voltage-controlled oscillator. See text for additional detail. Although a back-to-back pair of varactor diodes is shown, more may be required. It may also be useful to switch extra capacitance into the circuit with relays or PIN diode switches.



A working version of the Triad built in the UK. (TNX to George Fare, G3OGQ.)

call for full transceive capability.

The receiver performance has been outstanding with different triad members having obtained slightly varying results. With careful adjustment of the preselector and post filter amplifier, slightly under 10-dB noise figure has been measured in a receiver also showing an input intercept of +45 dBm. This is slightly under the early goal of achieving a 120-dB DR in an SSB bandwidth, but the ease of duplication of the FMT3125 mixer makes it preferable over one using the Si8901. That part had a 3-dB higher conversion loss, making it impossible to achieve a 10-dB noise figure without an amplifier in the "wide open" part of the front end. The present system with +45 dBm IIP3 and 10 dB NF (R = +35 dBm) will yield DR of 121.3 dB in 500-Hz BW.

There are some dramatic implications embedded within this work, ones that may well alter the way we design the next generations of receiver. It is clear that a lossy mixer can be followed directly by a narrow filter without compromising large signal performance. Use of the Engelbrecht technique is not new with filters, but it has not been routinely applied for experimenter equipment. The methods will work just as well with diode mixers as with FET mixers.

The typical high dynamic range receiver of recent vintage has consumed considerable power. This was generally accepted as the price one must pay for such performance. FET mixer based designs can, however, provide very high intercepts without high power. The oscillator powers are low, and with no early amplifiers, there is no compelling reason to use a high power amplifier anywhere in the system, especially if higher order, low loss roofing filters can be designed. Low loss and simplified matching should be possible with monolithic filter technology. We can now envision a very high dynamic range receiver that is as sensitive as we will ever need on the HF bands that operates efficiently with batteries.

But adequate challenge remains. The frequency synthesis problem continues to plague us. We certainly want new transceivers to include all of the refinements found in the older ones, and most of these features depend on frequency agility. The high phase noise of casual PLL synthesizers will drastically limit the performance. While somewhat better wideband phase noise is available from DDS, this is of little consolation when the noise is merely replaced by numerous coherent spurious responses. Some experimenters expect exciting things to happen in synthesis in the near future, which will help.[28]

But synthesis is not the major problem we face. Rather, it is the compromised nature of the transmitters that we usually encounter. It does little good to build a receiver that is so free of distortion that we become concerned about receiver damage when we measure it, only to find that the on the air signals we encounter are distorted.

Modern communications systems have been engineered with a sense of balance, using compatible transmitters and receivers. The receives have kept pace with the transmitters, but with little extra margin. The radio amateur service has not, however, grown in this way. Early stations had separate equipment for each function. We have had a DX based fetish for receivers, traditionally dealing with the classic axiom that "if you can't hear 'em, you can't work 'em." This left us ignoring our transmitters.

Many solutions to transmitter problems are found in the receiver design details. Improved receiver synthesizers will benefit our transmitter. High-level mixers, low-distortion amplifiers, and clean filters are elements common to both. The problem unique to the transmitter is in the higher power stages where distortion usually occurs. Even here, there is new technology that offers solution. *Feed forward* methods offer one route to reduced IMD.[29,30,31] Feedback and predistortion offer alternative routes.[32,33] Predistortion is discussed, with references, in Chapter 10.

6.6 TRANSMITTER AND TRANSCEIVER DESIGN

System Considerations; Transmitters with Mixers

A block diagram for a simple CW transmitter was presented at the beginning of this chapter, Fig 6.18. In the simplest form an oscillator is amplified, low pass filtered and applied to an antenna. The more elaborate scheme uses a frequency multiplier, allowing the use of a lower frequency oscillator, isolated from the higher power amplifiers later in the system. These represented the simple equipment that many of us used as we began our experimental efforts in radio. It remains a good design. Even with frequency multiplication, the only spurious responses are either harmonics of the output, or harmonics of the lower frequency oscillator. The former are substantially reduced with suitable low pass filtering while the latter are reduced through bandpass filtering immediately after the frequency multiplier.

The best frequency multipliers are those with balanced circuitry. Appropriate circuit symmetry will suppress the fundamental and some undesired harmonics. For example, a push-push doubler, a balanced circuit with two diodes, will suppress the fundamental drive component in the output by 30 to 40 dB. Selective circuits afford additional suppression. Multiple resonator filters are recommended over single tuned circuits.

We can calculate the performance of low pass filters that might appear in a transmitter output. **Table 6.1** shows the suppression at the second and third harmonics of a carrier that is passed through a low-pass filter with a cutoff frequency 10% above the input frequency. The fil-

Table 6.1

	Attenuation at	
N	2f	3f
3	10 dB	21 dB
5	30	50
7	51	79
9	72	108

ters were designed for a 0.1-dB-ripple Chebyshev response. Filters with 3, 5, 7 and 9 components are considered.

The simpler filters are poor performers. The N = 3 low pass with two capacitors and one inductor offers surprisingly little harmonic attenuation. Other passband ripples may enhance performance slightly, but the dominant effect is just the number of components.

The more common transmitter block diagram, Fig 6.19, uses two oscillators heterodyned together in a mixer to produce the desired output. A bandpass filter is again needed to select the desired output component while suppressing the image as well as various spurious products. While frequency multiplier balance enhanced performance, a balanced mixer does nothing to suppress an image. The filter must now do all of the work. Frequencies should be chosen wisely.

Although we occasionally see a heterodyne transmitter using nothing more than a single tuned circuit, two or three resonator filters offer much better performance with only slight added complexity. Intuition suggests that the added insertion loss of a third order filter would complicate design. But one can increase bandwidth with a triple tuned filter to realize the same loss with greater stability, better stopband attenuation, and ease-of-alignment. Some special cases, such as VHF applications demand even higher order filters.

An often abused, sensitive parameter is mixer drive level. A normal diode ring (+7 dBm LO) should generally be driven with an RF input less than –10 dBm. Third-order IMD is not excessive at this level (important in SSB transmitters) and high order mixer spurious products are low. However, spurious products grow at an alarming rate with greater RF drive.

Mixer drive level should be established through careful measurement. Even if the builder does not have a high frequency oscilloscope or spectrum analyzer, he or she can always build and use a low-level power meter, often used with a step attenuator. See the measurement chapter.

A high level (+17 dBm LO) diode ring functions well with an RF drive of 0 dBm. Higher-level mixers are capable of even greater drive. Diode mixers are usually 50-Ω parts and are aligned with substitutional measurements, outlined in the measurement chapter. A Gilbert Cell mixer (NE602, MC1496) is usually a high-input-impedance circuit. It operates with a single-ended local oscillator level of 0.3 to 0.6 V, peak-to-peak, usually established with an in-situ (*in place* within the circuit) measurement. This is measured with a 10× 'scope probe attached to the LO or RF input of the mixer IC. The measurement may also be done with an RF probe and high impedance dc voltmeter, although this measurement is rarely as accurate owing to levels that crowd diode thresholds. The allowed RF drive can be 0.3 V peak-to-peak for a Gilbert Cell used in a CW transmitter, also established with an in-situ measurement.

Transmit mixers are best driven with harmonically clean sources. It is often worthwhile to low pass filter the LO input to a diode ring mixer, mainly for reasons of waveform symmetry. Excess even-order harmonic distortion may unbalance the mixer. The clipping action of the mixer diodes will convert a sine wave drive into a square wave, rich in odd-order harmonics. The RF input signal should be low in harmonics, for they can mix to generate spurious outputs. The usual diode mixer does not generate these harmonics in the same abundance that it does odd-order LO products. Similar arguments apply to Gilbert Cell mixers.

The levels recommended are derived from our observations, and could vary with different mixers. Mixers in SSB equipment are driven at an RF level dictated by IMD requirements while mixers in CW rigs are only constrained by spurious outputs far from the desired output. These spurious products can and should be reduced with filtering, but that is not possible with the closely spaced IMD products in SSB. The levels given are conservative results based on our results. Clearly, spectrum analyzer measurements are always preferred over simpler power level determinations.

Linear Power Amplifier Chains

Design begins with a pair of equal IF signals, or two tones. Recall that the *peak envelope power* (PEP) of two identical signals or tones is 6 dB above one of the tones. The output from a normal (+7-dBm LO) diode ring mixer driven with RF = –16 dBm per tone is –23 dBm per tone, or –17 dBm PEP. A typical bandpass filter might have a 3-dB insertion loss, producing a –20 dBm PEP output. Assume this will be used in a transmitter with a 10 W PEP output (+40 dBm PEP or +34 dBm/tone). The output low pass filter usually has negligible insertion loss, so a net gain of 60 dB is required. This can be obtained with three stages, although four, each using negative feedback, would be preferred, especially if wide bandwidth was needed.

Design of the amplifier chain is based upon cascade intercept calculations if SSB or other linear modes are planned. Assume our design goal is IMD at least 40 dB below each output tone (46 dB below PEP) during two-tone transmitter testing. Each output tone will be 6 dB below PEP, or 2.5 W (+34 dBm) per tone. The related IMD must then be over 40 dB lower at –6 dBm per tone. The required output intercept must then be half of this ratio, or 20 dB above the output, +54 dBm. Such levels are obtainable with high-level class-A amplifiers. The block diagram for this amplifier chain is shown in **Fig 6.97**. We have assigned the gain-per-stage values shown across the top of the figure. The intercept values for the individual stages were then adjusted to meet the specification. The final calculated result of OIP3 = +54.2 dBm is less than the value

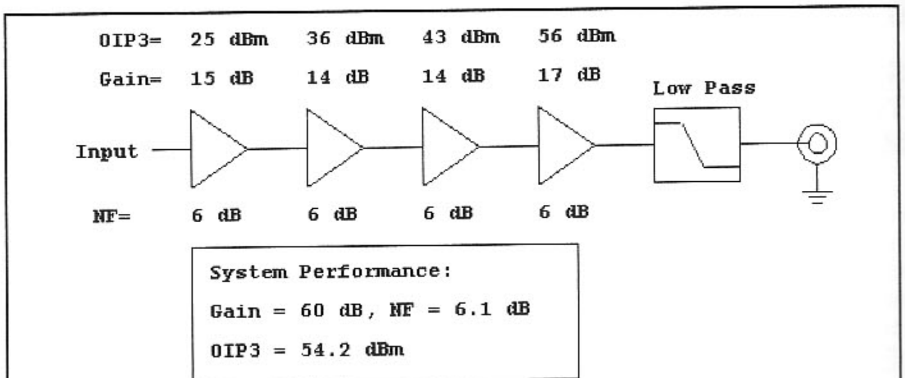


Fig 6.97—Individual stage parameters are combined for a cascade of four stages in an amplifier.

for the output stage itself of +56 dBm, allowing some of the distortion to occur in earlier stages. Increased output stage gain would relax the required earlier stage performance, but would reduce the margin for applying feedback in that stage. As in any practical design, this one is a collection of trade-off factors.

Noise figure is also calculated for the cascade, 6.1 dB based upon an assumed NF of 6 dB for each stage. If we assume a moderately low noise IF followed by a 10 dB loss in the mixer and bandpass filter, the output noise is essentially that of a resistor attached to the amplifier input. That noise is –174 dBm in a 1 Hz bandwidth. Adding 6.1 dB for the NF and 60 dB for gain, the wide band output noise density is –107.9 dBm/Hz. If this noise was to be sampled in a receiver with a 500-Hz bandwidth, total power would be –80.9 dBm. This is a very low power and would probably not be a problem for others using the same frequency. However, if another 20 dB of gain was added, bringing the output to 1000 W, the noise would be at –61 dBm. This noise would drop into the background at a distance, but could be troublesome for other stations in close proximity. This is a common difficulty with many stations in close proximity.

Transmitted phase noise is usually (much) greater than broadband amplifier noise. Consider a poorly designed transmitter with a synthesized LO generating phase noise of –120 dBc/Hz spaced 20 kHz from the carrier. If the carrier is amplified to a level of 1000 W (+60 dBm), the transmitted phase noise has a density 120 dB lower, or –60 dBm/Hz. If received with a 500-Hz-wide receiver, the noise is –33 dBm, or 0.5 μW. A low power transmitter of this level would probably not be heard at any distance, but can be copied by stations within a mile. The noise closer to the carrier will be much more evident.

The individual stages in the cascade of Fig 6.97 could be simple feedback amplifiers, biased to a high enough current that the individual stage intercepts are realized. The stages should present input and output impedances that match the adjacent stages, especially when wide bandwidth is desired. One may be more cavalier for a single-band CW design, although matched feedback amplifiers are still preferred, for they tend to preserve wideband stability. The emitter degeneration may be adjusted in a single band CW design to alter stage gain as needed for the desired output power. This practice should be used with more care when dealing with SSB.

A Class-A RF power chain can generally be built on a single board, for gain is modest. However, the board should end in a stage of around 1 to 10 W output. Higher-powered amplifiers should have separate power supply lines and an isolated thermal environment. A straight-line layout is recommended, separated from the band-pass filter that would normally follow the transmit mixer.

Fig 6.98 shows a two-stage class-A amplifier first presented over two decades ago. The design (like aging designers) is useful and robust in spite its age. The first stage uses a single TO-39 transistor biased to about 50 mA. Emitter degeneration and parallel feedback create low input and output impedance, presenting a good match at both ports. The second stage uses a parallel pair of TO-39 or similar transistors biased to about 250 mA. This circuit has a gain of 36 dB below 4 MHz, dropping to 29 dB at 29 MHz. The saturated output is a little over 1 W. IMD measurements at 14 MHz produced OIP3 of +43.5 dBm, making this a good starting point for low power SSB equipment. This circuit can also be used in CW applications by keying the positive supply to both stages with a robust PNP switch such as a 2N5322 or TIP-32.

A single-ended Class-A power amplifier is shown in **Fig 6.99**. This was built to investigate the performance of a variety of FETs as low distortion circuits. A 2N5947 bipolar feedback amplifier with measured OIP3 of +42 dBm preceded the circuit.

The first experiments used an IRF-510 HEXFET for Q1. With R2 = 1 Ω, an input network consisting of R1 = 47 with no input transformer, and with a 15 V power supply and bias adjusted for 0.5 A I_D, we measured OIP3 = +48 dBm. Increasing the

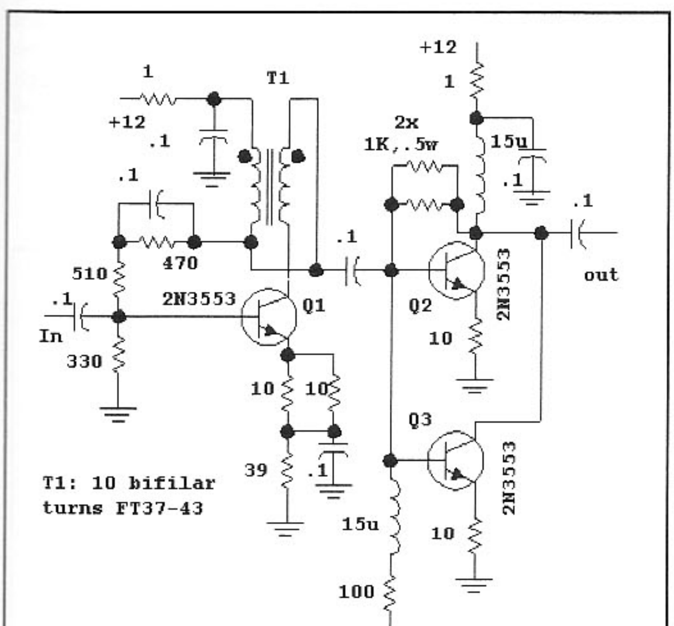


Fig 6.98—1-W power amplifier. Q2 and Q3 should have robust heat sinks if long operating periods are planned. If the 2N3553 is difficult to find, a Panasonic 2SC2988 can be considered for substitution. A single 2SC1969 might be a good substitute for the Q2 and Q3 pair.

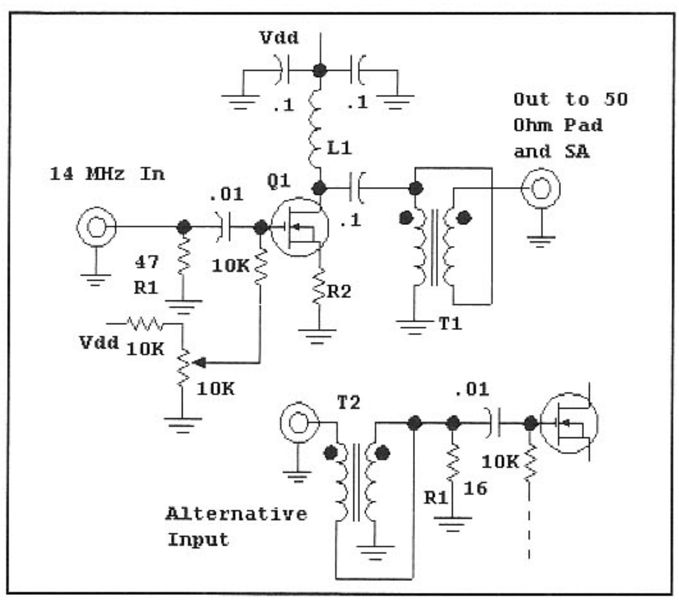
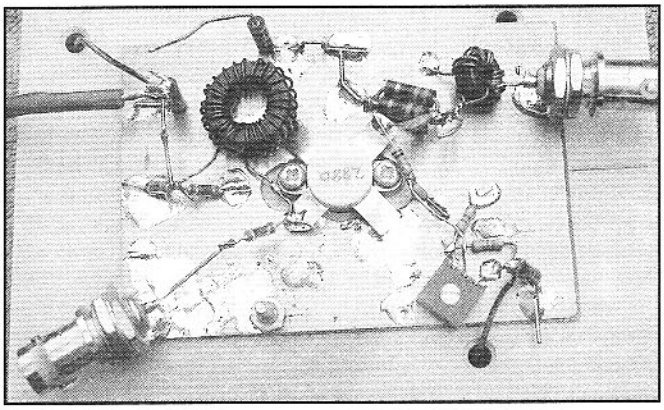
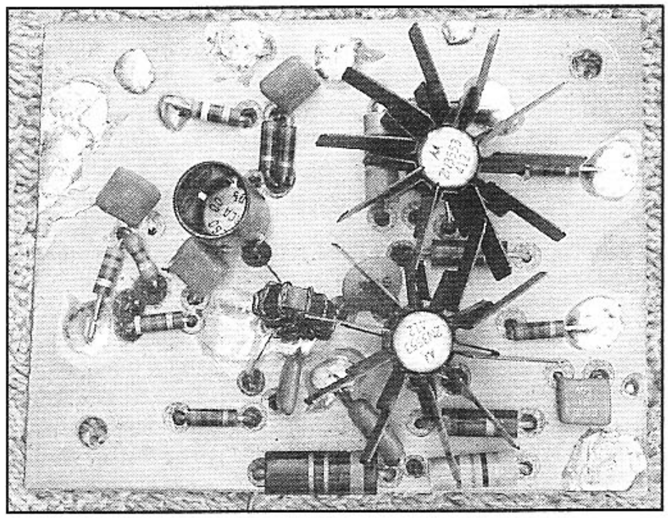


Fig 6.99—Class-A power amplifier experiment. Several MOSFET types were tried at Q1 while seeking high output intercept. L1 is 4 μH of #22 wound on a T68-2 toroid. T1 is 10 bifilar turns #18 on an FT-82-43 ferrite toroid. T2 is 8 bifilar turns #22 on an FT-37-43. R1 should have a 1-W power rating. Class-A amplifiers like this should be mounted on a large heat sink, for efficiency is not a feature of the design. See text for details.



Experimental Class-A FET RF power amplifier.



One-watt output Class-A bipolar-transistor power amplifier.

power supply to 25 V with I_D = 0.75 A yielded OIP3 = +51 dBm with 19-dB gain. The HEXFET seemed to *want* high drain voltage and did not provide low distortion performance with a 12-V supply. Experiments with the larger IRF-530 and the alternative input network produced similar results. The HEXFETs were thermally unstable at high drain current without the source degeneration resistance.

The next tests used a FET specified for RF performance, a now obsolete Siliconix DV-2880T. The alternative input network provided a lower driving impedance for the gate. High drain voltage was again required to obtain low distortion. With V_{dd} = 25 and I_D = 0.8 A, this device produced OIP3 = +57 dBm with 21-dB gain. The measurements were performed with outputs of +30 dBm per tone, or 4-W PEP. Slightly higher standing current should be used for a full 10-W PEP output.

The designer/builder could investigate other available FETs or power bipolar transistors. It appears that intercepts around +60 dBm will be available with moderately priced devices, allowing construction of Class-A power chains offering stellar performance at the 10-W PEP output level when compared with that offered by commercial transceivers. The experimental methods presented can certainly be extended to higher power levels.

Class-A power amplifiers are very inefficient with values of 25 or 30% being the best one can expect with reasonable distortion. Indeed, 50% is the theoretical maximum. Solid-state Class-AB amplifiers are also inefficient with values of 30% being typical. But the numbers obtained with two-tone testing are only part of the story. The Class-AB amplifier uses only enough bias to turn the devices on, perhaps to a maximum of 10% of the peak current used. With typical speech containing low average power compared to the peak value, average current is low. The average to peak power ratio is usually increased with speech processing, but net current is still far below Class A values.

An outstanding example of a medium power Class-AB FET amplifier was offered by Sabin.[34] That design is on the book CD.

Balanced Modulators

The voice signal from a microphone is amplified and converted to an intermediate radio frequency with a mixer. After up-conversion, it is usually processed with a crystal filter to eliminate one sideband. A balanced mixer is virtually always used in this application, a requirement to eliminate the local oscillator feedthrough. The mixer used in this application is usually described as a *balanced modulator*; the local oscillator that drives it is *the carrier*. All of the considerations presented earlier for mixers continue to apply. The popular diode ring mixers perform well in this application, often needing no adjustments for carrier suppression. The newer (physically smaller) TUF series parts from Mini-Circuits are preferred over the older and larger SBL-1, both for size and carrier suppression.

Fig 6.100 shows a simple balanced modulator design using two diodes. This is suitable for simple transmitters where the expense of a packaged mixers is to be avoided. The LO should be high enough to produce output that does not vary with LO drive, usually +7 to +10 dBm. Diode type is not critical. Silicon switching diodes such as the 1N4148 or similar will work well through the HF spectrum. Diodes should be matched for forward voltage drop with a current of a couple of mA.

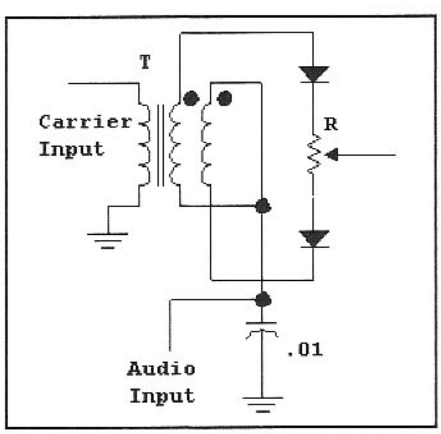


Fig 6.100—Simple balanced modulator for use in simple transmitters. R can be a small trim pot with R from 100 Ω to 2 kΩ. T is 10 bifilar turns on an FT-37-43 for HF applications.

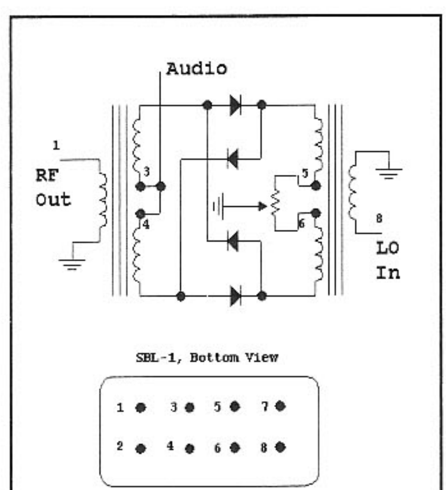


Fig 6.101—Adding balance adjustment to a balanced modulator using the SBL-1.

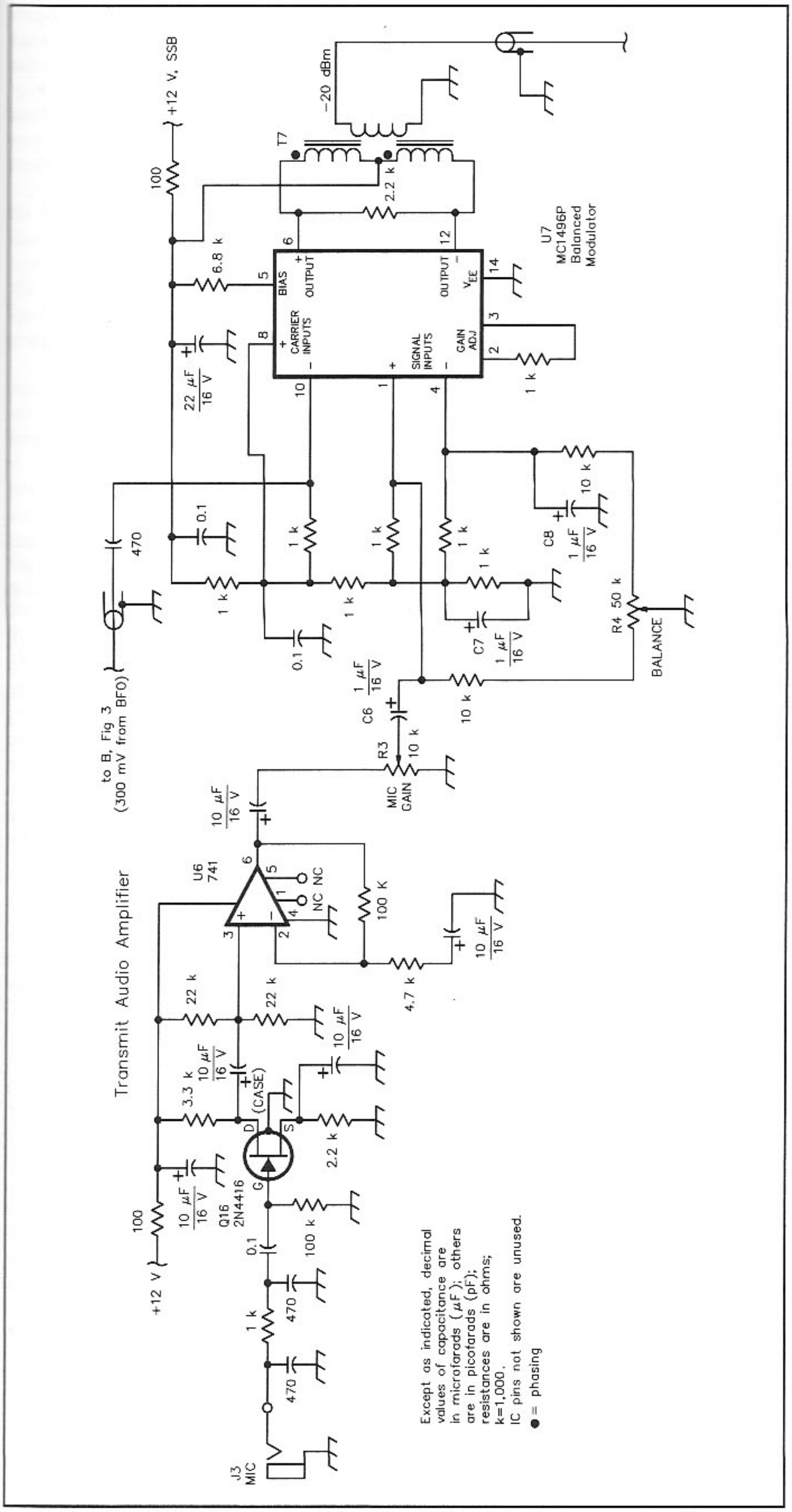


Fig 6.102—Speech amplifier and balanced modulator using an MC1496P. The transformer is 10 bifilar turns #28 on an FT37-43 with a 3-turn output link, used at 9 MHz. The carrier- balance pot is adjusted for minimum output at the carrier frequency. The dual in line version of the MC1496 is used here. Builders should consult manufacturer's data when using other variants.

Some builders have built very effective balanced modulators with the SBL-1 and similar Mini-Circuits mixers. But the topology is modified slightly from the expected where audio would be applied to pins 5 and 6, which were short circuited to each other. A modification used by W6JFR, shown in **Fig 6.101**, opens the short and inserts a low resistance (50 to 200 Ω) pot between pins. Adjustment of the pot allows the carrier to be nulled. Drive level considerations are still important.

The Gilbert Cell is an effective and popular balanced modulator. Fig 6.102 shows a simple speech amplifier and balanced modulator using the Motorola MC1496P. The internal circuitry for the MC1496 is found in the manufacturer's data, with fundamentals presented in Chapter 5. This circuit is capable of a carrier suppression exceeding 50 dB. Indeed, one can probably adjust it to even greater suppression, although it may be difficult to maintain this performance over time and temperature variations. The output with audio drive should be kept to about –20 dBm with this circuit. LO drive is 300 to 500 mV peak-to-peak, usually measured (in-situ) with an oscilloscope with a ×10 probe.

The speech amplifier used in **Fig 6.102** will accommodate both high and low impedance microphones. FET type is not critical. Most of the gain is provided by the op-amp. The builder may wish to use a dual op-amp with the other section configured as an active low pass filter. A project elsewhere in the book used this topology with a diode ring balanced modulator.

Transmitter IF Systems

The modulator output is routed to an IF amplifier. With a level of –20 dBm from the modulator and a requirement for only –10 dBm for a typical transmit mixer, little IF gain is needed. Indeed, most of the function of a transmit IF amplifier is that of signal conditioning and level control rather than gain. **Fig 6.103** shows an IF system. The first stage uses a common base amplifier, which provides good isolation between the modulator and crystal filter that follows. The amplifier also sets the termination impedance for the crystal filter. The amplifier and follower after the filter will establish the proper output level and gain. The follower provides a 50-Ω output impedance to drive a ring mixer while a 10-mA bias current sets low distortion.

A commercial crystal filter was used in the IF shown, part of an early transceiver.[35] The filter can be as simple as a 4th order Butterworth design. However, we have been disappointed with these simple filters. Filters with 6 to 8 crystals

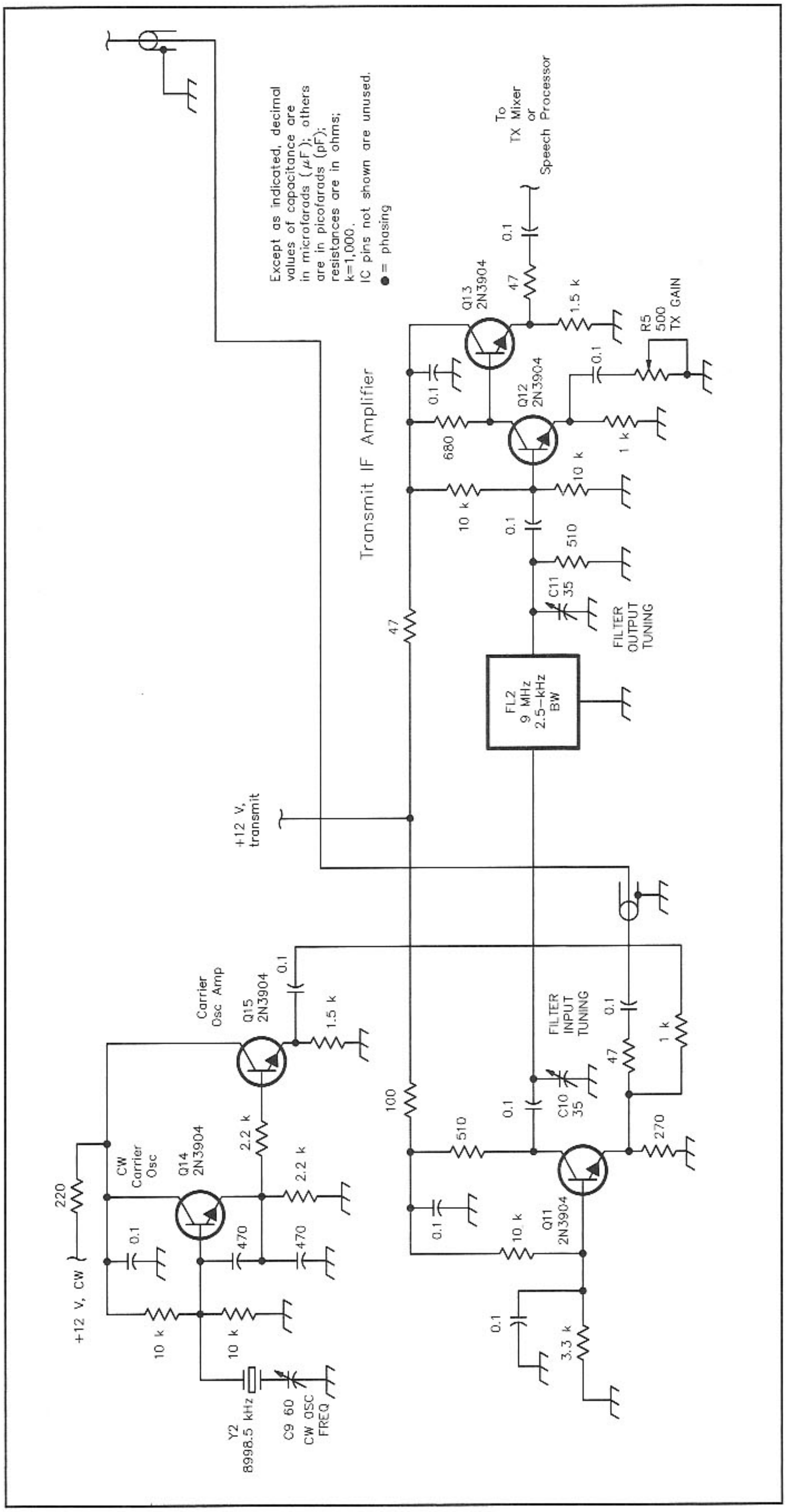


Fig 6.103—IF amplifier for an SSB transmitter. Very little IF gain is usually needed for this application. The trimmer capacitors were needed to terminate the crystal filter used on a transceiver using this amplifier, but may not be needed for other applications.

are little more complicated than a 4-pole circuit once the builder has been through the crystal characterization exercise needed when building filters. (See Chapter 3 for design details.) Yet the sideband suppression is dramatically better. Suppression is illustrated in **Fig 6.104** where overlapping 4 pole Chebyshev filter responses are presented. The level 6 dB down from the filter tops is marked, indicating the filter "passbands". The worst-case sideband suppression is about 30 dB, occurring for a 300-Hz audio note. Suppression approaches 60 dB at the highest audio input.

A Chebyshev filter shape is recommended for SSB applications over the simpler Cohn filter, which often suffers from poor passband shape. A comparison is made in **Fig 6.105**. The Cohn response, however, does have steep skirt attenuation, comparable to a 1.0-dB-ripple Chebyshev filter. Further, Cohn (equal coupling) filters built with lower Q_u crystals tend to have a smoother passband shape.

It is interesting also to compare available sideband suppressions with the responses of a phasing transmitter. The phasing system has the virtue of offering good suppression over the entire passband including the region close to the carrier. Hybrid systems with a phasing exciter followed by a filter could offer spectacular performance. (The same can be said for SSB receivers. See Chapter 9.)

CW Carrier Generation

The IF amplifier of Fig 6.103 includes a crystal-controlled carrier oscillator needed for CW generation. The oscillator and follower are relatively rich in harmonic energy, which might normally constitute a problem. However, the harmonics are removed by passing the signal through the crystal filter. The carrier is injected into the IF strip at the common base stage. The 1-kΩ resistor can be adjusted so the CW level is the same as the peak SSB power. An even simpler IF system is clearly in order for designs intended exclusively for CW. The important criterion is to provide the right level for the transmit mixer, but no more.

The CW carrier oscillator shown in Fig 6.103 functioned well in this application. This oscillator was turned off and on only at the relatively slow T/R rate. A faster rate is needed in many higher speed applications. But keyed crystal oscillators are subject to chirp, a change in frequency occurring as oscillation builds in the circuit. The problem often gets worse at lower frequency. There are several solutions to the problem. The crystal oscillator can be configured for lower loaded crystal

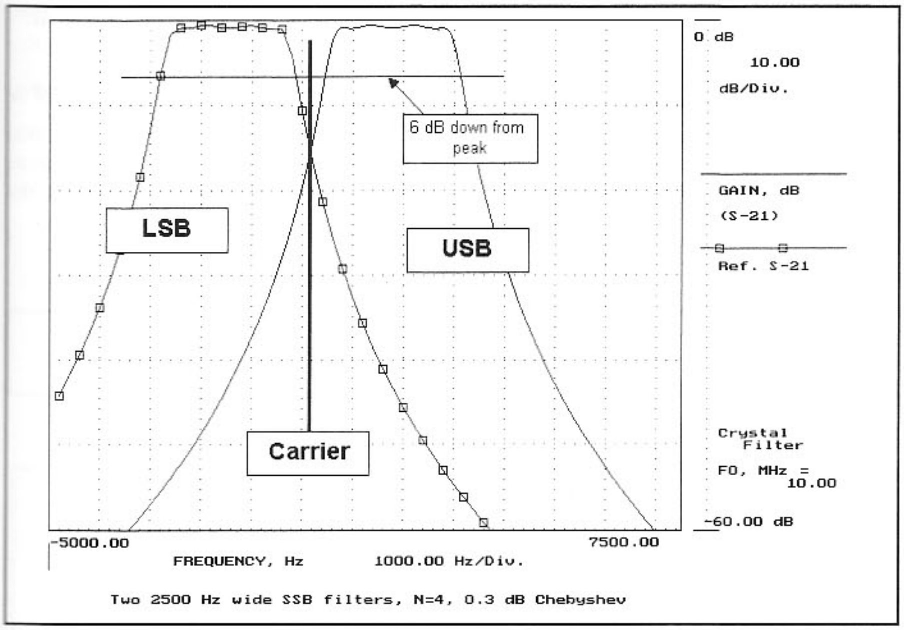


Fig 6.104—Two overlapping filters illustrating sideband suppression. See text. In a practical application, the filter response is measured and recorded in the builder/designer's notebook. The lower frequency 6-dB point is noted (for USB generation) and the carrier is placed 300 Hz below this point. The carrier is so marked in the figure.

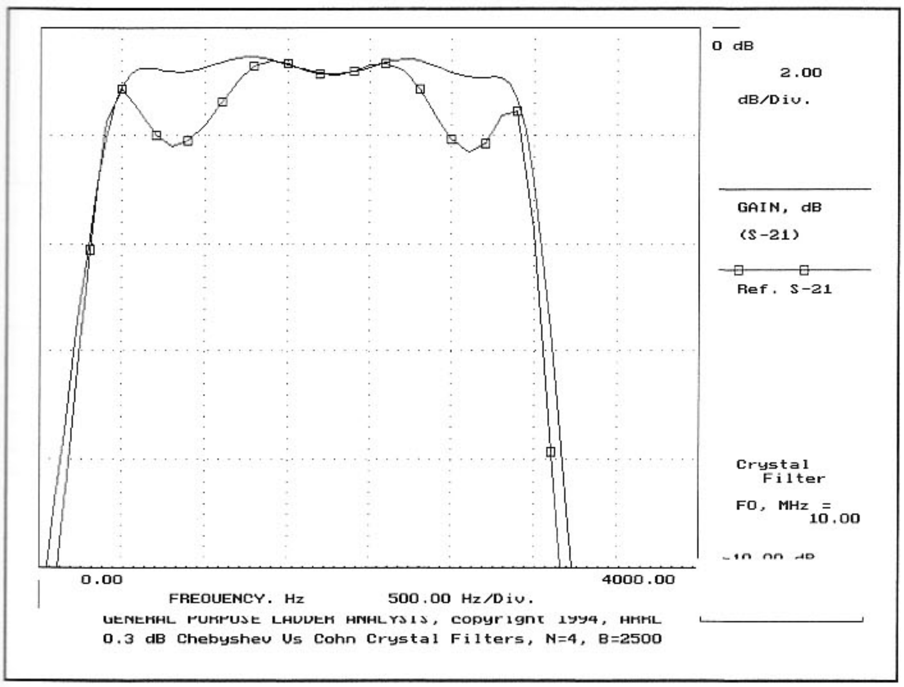


Fig 6.105—Two 4-element crystal filters are compared. The shape marked with small squares represents the Cohn filter while the other was designed for a 0.3 dB Chebyshev response. The two filters have similar skirt response, which is much better than a Butterworth shape, but much worse than a higher-order filter.

Q, often a difficult design task. A better solution uses an oscillator that is not keyed. The receiver BFO usually found in a transceiver is such an oscillator, but it is offset, operating at the wrong frequency. This slight change can be compensated with a suitable offset in the VFO. This is often a convenient solution, for RIT circuitry is already present in the transceiver.

Another alternative is a non-keyed crystal oscillator other than the BFO. But one can't normally use one within the receiver IF bandwidth, for it would be heard unless monumental efforts were taken to shield and isolate it from the receiver. Oscillator operation at a harmonic is often a convenient option. The signal is then divided with a digital divider during key down periods. One of our designs used a 5-MHz IF, but slight chirp was encountered when a 5-MHz crystal oscillator was keyed. The solution to the problem is shown in **Fig 6.106**.

Even though the free running oscillator in this scheme does not operate within the receiver IF, shielding is still required. A steady tone was heard when the 10-MHz oscillator was physically near the 5-MHz IF, a result of BFO second harmonic energy mixing with the higher frequency signal. Shielding and use of feedthrough capacitors for power and control eliminated the problem.

The non-integer frequency multiplication scheme described in Chapter 4 would also be well suited to generation of a CW carrier. That scheme divides a free running oscillator by 2, then uses one of the robust odd harmonics present in the square wave. In the prior example with a 5 MHz IF, a crystal oscillator at 3.3333 MHz could be used. It would be divided by 2 to produce a 1.667 MHz square wave that has a strong harmonic at 5 MHz. This could be filtered in a 5 MHz crystal or LC filter.

IF Speech Processor

The –10-dBm signal developed by the transmitter IF (Fig 6.103) is ready to drive a transmit mixer. Alternatively, it can be applied to an IF speech processor, shown in **Fig 6.107**.

The voltage related to a –10-dBm signal in a 50-Ω cable is only 0.1 V peak. This is not enough to turn on a diode. However, it can be increased with a transformer until diode clipping occurs. After the signal has been clipped, it is amplified and filtered. The filtering from the second crystal filter is necessary: without the filtering, intermodulation distortion products generated by the clipping circuitry would appear outside the IF bandwidth. Clipping cannot be done prior to initial filtering, for that clipping of the double sideband signal would create some distortion products within the eventual IF passband that would not otherwise occur.

The IF speech processor has the effect of increasing the average power within the speech sideband without increasing the peak. This higher average power increases intelligibility without excess distortion out of the normal passband. This processor, with the levels shown, increases the average to peak power by about 10 or 12 dB, readily observed with an oscilloscope.

The IF processor has a second advantage: It confines the IF level to prevent overdriving the transmit mixer. Without the processing, it would be desirable to add ALC, or "Automatic Level Control." This is an AGC loop in the transmitter that maintains the level through the overall power chain.

Intermodulation distortion is rarely a factor in a transmitter IF system. With so little gain required, the IF system can be simple. But the builder/designer should be careful to be sure that distortion is not an issue. It would be folly to design an extremely low distortion RF power chain only to feed it with the output of a mixer driven by a distorted IF signal.

Bidirectional Amplifiers

One view of a SSB transmitter says that it is nothing more than a superheterodyne SSB receiver with signals moving in the

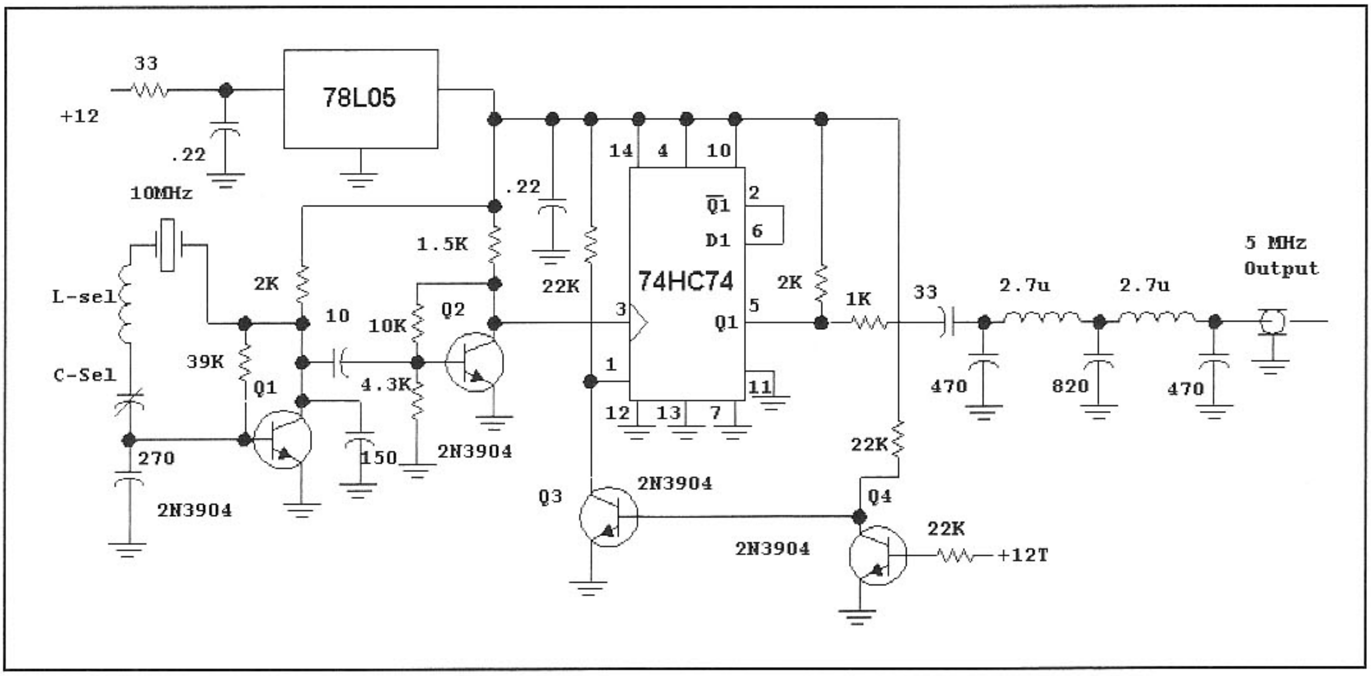


Fig 6.106—Alternative carrier-oscillator system for CW generation. A free-running 10-MHz crystal oscillator is divided with a digital divider to generate 5 MHz when needed. The divide-by-2 circuit is controlled with an IC reset line. See text.

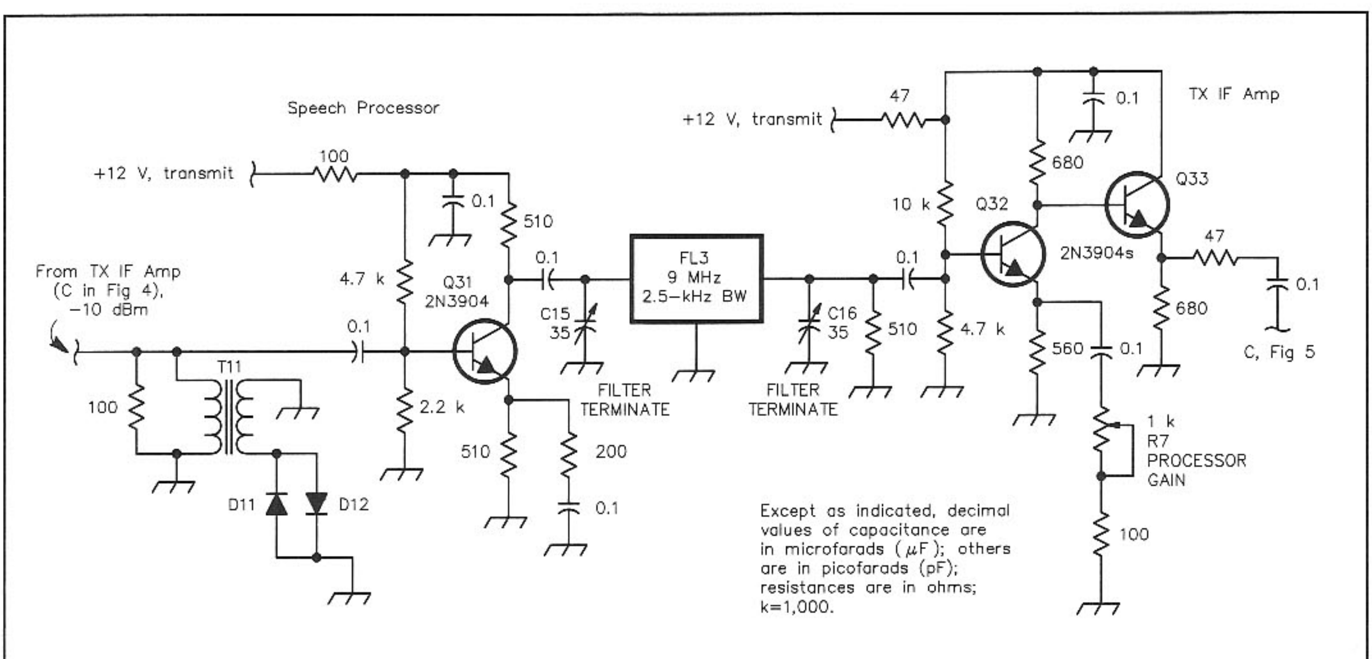


Fig 6.107—IF speech processor. Back-to-back diodes clip the IF signal. The resulting voltage is amplified and filtered in a crystal filter. It is then amplified and set to provide the desired –10 dBm to drive the transmit mixer. Schottky diodes are used in the clipper circuit. The diodes are driven by a 16-turn winding on an FT-37-43 ferrite toroid. The link on the 50-Ω line is 3 turns.

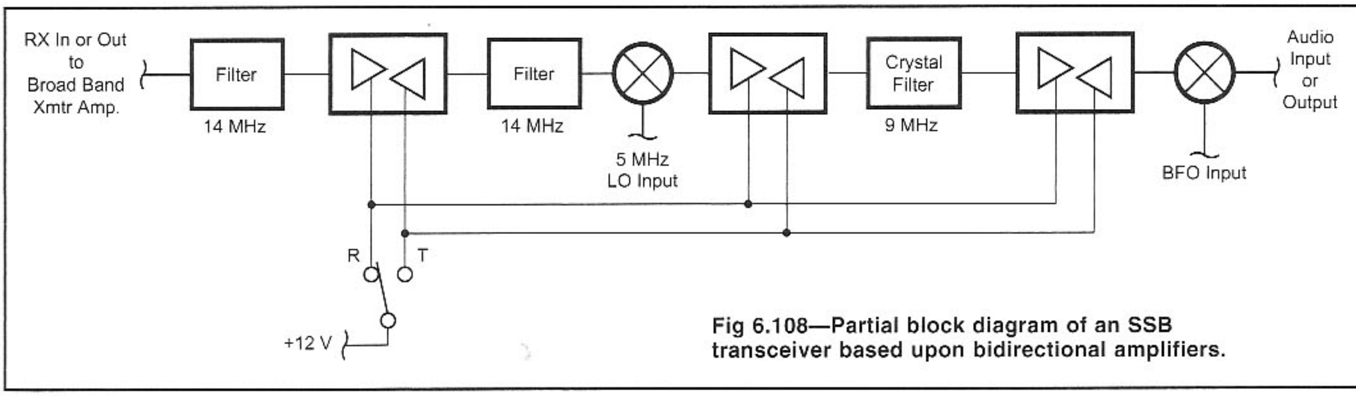


Fig 6.108—Partial block diagram of an SSB transceiver based upon bidirectional amplifiers.

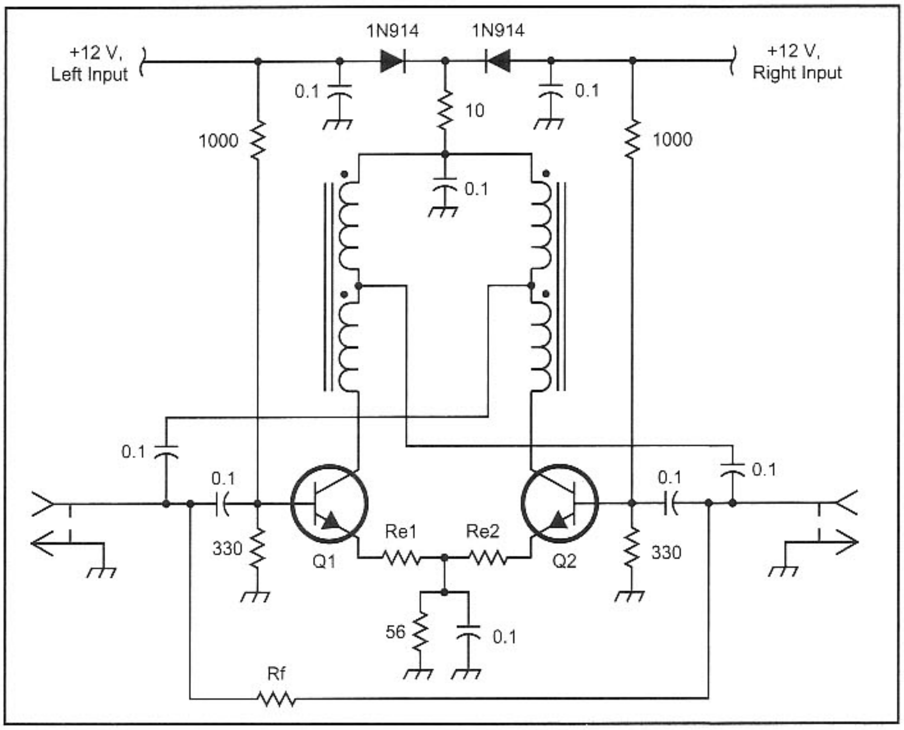


Fig 6.109—Bidirectional amplifier with bipolar transistors. Q1 and Q2 can be 2N5109s or similar parts. The input and output impedances are 50 Ω in both directions.

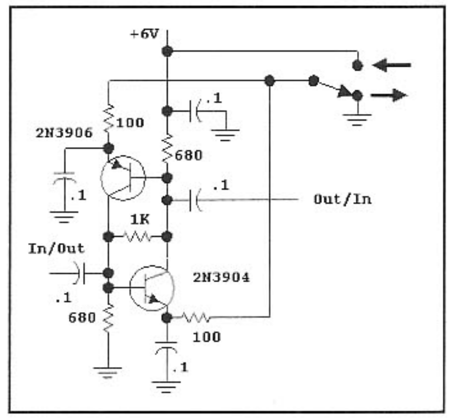


Fig 6.110—Bidirectional amplifier with complementary transistors. Only one transistor is on for each direction. Operation is clear if one of the transistors is mentally removed and the remaining circuitry is analyzed. See text for details.

opposite direction. The transmitter needs the same filters and oscillators as used in the receiver to create a SSB signal. Many transceiver designs have used this concept. A block diagram is shown in **Fig 6.108**. All of the RF and IF chain amplifiers are bidirectional; they provide gain to signals going in either direction when a dc control signal is changed. Diode-ring mixers are also bidirectional circuits, as are both LC and crystal filters. Audio signals can be switched with ease with integrated or discrete FET switches.

Fig 6.109 shows a circuit designed by the late Mike Metcalf, W7UDM. This circuit uses high F-t transistors biased to high current in the feedback amplifier circuit used throughout this book. The direction of operation is selected by applying Vcc to one of the two control inputs.

Very few of the components in the amplifier of Fig 6.109 are shared with switched directions. W3TS brought our attention to a simple bidirectional amplifier used in some "Manpack" transceivers built by Plessey.[36] We adapted this to the 50-Ω feedback circuit shown in **Fig 6.110**. The amplifier shown should be operated from a low V_{cc} to ensure that the emitter-base breakdown of either transistor is not exceeded. No emitter degeneration is used in the transistors, for each transistor is only biased to about 3.5 mA. Degeneration can be added for reduced gain or improved IMD. This amplifier will provide about 17 dB gain up to about 40 MHz. If redesigned for higher current, the 680-Ω resistors are replaced with smaller resistors in series with suitable inductors.

The junction field effect transistor is ideally suited to bidirectional amplifiers, owing to the usual symmetry of the physical device where the source and drain regions are identical. The drain only assumes drain-like properties when it is positively biased. A bidirectional amplifier using this is presented in **Fig 6.111**. A single-ended variation ("A" in the figure) shows the resonant drain network needed to generate high gain. This circuit appears twice in the bidirectional version ("B") of the circuit. A PIN diode short-circuits C-t when that portion of the circuit is used as an input. The low impedance then effectively short-circuits much of the tuned network. Input tuning can be implemented, if needed, by replacement of the RFC with small inductors. This circuit uses the metal can U-310 rather than the more common J-310, allowing a grounded gate with extremely low inductance, important for UHF stability.[37]

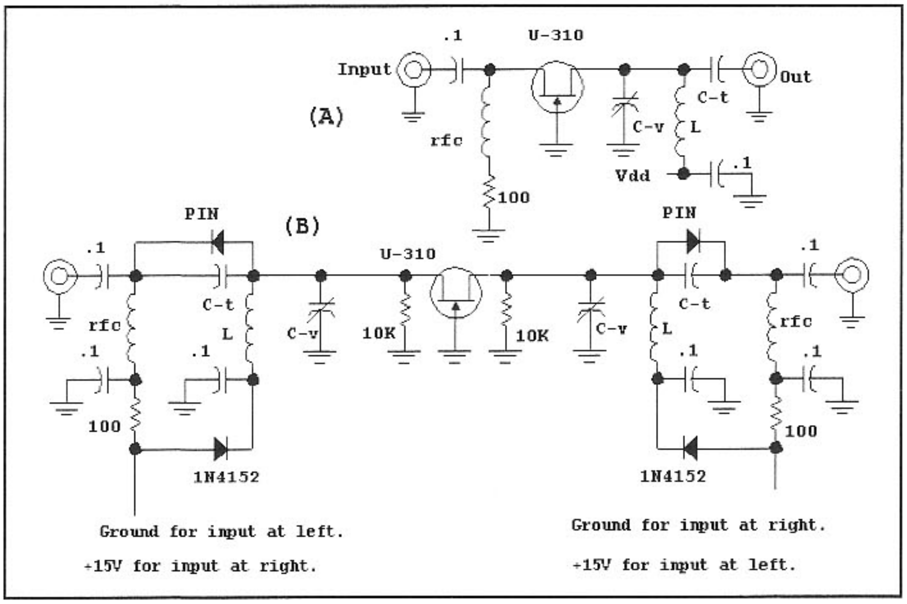


Fig 6.111—Bidirectional amplifier using a junction FET in a common-gate topology. Part A shows a single-ended amplifier where L, C-v, and C-t form a resonant network that presents a high impedance to the drain. Part B shows the bidirectional variation. See text.

Bidirectional Crystal Filter Circuits

Fig 6.112 shows a system with diode switching, allowing a crystal filter to be shared between receive and transmit functions. Diode D1 routes the signal to the filter input during receive while D2 connects to transmit circuits. R1 and R2 set D1 current during receive. The positive voltage developed across R1 serves to reverse bias the diode in the off path.

Part B of Fig 6.112 shows an option with an added transistor, Q1, in the receive path. Q1 helps to reverse bias the D1 anode and creates a low impedance to ground during transmit, both increasing the switch on to off ratio. Typical switch performance at 10 MHz will be a 45 dB on to off ratio with a 1 dB insertion loss.

While the diode switching looks simple enough, it is a critical transceiver circuit. The switching and the interfacing circuits should present the same impedance to the filter with switching to preserve filter performance. All components must be examined and, if needed, characterized for IIP3 as well as switching performance.

The best diodes to use in this application are PIN types. Lower cost high voltage rectifier diodes are often suitable, although they have higher off capacitance. We have measured IIP3 higher than +50 dBm for 1N647 and 1N4007 diodes. Less robust, but lower capacitance switching diodes are often used when crystal filters with a 500-Ω impedance are used. Careful experiments are then required to maintain IMD performance.

A scheme using a shared filter is shown in **Fig 6.113**. This method using NE602 Gilbert Cell mixers is the brainchild of K7RO.[38] Part A of the figure shows a partial schematic for a NE602. This part has good isolation between ports, a result of balance and the virtual cascode internal topology. This allows two mixers to be tied together to present a constant composite impedance to a filter, shown in part B of Fig 6.113. The mixer output impedance is 1.5 kΩ and remains even when the part is biased off. The input impedance is 3 kΩ, but is present only when the mixer is biased into operation. The output of U1, a receiver front-end mixer, and U2, a transmitter output mixer, are paralleled, pre-

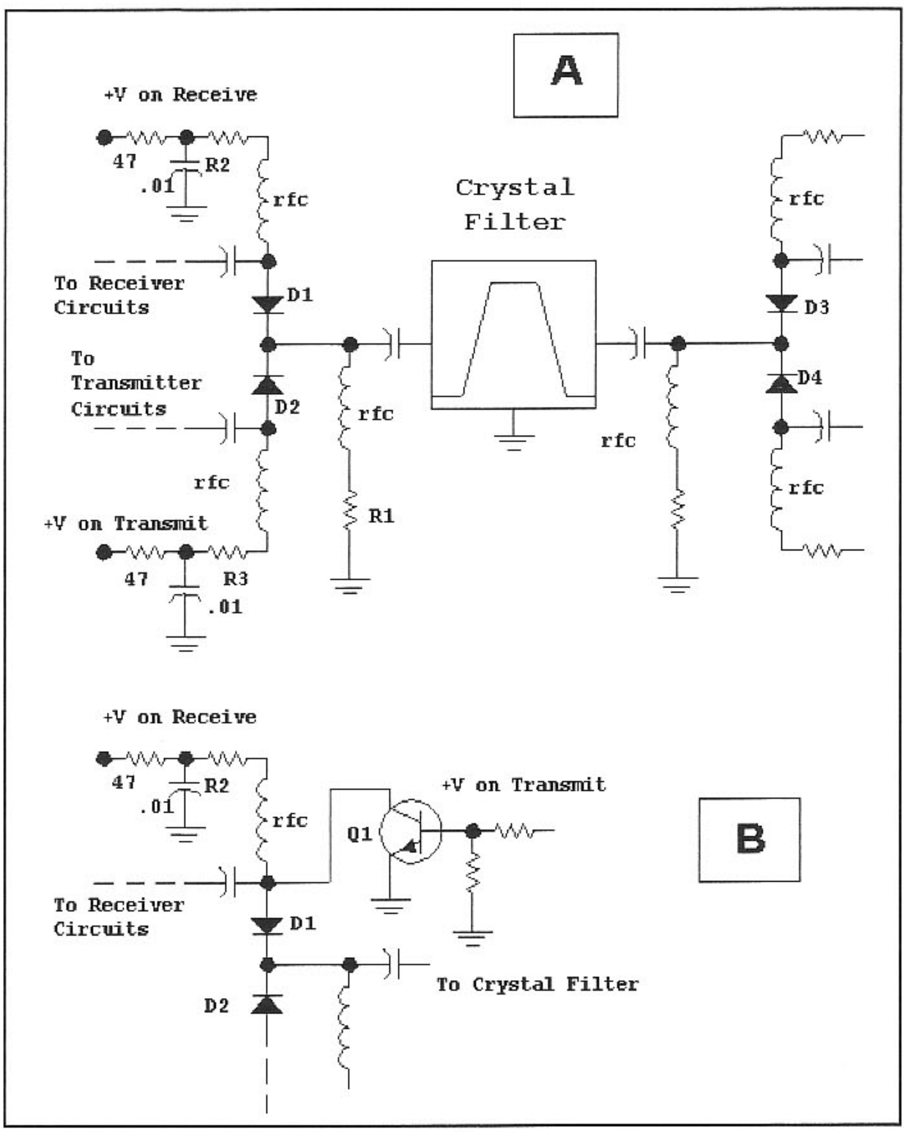


Fig 6.112—Diode switching of a crystal filter between transmit and receive functions. See text for details.

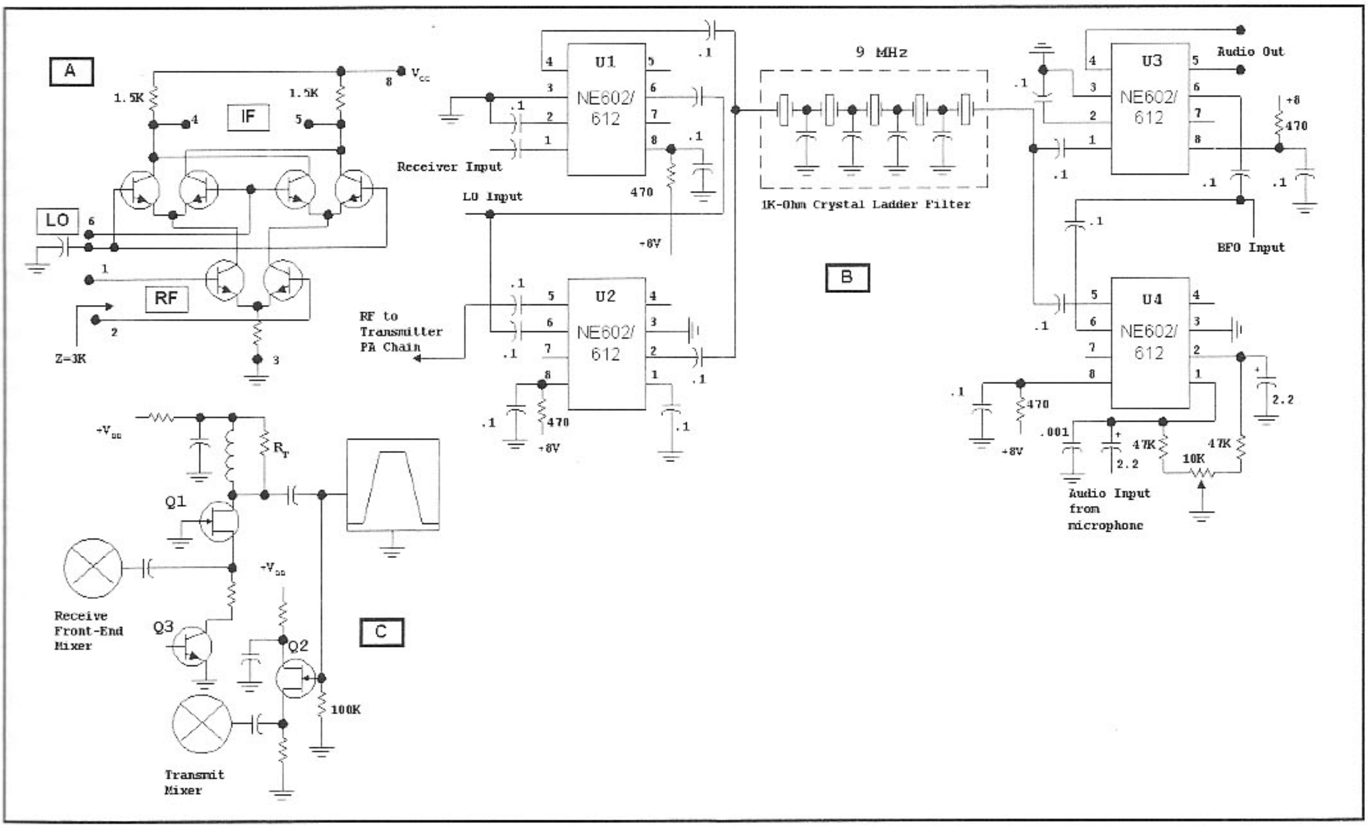


Fig 6.113—A scheme for sharing a crystal filter between functions. Part A shows a partial schematic for an NE602. Part B presents the basic scheme generated by K7RO while C shows FET buffers that allow other mixers and filters of many different impedances. The scheme in C has not been tried. See text for details.

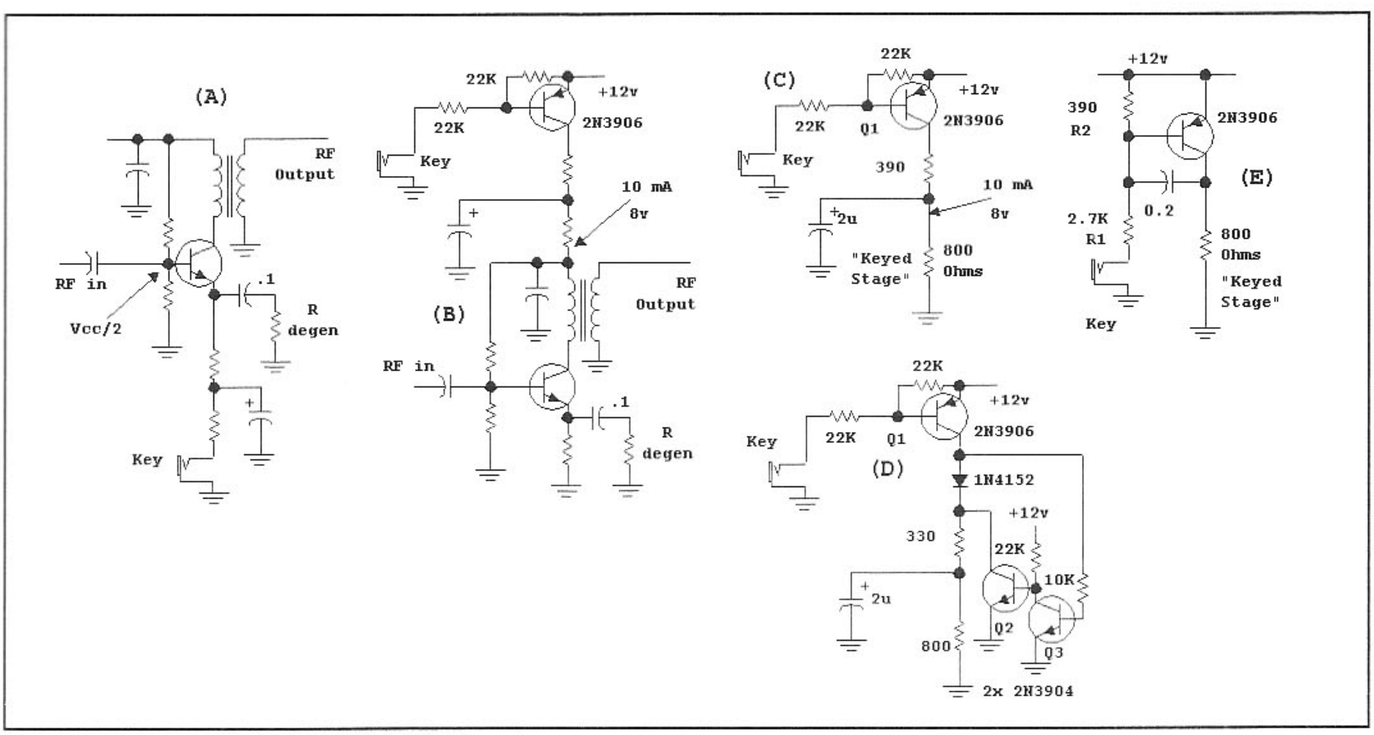


Fig 6.114—Circuits used to shape keying of a transmitter amplifier stage. Part A is a general case of switching an emitter current to ground. Part B uses a PNP switch to apply a keyed waveform to an NPN amplifier. If that stage draws 10 mA with 8 V applied, it is modeled as an 800-Ω resistor, as in Part C. Analysis of C shows an asymmetry. The rise is controlled by the equivalent of 390 Ω in parallel with 800 Ω while the fall is the result of the 800-Ω value alone. Part D provides nearly identical rise and fall times. E shows a modified switch where the PNP now functions not only as a dc switch, but as an integrator that shapes the rise and fall. See text for discussion.

senting a 1.0-kΩ impedance to the crystal filter. Local oscillator energy is simultaneously applied to both mixers.

Two more NE602 mixers are used with a similar connection to serve as a product detector (U3) and transmit balanced modulator (U4.) Biasing is slightly altered in Q4 to adjust balance.

One would ideally switch the mixers off and on to match their application. However, turning a mixer off that has an input that is shared with the output of another part will change the terminating impedance. The experimenter may wish to insert appropriate buffer amplifiers in the system to solve these problems.

The transceiver designed by K7RO used a crystal filter designed to have the impedance required by the mixers. Greater flexibility is afforded by the system in part C of the figure. Q1 functions as a common gate buffer amplifier, presenting a low input impedance such as might be needed for a diode ring receiver mixer. Q2 is a simple JFET follower to drive a variety of mixer types for the transmit function. Q3 is a dc switch that allows Q1 to be shut down during transmit intervals. Resistor R_T is the dominant element terminating the crystal filter.

Keying

Keying is the on-off control that is applied to a transmitter stage to generate RF in the pattern of International Morse Code. The keying circuitry can also control stages in a SSB transmitter when we wish to eliminate power consumption during receive periods. In principle, keying can be applied nearly anywhere in a transmitter. It is usually applied at an intermediate level and more than one stage is often keyed, especially when the following stages use linear amplifiers. It is acceptable to key just one stage when the following stages are nonlinear where bias is derived from RF input. The behavior we seek is a low *backwave*, meaning that the transmitted RF is low when the key is open. Backwave levels of –80 dBc are easily achieved.

Fig 6.114 shows several schemes for keying. Part A switches the emitter current, while the base is biased at about half the power supply. The electrolytic capacitor, the related stage current, and the resistor values time the rise and fall of the amplifier current. Both the rise and fall times should occur in a period of one or two milliseconds. Much shorter times allow key clicks to be created. Testing is normally done by examining the RF envelope with a high-speed oscilloscope, ideally while triggering the oscilloscope from the controlling dc.

The various parts of Fig 6.114 show a variety of shaping circuits, outlined in the caption. But the most popular is the simple integrator popularized by W7EL shown in part E.[39] The PNP transistor serves a dual role. The dc is switched, creating the basic function. But the transistor is also an amplifier that, in combination with the capacitor between base and collector forms an integrator circuit. No current flows when the key is up, bringing both base and emitter to +12 V, with the collector at ground. As soon as the key is pressed, current begins to flow in R1, causing the base voltage to begin to drop below +12 V. As soon as it gets to 11.3, base current begins to flow, forcing collector current to also flow which increases collector voltage. But the increasing collector voltage is coupled back to the base through the capacitor in a direction that "tries" to *reduce* the base current. This negative feedback does not let the collector voltage increase quickly, but forces it to ramp up at an approximately linear rate until the transistor begins to saturate.

The action is similar when the key is opened. The open R1 tries to reduce base current, which will let the collector voltage drop. But as that happens, base current will continue to flow through the capacitor as the collector voltage drops, again linearly, until the transistor finally turns off. R1 and C set the rising characteristic while R2 and C determine the fall. The traditional shapes of **Fig 6.115** approximate the linear ramp. Indeed it is the ramping part that is more effective in reducing clicks than is the rounded corners at the end of the shaping.

There are many methods that may be used to shape keying. In another W7EL creation (unpublished), a diode detector monitored the output of a transmitter. That signal was then compared with an ideal rise and fall in a dc-only circuit with an op-amp output controlling the gain of an amplifier. Shaping can even be done with DSP firmware, as presented in later chapters.

One sometimes sees simple transmitter circuits where a crystal oscillator is keyed. The result is often better than expected. This results from a general characteristic of oscillators—oscillation cannot start immediately, but must overcome the delay related to the bandpass filter intrinsic to all oscillator resonators. The resonator is the high Q crystal in this case. This behavior is usually not planned and should not be confused with design.

Although we emphasize shaping to reduce key clicks, some parts of the keying function must happen quickly. If an oscillator is keyed, it should occur quickly using circuitry isolated from shaping. The requirement for quick starting often precludes keying crystal oscillators. But keyed oscillators often suffer stability problems, adding challenge.

Generally, the following events must occur in sequence when a transceiver is keyed:

1. The receiver is operating normally.
2. The key is pressed to start a character.
3. The receiver is muted, preventing further audio from exiting.
4. Additional receiver muting is activated, preventing overload by strong transmitter signals.
5. The antenna is disconnected from the receiver input and is attached to the transmitter output. (In some cases, the transmitter output is already connected.)
6. Bias is established on important transmitter stages.
7. Oscillators are started and/or a frequency synthesizer is shifted and/or an RIT (detailed later) is shifted into transmit mode to establish the transmitted frequency.

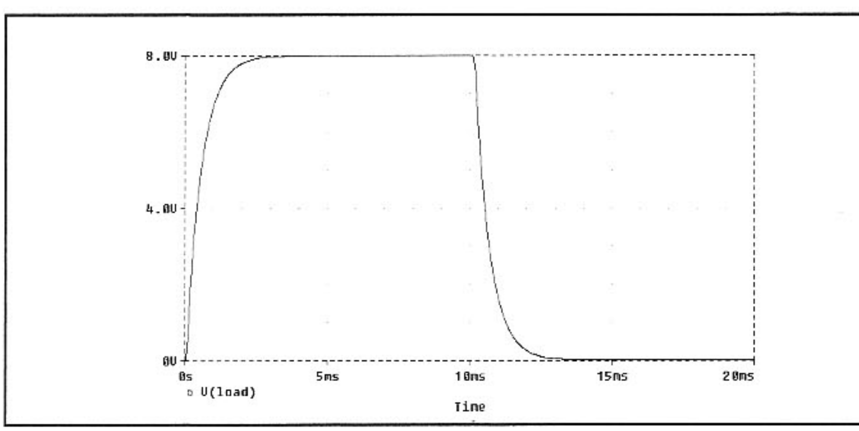


Fig 6.115—Desired waveform that should be applied to a keyed stage.

8. The keyed stages are supplied with the shaped dc that causes the desired waveform to be generated.
9. The dot or dash continues to be sent for the desired length.
10. The key is opened.

The sequence outlined is reversed, with the final event being the unmuting of the receiver, allowing the receiver function to return to normal.

Although not listed, it may be desirable to activate circuitry that "remembers" the gain state of a receiver at the exact beginning of a keyed interval so the receiver can immediately return to that state after the transmit interval is finished.

Muting a receiver can be a major challenge, especially if very high speed is desired. The high-speed operation is especially useful for QSK, or break-in CW operation where ideally a CW operator can hear other stations between high-speed dots. This facility is considered an advantage in competitive operations, but is also useful while exchanging routine or emergency traffic messages.

The simple way to mute a stage in a receiver is to remove the power supply. Unfortunately, this does not allow the gain to diminish or grow immediately, for bypass capacitors within the stage must charge and/or discharge with the switching. This process can often create transients that are as troubling as the presence of signal. The better method of muting a stage applies a gain altering bias that reduces gain without changing other dc parameters.

Even the "simple" circuit task of injecting an audio sidetone can be a challenge. Often a sidetone oscillator is keyed on or off in a way that creates a dc transient. That is, the "key down" waveform has an average value that differs from the value when the key is up. A better sidetone oscillator is one that has no change in dc level as it is turned on and off, and the best ones have shaping applied to the keyed waveforms.

6.7 FREQUENCY SHIFTS, OFFSETS AND INCREMENTAL TUNING

Oscillator Modifications

Both direct conversion and superhet transceivers usually include a provision to shift the frequency of the main oscillator when the key or push-to-talk button is pressed, causing the rig to shift from a receive to a transmit mode. There are various reasons for this shift, depending on the application.

Fig 6.116 shows several partial oscillator schematics that allow the frequency to be shifted in a discrete step as a control voltage is changed. The voltage changes between two well-defined levels producing two closely spaced output frequencies. The circuit in Fig 6.116A is an LC tuned VFO. The frequency is changed when a small variable capacitor, C_{var}, is shifted into the circuit with a diode switch. When the "control" signal is positive, dc current flows in the diode and C-var is part of the frequency determination. However, when the control voltage is set at 0, very little current flows in the diode switch, so C-var is removed from the circuit. The same coil tap used for oscillator feedback is used for offset. Additional capacitance, Cx, paralleling the diode will reduce shift, providing an adjustment.

A crystal-controlled oscillator with a diode switch is shown in Fig 6.116B. This circuit is ideal for shifts of only a few hundred hertz. The shift will depend upon the crystal parameters and the circuit design, so experimentation with C_{delta} is required.

A transistor is used as a switch in Fig 6.116C. The transistor saturates when the switch has base current applied, creat-

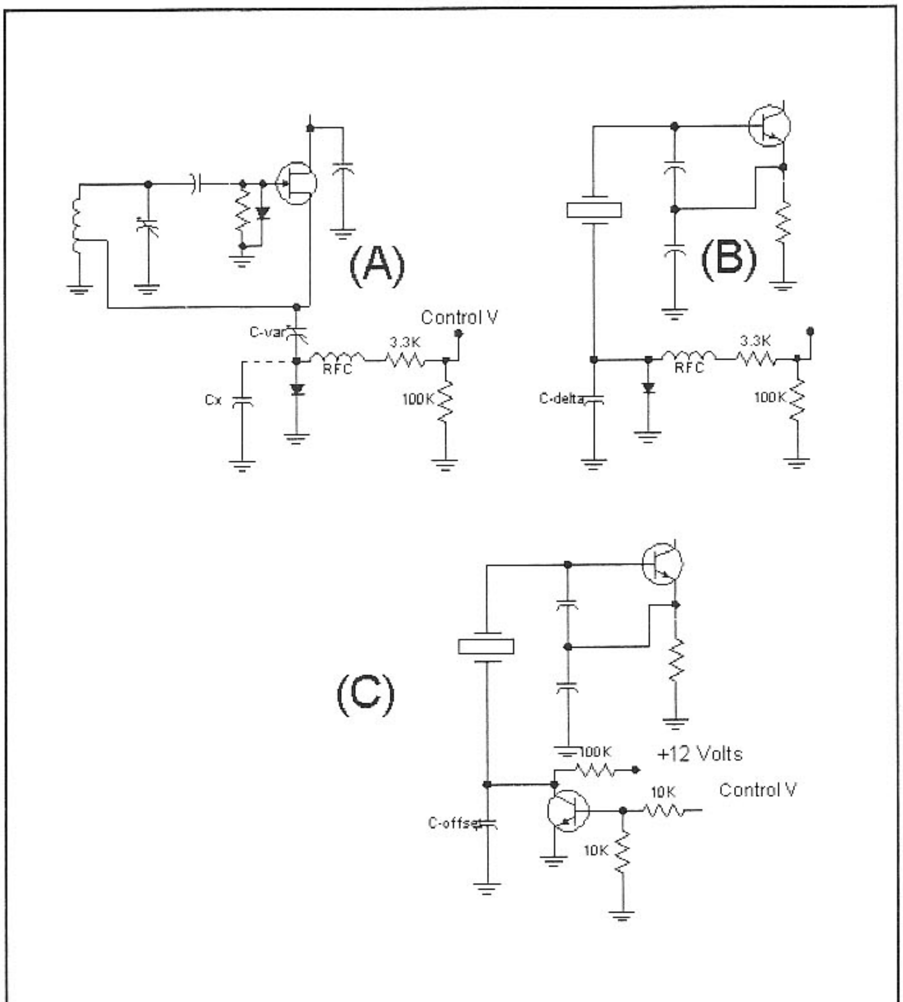


Fig 6.116—Oscillator circuits, including a means for frequency shifting.

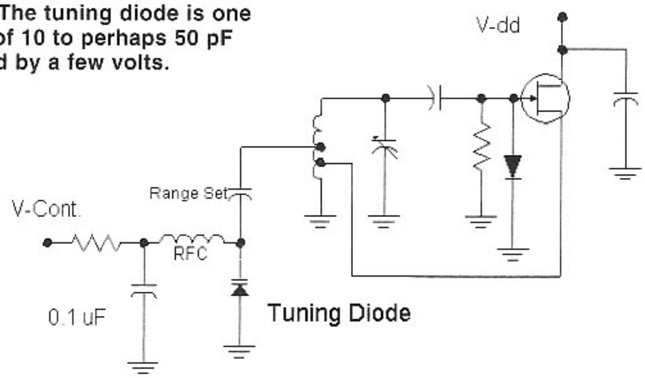


Fig 6.117—Modification of a classic LC oscillator for small tuning with a varactor diode. See the text for discussion of component values. The tuning diode is one with a capacitance of 10 to perhaps 50 pF when reverse biased by a few volts. Good choices for HF applications are the BB105 or BB109, or Motorola MV-209. Silicon power rectifiers or high-voltage Zener diodes are also sometimes used, encouraging experimentation.

ing effectively a RF short circuit. When base current is removed, the 100-kΩ collector resistor causes the collector voltage to rise, placing a reverse bias on the collector. The switch is then a small capacitor (a picofarad or two) that has less impact on the circuit.

A VFO example is shown in **Fig 6.117** where a traditional oscillator is modified with the addition of a varactor diode. For best stability, the "range set" capacitor is kept small, producing no more frequency shift than needed. Also, the voltage tuning range is picked to always reverse bias the tuning diode, even in the presence of large RF voltages. A typical circuit might have control voltage V_C that varies between 5 and 10 V dc. If the control drops close to zero, the RF will be rectified in the diode, altering V_C. This will often alter the Q of the oscillator tank and, in extreme cases, can cause oscillation to cease.

The bypass capacitor related to the tuning diode is shown as a 0.1 µF. A smaller value may be sufficient to decouple the RF. Values that are too large will slow the rate that frequency can change when the control voltage is altered, producing CW chirps or missed SSB syllables.

Superhet RIT

The most familiar application for the variable offset is *receiver incremental tuning*, or *RIT*, featured in most commercial transceivers. RIT is a simple function: During transmit periods, the transceiver frequency is determined by the main tuning system. But incremental tuning can become active during receive, allowing the user to adjust the received frequency by a small amount around the nominal transmit frequency. A typical range is +/– 3 kHz. Usual transceivers have a provision to turn the RIT function off, forcing the frequency of both transmitter and the receiver to be identical.

The RIT function might be controlled with the circuit in **Fig 6.118** where an operational amplifier determines the VCO control voltage. A 5-V regulator provides a stable voltage to drive the tuning pots and to power the oscillator. This is divided to provide 3 V for the noninverting op-amp input. A logic signal that is high during transmit periods is applied to the NPN, Q2. This saturates Q2 and cuts Q1 off, disconnecting the 10-kΩ summing resistor from the RIT pot, forcing the control to +7.5. The same result occurs when the "RIT-off" switch is closed.

The usual superhet transceiver generates the transmitted carrier by mixing the VFO output with a crystal controlled oscillator residing in the middle of a narrow IF bandwidth. During transceiver construction and alignment, the crystal oscillator is turned on and adjusted for a frequency that provides a desired beat note in the receiver, usually about 800 Hz. Then, during operation, the transceiver is tuned until an 800-Hz note is heard. Pressing the key then generates a signal that is exactly in zero beat with the received one.

Note that this operation and alignment is slightly different than that when SSB is generated in a superhet. In that case, the same circuit (usually crystal controlled) serves as the receiver beat frequency oscillator and the transmit suppressed carrier. It is important that an experimenter understand the frequency scheme used in his or her transceiver and the resulting operating mode. Also be careful to know when the RIT is active.

Offsets with Direct Conversion Transceivers

These basic superhet schemes will also work with direct conversion rigs. Consider a very simple 7-MHz direct-conversion

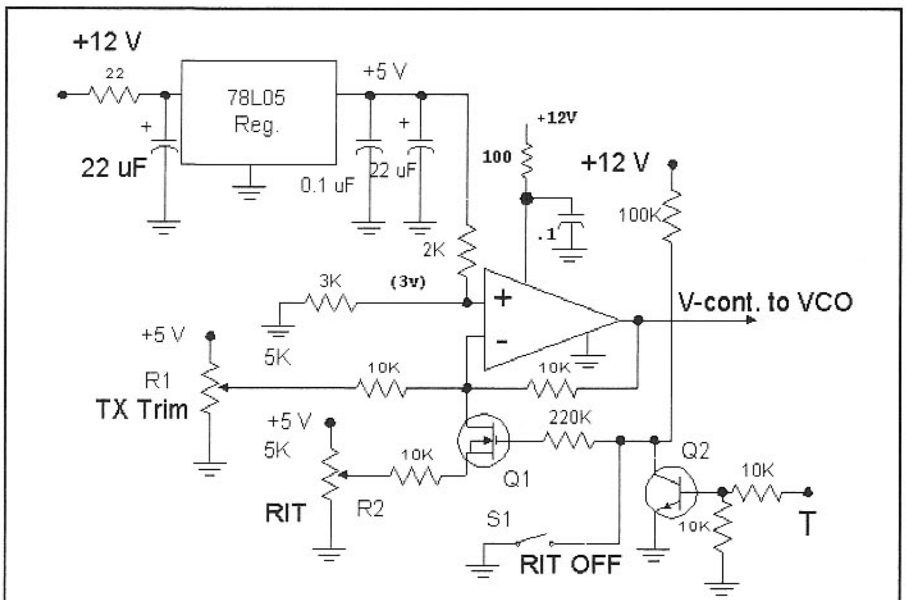


Fig 6.118—Circuitry to control RIT. Q1 is a TO-92 N-Channel MOSFET such as a 2N7000 or VN-10 or Zetex ZVNL-110A. Q2 = 2N3904 or similar. R1 sets the control voltage during transmit. The SPST switch is closed when the RIT is off. In this state, the control voltage should be approximately 7.5 V. The control voltage should vary between 4 and 10 with RIT on. Op-amp type is not critical; it could be a 741, half of a 5532 or 358, or similar.

CW transceiver using a VFO without offset or RIT circuitry. A simple switch transfers the antenna between transmit and receive functions, as needed. The transceiver is turned on and attached to a suitable antenna. The VFO is tuned, producing the expected collection of signals. A station is found calling CQ on 7040 kHz. Assume that you had been slowly tuning *up* the band when you heard this station. If you stopped tuning and listen to an audio note of 1 kHz, your VFO will be at 7039 kHz. If you tried to answer him, there is a high likelihood that he would miss you and would merely call CQ again. He will probably listen most intensely on his transmitter frequency of 7040 kHz.

A similar situation would have occurred if you had been tuning *down* the band. You would have stopped with your VFO at 7041 kHz to listen to a similar 1-kHz audio note, again transmitting off frequency.

Clearly, you must do something so that you transmit on the right frequency. One simple answer uses an offset generating circuit like that shown in Fig 6.116A. This circuit shifts the VFO *downward* by a fixed amount when the control is switched positive. The exact shift can be adjusted with a frequency counter, or by ear by listening to strong signals. The schematic is duplicated in **Fig 6.119**, which now includes needed control circuitry.

The system shown in Fig 6.119 is common for D-C transceivers. Pressing the key causes immediate PNP base current to flow. The collector goes up to +12 V, shifting the VFO frequency downward. When the key is let up, the frequency remains shifted for a short period controlled by the 10-µF capacitor and related resistors.

One tuning method emphasizes the SPOT switch. When a station is heard that you wish to call, the SPOT switch is closed and the station is tuned to zero beat (zero audio frequency.) This switch action is the same as pushing the key with the frequency shifted to the transmit state. Once the station is tuned to zero beat, the **SPOT** switch is opened. The station should then be heard with a 1-kHz note.

A second method is faster. When tuning and looking for stations to call, be sure that you are always tuning *down* the band, taking care not to tune through zero beat. It may be useful to mark the front panel with a small arrow next to the tuning knob, indicating the *proper* tuning direction. An error in picking the right tuning direction will now produce a 2-kHz error.

Extended use of a D-C transceiver reveals a subtlety: there is often interference when the VFO is on one side of the desired signal, but the other side is clear. It would be useful to be able to reverse the role of the offset. This leads to a modification of the usual scheme called "Almost Incremental Tuning," or AIT, shown in **Fig 6.120**.

Like the simpler system, the system with AIT is easy to use with a spot switch. Upon finding a station that you wish to work, tune to zero beat. Then throw the AIT switch. If there is interference, tune to zero beat and toggle the switch again.

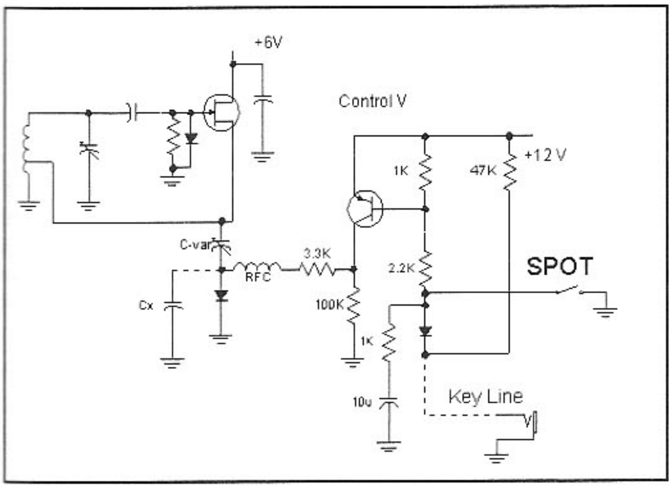


Fig 6.119—Offset system for a simple direct-conversion transceiver.

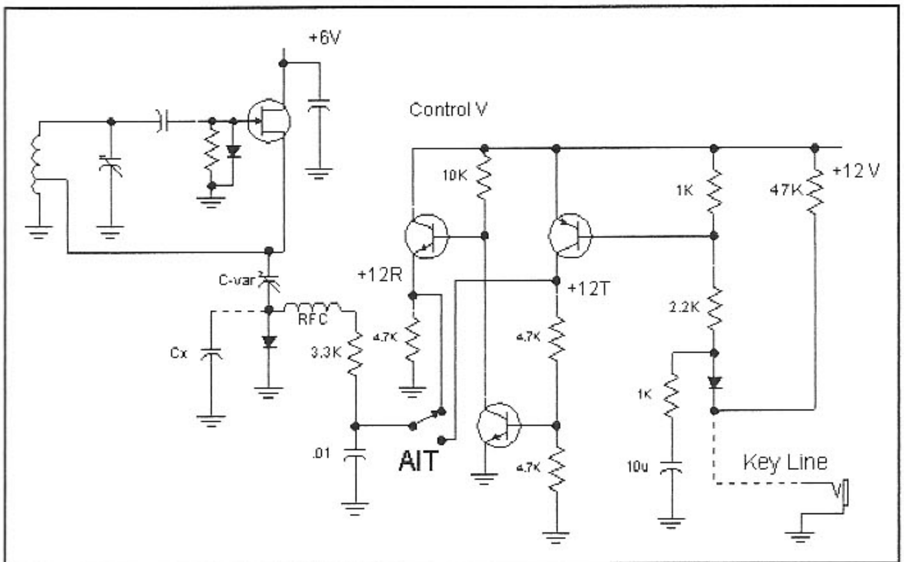


Fig 6.120—A VFO with offset capabilities and "AIT," Almost Incremental Tuning. This scheme allows the downward frequency shift in the VFO to occur on either transmit or receive, providing greater flexibility to avoid interference.

RIT with Direct Conversion

An RIT system is often included with a D-C transceiver. The utility of the feature helps immensely to overcome the deficiencies of the double-sided response. RIT can be accomplished at two different levels. W7EL popularized the simple scheme shown in **Fig 6.121**.[40]

A varactor diode is coupled to the oscillator through a small capacitor. During transmit or "zero" intervals, the bias on the diode is maximum at the level of the 9 V regulated supply. The voltage applied to the tuning diode during receive is less then the regulated supply, causing a downward shift in VFO frequency. The amount of the offset is tunable via the 20-kΩ RIT control. This scheme works well, providing all the adjustment needed for normal operation.

The complete superhet system can also be applied to a D-C rig.

One often encounters articles in the literature where VFO *offset* in direct conversion transceivers is discussed. The variety of offset options presented are sometimes referred to as having to do with "sideband selection." This term is not correct. The usual direct-conversion receivers using but one balanced mixer are not single sideband receivers (even though they can be used to receive SSB.) Moreover, they are usually used to listen to CW signals that do not include sidebands other than the closely spaced key clicks.

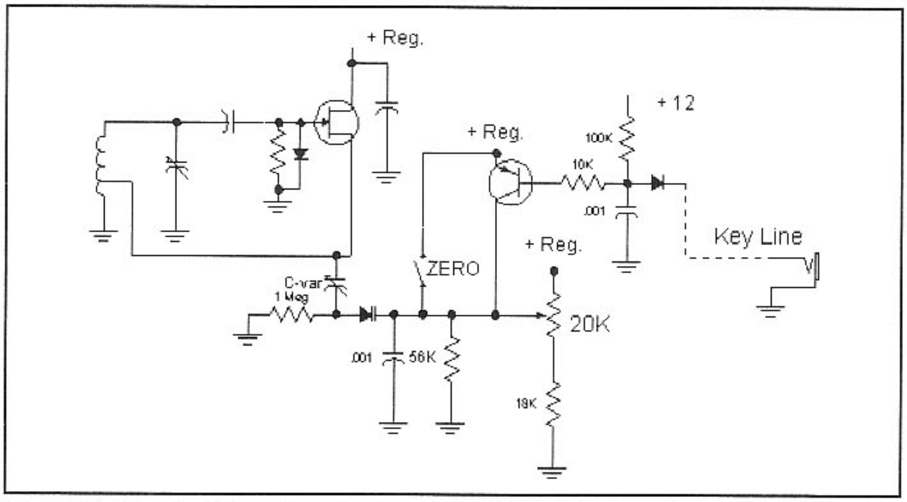


Fig 6.121—Simple RIT system developed by W7EL. This is a single-sided design where incremental tuning moves the VFO downward in frequency during receive periods, but only on one side of the transmit frequency. The general flexibility for effective RIT is retained. The tuning diode used by W7EL was actually a medium-voltage Zener diode, illustrating the simplifications that can be realized when one understands the behavior of the components. The system built by W7EL used a fixed capacitor where C-var is shown.

6.8 TRANSMIT-RECEIVE ANTENNA SWITCHING

An interesting design detail for a transceiver, and generally for any station is the way the antenna is switched between the receiver and transmitter. Something as simple as a manual switch will work and is used in some equipment in other chapters. However, the more common route uses either a relay or electronic switching methods. A traditional relay switch is shown in **Fig 6.122**. The RF part of the circuitry is presented in part A. The relay can be placed directly at the antenna terminal, but is shown here on the transmitter side of the usual low pass filter. Generally this scheme is preferred because the filtering is useful in both receive and transmit functions.

The example circuit in Fig 6.122B uses a 500-Ω relay coil. The relay current is switched with Q1, a saturated switch. Generally, the base current should be the collector value diminished by 10 to 20, so R1 is about 20 times the relay coil value. (The factor 20 is called a "forced beta" in this example.) Diode D1 serves to "catch" the voltage spike that will always occur when Q1 is turned off. Without the diode, the current that had been flowing in the inductive relay coil would "try" to continue flowing, generating the large spike as it charges the collector capacitance of Q1. This voltage surge can easily be large enough to destroy Q1.

If Q2 was not present, Q1 and the relay would be on. The base current in Q1 is shunted to ground through the collector of Q2 to control the relay. The Q2 base current is reduced by another factor of 20 over that in Q1. In the receive mode without the relay energized Q2 base current flows through R2 and the Zener diode, D2. The Zener voltage level is not critical, but should be near half the supply. The base current flows from the pot, R3, which provides a voltage from 6 to 12 V.

When the key is pressed, or a push-to-talk or VOX line go low, the base current in Q2 is diverted away from the base. Q2 then stops conducting, causing Q1 and the relay to switch on. Pressing the key, etc, also

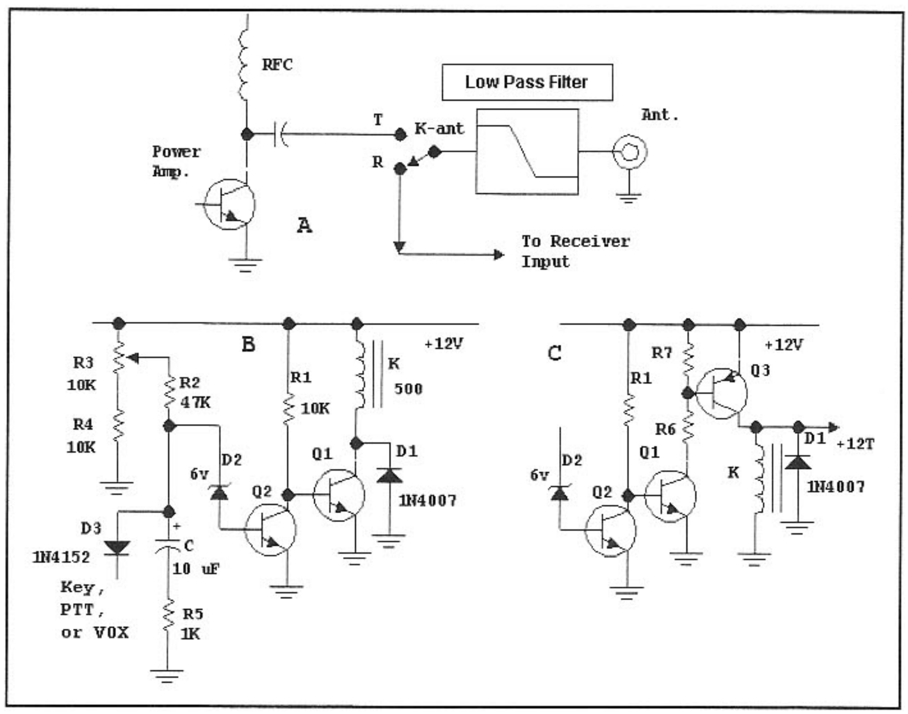


Fig 6.122—Relay T/R switching. The RF portion of the T/R switch is in part A while B shows a simple means for relay control. An expanded version is shown at C where higher relay current is allowed. Experimenters might wish to replace some of the transistors with some using built-in resistors found in parts catalogs, manufactured by Panasonic and others.

discharges capacitor C. Resistor R5 in series with C restricts the current that must be conducted in the key when switched. The circuit does not change states immediately when the key is released. Rather, switching is delayed by the time interval required for C to charge through R2 until it reaches the Zener voltage.

Plastic switching transistors such as the 2N3904 are fine for Q1 and Q2. Fig 6.122C shows a scheme with a PNP that can be used when the relay current is much higher, or when additional current must be supplied for other transmit circuit functions. R6 is picked to provide a Q3 base current of about 5 to 10% of the current that must be supplied by the Q3 collector. General purpose PNPs for this application are the 2N5322 or the TIP32.

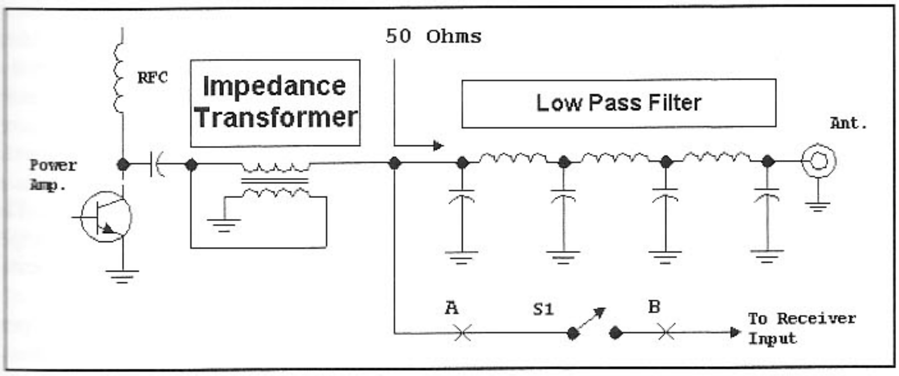


Fig 6.123—The RF portion of a T/R switch using a single switch. The transmitter is always connected to the antenna.

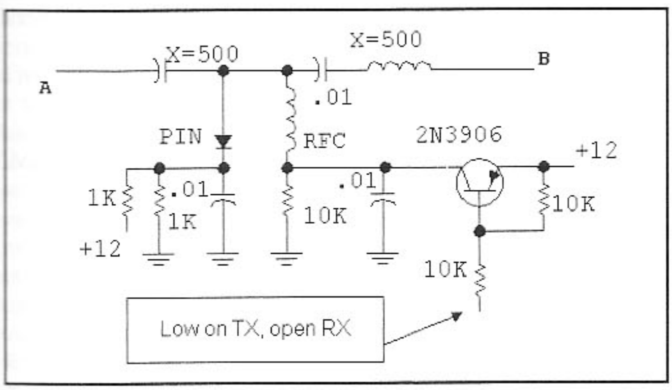
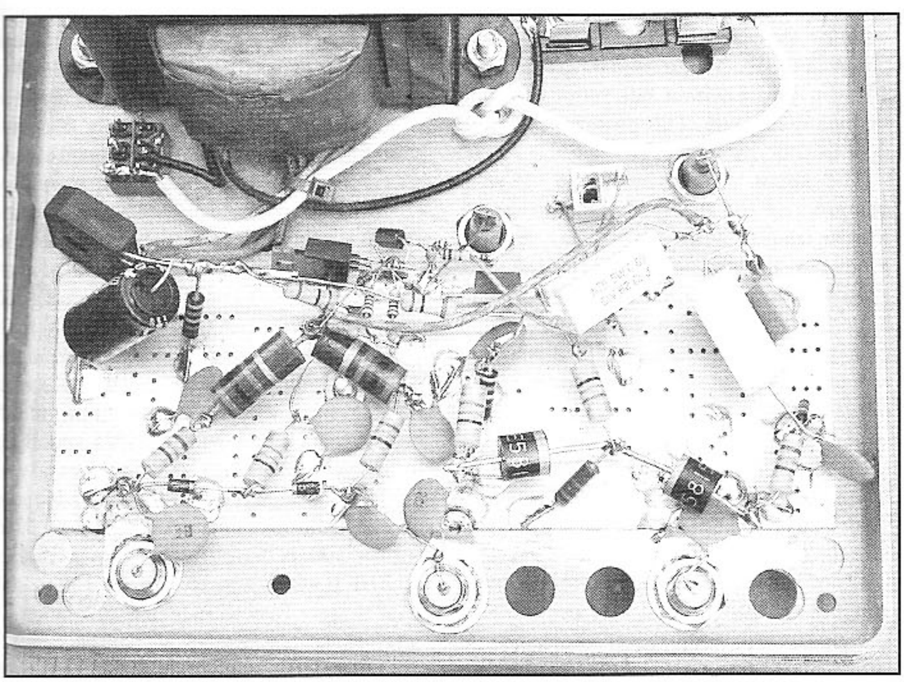


Fig 6.124—T/R switch with a shunt PIN diode.



Inside view of 100-W T/R switch using inexpensive diodes.

Fig 6.123 shows a common transmitter topology where the power amplifier (PA) is always attached to the antenna. The PA is cut off during receive periods, so it is essentially an open circuit with some parallel capacitance. Antenna energy is extracted through switch S1 to the receiver. This scheme is common, but it must be applied with care. The PA must not be conducting during receive; if it was, the collector resistance would absorb some of the signal that would otherwise reach the receiver. Also, conduction would generate excess noise that would compromise the receiver. It is also important to tap the receiver signal from a point in the low pass filter where the response will be maintained. For example, replacing the broadband transformer with a tuned network might lead to a shunt tuned circuit that would short some of the receiver energy to ground.

In some designs, a transmitter matching network might present an impedance lower than 50 Ω to the PA. This occurs when the output power is more than a watt or so from a 12-V supply. It is often tempting to tap the receiver signal from the PA collector. This may work, although if the impedance is much less than the receiver input impedance, the resulting mismatch can compromise performance. A matching network may be needed at the receiver input to increase the impedance back to 50 Ω.

The two sides of S1 are marked with A and B. A variety of switch circuits may be applied to generate the desired function. One is shown in **Fig 6.124**. Here, the switch is not a series element, but a shunt one realized with a PIN diode. The PIN diode is a common type used for RF switching. It departs from a normal PN switching diode with an intermediate region of *intrinsic* silicon. This has the effect of reducing switching speed, now a feature rather than a deficiency. The diode appears as a low valued resistor to radio frequency signals, but still as a diode for the dc controls. A PIN diode is capable of switching an RF current that is much larger than the dc current flowing. In contrast, a normal switching diode must be biased to a direct current that exceeds the peak RF current that is to be switched. The circuit in the figure biases the diode to 6 mA during transmit periods.

The shunt switch is effective in switching because it occurs within a tuned circuit. The usual capacitor at the end of a 50-Ω low-pass filter will have a reactance

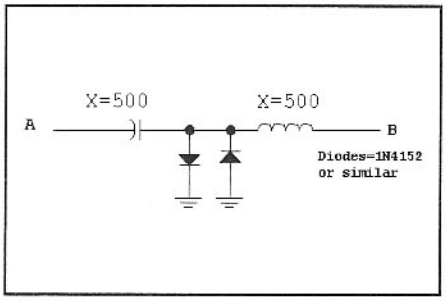


Fig 6.125—T/R switch with shunt PN diodes.

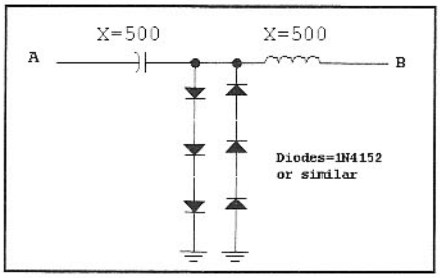


Fig 6.126—T/R switch with multiple PN diodes in each arm. This circuit features improved IMD. See text.

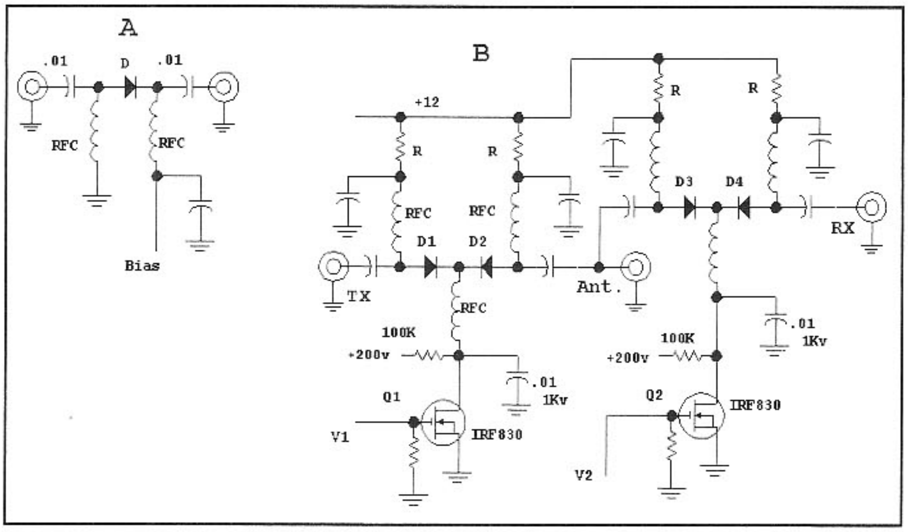


Fig 6.127—Part A shows the evaluation circuit. Poor "off" performance dictates the use of two series-connected diodes in each leg of the circuit in part B. Pick R to set the "on" current in the diodes.

around 50 Ω. The antenna signal is extracted from the low pass filter through a relatively small valued capacitor, one with a reactance of about 500 Ω. There is minimal receive loss, for it is tuned with a series inductor also with a 500-Ω reactance. When the junction of the two is switched to ground during transmit, the capacitor is merely paralleled with that in the end of the low pass filter, which will have little impact on transmitter performance. The inductance now in series with the receiver is useful in attenuating transmitter energy that might otherwise get to the receiver input.

A T/R switch of this sort is easily tested *before* a receiver is attached to guarantee that the power available to the receiver is low. The receiver end (B) of the switch is merely attached to a power meter and compared with the safe value for the receiver front end. A typical receiver with a diode ring as the first active element can usually tolerate 10 mW without damage.

The most common variation of the shunt T/R switch is shown in **Fig 6.125**.[41] Two common switching diodes (1N4152 typical) are placed in opposition. There is no controlling dc. Rather, when the transmitter is turned on, the RF causes the diodes to conduct, forming a relatively low impedance path to ground. We have measured this topology often (every time one is built) with the same result: The available output power at the receiver terminal is typically –10 dBm, easily within safe ratings for virtually any receiver. This power is independent of transmitter power.

The shunt diodes in Fig 6.125 can compromise the receiver dynamic range. Measurements with a 14-MHz example produced IIP3 of –3 dBm for the T/R switch, clearly a potential problem with high DR receivers. A solution is found in **Fig 6.126** where the single diodes are replaced by several series diodes. Two diodes per leg produced IIP3 of +7 dBm while three diodes per leg, the topology shown, yielded IIP3 = +13.5 dBm. The signals available at the receiver input increased to –4 and –1 dBm for the two and three diode per leg circuits. These levels will not cause damage to a receiver front end, but severe overload may occur.

Care is also required if these simple schemes are to be used at higher power. We have been able to extend the methods to the 100-W level, although only with circuit modification. The primary parameter to consider is the maximum current capability of the switching diodes. The 1N4152 that we have used in many circuits has a maximum current rating of 100 mA. The extended designs are discussed in a *QEX* paper.[42] This article is included in the CD that accompanies this book.

Another subtle, but significant problem occurs with this T/R scheme. The series-tuned LC is a tuned circuit that can interact with the tuned circuit(s) that follow to create a multiple-tuned circuit not in the designer's plans. The direct connection at (B) often leads to severe over coupling. The coupling can usually be adjusted to a proper level by inserting a suitable shunt capacitor at (B). Careful analysis is required.

Although the shunt diode switches presented are very useful for low power transceivers, they suffer from both IMD and power limitations, and are restricted to a single band. A wideband SPDT switch design with series diodes in the transmitter and receiver path would be more general. Our investigation of this topology begins with a simple single pole switch, shown in **Fig 6.127**, part A. This circuit is used to measure insertion loss and IMD with both forward and reverse diode bias. The IMD measurements should be done for both receiving conditions and at transmitter power levels when SSB use is planned.

High-power RF switching PIN diodes are available and discussed in the professional literature.[43] However, they are expensive and sometimes difficult to purchase. Our investigation, encouraged by K5CX, was directed toward inexpensive solutions. Many rectifier diodes are actually PIN structures, for this device topology tends to increase reverse voltage breakdown. The best inexpensive PIN diodes we found are the Motorola 6A6, a power supply rectifier specified for 6-A forward current and 600-V reverse breakdown. Diodes Inc manufactures similar parts. A forward bias current of 200 mA is enough for reliable operation at the 100-W level. We found identical performance with a NTE8515. We also got good results with the 1N4006, a 1-A, 800-V part.

While the forward biased performance was outstanding, the diode capacitance with reverse bias was relatively high, much higher than found with devices specified for RF switching. This made it necessary to put two diodes in series to obtain adequate reverse isolation. The SPDT topology used with a 100-W amplifier is shown in Fig 6.127B. It was necessary to go to 150 to 200 V of reverse bias to reduce capacitance of "off" diodes.

The reverse capacitance for the 6A6 diode was still 30 pF at 80-V reverse bias.

The 1N4006 dropped to 3.6 pF at the same bias. We also investigated a Motorola 1N4007, a 1-A, 600-V part and measured 2.1 pF at 80-V bias. In our final design we used the NTE8515 for D1 and D2 of Fig 6.125B, while 1N4006 diodes were used at D3 and D4. The 1N4006 was also satisfactory at D1 and D2 at the 100-W level, although this was not used for prolonged operation. The details of the T/R switch are shown in the *QEX* paper mentioned earlier. We used high-voltage HEXFETs for the bias switching. The switch insertion loss was so low that we could not measure it. Isolation was 56 dB between the TX and RX ports when the ANT port was 50-Ω terminated. IIP3 was greater than +40 dBm in the receive path. The IMD measurement was limited by the spectrum analyzer used and IIP3 may be even better.

We often wish to use a power amplifier driven by a transceiver. A suitable switching topology for this chore is shown in **Fig 6.128**. Three switches are shown. Only that at the PA output, SW3, would require the higher current diodes. SW1 and SW2 could use the less expensive 1N4006 or 1N4007.

Fig 6.129 shows a single band T/R switch using shunt PIN diodes, suitable for VHF as well as HF application. Quarter wavelength transmission lines interconnect the ports and switches. The diodes have reverse or zero bias during receive, but are forward biased during transmit. D1, behaving as an open circuit during receive, causes a short circuit to appear at the transmitter output. But open circuit D2 allows the nominal 50-Ω input of the receiver to appear at the antenna port. Switching to transmit forward biases both diodes. D1, now a short, reflects as an open circuit at the transmitter output. D2, also a short circuit, protects the receiver and presents an open circuit at the antenna port. The antenna impedance now appears at the transmitter output. This circuit can be implemented with true transmission lines or with pi networks as shown in Fig 6.129. The pi-network that behaves like a quarter wave 50-Ω line has L and C each with a 50-Ω reactance at the operating frequency. This circuit is used in a 17-m DSP-based transceiver presented later in the book.

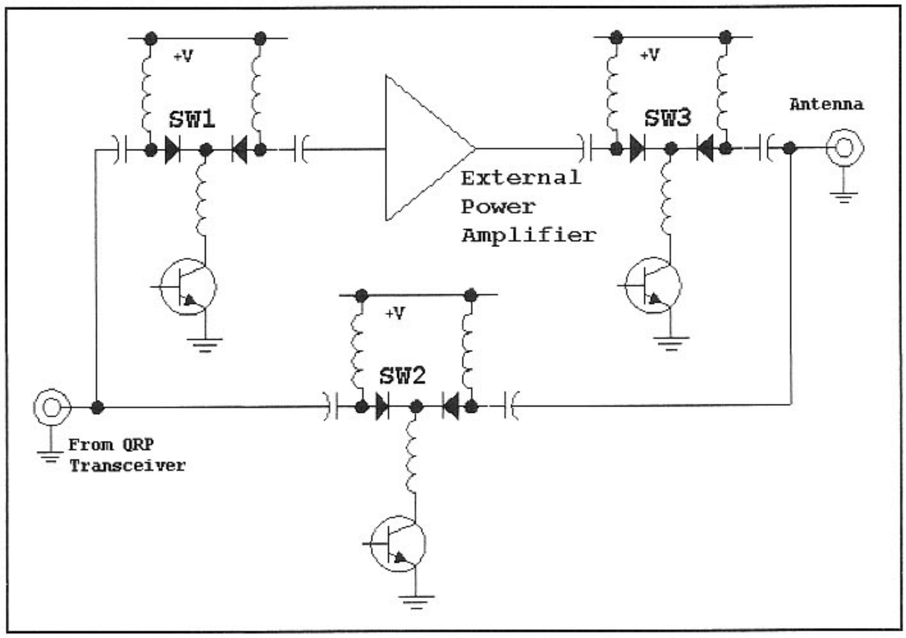


Fig 6.128—A T/R switch topology suitable for use following a low- power transceiver. We have not built this circuit.

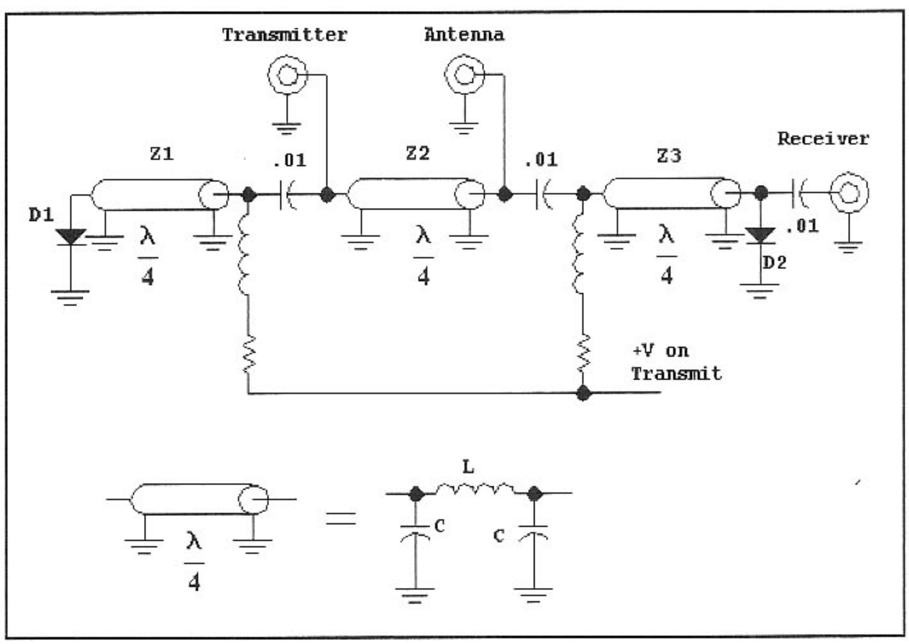


Fig 6.129—A T/R switch with shunt diodes using the impedance-reflection properties of quarter-wavelength transmission lines.

6.9 THE LICHEN TRANSCEIVER: A CASE STUDY

There are several suitable block diagrams for single sideband transceivers. The one we prefer shares only the oscillators, allowing receiver and transmitter optimization without compromise of interaction.[44] Although that scheme uses more parts, all basic functions are isolated with minimal interaction.

This transceiver, which is more efficient in its utilization of components, is an outgrowth of an architecture used by VE7QK in several versions of his *Epiphyte*.[45,46,47] This format, used in some early military SSB gear, shares many of the circuit elements between modes with signals flowing in the *same* direction in transmit and receive. The transceiver is presented here to illustrate design ideas and to present some of the steps needed to build such a transceiver.

Block diagram

The system with two mixers is shown in **Fig 6.130**. The first serves as the front end mixer during receive and as a transmit balanced modulator. The second is a receiver product detector and an IF-to-RF converter during transmit.

The original Epiphyte used NE602 mixers with no IF gain. The rig was intended for field use in the rugged mountains of the British Columbia Coast Range. The Lichen uses diode-ring mixers and includes IF gain. The 75-m-band Lichen can be adapted to many other bands.

The price of simplified signal flow is complex LO and carrier oscillator switching. The NE602 mixers used in the Epiphyte required little power in the 3-MHz LO, allowing switching with CMOS parts. The Lichen performs the switching with diodes, a scheme selected for compatibility with higher frequencies.

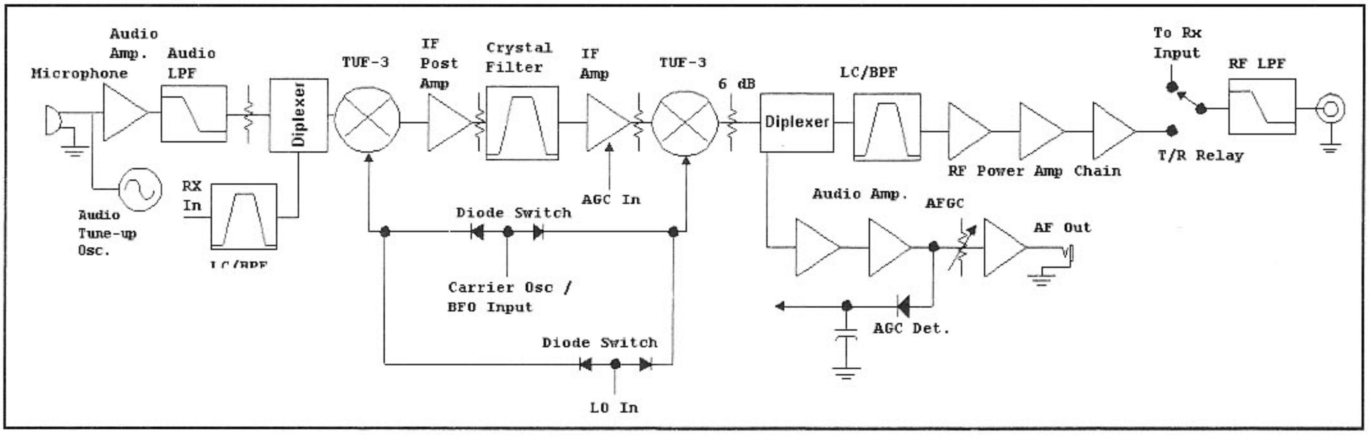


Fig 6.130—Block diagram for the Lichen transceiver.

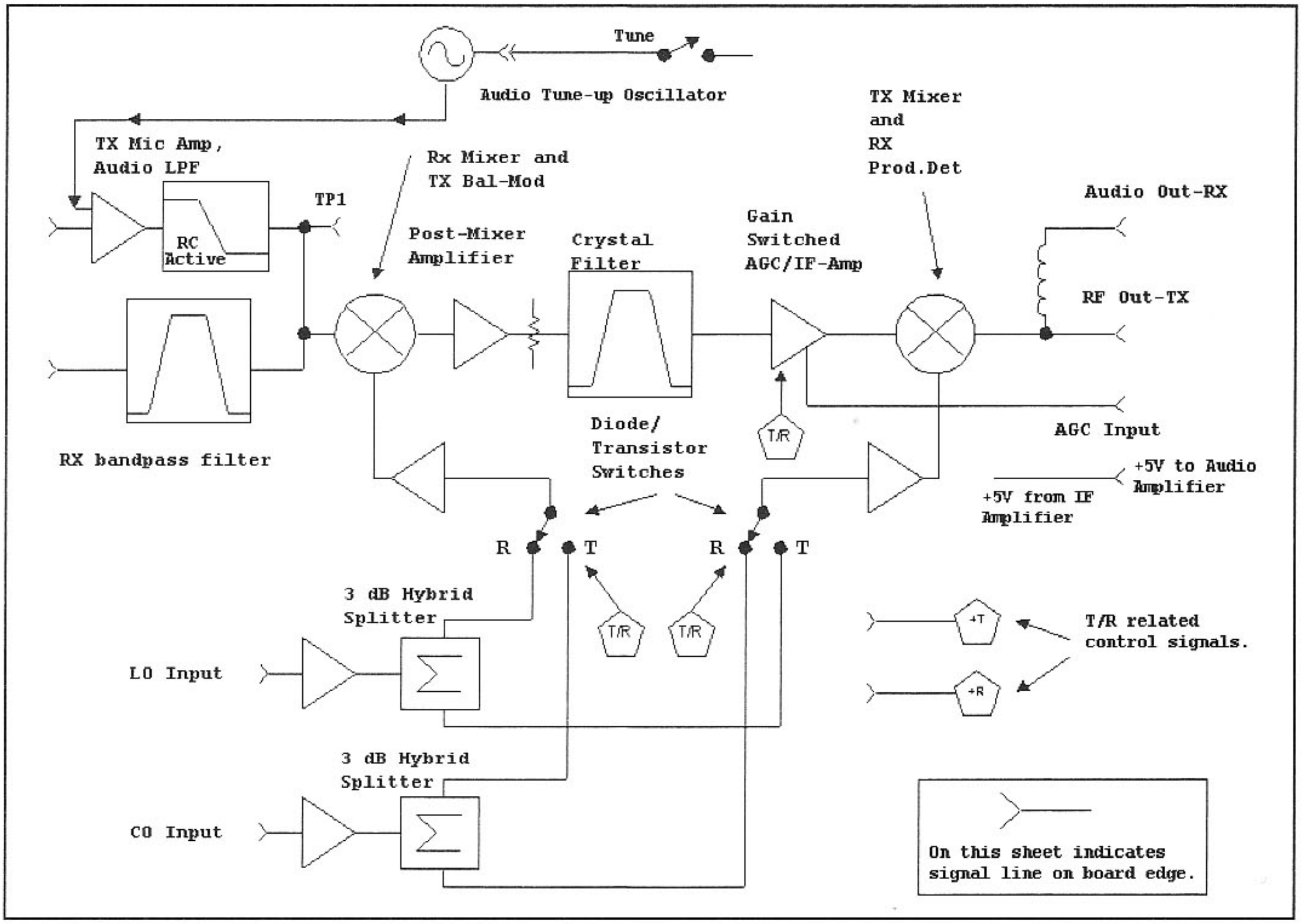
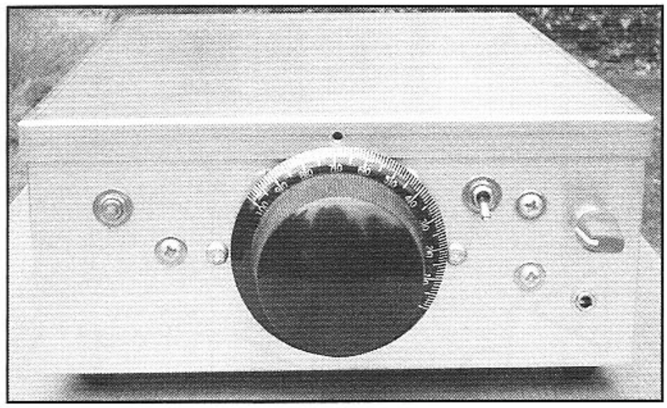


Fig 6.131—Block diagram for the transceiver main board.

Signal flow in the "Main Board"



Front panel view. Switch between tuning and audio gain is a sub-band switch. The push-button injects an audio tone for tuning.

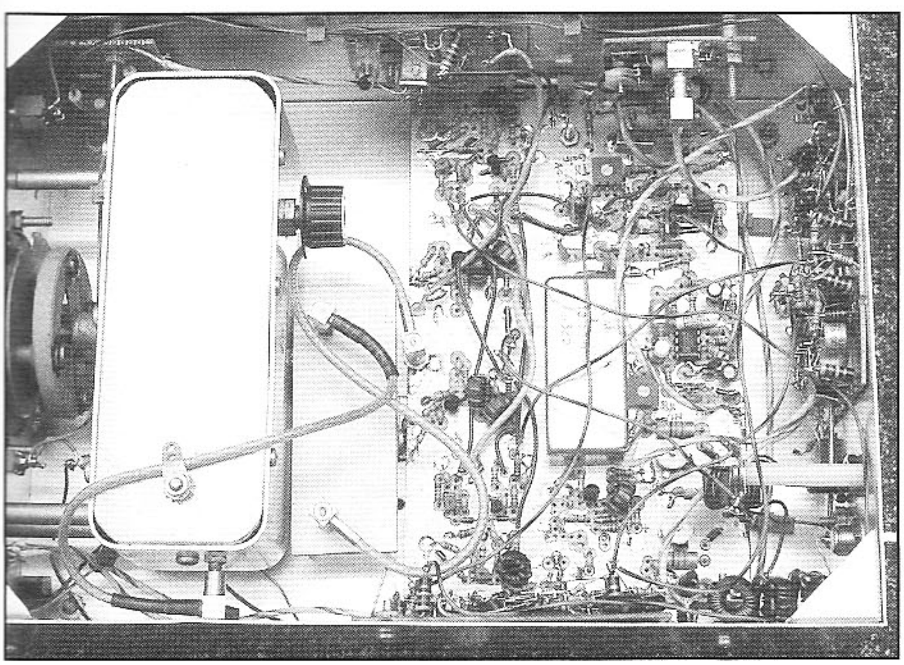
The transceiver is broken into several boards, a definite aid to the tedium of detailed measurements. The "main" board contains the receiver input preselector, a microphone amplifier, the two mixers, the IF system including crystal filter, and LO buffers and switching. The board includes an audio oscillator to facilitate testing. A block diagram is shown in **Fig 6.131**. The complete schematic is in **Fig 6.132**.

The main board begins at J2 where a signal enters the receiver input. ("J" numbers designate pads at the edge of a board.) The receiver preselector is a double tuned circuit using series resonators formed from molded RF chokes. The filter output is applied directly to the first mixer, U2. Bandpass filters for other bands are listed in **Fig 6.133**. The 160 and 80-m filters use Q_U = 50 RF-choke inductors while the higher bands use toroid inductors with Q_U = 200.

The microphone input is amplified and low pass filtered in U1. An RFC in the mixer line with capacitors in the receiver input filter form a diplexer to combine audio and receiver RF signals for the mixer. The microphone-amp is adjusted for a (lower than normal) signal of –20 dBm applied to the mixer.

The prototype transceiver used a commercial crystal filter while another (Fig 6.132) used a homemade 9.2-MHz crystal filter. The filter output drives a 2N3904 post mixer amplifier. *Post-amp* gain is 19 dB, reduced to 13 dB by the 6-dB pad, and has a 50-Ω input and output impedance. The sixth-order crystal filter is designed using the methods presented in Chapter 3.

An L-network (L4, C36) transforms the post-amp 50 Ω to the needed filter source impedance. Transformer T3 matches the relatively low filter impedance to the 2.2-kΩ input resistance of the following IF amplifier. T3 uses a 61-material ferrite core to keep the loss low. The filter should be built and measured before incorporation in the transceiver. The exact –6-dB filter frequencies should be recorded for later use. The designer/builder will have to design matching networks and transformers as well as the crystal filter.



Top view showing LO module with "main board" to the right. The small box built from scrap circuit board material contains the 14-MHz-LO bandpass filter.

JFETs Q6 and Q8 provide IF gain. These stages are gain switched by Q7 and Q9 with higher gain during receive. Reasonable IMD performance is vital, for the amplifier is in the transmit signal path. This system (Q6 and Q8) has a small signal receive gain of 27 dB with 70 to 80 dB of available gain reduction. Gain drops to 12 dB in transmit. IMD performance is good at OIP3 = +18.5 dBm, dropping to +14 dBm in transmit mode. IMD degrades with gain reduction, but the intercepts do not degrade as fast as the gain, a requirement to preserve output cleanliness. Receiver AGC is disconnected during transmit; R58 is switched in to establish a transmit level.

Transmit mixer, U3, should see maximum drive of –10 dBm for a spur free output, as discussed earlier. The post-amp, Q5, including pad has a gain of 13 dB while typical crystal filter loss is 4 dB. With a balanced modulator input of –20 dBm, the signal at the input to T3, just past the crystal filter, is –17 dBm. Transmit gain of 12 dB in the IF brings the level at U3 to –5 dBm. A slight IF gain reduction and a 3 dB pad in the IF output sets the –10 dBm level. If the balanced modulator had been driven at its nominal level of –10, the IF would be overdriven, resulting in overdrive for the second mixer.

This gain distribution degrades carrier suppression to 30 dB. If the post-amp gain could be reduced by 10 dB during transmit, the carrier suppression would be improved by a like amount.

With a U3 mixer drive of –10 dBm, the 6 dB conversion loss produces an output of –16 dBm. A 6-dB pad after the mixer and a bandpass filter (described later) with a 2-dB loss produce an eventual output of –24 dBm, established by R58.

The audio tune-up oscillator included in Fig 6.132 can be used during normal operation to generate a carrier for transmatch tuning. It is also available for

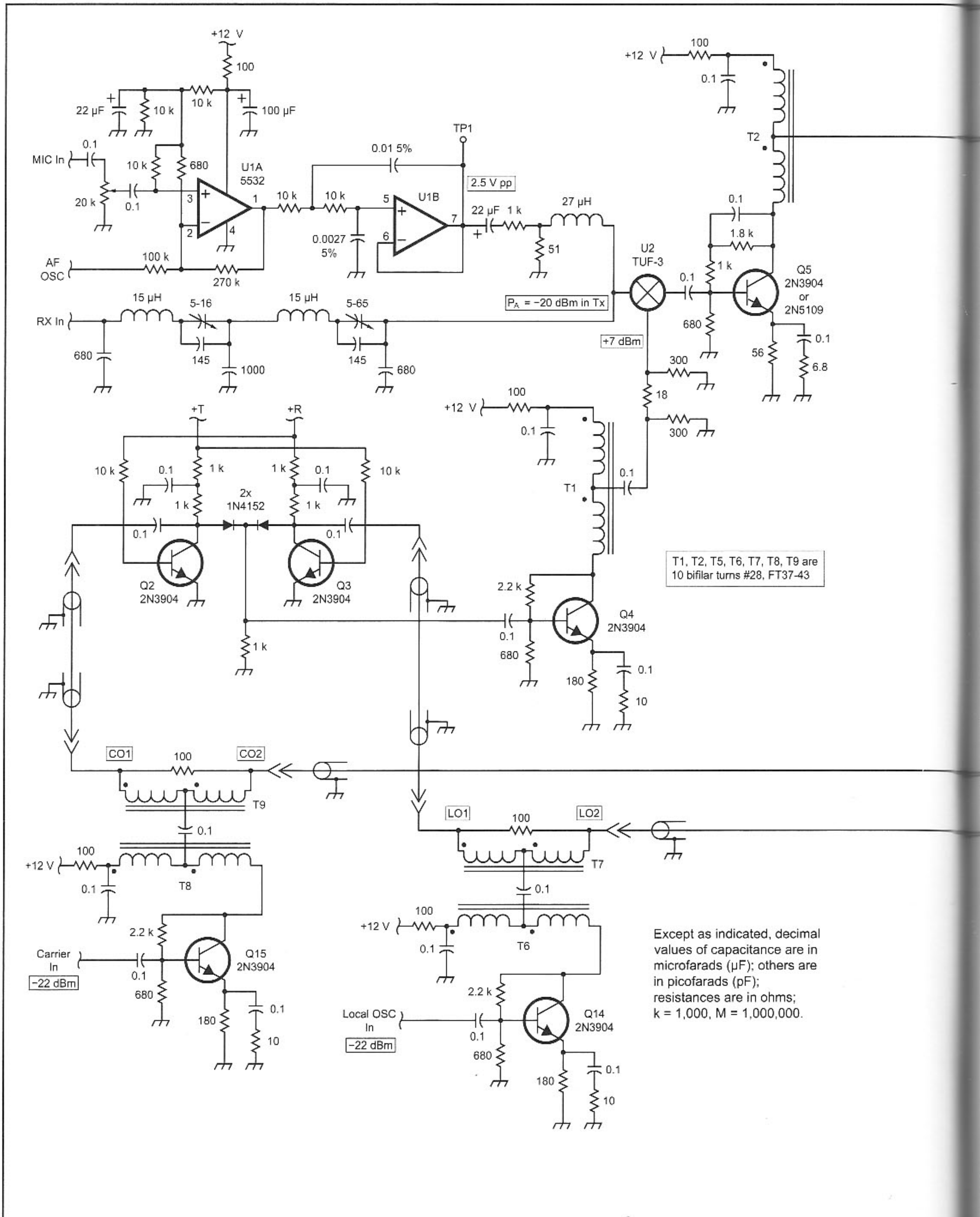
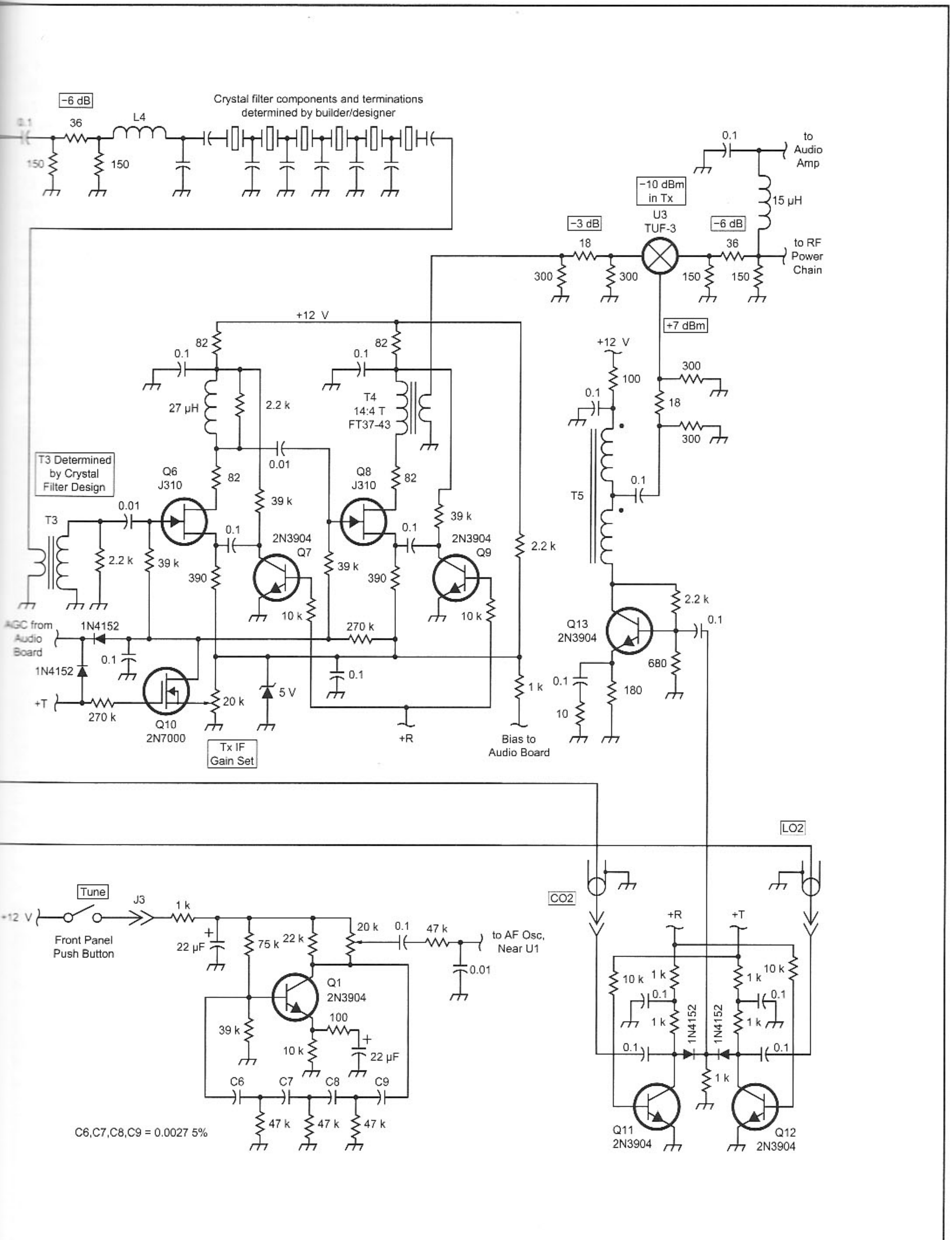
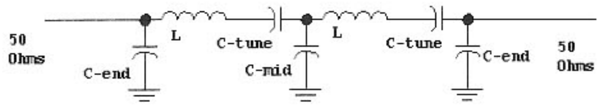


Fig 6.132—Schematic for the main board. See text for detailed discussion.



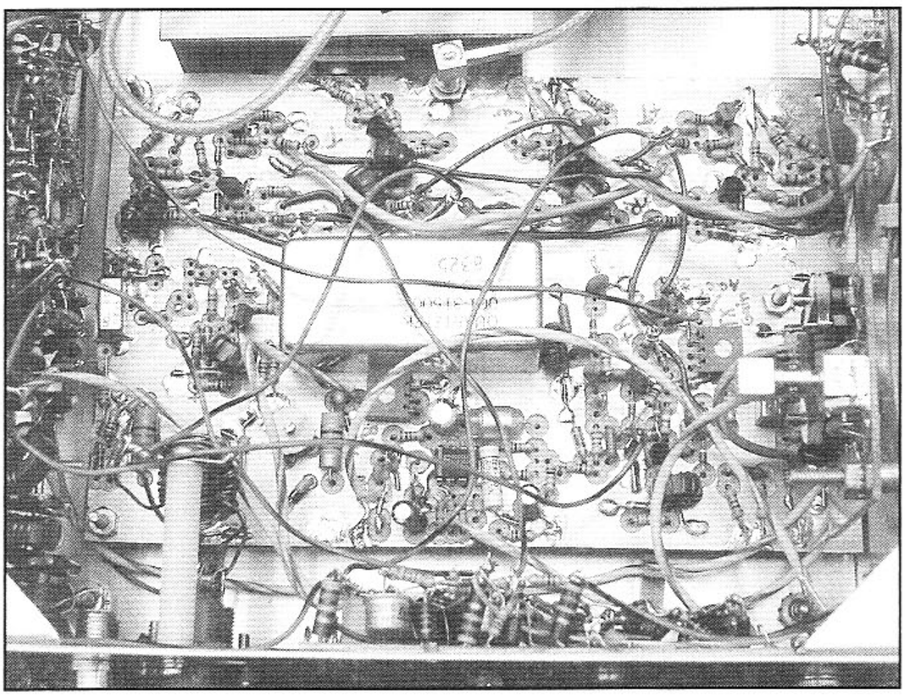
−6 dB
Crystal filter components and terminations determined by builder/designer
36
L4
150
150
0.1
to Audio Amp
−10 dBm in Tx
15 µH
−3 dB
U3
TUF-3
−6 dB
18
36
to RF Power Chain
300
300
150
150
+12 V
+7 dBm
82
82
0.1
0.1
+12 V
300
100
27 µH
2.2 k
T4
14:4 T
FT37-43
0.1
18
300
0.01
T3 Determined by Crystal Filter Design
Q6
J310
82
Q8
J310
82
0.1
T5
39 k
39 k
0.01
T3
0.1
2N3904
Q7
0.1
2N3904
Q9
2.2 k
2.2 k
39 k
39 k
390
390
2.2 k
10 k
10 k
Q13
2N3904
0.1
AGC from Audio Board
1N4152
270 k
0.1
680
0.1
1N4152
0.1
5 V
1 k
0.1
180
+T
270 k
20 k
10
Q10
2N7000
+R
Bias to Audio Board
Tx IF Gain Set
LO2
CO2
Tune
J3
+12 V
1 k
+R
+T
Front Panel Push Button
22 µF
75 k
22 k
20 k
0.1
47 k
to AF Osc, Near U1
0.01
10 k
1 k
1 k
10 k
Q1
2N3904
0.1
0.1
39 k
100
1 k
1N4152
1N4152
1 k
10 k
22 µF
0.1
0.1
C6
C7
C8
C9
1 k
47 k
47 k
47 k
C6,C7,C8,C9 = 0.0027 5%
Q11
2N3904
Q12
2N3904

testing during board development. The microphone is attached at the amplifier input, J1, and the level at test point TP1 is observed. Audio gain (R1) is adjusted for 2.5 V peak-to-peak at TP1 on voice peaks with a normal voice into the microphone. The tune-up oscillator level, R14, is then set to produce the same level.



Freq., MHz	BW, MHz	C-end, pF	C-mid, pF	C-tune, pF	L, uH	Q-u	IL, dB
1.9	0.3	2100	2300	321	27	50	2.5
3.75	0.6	680	1000	145	15	50	2.4
7.15	0.4	1000	1750	78	7	200	1.7
10.1	0.3	1200	2300	53	5	200	3.5
14.2	0.5	680	1200	34	4	200	2.8
18.1	0.4	680	1200	20	4	200	5.1
21.3	0.6	510	1000	20	3	200	3.7
28.4	1.0	270	390	11	3	200	2.8

Fig 6.133—Receiver bandpass filters using series resonators.



Close up of the main board.

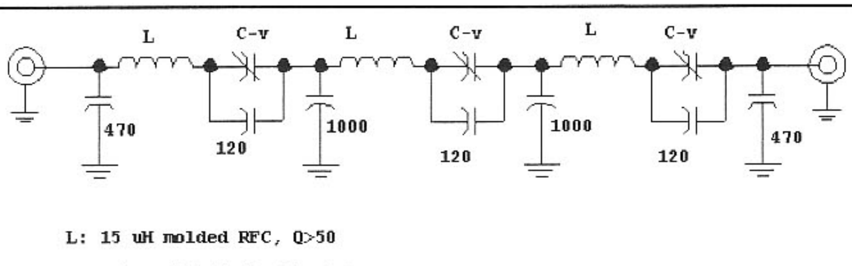


Fig 6.134—Triple-tuned 3.5 to 4-MHz bandpass filter for the output of the transmit mixer.

Mixer Injection Switching

The 10-MHz IF version uses a 13.5 to 14-MHz LO and a 10-MHz carrier oscillator (CO.) The LO must be applied to U2 in receive while the CO drives U3. Roles are then reversed in transmit with the LO driving U3 and the CO driving U2.

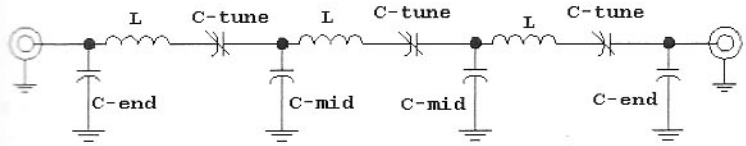
Each ring mixer requires nominal LO power of +7 dBm. But lower power levels are switched. Drive amplifiers Q4 and Q13 reduce the switched power to –9 dBm, easily controlled with normal silicon diodes biased for modest current. Diodes D1 and D2 switch the signals going to U2 while D6 and D7 route energy to U3. These switches are controlled by signals labeled with T or R, indicating positive bias on either transmit or receive. These signals, appearing often throughout the transceiver, are generated on the RF power amplifier board. The diode switches route a desired signal to an intended load, but do not present as much attenuation of the *off* path as we would like. Shunt transistor switches Q2, Q3, Q11, and Q12 were added to provide about 50 dB reduction in the *off* paths.

Although the shunt transistor switches improve performance, they add a complication: Each input (LO and CO) is amplified and buffered in an amplifier, Q14 and Q15. If those amplifier outputs were routed directly to the composite diode/transistor switches, they would always be short circuited. Isolation results from transformers T7 and T9 which function as a splitter-combiner, described in Chapter 3. These switching methods can be extended to UHF. LO and CO signals are required at the board inputs with a power of –22 dBm.

The circuit board contains short lengths of coaxial cable to route the LO and CO signals. The two LO components, LO1 and LO2, move respectively from J19 to J5 and from J20 to J14 on cable. The CO signals CO1 and CO2 move respectively from J16 to J4 and J17 to J13. The best place to measure LO chain power is just before the mixers. Lift C29 or C59 at the pad ends and measure the power coming from the LO system. Those powers should both be close to +10 dBm. The LO amplifiers use 2N3904s, but the less robust MPS3904 is *not* suitable. The MPSH10 (Fairchild and Philips) is also an excellent choice.

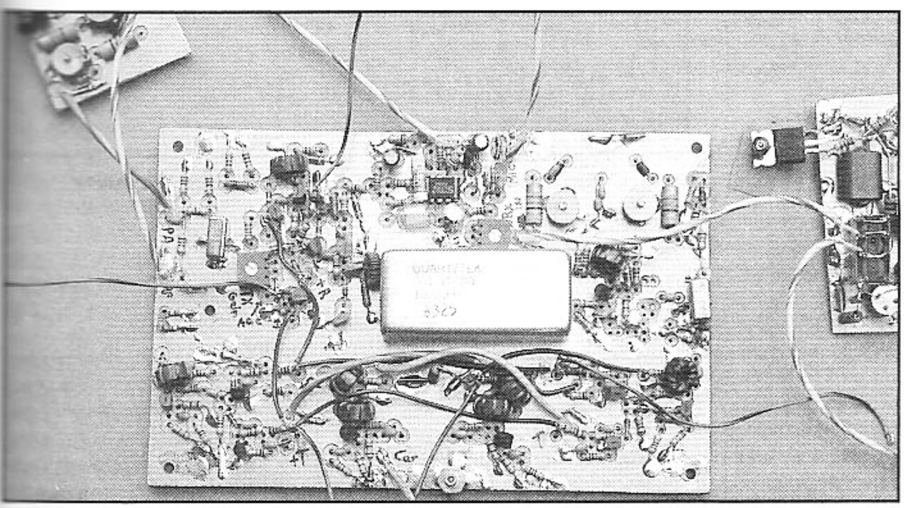
Transmit Bandpass Filter

The Main board RF output at 3.5 to 4 MHz has a 23.4 to 24 MHz image. The lower range is selected with the filter shown in **Fig 6.134**. This circuit is best



Center Freq.	B.W.	C-end	C-mid	C-tune	L	Q_U	I.L.
MHz	MHz	pF	pF	pF	µH		dB
1.9	0.22	2200	3300	307	27	50	3.6
3.75	0.7	470	1000	148	15	50	2.1
7.15	0.4	820	1750	78	7	200	1.7
14.2	0.55	500	1200	34	4	200	2.5
21.2	0.65	390	820	20	3	200	3.2
28.4	1.1	180	390	11	3	200	2.4

Fig 6.135—Triple-tuned bandpass filters for several HF bands. The required unloaded Q (vital) is also given.



Main board removed from cabinet. Circuitry below crystal filter is for the LO and carrier oscillator buffers and switches. Upper right corner contains RF Input bandpass filter.

assembled and tested in a 50-Ω environment prior to use in the transmitter. A table of computer generated values is given in **Fig 6.135** for several additional bands.

The Local Oscillator

The LO tunes from 13.5 to 14 MHz with the heterodyne system of **Fig 6.136**. Q402 is a 2.5 to 3-MHz Colpitts oscillator buffered with a common-base amplifier, Q405. Output is kept low, for only –10 dBm is needed by diode ring mixer U402. The output is established with the pad driving the RF port. This level, and that at the mixer LO port should be measured during construction.

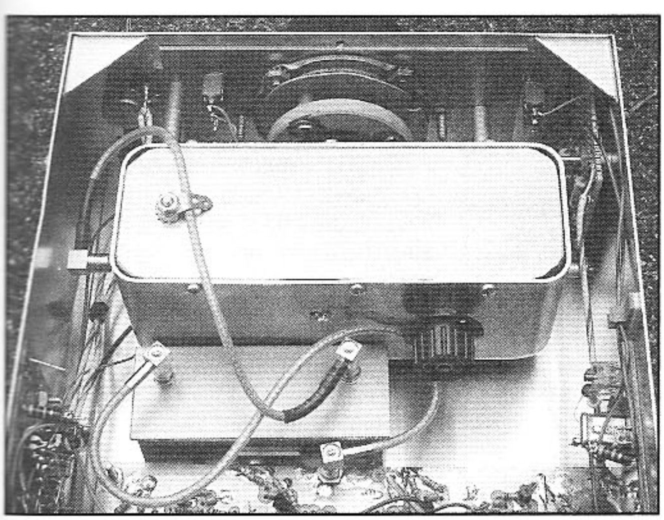
A 365-pF variable capacitor tunes only half of the range. The other half is tuned by switching in an additional capacitor, C402. The switching is performed with a pair of PIN diodes, D401 and D402. When a positive voltage is applied to J401, Q401 is saturated, causing both PIN diodes to conduct.

A crystal controlled 11-MHz oscillator provides the drive for the diode-ring mixer. The two oscillators are both placed inside the shielded LO enclosure, along with the ring mixer. The output is then routed through coaxial cable to a triple-tuned LC bandpass filter, **Fig 6.137**.

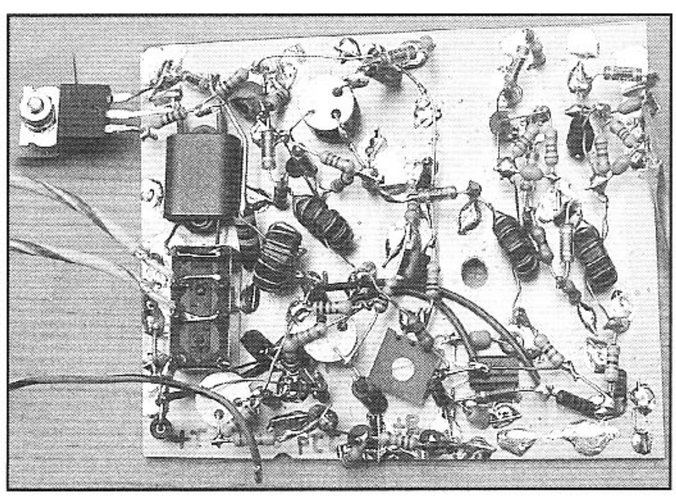
A change in IF from 10.0 MHz will result in the need for a new LO frequency on the part of the designer/builder.

The Carrier Oscillator

A carrier oscillator (CO) drives the balanced modulator in transmit and the BFO in receive. The CO must have the same –22 dBm level as the LO when applied to the Main board. The CO circuit is shown in



View of LO.



RF Power chain. The HEX-FET PA is normally attached to the cabinet that serves as a heat sink.

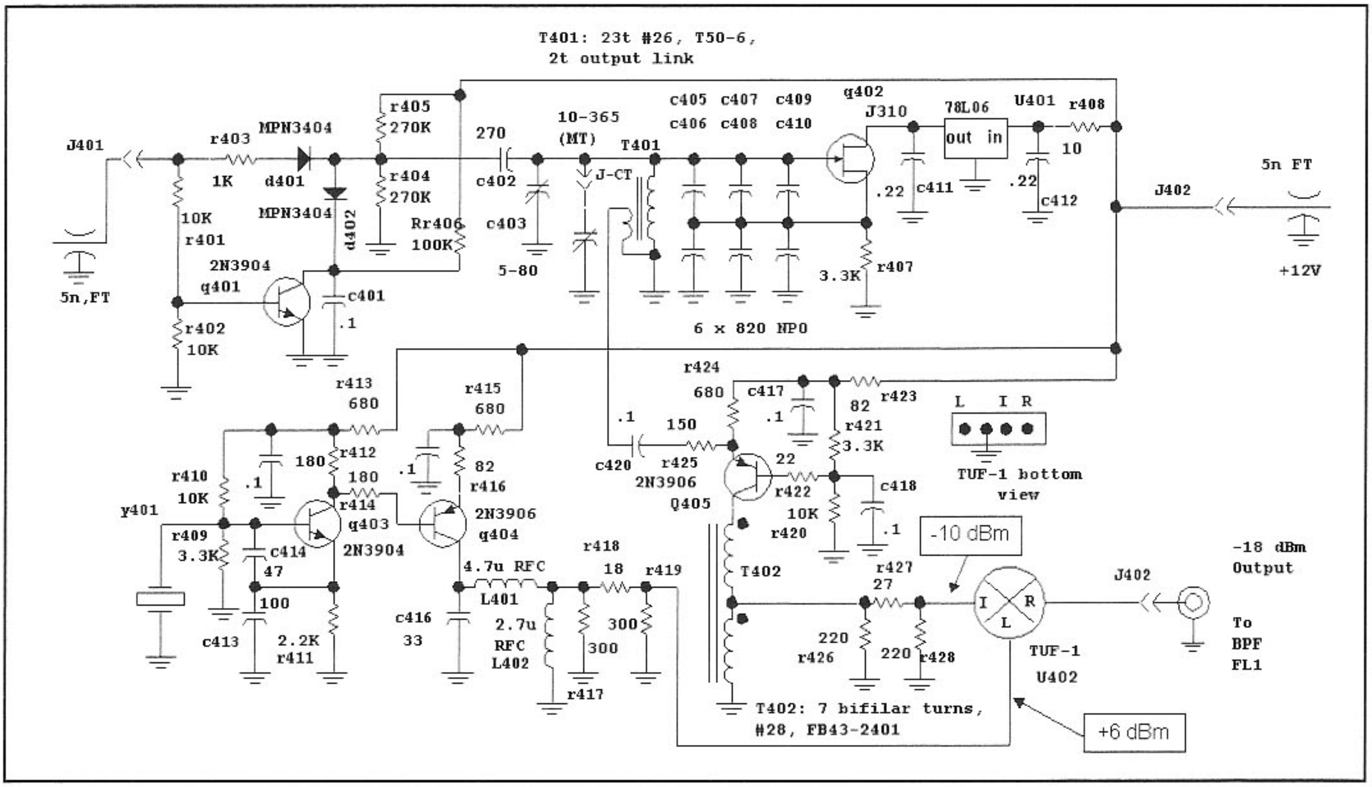


Fig 6.136—Transceiver LO system produces output at 13.5 to 14 MHz. The bandpass circuit of Fig 6.137 filters the mixer output.

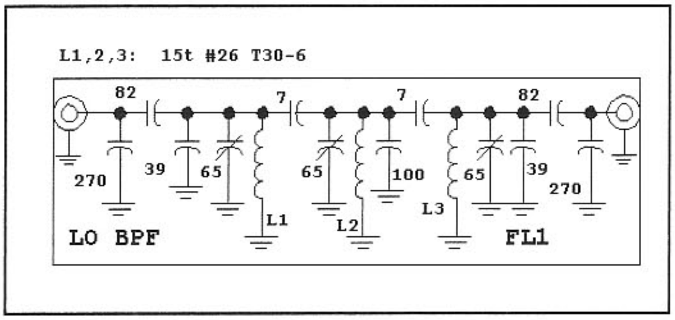


Fig 6.137—LO bandpass filter.

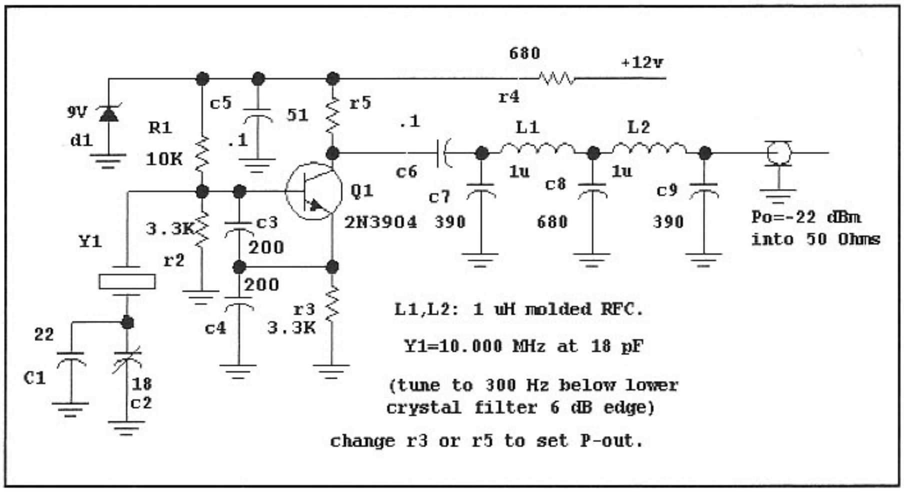


Fig 6.138—Carrier oscillator.

Fig 6.138. The output power is set at –22 dBm by adjustment of R5 in the oscillator collector. The power supply is regulated more as a means to stabilize amplitude than frequency.

We measured the crystal-filter response during circuit development. Knowing the exact lower 6 dB passband edge, we placed the carrier oscillator at a frequency 300 Hz below that edge. The resulting 10-MHz USB signal is inverted to become a LSB output at 3.8 MHz. Slight frequency adjustment may be done to optimize signals.

The Receiver Audio System

Fig 6.139 shows the audio system. The product detector output reaches the board via coaxial cable where it is amplified by Q301 and Q303, and applied to an off board audio gain control. The result is then amplified in two op-amp stages, U301, and applied to headphones.

The signal at the gain control is sampled and routed to op-amps U302 for full wave rectification. This charges the AGC sampling capacitor, C315, a 1 µF stacked metal film type (Panasonic V-series or similar.) R325 controls attack time while R324 sets recovery. U303A is a follower to drive the IF system with dc. Normal audio muting is not required. AGC was disconnected from

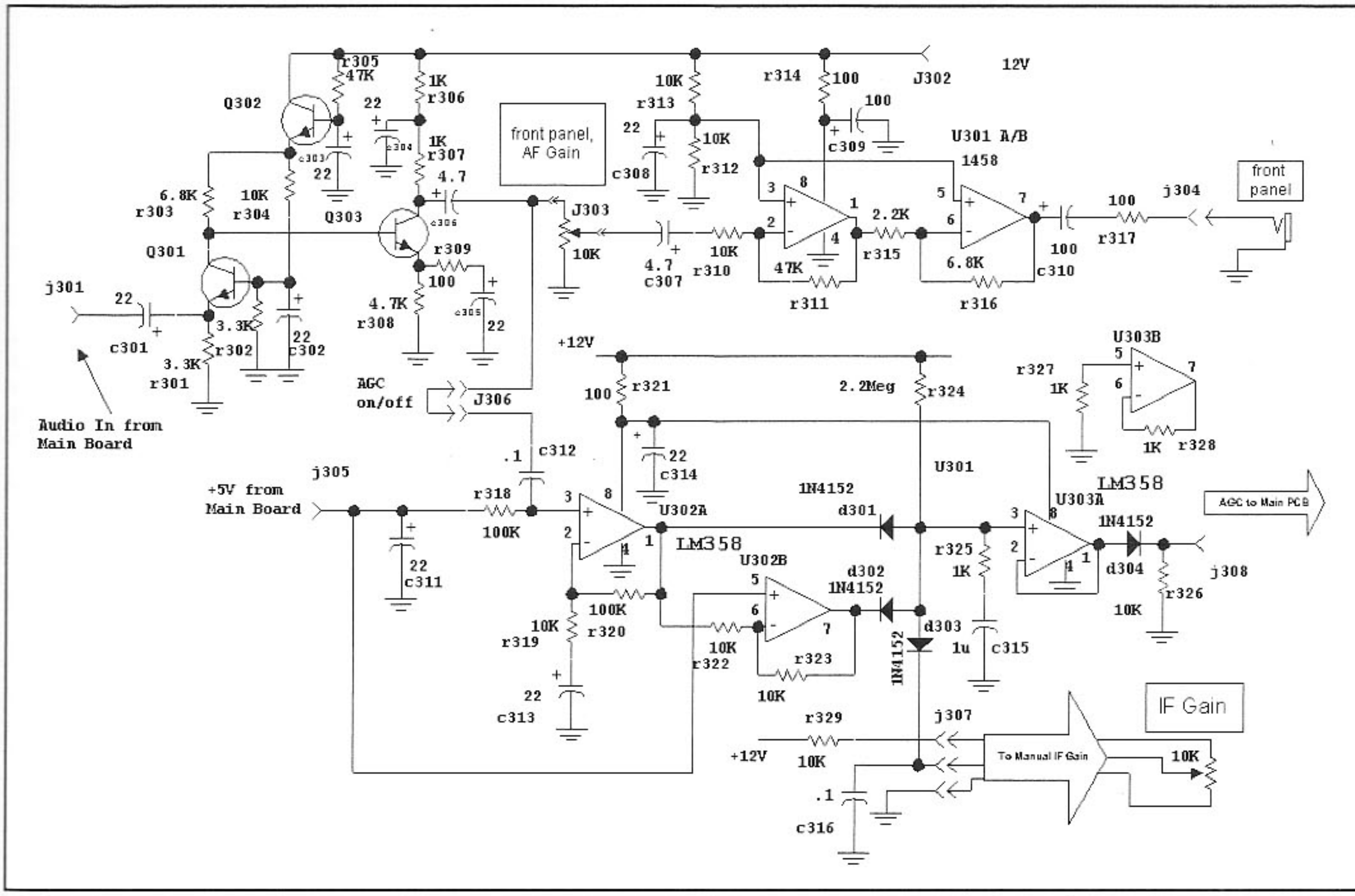
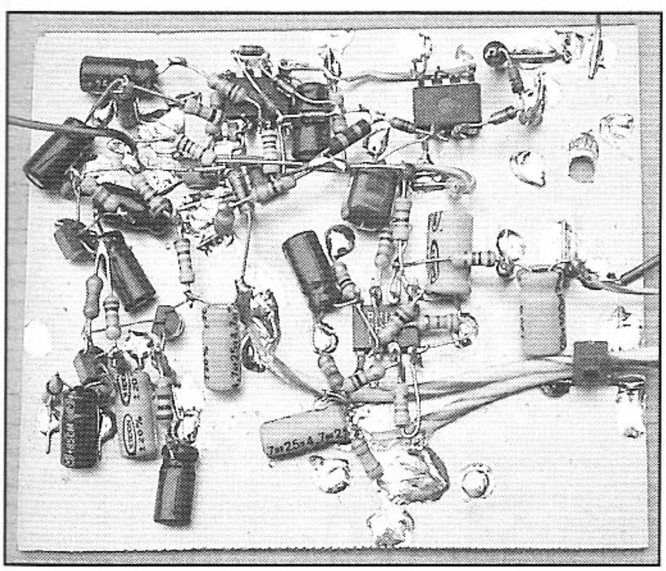
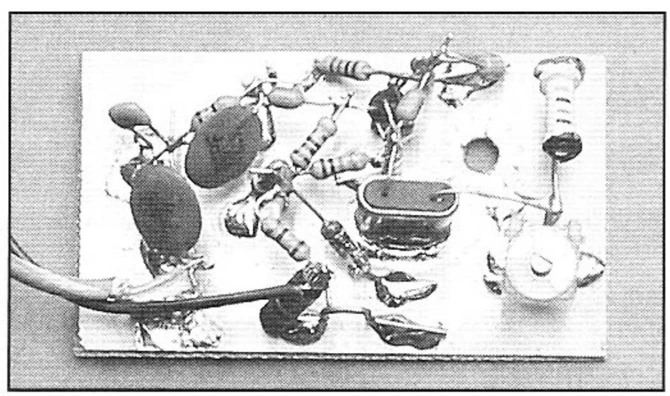


Fig 6.139—Audio system and AGC detector.



Audio Amplifier.



Carrier Oscillator.

the IF during transmit with D4, D8, and Q10 on the Main board.

The RF Power Chain

A four-stage RF power chain, **Fig 6.140**, completes the transceiver. Three bipolar transistors drive a HEXFET PA for a 5-W output.

The first two stages use a 2N3904 while the third uses a 2N3866 with a small heat sink. The three are respectively biased at 10, 17 and 50 mA. A 6-dB pad is placed after the first stage, providing a convenient place to alter gain for use on other bands. **Fig 6.141** shows gain vs frequency for the three stage bipolar driver. Although gain is dropping, the driver chain is useful through the entire HF spectrum. We realized another 3-dB gain at 50 MHz when Q101 and Q102 were changed to MPSH10s. IMD was measured at 14 MHz for the driver chain, producing OIP3 = +39 dBm with either transistor type in the first two stages. The nominal output for Q103 is +10 dBm per tone with a two-tone test, or +16 dBm (40 mW) PEP.

The PA, an IRF-510 HEXFET, is biased

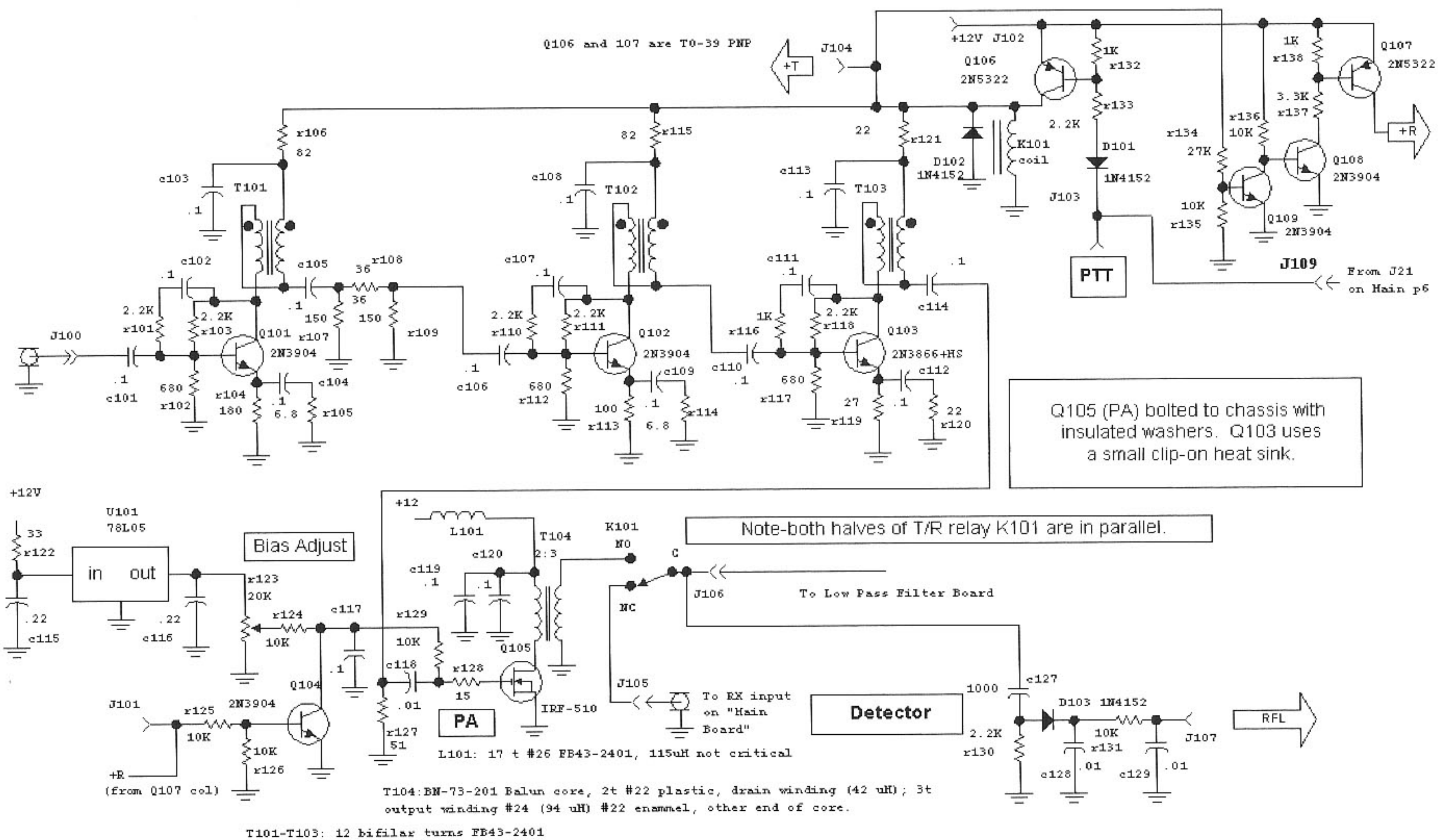


Fig 6.140—RF driver chain for the Lichen transceiver uses 4 stages for an output of 5 W. The T/R relay is a Nais DS2Y-S-DC12V or similar.

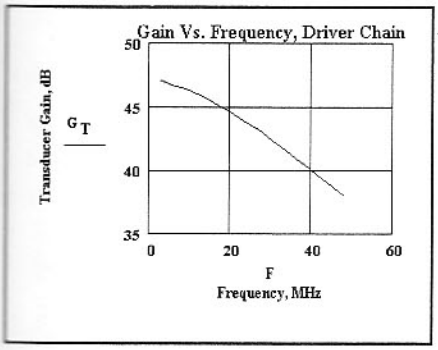


Fig 6.141—Small signal gain vs frequency for the three-stage bipolar driver chain.

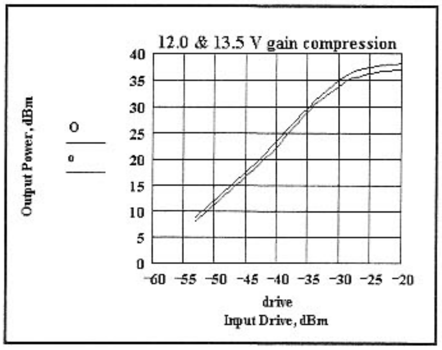


Fig 6.143—Gain compression measurement for complete RF power chain.

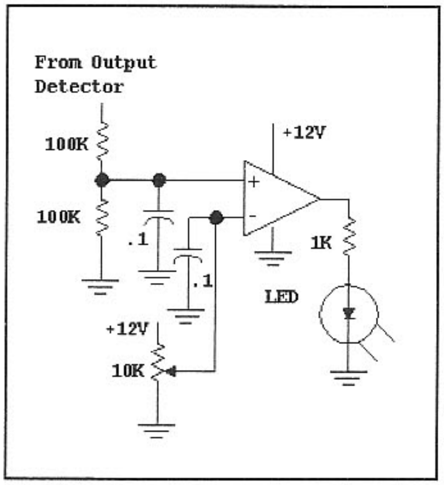


Fig 6.144—LED driver circuit that can be driven by the output peak detector. Op-amp is a 741, 1458, LM358, LM324, or similar part.

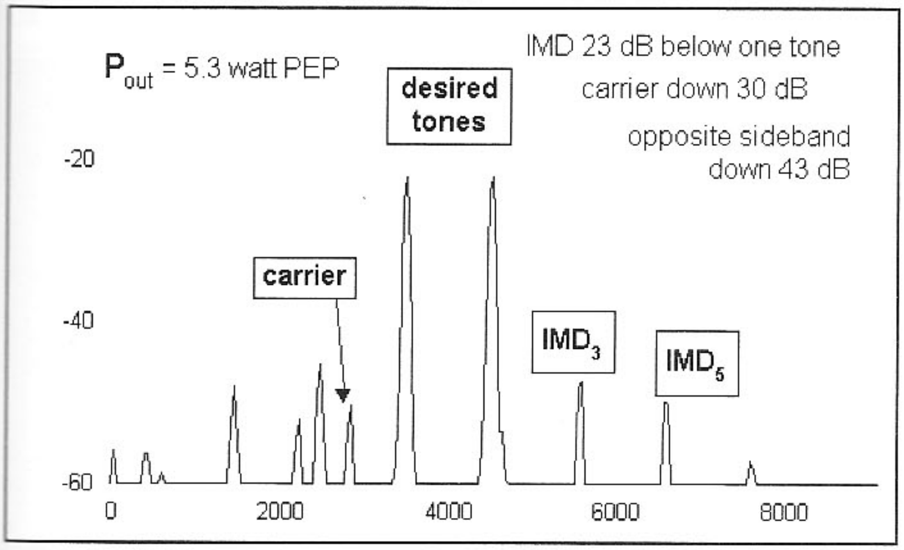
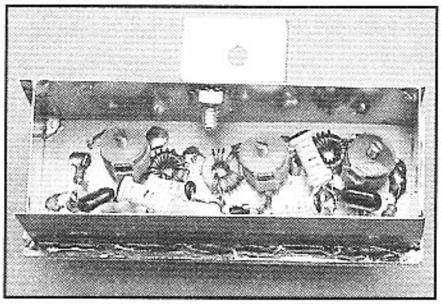


Fig 6.142—Spectrum analyzer view of transmitter output under two-tone testing. For software, see **www.monumental.com/rshorne/gramdl.html**.



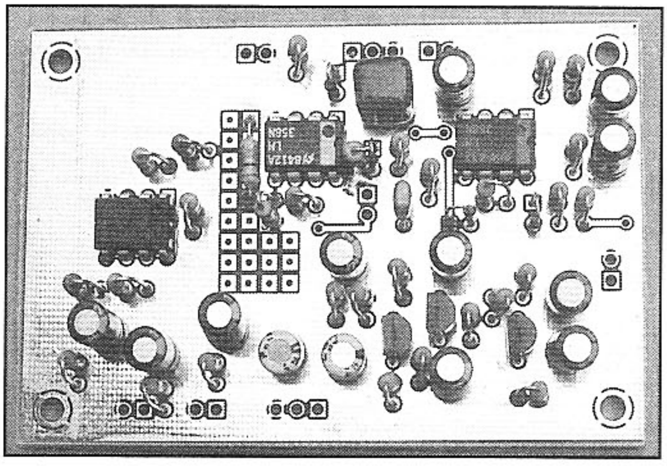
A view of the 14-MHz bandpass filter used for LO in transceiver.

from a pot driven by U101, a 78L05. Bias current with no drive is set for about 40 mA, a level producing excellent gain and distortion acceptable for QRP efforts. Transmitter output is shown in the two-tone test spectrum of **Fig 6.142**. This was obtained with a FFT spectrum analysis program, Spectrogram, running on a laptop computer, augmented with a converter. (See spectrum analysis discussion in Chapter 7.) Third order IMD is only 23 dB down from each tone, or 29 dB below PEP. The 30-dB carrier suppression is also shown. Opposite sideband suppression was 43 dB for a 1700-Hz single audio tone. Earlier driver chain measurements confirm the FET PA as the distortion source.

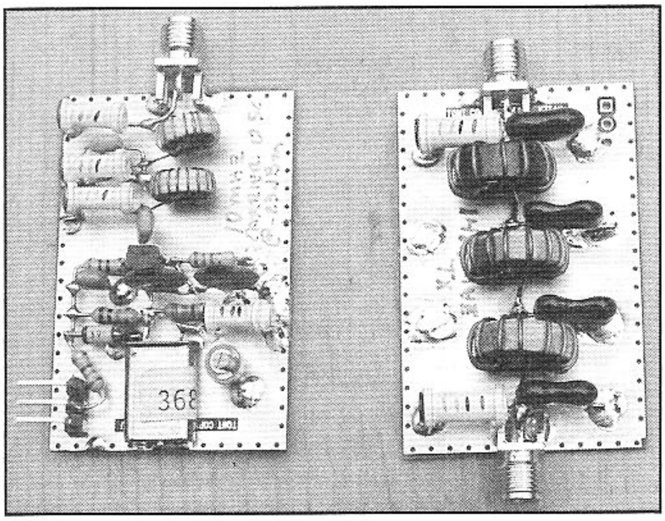
Fig 6.143 shows power chain output power as a function of drive power. This gain compression measurement was done with single-tone drive. The amplifier is relatively linear up to the +33 to +35 dBm output. This is a measurement that can be performed in the home lab that has yet to include a spectrum analyzer.

A peak detector is included at J107, useful during transmitter setup. It can also be used to drive a front-panel LED through a circuit like that shown in **Fig 6.144** where an op-amp serves as a comparator. Alternatively, the detector could drive an auto level control (ALC) circuit to provide negative feedback to the IF.

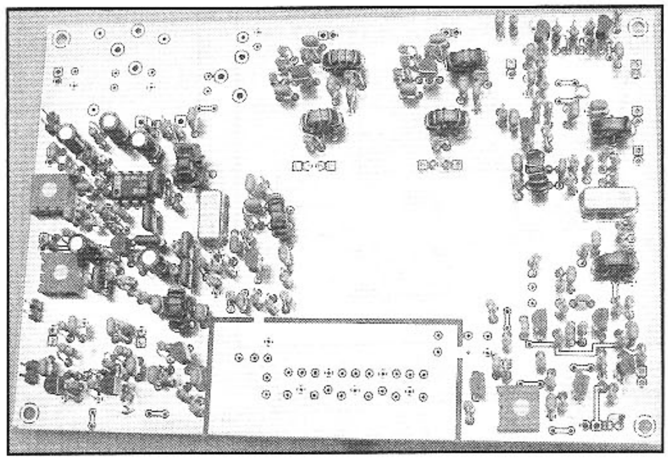
An IF speech processor was described in an earlier section where limiting within the IF constrained the output level. That scheme had the added advantage of preventing excessive levels in the transmit mixer and following amplifiers, eliminat-



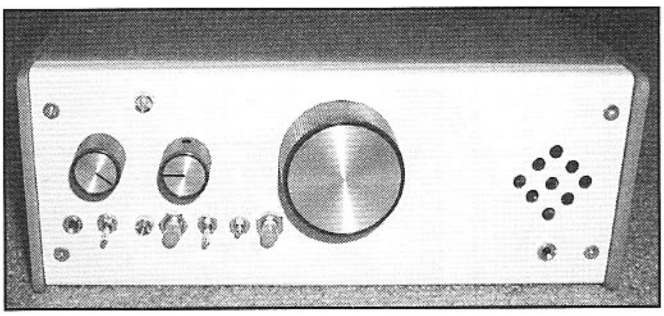
Printed circuit audio amplifier. (TNX to K7TAU)



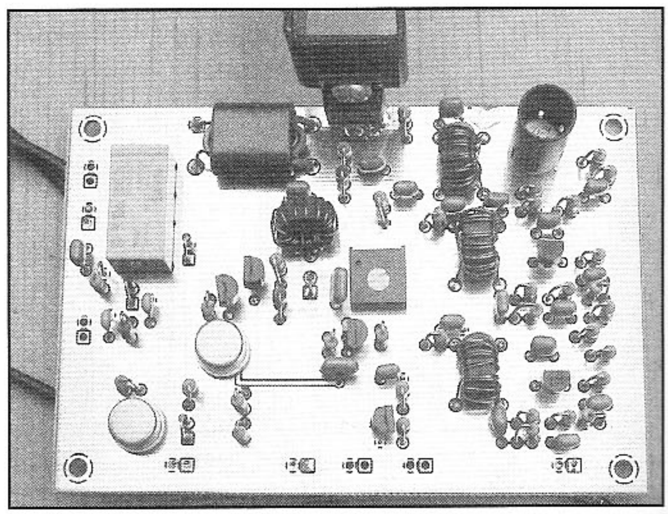
Breadboarded carrier oscillator and TX low-pass filter for a 14-MHz version of the transceiver by K7TAU.



Partially built printed main board.



Front panel of 75-meter version built by AA7QU. One of the buttons activates a "Freq-Mite" frequency keyer that then reads the frequency and presents it in morse code.



Printed Circuit Version of RF Power Chain. (TNX to K7TAU)

ing the need for ALC. The IF limiter has the minor disadvantage of requiring another crystal filter. However, it would be a dramatic virtue in this transceiver. Not only would it enhance transmitter performance, but it would generate excellent receiver skirt selectivity.

A seventh-order low pass filter follows the FET power amplifier, as shown in **Fig 6.145**. The filter is built on a separate board, isolated from the rest of the PA.

Control Circuits

The transceiver uses push-to-talk (PTT) operation, realized with the control circuitry included in Fig 6.140. When the microphone PTT button is pushed, a line goes low at J103 to saturate PNP switch Q106. That transistor powers antenna relay, K1, and feeds a +12V-T signal to the many places in the transceiver marked with "T." Q107, 108, and 109 then provide a similar +12V-R to control the receive function. PA bias is shorted with Q104 during receive periods. Both sections of the DIP antenna relay are paralleled for the T/R switching.

Extensions and Results

Once the boards are built and measured, they can be assembled and combined.

The system using a 10-MHz IF is reasonably clean with the second harmonic at –57 dBc as the dominant spur. Three nonharmonic spurs were found with strength from –67 to –62 dBc. A 9.2-MHz IF version (built by AA7QU) had similar performance. We were disappointed in the IMD performance offered by the HEXFET PA.

Receiver performance was adequate for the 75-m band. The relatively high noise figure of 18 dB is not a problem for this frequency. Measured IIP3 was +16 dBm and two-tone DR was 92.7 dB. The dynamic window is skewed to favor high intercept rather than low noise. A low-

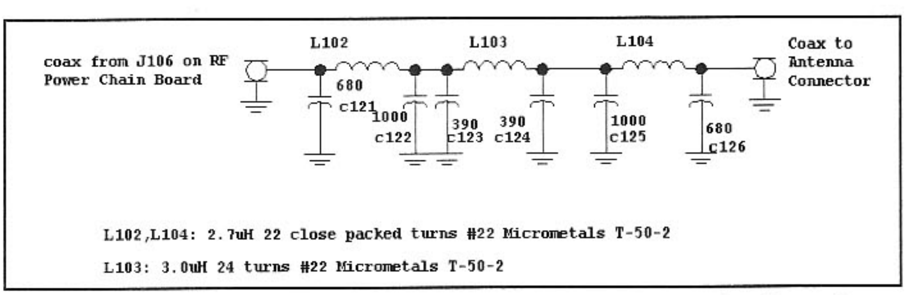


Fig 6.145—Low-pass filter for the 75-meter Lichen. Capacitors can be silver mica or ceramic.

noise RF amplifier with modest gain would substantially improve noise figure with little DR penalty, making this general topology useful at higher frequency.

Several boards were used in favor of a few, allowing the designer/builder to measure those parameters so critical to success. If the Main board was built without the input preselector filter, it would contain no band-specific components. The RF power chain and audio board are also band-independent, suggesting a multi-band design. Relay switching is recommended in the receiver front-end over PIN diodes to avoid second-order distortion problems.

6.10 A MONOBAND SSB/CW TRANSCEIVER

Although this transceiver was designed for operation on any single band within the HF spectrum, there is no fundamental reason it will not also function at VHF. Like the Lichen presented earlier, it is based upon homebrew crystal filters fabricated by the designer/builder.

This radio was designed for flexibility and performance. A common local oscillator system and common BFO/Carrier Oscillator are shared between the transmit and receive functions. The other functions are independent, allowing each to be optimized to meet the needs of the designer/builder/user. This seemingly inefficient approach becomes practical and inexpensive when one builds his or her own crystal filters. Although more extensive, the project is often less tedious than other sideband transceivers, for the receiver can be finished and made operational before dealing with the transmitter.

A collection of small circuit boards was used. Some were etched while others were merely breadboarded. The use of many small boards rather than just a few large ones provides improved isolation between functions and enhanced testability. A transceiver block diagram is shown in **Fig 6.146**.

The block diagram includes some shaded areas where circuit modules already presented are applied. The receiver begins with the "General Purpose Monoband Receiver Front-End" of Fig 6.68. That board includes a crystal ladder filter with up to 6 resonators. The next block is an IF amplifier. The recommended design here is that presented in Fig 6.50 using cascode connected J310 JFETs. Designs using some of the more up-to-date integrated circuits from Analog Devices should also be considered. Neither the front-end nor the IF will be discussed here.

The RF power chain is also shaded in the block diagram of Fig 6.146. A similar module developed for the Lichen transceiver would be suitable. Substitution of a different PA is recommended if the system is built for bands at the high end of the HF range, or for VHF. The poor IMD performance of the IRF510 would also be justification for a new PA design.

The monoband transceiver version

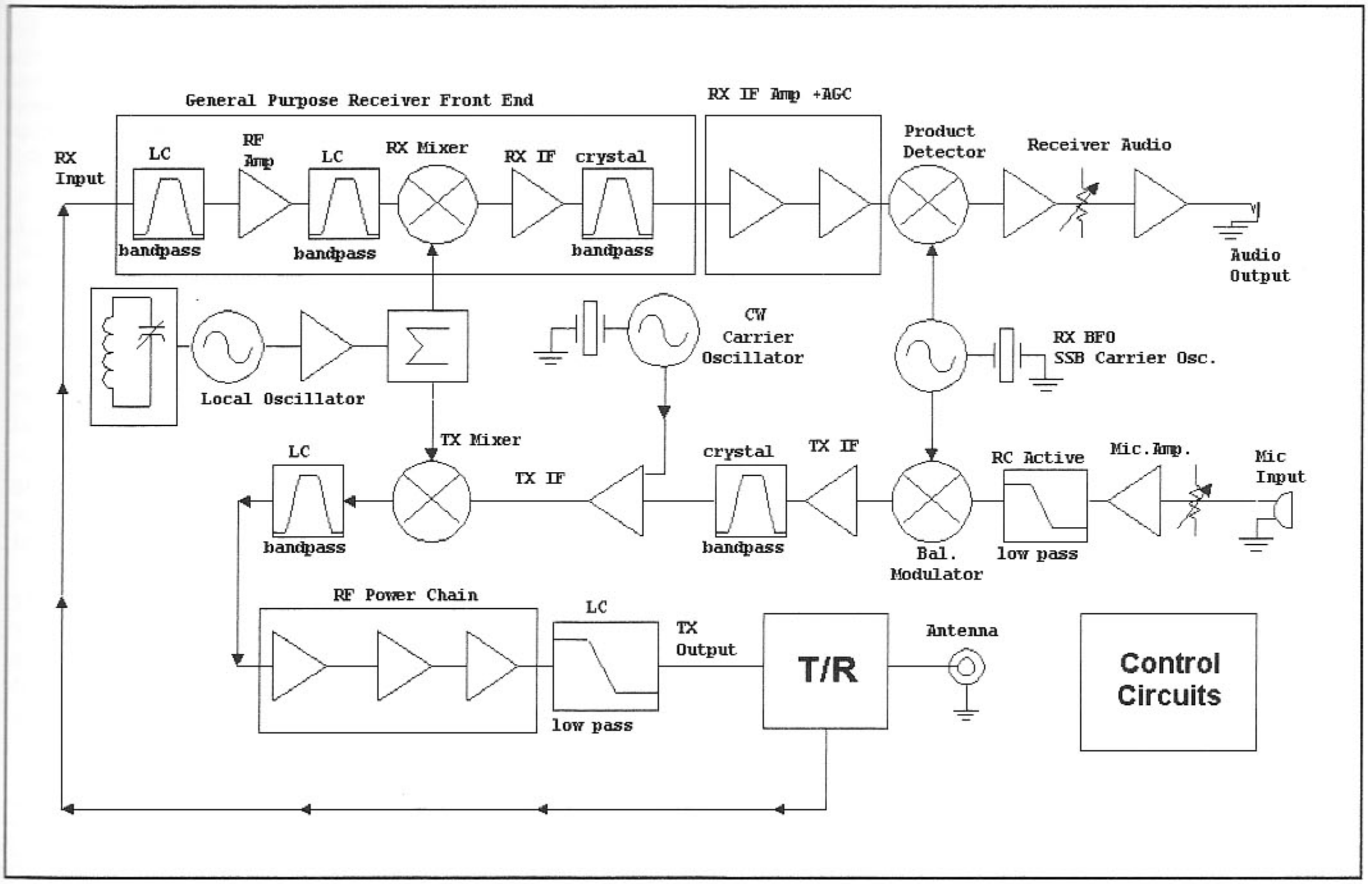


Fig 6.146—Block diagram for the SSB/CW transceiver. The version we built is for the 6-m band, but can be adapted to any band from 1.8 to 144 MHz. The system shown in the block diagram uses a non-heterodyne VFO system.

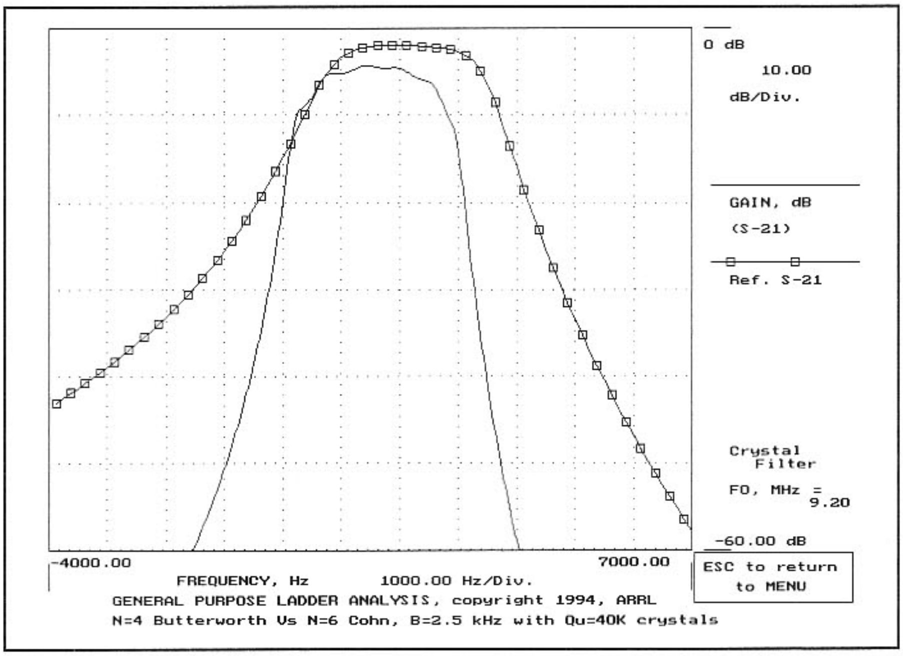


Fig 6.147—Crystal filter responses for two crystal filters. The Cohn is the preferred design for this transceiver even though the low crystal Q_u rounds the passband corners. See text.

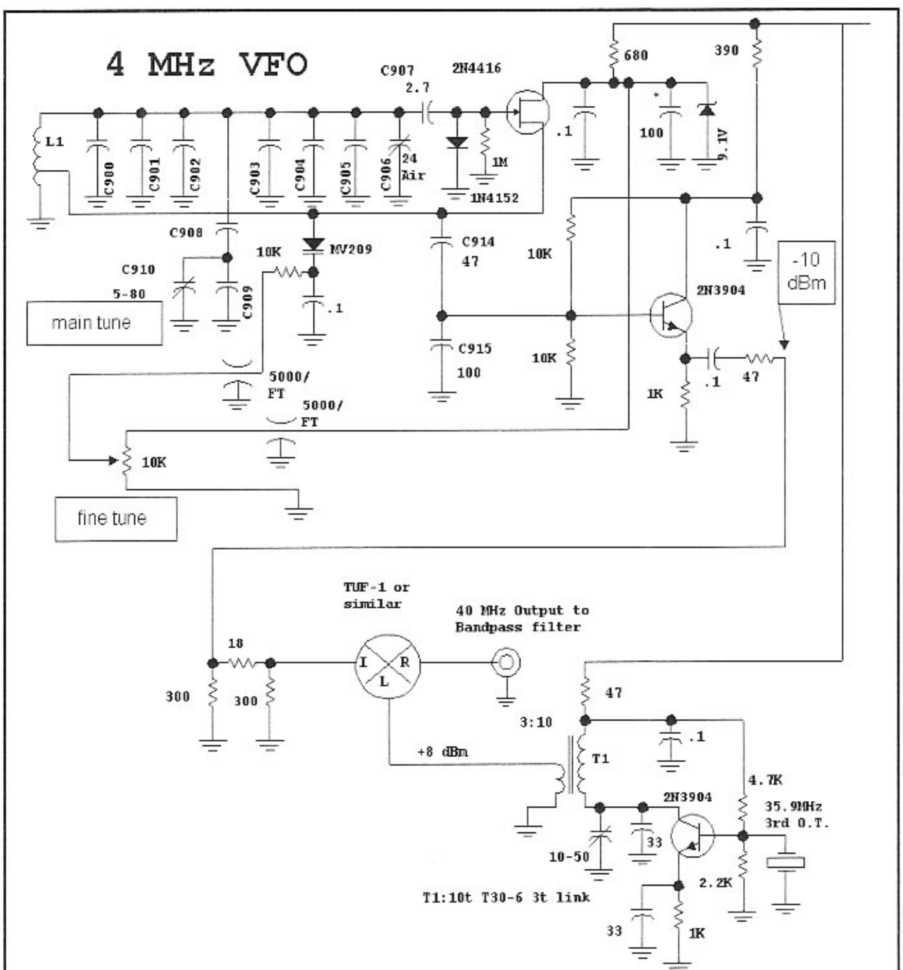


Fig 6.148—VFO for the 6-m transceiver. L1 is unspecified, but will generally be around 5 µH. The many resonator capacitors allow flexibility in setting the frequency. Details are set by the designer/builder.

described here was built for the 6-m VHF band using a 10-MHz IF. However, there is nothing special about that frequency. 10.7 MHz is a good general purpose IF suitable for both HF and VHF. 4.915 MHz has been used in several HF QRP transceivers with good success, based upon available computer crystals.

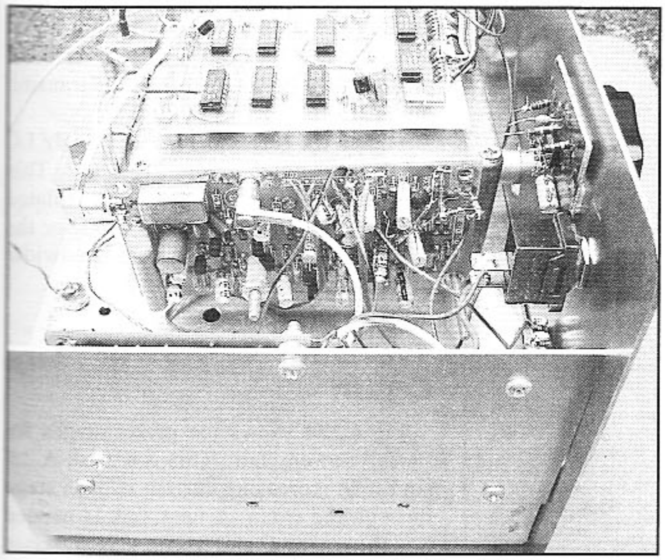
Our 6-m transceiver initially used only 4-pole crystal filters. They were cut for a 2.5-kHz bandwidth with 500-Ω terminations and a Butterworth shape. While the filters performed well, we often wished for better stopband attenuation in both functions. The original thought, that a casual 4-pole filter would be suitable for VHF applications, was clearly not valid when the 6-m band opened in the spring months! **Fig 6.147** shows the calculated response of a 9.2-MHz sixth-order Cohn filter with a 2.5-kHz bandwidth. This is an easy filter to build and duplicate for both functions. The plot also includes a plot for a Butterworth filter with four crystals. The aggressive designer/builder might expand his or her filter efforts to include extra filters to enhance receiver performance and for transmit IF speech processing.

LO System

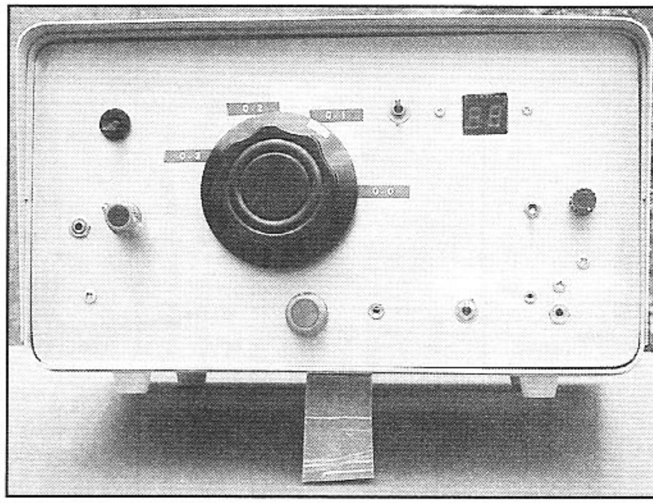
The local oscillator system for the 6-m transceiver is shown in **Fig 6.148**, beginning with a conventional 4-MHz Hartley VFO. An emitter follower buffers the output to a diode ring mixer. A capacitor (C915) is selected to establish a follower output of –10 dBm. The VFO uses a 9-V regulated power supply established with a Zener diode. That regulated voltage is routed out of the shielded enclosure on a feedthrough capacitor to a front panel pot. The voltage generated is run back inside the shield where it controls bias on a varactor diode, D900. The diode tuning range is set up to be about 10 kHz. The main tuning cap, C910, uses a large knob with no vernier drive, offering mechanical simplification. This scheme has been surprisingly effective, even with a tuning range of 350 kHz, a direct result of a large tuning knob on a smooth capacitor. Digital readout provides the needed resetability.

The diode ring mixer and a 35.9-MHz third-overtone crystal oscillator occupy the same enclosure with the VFO. The mixer output is then applied to a coaxial connector through a short run of coax cable. The LO box output is routed on coaxial cable to a 40 MHz bandpass filter, shown in **Fig 6.149.** A triple tuned filter is used to enhance spectral purity. We measured 80-dB rejection of the 35.9-MHz component and the 32-MHz image.

The filtered LO signal is relatively weak



The audio amplifier and product detector board for the Universal Monoband Transceiver.



Front panel of the 6-meter transceiver. The very large tuning knob allows surprisingly smooth tuning without a vernier drive. The knob below the main tuning controls a varactor fine tune function.

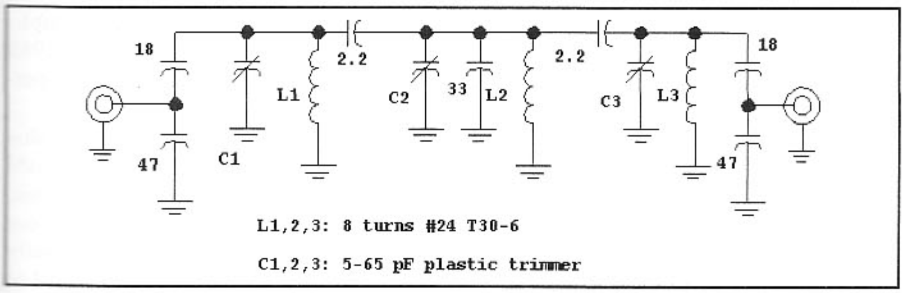


Fig 6.149—Triple-tuned 40-MHz bandpass filter. This circuit was built on a small scrap of circuit board material (approximately 1 × 3 inches) with coaxial connectors mounted at each end. After the filter was tested, a wall was built from ¼-inch brass sheet and soldered to the board. A lid was soldered to the brass walls after filter tuning. The filter was designed for a 2-MHz bandwidth. The inductors had an unloaded Q of 130 at 40 MHz.

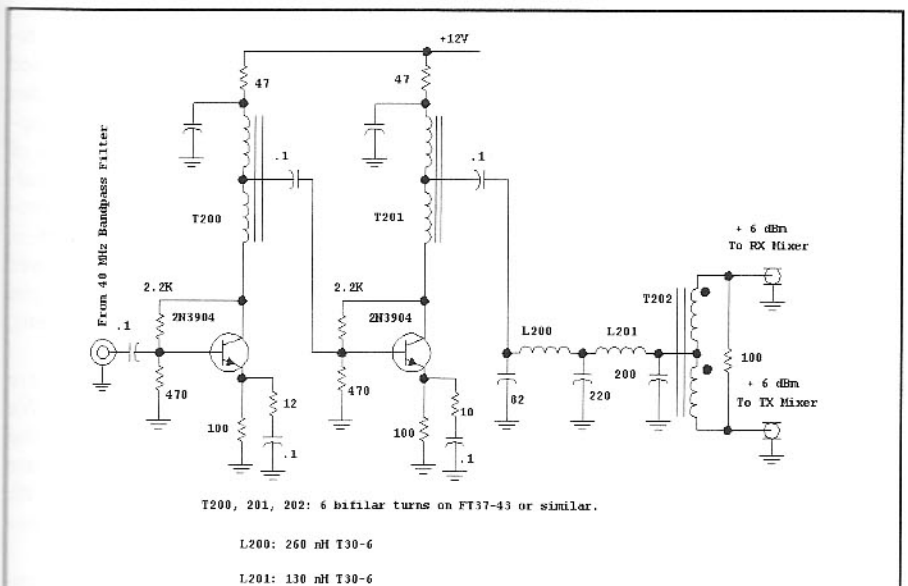


Fig 6.150—LO amplifier feeding 40-MHz energy to the two ring mixers used for the receiver front end and the transmit mixer. T200, 201, and 202 are all 10 bifilar turns #28 on a FT-37-43 toroid. L200 is 8 turns of #24 on a T30-6 core. L201 is 6 turns of #24 on a T30-6.

(about –20 dBm) as it exits the ring mixer and bandpass filter. The level is increased with the two-stage feedback amplifier shown in **Fig 6.150**. The second-stage output is low-pass filtered and applied to a hybrid splitter that delivers two isolated signals, each with a power of +7 to +8 dBm. The hybrid input impedance terminated in a pair of 50-Ω loads is 25 Ω. A low pass filter, initially designed for 50-Ω terminations, was then modified for a 25-Ω load using the procedure of Chapter 3.

BFO/Carrier Oscillator

A traditional Colpitts crystal controlled oscillator generates 10-MHz energy, shown in **Fig 6.151**. The oscillator was modified with inductor L300 allowing oscillation below crystal resonance. Two buffered outputs are available, providing +7 dBm to the product detector and the transmitter balanced modulator. A +12 T supply is applied to only one buffer during transmit periods.

SSB Generator

The SSB Generator board, **Fig 6.152**, begins with an op-amp speech amplifier followed by an RC active low pass filter. A test point allows the audio signal to be monitored to prevent overdrive of the balanced modulator. The peak-to-peak audio signal at TP600 should be 0.4 V for –10 dBm available at the balanced modulator input, which uses a TUF-1 or SBL-1 mixer.

Q600 amplifies the DSB signal from U600 and also sets the driving impedance for the crystal filter. R617 is picked to have the same value as R615, which is the desired termination value for the crystal fil-

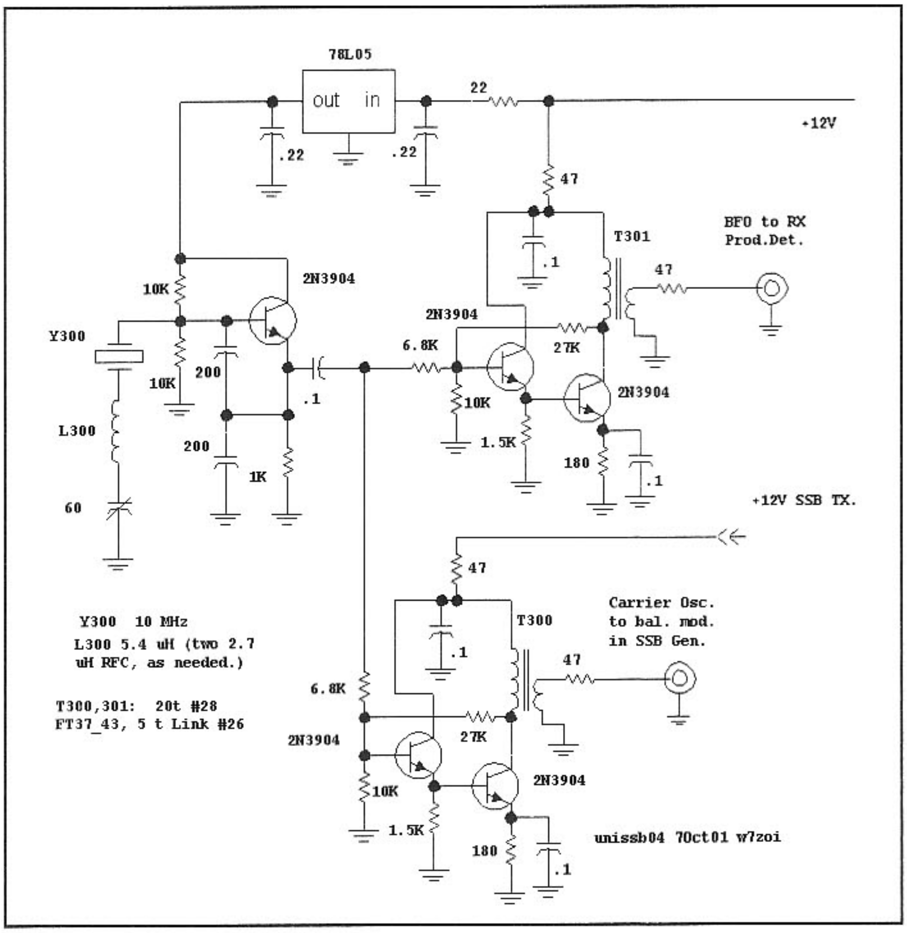
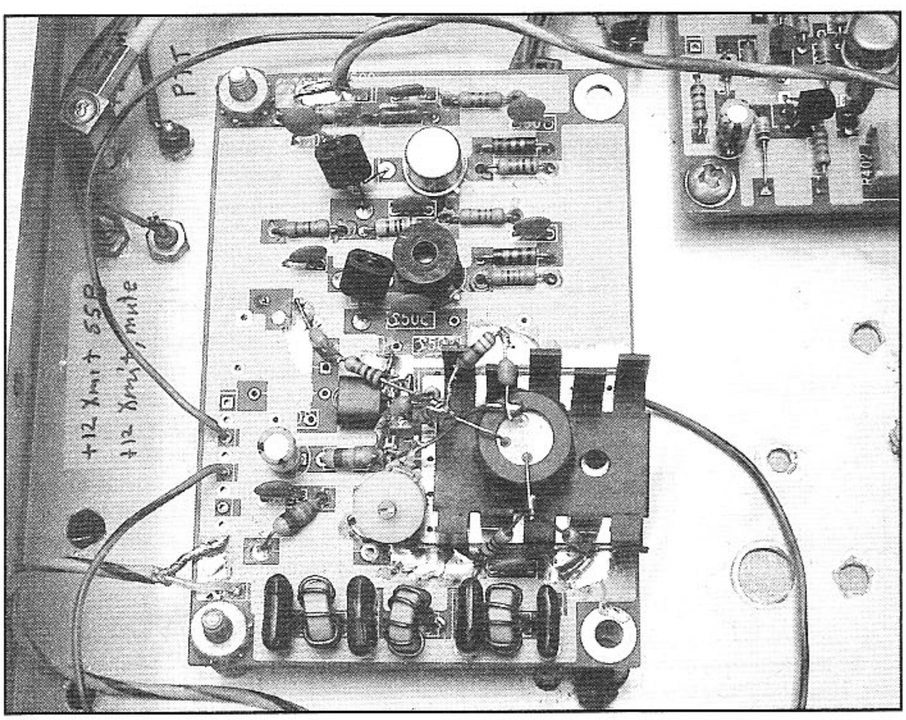


Fig 6.151—BFO and carrier generator. T301 and T300 each have a 15-turn primary with a 5-turn secondary on FT-37-43 cores. The amplifier input resistors, now 6.8 kΩ, can be changed to set the output power.



The RF power amplifiers up to about +23 dBm output.

ter. Further gain is obtained with Q603, 604, and 605. R635 allows a level to be picked that will not overdrive the transmit mixer, U601.

The mixer output drives a 50-MHz LC bandpass filter shown in **Fig 6.153**. This triple tuned filter is build in an isolated box with the same methods used for the LO filter of Fig 6.149 and has a bandwidth of 2.5 MHz.

Transmitter Power Chain

Fig 6.154 shows the driver stages for the RF power chain. This is a class-A design with increasing current in each stage through the chain. A heat sink is needed for the second and third stages. Gain for the chain is 47 dB with an output of 300 mW. The output low pass filter was included for QRP use before a "brick" was added. The low pass could be eliminated (or abbreviated) if a higher power amplifier is planned to follow Q3. A 2SC2988 might be a suitable substitute for Q3 operating at 50 MHz.

The power amplifier used with this transceiver is based upon the Mitsubishi M57735 hybrid integrated circuit, **Fig 6.155**. The hybrid (obtained from Down East Microwave) is an especially convenient part to use, providing 21 dB of small signal gain from a two-stage class-AB circuit. Power output is 14 W for the IC. The chip, which includes a built in low pass filter, is built on a flange that bolts directly to a grounded heat sink. A strip of scrap circuit board material is bolted next to the IC, offering a convenient place for additional circuitry.

Three terminals on the RF module require a power bias. Two use 12 V and feed the two collectors while the third provides base bias networks with 9 V. The 9-V supply should be regulated. In the process of setting up a LM-317T regulator, we realized that it could also function as a programmable circuit. This modification is included in **Fig 6.155** for complete power control over the amplifier. The bias on pin 3 of the IC module is 9.1 V in transmit, dropping to 1.27 V during receive.

The decoupling capacitors used are those suggested by the manufacturer. We measured these networks, finding that the 22-µF electrolytic capacitors we used are modeled with an inductance of 65 nH with very low Q. A better wideband bypass might be several parallel 0.01 µF.

Although the M57735 is ideal for general-purpose applications, it is an expensive part. **Fig 6.156** shows a QRP power amplifier that can be used in place of the hybrid. The output from this stage is 3 W

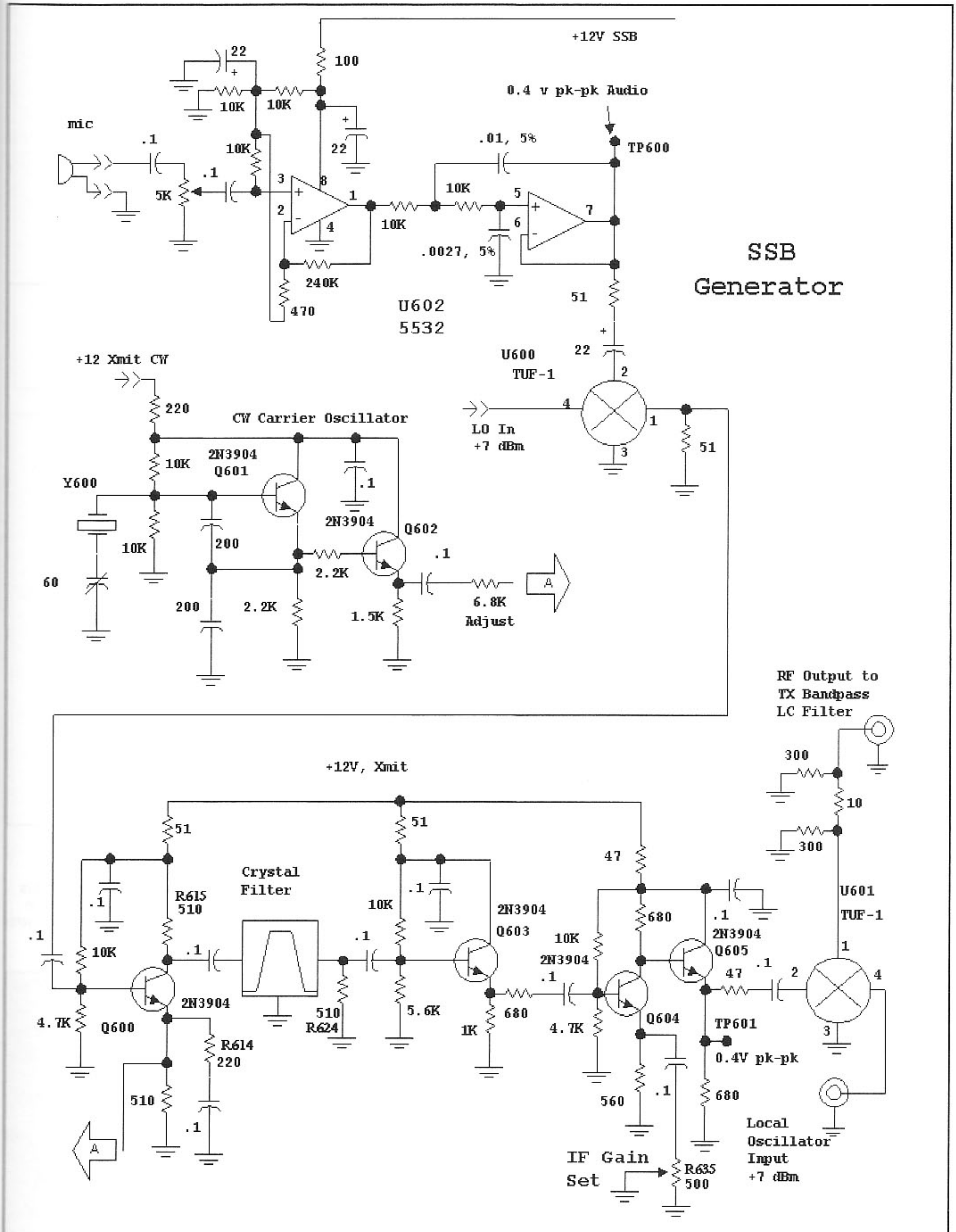


Fig 6.152—SSB generator. R615 and R624 should be picked to equal the desired terminating resistance for the crystal filter, which is a designer/builder-determined element. R614 can be varied to change gain, if needed. R635 is adjusted for 0.4 V peak to peak at TP601 during transmit. That level should be identical in CW and SSB.

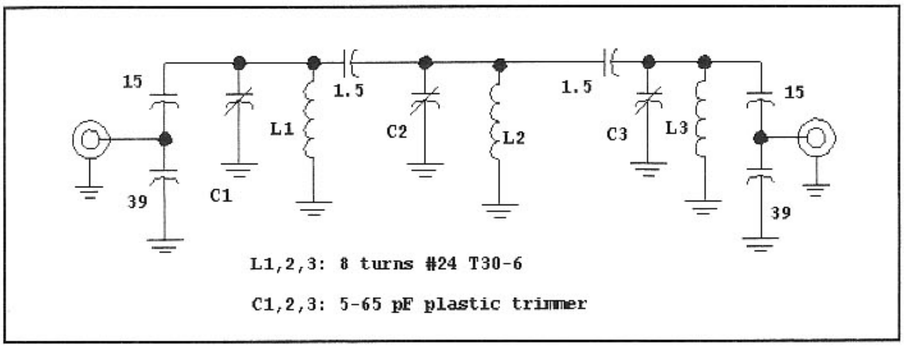


Fig 6.153—Triple-tuned 50-MHz bandpass filter.

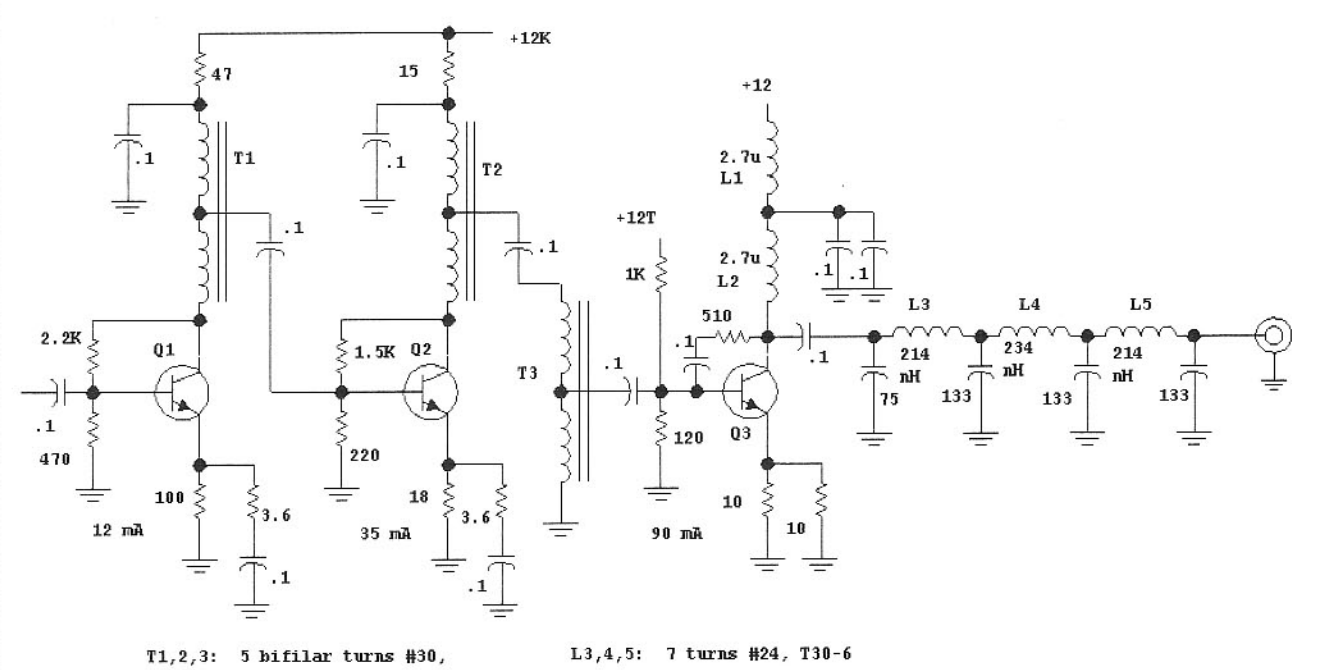
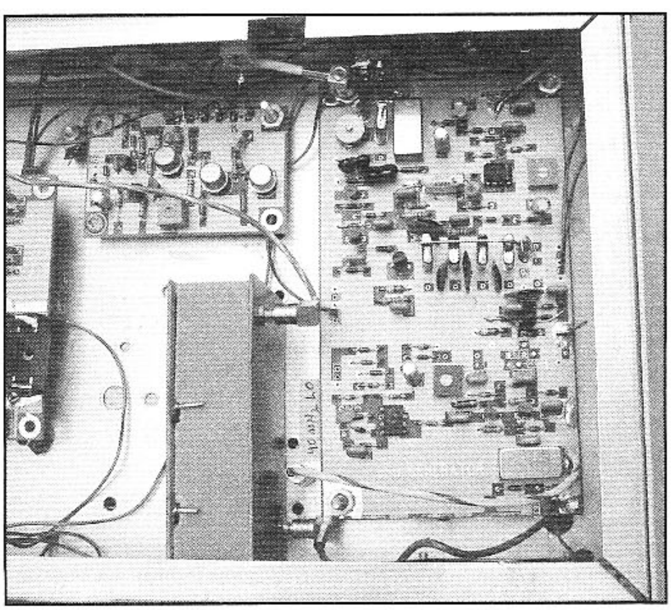
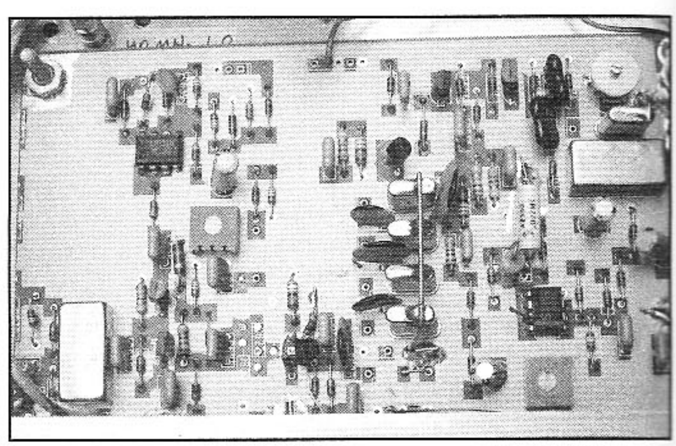


Fig 6.154—Transmitter chain.



The large board is the SSB generator and transmit mixer. This version used SBL-1 mixers. The transmit bandpass filter is in the box fabricated from scrap circuit board material. The control board is above the bandpass filter.



Close up view of SSB generator.

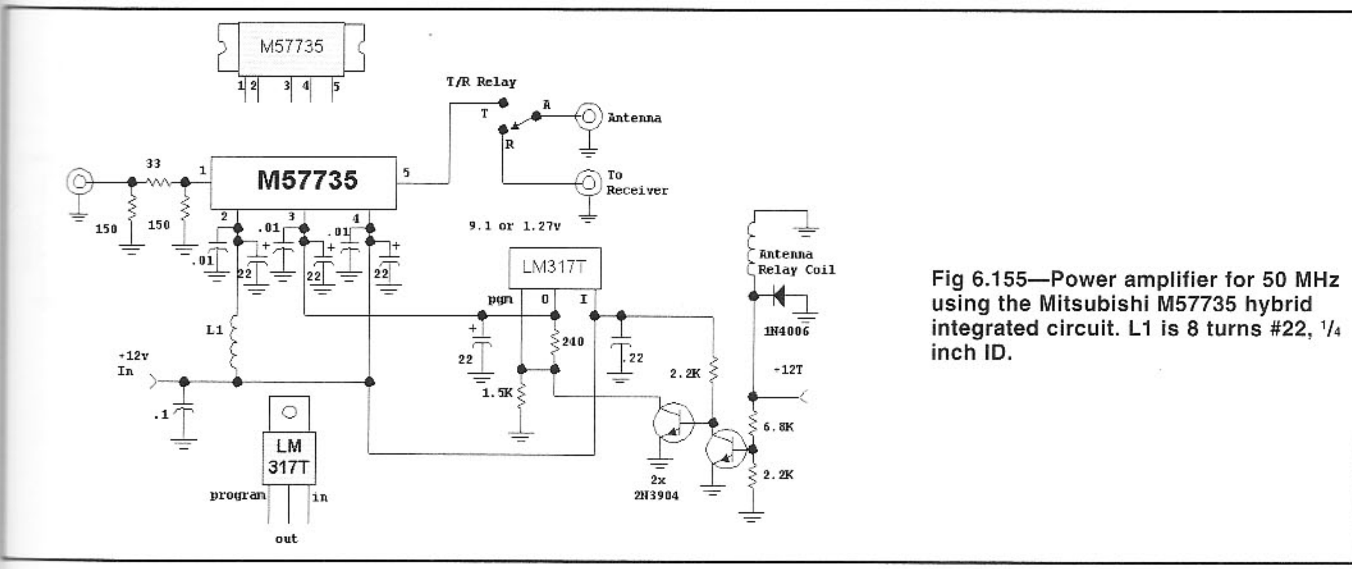


Fig 6.155—Power amplifier for 50 MHz using the Mitsubishi M57735 hybrid integrated circuit. L1 is 8 turns #22, 1/4 inch ID.

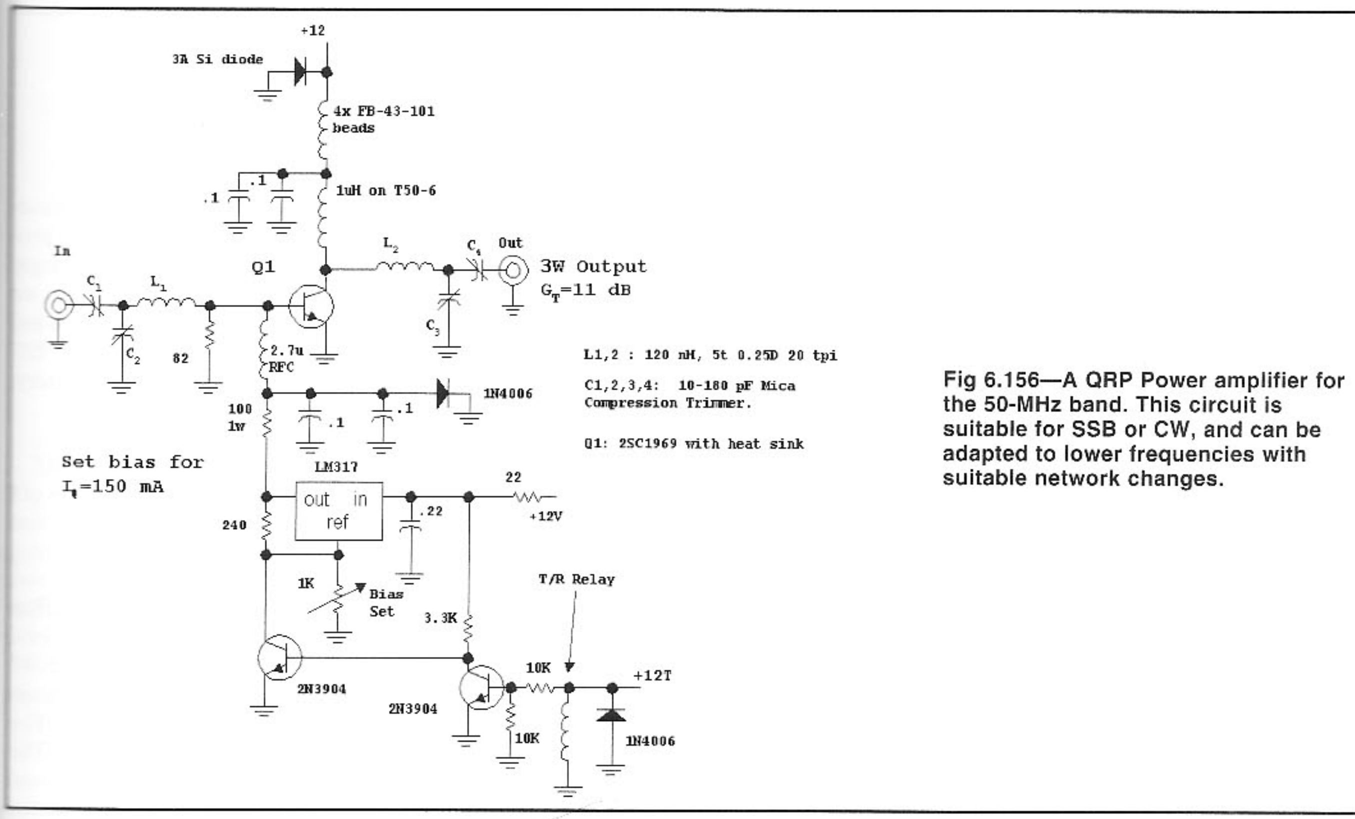
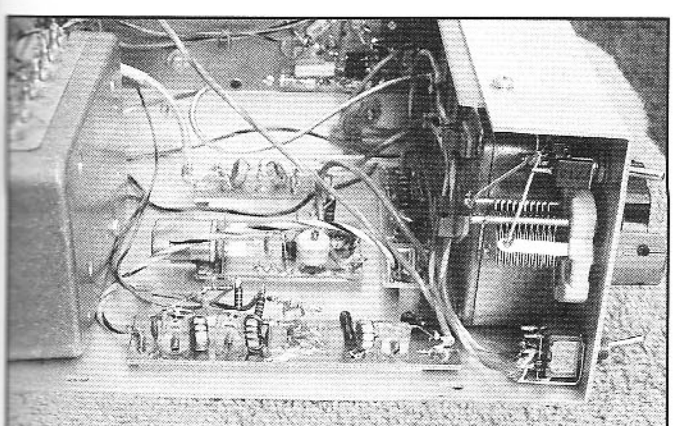
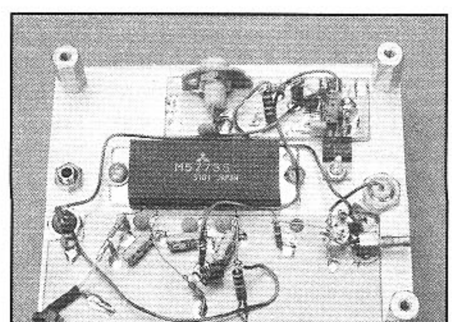


Fig 6.156—A QRP Power amplifier for the 50-MHz band. This circuit is suitable for SSB or CW, and can be adapted to lower frequencies with suitable network changes.



Receiver RF amplifier and preselector filter for the 50-MHz portable station. The variable capacitor tunes the transmitter VXO.



View of RF power amplifier using the Mitsubishi Hybrid. Output is up to 14 W.

with a power gain of 11 dB. This circuit can be adapted to any of the lower-frequency bands, with higher power gain expected. The 2SC1969 transistor is very robust, modestly priced, and available from Mouser.

Receiver Circuits

The receiver circuits resemble others used in this chapter and will not be repeated here. This transceiver uses a low gain RF amplifier, which would not be required for the lower HF bands. We used a shielded double tuned circuit built as a small, measurable filter module as the preselector ahead of the diode ring mixer. The post-mixer amplifier was a 2N5109 with 30-mA bias.

Control Circuits

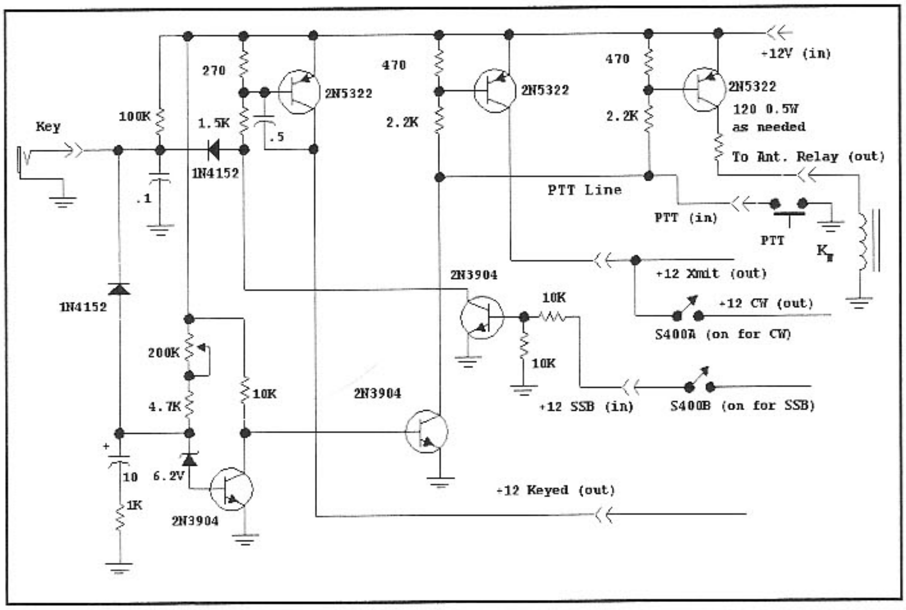


Fig 6.157—Control circuits for the SSB transceiver.

Fig 6.157 shows the control circuitry used with this transceiver. The design is quite general and is suitable for any transceiver with a relay for T/R. With some modification, it should also be suitable for use with PIN diode antenna switching.

The board generates three outputs: +12 relay, +12 transmit, and +12 keyed. These are produced by TO-39 PNP transistors. We have used 2N5401 and 2N5322 in this application. About any PNP capable of switching about 500 mA (often less) will do as well. The TIP-32 should work. Q403, which provides the +12-V keyed signal, generates the shaping required to suppress clicks.

Most of the signals available at the board are inputs. These include a +12 V supply, a ground-active key line, a similar ground-active push-to-talk (PTT) line, and a +12 SSB line. S400B is a DPDT front panel switch that provides +12 SSB during receive and transmit while in SSB, and +12 CW while in transmit mode in CW.

Results

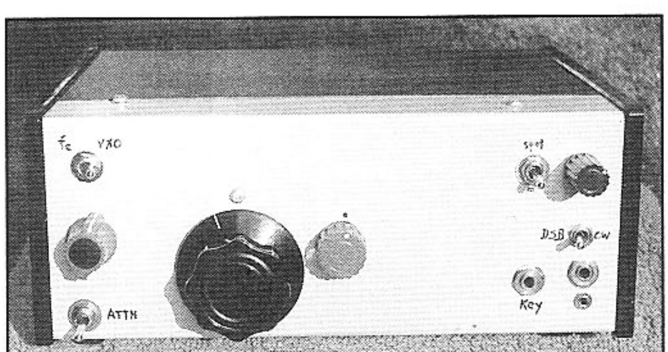
This transceiver has generally been a useful and enjoyable addition, having provided an enjoyable sampling of "The Magic Band." But it is an evolving design that we plan to modify with better crystal filters and a different receiver IF amplifier. The circuit is suitable for operation from a battery, allowing some portable activity.

6.11 A PORTABLE DSB/CW 50 MHZ STATION

A favorite activity for all three of us is VHF operation from interesting locations, usually areas inaccessible to all but one traveling on foot or kayak. Equipment must be fairly light weight. This 6-m transceiver weighs 3 pounds and has an output of 0.3 W.

The rig uses a VXO-controlled DSB and CW transmitter. An 8-MHz direct-conversion receiver is coupled with a simple converter. The transmitter VXO, shown in **Fig 6.158**, uses an off-the-shelf 14.318 MHz color burst crystal. This oscillator is on at all times, but no output is present at 50 MHz until the key or push-to-talk (PTT) switch is closed. U1 then divides the signal by two, producing a 7-MHz square wave from circuitry presented in Chapter 5. The seventh harmonic, occurring in the desired part of the 6-m band, is selected with a double-tuned circuit, amplified with a Mini-Circuits MAR-2 amplifier and further filtered in a second bandpass. The filter output is –3 dBm with the worst spurious response at –64 dBc.

The VXO output is now routed to the transmitter circuit (**Fig 6.159**) where it is increased to +8 dBm with U4, a MAR-3 amplifier, and applied to a TUF-1 operating as a balanced modulator. U4 is driven with either audio from a microphone or dc to provide a CW signal. The –16 dBm modulator output is increased to +14 dBm through MAR-3 and MAV-11 amplifiers, U6 and U7. This then drives a 2N5947 class A amplifier. Suitable substitute transistors would include a 2N5109. The output is about 0.3 W in CW or DSB. The PTT switch on the microphone will ground the key line that also activates the antenna relay circuitry.



Front panel view of the portable DSB/CW transceiver.

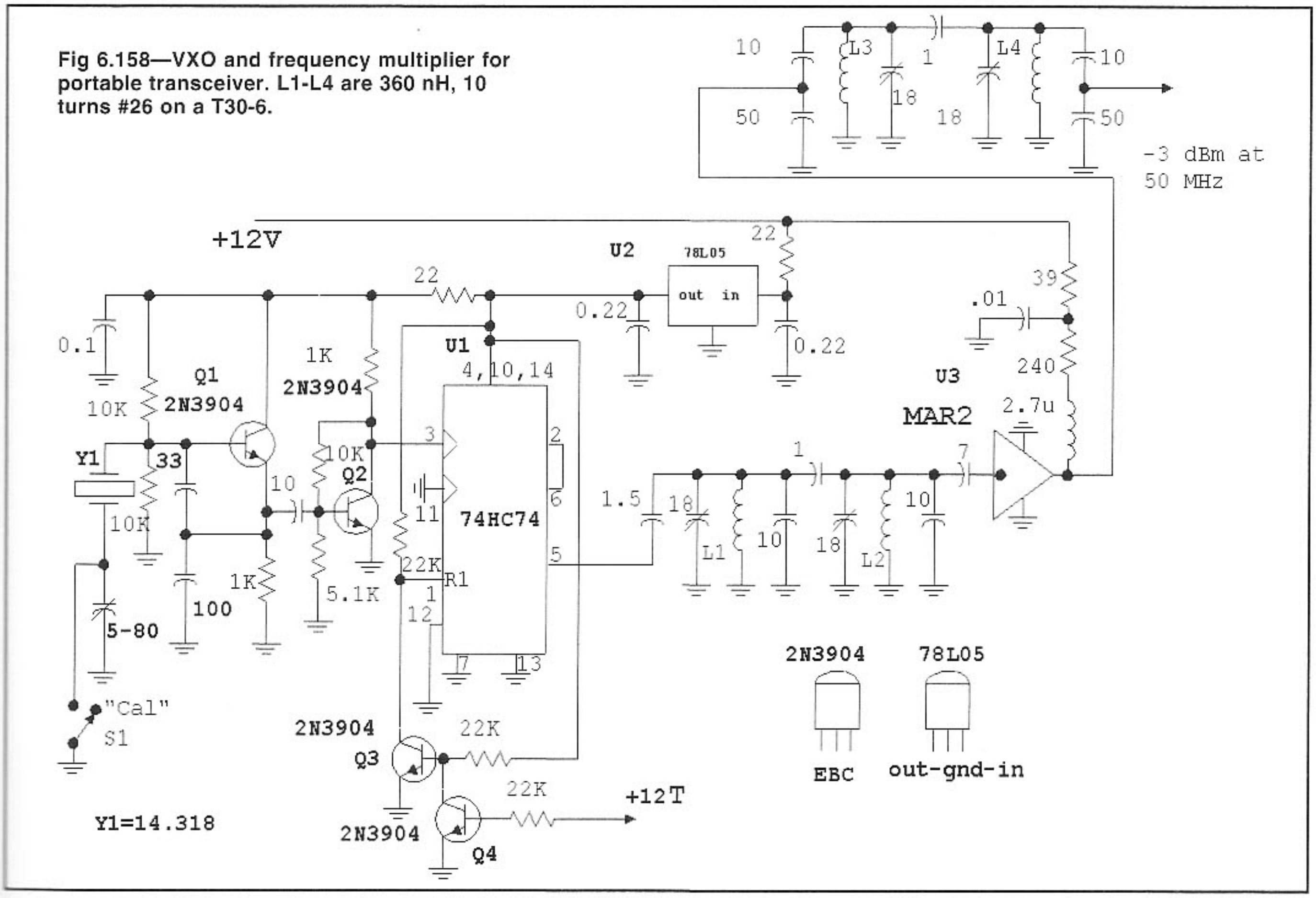


Fig 6.158—VXO and frequency multiplier for portable transceiver. L1-L4 are 360 nH, 10 turns #26 on a T30-6.

The receiving converter, shown in **Fig 6.160**, begins with a single tuned circuit driving a MAR-2 RF amplifier with a gain of about 12 dB. A double tuned circuit then preselects the signal before it is applied to a TUF-1 mixer followed by a 2N5109 post mixer amplifier. A switched 20-dB pad can reduce the signal before the product detector. A PIN diode at the mixer offers additional attenuation.

The converter output is 8 MHz, used merely because a 42-MHz crystal was available in the junk box. A better choice would be 43 MHz. The D-C receiver could then function on the 7-MHz band. The MAR-2 RF amplifier with its input filter could also be eliminated for typical applications, keeping only the double tuned circuit preselector.

Fig 6.161 shows the 8-MHz VFO used with the receiver. This circuit drives a fully shielded board containing the product detector, audio amplifier with switched attenuator, and sidetone oscillator. This module is described in Chapter 12.

Double sideband offers a very simple way to get a phone signal on the VHF bands, one that is compatible with SSB. If we were building this station anew, the minimalist phasing SSB transceiver described in Chapter 9 would probably be used. The VXO used with this rig would provide the needed 50-MHz injection.

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2. An excellent summary of modulation is given in Krauss, Bostian, and Raab, *Solid State Radio Engineering*, Wiley, 1980, Chapter 8.

3. W. Hayward, *Introduction to Radio Frequency Design*, ARRL, 1994, pp 205 and 349.

4. H. T. Friis, "Noise Figures of Radio Receivers," *Proceedings of the IRE*, 32, 7 (Jul, 1944), pp 419-422, or R. Pettai, *Noise in Receiving Systems*, John Wiley & Sons, 1984.

5. **www.ham-radio.com/n6ca/50MHz/50appnotes/U310.html**; See also Gonzalez, *Microwave Transistor Amplifiers, Analysis and Design*, Prentice-Hall, 1984 for designing for lowest noise.

6. W. Carver, "A High-Performance AGC/IF Subsystem", *QST*, May, 1996, pp 39-44.

7. Ibid.

8. For further discussion of AGC loop dynamics, see U. Rohde and T. Bucher, Chapter 5, *Communications Receivers: Principles and Design*, McGraw-Hill, 1988.

9. W. Hayward, "A Competition-Grade CW Receiver," *QST*, Mar, 1974, pp 16-20, 37 and Apr, 1974, pp 34-39. Also see W. Hayward and J. Lawson, "A Progressive Communications Receiver," *QST*, Nov, 1981, pp 11-21.

10. W. Carver, "A High-Performance AGC/IF Subsystem", *QST*, May, 1996, pp 39-44.

11. Personal correspondence between the author and Ulrich Rohde, 1997.

12. W. Hayward, *Introduction to Radio Frequency Design*, pp 219-232. Also see K. Simons, "The Decibel Relationship Between Amplifier Distortion Products," *Proceedings of the IEEE*, 58, 7 (Jul, 1970), pp 1071-1086.

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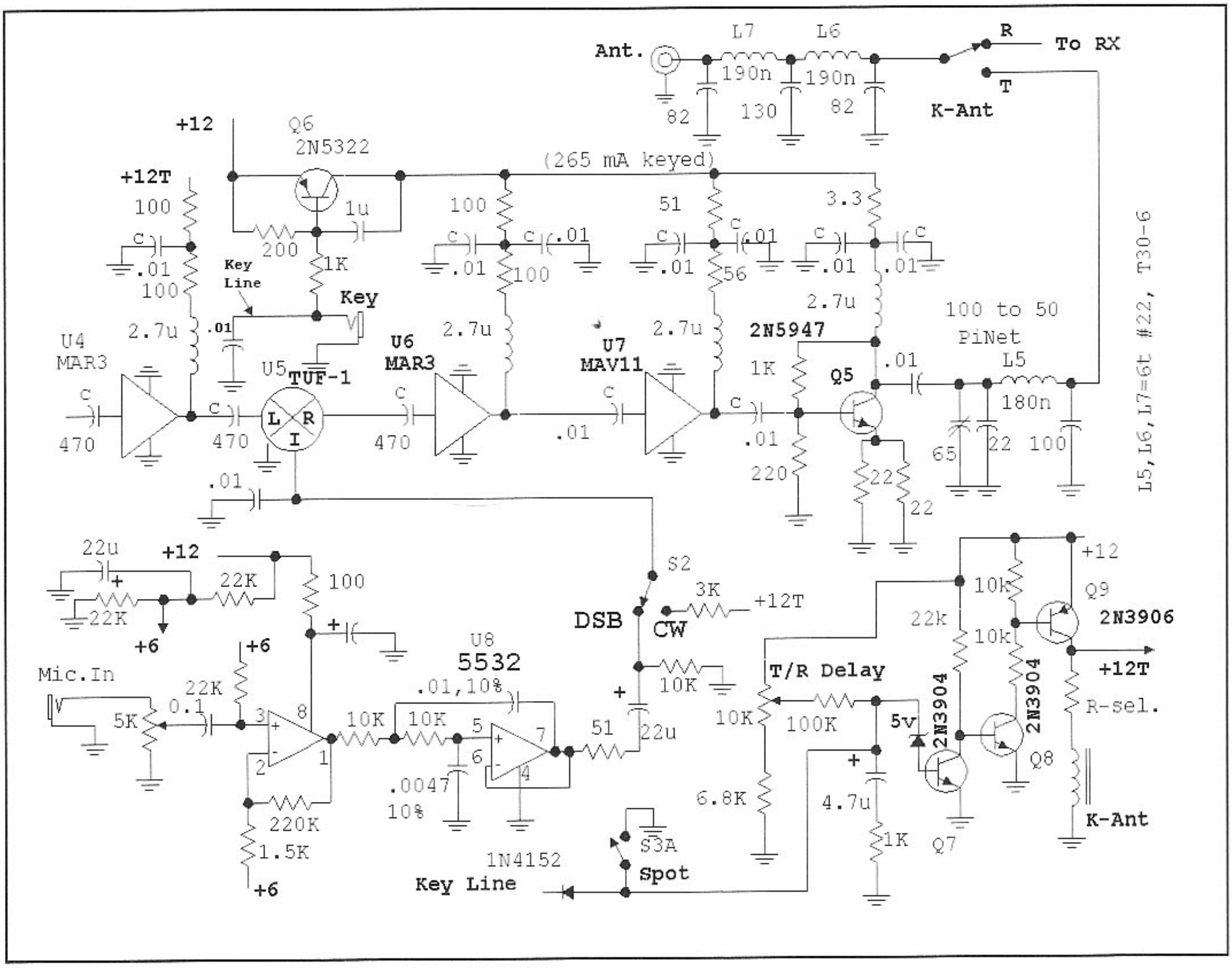
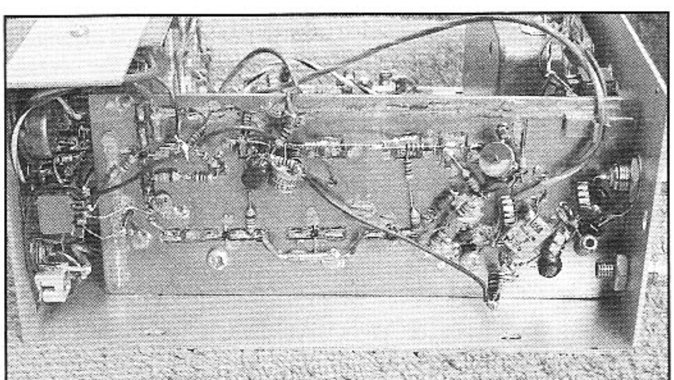
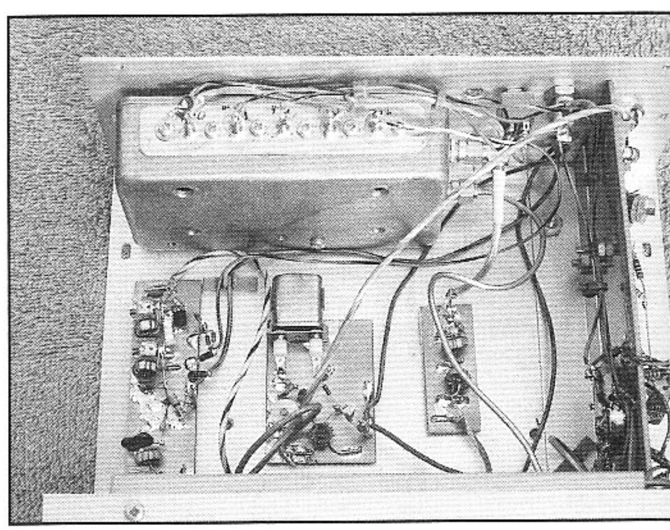


Fig 6.159—Transmitter portion of the 6-meter station.



Transmitter chain for portable rig. Audio microphone amplifier is on the other side of the board.



The audio amplifier and product detector for 8-MHz direct-conversion IF system are all in a Hammond 1590B box with coax and feedthrough capacitor interface connections. The 42-MHz crystal oscillator and 8-MHz low pass filter are on the small boards. The long board across the bottom of the figure is the VXO and ×3.5 frequency multiplier chain.

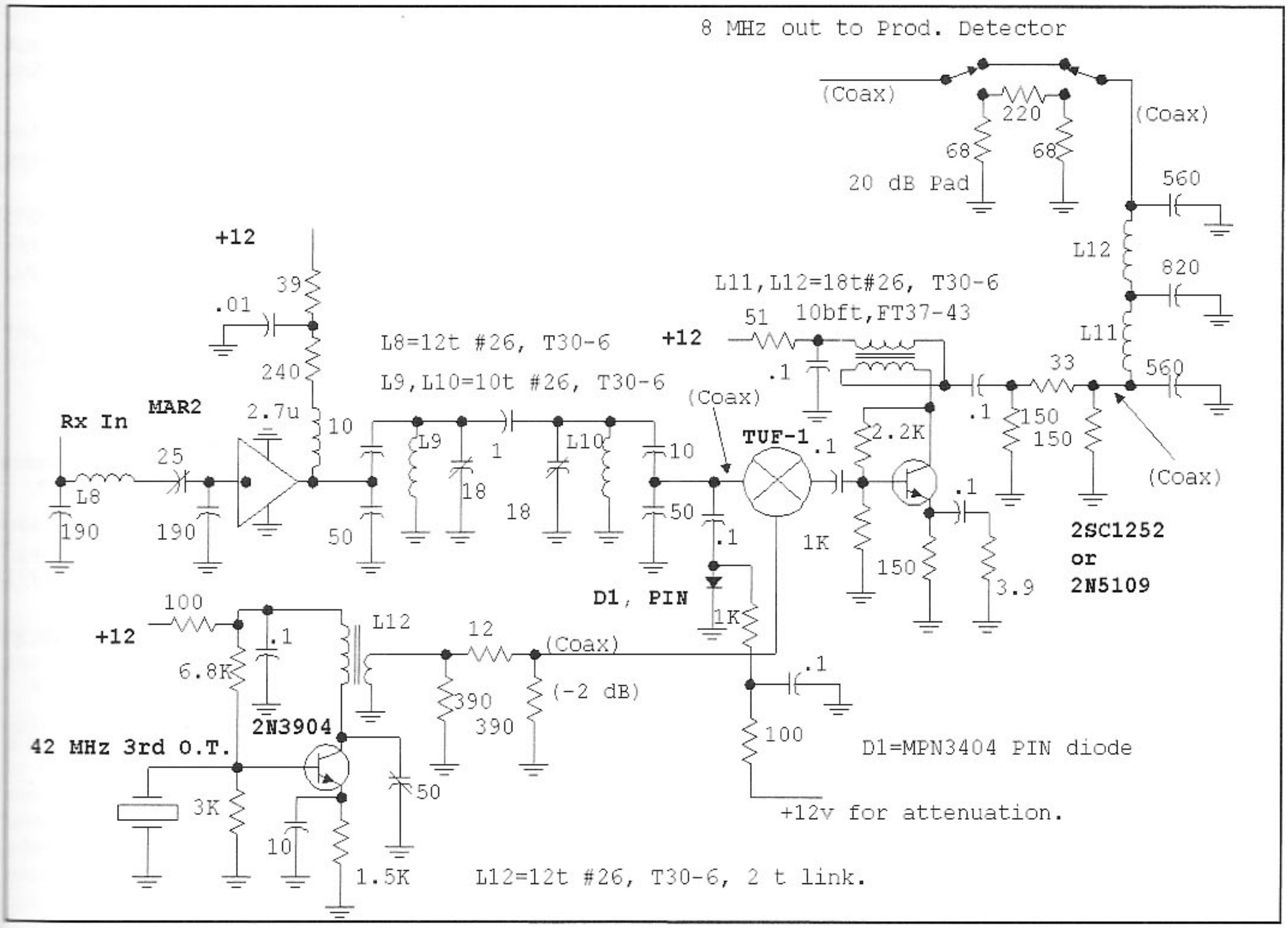


Fig 6.160—Receiving converter used with the 6-m portable station.

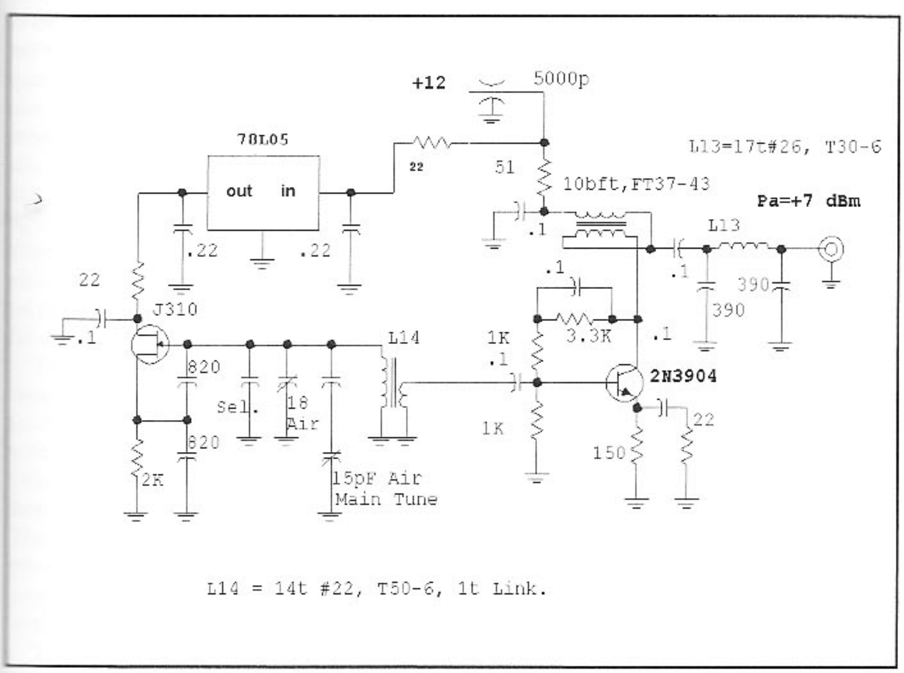


Fig 6.161—Eight megahertz VFO for the 6-m station receiver. Tank capacitors are selected to establish resonance at the desired operating frequency

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CHAPTER 7

Measurement Equipment

7.0 MEASUREMENT BASICS

Measurements are fundamental to all that we do as radio experimenters. The beginner needs a voltmeter to debug the kit he or she has just built, a simple power meter to evaluate it, and a bridge to use in setting up an antenna to use with it. At the other extreme is the designer/experimenter who lives with the equipment needed for the design efforts.

There was a time when the test equipment used by the radio amateur was no more than the indicator level gear needed to build basic gear (VOM and dipper) with the "high end" consisting of service equipment. Today's expectations demand more. Not only do we wish to build some of the equipment that we use, but we want to understand the performance. Our questions probe further as we seek to design that equipment, placing greater demands on measurements. Traditional service gear is usually inadequate, lacking range and accuracy. But it is impractical to purchase the laboratory equipment we would really like to have. The experimenter's measurement gear is often specialized, aimed at performing a few fundamental measurements, but doing so with meaningful accuracy.

This represents a research attitude, emulating the way we might examine a new field where no instrumentation exists, but where the questions must still be answered. The researcher expects to develop new skills as he attacks his or her work. The usual engineer is only expected to possess the skills at the beginning of a project, willing to deal with technology, without an expectation to develop it.

This chapter addresses measurement needs by describing some fundamental test equipment. We begin with some of the equipment needed by the beginner mentioned above, but expand to include the gear needed by the hard-core experimenter. This equipment is based upon some specific guidelines:

1. The experimenter should measure everything that he or she can. Even if you do not have the "right tool," you can often perform an approximate determination. The most casual measurement is still more informative than none.
2. Test equipment need not be refined. That is, simple equipment is still adequate if you can perform a calibration that provides information.
3. The equipment in this chapter is designed for the RF experimenter with a primary interest in building radio equipment. It is easy to become a "test equipment junky" by building and purchasing a great collection of good test gear, with no resources left for the original experiments. Individual goals must be the guideline.

In Situ vs Substitution Measurements

Measurements usually fall into two classes. The in situ or *in place* measurement is one where instruments are attached to a working system. A goal is to extract as much information as possible without disturbing the system any more than is absolutely necessary. Most of the measurements we do with an oscilloscope or a voltmeter occur in situ. Such measurements are the basis of analog electronics.

The contrasting measurement uses a substitution. In this case, part of a system is examined in isolation from the rest, with test equipment substituted for some components. Clearly, this is a major disturbance; the studied system ceases to function during the measurement. However, things can be evaluated that cannot be measured in situ. An example of a substitution measurement would be determination of receiver sensitivity. An oscilloscope or voltmeter can't measure the sub micro-volt signals that are applied to the antenna terminal of the receiver. So, we examine the receiver output while applying a calibrated signal source to the input. Substitution measurements provide the basis for radio frequency electronics.

We *will often describe the measurements we discuss as being substitution* or *in situ*. It is important to isolate the two, for a piece of equipment suited to one mode may be useless for the other. Some equipment can move into both worlds so long as it is applied with care.

Using This Chapter

We will describe a variety of test equipment in the following pages. Some is simple while some is more complex. The order of presentation does not generally coincide with complexity or utility, leaving the beginner searching for the suitable starting point.

The novice experimenter should begin with the simplest gear such as a voltmeter for kit building. Add an instrument for measuring inductor and capacitor values as you progress beyond these beginnings. If you are building any RF communications gear you will want a power meter or some other means for power determination.

As your commitment to experimentation deepens, you will want more test equipment. An inexpensive oscilloscope is probably one of the most useful tools one could acquire. It is useful for the classic in situ analog measurements, the substitution measurements of RF, and even the timing measurements of digital electronics. The oscilloscope then becomes the foundation for numerous other measurement tools.

But, no matter what equipment is being used, simple or sophisticated, keep your goals in mind. Our goal is to understand: Does the gear we build perform as fundamental concepts tell us that it should? This means that the test equipment is in constant use during construction of a project. Each stage in a complicated system is evaluated and confirmed as the system grows. The user should divorce himself from the oversimplified idea that test equipment is merely a tool for final evaluation.

7.1 DC MEASUREMENTS

The most basic instrument of electronics is the galvanometer of fundamental physics. Current flows in a coil to produce a magnetic field, interacting with another field to cause force against a spring. The resulting motion has an attached scale to indicate current.

The simple 0 to 1 mA meter movement is a modern equivalent. This meter usually has a very low internal resistance of 25 to 100 Ω. Larger currents are measured with meter "shunt" resistors while voltage is measured with a series "multiplier" resistor. A 1 mA meter movement would need a 10-kΩ resistor to measure 10 V. Hence, a voltmeter so built would load the circuit being measured as if a 10K resistor was attached to ground. See **Fig 7.1**.

The loading problems are significantly reduced when active circuits append the meter movements. The traditional active instrument is the classic VTVM, or *vacuum tube voltmeter*. A modern equivalent is a voltmeter using an op-amp with an example shown in **Fig 7.2**. The input signal is applied to a very high impedance voltage divider, resulting in a signal to the non-inverting input of an op-amp. The 1kΩ in series with the meter, R_{CAL}, can become a 2-kΩ pot if calibration is required.

Most experimenters tend to purchase general-purpose meters rather than build them from scratch. The typical unit is a digital voltmeter, or DVM that will measure dc and ac voltage and current and dc resistance. Some have become so good and so inexpensive that it is justified to purchase a general-purpose instrument to build into a special application.[1] The typical DVM will have an input resistance of 10 MΩ when measuring dc voltage. Some traditional VTVMs also had a 10-MΩ input resistance, but also had a resistance (1 MΩ or more) built into the tip of the probe used with the instrument. This allowed the probing of sensitive circuits with little loading, even at high frequencies. While the modern DVM will not cause problems with dc loading, the long test lead can certainly cause problems for circuits containing signals at audio or higher frequencies.

While the resolution and accuracy of a modern DVM is outstanding, many users still prefer an analog indication when a circuit is being adjusted. Some DVMs approximate an analog meter movement with a digital bar graph.

In spite of their justified popularity, the user should be careful when using DVMs, for they create some unique problems. Probably the greatest is the assumption that they are as accurate as their resolution. We should not assume that a meter reading a voltage to 1 mV or better is accurate to that level. See the meter's manual. Another often-overlooked problem is the "burden" of these meters when measuring current. Burden is the voltage drop across the meter when measuring current. This can often be several tenths of a volt for high currents, a departure from the classic multimeters of the past.

We often wish to measure audio signals from the output of receivers. This is best done with a true RMS responding voltmeter. Some of the newer DVMs from Fluke and other vendors include this highly useful feature. The user without older meters can still perform true RMS audio measurements by building an appropriate adapter.[2] This paper is included on the CD that accompanies this book.

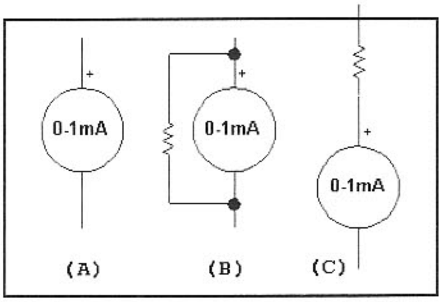


Fig 7.1—A basic 0-1 mA meter (A); measures higher current (B), or voltage (C) with the addition of resistors. Resistance can be measured with these through application of Ohm's Law.

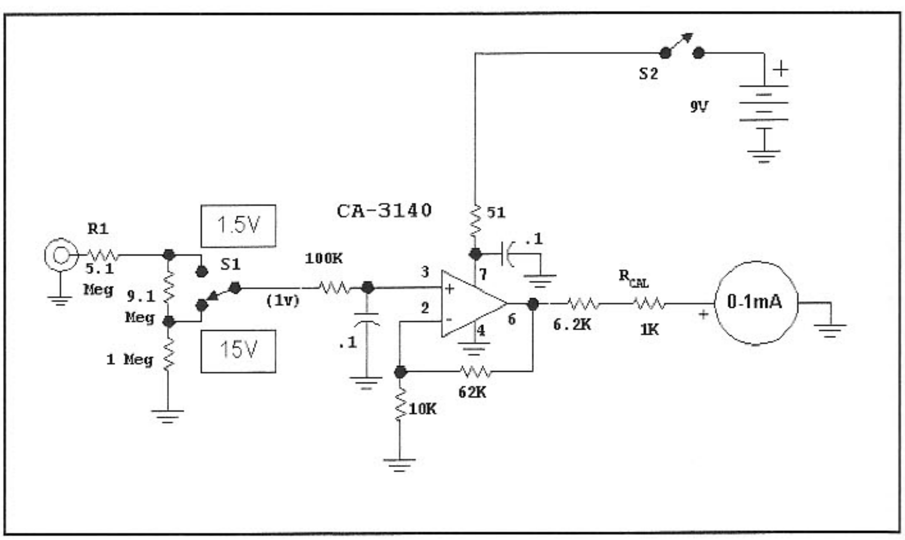


Fig 7.2—A simple op-amp based voltmeter. The meter is one normally intended for use as a 0-15 V meter where a 0-1 mA movement is used with an external 15-kΩ multiplier. The 0 to 15 indication on the meter is now used to register 0 to 1.5 or 15 V, but with a 15-MΩ input resistance. This circuit operates with an op-amp voltage gain of about 7, generating an output of 7 V for a full scale response. With a 9-V supply it becomes virtually impossible to damage the meter movement with excess voltage.

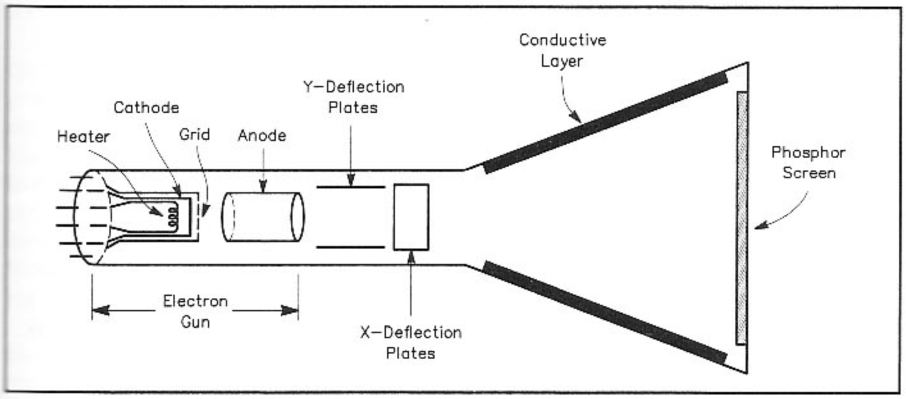


Fig 7.3—Cross section view of a cathode ray tube.

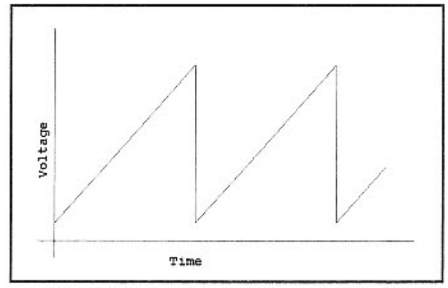


Fig 7.4—Linear ramp applied to the X axis of a CRT. A repeated ramp is called a *saw tooth* waveform.

7.2 THE OSCILLOSCOPE

The ultimate measurement tool for the time domain (explained later) is the cathode ray oscilloscope, or just oscilloscope or scope. This is an instrument that usually measures a voltage that varies as a function of time and displays the result as a time graph. Other measurements are also possible and will be outlined.

The basis for a traditional oscilloscope is the cathode ray tube, shown in **Fig 7.3**. This device begins at the left with a heater and a cathode, the electron-emitting element in the structure. Those unfamiliar with the basics of vacuum tubes can examine their construction and operation in the *ARRL Handbook*. The CRT cathode is much like that in any other vacuum tube, although it is usually a flat or planar surface. Directly to the right of the cathode is a grid. Normal bias slightly negative with respect to the cathode prevents the electrons from leaving the region close to the cathode. Changing the grid bias slightly in a positive direction allows some electrons to escape. They are then accelerated toward an element called an anode. This, plus other electrodes not shown, causes the electrons to be formed into a beam, or *ray*, in accordance with the classic name. The part of the CRT described is called the *electron gun*.

The region after the electron gun contains the deflection electrodes. These will alter the beam direction and allow it to eventually strike the faceplate where it will impinge on a phosphor, a material that gives off light when struck by energetic particles.

Most of the electron gun is biased negatively at a potential of –500 to –2000 V while the deflection region is close to ground. The rest of the CRT is also near ground potential for simple 'scopes. Higher performance instruments often include a high voltage post deflection acceleration (PDA) region for greater brightness.

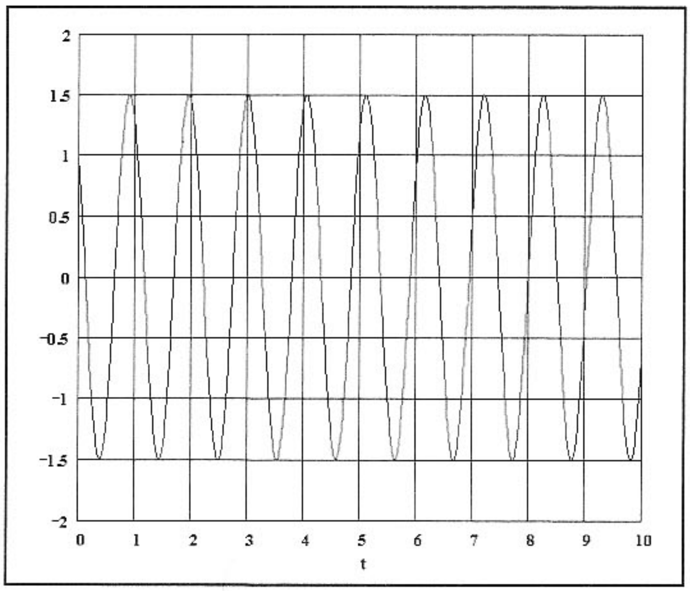


Fig 7.5—The appearance of an oscilloscope faceplate while examining a sinusoid. This ac voltage moves from zero to 1.5 V, back to zero, to –1.5, and again to zero, with the sequence repeating for a long time. This signal is measured as 3 V peak-to-peak. The display shown has a vertical sensitivity setting of 0.5 V per division.

The electron beam leaving the gun passes between deflection plates, often no more than parallel sheets of metal. The beam passes first through the vertical, or Y plates, and then enters the horizontal or X deflector. A voltage applied between the plates generates an electric field causing the electrons to move toward the more positive plate. The electrons are moving quite fast as they enter the deflection region, so the change in direction brought about by the deflectors may be slight. But a few volts across the horizontal plates will cause a beam originally headed for the faceplate center to strike at the edge.

The voltage applied to the X plates will cause the beam position to vary with the applied voltage. If we apply a voltage that is a linear ramp with time, shown in **Fig 7.4**, the result is a horizontal line across the faceplate.

The electrons move predominantly along what is usually referred to as the "z" axis. A signal applied to the grid next to the cathode is called a z axis or intensity modulation.

There are numerous applications for this versatile configuration. For example, if a fast ramp is repeatedly applied to the X axis (called a raster) while a slow one drives the vertical, the entire faceplate area is scanned. Modulation applied to the intensity controlling grid then allows television to be displayed.

Oscilloscope measurements usually begin with a ramp, a voltage that grows linearly in time, applied to the X axis. A signal being studied then drives the Y axis. If that signal, for example, is a simple sine wave, the user sees a sine pattern on the face of the CRT. This result is shown in **Fig 7.5**.

The operation just described would work well if the CRT was very bright and just one sweep occurred. The sinusoid would be seen right after it occurred, but would then decrease in intensity as the phosphor decays in time. Most of the signals we study are repeated in time and we use a low intensity beam that appears again and again. If we did this without doing something special to force the horizontal sweep and the vertical excursion to synchronize, we would have a display like that of **Fig 7.6** where no information is conveyed.

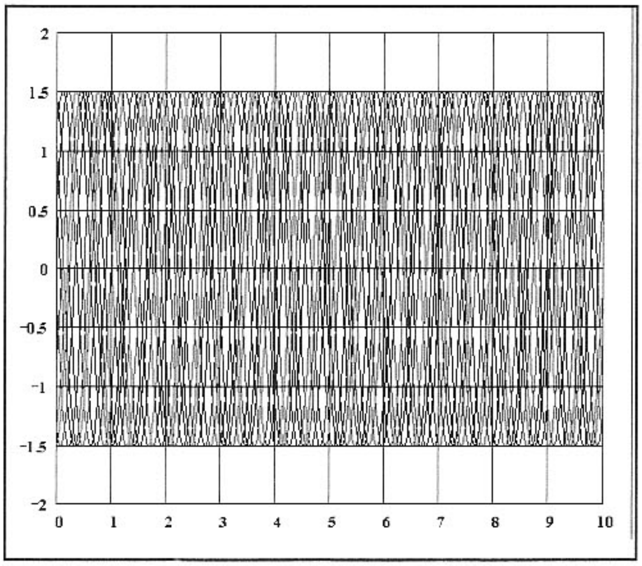


Fig 7.6—The sine wave of Fig 7.5 viewed without triggering. See text.

The elements that cause this synchronization are called trigger circuits, critical parts of an oscilloscope now shown in greater detail in the block diagram of **Fig 7.7**. The trigger is a circuit that looks at the signals present in the vertical channel. Once a predetermined level set by a front panel control (trigger level) is reached, a pulse is generated that is sent to two parts of the system. The pulse reaching the sweep circuit where the sawtooth wave is generated starts the ramp. The pulse reaching the Z-axis system *unblanks* the electron gun, turning on the electron beam. Once just one sweep is finished, it terminates, but starts again when a new trigger pulse is generated.

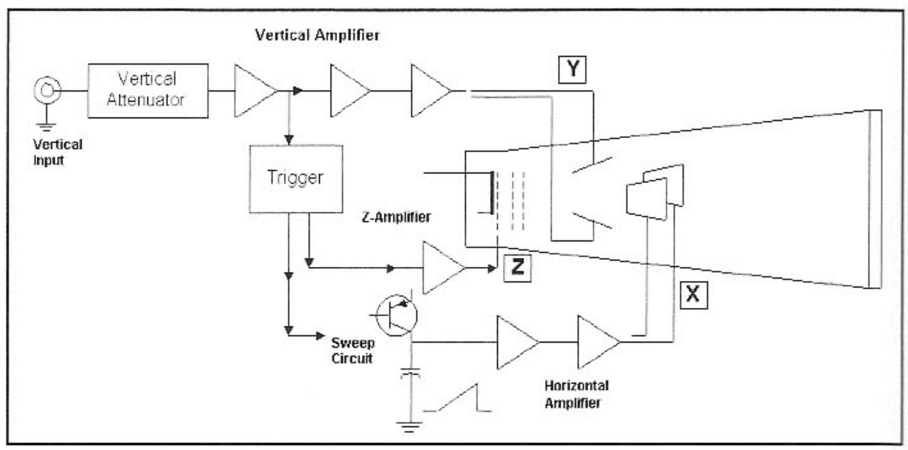


Fig 7.7—Partial block diagram for an oscilloscope. See text for details.

Most 'scopes have an automatic trigger mode that causes a continuous sequence of sweeps to occur. However, as soon as a valid trigger pulse is generated by a vertical signal, that action dominates. While the vertical signal is the most obvious and useful source for triggering, others can also be used. An external trigger terminal is useful for sources that have a well defined associated signal. It is also useful to trigger from the 60 Hz line, allowing related (hum) signals to be examined.

The scope vertical input drives a resistive attenuator that establishes vertical sensitivity. The most sensitive position is typically 10 mV per division, increasing to 10 V per division in a 1-2-5 sequence. All modern scopes are dc coupled, although the user has the option of ac coupling. That is, applying a dc voltage will produce change in the sweep position that remains as long as the dc is present.

The availability of two or more vertical channels is also common. A variety of schemes are used to share one electron gun with the two.

The horizontal sweep is usually calibrated with a wide range of sweeps. One of the instruments used for much of our work is a Tektronix 453 with sweep rates of 0.5 second to 0.1 microsecond per division. Both the vertical and the time base can be operated in un-calibrated modes in most scopes. Further, both X and Y channels have related position controls, allowing the display to be moved to fit the incoming data.

The input impedance of the typical vertical channel is 1 MΩ paralleled by about 20 pF. As such, the loading imposed by the 'scope is not severe. However, it can still be substantial, often dominated by the capacitance of the cable needed to connect the instrument to a circuit being tested.

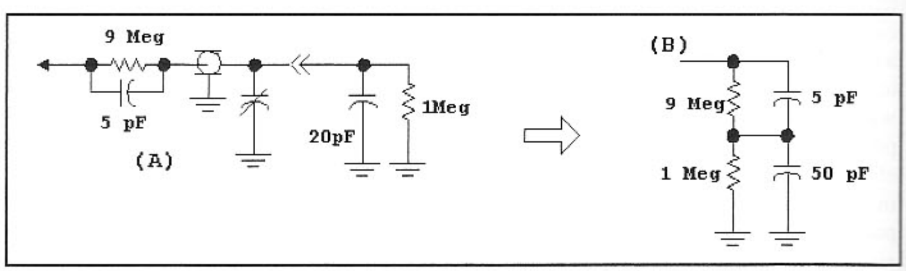


Fig 7.8—A 10X oscilloscope probe. Part A shows the probe and the input to the attached scope while B shows an equivalent circuit. See text.

A typical oscilloscope accessory is a 10X probe, used to reduce the capacitance seen by a circuit being tested. A 10X probe circuit is shown in **Fig 7.8**. A fixed capacitance parallels a 9-MΩ resistor to drive the cable and a variable capacitor. The combination drives the input RC of the scope. The capacitor is adjusted to produce clean, sharp edges when driving the probe from a 1-kHz square wave, the usual calibrator built into most oscilloscopes. Without the 10X probe, the scope input has a low pass characteristic formed by the circuit resistance and the scope input capacitance. The two capacitors of Fig 7.8B form a low pass – high pass combination with effects that cancel (an all pass filter), extending performance to the probe tip.

It is common to find beginners who acquire a new oscilloscope, but do not get the probes to go with it. Don't! The 'scope without the 10X probes is an invitation to misleading measurement attempts resulting from the loading from high oscilloscope input capacitance. Almost all high frequency measurements done with a 'scope are performed with the 10X probe. Even this loading is extreme in many applications.

Most oscilloscopes also have an X-Y mode where one vertical channel drives the Y axis, but the other is attached to the X axis. If you use this setup with two sine waves, you can infer something about the phase relationship between them. Two sine wave signals of the same frequency will produce a slanted, 45 degree line if they are in phase with each other. But a 90 degree phase difference will produce a circle when both have the same amplitude. These are called Lissajous patterns. The X-Y mode is also useful with other instruments that include their own time basis (sweep,) such as a homebuilt spectrum analyzer discussed later.

The up-to-date oscilloscopes offered for industrial and research applications differ from the traditional picture we have painted. While many of the changes relate to extended features, others deal with the very nature of the products. Modern scopes rarely feature the high performance CRTs of earlier times. Rather, the input connectors drive amplifiers that then drive high speed Analog to Digital converters, producing a digital version of a picture that is eventually presented for viewing on an inexpensive display. The performance is often impressive, as are the prices.

As you become accustomed to a new oscilloscope, you will find numerous ways to apply it. It is effective in measuring dc levels as well as the ac signals within a circuit. Careful triggering and setting of horizontal position will allow surprisingly accurate frequency measurements, although not up to counter standards. We will comment on various applications throughout the rest of this chapter.

A good general purpose reference on traditional oscilloscope measurements is the paper by K7OWJ, which is included on the CD that accompanies this book.[3]

7.3 RF POWER MEASUREMENT

One of the first things the beginning communications experimenter wishes to measure is radio frequency power, usually from a transmitter. Although not hard in concept, it can be a difficult measurement to perform with good accuracy.

The simplest way to measure RF power uses a termination with a dissipation exceeding the highest power to be measured, a diode, and a capacitor in a peak detector, shown in **Fig 7.9**. A transmitter to be tested is attached to the load and the signal is rectified by the diode, which then charges the capacitor. The capacitor will reach a voltage nearly equaling the peak ac value. Although virtually any meter can be used, one with a high dc impedance is preferred. A DVM works well, although if adjustments are being done, analog action is still useful.

Assuming a diode drop of 0.6 V, the RF power is given by **Eq 7.1** where R is usually 50 Ω. The breakdown voltage for the 1N4152 diode is 100 V, so dc levels of 50 V can be measured, corresponding to a little over 25 W. One can use higher breakdown diodes or tap the diode part way down the resistor to measure higher power, shown in Fig 7.9B. One must, however, alter the equation to reflect the voltage division.

$$P_{watts} = \frac{(V_{dc} + 0.6)^2}{2 \cdot R} \qquad \text{Eq 7.1}$$

R1 can be a parallel or series combination of resistors to reach the needed dissipation. Two or three watt resistors can be stacked between parallel sheets of circuit board material to reach the 100-W level. If the resistors are spaced from each other, and open to the air, they can be stressed beyond their normal rating for short intervals. One termination we use for 100-W measurements consists of 30, 1.5-kΩ 2-W resistors. These methods are generally confined to 50 MHz and lower.

We can add a voltmeter to the circuits of Fig 7.9 for a stand alone instrument requiring no external meter. Two versions are shown in **Fig 7.10**. The one at (A) uses a 1-mA meter movement with a 15-kΩ resistor to form a voltmeter with a maximum of 15 V. Using Eq 7.1, the maximum power would then be 2.43 W, so the 50-Ω load resistor should have this dissipation rating or greater. A valid choice would be two parallel 100-Ω, 2-W resistors. In practice, 1-W resistors would work well for short tests. The circuit at (B) is actually two power meters with one meter movement. This scheme functions because the typical milliampere meter has a low internal resistance.

The two ranges of the meter at Fig 7.10 are quite different. The one at the right hand input is much like the others discussed while the left input has a 50 mW full-scale reading (+17 dBm). This range is best calibrated against a calibrated signal generator. Alternatively, a higher power meter can be used to measure a

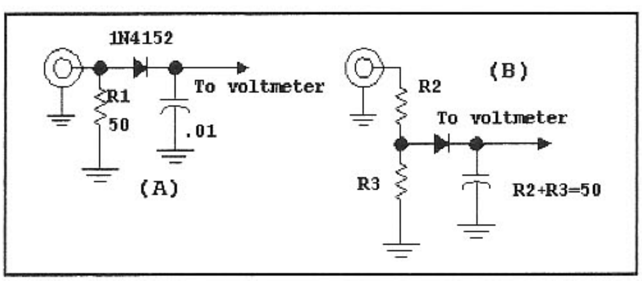


Fig 7.9—A peak detector (A) measures the peak RF voltage across a load, allowing calculation of RF power. The scheme at (B) allows higher powers to be determined without taxing diode breakdown voltages.

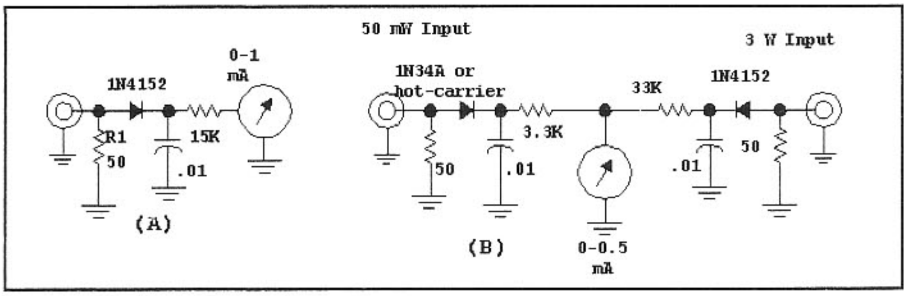


Fig 7.10—(A) shows an instrument with built in meter while the version at (B) has two RF inputs available. See text for details.

dB Arithmetic

Two RF powers are compared as a ratio, or in dB form with

$$dB = 10\ \mathrm{Log}\left(\frac{P_1}{P_2}\right)$$

...where the powers P_1 and P_2 are both in the same units of W, mW or µW. The dB, as well as other logarithmic forms is useful because a change in power ratio is analyzed with addition or subtraction. dB is defined only when *two* powers are considered.

We often specify a power in dB terms with respect to some reference. dBW is dB with respect to 1 W. The familiar dBm is power referred to one mW. These are both ratios, with the 1 (mW) understood. While many power measurements we perform that read out in mW happen in 50-Ω systems, this is certainly not necessary. There is nothing to preclude us from referring to 1.5 peak V across a 150-Ω resistance (7.5 mW) as +8.75 dBm, even though this is not the result we would read if the related power source was applied to a 50-Ω power meter.

With most measurements, an increment from one value to another occurs with a step value of the same units. For example, we change the length of a 50-inch antenna by one inch to becomes 51 inches. The inch unit is used in all cases. But this is not the case with dB and dBm. An absolute power of 20 mW (+13 dBm) is increased with an amplifier by a factor of 5 (7 dB) to 100 mW (+20 dBm.) A dBm value is altered by adding a dB value to become a new dBm value. The ratio of two powers is obtained by taking the difference of their dBm values to get a power ratio in dB.

It is usually not correct to "increase a +27 dBm power by *3 dBm*," which would literally mean increasing 500 mW by 2 mW. What was probably intended was to double (3 dB increase) the power of a +27 dBm (one half watt) source (500 mW) to 1000 mW (+30 dBm or one watt.)

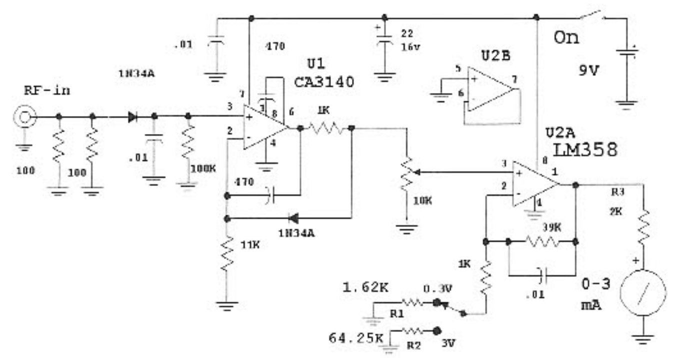
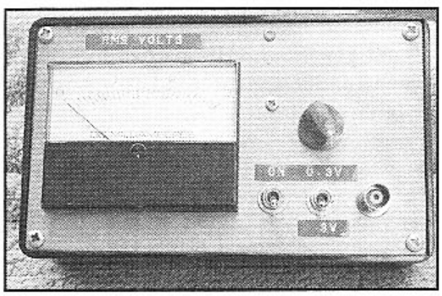
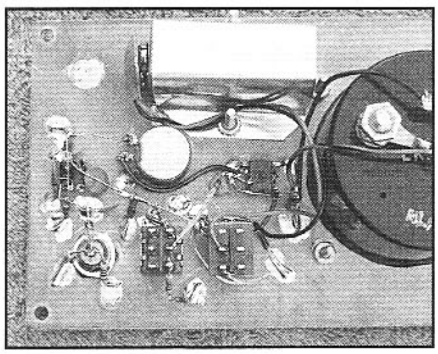


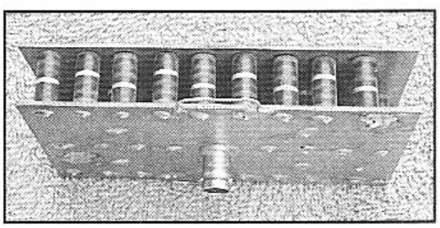
Fig 7.11—This power meter, based on the work of W7EL, has full scale readings of 0.3 and 3 volts RMS with sensitivity of less than –10 dBm. The circuit can be adapted to other ranges. R3 can be changed to 6 kΩ if a 0-1 mA movement is used. See text for details.



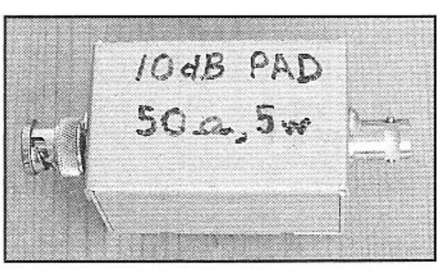
50-Ω power meter using the compensation method of W7EL.



Inside view of the W7EL type power meter.



Thirty parallel 2-W, 1.5-kΩ resistors sandwiched between postcard-sized pieces of circuit board material form a medium power termination. Although the rating is only 60 watts, the wide spacing between resistors allows 100 watts to be dissipated for modest times. The wire hooks are convenient places to attach an oscilloscope 10X probe.



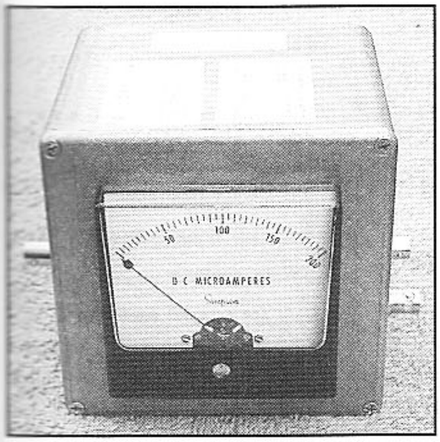
A 10-dB pad built into a small box is a valuable piece of test equipment as well as a station accessory suitable for reduced power experiments.

suitable source such as a QRP transmitter. A step attenuator is then used to decrease the power in known steps to calibrate the 50-mW input. The more sensitive meter can detect powers as low as 1 or 2 mW.

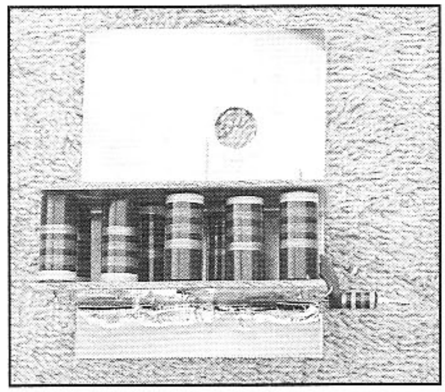
The intended purpose of power meters with small maximum power is not to test very small transmitters. Rather, it is to measure RF power in the early stages of transmitters or in receiver LO systems. A very common example is when setting up a diode ring mixer using hot carrier diodes for LO power of +7 dBm (5 mW.) This is a substitution measurement where a source is set for an available power of 5 mW into 50 Ω, even though it is attached in practice to a less ideal termination.

Microwatt Meter Circuits

Several methods can extend the sensitivity of power measurements, allowing lower levels to be read. One uses an op-amp to follow the RF detector. This guarantees a high impedance load for the detector. Then a matching diode is placed in the op-amp feedback path, which essentially removes the effects of diode



One box contains three power meters with full scale responses of 100 mW, 2 W, and 20 W.



Nine parallel 470-Ω resistors form the RF load for the 20-W power meter. The diode detector and meter multiplier hang on one side. The BNC connector mounts the board to a wall.

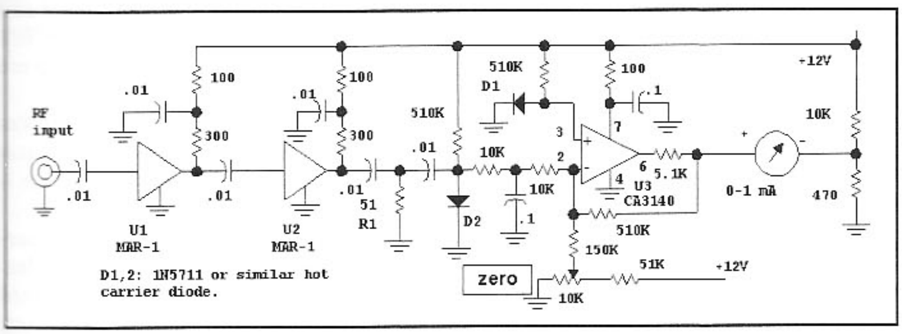


Fig 7.12—Low-level power meter capable of well under 1 µW full scale. This circuit is calibrated against a calibrated signal generator, or against an attenuated QRP transmitter that has been measured with a simple power meter.

offset. This method was presented by Grebenkemper in 1987 and then applied to an in-line QRP power meter by Lewallen in 1990. Both papers are outstanding and are included on the book CD.[4,5] Both instruments included built-in directional couplers that allowed them to be used for in-line power and VSWR measurement.

The simple power meter shown in **Fig 7.11** was adapted from Lewallen's design. The input is a 50-Ω termination followed by the detector. The following op-amp includes a diode within the feedback path. The major effect of this diode is to cancel the effect of the voltage drop across the detector diode, forcing the meter to generate a reading closer to the RF value. The panel meter available when this was built had a 0-3 mA movement, so the instrument was set up for full scale readings of 0.3 and 3 V, RMS. This does not mean that a true RMS voltage is being read. It's still essentially a peak reading circuit, but is calibrated with regard to the related RMS value. Resistors were selected at R1 and R2 to establish the ranges. Lewallen used pots in his meter. The circuit in the figure easily responds to signals less than −10 dBm.

Fig 7.12 shows a power meter using two other methods to obtain greater sensitivity. The first is bias: The diodes are biased at about 20 µA in this system. Two diodes are used in a differential arrangement to reduce temperature drift. The bias

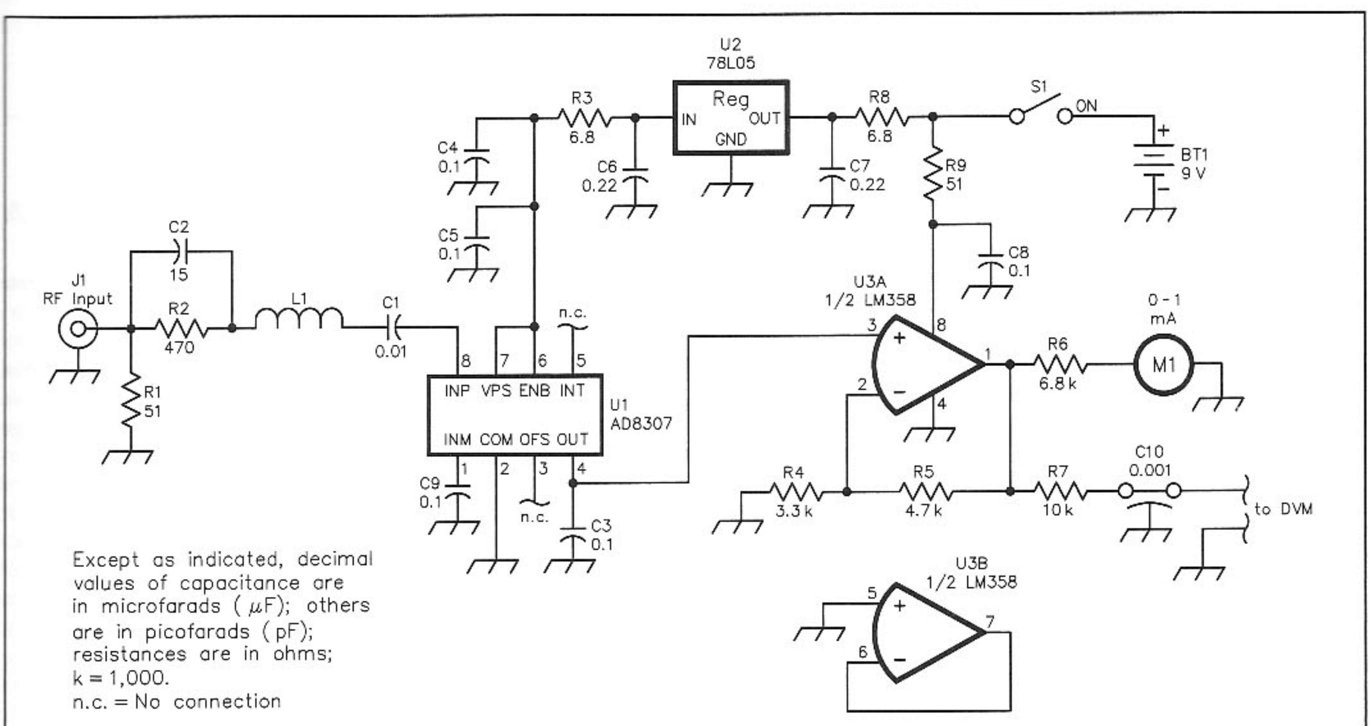


Fig 7.13—Logarithmic power meter capable of reading signals from −80 to +13 dBm.

allows us to see signals of –10 dBm or better at R1. Leaded or surface mounted hot carrier diodes are used. This circuit worked with 1N4152 diodes, although the sensitivity was reduced by a couple of dB. This detector functions well to over 1 GHz. An op-amp provides an interface between the diodes and the meter, and protects the meter against damage from overdrive.

Second, we enhance sensitivity with amplifiers before detection. Here, we use some of the inexpensive monolithic microwave integrated circuits (MMICs) from Mini-Circuits. Discrete feedback amplifiers could also be used.

This power meter will detect signals as low as –40 dBm full scale. This circuit displays about 10 dB of change in the meter motion, making it ideal for careful adjustment of filter circuits. The simpler peak detector power meters (Fig 7.9) typically had 18 dB or higher scale range.

Even greater sensitivity is available from the circuit of **Fig 7.13**. This power meter is based on a logarithmic amplifier integrated circuit from Analog Devices, the AD8307. This circuit functions as a logarithmic detector, accepting signals from audio up to 500 MHz over a power range from around –80 dBm up to over +10 dBm. The output is then a dc signal that tracks with spectacular accuracy, changing by 25 mV for each dB input change. The chip has a sensitivity that drops with frequency, but the circuit shown is compensated to be flat to beyond 500 MHz. This power meter is described in detail in a paper on the CD that accompanies this book.[6]

Any of the low level power meters described can be extended to higher levels with a variety of methods. One is a power attenuator, described later. Another is the 40 dB "tap" shown in **Fig 7.14**. This is essentially a small metal box with a wire connection through to an output attached to a high power termination, or *dummy load*. But the path is sampled with a large value resistor that then drives a 50-Ω terminated connector leading to the power meter. The power available at the tap is, in this example, 40 dB below that flowing in the main path. The *wire* between J1 and J2 is actually a piece of metal, approximately 1 x 1.5 inches, trimmed to fit the box, a Hammond 1590A. With the compensated power meter of Fig 7.13 with a maximum power of +13 dBm, signals beyond +50 dBm, or 100 W can be measured with the tap. The designer/builder should run the circuit only for short periods at full power, for the resistors used in the tap are otherwise taxed.

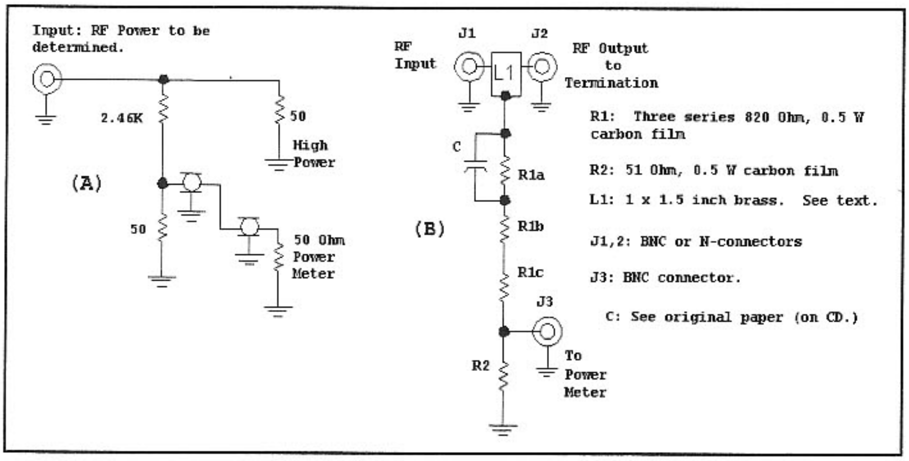


Fig 7.14—Power tap with 40-dB attenuation. Part A shows the basic concept while B shows the version built. See text and original paper on the book CD.

The power meter using the AD8307 was originally described in a *QST* article that is included on the CD. The tap information is in that paper.[6]

The in-line power meter referenced earlier by Grebenkemper used two simultaneous detectors attached to the forward and reflected ports of a directional coupler. This allowed both components to be displayed at once. Further, calculations could be performed on the resulting data. (Op-amps would probably be used.) N2PK has used a pair of AD8307 ICs to obtain similar performance with reduced powers.

7.4 RF POWER MEASUREMENT WITH AN OSCILLOSCOPE

Fig 7.15 shows how RF power is measured with an oscilloscope. A key element is the 50-Ω terminator. This is a 50-Ω resistance that can be paralleled with the oscilloscope input connector. The usual 'scope vertical input is 1 MΩ paralleled by 20 pF, essentially an open circuit for low impedance RF. The terminator is effective in setting impedance to 50 Ω. A terminator used for power measurement should *always* appear at the scope end of the coax cable and never at the transmitter end.

This method is limited to the power dissipation of the terminator used and by the vertical input limits. Higher powers can be measured by adding a 50-Ω attenuator in the line. Much higher power can be measured by routing a transmitter output to a 50-Ω load through a directional coupler or tap (described earlier) in the interconnecting cable.

A 10X probe forms the second recommended method for RF power measurement, shown in **Fig 7.16**. A power termination (dummy load) is connected to the transmitter with a coaxial cable. The voltage across the load is then measured with the probe. This method is generally suitable for powers up to 100 W at HF, 3 to 30 MHz. The ground lead should be clipped to the ground part of the load.

Voltages exceeding around 300 V can damage the usual oscilloscope probe, and additional de-rating is required above 10 MHz or so. For example, a 10-X probe may well present an impedance of only 5 kΩ by the time you reach 10 MHz, even though the resulting voltage measurement is accurate.

An often used, but generally inaccurate measurement is shown in **Fig 7.17**. An external dummy load is used, but the interconnect is realized with sections of 50-Ω cable. The difficulty results from transmission line behavior. We wish to examine the voltage across the 50-Ω termination while configuring the lines so that a 50-Ω load is presented to the transmitter under test. A 50-Ω load at one end of a coaxial cable with 50-Ω characteristic impedance presents 50 Ω at the other end. These measurement requirements are satisfied by the setup of Fig 7.15, but not with that of Fig 7.17.

Once a voltage measurement has been performed, it is easily converted to power with one of several equations, shown in **Fig 7.18**.

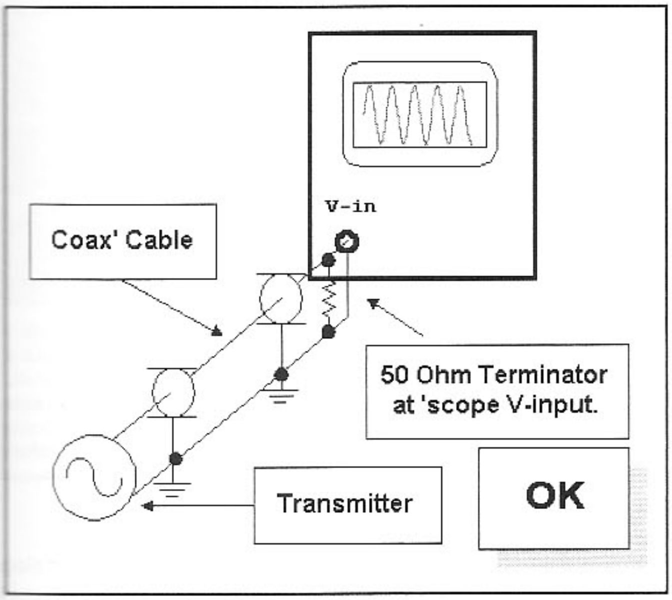


Fig 7.15—Power is measured with an oscilloscope and a 50-Ω terminator at the scope input connector.

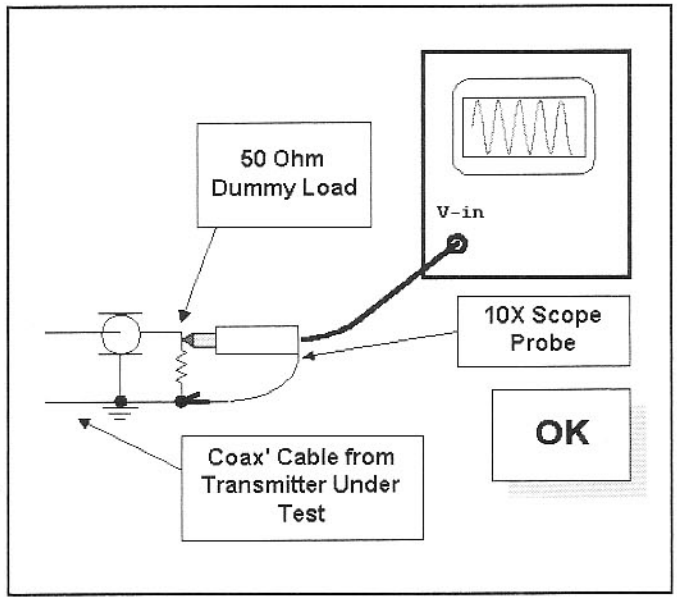


Fig 7.16—A 10X probe is used with an oscilloscope for power measurement.

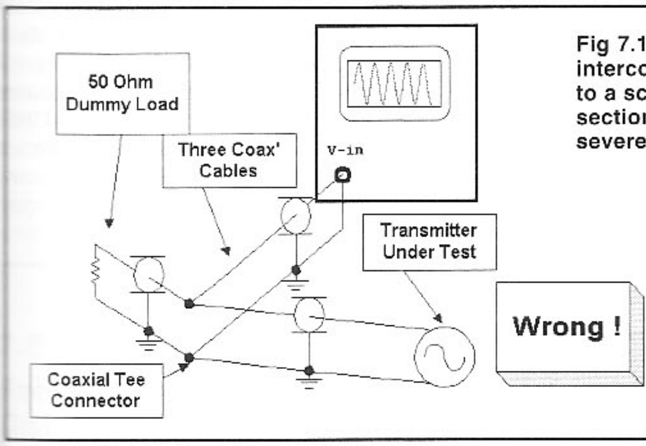


Fig 7.17—Random interconnection of a load to a scope with coax sections can produce severe error. See text.

$$P(\text{watts}) = \frac{V_{RMS}^{\;2}}{R}$$

$$P(\text{watts}) = \frac{V_{peak}^{\;2}}{2 \cdot R} = \frac{V_{pk\_pk}^{\;2}}{8 \cdot R}$$

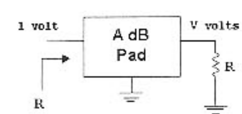
$$P(\text{mW\_50\_}\Omega) = 2.5 \cdot V_{pk\_pk}^{\;2}$$

Fig 7.18—Equations used to calculate power from oscilloscope readings.

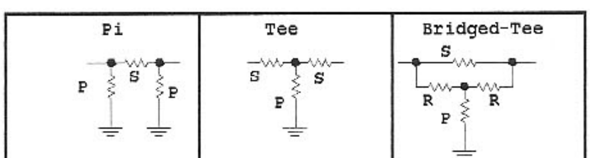
Attenuators

Attenuators form one of the most important and useful components in any RF measurement laboratory. They become especially useful in a home lab, for they are easily constructed and calibrated with dc. Once available, they can be used to extend numerous measurements to lower or higher levels.

Three attenuator network forms are shown in **Fig 7.19**. The series resistors have value S and the parallel ones a resistance P. When terminated in R (usually 50 Ω) at the right, the input resistance looking in at the left will also be R. This condition leads to a mathematical relationship between the series and the parallel resistors. Setting the attenuation, which established the output voltage V for a 1 V input, allows another equation for each type to be derived. Solv-



$$V = 10^{\frac{-A}{20}} \qquad V < 1$$



Pi	Tee	Bridged-Tee
$P = \frac{R \cdot (1 + V)}{1 - V}$	$S = \frac{R \cdot (1 - V)}{1 + V}$	$S = \frac{R \cdot (1 - V)}{V}$
$S = \frac{2 \cdot R^2 \cdot P}{P^2 - R^2}$	$P = \frac{R^2 - S^2}{2 \cdot S}$	$P = \frac{R^2}{S}$

Fig 7.19—Schematics and design equations for three popular attenuator forms. To design any of the attenuators, pick R and A in dB and calculate V with the formula shown. The parallel resistor, P, and the series one, S, are then calculated with the equations.

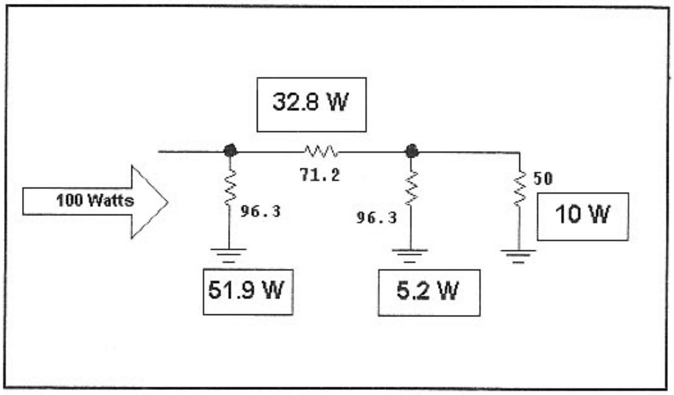


Fig 7.20—Power dissipated in each resistor is shown for a 10-dB pad with 100 W applied. The numbers are also percentages.

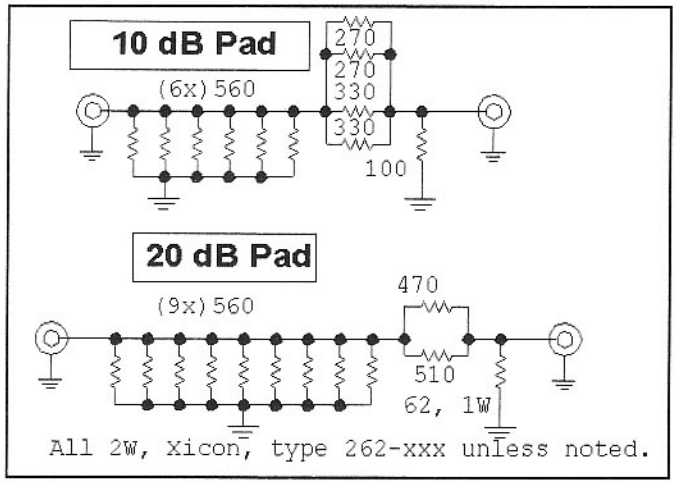


Fig 7.21—Power π attenuators built by Fred, W2EKB. The resistors were purchased from a catalog of electronic components. The 262-xxx numbers are from a Mouser catalog.

ing these two produces design equations included in Fig 7.19. If we pick A=4 dB as an example, V will be 0.631, resulting in P=221 Ω and S=24 Ω for the pi, P=105 Ω and S=11.3 Ω for the Tee, with P=85.5 Ω and S=29 Ω for the Bridged-Tee.

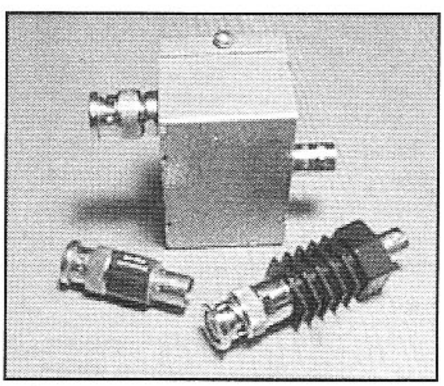
The pi and Tee both use three resistors and are equally useful. The pi may fit better with switches (described below.) The bridged-Tee uses 4 resistors, but only two need changing for different attenuation, so it tends to be a good topology for further design of adjustable circuits.

The dB attenuation value is a weak function of the actual resistance values, allowing one to use close 5% values to build practical circuits. For example, building the 4-dB Tee pad mentioned earlier with 12-Ω series resistors and a 100-Ω shunt would produce a 4.2 dB attenuation with input resistance of 50.3 Ω.

One must use care when designing attenuators for use with transmitters delivering modest to high power. **Fig 7.20** shows a 10-dB Pi-pad with 100 W applied to the input. The powers dissipated in the output and the three resistors are shown. The numbers are also the percent of the input power dissipated in each element. Clearly, for example, over half of the applied power appears in the first resistor. Analysis of this sort will allow one to design higher power attenuators. Two high power pads built by W2EKB are shown in **Fig 7.21**. When asymmetric pads are built, the input should be carefully labeled.

Care must be exercised when picking resistors for attenuator applications. Many power resistors use wire wound construction, often hidden in ceramic, making them too inductive for RF use. Carbon composition and the various types of film resistors are generally suitable for RF through UHF.

Fixed attenuators have two significant applications for the experimenter. The ob-



This photo shows some typical terminators. The smaller two are surplus with power dissipation of 2 and 5 W. The box is a homebrew terminator containing four paralleled 200-Ω, 2-W resistors.

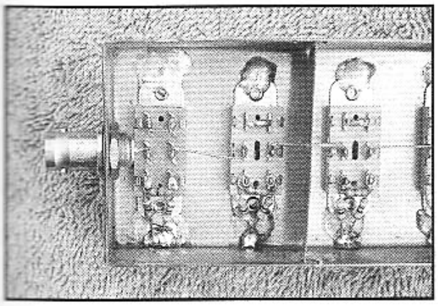
Power Resistors at Radio Frequency

Several resistors were evaluated with an HP-8714 network analyzer to establish suitability for use as RF terminations or as elements in attenuators. The results are shown in the attached figure. The RF measurements were performed at the listed measurement frequency, establishing RF resistance and inductance. A maximum frequency was then calculated as that where the inductive reactance goes up to half of the RF resistance. Clearly, traditional wire-wound power resistors are not suitable as RF loads.

Spec. R = RF R L at RF

Part	Spec. R	DC R	RF R	L at RF (μH)	Freq. for RF Measurements (MHz)	Maximum Frequency (MHz)
A	50	52.2	51.5	6.4	3.5	0.64
B	100	99.6	99.4	0.194	30	40.8
C	50	56.2	59	0.24	30	19.6
D	47	47.2	49	0.0099	250	395
E	47	46	47	0.0095	250	394

Parts Key
A: Lectrohm 10W Wirewound
B: Tru-Ohm 20W Non-Inductive
C: Sprague KookOhm 5W
D: Xicon 3W Metal Oxide
E: Allen Bradley 2W Carbon Composition



A step attenuator for the HF spectrum is easily built with slide switches and 1/4-W resistors. This design used a brass box with the switches soldered in place. This was hard on the plastic parts of the switches, making hardware mounting preferred.

vious one is that of reducing power by a known amount. The other, often just as important, is that they serve to establish impedance level. Assume you have a receiver that you wish to use for measurements in a 50-Ω system. The input impedance of the typical receiver is rarely well matched to 50 Ω, even if it was designed for use with a 50-Ω antenna. However, inserting a suitable pad alleviates the problem. If, for example, we used a 10-dB pad, the return loss we would measure looking into that pad would be 20 dB when the output was left open, and would improve with any termination. A 20 dB return loss corresponds to VSWR=1.2. The receiver with the pad is now a good impedance match. We often use pads in the output of signal generators to force a clean output impedance.

The Step Attenuator

The core of many basement RF laboratories is a step attenuator. Although simple and even relatively inexpensive, such an instrument allows measurements performed at a modest level where they are easy to be extended to other powers where they are difficult. A step attenuator consists of fixed pads that are attached to a switch. Each pad is then switched in or out of a signal path, allowing a total attenuation to be established by adding the individual values.

Several switch types can be used. Most of our experience is with inexpensive DPDT slide switches (eg, CW Industries G and GF series) found in component catalogs. Use those with mounting flanges. The attenuator is built in a trough-like enclosure fabricated from scraps of PC board material. Rectangular holes are cut for the switch handles and the switches are mounted in a line. The resistors are then mounted with very short leads. Short wires are attached to extend one switch section to the next. WB6AIG and WA6RDZ described this circuit in a classic paper and found that vhf performance was improved by adding shields across the center of each switch section.[7] Shriner and Pagel built a similar design, using shields between sections. Bramwell did a more recent version of this classic where careful attention was devoted to maintaining the 50-Ω characteristic impedance within the trough structure.[8] The last two papers are included on the CD that accompanies this book.

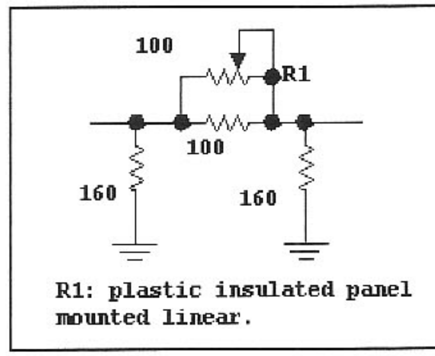


Fig 7.22—Continuously variable attenuator with about a 4-dB range.

It is sometimes useful to have a continuously variable attenuator. **Fig 7.22** shows an attenuator that we have used in the output of homebrew signal sources. This design has an attenuation ranging from 2.5 to 6.7 dB. The exact range obtained will depend on the surrounding impedances. This design will certainly be compromised at higher frequency.

7.5 MEASURING FREQUENCY, INDUCTANCE AND CAPACITANCE

Frequency Determination

The frequency counter is now the most practical instrument for measurement of frequency up to a few GHz. The ICs that form the basis for such measurements are available in virtually all digital formats and are all relatively easy to use in this application. We are not going to say much about counters in this chapter, but note that a simple and inexpensive counter was described in Chapter 4. That circuit could be adapted for general purpose counting with little additional effort. We have built versions with 2, 3, and 4 digits, but would recommend 6 or 8 for a general purpose lab instrument.

Counters are available in all price and frequency ranges, often at less than $100 for a unit that will count to beyond 1 GHz. Resolution at low frequency is typically 10 Hz, although some units are found that will count to 1 Hz. The higher resolution is easy to build if one is brewing an instrument for the home lab and is well worth the extra effort for those cases when it is needed. We find that 1 Hz or better resolution is especially useful when measuring parts for use in crystal filters.

Battery operation is also a useful feature. A battery operated counter will let one build numerous simple instruments that can then be carried into the field for antenna measurements.

It has become popular to build counters from single chip microprocessor of the PIC or BASIC Stamp variety. This offers some hardware simplification and a useful task to use as a mechanism to learn more about the use of these processors. It also offers some unusual possibilities. For example, one kit vendor (Small Wonder Labs) offers a frequency counter designed for use with low power transceivers where the counter uses no visual frequency display. Rather, when a button is pushed to start the circuit, the frequency is counted with the value sent to the user in Morse code. In another design, a single digit display is used sequentially to read up to 8 digits, offering economy and simplicity.[9]

Some inexpensive counters only have high (1 Hz) resolution when digital circuits are investigated. An example is from RadioShack, catalog no. 22-306. A simple interface can be built that will accept a low level RF input while providing a TTL or CMOS compatible output, shown in **Fig 7.23**. This circuit will usually function

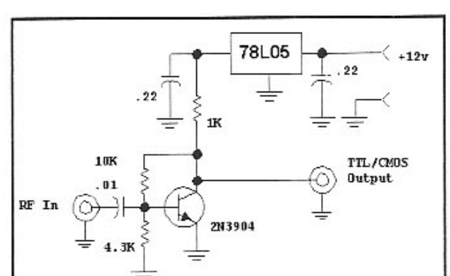


Fig 7.23—Low-level RF to TTL/CMOS converter for simple counting applications. The 10kΩ/4.3kΩ resistive divider sets the collector voltage at about 3 times the 0.7 V emitter-base offset, guaranteeing bias in the active region.

with inputs of −20 dBm at 10 MHz or −10 dBm at 30 MHz (substitution measurements from a 50-Ω signal generator).

Using counters is not difficult, although it is always useful to read the manual. The longer gate times, sometimes controlled by the user, will provide greater resolution, but with longer time between readings. Many counters have a 50-Ω input impedance, but also have a maximum input power. Don't over drive them for it will damage the counter. Instead use an attenuator after you have used a power meter to examine the source you plan on counting. Often a 10X 1-MΩ oscilloscope probe works very well at the input to a counter, even with 50-Ω inputs.

Some users will attach a small link to a piece of coax driving the counter. The link is then used to sniff the circuit under test. This may work, although the power to the counter is not well defined. Moreover, if the source is rich in harmonics, you can end up counting a harmonic instead of the fundamental. Don't try to use the counter as a spectrum analyzer; it may be an interesting measurement anomaly, but it is not a good method.

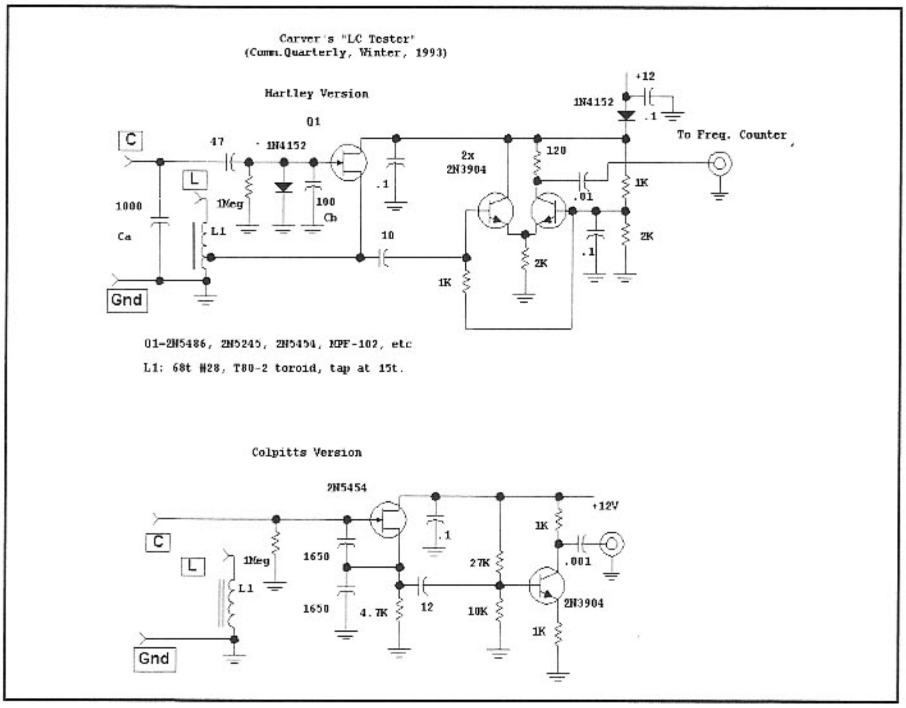


Fig 7.24—"The LC Tester" offered by Bill Carver, W7AAZ, in *Communications Quarterly*, Winter, 1993. The two modes essentially offer identical performance. See text.

L and C measurement

The traditional experimenter measured inductance or capacitance by finding a resonant frequency with a dip meter. An unknown C was paralleled by a known inductor, the combination was "dipped," and the value was calculated. An identical process measured an unknown L. But the frequency measurement was poor, leaving the experimenter wondering about his or her results.

The same general method can be applied today, but the dipper is completely eliminated from the measurement. A stable LC oscillator is built in its place with a buffer to drive the frequency counter. Unknown components are then attached to the oscillator to alter its frequency. This produces the data needed to obtain the L or C. This method was the basis for a simple instrument built by Bill Carver.[10] This instrument is shown in **Fig 7.24**.

The instrument is ruggedly built with three binding posts labeled *L, C* and *Ground*. Operation always begins by placing a wire between the L and the C terminals and measuring frequency. Calibration can then be performed, (not necessary with every measurement) by placing a known capacitor between the C and the ground posts with L and C still shorted. A good calibration value would be a 1000 pF 1% capacitor. A new frequency is measured with the CAL cap in place. From the two frequencies and the known CAL capacitor value, the net fixed capacitance and the inductance value can be calculated, C_0 and L_0.

Measurements are now performed by parallel or series connections of the unknown components. The instrument is turned on and an initial frequency, F_1, is counted. An unknown inductor is then attached either between C and ground, or between L and C. The new frequency, F_2, is measured. Knowing Co, a new inductance can be calculated. If a series connection was used, $F_2<F_1$ and L is found by subtracting L_0 from the measured value. If a parallel connection was used, $F_2>F_1$, and the measured L will be less than that of the one connected. The same resonance concepts give capacitance results.

Carver's original circuit used the Hartley circuit shown. When we breadboarded the circuits, we also tried a Colpitts variation that allowed larger capacitor values to be determined. Either large C or small L between the C and ground terminals can cause oscillation to cease. The two topologies are otherwise identical.

Once the instrument is built and in use, a computer or calculator program can be written to expedite calculations. Carver includes such a program in his paper.

Carver's paper also mentioned a preliminary version of the instrument that used a PIC microprocessor, performing the counting function as well as the calculations. Since that paper was published, a similar instrument has arrived on the market by Almost All Digital Electronics, which is offered as an easily constructed kit. (**www.aade.com/**)

The experimenter has a choice of building his or her own LC Tester or purchasing the kit from AADE. Whatever the choice, the modern experimenter cannot afford not to have this measurement capability. This instrument essentially replaces the classic grid dipper for the electronics experimenter of the 21st century!

7.6 SOURCES AND GENERATORS

A signal source or generator is needed to align and adjust most projects, or for most fundamental circuit experiments. Two or more are required for many other experiments. In this section we present a wide variety of sources

The one instrument that would do most of what we need is a "lab quality RF signal generator." But there is more to the name than suspected. A traditional signal generator used for servicing consumer radio and TV receivers consisted of a wide tuning range oscillator covering all input and intermediate frequencies that the service person might encounter. These boxes usually had modulation capability, allowing the user to align AM receivers. However, they did not qualify as the lab quality instrument we really want. A good signal generator will have the mentioned characteristics plus accurate frequency readout, a 50 Ω output impedance, low phase noise, low spurious outputs close to the carrier frequency, excellent buffering, good isolation from the power supply, and un-compromised shielding. Long term stability and low harmonic content are also useful, but are not dominant specifications.

Many instruments presented as *signal generators* don't qualify because they can't be made weak enough to test a receiver that is useful for communications. When you disconnect the generator, but perhaps attach an antenna to a receiver under test, the generator is still heard. The problem may be poor shielding, signal conduction through the power supply, or both.

The sources we describe in this chapter will not result in a lab quality instrument. Rather, we will describe specialized sources that will satisfy some of these needs, but not in one instrument. The surplus market is full of good equipment that will fulfill many of the experimenter's needs. Having one of these is useful as a means to calibrate home built sources.

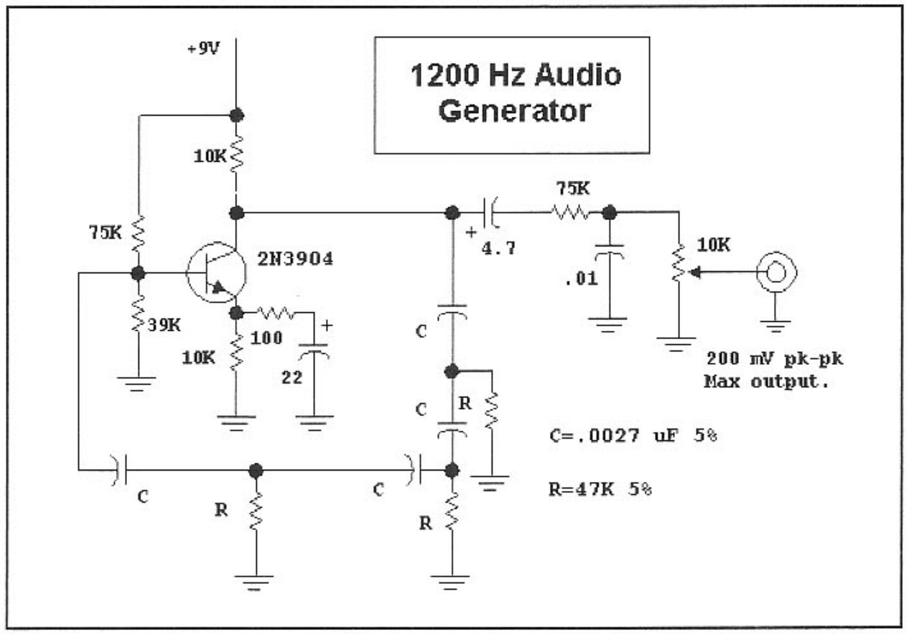
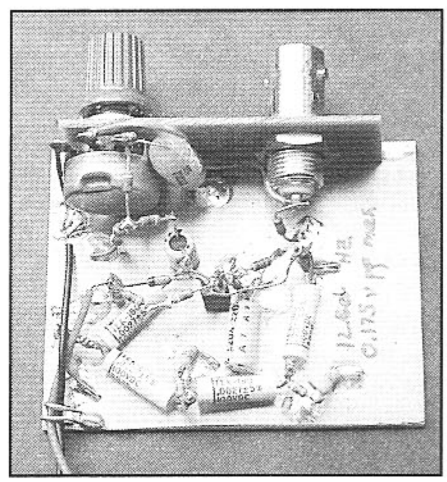


Fig 7.25—A simple audio generator for transmitter testing.

Audio sources

A whistle or a few words spoken into a microphone may serve as a first functionality test for a phone transmitter. However, we need something more when testing a transmitter. A simple generator is shown in **Fig 7.25**. This circuit is battery operated from a 9-V cell, a very convenient feature when seeking good isolation from other sources. This topology is called a phase shift oscillator. The transistor is biased as an inverting amplifier (180 degree phase shift) with a voltage gain of just under 50, established with feedback and biasing. The output is routed back to the input through an RC high pass filter. Oscillation occurs at the frequency where the total phase shift is 360 degrees, half provided by the frequency dependant feedback network. Output is extracted from the collector, attenuated, low pass filtered, and applied to an output level control. This oscillator operates at 1200 Hz. There is nothing special about the exact component values. This one was based upon a handful of 0.0027 µF capacitors on hand. The measured 2nd harmonic was 40 dB below the desired output.

The circuit is built on a small scrap of circuit board material. Another board scrap is mounted to the original to hold a BNC output connector and a level control.



A simple audio oscillator for transmitter testing.

The maximum output from this circuit is about 200 mV peak-to-peak, more than that supplied by most microphones. Use begins by attaching a microphone to a speech amplifier in a transmitter. A few words into the microphone while looking at the amplifier output with an oscilloscope allows us to set audio gain. The microphone is then replaced with the audio oscillator with the level set to establish the same maximum level. This can then be used for extended bench testing.

Fig 7.26 shows a two tone generator useful for testing SSB transmitters. One generator operates at about 650 Hz while the other is at 1650, a non-harmonic higher frequency. A Wien Bridge circuit, shown in the inset, is used for each source. Each oscillator had a measured third harmonic that was only suppressed by about 30 dB, so suitable active low pass filters are added. The two signals of about 3 V peak-to-peak are added and attenuated in U3A while U3B provides a 600-Ω output impedance.

There are many other ways to build audio sources including some special function generator ICs. These are circuits intended to generate triangle and square waves, but with modifications to also approximate a sine wave. The Exar XR-2206 and the Maxim MAX038 are examples. A DSP-based solution is also presented in Chapter 11.

The two-tone generator is attached to a transmitter mic input and the level is adjusted for the desired output. One tone can

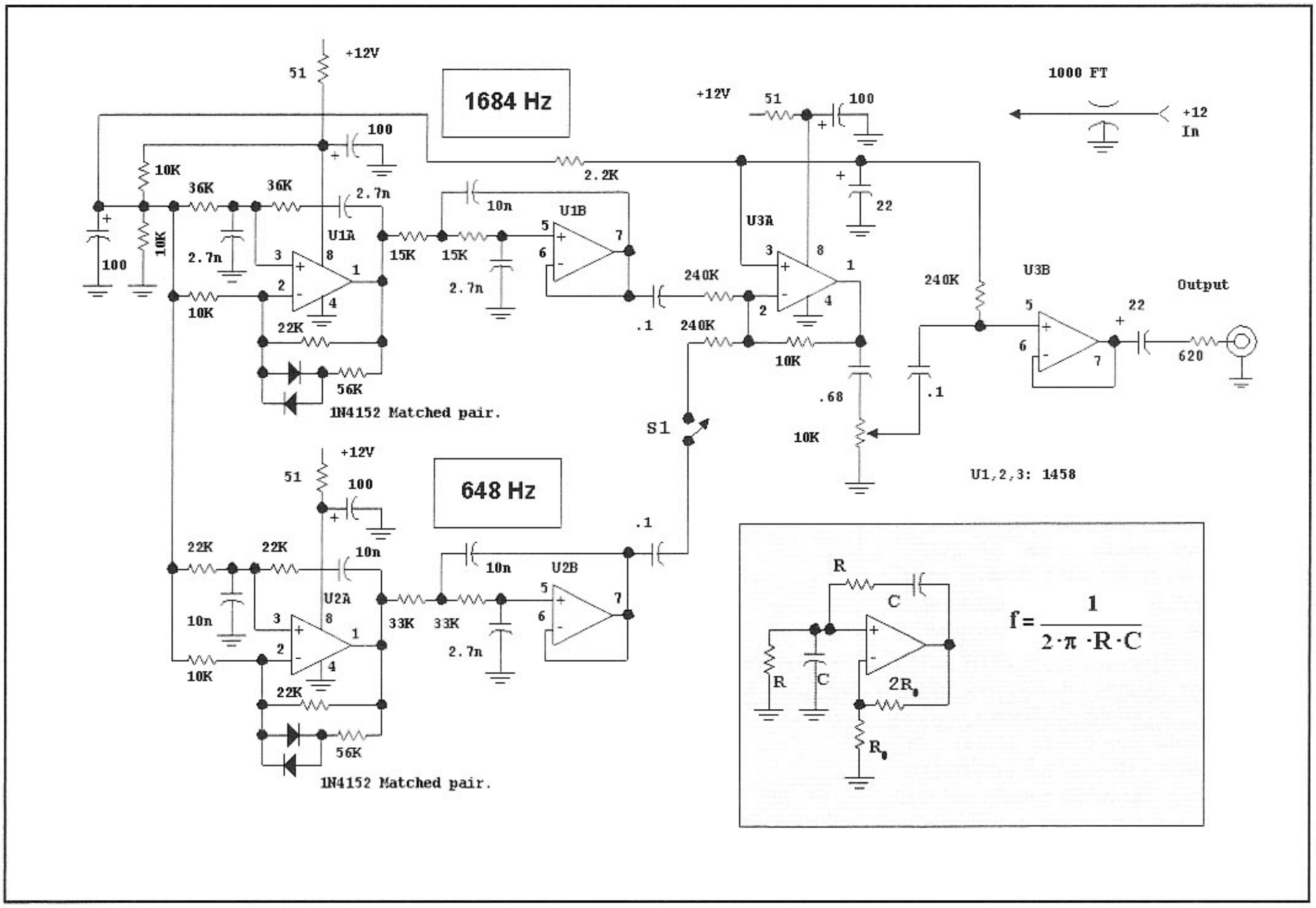
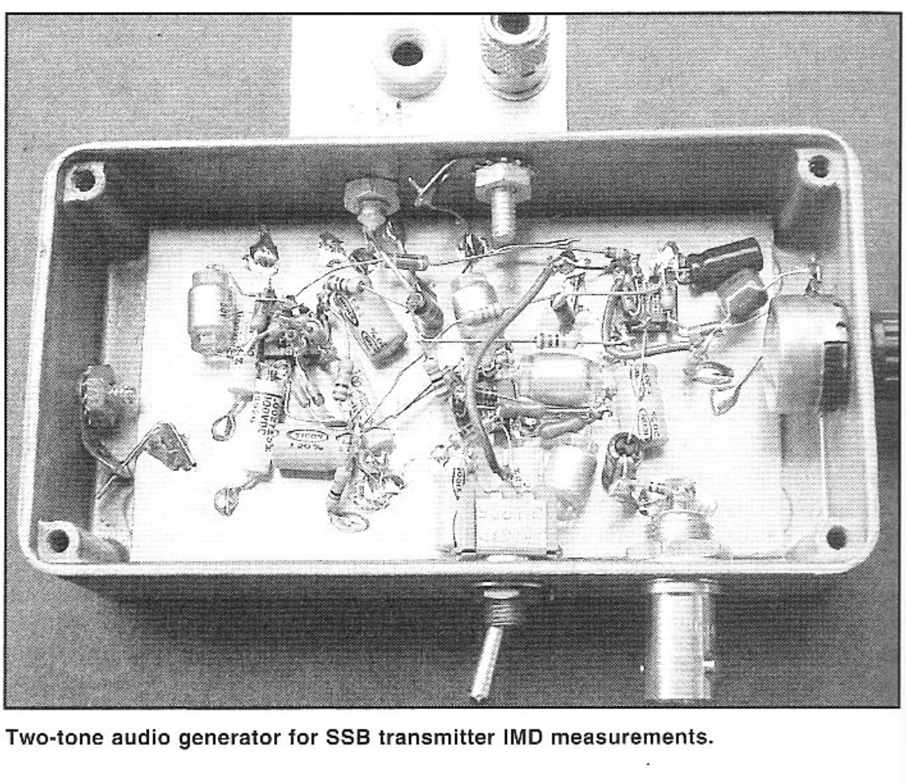


Fig 7.26—Two tone audio source. Each oscillator uses a matched pair of diodes with matching done with a DVM in the diode test position. Matching was done to 10 mV.



Two-tone audio generator for SSB transmitter IMD measurements.

be turned off with S1 so single tone power can be measured. With two tones present, the composite signal moves through all stages of the SSB transmitter to produce a two tone output that can be observed with an oscilloscope or spectrum analyzer, or ideally, both. The intermodulation distortion products (or flat topping in a 'scope display) are then the result of distortion in the transmitter. It is vital that the source be free of these products.

General Purpose RF Sources

No lab is complete without a general purpose RF generator. Like power supplies and step attenuators, one more is always useful. The early sources we built consisted of an LC oscillator, link coupled to a feedback amplifier and pad to provide an output power of +5 dBm or more, enough to drive a diode mixer. Although the design was useful, the buffering was sometimes inadequate, especially for crystal filter testing. The addition of a com-

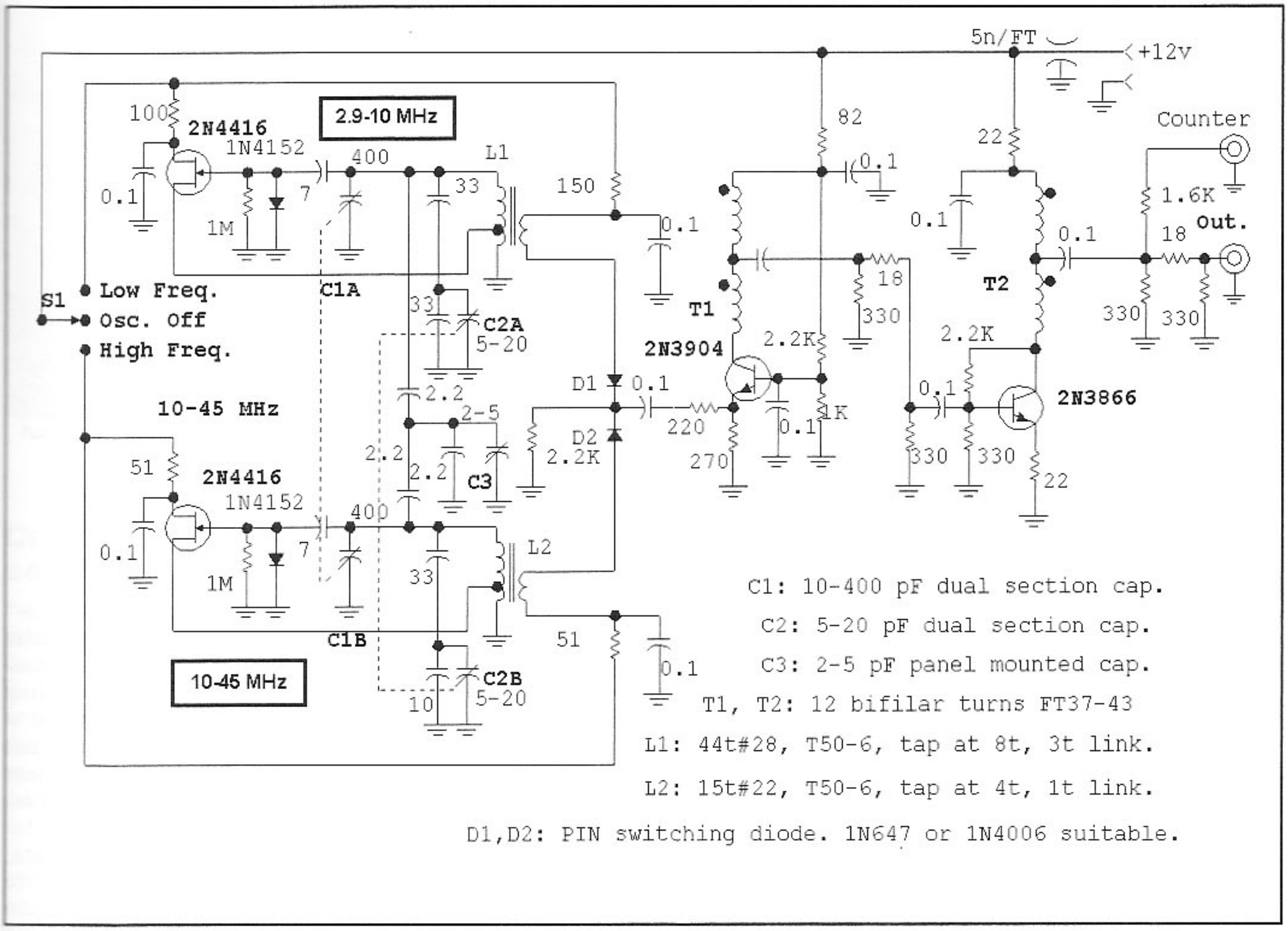
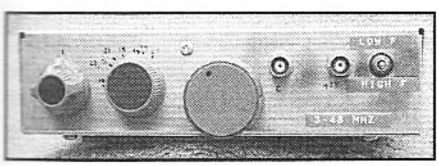
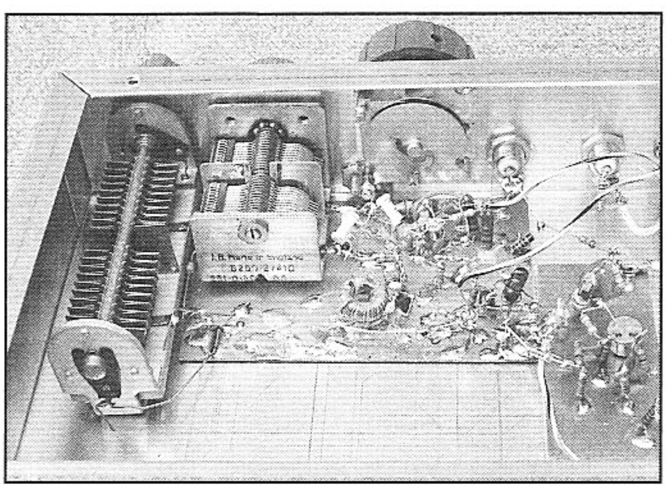


Fig 7.27—General purpose oscillator tuning the range from 3 to 45 MHz in two ranges. See text for details.



General purpose RF source tuning from 3 to 45 MHz.



Inside view of 3-45 MHz RF Generator.

mon-base buffer amplifier has solved these problems.

A wide tuning range oscillator is shown in **Fig 7.27**. Two Hartley oscillators are tuned by dual section capacitors, C1 and C2. The Hartley topology is optimum, for it uses an inductor tap to obtain feedback. As such, all resonator capacitance can be variable, providing the widest possible tuning range. This circuit achieves 2.9 to 10 MHz in one of the oscillators with the other tuning 10 to over 45 MHz. C1 is the main tuning while C2 provides bandspread. Even greater bandspread is provided by C3, now a single section capacitor. C3 is coupled to both resonators in such a way that the inoperative oscillator does not disturb the other. The bandspread afforded by C3 allows the generator to be set accurately, even at the high end.

Another scheme that could provide bandspread would add a variable capacitor from the cathode of the PIN diode switches to ground. This capacitor would then be switched between oscillators with the diodes. But because it reaches the resonator through a link, it tunes over a proportionally smaller range.

Band switching is performed with a SPDT toggle switch with a center-off

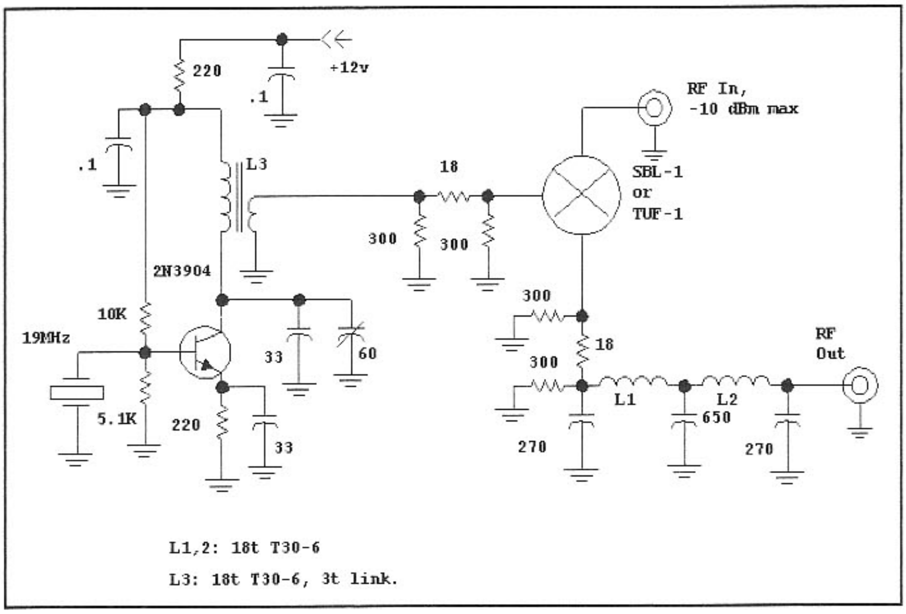


Fig 7.28—Signal Generator Extender.

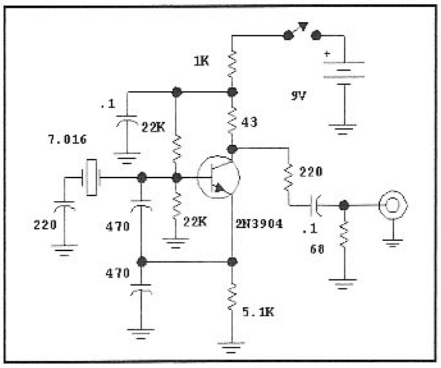


Fig 7.29—Crystal controlled oscillator used for receiver testing. This unit doubles as a spectrum analyzer calibration source with a 7-MHz output of –20 dBm.

position. The "off" mode has been useful to completely extinguish a signal without changing other settings. The toggle switch applies power to one of the two oscillator circuits and biases a PIN diode that routes the output to the buffer amplifiers. A high speed switching diode (1N4152, etc) should *not* be substituted here, although many rectifier diodes work well. The diode switch output is applied to the common base buffer amplifier, preferred over a common emitter amplifier or an emitter follower. The output stage is a 2N3866 common emitter feedback amplifier with a 3-dB pad. A bit of the output energy is tapped and supplied to an auxiliary output feeding a frequency counter. The output power from this source is around +10 dBm on both bands, although it is not as flat (constant amplitude with frequency) as we would like. But this is also the case with many very good signal generators, such as the classic HP-608 series and the surplus URM-25 line. A PIN diode leveling loop could be added to solve this problem, but should be done with considerable care, for such loops can generate additional problems.

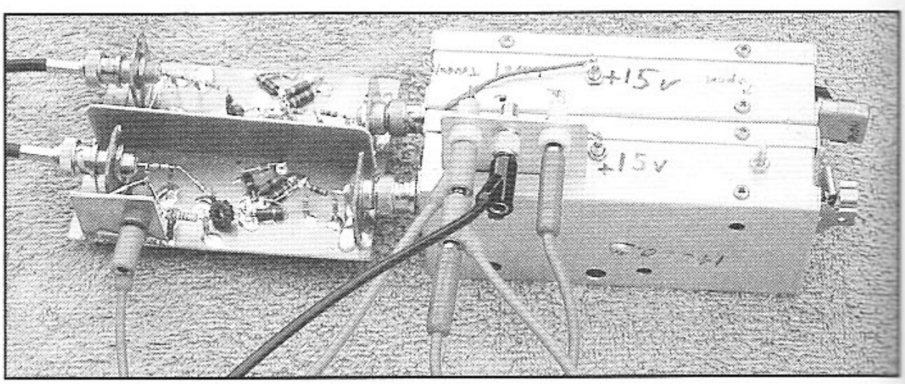
Single band variations of the oscillator of Fig 7.27 have been built, all with a virtually identical circuit. One version was built into the remains of a surplus BC-221 frequency meter. The tuning range was purposefully restricted to about 30 kHz around 5 MHz. The oscillator is then used for crystal and crystal filter measurements.

These RF generators do not lend themselves to easy duplication owing to the unique components used. The junk box is the basis for much of our test gear. If dual section capacitors are not available, single range versions of this oscillator may be built. The circuitry is generally simple, tolerant of component value changes, and inexpensive except for the variable capacitors. These oscillators are running at moderately high power with over 10-V peak-to-peak across each resonator. While this is ideal for low phase noise, it means that one *cannot* casually substitute a varactor diode in these circuits.

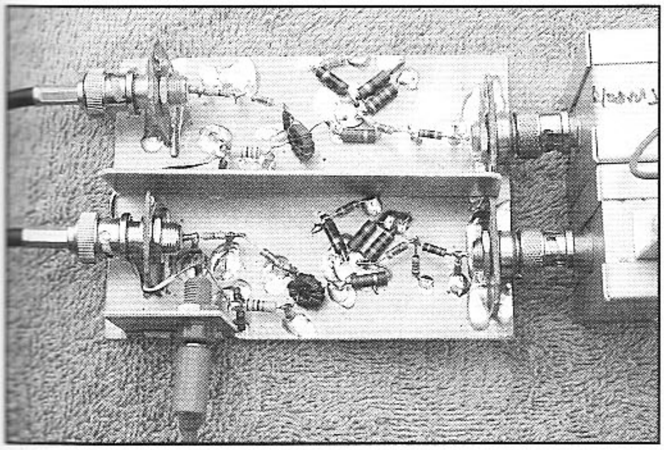
The dual range source has been used for numerous applications, ranging from antenna measurements to IMD testing.

There are many generators found on the surplus market that cover ranges from 10 MHz upward. Examples include the HP-608 and HP-8654. A useful lower range may be added with the "extender" shown in **Fig 7.28**. An available 19 MHz junk box crystal was used in a crystal controlled oscillator driving a diode ring mixer. The signal generator is applied at the input above the crystal frequency and at a level of –10 dBm or less. The mixer output is attenuated in a pad and low pass filtered. This unit is especially useful, for the original generator amplitude calibration is retained with a 9-dB offset. We have also used this same box as an audio source. A 19-MHz VXO can then be used in place of a signal generator. The low pass filter following the mixer has a cutoff just above 10 MHz, the maximum output frequency for this box.

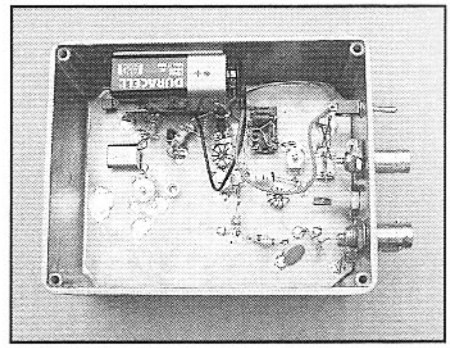
A useful variation of this instrument would use a high level (+17 dBm LO) mixer. More 19 MHz LO energy would be required. This would then allow operation at 10 dB higher levels, needed for some IMD measurements.



Outside view of matching crystal controlled RF sources used for receiver testing. The outboard amplifiers provide the higher signals needed for testing mixers and high-level amplifiers.



Close up view of outboard amplifiers for IMD testing.



An off-the-shelf 14.318 MHz color burst crystal becomes a convenient RF source for the 50-MHz band. Built by KA7EXM.

Crystal controlled sources

Most of the careful receiver measurements we do require good stability in both the receiver and the equipment used to test it. The ideal (affordable) solution uses crystal controlled test oscillators. **Fig 7.29** shows a general purpose source that was originally built as a spectrum analyzer calibration source. The crystal chosen lies within the 7 MHz amateur band, so it serves well as a general alignment tool. The harmonics at 14, 21, and 28 MHz are also useful. The 7 MHz output is –20 dBm. This unit is built into a Hammond 1590B box with a battery contained on the inside, providing the ultimate power supply filtering.

VHF experimenters are always in need of a source to test their equipment, and a crystal controlled oscillator will often serve this need. **Fig 7.30** shows a source using an inexpensive, standard "color burst" crystal to generate signals at 7.16 MHz and at 50.125 MHz. The marked crystal frequency is 14.318 MHz. This is frequency divided in a 74HC74 divider circuit to produce a squarewave at 7.16 MHz. Some low pass filtering strips most of the harmonic energy away for use at 7 MHz. The 7th harmonic of the square wave is extracted with a double-tuned circuit to provide the needed source for the 6-m band. This source was built by KA7EXM.

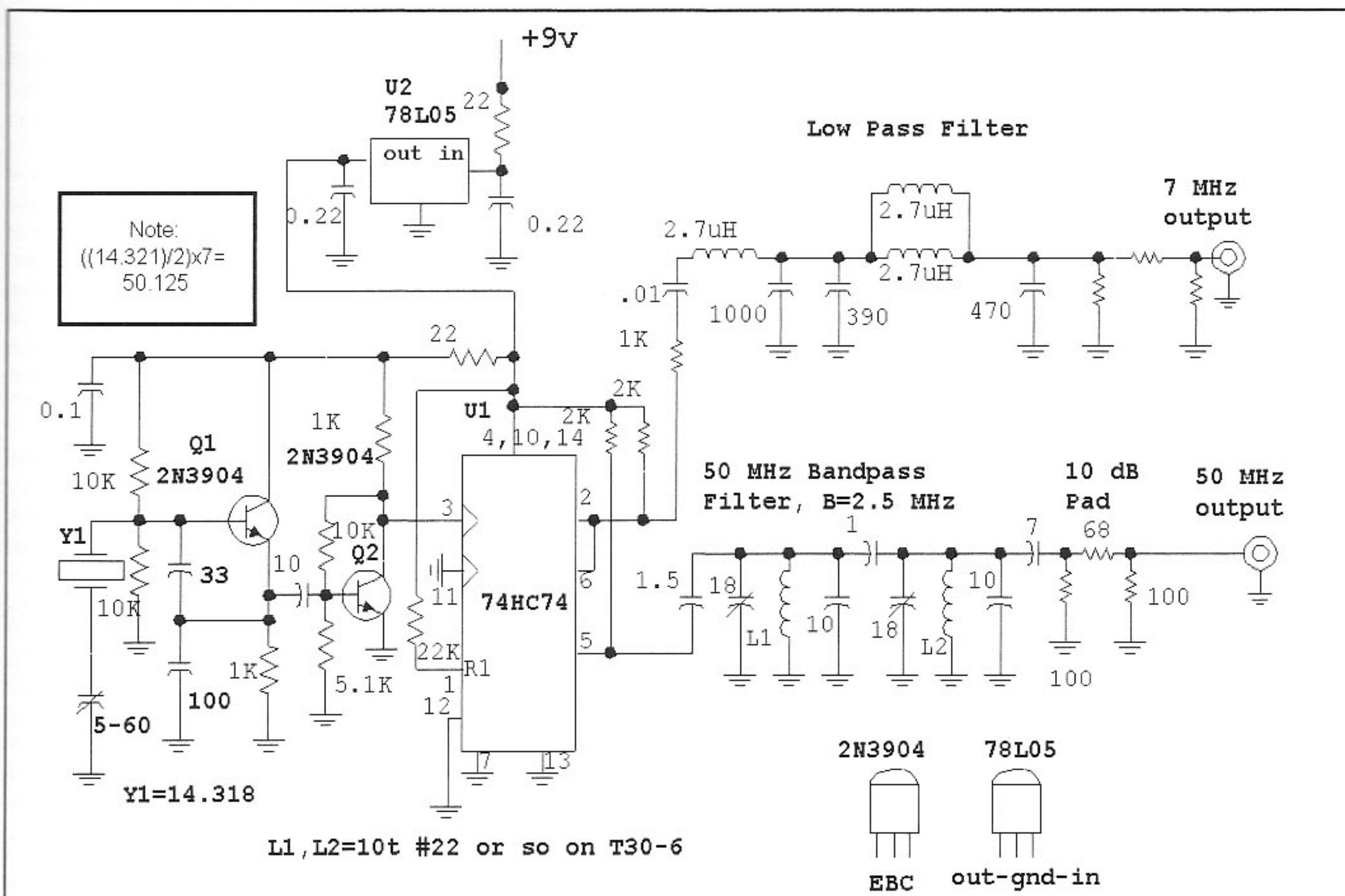


Fig 7.30—Crystal controlled source providing output on the 7 and 50-MHz bands.

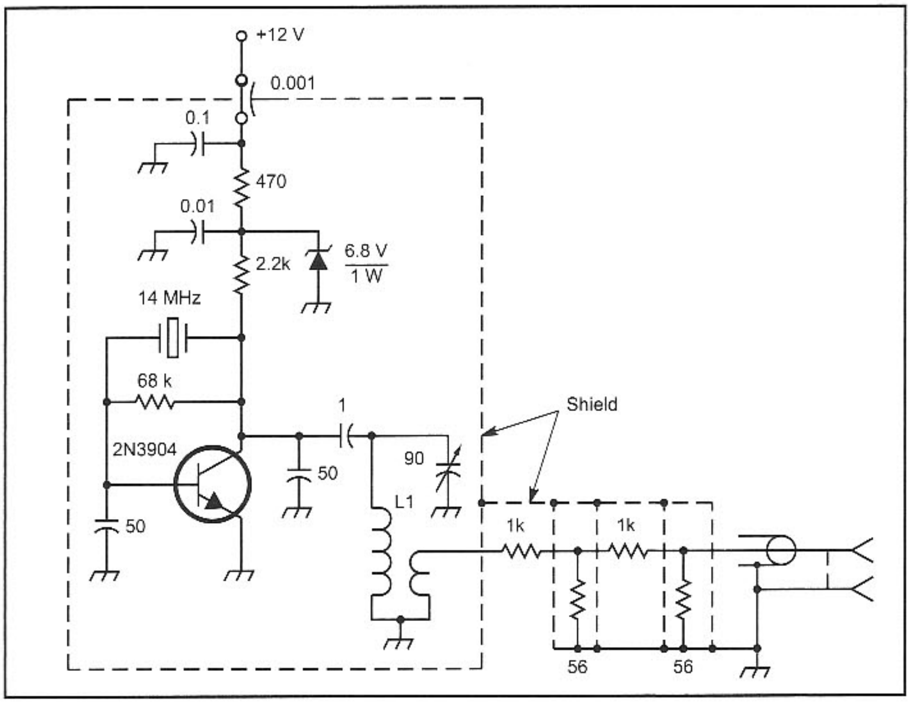


Fig 7.31—Crystal controlled oscillator for receiver MDS measurements. The output is set for about –100 dBm. A builder may wish to add a small resistor or an inductor between the feedthrough capacitor and the 0.1 µF capacitor. A few turns on a ferrite bead should work well. L1 is chosen for resonance at the crystal frequency—the one or two-turn link provides output.

A Weak Signal Source for MDS measurement

The source shown in **Fig 7.31** is similar, but has considerable attenuation included within the box. This unit is predominantly used as a weak signal source for receiver *minimum detectable signal* (MDS) measurements. The oscillator is built at one end of a narrow box fabricated from scrap PC board. Shields are then added with sections of attenuation between. The attenuation is then set to establish the desired output. Levels around –110 to –100 dBm are good, for they are easily attenuated further in a step attenuator to drop to the MDS levels often found with HF receivers. After the output is set, a shield lid is soldered to the box. If double sided board is used, be sure that the inside and outside are attached to each other at the lid.

The unit is calibrated with a CW receiver and another signal generator. The crystal oscillator is tuned with the receiver (AGC off) and the output is measured with an audio voltmeter. The signal generator is then tuned to the same frequency and the amplitude is adjusted until the same output response is observed. The level is noted in your notebook and is marked on the outside of the MDS generator.

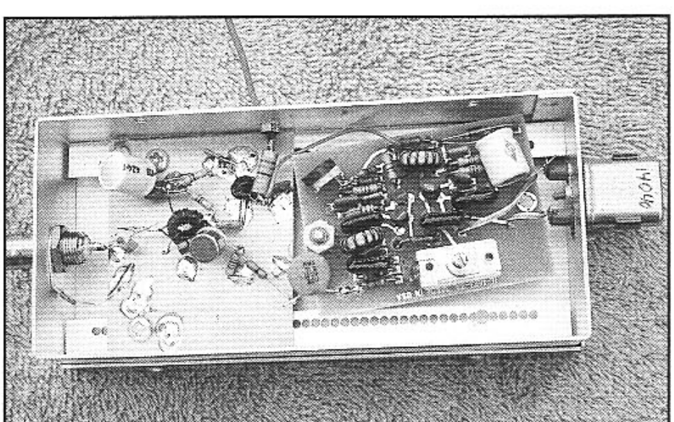
MDS can then be measured with the oscillator and a step attenuator. The source is attached to the receiver (AGC still off) and the receiver is tuned to the generator frequency. Attenuation is then added to weaken the source. The source is momentarily turned off and the noise level is noted in the audio meter. The source is turned on again and the attenuation is adjusted until the meter response is 3 dB above the noise. The strength of the source less the added attenuation is then the MDS.

It's worthwhile to listen to the receiver as a means for growing a "calibrated ear." Although this signal is weak, it is clearly audible above the noise, even if the bandwidth is a kHz or more. As receiver bandwidth drops, the MDS will become smaller but there is less difference between the measured MDS and that perceivable by ear. When running a relatively wide SSB bandwidth, a signal at measured MDS sounds rather loud. It is not surprising that many weak signal VHF enthusiasts including EME aficionados will use the wider bandwidths when QRM is not an issue.

Crystal Oscillators for Intercept Measurements

Having measured receiver MDS, we now need "loud" generators that can be used to measure the strong signal performance, the receiver input intercept, IIP3. The measurement was described in detail for an amplifier or mixer in Chapter 2 and then applied to a receiver in Chapter 6. The basic source we use for receiver testing is shown in **Fig 7.32**. The crystal oscillator is carefully tailored to operate with current limiting, avoiding the Q degrading voltage limiting. The following buffer has an input impedance dominated by a single resistor, but then operates as a limiter, developing an output substantially independent of drive level. That output is low pass filtered and attenuated in a 6-dB pad and then applied to a common base output amplifier, picked for good reverse isolation.

We use two identical versions of the source of Fig 7.32, usually separated by about 20 kHz. The sources are always checked ahead of each use, confirming power and match between units. The output level chosen is 0 dBm for each source. These are usually applied to 6 dB-pads and then to a 6-dB hybrid combiner. The combiner, described later, is a return loss bridge used in a different way. The hybrid output is attached to a 15 MHz low pass filter and then to a step attenuator. This setup, shown in **Fig 7.33**, provides signals of –12 dBm per tone and lower. The role of the hybrid is to add the two signals while preventing the output of one source from reaching the other. If the output from one oscillator reached the other, inter-



Inside one of the crystal controlled RF sources.

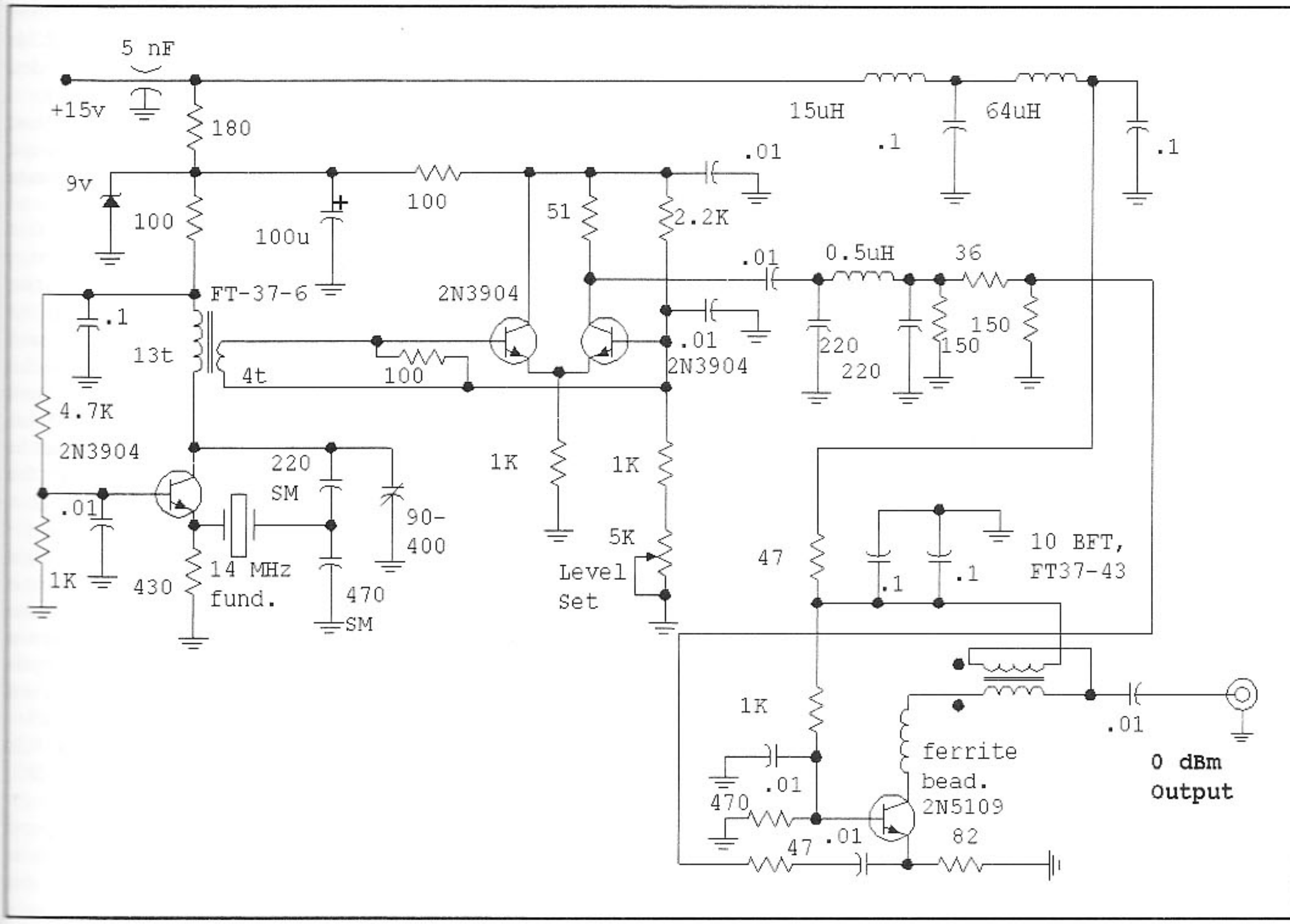


Fig 7.32—A source with an output of 0 dBm suitable for receiver testing. See text for discussion.

modulation could occur, creating spurious signals at the same frequencies as produced by the third order IMD that is usually measured with this system.

There are alternatives to the 6-dB hybrid. A 3-dB Splitter-Combiner is sometimes used and can offer excellent performance. Some experimenters will even use a 50-Ω power divider, which preserves impedances but provides no isolation. A 50-Ω power divider consists of three 50-Ω resistors in a "Δ" configuration, or three 18-Ω resistors in a "Y." The 6-dB hybrid is recommended.

Assume that the two generators have crystals to put their frequencies at 14.03 and 14.05 MHz. Tuning to either of these signals produces a large meter response. These signals impinging on the receiver front end will intermodulate, generating distortion products above and below the two desired signals, at 14.01 or 14.07 MHz. These products are created within the receiver, usually in the circuitry ahead of the main IF filter. With the two test signals separated by 20 kHz, the distortion signal will be 20 kHz above the upper desired signal and 20 kHz below the lower one.

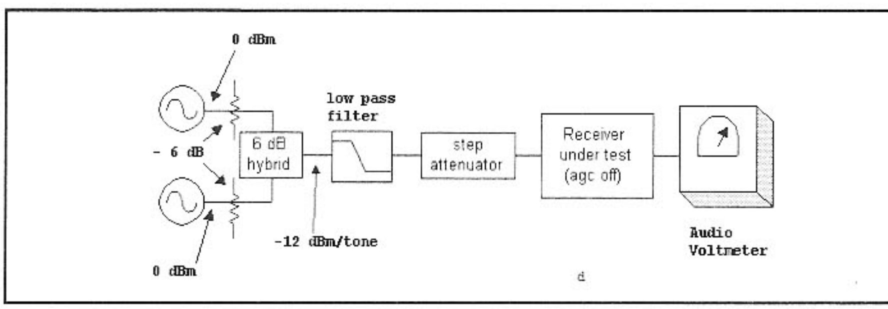


Fig 7.33—Test setup for determining a receiver IIP3, or "input intercept." See details in Chapters 2 and 6.

We tune to either of these IMD responses to measure them, seeing a loud, but still manageable response. Assume an audio signal of 50 mV when tuned to one of the distortion frequencies and that this occurs with the step attenuator set at 30 dB. The signals are then –42 dBm/tone at the receiver antenna terminal. But how strong is this response compared with the input signals? We find an answer by tuning the receiver to one of the main tones and increasing attenuation. When the net attenuation inserted is 110 dB, the audio output is again 50 mV. We have increased the attenuation by 80 dB to depress the main signals to the point where they produce the same response as was seen from intermodulation. The intermodulation distortion ratio, IMDR, is then 80 dB. The input intercept is then given by

$$\mathrm{IIP_3}\left(\mathrm{dBm}\right) = \mathrm{P_{in}}\left(\mathrm{dBm}\right) + \frac{\mathrm{IMDR}\left(\mathrm{dB}\right)}{2} \qquad \text{Eq 7.2}$$

For this example, P_{in} = –42 dBm and IMDR=80 dB, so IIP3= –2 dBm.

Let's repeat the experiment, but start with less attenuation. Instead of 30 dB in the beginning, start with 24-dB attenuation to apply signals that are 6 dB stronger. The response at the distortion frequencies is now much larger, significantly more than the 6 dB increase in the main tones. Assume that it's about 400 mV in the audio voltmeter. We record this level and then tune the receiver to one of the main signals and increase the attenuation. After adding 68-dB attenuation, for a net attenuator setting of 92 dB, we observe 400 mV of audio. The applied power is –36 dBm/tone and IMDR=68 dB, so **Eq 7.2** predicts IIP3= –2 dBm.

This example illustrates the utility of the intercept concept. If we know the input intercept for the receiver, we know what the response will be to any input signals. If we allow the mathematics to get a little more complex, we can even predict the response to input signals of unequal amplitude.[11]

Let's say that this receiver had MDS of –139 dBm, a reasonable sensitivity for a CW receiver with a bandwidth of perhaps 500 Hz (NF=8 dB). The two-tone DR would then be

$$\mathrm{DR}\left(\mathrm{dB}\right)=\frac{2}{3}\cdot\left(\mathrm{IIP}_3\left(\mathrm{dBm}\right)-\mathrm{MDS}\left(\mathrm{dBm}\right)\right)$$

Eq 7.3

or, 91.3 dB in this example. But what does this mean?

The meaning of two-tone DR is clarified with a more direct measurement, still using the example receiver we have been examining. First, we use our weak signal source with the step attenuator to measure MDS. Assume that the receiver gains are set to produce an output of 10 mV with the weak signal source. When we turn the source off, the level drops by 3 dB to 7 mV. Receiver AGC is still off and we don't touch any of the gain controls.

We now replace the weak source with the two tone generator setup of Fig 7.33. We tune the receiver to one of the distortion product frequencies and adjust the attenuator until we get the same response we saw with the MDS measurement, 10 mV on the meter. We tune the receiver to one side and the other of the distortion product to be sure that the response drops to the noise floor of 7 mV. This happens in our example with the attenuator at 36 dB, which places a strong signal of –48 dBm at the receiver input. We record these levels in our notebook and then retune the receiver to one of the strong tones. (It's a good idea to *not* have the headphones on during these experiments!) We now add attenuation until the response from a strong tone is again 10 mV. This occurs with a total attenuation of 127 dB. This is 91 dB lower than the signals that produced the distortion responses.

This experiment has illustrated the real meaning of receiver two-tone dynamic range: DR is the difference between the weakest signal we can hear with that receiver and the strength of one of a pair of signals that will produce intermodulation distortion at the same level as that minimum. This is a severe test, but it is measurable with carefully built test equipment.

The high attenuation levels needed for DR measurements, especially the direct one, may be intimidating. It's hard to obtain over 100 dB of attenuation, especially in casual homebuilt designs. For this reason, an indirect measurement is often easier. That is, measure IIP3 with two moderately well shielded strong sources with levels that can be confirmed with a power meter, a spectrum analyzer, or terminated oscilloscope measurement. Perform an independent measurement of MDS with a special generator you have built for just that purpose. Then calculate DR from **Eq 7.3**. It is, however, best to work with weaker "strong" signals, for most receiver mixers will then be "well behaved," as defined in Chapter 6.

The procedure we recommend eliminates the MDS measurement, replacing it with a noise figure determination. This will be discussed later.

Component Intercept Measurements

While the receiver builder may wish to perform IIP3 and MDS measurement to obtain DR, the designer is equally interested in evaluation of component parts of a receiver or transmitter. The two tone source is again used, driving the component, followed by a spectrum analyzer. (Analyzers and their design are described later.) The test setup is given in **Fig 7.34**. Frequency spacing is adjusted as needed for the component being investigated.

The test setup is more illuminating than the receiver evaluation, for it is a swept measurement showing the main signals and the distortion products on a calibrated screen, all at the same instant. A step that should always be done is to apply the signal from the step attenuator directly to the spectrum analyzer, prior to inserting the component. Any distortion seen would then be occurring in the analyzer or in the generators. Once a distortion-free test setup is confirmed, the amplifier is inserted, the analyzer input attenuation is readjusted to keep the main signals on the screen, and the data is recorded. The gain of the amplifier (or whatever) is now observed, equal to the change in spectrum analyzer sensitivity needed to keep the main signals in the same position on the screen. We know the input levels, for we measured them before inserting the amplifier, and the IMD ratio can be observed directly on the screen, so the input intercept, IIP3, can be calculated from Eq 7.2. The corresponding output intercept, OIP3, is just IIP3 plus the amplifier gain.

It is very informative at this time to vary the strength of the input tones used to test the amplifier, achieved by adjusting the step attenuator. The desired output signals should change on a dB-for-dB basis with the inputs. However, the distortion products above and below the desired two signals will move on a 3 dB per one dB input change rate. It is not necessary to collect all of the data to actually plot traditional intercept curves, such as were shown in Chapter 2 of this book.

Measurements normally performed with a spectrum analyzer can also be done with a receiver. It will be necessary to put an attenuator ahead of the receiver to control the levels reaching it, always taking care that IMD in the receiver is not dominant. One then proceeds to add an am-

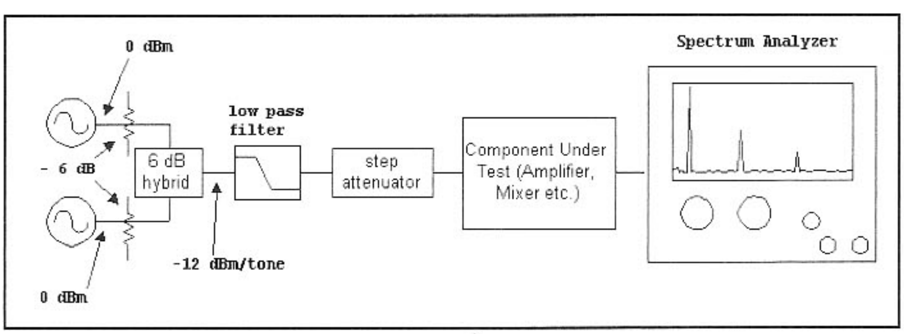


Fig 7.34—Test setup for testing components.

plifier, followed by further attenuation to maintain signal levels at the receiver input. If a receiver is to serve this function, it must have much better shielding and decoupling than it would for normal use, for we don't want signals from our generators to enter the receiver via any path other than the antenna terminal.

It is even possible to test receiver components (mixers, amplifiers, etc) that are part of a receiver while using that receiver for the measurements. Essentially one does intercept measurements as described, followed by a repeat measurement with a fixed attenuator added between stages. If the IMDR does not change when the pad is added, the distortion is occurring before the pad location.

Some components may required larger signals for testing, a prime example being high level switching mode mixers. Such circuits may have IIP3 of +30 dBm or more. To examine such circuits, we place an amplifier after each generator. **Fig 7.35** shows some sample feedback amplifiers while the application is shown in **Fig 7.36**.

Even greater power may be obtained with another stage or by eliminating the output pad. Eventually the point is reached where IMD in other elements may come into play. W7AAZ and the other members of the "Triad" (see Chapter 6) reported seeing IMD in hybrid combiners.

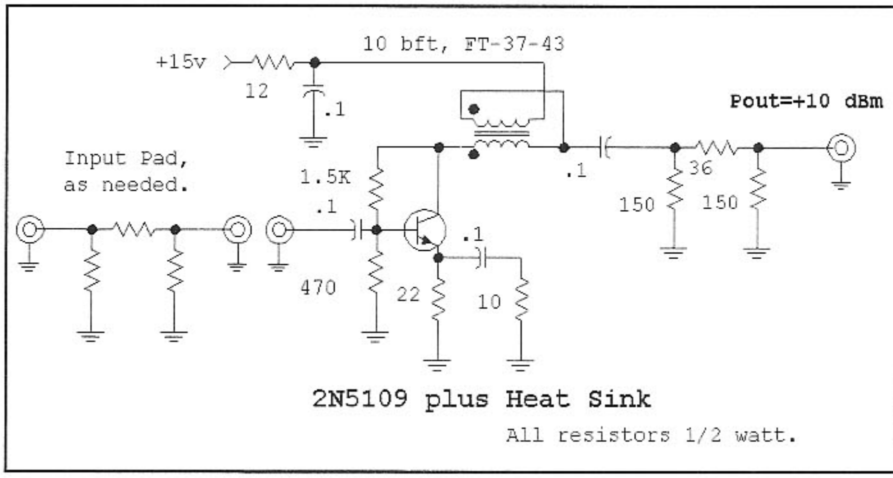


Fig 7.35—Feedback amplifier used following *each* IMD generator to increase the power to +10 dBm per tone. Amplifier gain is 22 dB at 14 MHz, which is reduced to 16 dB with the output pad. Using a 6-dB input pad with the source of Fig 7.8 provides +10 dBm/tone output.

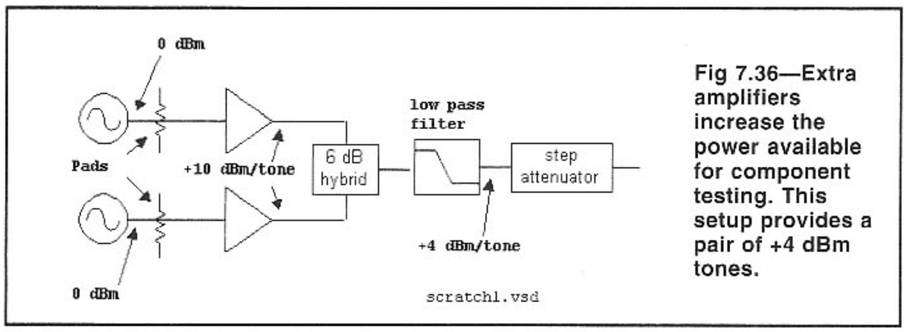


Fig 7.36—Extra amplifiers increase the power available for component testing. This setup provides a pair of +4 dBm tones.

7.7 BRIDGES AND IMPEDANCE MEASUREMENT

We are always interested in measuring impedance, be it for antenna experiments or to set up a termination for a filter. These measurements are difficult with homebuilt equipment, but they are becoming less so with the changing technology we enjoy. Traditional bridge circuits included built-in diode detectors, a restriction that is no longer necessary or even desired.

Shown in **Fig 7.37** is the circuit for a basic Wheatstone bridge. Assume that 1 V is applied to the RF input. If R1 and R2 are equal, point "x" will be at 0.5 V. Point "y" will also be at 0.5 V if the unknown impedance is 50 Ω resistive. A detector between x and y will show no output and a null is detected. If the unknown departs from 50+j0 in either the real or imaginary part, the null is not complete and an error appears at the detector port.

There are two ways that the bridge circuit can be used. The traditional examines the "detector" port between x and y as a place to seek a null. The bridge elements are adjusted to produce the desired perfect null. The alternative places meaning on the indication at the detector port. We will examine both applications here.

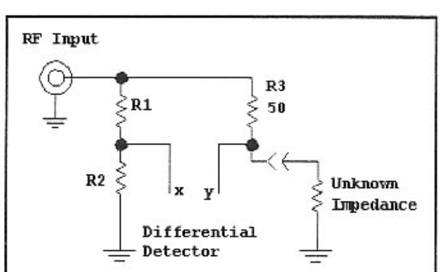


Fig 7.37—A basic bridge circuit.

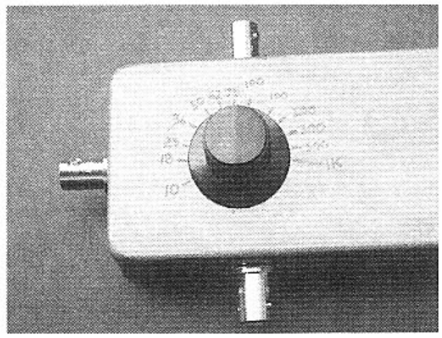
We can form simple bridges with the circuit shown in Fig 7.37. (This one even works with dc.) When all three resistors are 50 Ω (use 51 if building one), the input will appear as 50 Ω to the RF source when the unknown becomes 50 Ω. The voltage between points x and y is roughly the voltage reflection coefficient, which goes to zero for a perfectly matched 50-Ω unknown Z. Such a bridge can be used to tune an antenna or transmatch. We will show some practical examples later.

A useful variation is adjustable. In this form, R1 and R2 are replaced by a 100-Ω pot with the arm serving as "x." Assume the bridge is loaded with 25 Ω as the unknown and the pot is tuned until a null is produced. Analysis shows this to occur when the pot arm is 1/3 of the way up from the ground end.

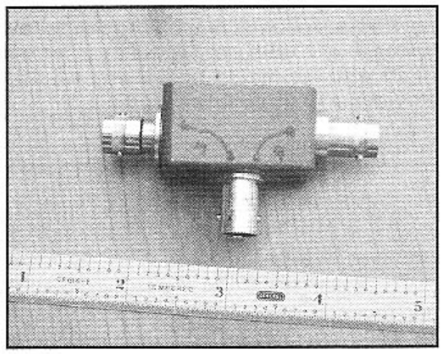
RF bridges with variable resistors have long been popular with the experimenter. The traditional instruments included a built-in diode detector and meter as the null indicator. They suffer a common problem: the sensitivity suffers with low RF drive owing to the threshold voltage presented by most diodes. Measurements that do not rely upon diode detection of a low level RF signal are preferred.

Fig 7.38 shows an RF resistance bridge with an external detector. This circuit was designed to measure RF resistance while using a sensitive power meter, spectrum analyzer, or 50-Ω terminated oscilloscope as the detector. An unknown resistive impedance is attached to the bridge and R1 is adjusted for a minimum response. The bridge is normally driven with a low level source of around 0 dBm. Less power is used when the termination will be an active circuit; more may be appropriate for antenna measurements. When working with antennas, it is useful to alternately tune the signal frequency and pot R1 to get the deepest null.

The instrument was calibrated at 14 MHz with resistors from 10 to 1000 Ω. T1 is wound on a low permeability, low loss core. Primary inductance was about 50 μH, allowing operation down to 2 MHz or less. Transformer T2 is a common mode choke with about 20 μH per winding that isolates the T1 secondary from ground. This bridge had over 30 dB directivity over the HF region. Directivity is the change between the open circuit response and that when the unknown-Z port is terminated in 50 Ω.



Exterior view of RF Resistance bridge after calibration.



Exterior view of return loss bridge.

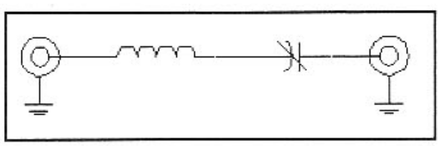


Fig 7.39—Tuned circuits can be added to the bridge to extract complex impedance information.

Performance was flat throughout the lower part of the HF spectrum. However, as the frequency moved toward 30 MHz and higher, the 50-Ω point on the scale moved toward the high R end. Further refinement is required.

A series-tuned LC circuit can be cascaded with the unknown port for the measurement of reactive impedances, shown in **Fig 7.39**. The capacitor (or inductor) is then adjusted to deepen the dip. Repeated R1 adjustment may be necessary. A traditional instrument would have suitable scales, but that is not necessary. Rather, after adjustment of a trimmer capacitor, it could be measured with an instrument like the W7AAZ LC tester or the similar instrument from AADE. See Fig 7.24.

If the resistance bridge is used without the auxiliary tuned circuit, complex terminations will produce shallow dips. It's common to look at the meter scale and erroneously conclude that the impedance has a magnitude close to the value shown. This is rarely a valid interpretation, further justifying the reactance measuring options.

The bridge of Fig 7.38 was calibrated at 14 MHz with a handful of carbon resistors with the values then marked on the panel. While this is handy, it may not be necessary. Consider the variation shown in **Fig 7.40**. This is equivalent to the other bridge at RF where the capacitors are virtual short circuits. However, the design with capacitors can be measured with a digital voltmeter attached to the "unknown" port. The dc measurement tells the user the status of the pot, allowing the RF resistance to be inferred.

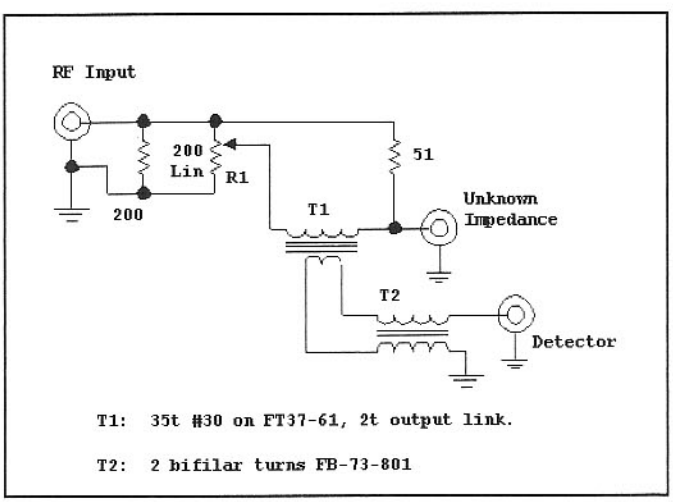


Fig 7.38—RF bridge for HF measurements. R1 is ideally a 100-Ω linear pot, but all we had was 200 Ω. The Clarostat 1/2-inch diameter conductive plastic parts should offer reasonable performance, although we have not used them in this application.

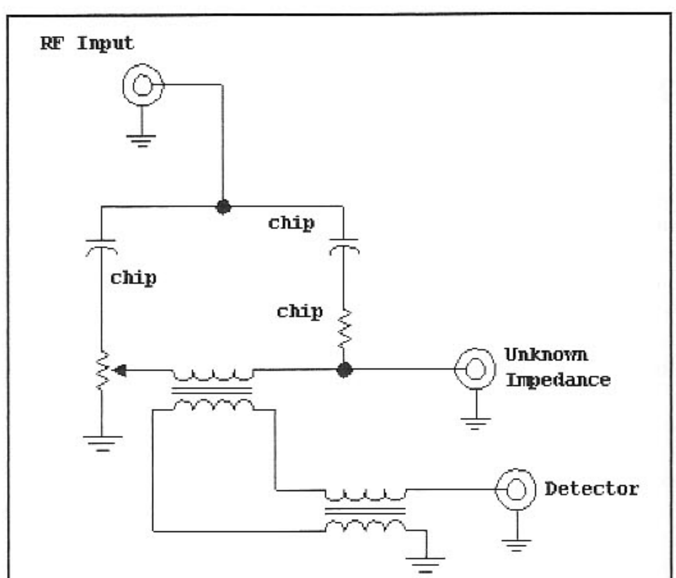
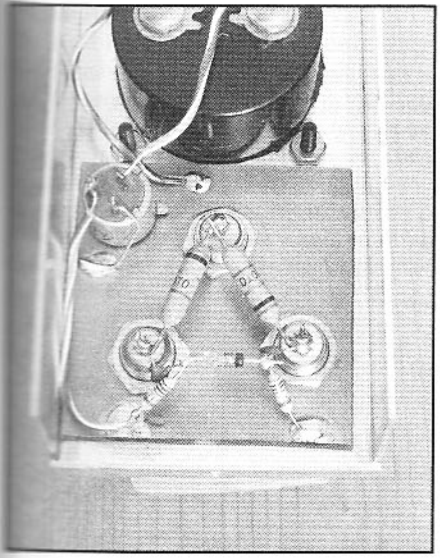
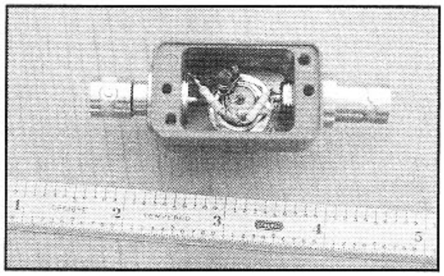


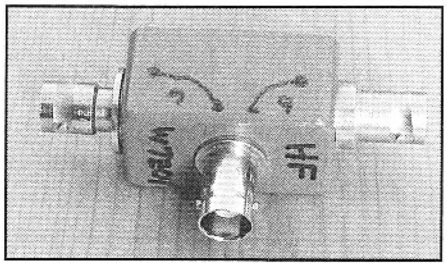
Fig 7.40—Optional variation of the resistance bridge.



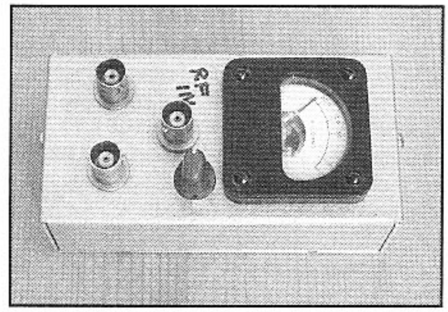
Interior of the RF impedance bridge. Symmetry and short lead lengths are maintained during construction. Long leads are okay with the dc parts of the circuit.



Interior view of return loss bridge. This one is built with 49.9 Ω, 0.1 W, 1% resistors.



Return loss bridge for HF range.



RF impedance bridge with built in meter. A reference must be attached for measurements.

Table 7.1

Frequency (MHz)	*D (dB)*	*O/S (dB)*
2	44	0
10	47	0
20	44	0
30	41	0
50	36	1
144	23	2

There is virtue in the modified bridge: Since a calibrated dial is not needed, it can be built with small trimmer pots with much better RF characteristics than encountered with pots with shafts. This will allow these traditional methods to be extended to higher frequency. (These experiments remain on our "to do" list at this writing.)

Fig 7.41 shows a return loss bridge (RLB,) a circuit with no adjustable elements. The signal coming from the detector port indicates the quality of the impedance match. Bridge use begins with a calibration, which places an open circuit at the unknown port. The detector level is carefully noted in dBm. Then the unknown termination is attached and the new detector level is recorded, again in dBm. The difference between the two in dB is called the return loss.

It is also interesting to observe voltage (rather than power) at the detector port. Assume we observe V_0 when the bridge is terminated in an open circuit and a smaller V_1 when loaded. The ratio V_1/V_0 is termed the voltage reflection coefficient, often signified with an upper case Greek Gamma, Γ. Return loss is related to Γ by RL=–20 log(Γ). Also, Γ is directly related to VSWR by VSWR=(1+|Γ|)/(1–|Γ|). Hence, VSWR=2 corresponds to Return Loss = 9.54 dB and Γ=0.333.

One can use a short circuit instead of an open for calibration. In principle, the two responses will be identical.

There are two frequency dependent RLB characteristics that indicate performance. One is called directivity (D, dB), which is the indicated return loss when a good 50-Ω termination is attached to the unknown port. The other is the dB difference between an open circuit and a short circuit (O/S) at the unknown port. These parameters define the experiments we do when building a bridge.

Table 7.1 shows the results obtained with an experimental RLB. This represents the best transformer(at HF) we found after examin-

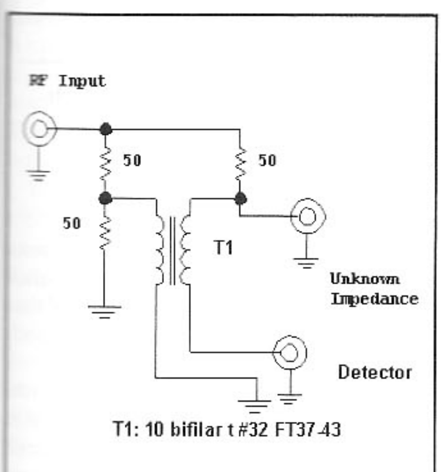


Fig 7.41—A return loss bridge is also known as a 6-dB hybrid. The detector impedance should be 50 Ω for accurate calibration.

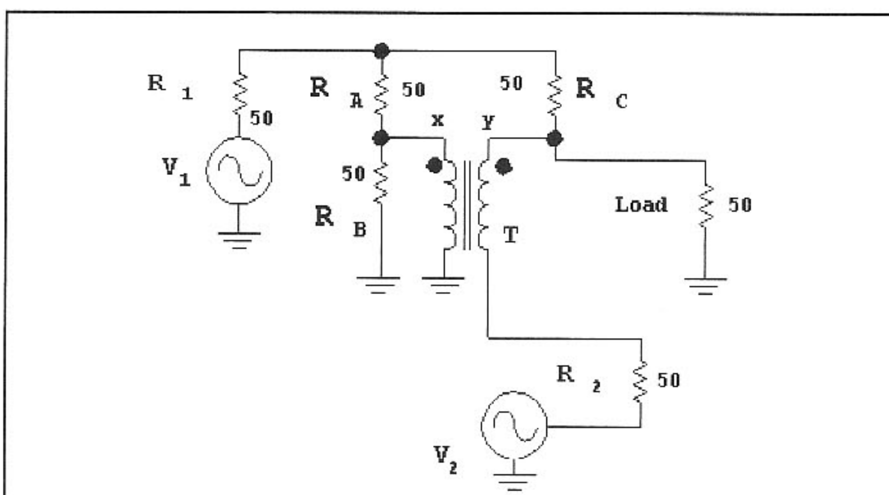


Fig 7.42—A RLB also finds use in combining two signal generators. The power delivered to the load is 1/4 of that available from each generator when the bridge is balanced.

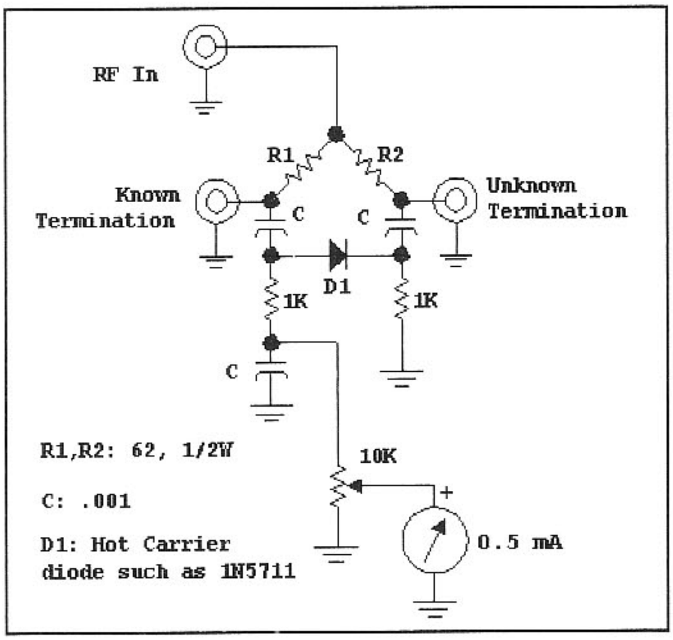


Fig 7.43—A bridge suitable for use through UHF. The symmetry of the schematic should be followed when building the instrument. Drive is 100 mW to 1 W. The "known" termination usually used is a 50 or 75-Ω coaxial terminator.

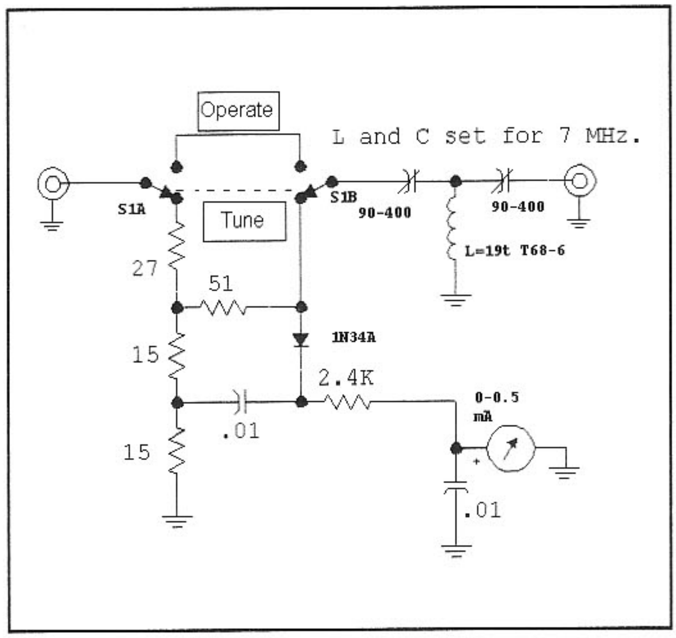


Fig 7.44—7 MHz transmatch and resistive bridge for portable operation. Variable capacitors are screwdriver adjusted, mica compression types. All resistors are 1/2 W. S1 is a DPDT slide or toggle switch. This design is suitable for transmitters up to about 3 W if the tuning is done quickly.

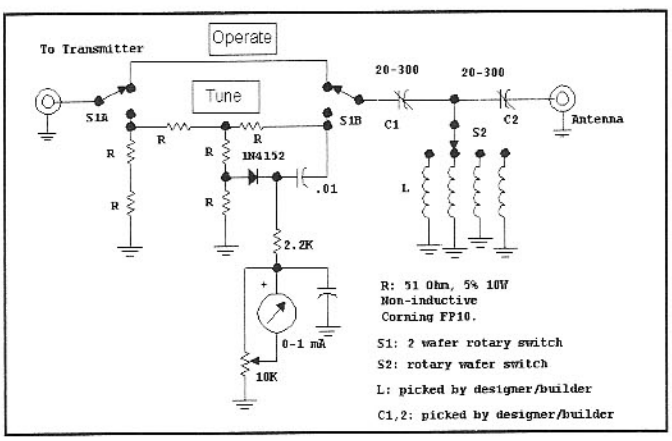


Fig 7.45—Higher power version of a transmatch with a resistive bridge. This unit is rated for 40 W or slightly more for short periods. The topology shown presents a 50-Ω load to the transmitter while attenuating the signal put on the air by 12 dB. If the resistors specified for R cannot be purchased, parallel combinations of 2-W resistors can be used.

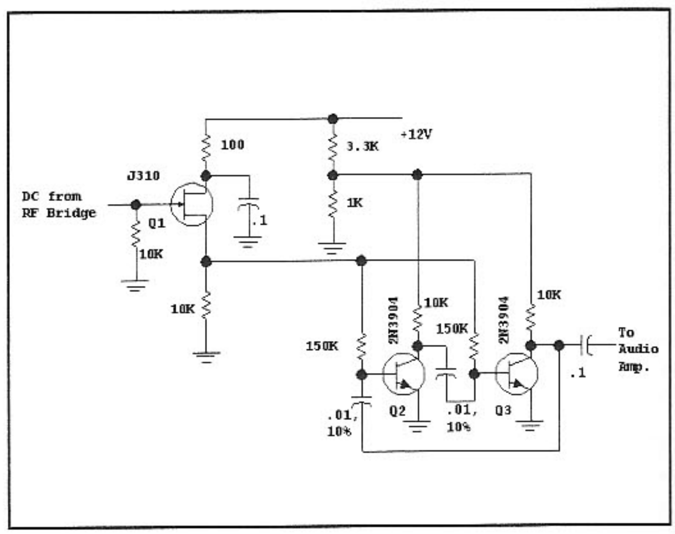


Fig 7.46—Audio meter replacement scheme for transmatch tuning. See text.

ing several. This bridge used 51-Ω, $^1/_{10}$-W resistors and a transformer consisting of 10 bifilar turns of #28 on a FB73-2401. The high permeability core is preferred, providing an inductance of 175 µH for each winding.

A different transformer improved VHF performance at the cost of HF directivity. We saw 30-dB directivity at 144 MHz when the transformer used 5 of the 6 holes in a multi-hole bead, a FB43-5111. This configuration produces an inductance of 8.4 µH per winding.

The hybrid qualities of the return loss bridge are illustrated in **Fig 7.42**. Generator V_1 causes voltages x and y to be equal and in phase if the bridge is balanced. Hence, none of this power ends up in R_2, the impedance of the other source. But V_2 also sees a balanced bridge. The power delivered by V_2 forces the node with R_1 to be at signal ground, so none of the V_2 power ends up in R_1. These characteristics provide the isolation that we need when combining signals from two generators for IMD testing.

A conventional resistance bridge circuit with built-in detector is shown in **Fig 7.43**. This circuit functions into the UHF area, realized by small lead length and careful symmetry. A photo shows the inside of the circuit. This bridge works well at 144 and 432 MHz, as well as the HF spectrum.

A simple resistance bridge with included detector is often used for the adjustment of low power antenna tuners. This is often preferred over an in-line directional power meter, for the transmitter is always properly terminated during tuning. A circuit used with portable transceivers is shown in **Fig 7.44** where the compoments are appropriate for the 40-meter band. The slide or toggle switch is put into the "tune" position to adjust the circuit for best null. It is then returned to the "operate" position. A higher power home station version is shown in **Fig 7.45**. The low power variant uses a germanium diode while a silicon switching diode is used at higher power.

Some builders have used a light emitting diode to replace the meter indicating bridge balance. Performance is poor, especially for low power transmitters, for visual output is zero until about 1.6 V biases the LED. But meters are often heavy, difficult to find, and expensive. Some refined circuits use ferrite transformers for greater sensitivity.

An alternative scheme is shown in **Fig 7.46** where an audio oscillator replaces the visual output. The oscillator, a simple multi-vibrator using Q2 and Q3, is frequency modulated by the bridge signal with the *pitch* becoming higher with greater mismatch. The circuit is used by sending a string of dits into the transmatch. The pitch becomes identical for key up and key down when the match is perfect. The primary purpose of Q1, the JFET input, is to generate a dc offset from ground, so JFET type is extremely non-critical. An op-amp would also serve this function.

7.8 SPECTRUM ANALYSIS

What is a Spectrum Analyzer?

One of the most useful instruments the radio experimenter could have is the spectrum analyzer. Commercial versions are sophisticated and expensive, but excellent examples are beginning to appear on the surplus market. And there are now many available components that allow the enterprising experimenter to build his or her own spectrum analyzer.

The first question we must address is the most fundamental: What is a spectrum analyzer? In the general mathematical sense, the signals we encounter are generally collections of sine waves of the form:

$$A \cdot \sin(2 \cdot \pi \cdot f \cdot t)$$

…where A is an amplitude, f is frequency in Hz and t is time in seconds. We can regard this function as either one of time, t, or of frequency, f. In the most general sense, any function of time has a related spectrum or frequency domain representation. The two domains or viewpoints are related through a mathematical operation called the Fourier Transform.[12,13] Also see Chapter 10 of this volume.

Setting formalities aside, we look at electronic signals in the time domain with an oscilloscope or examine them against frequency with a radio frequency spectrum analyzer. We are already familiar with radio frequency spectra of several sorts, although they may not have been presented as such. A rudimentary spectrum analyzer, albeit un-calibrated, is shown in **Fig 7.47**. We have extracted our communications receiver from normal service and opened it to attach wires to the S-meter, a panel meter indicating the strength of received signals. This voltage is usually derived from the receiver AGC. The receiver is set to a band of interest and the tuning control is attached to a motor through a suitable pulley. The motor also drives a potentiometer that develops a voltage proportional to the frequency. The voltage from the pot indicating frequency is routed to the horizontal axis of an oscilloscope while the signal from the receiver's AGC, indicating signal amplitude, is applied to the 'scope vertical. The result is our spectrum analyzer.

The usual spectrum analyzer is calibrated in frequency, so we know the frequency representing the screen center. We also know the frequency span, the number of kHz or MHz associated with the dot as it sweeps from left to right.

The on-screen vertical position is also calibrated in the laboratory spectrum analyzer. While we obtained a *voltage* from the receiver to apply to the 'scope vertical axis, we calibrate with regard to the *power* related to the signal that developed that voltage. The *top* of the screen is called the *reference level,* leaving the bottom with no special significance. When we see a signal on screen that just reaches the reference level, we know that it has a strength equal to that level. The usual spectrum analyzer displays signals logarithmically, so the calibration will be in terms of a num-

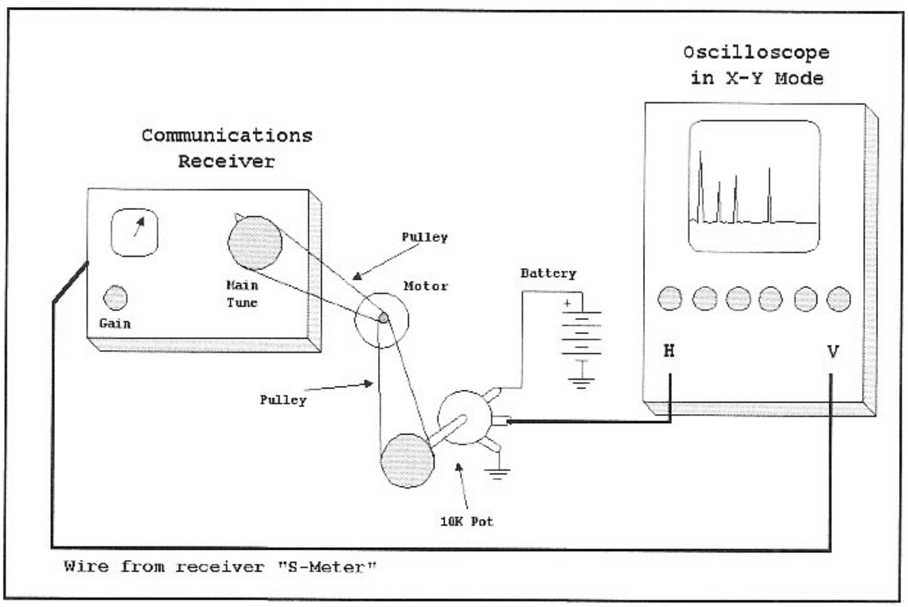


Fig 7.47—A rudimentary spectrum analyzer formed by applying motor drive to receiver tuning and to a pot that generates a voltage that indicates the tuned frequency. This voltage controls the X axis of an oscilloscope. The vertical Y axis is derived from the receiver S-meter circuitry. (Thanks to Bob Bales from Tektronix, Beaverton, OR who suggested this explanation.)

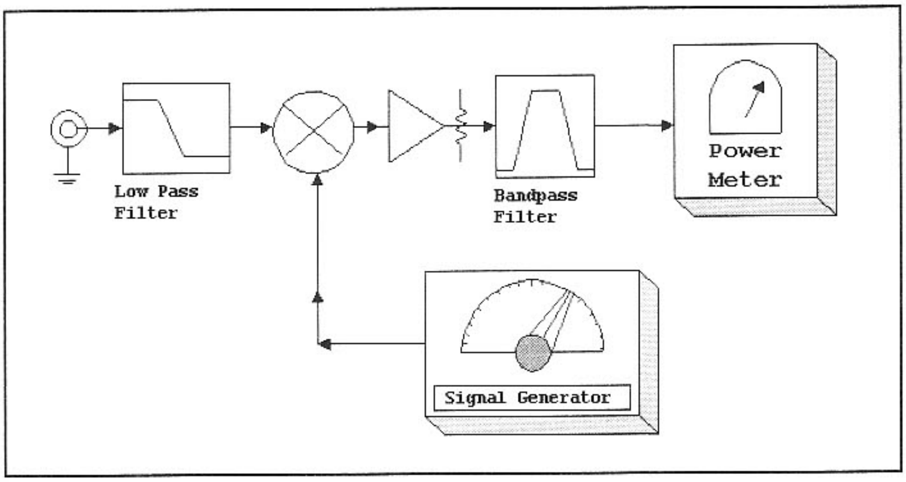


Fig 7.48—Measurement receiver allowing rudimentary spectrum analysis. Although this instrument is presented primarily to illustrate concepts, this unit could be built and would be useful. The amplifier could be a MAR-3 driving a MAV-11 (both from Mini-Circuits) with a 6-dB pad. The mixer might be a TUF-1 or similar part.

ber of dB per vertical division, for the decibel is also a log function. If we have our spectrum analyzer set up for 10 dB per major division, have a reference level of –30 dBm, and see a signal peak two divisions below the top, we conclude that the signal power is –50 dBm.

Spectrum analyzers come in many forms to cover many different frequency ranges. One that we will discuss in more detail tunes from 0 to 70 MHz. Instruments continuously sweeping and tuning from 0 to 2 or 3 GHz are common. Band switching units often tune from 0 to 21 GHz or even more.

The property of *selectivity* in a receiver becomes *resolution* in a spectrum analyzer. Resolution is the ability of an analyzer to resolve two signals that are close to each other in frequency. This is specified by the analyzer resolution bandwidth, RBW, usually equal to the 6-dB width of the filter in use. It is common for high performance spectrum analyzers to have resolution bandwidth selectable from 3 MHz down to 10 Hz. The extremely narrow bandwidth is useful for such tasks as examining 60 Hz sidebands on carriers or for digging way into the noise.

The typical analyzer is not a very sensitive instrument when compared with our receivers. A routine communications receiver might have a noise figure of 10 dB to yield MDS of –137 dBm in a 500 Hz bandwidth. A typical NF might be 25 dB for an analyzer, resulting in MDS of –119 dBm in a 1 kHz RBW. The analyzers are not lacking in dynamic range though. A typical analyzer will have a basic reference level of –30 dBm, but will include an input attenuator with a 60-dB range, allowing the reference level to be extended to +30 dBm, or one watt. A "proper" spectrum analyzer uses a front end that is strong enough to produce no internally generated third order IMD when all input signals are kept below the reference level, or "on screen."

Analyzers the Experimenter can Build

The equipment described above is not the ultimate, but merely the norm, representing what has been common within industry for the past 20 years or more. Equipment offering this performance is still rare in the basement lab of the typical experimenter. It would be a monumental task to duplicate a high performance laboratory instrument. But that is not our goal. Rather, all that we ask is to do some of the measurements, as needed for our experiments, with instruments that are simpler, but manageable. The concepts and some of the methods of the high end instruments will be applied to realize these goals.

Consider a very simple spectrum analysis receiver, shown in **Fig 7.48**. This is based upon a power meter that was described earlier in the chapter in Fig 7.13. The meter measured signals from approximately –80 to +10 dBm. We precede this meter with a 2 MHz wide bandpass filter at 110 MHz center frequency.[14] A remote signal generator is the local oscillator signal for a diode ring mixer followed by an amplifier and pad. The amplifier terminates the mixer and adds gain, allowing smaller signals to be seen. A low pass filter with a 70-MHz cutoff precedes the instrument, eliminating images.

Let's assume that we inject a 30-MHz signal from another generator into the input. We see no output until the *local* signal generator is tuned to 140 MHz when the input signal is converted to the 110-MHz IF. Changes in the input amplitude can easily be observed. We could use this instrument to tune a 30-MHz filter or amplifier. Tuning the local generator to 170 MHz allows 60 MHz to be received, allowing us to measure the second harmonic of the input signal. The 90-MHz third harmonic could be measured with the LO set to 200 MHz except that the 70-MHz input low pass filter would attenuate this response. (We could eliminate the input low pass filter from this instrument to produce an instrument that would allow the entire HF and VHF spectrum to be seen, although results would now be obscured by images.)

We now attach an antenna to the receiver and see considerable energy when tuned to the AM broadcast band around 1 MHz. However, we can't isolate one signal from the other because the 110 MHz bandpass filter is 2 MHz wide. The entire BC band fills the filter at once. This deficiency is altered with a 110 MHz filter with a narrower bandwidth. While crystal filters are possible at VHF, the more practical solution converts the signal to a second, lower IF.

A second problem occurs when we tune the analysis receiver to look at a low frequency: A spurious response is observed even with no applied input signal. This occurs because the LO is at 110 MHz, the intermediate frequency. This is a common characteristic of most swept front-end spectrum analyzers. Improved balance in the input mixer increases mixer LO to IF isolation to reduce the "zero spur" response.

Another subtlety becomes apparent when we actually build the analysis receiver of the figure: The balanced mixer must be reversed from the normal application. Most balanced diode ring mixers, such as the TUF-1 or SBL-1, have transformer coupled "LO" and "RF" ports with a dc coupled "IF" port. If low input frequencies are to be examined, the dc coupled port must be used as the RF input.

The instrument of Fig 7.48 is not a spectrum analyzer, for it lacks a graphic display. This is usually obtained by sweeping the frequency with time in unison with a sweep of the display. This begins by replacing the signal generator LO with a voltage controlled oscillator. The VCO is then swept with suitable circuitry. VCO design was discussed in Chapter 4.

A basic swept voltage generator is shown in **Fig 7.49**, beginning with the integrator circuit of part A. Starting with the capacitor discharged, apply a negative

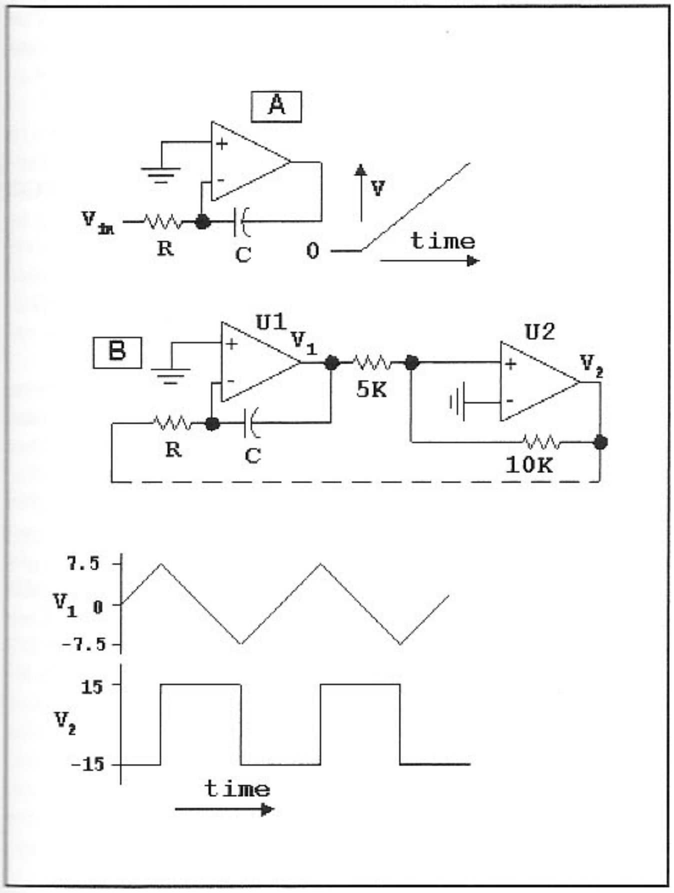


Fig 7.49—Part A shows an integrator circuit. This drives a level detector with hysteresis, U2 in part B. Feedback then creates a sawtooth generator. See text.

voltage to V_{in}. This is coupled to the inverting input, which causes U1's output to begin increasing. But this is coupled back to the inverting input through the capacitor. The equilibrium we require of a closed feedback loop in an op-amp is realized when the U1 output voltage ramps linearly upward. The current in the capacitor then equals that in the resistor, V_{in}/R. Had we applied a positive input we would generate a negative going ramp.

In part B of the figure, we drive the input of the next stage with the ramp. Assume U1 is ramping upward and that the output of U2 is negative against the −15 V power supply. The non-inverting input of U2 reaches 0 when U1's output is +7.5 V, a consequence of the voltage divider action. At this instant, the output of U2 changes state, now "slamming" against the +15 V power supply. If the U2 output becomes the driving source for the integrator input with the dotted connection, we obtain the sawtooth waveform shown for V1.

A practical sweep circuit grows slightly from that described. Diodes provide different slopes for the positive and negative going portions, for we use the left-to-right as the sweep and the other as a retrace. Potentiometers or switched resistors and/or capacitors are added to change sweep

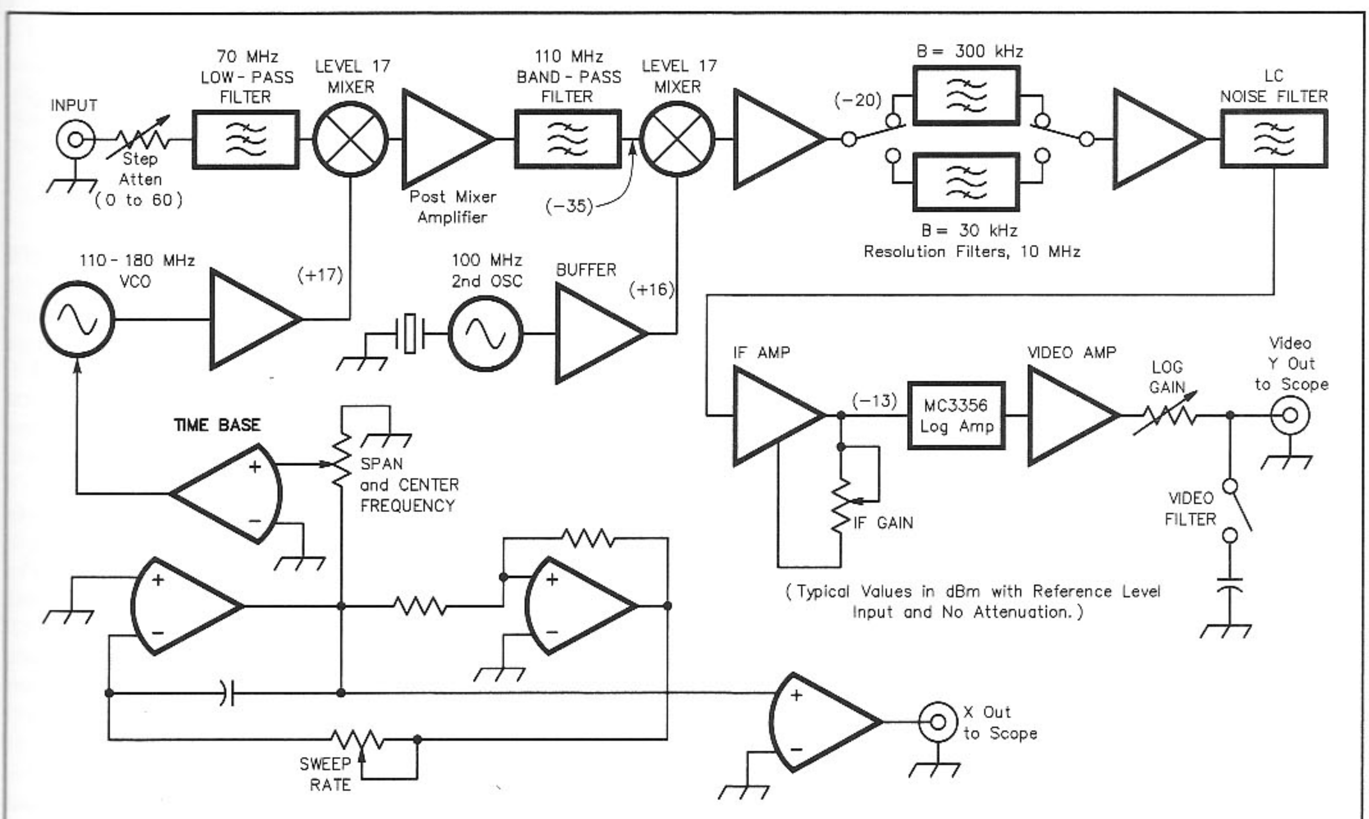


Fig 7.50—Block diagram of a spectrum analyzer the experimenter can build. A practical realization of this design is on the book CD. The 60-dB step attenuator can be an external accessory or built into the instrument.

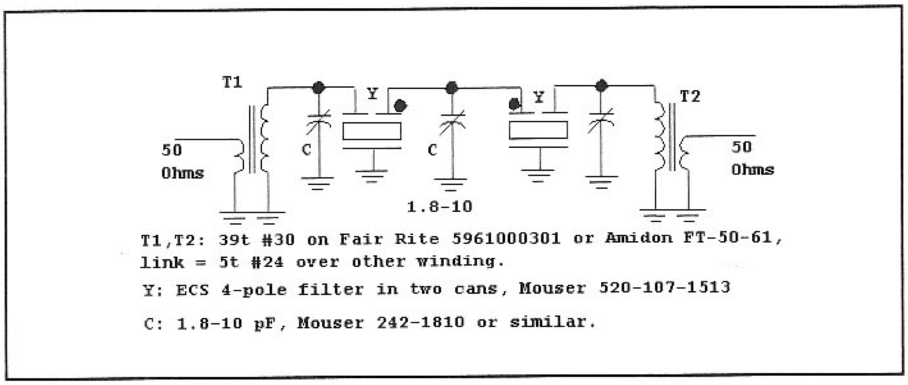


Fig 7.51—4th order monolithic crystal filter.

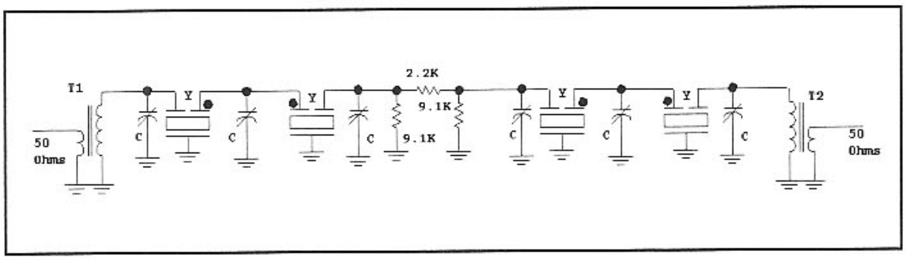


Fig 7.52—8th order crystal filter using two of the filters from Fig 7.51. Each filter block consists of a capacitor-filter element-capacitor-filter element-capacitor combination. These filters were the efforts of Jack Glandon, WB4RNO, and Fred Holler, W2EKB.

rates. V1 is ready to drive the X-axis of an oscilloscope while additional op-amps buffer the ramp and offset it as needed to drive the VCO.

An analyzer begins to emerge, shown in the block diagram of **Fig 7.50**. A commercially available varactor tuned VCO serves the LO function, with buffering to reach a level of +17 dBm. Dual conversion is employed to obtain a resolution bandwidth narrower than afforded by the VHF filter. High level mixers are used for reduced IMD.

This is a practical design that that has been widely duplicated.[15] Details are presented in the articles, which appear on the CD that accompanies this book. The rest of our discussion of spectrum analyzers is confined to general comments and thoughts for refinements of the *QST* design.

Two resolution bandwidths are available in the *QST* spectrum analyzer. One with a bandwidth of 300 kHz uses an LC filter while the other uses a commercial 30 kHz bandwidth crystal filter. Our 1-st and 2nd IFs were 110.0 and 10.0 MHz, but 110.7 and 10.7 allow commercial crystal filter elements at 10.7 MHz to be used. These are ECS types X703ND and were purchased from Mouser or DigiKey.

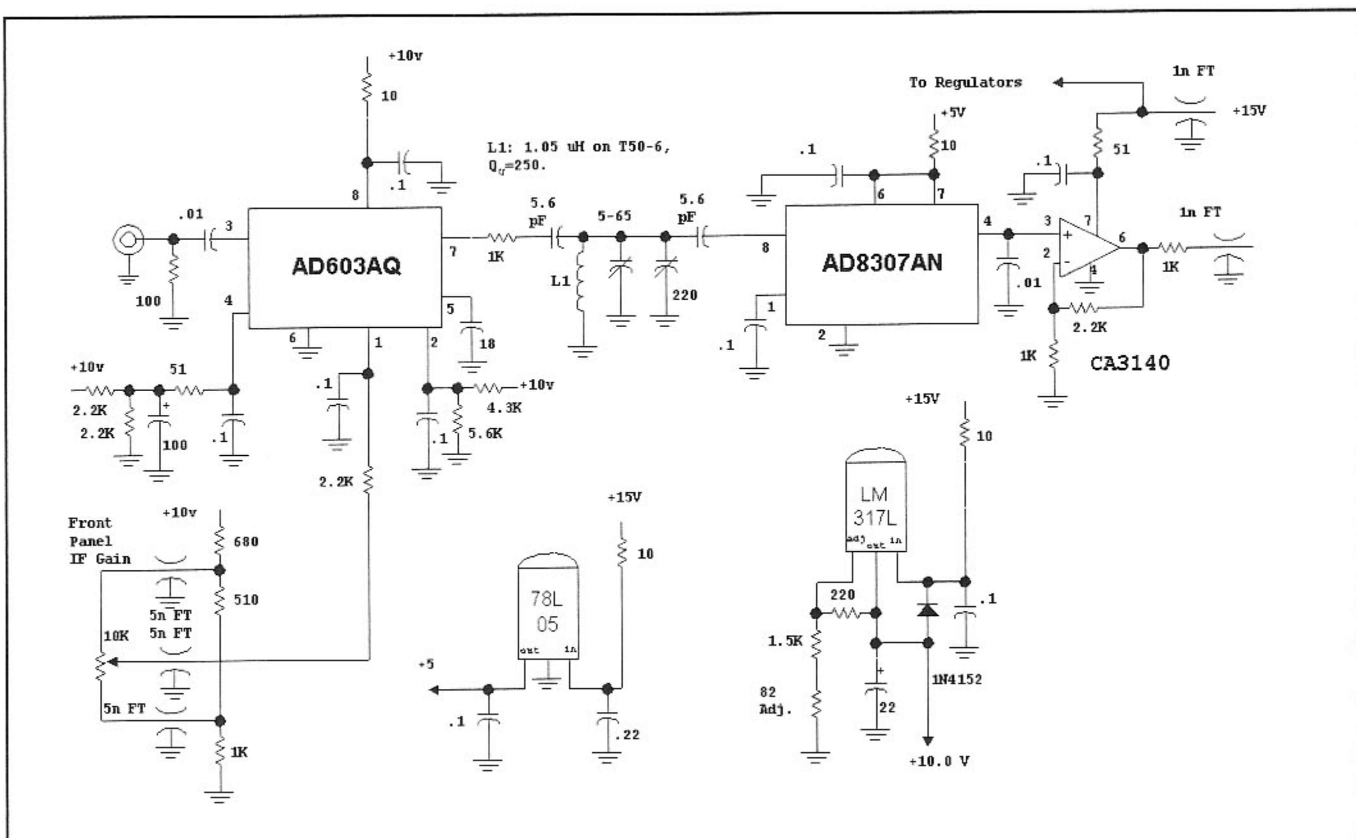


Fig 7.53—This IF and Log Amp section using more accurate integrated circuits and replaces all circuitry of Fig 5 of the original article (see the book CD.) IF gain is variable from 10 to 50 dB. Resistors around the LM317L can be adjusted to set the 10 V level.

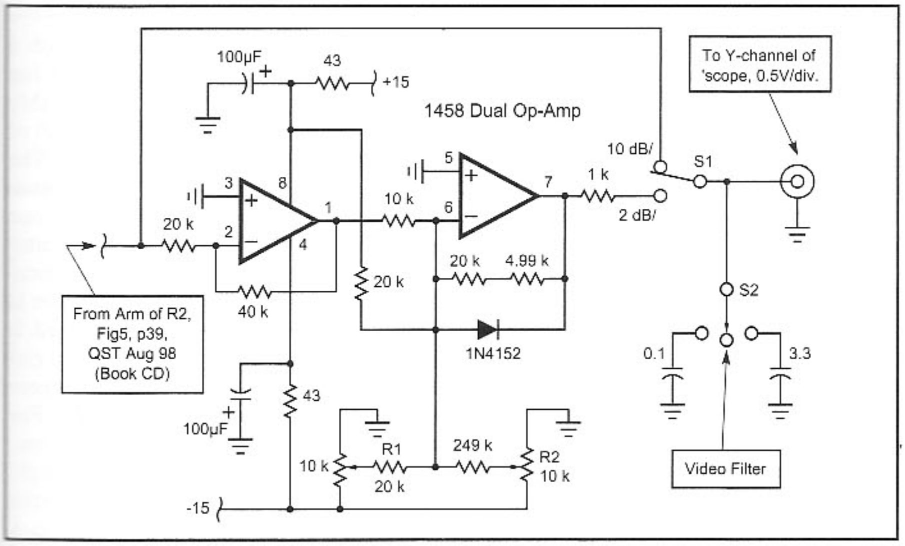


Fig 7.54—Circuit adding 2 dB per division to the spectrum analyzer. The video filter circuit is also included.

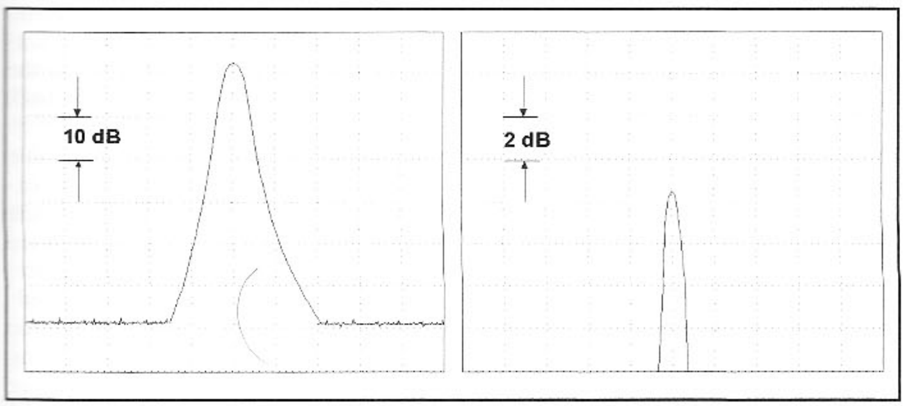


Fig 7.55—A 10 dB/div. signal at the left is adjusted to fill much of the screen. Switching to 2 dB/div. produces the display at the right. Adjusting the offset controls R1 and R2 allows moving the response anywhere on the CRT screen.

Fig 7.51 shows the schematic for a 4 pole filter using two packages. (One "product" from the catalogs includes two filter packages.) The termination for this filter is 3 kΩ at each end, realized with ferrite transformers. Owing to filter loss considerations, a Type 61 core is preferred over the higher permeability cores.

Although the performance was impressive, the stopband attenuation for the 4-pole filter was not adequate. Two stages of the circuitry of Fig 7.51 are cascaded to form an 8th order filter, shown in **Fig 7.52**. This filter has a stopband attenuation in excess of 90 dB, allowing a wide range of measurements. The filters are aligned for a compromise of rounded peak shape, low insertion loss, and stopband attenuation. Alignment can be done with the working analyzer and any convenient input signal.

IF filters for spectrum analyzer use are more critical than those used in a receiver. The analyzer operation essentially paints a picture of the filter shape over the complete dynamic range of the analyzer, so the filter should have a clean, spur-free response over this range.

The *QST* analyzer used the received signal strength indicator (RSSI) function from an early Motorola IC for the log amplifier. The parts were inexpensive and available at the time of publication. The AD8307 from Analog Devices is now commonly available and offers significantly better performance. The AD8307 has a wider dynamic range, improved accuracy, better temperature stability, and is the recommended part. However, it is not a pin-for-pin replacement, and it uses a different input power window, so the designer/builder will have to do some circuit development. The original system used discrete parts for the IF amplifier. An updated version that includes an AD603 as the IF amplifier, is shown in **Fig 7.53**. This circuit drives R2, the "log amp cal" pot, which then is routed to the added 2 dB per division board (described below) and then to the oscilloscope Y axis.

The analyzer contains a video filter, which consists of nothing more than a capacitor that is switched to parallel the video line to the oscilloscope Y-axis. This component, with the driving output resistance, serves to smooth the noise that otherwise creates a fuzzy line. The original video filter used a SPST toggle switch and a 0.1 µF capacitor. This has been replaced with a SPDT/Center-off toggle switch. Two capacitors of 0.1 and 3.3 µF are available, shown in **Fig 7.54**. The heavily filtered response is especially useful for noise measurements. Either filter may be useful in creating a trace that is more easily read on screen.

The spectrum analyzer user soon notices that the sweep rate must be changed with changes in filtering. This is usually a consequence of sweeping. The signal coming out of a filter can respond only as fast as the bandwidth of the filter allows. If, for example, our analyzer had a bandwidth of 1 MHz, we would expect to see output changes at the log amp commensurate with 1 µS. Any sweep rate available in the *QST* analyzer would be slow enough to keep up with such a bandwidth. But switching a 30-kHz filter into the system will cause the response shape to distort, never reaching the peak response seen with a slow sweep. Narrow video filtering does the same thing. Modern analyzers will automatically adjust sweep rates to match the selected resolution and video bandwidths.

Our spectrum analyzer is configured to produce 10 dB of change for every major division on the CRT screen, assuming an 8 division vertical range. This is in line with many traditional instruments. There are many situations when greater amplitude resolution is needed. One might be, for example, a measurement of resonator Q where one needs to accurately see a 3 dB change. This measurement is facilitated with the circuit of Fig 7.54. A front panel switch is added that allows the user to toggle between 10 and 2 dB per division.

The first op-amp of Fig 7.54 is set for an inverting voltage gain of 2 while the second has an inverting gain of 2.5 for a net of 5. The circuit can be offset by a large amount, which can be dialed in with R1 and R2. Any signal that appears on the screen in the 10 dB/div mode can be offset to appear anywhere on the screen with the 2 dB per division mode, illustrated in **Fig 7.55**.

A crystal oscillator presented earlier (Fig 7.29) is useful as a calibrator for the analyzer. It could be built in with a front panel BNC connector, or as a battery powered stand alone unit. The calibrator amplitude is adjusted with circuit component changes to deliver a level of –20 dBm while using a calibrated source as a "standard."

The calibrator or a signal generator can be used to calibrate the instrument. A signal of –20 dBm is applied to the analyzer input, which is usually run with at least 10 dB of input attenuation. The IF gain is set to generate a reference level response. The attenuator is then switched in 10 dB steps to move the response down the screen. If the signal does not line up on the major screen markers the log amp gain is changed and the process is repeated until reasonable log accuracy is realized. Analyzers using the AD8307 log amp are so accurate that the oscilloscope's vertical position control functions much like the IF gain control. There is no significance to the "screen bottom" setting in the 'scope in this application.

The AD8307 log amp accuracy is as good as or better than that of log amps in many spectrum analyzers found on the surplus market, allowing the builder/designer to realize outstanding performance with modest cost. Consumer communications ICs with built-in RSSI functions do not fare as well. But moderately accurate measurements are still possible by careful application of the step attenuator.

Consider a spurious response evaluation of a transmitter as a typical example of a measurement that asks for a dB ratio between two power levels. The transmitter is applied to the analyzer, taking care to keep all signals on screen. An extra attenuator or power tap may be needed to safeguard the analyzer from the high outputs available from a transmitter. The display level of the spur is carefully noted, perhaps by using the 2 dB/div mode for improved accuracy. The analyzer is tuned to the carrier signal and attenuation is added until the on-screen response equals that observed for the spur. This procedure is enhanced if 1 dB steps are available in the step attenuator. The spur level in dB with respect to the carrier (dBc) is then the amount of attenuation added. This measurement is as accurate as the step attenuator and has little to do with the analyzer characteristics. Harmonic distortion is a special case discussed later.

Shielding

One of the first questions ask when a designer embarks on the construction of a spectrum analyzer is "how much shielding is needed." While difficult to quantitatively answer, a little thought shows that shielding must be very good. The *QST* analyzer we have discussed has a minimum bandwidth of 30 kHz and a noise figure around 20 dB, so the minimum discernable signal is around –109 dBm. Yet we routinely use this instrument with 100 W transmitters. That power is +50 dBm, 159 dB above the analyzer MDS. This is the attenuation that must be provided in the overall measurement setup to be able to do good measurements. Part of this results from shielding and part comes from testing the transmitter with a non-radiating termination.

The popular boxes offered by Hammond, available in many catalogs, afford excellent shielding. These cast aluminum boxes have tight fitting bolt on lids and are easily drilled. A box is used for each major block in the RF chain, so one box contains the first mixer, post mixer amplifier, the VCO, and its buffer amplifier. The input low pass resides in a separate box with the 110 MHz first IF filter in another. The only "open" board in the analyzer contains the time base. Signals move into and out of the box on coaxial cable while dc bias and gain control lines are attached to feedthrough capacitors. The VCO tune line is on coax. Wires extending through rubber grommets in box walls are *not* suitable and should never be considered for RF application.

Use what is available for coaxial connectors. SMA or SMB are excellent, but expensive and not generally required for HF and VHF. BNC cables have become more affordable with the popularity of computer networks. A crimping tool is needed to take advantage of these parts. Inexpensive phono plugs and sockets (RCA) are suitable if carefully applied.

Application Hints

The spectrum analyzer is not merely an evaluation tool to test the rigs that are finished, although many folks treat it as such. Rather, the SA is used to measure things throughout the experimental experience. First and foremost, it is a sensitive meter used to examine signal levels, even when they are too weak to be seen with an oscilloscope. The sensitivity is the result of narrow bandwidth. Utility is maintained as a result of sweeping, eliminating the need to retune for various signal components.

The spectrum analyzer is almost always a tool for substitution measurements. As such, it is usually necessary to break a 50-Ω signal path and attach the spectrum analyzer. This is done in a breadboard by bolting a BNC connector to a ground lug and then soldering that lug to the ground foil near the circuit under test. The connector can be moved later, so it can be placed close enough to maintain short leads.

In other cases it is handy to attach a BNC chassis connector with ground lug to a short length of small coaxial cable (RG-174 or similar) with the other end of the cable soldered into the circuitry. The probing end should have a maximum ground length of perhaps one half to one inch with a similar length for the center conductor for HF and low VHF applications. The end of the center insulation is removed and soldered to a circuit board. It is vital to solder the cable ground to a circuit board ground close to the place where the measured signal currents flow. For example, if the output of a feedback amplifier was to be examined, you might "lift" a blocking capacitor from the output signal line. That capacitor can then be tack soldered to the cable center conductor. The ideal place for the cable ground is the board ground foil directly under the capacitor position. Removal of solder masking may be required in some cases. Alternatively, the ground connection for the bypass capacitor related to the feedback amplifier output could be used.

It is rarely valid to merely attach a cable ground at the edge of a board at, for example, a mounting hole. This procedure works well enough for high impedance probes from an oscilloscope while performing *in-situ* measurements. The feedback amplifier, in that case, still has the output currents flowing to a following stage. That termination was broken for our substitution measurement. Examine the complete loop starting and ending with the place where the center conductor and coax cable braid split. That loop should generally be small. If you are trying to evaluate the presence of spurious signals, you should not allow the loop to contain extra stages that might be carrying some of the contaminating signal.

Some applications are presented in the paper on the CD that accompanies this book.[16] The applications related to power meters, again on the CD, are also generally useful with spectrum analyzers.[17] Spectrum analyzer measurement of intermodulation distortion was discussed earlier in this chapter in the section on signal sources.

A common problem encountered when breadboarding a new circuit is a spurious oscillation. More often than not, this will occur at very high frequencies, often approaching the F_T of the offending transistor. A spectrum analyzer tuning only to 70 MHz will never see this directly, but the result is often still apparent on screen. This appears as a low level signal that moves in frequency as a hand or tool is placed close to the circuit. This is the re-

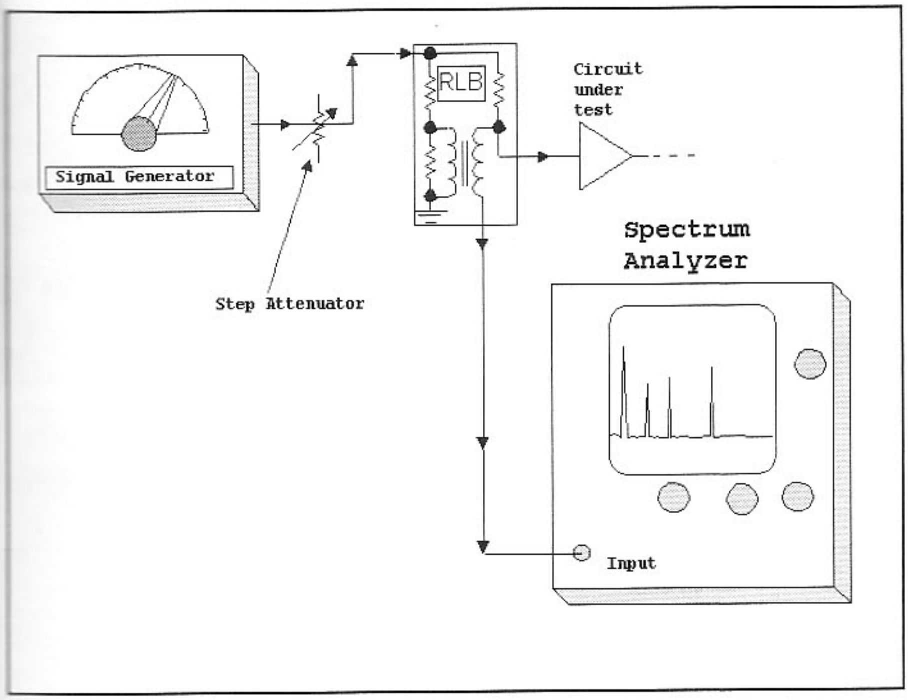


Fig 7.56—Return loss (VSWR) is easily measured during bench testing with a simple bridge.

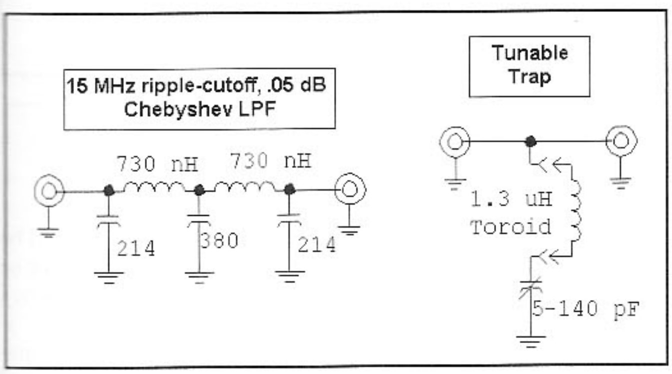


Fig 7.57—Low pass filter and tunable trap are used to evaluate harmonic distortion in the front end of an analyzer. These circuits were used to evaluate analyzer performance for measurement of 14-MHz harmonics from a transmitter.

sult of mixing between the spurious oscillation and harmonics of signals that excite the circuit.

It is often useful to investigate the quality of impedance match, even with small signal amplifiers. A return loss bridge (discussed earlier in this chapter) is driven by a signal source and applied to a circuit under test. The generator power is turned down to a level that will not overdrive the amplifier under test. The return loss, which is directly related to VSWR, is then measured as shown in **Fig 7.56**.

Calibration During Measurements

A calibrator circuit was described earlier, a convenient means for checking analyzer amplitude and frequency calibration. But there is more to calibration for RF measurements.

Generally, the best procedure is to place no trust in the equipment that has not been earned. This applies especially to the homebrew spectrum analysis equipment described in this book, but is also important for the best laboratory instrumentation available.

Assume that we plan to measure the gain of an amplifier, and that we wish to get the most accurate number possible. The amplifier is set up with the appropriate power supply, a signal generator, and the spectrum analyzer or power meter. The set up is turned on and generally checked. The calibrations that have already been done for the analyzer are enough to get things started.

Once the system is working as expected, we now do a test set-up calibration. The amplifier is disconnected from the two coaxial cables and replaced with a through connector. This is a barrel or bulkhead connector in BNC cables or the equivalent in other cable types. It is important to use the same cables for the calibration as are used with the amplifier. The response is noted with the through connector. The amplifier is then inserted in its original position and the new response is noted. 2 dB per division is used for both measurements. The gain is then the difference between the two levels.

Newer commercial equipment is usually fairly accurate in the 1 or 2 dB per division ranges, so log errors are not major. However, when a homebrew analyzer based upon an IC RSSI function is used, the measurement should be done with a step attenuator rather than with numbers from the screen. This is a wise procedure with older commercial analyzers or with any measurements performed near the bottom of the log amplifier ranges, or with any measurements where noise levels are being compared.

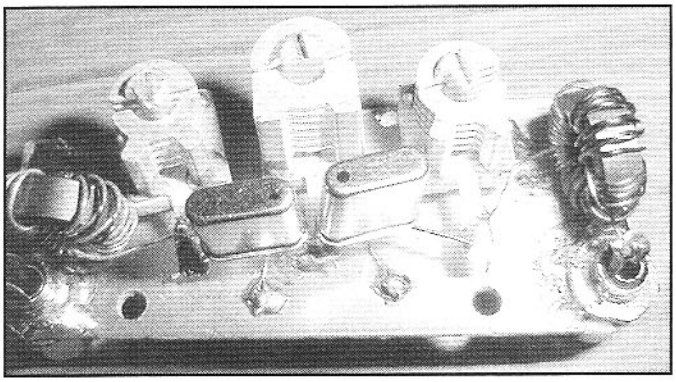
Commercial spectrum analyzers feature highly refined frequency readouts. A cursor function can be activated that marks a trace on screen. The exact frequency is then displayed. Some instruments can be extremely accurate in this mode. The procedure is much more casual with the *QST* and other simple homebrew instruments. When we see a signal on screen with an unknown frequency, we carefully note the horizontal position, disconnect the input cable and attach a signal source adjusted for the same response, and read the frequency from a counter attached to the source.

The analyzer can be modified to incorporate a frequency counter. The frequency sweep would be stopped by opening the line from the center arm of the sweep rate pot.[18] There would still be horizontal motion on screen, but the amplitude would be fixed at that corresponding to screen center. This is called a "zero span" mode. The VCO could then be counted. Subtracting the first IF from this value gives a "center frequency."

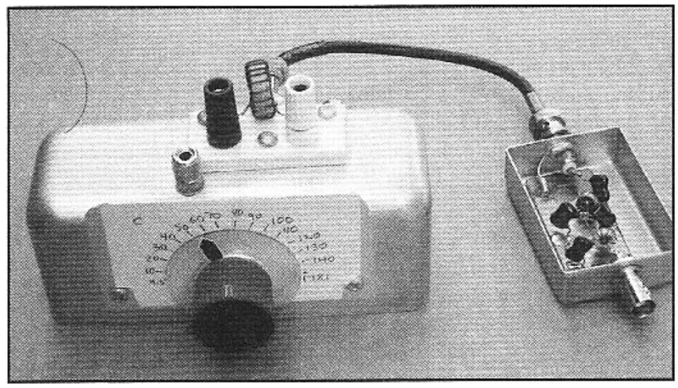
Harmonic distortion measurements

Although common, this seemingly simple chore can be complicated by harmonics created within the spectrum analyzer. Measurements are meaningful only when we have confirmed the analyzer performance.

The evaluation can be done with several experiments. The first applies a signal to the analyzer from a generator and looks at the harmonic levels. The attenuation in the analyzer front end is changed. If both the



A close up photo of a 4th order filter built by WB4RNO. Any small trimmer capacitor with a suitably low minimum capacitance can be used.



Clean-up gear to reduce the harmonic content of a signal source. This is used when evaluating a transmitter or other source for harmonic distortion.

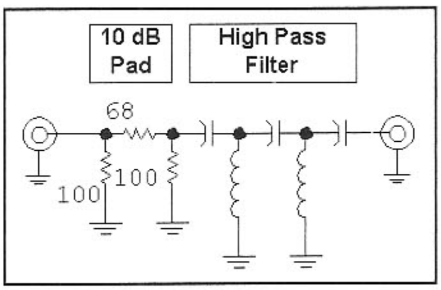


Fig 7.58—High-pass filter used for harmonic measurement. See text.

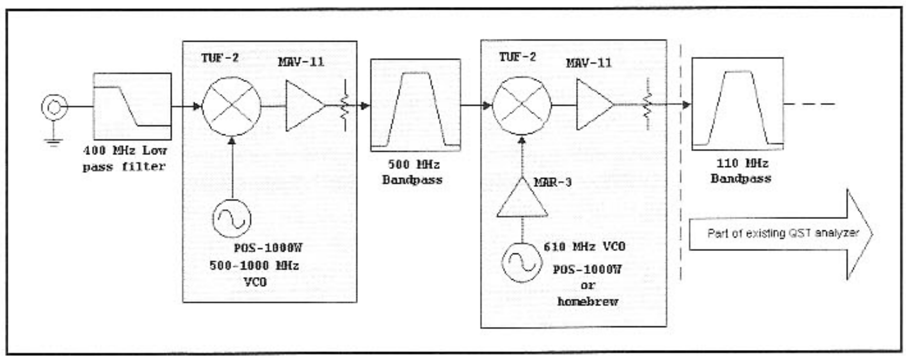


Fig 7.59—Front end for a triple conversion spectrum analyzer tuning to the low UHF spectrum. This analyzer has yet to be built, but is planned.

fundamental and the indicated harmonic change in unison, the distortion is probably real and not an analyzer spur.

A second experiment places a low pass filter in the line from the generator to the analyzer. This will improve the generator performance, allowing the first experiment to be repeated with greater sensitivity. Again, identical tracking of fundamental and distortion tend to vindicate the analyzer, now at a level commensurate with the new harmonic attenuation level.

Traps can be used for further analysis. A tunable trap is shown in **Fig 7.57**. The trap is placed in the line between generator and analyzer and is tuned to attenuate the fundamental signal. If the trap is sharp, it can dramatically attenuate the fundamental with little impact on the harmonics. A 20 dB or greater attenuation of the fundamental without altering the harmonic guarantees the fidelity of the analyzer.

An analyzer can still be useful for analysis even when it is generating harmonics of its own. All that is required is to reduce the fundamental signal reaching the analyzer without altering the harmonic energy. This can be done with a high pass filter, shown in **Fig 7.58**. The high pass is preceded by a 10-dB pad, establishing a proper impedance environment for the generator (or transmitter) being evaluated for distortion. A measurement is performed without the trap to establish the fundamental power. The trap and pad are then inserted and the analyzer sensitivity is increased by the pad loss. The harmonic power is read to calculate a dBc value. If necessary, the trap can be cascaded with the high pass for further attenuation of the fundamental.

Expanding Performance

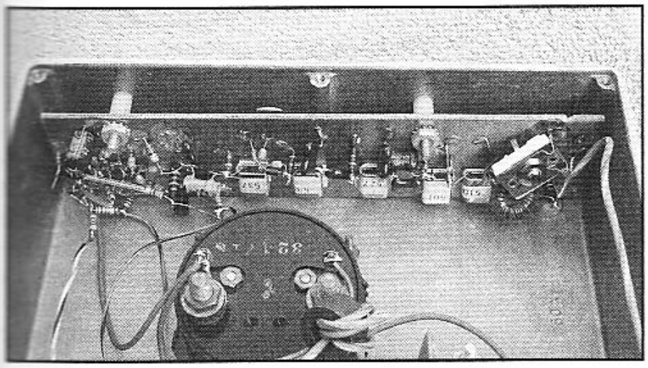
The *QST* spectrum analyzer tuned over a restricted range of 0 to 70 MHz with only two available resolution bandwidth positions. The VHF experimenter will want higher frequency performance.

Expanding the tuning range to higher frequency is easily realized, beginning with a review of the latest catalogs from Mini-Circuits and other vendors. A 100-200 MHz VCO was the basis for the *QST* design (Fig 7.50), but this could be replaced with other parts. One variation would use the POS-535 tuning from 300 to 525 MHz as the first LO. The first IF would become 300 MHz. A good choice for a second IF would then be 21.4 MHz where commercial monolithic crystal filters are available. A VHF 2nd LO will be needed, which could be free running or be multiplied up from a lower frequency crystal oscillator.

A triple conversion version of the analyzer is shown in the block diagram of **Fig 7.59**. This version tunes to 400 MHz with a first IF at 500 MHz. The second IF is then 110 MHz using the circuitry from the original design. This upgrade could be built as a supplement to the *QST* analyzer without disturbing the functionality of the original. This UHF extension uses only +7 dBm mixers, so the new design will not be as strong as the first with regard to distortion measurements. The 2nd LO could be homebrew or might use a second Mini-Circuits part.

The present analyzer can be supplemented with a block converter in much the same way that we add converters ahead of receivers for the higher HF or the VHF bands. A very simple block converter that we built uses a POS-200 (100-200 MHz) VCO driving a TUF-1 mixer. A 4 dB pad in the signal path sets the overall conversion gain at –10 dB. The 144 MHz amateur band is converted to 30 MHz when the LO is at either 114 or 174 MHz. Recall that the 3rd harmonic of a LO is generated within a diode ring mixer, often creating spurs, but also allowing third harmonic mixing. So setting the VCO to 157.3 MHz creates an effective LO of 472 MHz, which will convert 432 MHz to appear as 40 MHz. Mixer conversion gain is less with harmonic mixing and depends on the harmonic being used. The block converter output is filled with numerous spurious responses, but is nonetheless a useful and simple tool.

Figure 7.60 shows a narrow tuning range approach to spectrum analysis. This circuit was configured as a measurement receiver. It uses an outboard local oscillator to drive a diode ring mixer followed by a traditional post-mixer amplifier. The post-amp output is then applied to a narrow bandwidth 5 MHz crystal filter that then drives a log amp. There are two outputs. One is a built in meter while the other is a jack to drive a DVM. This instrument was originally configured to measure carrier and sideband suppression in single sideband transmitters, but has also found use in the pursuit of spurs from frequency synthesizers using direct digital synthesis. The instrument could also be configured for baseband measurements close to dc. It



Crystal filter, log amp, and output driver for "Measurement Receiver."

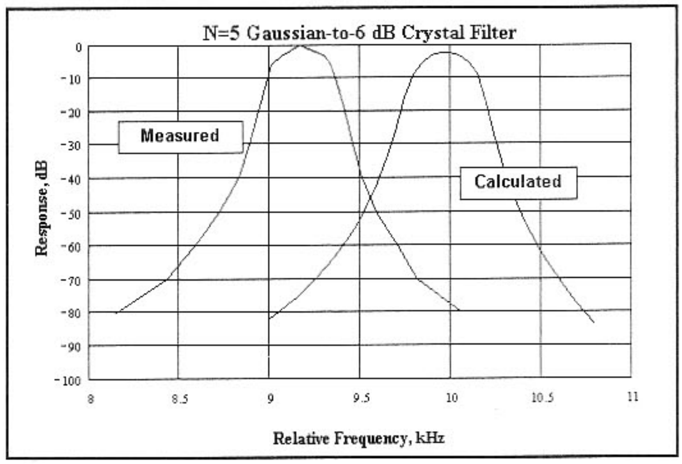


Fig 7.61—Crystal filter response for the circuit used in the measurement receiver. See text.

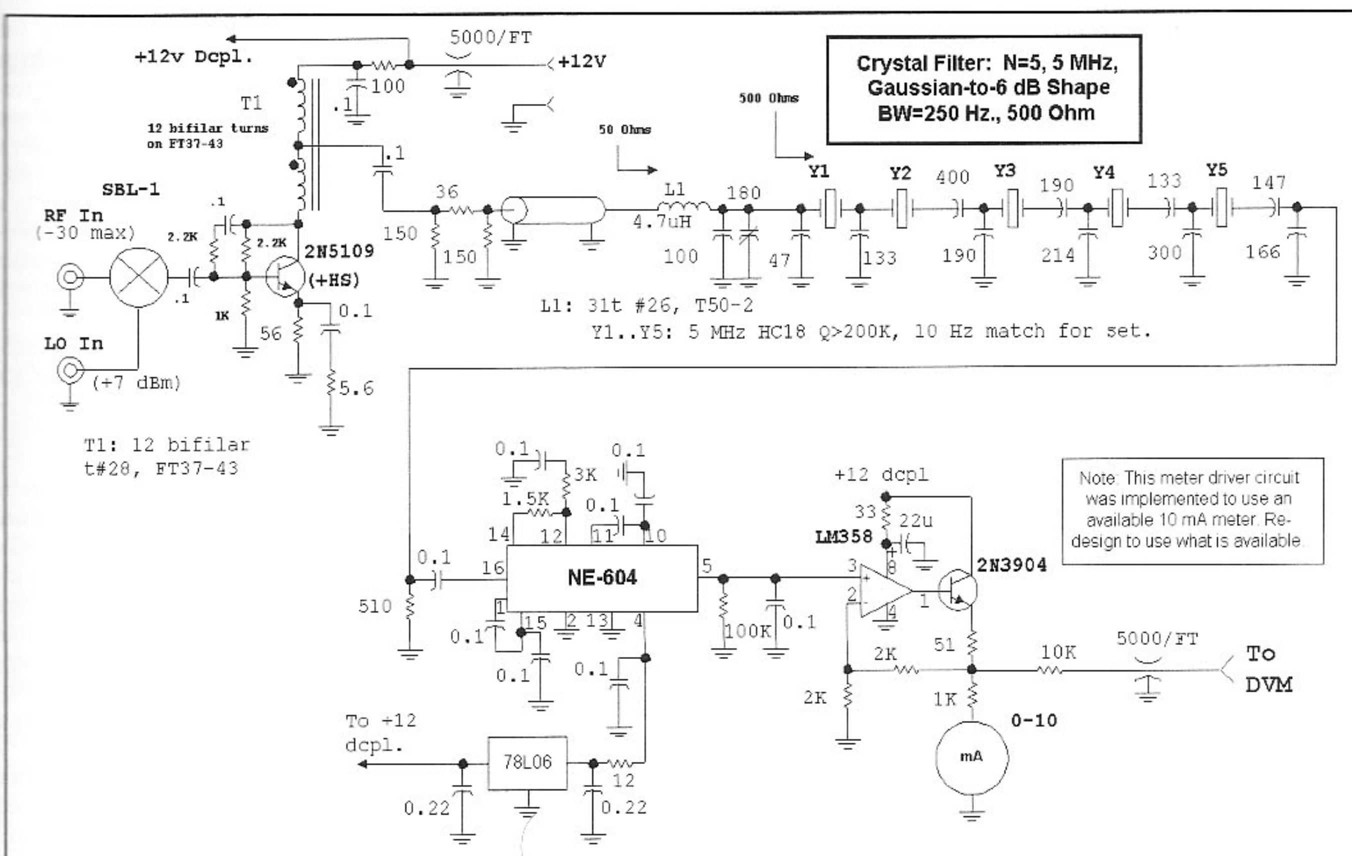


Fig 7.60—Measurement receiver for measurement of SSB transmitters. This unit used an available 10-mA meter movement with a high resolution scale, but can be adapted to available meters. This instrument can be adapted as a narrow tuning range spectrum analyzer, a refinement that we have yet to complete.

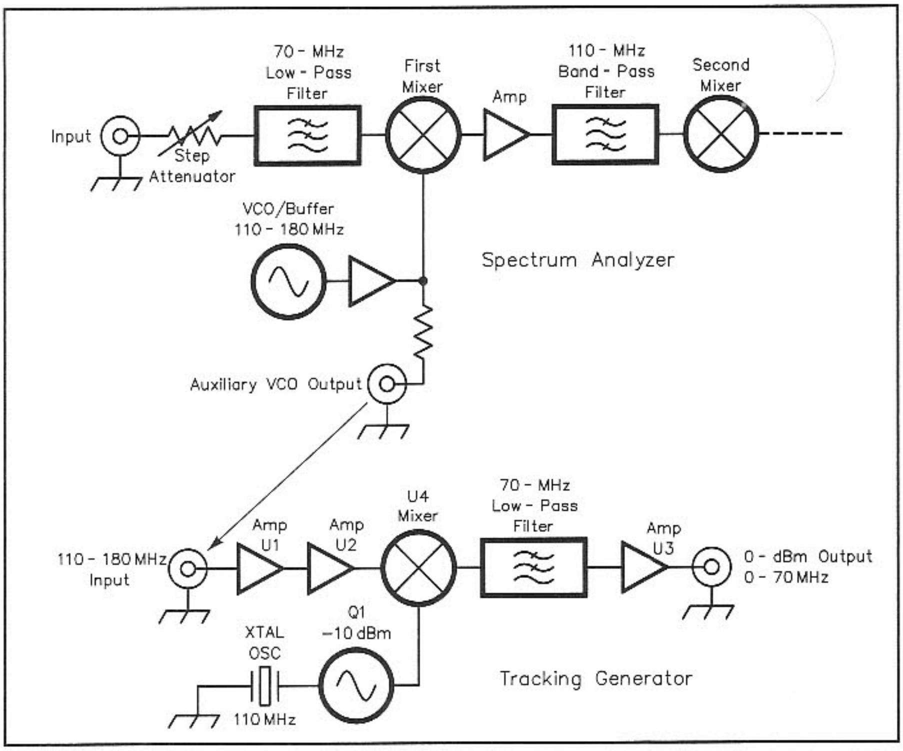
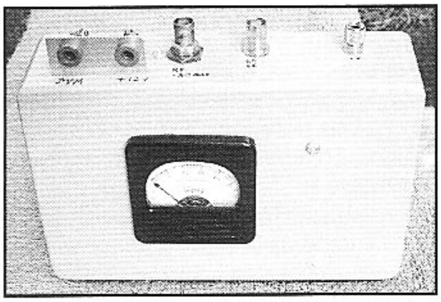
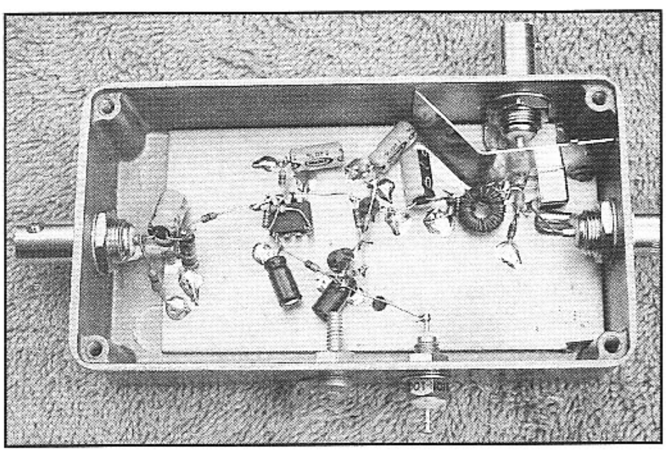


Fig 7.62—Functionality of a tracking generator and the mating spectrum analyzer front end. The complete design is included on the book CD.



Outside of measurement receiver.



Converter for baseband spectrum analyzer on a PC. Used for evaluation of IMD in an HF transmitter.

would then be useful for noise measurements in connection with oscillator phase noise evaluation.

The narrow crystal filter used in the measurement receiver is designed for a Gaussian-to-6 dB shape. Measured and calculated responses are shown in **Fig 7.61**. This filter shape is ideal for measurement applications, a consequence of the rounded, unambiguous peak with reasonable skirt response. The prospective builder is encouraged to design his or her own filter, for the component values will depend on crystal characteristics. The crystal used in this filter had a motional inductance of 98 mH and average unloaded Q over 200,000. The crystals were matched within 10 Hz. This response shape is generally very tolerant of component variations. Note that the traditional symmetry in component values is not present in this filter, even though the terminations are equal at 500 Ω at each end. Avoid narrow Chebyshev filters in analyzer applications.

This measurement receiver could be reconfigured as a spectrum analyzer with relative ease. A simple way to do this would be to modify the existing *QST* analyzer. Power supply and a sweep voltage from the analyzer could be brought to suitable connectors to drive the narrow bandwidth unit. The video output could be routed directly to the Y axis. The same sweep circuit and related panel controls would then control both spectrum analyzers.

A stand-alone swept VCO would be needed for the narrow bandwidth adapter. This, however, is not a difficult design task. It is wide bandwidth VCOs that offer greater challenge.

Tracking generators and filter measurements

Swept instruments are ideal for the alignment of filters of all types. Having a swept signal means that the entire frequency response can be displayed at one time. A tracking generator (TG) converts a spectrum analyzer to perform this task.

If we think of a spectrum analyzer as a special purpose receiver, a tracking generator is nothing more than a transmitter that transceives with the receiver. A block diagram is shown in **Fig 7.62**.

A sample of the swept first oscillator from the spectrum analyzer is required for the tracking generator. This signal is amplified and becomes the LO for a high level mixer, U4. The RF input for that mixer is a crystal controlled signal *exactly* at the spectrum analyzer first intermediate frequency. This frequency is easily measured by injecting a signal from a generator into the first IF with the spectrum analyzer set for the narrowest possible resolution bandwidth. This measurement needs to be done after the analyzer is finished and working, but prior to ordering a crystal for the TG.

This TG has an output of 0 dBm. This signal is a swept one that is always tuned to the same frequency that the analyzer sees. The great utility of a tracking generator over a simpler stand-alone swept oscillator is that the SA-TG combination allows observation in the narrow bandwidth of the analyzer. This results in a dramatic increase in measurement dynamic range. The evaluation of filter stopband attenuation details at levels well below the −100 dBc levels are possible with a SA-TG combination. Full details of the TG are included on the CD that accompanies this book.

The extreme dynamic range comes with a price: The shielding of both the tracking generator and spectrum analyzer must be *very* good. As mentioned earlier, the SA-TG combination behaves like a transceiver. However, unlike the usual transceiver we might build for communications, the receiver and transmitter must both function *at the same time!* Signals that might leak from the TG to the SA will interfere with the intended one when testing filters. The observed result will often be a distorted filter shape with the edges of the filter skirts dipping into the analyzer noise floor. Another tell-tale indicator of these problems is a filter shape that changes with the position of some of the interconnecting coaxial cables.

As useful as the SA-TG combination can be, it presents a problem for the serious experimenter: Filters are so easily "tweaked" that builders may be tempted to ignore designing the filters in favor of empirical methods. Don't fall into this trap!

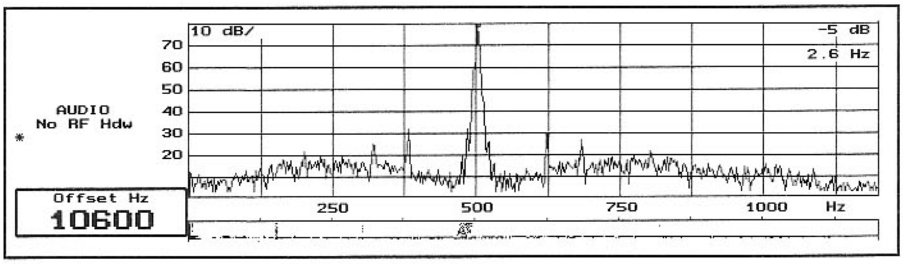


Fig 7.63—High resolution spectrum of a signal generator. The noise is phase noise on the generator. 120-Hz hum modulation is readily observed as well.

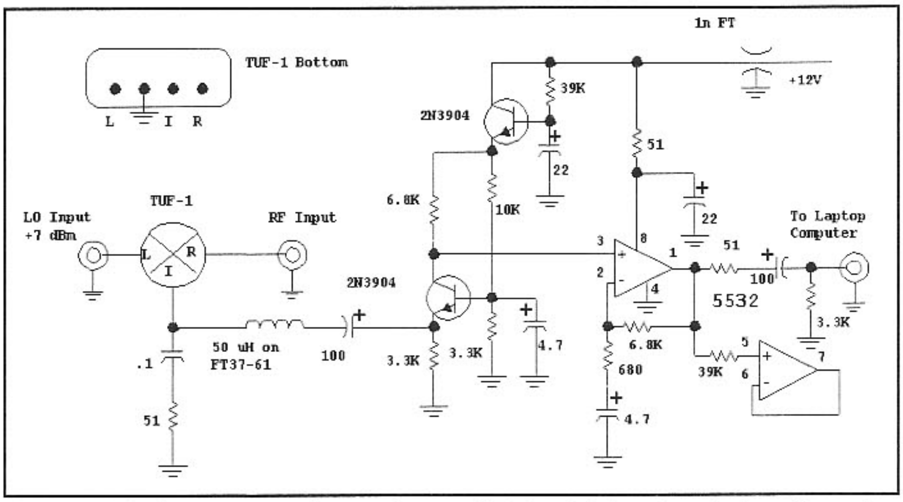


Fig 7.64—Block converter to heterodyne an RF signal to baseband where it can be observed with a spectrum analyzer running on a PC.

DFT Spectrum Analysis

The spectrum analyzers discussed so far have been of the *swept front end* type. The case where a block converter preceded a swept front end analyzer produced a *swept IF analyzer*. There is another popular analyzer that has become very common in recent times, the Fourier Transform Spectrum Analyzer. In this type, an incoming signal is converted to a digital stream of data with an analog to digital converter. The analog data feeding the converter is filtered with a low pass or bandpass filter to restrict the resulting digital data. The time domain representation is then subjected to mathematical calculations resulting in a frequency domain representation of the signal, a spectra. This is then graphically presented. The analysis used is a Discrete Fourier Transform, or DFT. The most popular DFT form is the so called Fast Fourier Transform, or FFT.[19]

The radio amateur is familiar with this method as a software technique. Audio signals are presented to the sound cards of personal computers. The resulting digital data is Fourier transformed in suitable software programs and displayed in one of several forms including the "waterfall" popular with digital communications modes.

DFT spectrum analyzers have two major advantages over swept tools: First, they are capable of very high resolution (narrow bandwidth). Second, the spectrum shown represents the spectrum at one instant in time.

A FFT analyzer is very useful as a measurement tool. **Fig 7.63** shows an example where a signal generator was being investigated for phase noise. The noise shown in the figure is indeed noise, for a cleaner oscillator operating with the same analyzer parameters produced a similar spectrum, but without the noise. The resolution bandwidth for this example is 2.6 Hz! The hardware and software used for this example are discussed in much more detail in Chapters 10 and 11.

Although FFT methods often concern audio or "baseband," the concepts are capable of much more. So long as a signal can be sampled in time and converted to digital data, it can be transformed to the frequency domain. Many modern oscilloscopes are built with relatively low speed displays. But the incoming analog signal is anything but slow. The incoming data is amplified and/or attenuated and presented to a high speed "scan converter," essentially an A to D converter. Once the high speed signal is remembered, it can be read at a lower speed and displayed as a time signal. The data can also be presented to a FFT "engine," or computer to generate a corresponding spectra. While usually lacking the dynamic range of an analog spectrum analyzer, a spectra with a dynamic range of 50 dB or better is common with such oscilloscopes.

A block converter can be used to move part of an RF spectrum down to audio where it can be examined with a FFT type spectrum analyzer with an example shown in **Fig 7.64**. An external step attenuator and (optional) bandpass filter precede the converter. A diode ring mixer then moves the signal down. The rest of the circuitry is very much like that found in direct conversion receivers. This converter can be used ahead of the FFT analyzer implemented with the DSP hardware from Chapters 10 and 11. We have also used it with a personal computer sound card and modest cost software.[20] One must be careful with any of these schemes to avoid overdriving the A-to-D converter; overdrive can turn the entire screen to unrecognized gibberish! Sound card solutions seem less robust than the devoted DSP tools.

A block converter and a baseband FFT analyzer are ideal for evaluation of SSB transmitter IMD. What had always been a difficult laboratory measurement is now available to almost all experimenters. A traditional two-tone audio generator was included earlier in this chapter.

The narrow resolution available from an FFT based analyzer will also allow the experimenter to measure in-band transmitter distortion. A tone spacing of around 100 Hz then becomes appropriate. In-band performance becomes important when a SSB transceiver is used to process narrow bandwidth information such as encountered in PSK31. Again, the availability of measurement tools provides the experimenter with great opportunity.

7.9 Q MEASUREMENT OF LC RESONATORS

Several schemes have been used for Q measurements over the years. They can all work well when carefully executed. Two schemes are presented here for LC tuned circuits. The first method measures the bandwidth of a tuned circuit configured as a symmetrically loaded bandpass filter with very high insertion loss. The schematic is shown in **Fig 7.65**.

The two coupling capacitors should be approximately equal. This prevents heavy loading by the input with weak output coupling which could create high insertion loss with a wider than minimum bandwidth. Equal values guarantee that the input and output each contribute equally to the loading. High insertion loss then ensures that the external loading is light so that bandwidth is determined only by resonator loss.

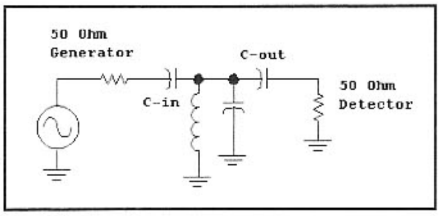


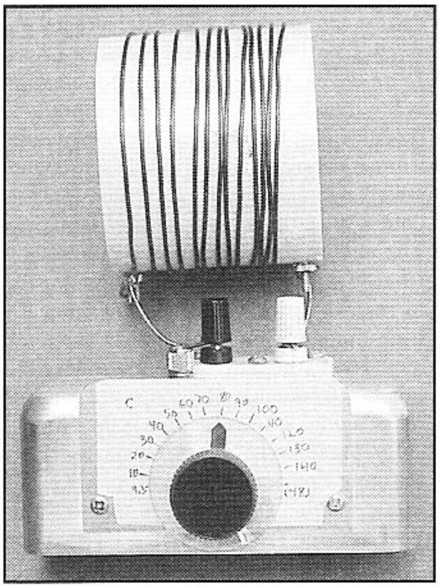
Fig 7.65—Measuring Q by determination of 3-dB bandwidth. The coupling capacitors, Cin and Cout, should be approximately equal and should be small enough that the insertion loss is 30 dB or more.

The measurement is done with a signal generator and sensitive detector such as a spectrum analyzer, a 50-Ω terminated oscilloscope, or one of the power meters described earlier. The generator is tuned for a peak response and the center frequency, f_0, is read with a counter attached to the generator. The output amplitude response is also noted. The signal generator drive is then increased by 3 dB, causing the output to increase by the same amount. The generator is then tuned first above, and then below the peak until the response is identical to the original amplitude. The frequencies of the upper and lower –3 dB points are noted and the difference is calculated as the BW. Then $Q_u = f_0/BW$ where both are measured in the same frequency units. If the insertion loss is 30 dB or more, the measured Q is very close to the unloaded value. See section 3.3. The measurement can be done with lower IL, but corrections will then be required to calculate Q_u from the measurement Q.

Another scheme for Q measurement uses resonator elements in a *trap* circuit, shown in **Fig 7.66**. Again, a tunable generator and a 50-Ω detector are used. However, instead of configuring the resonator as a lossy filter, we now configure it as a trap, a circuit that produces high attenuation at one frequency. The generator is tuned to find the null in the output response. The null depth, which can be very large, becomes a measure of the resonator Q.

Either a parallel connected series-tuned circuit or a series connected parallel-tuned circuit can be used as traps. There is usually little virtue of one type over the other. We generally prefer the series-tuned circuit because a grounded and calibrated variable capacitor can be used in the resonator. A photo shows a test fixture with a 140-pF variable capacitor and binding posts.

The generator is tuned to find the null response and the level is carefully noted. A spectrum analyzer is ideally used as the detector and should be in a 1 or 2 dB per



A test fixture simplifies Q measurement with the parallel connected series tuned trap method. The inductor shown was 13 turns of #14 enamel-covered wire wound on a 3.5-inch-diameter PVC pipe fitting. This coil had a measured Q of 371 at 7 MHz. The test fixture includes a grounded post allowing additional fixed capacitance to be added.

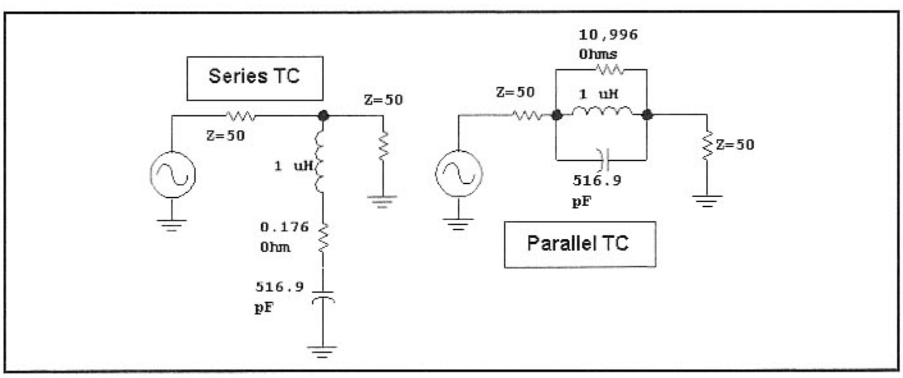


Fig 7.66—Measuring Q by determining the attenuation of a trap. A 7-MHz tuned circuit is used in this example with L=1 µH. The 0.176-Ω resistor in the series-tuned circuit and the almost 11-kΩ resistor in the parallel tuned circuit are models representing a 7-MHz Q of 250. The series-tuned circuit (STC) will have an attenuation of 43.1 dB while the PTC has 40.9 dB.

division sensitivity to provide amplitude resolution. The resonator is then disconnected and the generator is connected to the detector through a step attenuator. The attenuation is adjusted until the analyzer response is exactly the same as produced at the null. The attenuator value is then the null attenuation, A, in dB. Values of 60 dB or more are possible with some high Q tuned circuits.

This same measurement setup can be used to determine inductance if a calibrated capacitor is used. The unloaded Q is related to attenuation by

$$Q_S = \frac{4 \cdot \pi \cdot f \cdot L_u}{Z} \cdot \left(10^{\frac{A}{20}} - 1 \right) \qquad \textbf{Eq 7.4}$$

f, MHz; A, dB; L_u, uH; Z, Ohms

if the series tuned circuit form is used, or

$$Q_P = \frac{Z}{\pi \cdot f \cdot L_u} \cdot \left(10^{\frac{A}{20}} - 1 \right) \qquad \textbf{Eq 7.5}$$

f, MHz; A, dB; L_u, uH; Z, Ohms

...if the parallel tuned circuit is applied. Frequency is measured in MHz, A is in dB, and inductance is in µH for these equations. Z is the characteristic impedance of the measurement environment, usually 50 Ω.

It is useful to plot series resistance against attenuation for the parallel connected series impedance. This is shown in **Fig 7.67**. The experimenter may wish to build a similar curve for the series connected parallel impedance.

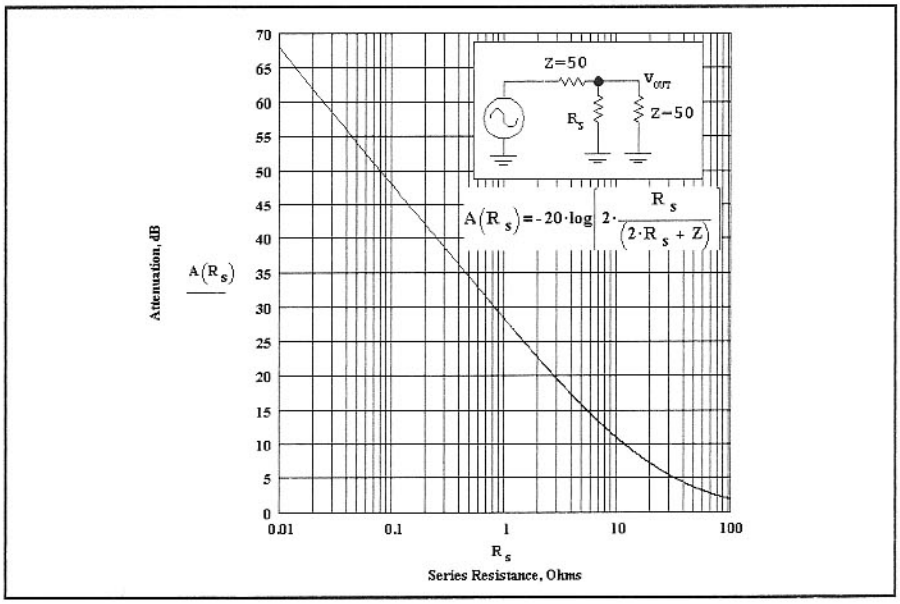


Fig 7.67—Attenuation vs R for the series impedance. See text.

It is important that a solid 50-Ω load and source impedance (Z in the equations) be used in this measurement. If the impedance is in question, use a 10 dB pad at both the generator and detector.

It is also important to prevent harmonics from confusing the results. This is guaranteed if you use a narrow bandwidth detector such as a spectrum analyzer. A wideband detector (a power meter or a 50 Ω terminated oscilloscope) will respond to harmonic energy that is not attenuated by the trap. The spectrum analyzer used for Q measurement could be very simple. Something as simple as a single tuned circuit preceding an oscilloscope would work so long as a pad was used to establish impedance. Alternatively, a very well low pass filtered signal generator could be used with any detector with adequate sensitivity.

The virtue of the trap scheme becomes apparent as soon as the two methods are compared. The traditional 3-dB bandwidth measurement depends on precisely establishing the 3-dB down level. A fraction of one dB error could still impact accuracy. In contrast, the depth of a null is often quite large for high Q resonators, and is easily measured with a step attenuator.

An accurate capacitance measurement tool such as the AADE or W7AAZ meters mentioned earlier is quite useful as a supplement to a Q measurement setup. With such a tool, accurate calibration of capacitors is ensured.

7.10 CRYSTAL MEASUREMENTS

A quartz crystal is modeled as a series RLC paralleled by a capacitance, **Fig 7.68**. Crystals are of special interest, for they are often used in construction of narrow filters. For this purpose, we need to know all of their parameters. Great precision is needed in knowing resonant frequency, for that strongly controls filter tuning. The knowledge of the other parameters is needed at an accuracy similar to that encountered in an LC filter.

There are numerous measurement schemes that will produce the four values. A 50-Ω measurement setup was presented in Chapter 3. Results from it are informative, especially if a batch of "junk box" crystals is encountered. However, more refined measurements are desired for filter design. An extremely useful, yet simple oscillator was also presented in Chapter 3 and is repeated here as **Fig 7.69**. A Colpitts oscillator with an emitter follower drives a frequency counter. A capacitor in series with the crystal, C_s, may be short circuited with a toggle switch. This produces a change in frequency that, when combined with the frequency and capacitor value, yield the motional capacitance, C_m. The motional inductance, L_m, is then calculated from series resonance, which is well approximated by the oscillator frequency when the switch is closed. The design equations are included in the figure. F is the frequency while DF is the frequency shift, both in Hz, when the switch is toggled; C_S and C_P , in Farads, are from the circuit. And as usual, ω=2πF.

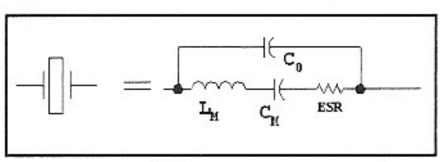


Fig 7.68—Model for a quartz crystal.

If this test oscillator is built with Colpitts capacitors of C_p=470 pF and a series capacitor of C_s=33 pF, the circuit will function (fundamental mode) with crystals from 2 to 25 MHz. Simple equations are valid when C_p is more than 10×C_s. It is

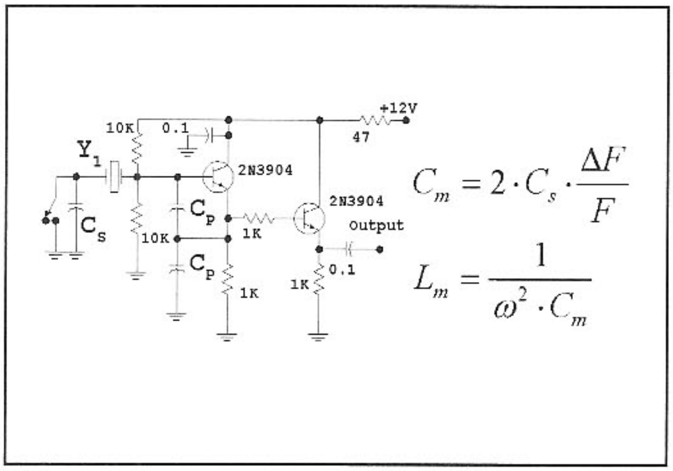


Fig 7.69—Colpitts oscillator for crystal testing, based on an insightful suggestion by G3UUR.

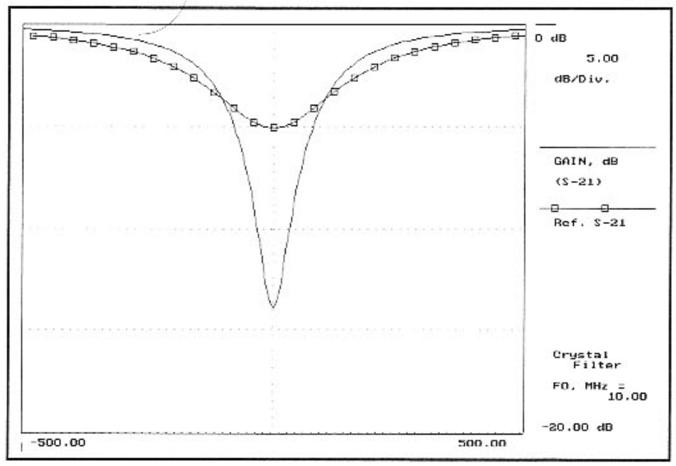


Fig 7.71—Sweeping two crystals while investigating their properties as traps. One has a Q of 40,000 while the one producing the deeper notch has a Q of 200,000. Notch depth is measured to determine Q.

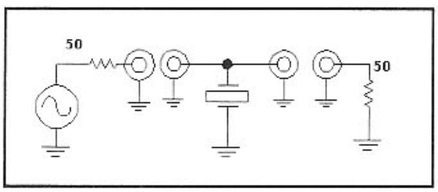


Fig 7.70—Using the trap nature of the crystal for a Q measurement.

also important that the C_s value be determined by measurements that include the switch. The 33 pF capacitor in our test set plus switch capacitance produced a net C_s=41 pF.

The crystal is essentially a series tuned circuit when operating near series resonance, so the series trap scheme described earlier for LC tuned circuits will also provide Q_U, as shown in **Fig 7.70**. Computer generated plots are shown for two different 10 MHz crystals in **Fig 7.71**. The shallow notch represents a low Q crystal with Q_u=40,000. The deeper and narrower notch corresponds to Q_u=200,000. The crystal Q relates to attenuation A in dB, motional L in Henry, frequency in Hz, and terminating resistance Z in Ω with...

$$Q = 4 \cdot \pi \cdot f \cdot L \cdot \frac{\left[10^{\left(\frac{1}{20} \cdot A\right)} - 1 \right]}{Z} \qquad \textbf{Eq 7.6}$$

We performed an experiment with a crystal that had also been measured with earlier methods. The notch method for Q measurement yielded QU=202,000 with ESR=17.5 Ω. This was within a few percent of the earlier measurements. The ESR values for crystals are higher than we usually see with an LC resonator, so the notches are not as deep. This allows measurement with a power meter such as the AD8307 based design described earlier; a spectrum analyzer is not necessary. ESR can be 100 to 1000 Ω for very low frequency crystals, so the series connected parallel tuned circuit method might offer better measurements here.

Parallel capacitance, C_0, is easily measured with other tools such as the AADE or W7AAZ circuits. They are effective because those instruments operate at low frequency, around 1 MHz, well away from typical crystal resonance. With all four crystal parameters available, the designer/builder can proceed with the filter designs presented in Chapter 3.

The equipment described has also been used to evaluate HF ceramic resonators. In one measurement on an ECS type ZTA358MG (from Mouser) we saw L_M=761 µH, C_M=2.74 pF, C_0=31 pF, and Q_U=636. Series resonant frequency was well below the marked 3.58 MHz frequency at 3.38 MHz. The part is normally used in oscillators with a series capacitance.

7.11 NOISE AND NOISE SOURCES

Noise is generally the part of the response generated by our receivers that is undesired. However, we can also use noise as a measurement tool. By injecting noise into a communications system or component and examining the response, we can extract information about the system.

Figure 7.72 shows a simple noise source that is quite strong. This circuit delivers a noise output reaching –50 dBm at 10 MHz on a spectrum analyzer with a 300 kHz resolution bandwidth. This is more than 40 dB above the analyzer noise floor. If we apply this noise source to a filter, the signal appearing on screen is a picture of the filter response. While not nearly as useful as a tracking generator, it is still a simple and useful way to examine a filter. Gain stages can be added to the design to obtain even higher noise output.

The noise source of Fig 7.72 is not very flat with frequency. An improved source could be built with a Zener diode biased for a current of a few mA, with coupling into a high gain amplifier designed to have gain that is flat with frequency.

A noise source suitable for noise-figure measurement is shown in **Fig 7.73**. This circuit was designed by WØIYH and described in a paper included on the CD that accompanies this book.[21] The noise is generated by current flowing in D1 with S1 in the position shown in the figure. When the switch is toggled, current flows to forward bias the diode, preserving the source output impedance in the "off" state.

Paul Wade, W1GHZ, has also done some excellent work with noise generation, which is also included on the book CD.[22] Wade noted that an excellent noise source can be built with the emitter-base junction of a microwave transistor, using the diode as a Zener. Wade reports good results with the noise diodes operating as series elements.

The noise source of Fig 7.73 had an *excess noise ratio* (ENR) of 178 in the HF spectrum. This means that the noise power available from the source is 178 times (22.5 dB) stronger when the diode is biased into avalanche breakdown (Zener action) than when it is forward biased. If we were to attach this source to a perfect amplifier, one with no noise of its own, the resulting output noise would also change by 22.5 dB as the switch is toggled. An imperfect, real world amplifier will generate some noise of its own, so the output noise change will be *less* than 178 times when the diode is toggled. The output noise change is called the Y-factor and this measurement technique is called the Y-factor method. Noise factor is related Y factor by

$$F = \frac{ENR}{Y - 1} \qquad \text{Eq 7.7}$$

...where both ENR and Y are power ratios rather than dB values.

The noise sources are generally not difficult to build. However, calibration can be difficult. We borrowed a noise source to calibrate ours. See the two CD noise papers for more calibration information.

Noise figure for a receiver is measured with the test setup shown in **Fig 7.74**. The noise source is attached to a receiver antenna port with receiver AGC is turned off. The audio output is then applied to a true RMS reading voltmeter. We have used a surplus HP3400A and the Fluke Model 89 DVM. Alternatively, one can build an instrument using an Analog Devices AD636 that converts an arbitrary ac wave form to a dc signal proportional to the true RMS of that waveform. A paper describing this instrument is included on the book CD.[23] True RMS measurements are also done with relative ease with DSP software; see Chapter 11.

Consider an example: We toggle the switch to observe a 15.6 dB increase in audio output. This corresponds to a Y factor of 36.3. From **Eq 7.7**, the noise factor is then 5.04, which is a noise figure of 7 dB.

A practical detail complicates noise measurements when the bandwidth is narrow, such as the 500 Hz found in many CW receivers: The statistical variation with time of the noise from the receiver causes many meters to vary, making it difficult to obtain an accurate reading. The video filter of **Fig 7.75** averages the noise to reduce this problem. The dc output is applied to a high impedance voltmeter or oscilloscope.

The noise figure of amplifiers may be

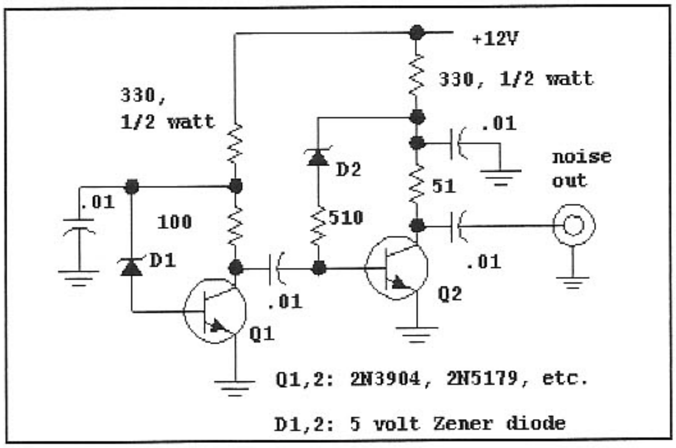


Fig 7.72—Noise in D1 is amplified in a two-stage amplifier, resulting in a strong noise source suitable for measurements. Virtually any diode or transistor types can be used in this source.

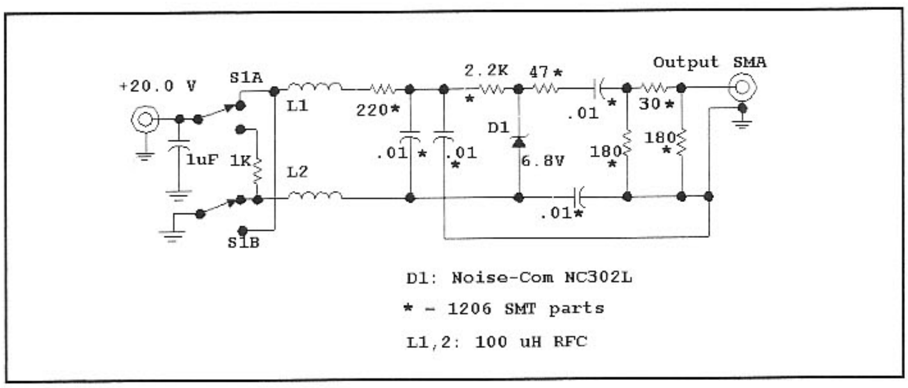


Fig 7.73—Noise source providing a flat frequency response over a wide bandwidth. Our source was built with surface-mounted components where possible. The diode was purchased from Noise Com, East 64 Midland Ave, Paramus, NJ 07652; tel 201-261-8797.

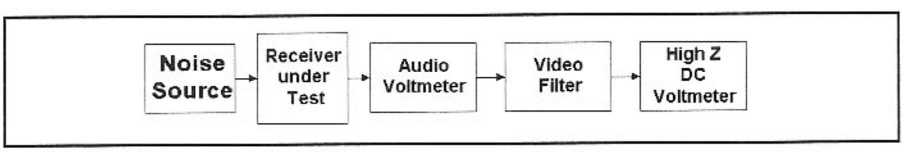


Fig 7.74—Test setup for noise figure measurement. The HP3400A is a true RMS audio voltmeter. This setup includes a video filter driving an oscilloscope, a refinement that may not be required. See text.

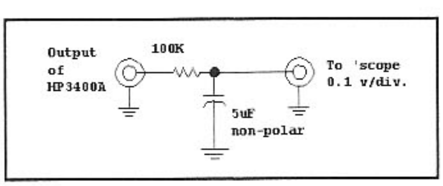


Fig 7.75—A simple video filter reduces meter-reading errors when working with narrow bandwidths.

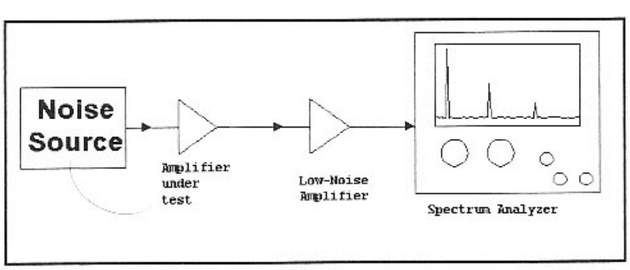


Fig 7.76—Test setup for noise figure measurement of an amplifier or other component.

evaluated with a spectrum analyzer in the test setup of **Fig 7.76**. The key element here is an auxiliary low noise amplifier (LNA) placed between the amplifier under test and the spectrum analyzer. This is needed because the noise figure of the typical analyzer is quite high. The LNA produces a cascade with a low combined noise figure not compromised by 2nd stage noise. See the discussion of noise figure in Chapter 2.

Begin a measurement with the noise source and both amplifiers off. Applying power first to the auxiliary LNA should produce an increase in output noise. Powering the amplifier under test should again increase the on-screen response. Switch the spectrum analyzer to a 1 or 2 dB per division vertical sensitivity and use extensive video filtering to replace trace "fuzz" with a smooth line representing averaged noise. Carefully note the on screen level of the noise. Then switch the noise source to the high noise position. Rather than reading a level from the screen, add attenuation in the analyzer front end until the trace is at the level seen earlier. An attenuator with 1 dB steps (or less) is preferred for this measurement. The amount of added attenuation is then the Y factor in dB. Converting this to a power ratio allows Eq 7.7 to be used.

The auxiliary low noise amplifier we used consists of a MRF544 followed by a Comm-Linear CLC425 operational amplifier.[24] Another suitable amplifier could be built with a cascade of MiniCircuits MAR-3 amplifiers, or similar parts, with a MAR-6 input stage, a configuration that should have a noise figure around 3.5 dB. Low noise figure designs were described in Chapter 6.

7.12 ASSORTED CIRCUITS

Testing AGC in receivers

The circuit shown in **Fig 7.77** is useful when observing the dynamics of a receiver AGC system with an oscilloscope. Named the "ditter," the circuit is an electronic switch with an off-to-on ratio of 80 dB at 14 MHz. The switching elements are inexpensive PIN diodes that are cascaded to obtain the desired off-to-on ratio. The circuit is balanced for the RF signal. However, the dc drive that turns the RF on and off is single ended. This prevents the control signal from creating a click that overwhelms the receiver. The topology was suggested by K7RO.

A slow pulse generator using a 555 timer drives the RF switch. Capacitor C1 controls the timing while the pot sets a duty cycle. A sample of the pulse provides a trigger signal for oscilloscope control. The signal biasing the diodes is filtered with C2 to prevent key clicks from an otherwise too fast rise and fall time. The circuit as shown has about a 1-mS rise, but a longer fall. Although the circuit was useful in studying some of the receivers in this book, a better timing circuit would be useful. One could use an external pulse generator or build a more refined one, probably using more than one timer.

The drive level should be confined to 0 dBm or less, which is adequate to overwhelm almost any receiver. Larger signals are partially rectified with the chosen PIN diodes.

A Experimenter's Receiving Converter

There are many situations where one wishes to receive signals at VHF to facilitate an experiment. A junk box crystal and diode ring mixer form the basis for the circuit of **Fig 7.78**. The crystal controlled oscillator at 25 MHz drives the diode ring at the standard +7 dBm level. Clearly, whatever crystal is available would be suitable.

In one application, we wished to check a 7-MHz transmitter for chirp, or slight change in frequency with keying. The best way to detect this is to listen to a harmonic. The receiver was attached to one of the mixer ports (either one is okay) and a 10X oscilloscope probe was attached to the other through a step attenuator. The transmitter, set for output at 7.04 MHz, was terminated in a load and the probe was attached to the termination. Third harmonic mixing was to be used, so we depend on a 75-MHz LO injection, a naturally strong response with a diode ring. The receiver was tuned to 2.44 MHz. This is the result of the 11th transmitter harmonic beating with the third LO harmonic. A chirp-free response was confirmed. A preselector filter can be used to reduce spurious responses for many applications.

Evaluating Noise in Local Oscillator Systems

The "critical path" for the construction of better communications equipment today is the local oscillator system in use.

Low distortion receiver front ends are becoming easy to build. Crystal filters with

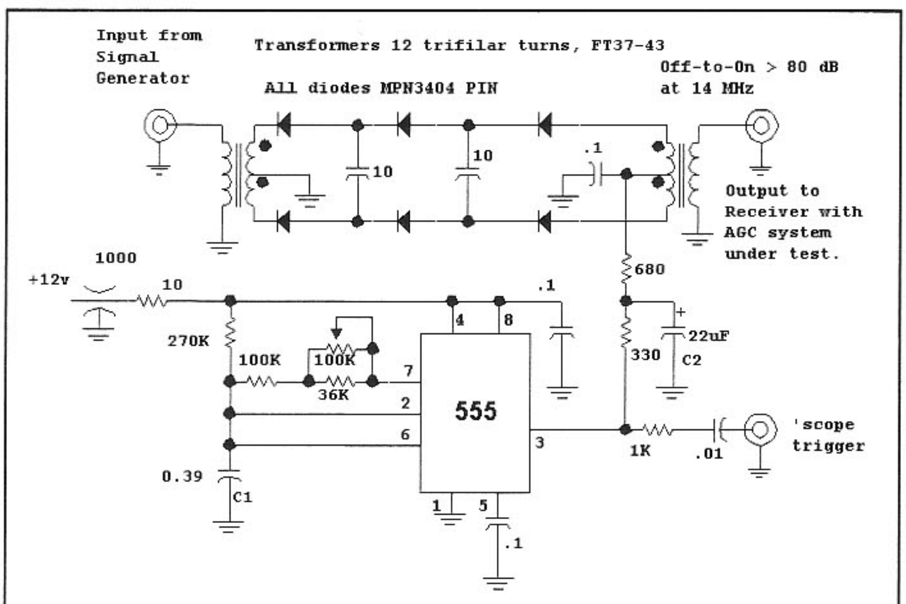


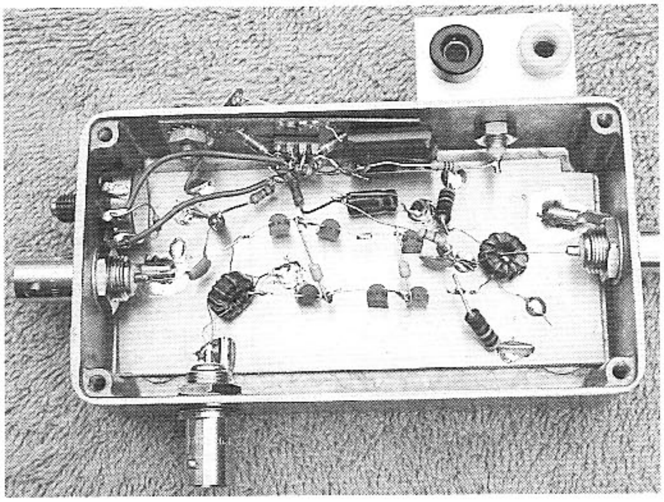
Fig 7.77—The *Ditter*, a circuit for generating keyed receiver input from a signal generator.

additional signal processing can provide outstanding selectivity, both close to a signal and well away from it. The various forms of frequency synthesis available to the builder all offer good frequency stability with the added bonus of electronic tuning. But the LO systems are compromised. Phase locked loop (PLL) systems tend to be plagued with phase noise. Synthesizers using direct digital synthesis (DDS) are often dominated by coherent spurious responses.

Although difficult problems to solve, the measurements are not that difficult. We illustrate the problem here with two measurement examples, the first with a commercial receiver using a synthesizer with both DDS and PLL. A crystal controlled oscillator (Fig 7.29) built with an internal battery, all housed in a well shielded box, was attached to the receiver input through a 10-dB pad and a step attenuator, initially set to 0 dB. The available input signal was confirmed to be –30 dBm at 7.018 MHz. The receiver, in CW mode, was tuned to this frequency with the setting stored in receiver memory. The receiver was then tuned downward while listening for responses with a well defined tone. AGC was on, for there is no provision to turn it off in the compromised receiver. A spur was found within a couple of kHz. The spur frequency was recorded in out notebook. The amplitude response was noted on an audio voltmeter attached to the receiver output. The tuning was then returned to the main signal and attenuation was inserted until the audio output equaled that seen with the spur. This occurred with 58-dB attenuation, so we infer the LO spurious response to be at 58 dB below the carrier, or at –58 dBc. This procedure was repeated as we found a large collection of spurious responses above and below the desired signal with results plotted in **Fig 7.79**.

There are difficulties encountered with this procedure. One must be sure the source is spur free. This was confirmed by repeating the experiment with a receiver using a traditional LC oscillator. You must also be sure that the signal from the source oscillator is not reaching the receiver by routes other than the antenna terminal. This can be confirmed by disconnecting the source from the attenuator to confirm that the signal disappears, or drops well below the level of the measured spurs.

Our second example evaluates phase noise with essentially the same procedure. Again start with a very strong signal, a –30 dBm input. Then tune away from the source frequency to a spacing of, for example, 10 kHz. Note the response in a true RMS reading audio voltmeter attached to the receiver output. Turn the source off momentarily to be sure that the noise decreases, for we wish to measure



The *Ditter* for AGC testing in a receiver. (Thanks for circuit suggestion from K7RO)

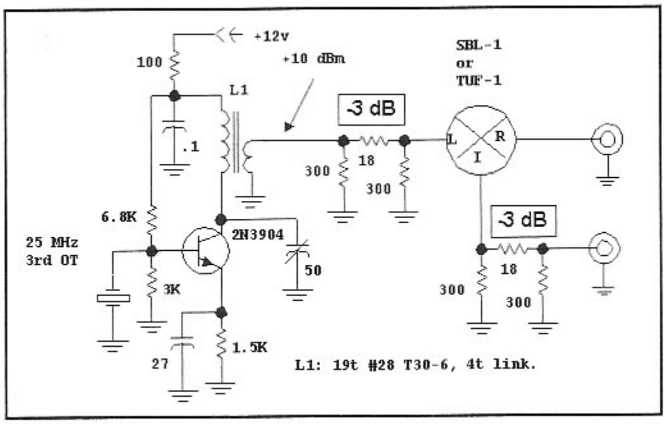


Fig 7.78—Receiving converter for experiments.

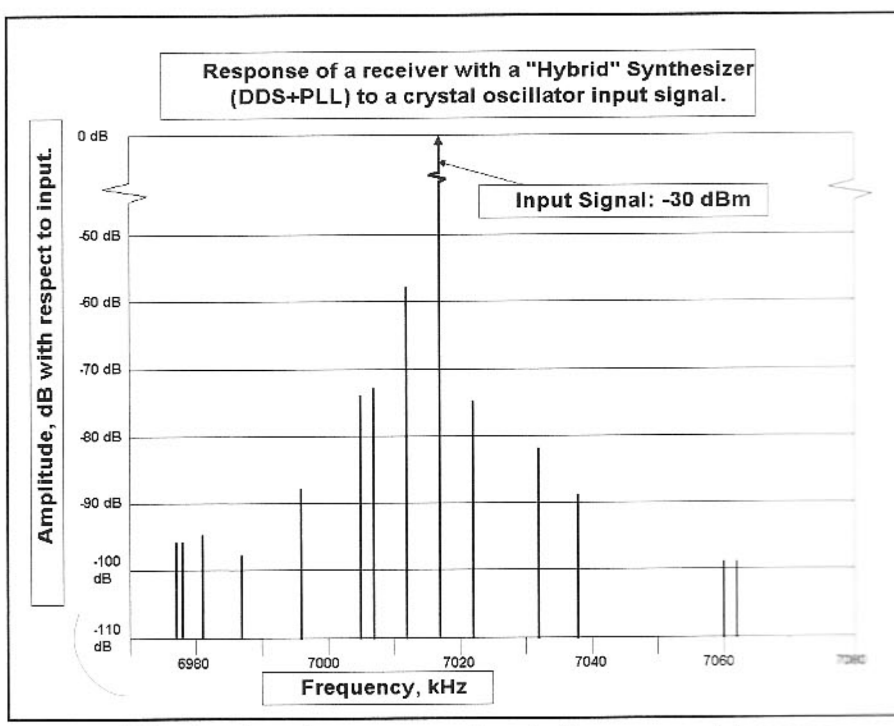
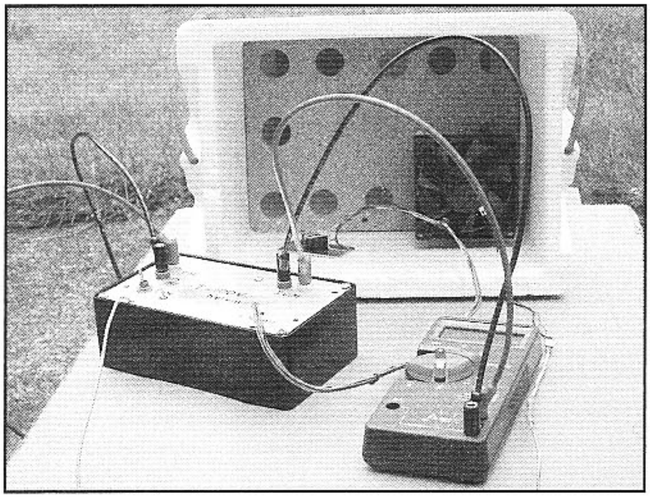


Fig 7.79—DDS-related spurious responses found with a commercial receiver.



Control box, DVM, and "environmental chamber" for oscillator testing. The chamber has the lid removed so an oscillator can be placed inside. The lid is then place on the box. A light bulb heater resides under the press wood base with holes. A 12-V fan moves the air within the box. Cables to the oscillator under test and the IC used for temperature measurement are routed under the lid edge.

the noise above the normal receiver background floor. Having recorded the response at 10 kHz offset, we return the tuning to the input signal. Attenuation is then added to bring the response down to the noise response level. In one measurement of this type reported in Chapter 4, we observed a noise response 110 dB down at a 5 kHz spacing. The receiver being measured had a 500 Hz noise bandwidth, so the spectral density of noise was 27 dB (10×Log[BW]) lower on a per Hz basis, or −137 dBc/Hz. It is necessary to normalize the response related to white (evenly distributed) noise, for that noise will change in proportion to bandwidth. In this example we attributed the observed noise to a VCO being tested, although it could have been the receiver LO. It was still a clean response compared with a typical DDS system like the one of Fig 7.79.

We often see equipment reviews where plots appear showing phase noise. Coherent spurs also appear in these plots. A per-Hz normalization is usually applied to the plot, for that is the most useful information form for pure noise. That normalization may or may not also be applied to the coherent spurs. The normalization, if applied, is not always stated in reviews. This problem disappears when you do your own measurements.

An Oven for Drift Compensation

A photograph shows an oven that we use for the evaluation and compensation of oscillators. The basic measurements were outlined in section 4.2. The "oven" is quite simple, starting with a Styrofoam box purchased at a local super market. The volume is approximately 600 cubic inches. The lower half of that space is occupied with a 60-W light bulb mounted in a ceramic socket attached to a wood strip. The cord for the bulb is run through a hole in the box.

A wood shelf with numerous 1-inch holes divides the box. The upper region contains a small dc fan that can be turned on to circulate the air and enough room for the oscillator module being tested and the temperature measuring circuitry. This oven measures temperature with a National Semiconductor LM3911 integrated circuit that is mounted in a small heat sink and then attached to a small circuit board. The LM3911 has been discontinued, replaced by a much better part from National, the LM45 that is supplied in a SOT-23 surface mount package. The part can be soldered to a small scrap of circuit board with a suitable bypass capacitor and the three wires needed to both power the device and to extract a signal. The output is read with a standard DVM with a sensitivity of 10 mV for each degree C change in temperature.

The oscillator under test is placed in the chamber and the lid is put in place. The oscillator is allowed to warm up while viewing output frequency on an external counter and initial temperature data is read. The light bulb is then turned on, allowing the temperature to climb. It's useful to cycle the bulb off and on, forcing the temperature to increase slowly. Once you have increase T by perhaps 20 degrees C, the fan is turned on for a short burst and the bulb is turned off, forcing the temperature to stabilize. If T seems fairly stable, new frequency data can be measured and **TCF** (**T**emperature **C**oefficient of **F**requency) can be calculated. It is not generally necessary to reach high temperatures, although an initial run up to perhaps 80C will serve to relieve stresses in the inductors resulting from the toroid winding. After a little data has been obtained, the lid can be removed, the bulb turned off, and the fan turned on. This will force the temperature to drop to room value in just a few minutes. The time is used for calculating the value of the temperature compensating capacitors needed.

The temperature compensation process is one that has left us with some very strong impressions:

1. An oscillator that we had regarded as being "pretty stable" with normal components drifts dramatically with the simple oven. This is *not* a minor, subtle effect, but dominant behavior.

2. Once we begin to apply compensation to the oscillator, just 2 or 3 runs will be enough to produce excellent stability.

3. A circuit that started as a "pretty stable" circuit is easily converted to "rock solid."

4. Circuits using really bad components regarding drift (such as varactor diodes) can still yield practical performance.

The whole process is an easy one. The one drawback is that it is somewhat time consuming, so we integrate it with other casual activities.

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CHAPTER 8

Direct Conversion Receivers

8.1 A BRIEF HISTORY

In the early days of radio, signals were collected on a wire, converted from RF voltage and current to audio voltage and current with a crystal detector, and converted to acoustic energy with headphones (**Fig 8.1**). This worked well for spark and later AM broadcast signals, but with continuous waves, the output of the crystal detector was just a very weak dc voltage. A number of schemes were used to convert the CW to AM at the receiver, but the most sensitive method for detecting CW signals on a crystal detector required the use of an oscillator located near the receiver, as shown in **Fig 8.2**. When the oscillator was tuned close to the transmitted signal frequency, audible beats were produced by the crystal detector. The use of a "local oscillator" has been standard in receivers ever since.

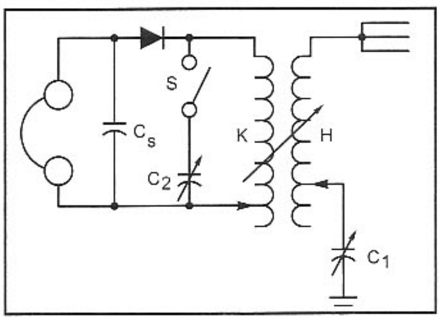


Fig 8.1—A fundamental crystal radio design.

The audible beat signal at the crystal detector is very weak. Early experimenters purchased the most sensitive headphones they could afford, and erected large antennas to collect as much signal as possible. Tuners included adjustments for both peaking the desired signal and achieving maximum power transfer between the antenna and detector. The technology for building highly sensitive headphones was already mature in the early days of radio, because the telephone system predated vacuum tube amplification by several decades. The first application of vacuum tubes in receiver circuits was for audio amplification. The "crystal detector" diode is considerably less sensitive as an envelope detector for AM than it would be with sufficient LO injection to serve as a product detector for CW, but early receiver lore involved using very low level LO injection. RF amplification was needed for AM, and (inevitably) early RF amplifiers using vacuum tubes were marginally stable, which lead directly to the discovery of regenerative receivers. Some RF amplifiers oscillated at two frequencies at once—which lead directly to the discovery of the superregenerative receiver. Cascading two regenerative detectors, one at HF and one at a superaudible frequency around 30 kHz, resulted in the superaudioheterodyne receiver, which was tricky to adjust and received every signal at two places on the dial.

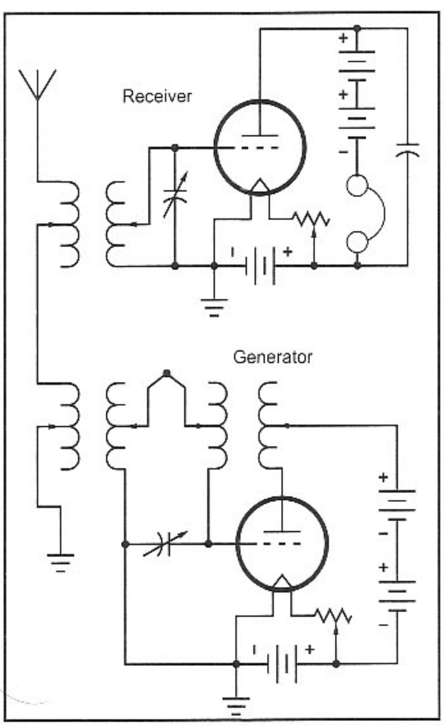


Fig 8.2—A classic radio enhanced with a local oscillator.

Regenerative receivers were simple, inexpensive and worked well enough for amateur AM and CW work that receiver innovation stalled for more than a decade, until the bands became crowded enough that more selectivity was needed. The superheterodyne had been further developed for AM broadcasting, and by the mid 1930s, the transition to the superheterodyne for amateur high frequency work was nearly complete. High Frequency Regenerative receivers remained in *The ARRL Handbook* until the mid 1960s, and superregens are still widely used in toy walkie-talkies, radio controlled cars, and garage door openers.

Signal gain ahead of the detector is desirable if a diode is used to envelope detect AM, but for the linear modes, SSB and CW, the first stage of the receiver may be a lossy frequency converter, directly to audio. Such receivers are capable of outstanding performance at very high frequencies—something to think about the next time a State Patrolman recovers a weak echo from your speeding vehicle with a direct-conversion microwave receiver.

All of the technology—diodes, transformers, local oscillators and audio amplifiers— was available by 1920 to build high-performance direct conversion receivers for CW. There was little motivation for amateurs to develop such receivers at the time because regenerative receivers were adequate, simple and inexpensive. There was also a perception in that era that voice modes were the realm of experimenters and CW the realm of practical communicators. The situation is reversed today, with most technically advanced amateurs experimenting with non-voice modes, from minimalist HF CW stations through microwave systems for 1000-km tropospheric paths.

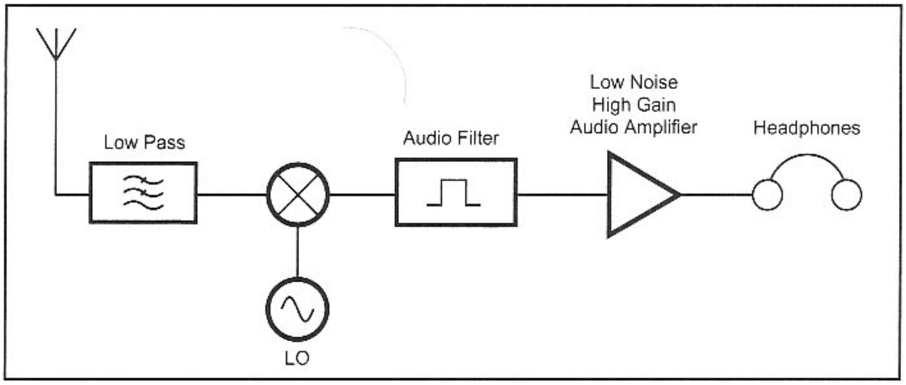


Fig 8.3—A block diagram of a basic direct-conversion receiver.

A radio experimenter is driven not by the desire to duplicate existing circuitry, but by the need to put a station on the air using whatever means are available, preferably without making expensive trips to the parts store. Marginal finances often unleash a wealth of ideas (the philosophy behind PhD programs and other monastic experiences). In the 1960s, when most HF stations operated at the 100 W level, the QRP Society embraced the philosophy of putting simple radio stations on the air and working DX using operator skill instead of transmitter power. Radio experimenters quickly expanded the QRP skill set to include radio design and construction, with an emphasis on elegant simplicity. With the disappearance of AM from the bands, and the emergence of CW as the experimenter's favored mode, the time was ripe for a reexamination of basic receiver circuitry. The '60s implementation of the direct conversion receiver was developed in parallel by a number of independent experimenters. All of the pieces were described in the mid '60s *ARRL Handbook*, but the editors clearly did not envision connecting them together into a receiver without an IF. Even the 1970s *ARRL Handbook* description of direct conversion receiver dynamic range and sensitivity exhibits gaps in understanding.

While the QRP Society provided the direct conversion receiver with a home, their fundamental philosophy also hampered its development. The QRP community embraces simplicity, and many of their designs are indeed simple and only just adequate. Examples of optimizing for simplicity are the numerous NE602 receiver circuits, which have surprising performance for so few parts. The usual first impression upon listening to a simple direct conversion receiver is that it sounds very good, but after making a few contacts most operators want something better. The something better is almost always a superhet. Wes Hayward correctly stated in *Solid State Design for the Radio Amateur*[1] that a direct conversion receiver with audio image rejection is at least as complicated as a simple superhet. This is even truer today, after another quarter century of superhet receiver evolution. The maturity of crystal ladder IF filter design has eliminated IF filter cost as a drawback for superhets, and easy-to-use ICs have reduced parts count below what was possible in the mid '70s.

A small group of experimenters stubbornly continued to develop the direct conversion receiver. Roy Lewallen's design[2] from 1980 is a timeless example of an optimized DSB design with CW filtering, and Gary Breed's 1988 design[3] nicely illustrates the practicality of eliminating the audio image. The KK7B designs published from 1992 through 1995[4-15] were originally intended to serve as VHF tunable IFs with microwave no-tune transverters, but were designed for broadband operation at any frequency from 25 kHz to 5 GHz. These designs have more components than the simplest superhets, but offer several performance advantages including freedom from birdies, ease of use throughout the radio spectrum, and superb in-channel audio fidelity.

By the year 2000, direct conversion receiver designs (**Fig 8.3**) pioneered by amateurs were making significant inroads into practical communications gear including family radio service transceivers, cordless phones, and cellular handsets. The number of papers on direct conversion presented at professional conferences has jumped from a few per decade to over a hundred in one year.

8.2 THE BASIC DIRECT CONVERSION BLOCK DIAGRAM

Fig 8.4 is the block diagram of a direct conversion receiver system for 40 meters. Unlike other figures in this text, the antenna and headphones are included in the diagram. The first block is the antenna. Its function is to collect as much of the desired signal, and as little noise and interference, as possible. While this seems obvious, few amateur or professional engineers actually think about the antenna when designing a receiver system. A 40-M dipole may provide a 1-mV rms noise floor in a 2-kHz bandwidth, during the evening, in the north central United States. Strong foreign broadcast stations may reach millivolt levels. Computer noise and "touch lamp" interference can reach 100-mV levels if the offending appliances are in the near field of the dipole. All of these signals are present at the downconverter.

Another important set of signals present at the downconverter input are FM broadcast stations. In urban areas, FM broadcast signals can produce signals of tens of millivolts in a few meters of wire. The 13th and 15th harmonics of 7 MHz are in the FM broadcast band, and most wideband mixers will downconvert signals near odd harmonics of the LO. The TUF-1 mixer recommended for several projects in this book has 34 dB more loss as a 13th or 15th harmonic mixer than as a fundamental mixer, when measured using a 7-MHz LO. A 1-mV signal at 91.5 MHz (easily obtained on a few meters of wire at KK7B,

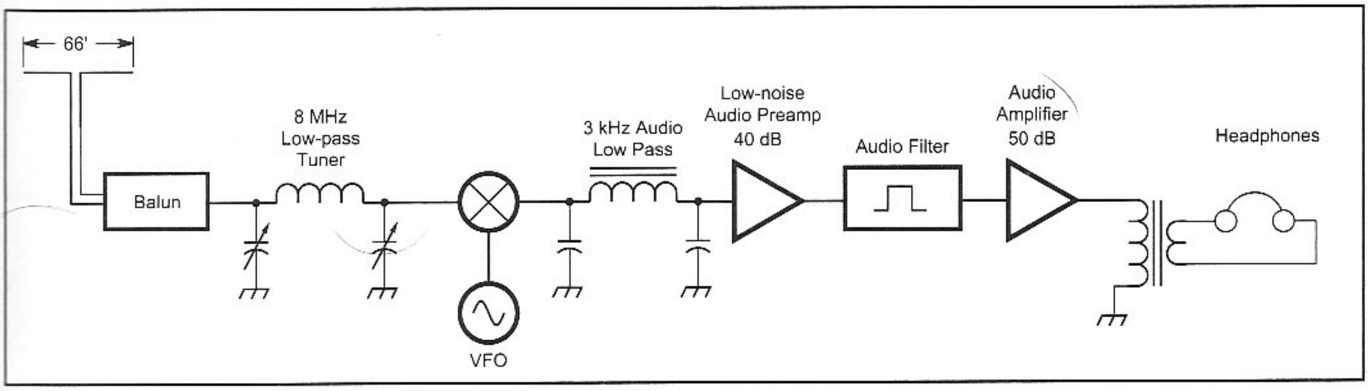


Fig 8.4—Block diagram of a 40-meter direct-conversion receiver.

Portland) is zero beat when the direct conversion receiver LO is tuned to 7.038 MHz, and the 34 dB of excess conversion loss reduces it to the equivalent of a 20-µV 40-meter signal at the antenna. It is easy to prevent these signals from arriving at the RF port of the mixer by using a low pass filter right at the mixer. They are VHF signals, so VHF construction techniques must be used. It is also important to prevent these FM broadcast signals from entering the receiver cabinet on power supply wires, speaker wires, headphone leads, CW key leads and microphone cords—all of which tend to be the right length to make efficient FM broadcast antennas.

The mixer itself can be any of several types, but the diode ring is a good choice for people who want simplicity, good performance, and understanding of how the mixer works. The details of the NE602 schematic are unpublished, and the bias controls to improve its performance are locked in place on the die.

Commonly used mixers have noise figures between 6 and 10 dB, and may have either conversion gain or loss. At first glance, conversion gain would seem to be an advantage. A receiver needs about 100 dB of gain between the antenna connector and headphones, and mixer gain makes the rest of the receiver easier to design. But there is a catch. Mixer gain occurs before any channel selectivity. The filter before the mixer in a direct conversion receiver passes an entire band, and the filtering after the mixer selects the desired signal. The mixer must linearly handle all of the strong and weak signals in the entire band, without distortion. If the mixer has gain, it amplifies all of the strong, undesired signals right along with the weak desired signal. High performance receivers, whether superhets or direct conversion, limit the amount of gain before the channel filter. Thus, minimum-parts-count casually designed receivers tend to have mixers with conversion gain, and more serious receivers have mixers with conversion loss.

Lossy mixers may be either the common diode ring and variations, or made up from transistors used as switches. A number of excellent passive FET mixers have been designed in the past few years, and they are now widely used in a variety of applications.

Mixer gain or loss does not affect receiver noise figure as much as might be suspected. Compare two receivers, each with a 2-dB noise figure, 13-dB gain RF preamplifier. Receiver #1 in **Fig 8.5** has a Mini-Circuits TUF-1 mixer with 5.7-dB loss and 7-dB noise figure, followed by an audio stage with 5-dB noise figure. Receiver #2 in **Fig 8.6** has the same RF preamplifier in front of a Gilbert Cell mixer with 8-dB noise figure and 10-dB gain, driving the same 5-dB noise figure audio amplifier. Using the cascaded noise figure formula presented elsewhere, Receiver #1 has a calculated 3-dB noise figure, and Receiver #2 has a 2.5-dB noise figure.

Now consider the fact that the Gilbert Cell receiver has 23-dB gain before any selectivity, and remember that short-wave Broadcast signals often reach millivolt levels. After the mixer downconverts the entire frequency spectrum present on the antenna and folds it in half around zero Hz, the circuitry connected to the IF port of the mixer selects a narrow portion of the spectrum and then amplifies it. Selectivity between the mixer and first audio amplifier is needed so that the first stage of audio does not have to linearly amplify the entire HF spectrum at once. A simple 10-kHz low-pass filter will narrow the frequency range to just 20 kHz centered around the LO frequency. Further band-limiting is normally included in the audio amplifier stages, but a wide-open direct conversion receiver sounds better on CW and SSB signals than any other receiver type, and should be experienced as a baseline for further receiver experimenting.

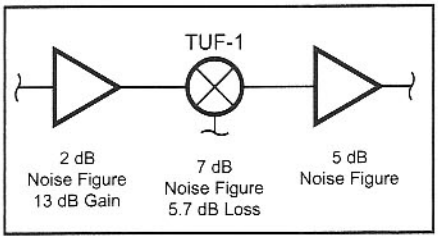


Fig 8.5—A preamp diode ring direct-conversion receiver.

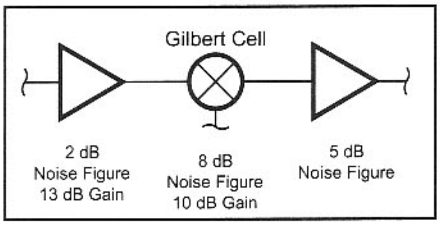
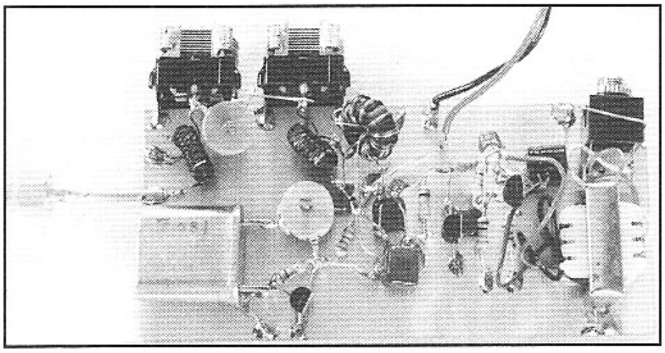
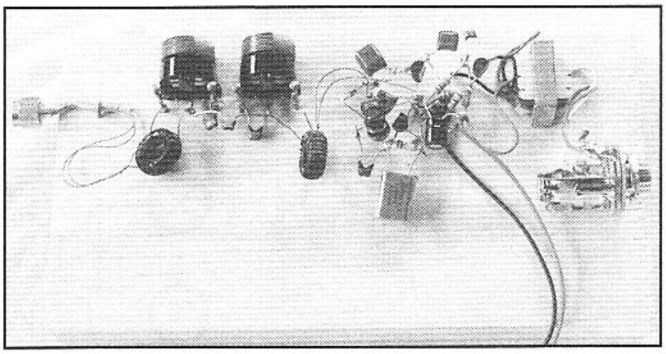


Fig 8.6—Block diagram of a preamp Gilbert direct-conversion receiver.

If the antenna provides 1 µV of noise floor and the headphones require 10 mV for comfortable listening, the receiver needs 80-dB gain. Very quiet locations may have a 0.1-µV 40-m noise floor, and low-sensitivity headphones might require 100 mV—which increases the gain requirement to 120 dB. Receivers without AGC require less gain than receivers with AGC, and also need a different listening style. A receiver described in the next chapter has more than 80 dB of undistorted headroom above the receiver noise floor. Some operators are accustomed to listening for weak signals with the receiver gain turned all the way up, and the receiver noise floor just below the pain threshold. If a click, pop or loud signal suddenly appears in the passband, the receiver is (theoretically) capable of providing an output that will break eardrums and melt headphones. Human ears have remarkable



The "ugly" MicroR1.



The MicroR1 built on a board.

dynamic range. It is far more natural to listen to weak signals 60 dB below the pain threshold and match the receiver in-band dynamic range to the ear's capability.

In previous years the author has merely acknowledged that there are different listening styles, and some styles of listening require AGC more than others. However—two of our close friends (and strongest advocates of AGC), are nearing retirement with serious hearing loss. Both were licensed as novices in the early 1950s, and have spent half a century depending on receiver AGC to protect their ears. Setting receiver gain so that the noise floor from the antenna is well below the pain threshold and training the ears to listen is good hygiene. Weak signals will then be weak, strong signals will be strong, and only rarely will AGC be desired.

A Minimalist Direct Conversion Receiver

Not all direct conversion receivers have to be designed for high performance. Since the historical appeal of direct conversion is simplicity, it is appropriate to present a strict minimalist design. Simple NE602 based circuitry is presented elsewhere in the text. For this circuit, the use of specialized components is avoided. The receiver in **Fig 8.7** has each of the functional blocks from Fig 8.3. Q1 and its associated components is a simple Pierce oscillator. With the component values shown, it oscillates with every crystal tried from the author's junk box. The frequency may be trimmed a few kHz with a small (about 20 pF) trimmer capacitor in series with the crystal. Since both ends of the trimmer capacitor are floating, an insulated tuning tool or shaft should be used.

T1 is 10-trifilar turns of enameled wire on a FB 2410 ferrite bead. A transformer made of ten trifilar turns of plastic covered bell wire on a large ferrite RFI suppression core salvaged from a computer printer cable also works well. Diodes are 1N4148 or similar, and the three transistors are 2N3904 or similar small-signal NPNs.

The two stage audio amplifier has more than enough gain to bring the 40-m band US West Coast noise floor up to the audible level in portable CD player headphones. Coupling and feedback capacitors were selected by ear and back-of-the-envelope calculations from available values in the author's junk box. Gain is intentionally kept low for ear protection, and to eliminate the need for special construction techniques, a volume control, or shielding. The double tuned circuit on the RF input solves any harmonic mixing or AM broadcast detection problems, and the three adjustments may be tweaked to optimize signal power transfer from the antenna to the receiver. When signals are strong and shortwave broadcast interference is a problem, the coupling capacitor may be reduced and the input circuit optimized for desired signal-to-interference ratio rather than just maximum signal strength. The independent 9-V battery supply, balanced antenna and headphone connections, and no external ground connection eliminate ground loops and common mode problems. Current drain from the 9-V battery is about 8 mA.

This simple receiver is fun to listen to, particularly when it is open on the bench with all parts visible, and signals from 10,000 km away are rolling in. The accompanying photos show two different construction styles. Parts may be purchased new, or salvaged from old computer boards and transistor radios.

The receiver described in the preceding paragraphs is a nice illustration of how simple a "real" communications receiver can be. It also illustrates some of the challenges of simple receivers. Crystal control strictly limits tuning range, and limited selectivity requires skill in digging signals out of crowded bands. The challenges inherent in simple equipment are not necessarily disadvantages—it takes more skill to cross a harbor in a sailing dinghy than a motor boat. Copying signals from across the oceans with a three transistor circuit is similarly rewarding.

Just as sailors always want a bigger boat, radio experimenters always want to improve their receivers. The following paragraphs dig into the technical fundamentals needed to understand direct conversion receivers at a depth that allows performance to be pushed to superhet levels and beyond.

Direct Aversion

Before proceeding with the technical discussion, it is worthwhile to note that many otherwise rational human beings have an emotional aversion to direct conversion receivers. The basic block diagram is so simple and appealing than many unsuspecting designer-builders and engineering managers have fallen into the trap of believing that direct conversion is the "holy grail" of receivers, able to outperform the old, obsolete superheterodyne architecture at a fraction of the cost. Most attempts to build something cheaper and better than an existing, mature technology will fail. When the holy grail turns out to be a cracked clay cup, the designer involved may end up with a lingering bad taste in his mouth. Experienced professional and amateur technical writers tend to either love or hate direct conversion receivers, and this bias has often appeared in print.

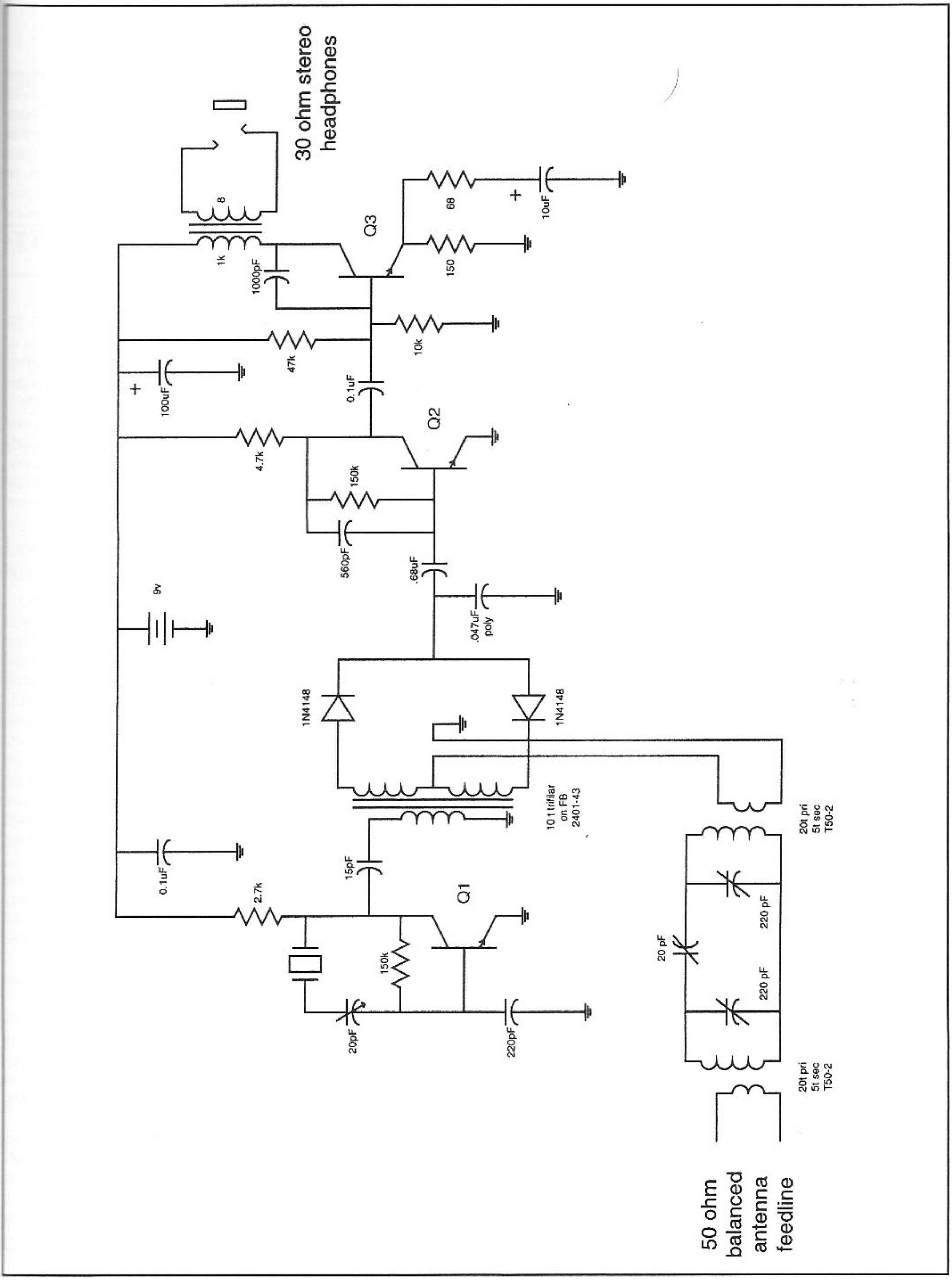


Fig 8.7—The schematic of the MicroR1.

8.3 PECULIARITIES OF DIRECT CONVERSION

The level of understanding represented in the preceding paragraphs is enough to build direct conversion receivers and use them to make contacts on the amateur bands, but they will exhibit some strange behavior that is not explained by conventional superhet thinking. Explaining the peculiarities of direct conversion receivers, and more importantly, designing and building a new generation that outperforms previous attempts, requires further study and a deeper understanding.

High Audio Gain

There are significant differences between the block diagrams and gain distributions of superhets and direct conversion receivers. Direct conversion peculiarities fall into two classes: problems from high audio gain and the effects of local oscillator radiation. AM demodulation, a common problem with direct conversion receivers, is a symptom of both high audio gain and LO radiation.

A typical direct conversion receiver has about 100 dB of gain from the mixer to the output. The output might be a 1-mA current flowing in a wire to the headphone jack. The ground wire coming back from the headphones also carries 1 mA. If the ground wire has 1 milliohm resistance, the voltage drop will be 1 µV, which is 100 times larger than the weakest audible signals. This sets up an ideal condition for audio oscillation or regeneration. Since it is impractical to reduce the resistance of all ground wires (#24 copper wire has about 2 milliohms per inch), it is very important that any ground return carrying output signals be separated from any input signal ground return. The easiest way to insure this is to use a separate ground wire for every component, and connect them all together a single point. It is particularly important to treat the speaker or headphone jack as a component, and bring it's ground lead all the way back to the common ground connection rather than just grounding it to the radio case. This bears repeating: use two wires, a signal and a ground wire, to connect to the headphone jack or speaker, and do not ground the speaker or headphone jack to chassis ground. With a simple receiver, it is possible to actually connect the grounded leads of all components to the same point. **Fig 8.8** is a schematic showing how this can be done with the receiver in Fig 8.7. There are also magnetic and capacitive feedback mechanisms that become important at audio with 100 dB of gain. Often oscillations can be cured by moving around the wires carrying audio signals and power.

Inductors in the early stages of a direct conversion receiver should be of a self shielding type. Conventional Iron E core audio transformers are best avoided, although they have been successfully used on the input to high gain audio amplifiers in direct conversion receivers with several layers of magnetic shielding. The Toko 10 RB series of shielded inductors has been used for years, although the shielding is not perfect and they will pick up hum from nearby transformers. A small steel or mumetal enclosure around the audio preamp stages of a direct conversion receiver can reduce hum pickup by many

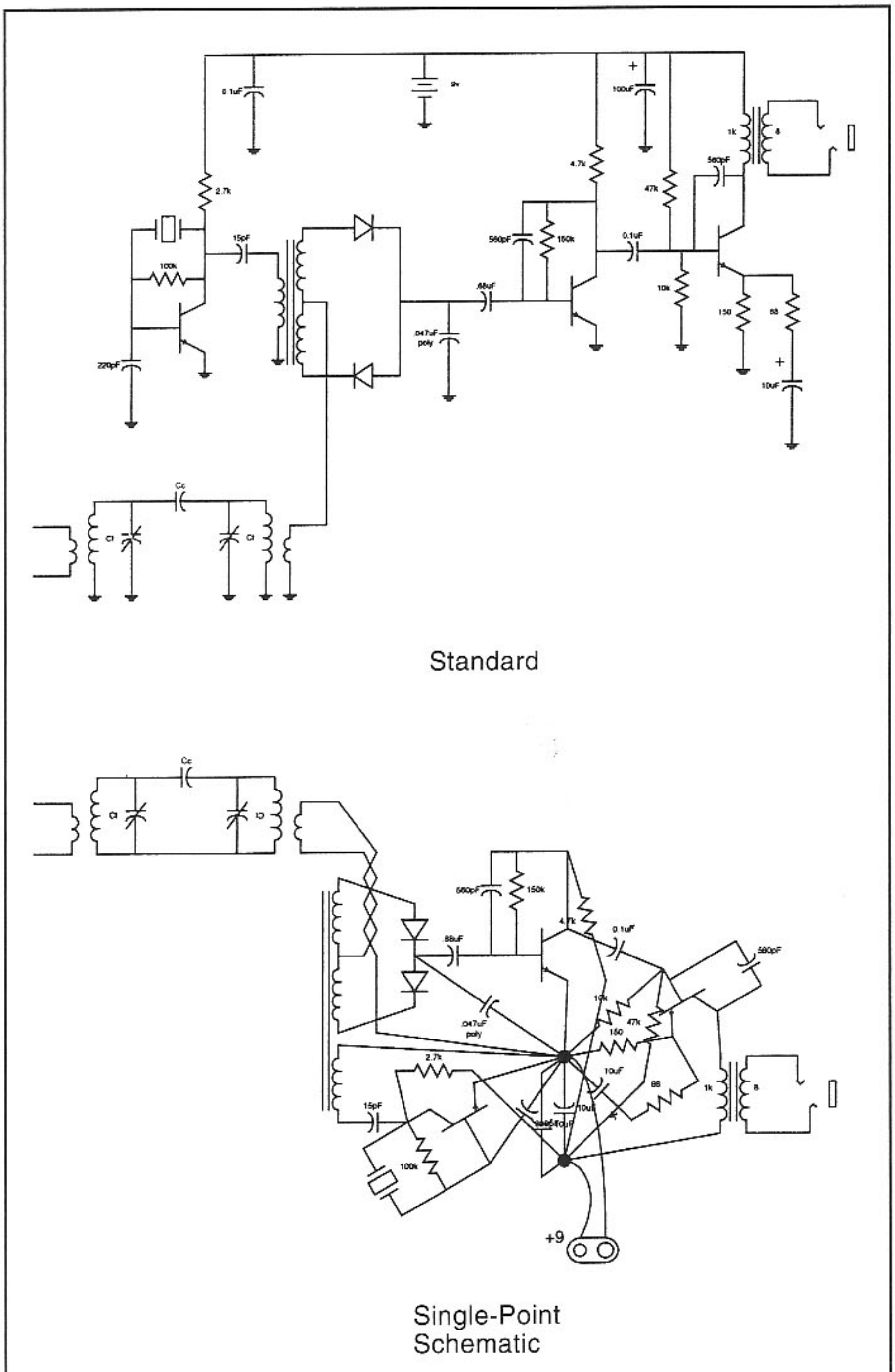


Fig 8.8—Compare the "standard" MicroR1 schematic above to the single-point schematic below.

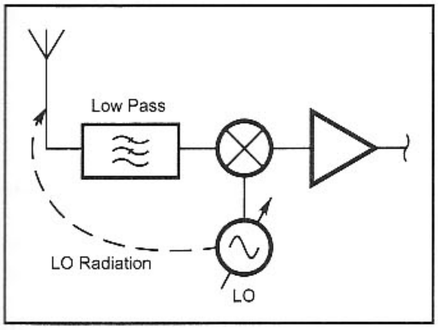


Fig 8.9—Local oscillator radiation.

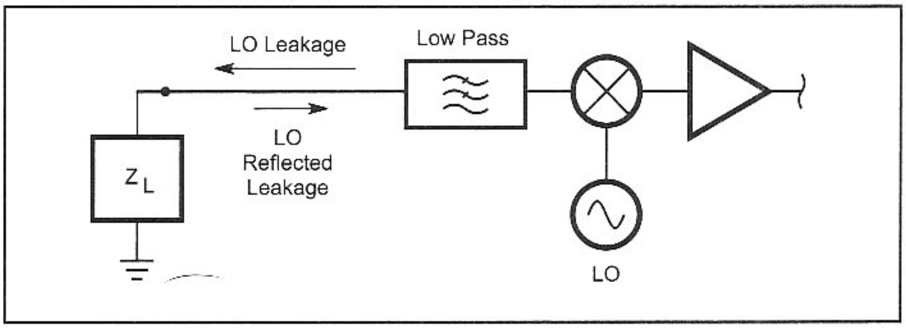


Fig 8.10—A mixer/LO with reflection coefficient.

dB. Good direct conversion receivers tend to include high-pass filters in the audio chain, aggressively rolling off the audio response below about 300 Hz.

Microphonics, the loud clicks and pops when the receiver is bumped, are often blamed on high audio gain, but they are actually a symptom of Local Oscillator radiation, and can often be cured by improving receiver shielding.

Local Oscillator Radiation

Local oscillator radiation raises a whole new set of problems. **Fig 8.9** shows a simple direct-conversion receiver front end with local oscillator radiation arriving at the RF input port of the mixer. Since the LO is at the RF frequency, there is no possibility to use RF selectivity to reduce the level of LO at the mixer RF port (in a superhet, the LO and RF are separated by the IF, so the RF selectivity necessary for image rejection usually reduces the LO signal between the antenna and RF port of the mixer). At first glance, it appears that the LO signal at the mixer RF port will have no practical effect, because it is exactly zero beat. The mixer multiplies the RF port LO signal with the LO, and the output is pure dc:

$$\text{low pass}\ \{a \cos(2\pi f_0 t + \phi) \cos(2\pi f_0 t)\} = a/2 \cos \phi \qquad \textbf{Eq 8.1}$$

...where f_0 is the LO frequency, a is the amplitude of the LO leakage, and ϕ is the phase difference between the LO and LO leakage.

DC at the IF will unbalance a balanced mixer, which causes it to radiate more LO. The additional LO radiation might be reflected by nearby objects or an imperfect antenna match. If the new term is in phase with the original radiated LO, this will further unbalance the mixer. Thus the amount of LO radiation is a function of the physical environment near the antenna. This is not usually a problem at HF with large outdoor dipoles, but HF direct conversion receivers commonly exhibit disappointing performance with wire antennas connected directly to the back of the radio. A changing local electromagnetic environment around the antenna can be a particular problem at VHF and microwaves where antennas are small and good reflectors are numerous.

LO radiation and pickup by the antenna becomes more significant when either the amplitude or phase of the LO signal at the RF port of the mixer is time dependent. There are three major classes of time variation in the LO signal: transients, Doppler and modulated scatterers. Each of these will be treated separately.

Transients in LO radiation and reflection

One of the major annoyances with direct conversion receivers is microphonic clicks and pops when anything in the system experiences a mechanical change. **Figure 8.10** shows a mixer and LO system connected to a high-gain audio frequency IF amplifier and a load with some arbitrary reflection coefficient. As an example, suppose that the mixer is a Mini-Circuits TUF-1 and the LO is at 50 MHz. The data sheet shows 57 dB of LO to RF port isolation in this mixer at 50 MHz. With a +7 dBm LO, –50 dBm of LO power leaves the RF port of the mixer and is reflected from the load connected to the RF port. Let's pick an arbitrary reflection coefficient, say 0.2 at an angle of 45 degrees, for the load. The magnitude of the reflection coefficient will stay the same, but the angle will change as we vary the length of 50-Ω transmission line connecting the mixer to the load. –50 dBm in a 50-Ω system is 1 mV peak. The magnitude of the reflection is (0.2)x1 mV or 200 µV. The 200-µV signal reflected from the load arrives at the mixer, and with 6-dB conversion loss and the appropriate phase, becomes a 100-µV dc voltage at the IF port of the mixer and input to the audio amplifier. This voltage is too small to seriously unbalance the mixer, and is blocked from the following audio amplifier by the series input capacitor. However, if the connection to the load is broken, for example, by disconnecting the BNC connector, the reflection coefficient jumps from 0.2 at 45 degrees to 1.0 at some other angle. The signal at the RF port of the mixer jumps from 200 µV at some phase to 1 mV at some other phase. At the IF port, the signal jumps from 100 µV dc to 500 µV dc. The "before" and "after" voltages are both dc, but the jump between them is a transient, and is amplified by the audio amplifier. The output of the audio amplifier with a short transient into the input dc blocking capacitor is the impulse response of the amplifier. (If we recorded the shape of the amplifier output pulse on a digital oscilloscope, we could then perform an FFT and see the frequency response of the amplifier.) 400 µV is a big signal, and probably drives the amplifier into saturation. The output is a very loud pop in the headphones. The level of LO isolation in a direct conversion receiver can be quickly judged by simply disconnecting the antenna while listening. A loud pop indicates poor LO isolation.

As shown in equation **Eq 8.1**, the dc output of the mixer depends not only on the level of the LO signal at the RF port, but also on its phase ϕ. An abrupt change in phase with no change in reflection coefficient magnitude will also induce a pop in the headphones.

Mixer LO port to RF port isolation is only one way for LO to leak out of the system and return to the RF port. Any leakage from the LO compartment results in a signal that may be picked up by the antenna. Often a direct conversion receiver that works exceptionally well in the lab when connected to signal generators exhibits all manner of peculiar behavior when connected to an antenna. As long as the LO leakage is small and doesn't change with time, there will be no observable effects. If the LO leakage changes suddenly, however, there will be an audible response. A loose screw in a metal radio cabinet can cause a scratching sound when the radio is tuned, by changing the amount of LO that leaks out of the case and is picked up by the antenna. Direct conversion receivers that work well when first packaged in a shiny new aluminum enclosure often become microphonic as they age and the mating surfaces corrode. Direct conversion receivers soldered up in boxes made from copper-clad PC board age more gracefully.

Doppler Effects

Since direct conversion receivers can detect differences in the phase of a reflection, they are very sensitive to reflections from moving objects. Doppler becomes most important when the motion is fast enough that the Doppler modulation on the radiated LO signal is in the audio amplifier passband (**Fig 8.11**). The maximum Doppler shift for a signal radiated from point A, reflected from a moving object at point B, and received again back at point A is:

$$\text{Doppler Frequency} = 2\,V_o/\lambda \qquad \textbf{Eq 8.2}$$

At 40 m, an airliner (**Fig 8.12**) passing directly overhead at 500 miles per hour (220 m/s) would induce a Doppler shift of 2x220/40 = 11 Hz. Airliners don't normally fly that fast when they are close to the ground, and 11 Hz is well below the audio range of interest, so we can ignore Doppler effects at HF. At 2 m, the Doppler shift from a 500 MPH airliner is 220 Hz, but airplanes flying that fast are normally a long way from the antenna. At microwaves, however, the story is entirely different. A 10368 MHz direct conversion CW receiver with LO leakage can detect all kinds of moving objects. With 3 cm wavelength, the Doppler shift from the airliner becomes 2 × 220/0.03 = 14.7 kHz which is at the top of the audible range. Cars at 50 MPH, however, have 1.47 kHz echoes, right in the middle of the audio passband for a conventional receiver. An audio phase-locked loop to recover the weak echo and an audio frequency counter can be used to remotely measure the speed of automobiles at ranges out to a mile or so, with very little radiated LO power. The direct conversion microwave receiver is sensitive not only to constant motion, but to vibration as well. Above 1 GHz extra care should be taken to make antennas for direct conversion receivers mechanically rigid. Some types of antennas, like horns, are less susceptible to reflecting surface vibrations than dish antennas, and Yagi antennas with mechanically resonant elements will induce spectral lines in the receiver audio output that can be seen using an audio FFT analyzer.

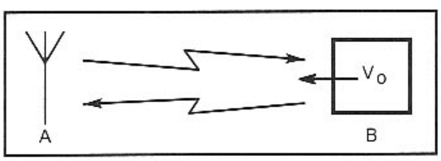


Fig 8.11—An illustration of RF Doppler.

It is a useful exercise to estimate how far away objects can be and still produce Doppler effects in a receiver. Assume we have a 2-m receiver with very poor LO isolation, radiating 0 dBm from the antenna. Radiated power density (in watts per square meter) falls off as the surface of an expanding sphere:

$$\text{Power Density}\left(\text{watts/meter}^2\right) = \frac{P_o}{4\pi R^2} \qquad \textbf{Eq 8.3}$$

where P_o is the total radiated power and R is the distance between the source and the power detector

At 1 km, the power density is about 10^{-10} watts/m^2. Suppose this radiated LO energy bounces off of an airliner 1 km away with an effective radar cross section of 100 m^2. 10^{-8} watts will be bounced of the airliner. The spherically expanding scattered wave will have a power density of about 10^{-15} watts/m^2 after traveling the 1-km distance back to the receiver. A 2-m dipole has an effective capture area of about ½ m^2, so the signal bounced off of the airliner is about 5 x 10^{-16} watts, or −123 dBm at the receiver antenna terminals. This is about 10 dB above the noise floor of a typical SSB receiver.

A more typical receiver will have much lower LO radiation, but moving objects within 10 meters of the antenna often result in a detectable output in the antenna. A half-wave dipole with a toggle switch in the middle is a useful VHF direct conversion receiver diagnostic tool. If you can hear the switch click in the headphones, you are detecting LO radiation.

Tunable or Common Mode Hum

One of the direct conversion receiver peculiarities that puzzled early workers is the phenomenon of tunable hum. Receivers would have a particularly ragged sounding ac line noise hum that varied with changes in receiver tuning. This hum was particularly annoying in receivers that used a single high-Q tuned circuit at the RF port of the mixer—the common form of early direct conversion receiver. There were numerous theories for tunable hum—a few of them humorous in hindsight. In typical amateur fashion, lore developed that offered a set of fixes for tunable hum, including using an outdoor balanced antenna, using ferrite beads on the power supply leads, and using a battery power supply.

There is a difference between wisdom (don't eat raw pork) and understanding (Wow! Look what we see under the microscope!). Wisdom comes from experience, and understanding comes from study. For practical people like radio amateurs, wisdom usually comes long before complete understanding. Unfortunately, with the

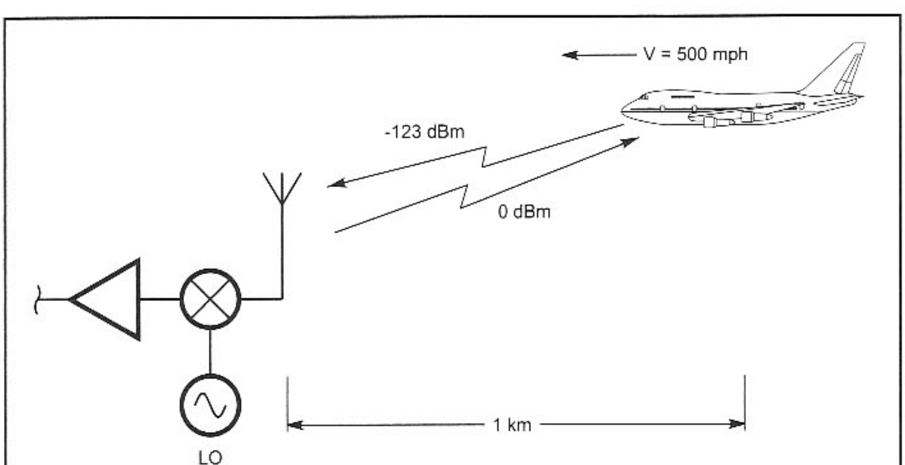


Fig 8.12—2-m radiation from an airplane 1 km away.

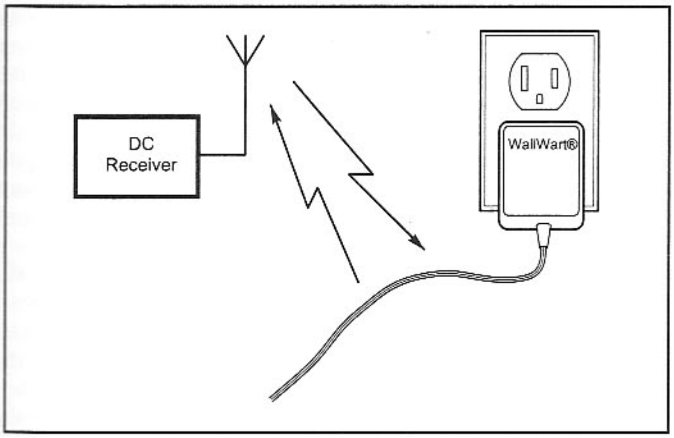


Fig 8.13—A tunable hum experiment.

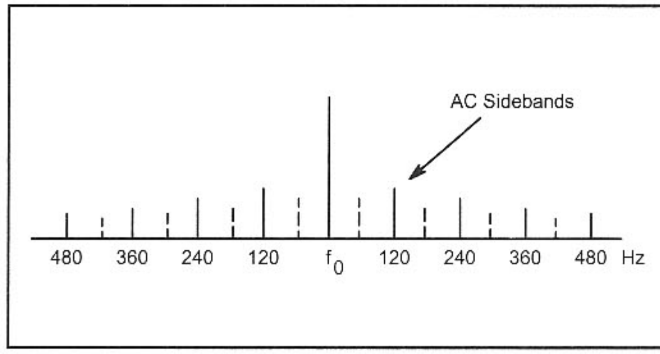


Fig 8.15—The spectrum of a re-radiated LO.

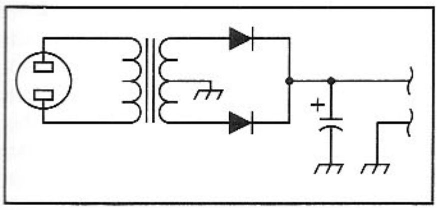


Fig 8.14—A power supply schematic.

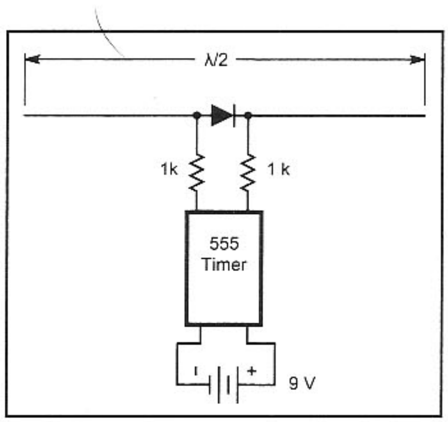


Fig 8.16—A hum probe.

proliferation of computer design, we are entering an age where folks are reluctant to do anything that can't be modeled mathematically and simulated. It is a good thing our ancestors weren't saddled with such nonsense, or they would have continued sticking their hands in the fire until medical science told them to stop. On the other hand, it is understanding that permits us to push the state of the art.

We now understand tunable hum well enough to dispense with the ferrite beads on battery power supplies and use indoor antennas on direct conversion receivers if we must, but much of old lore is still good. Battery supplies and a full-size outdoor antenna are recommended for reasons other than hum elimination.

Fig 8.13 shows a typical tunable hum experiment. The direct conversion receiver is connected to an antenna directly on the back panel. Right next to the antenna is a power cord going to a plug-in dc power supply. The power supply cord is a parasitic element of the antenna system. The power supply schematic is shown in **Fig 8.14**. Note that the power supply schematic is almost identical to the diode balanced modulator in the previous chapter. The modulating frequency is 120 Hz, due to the full-wave rectifier. The LO is picked up from the antenna wire, and then re-radiated with the 120-Hz sidebands. This wouldn't be much more than an annoyance, except that the 120-Hz modulating waveform is very rich in harmonics. The spectrum of a typical re-radiated LO signal is shown in **Fig 8.15**. The LO signal itself is at dc, and doesn't make it through the audio amplifier (although it may unbalance the mixer—increasing the strength of the radiated LO), but the sidebands are recovered by the mixer, and particularly the higher harmonics at 240 Hz, 300 Hz, 420 Hz etc. are subject to the full gain of the audio amplifier.

This explains the hum, and the harmonic content explains the raunchy sound, but why is it tunable? Refer again to equation **Eq 8.1**. The IF output of the mixer is a function not only of the amplitude of the signal at the RF port, but the phase ϕ. In fact, if the phase of the LO signal at the RF port is exactly 90° different from the LO drive, there is no detection of the sidebands at all. With a sharp single tuned circuit on the RF port, the phase varies more rapidly than the amplitude response as the tuning moves through resonance. At resonance, the phase shift through the tuned circuit will be zero, but off resonance the phase will smoothly tune from +90° to –90°. If there is some other phase shift path from the LO to the RF port of the mixer (there usually is), then at some point in the RF tuning, the hum will drop into the noise floor. Often the hum is eliminated at a point in the tuning where the sensitivity has been reduced to an unacceptable level.

It is interesting to observe that tunable hum is absent from *image-reject* direct conversion receivers. Common mode hum may still be present, but it is not tunable. An image-reject direct conversion receiver has two mixers with LO (or RF) ports 90° out of phase. After some baseband phase shifting, the IF outputs of these two mixers are added. If one mixer has zero common-mode hum, the other will have maximum hum. The sum will then have constant common-mode hum, regardless of any phase shifts in space or in the receiver RF path. Experimenters with image-reject direct conversion receivers who break the I and Q signal paths and listen to each channel separately often complain that "one channel has a lot of hum, but the other is fine" and try to eliminate the hum in the "bad channel" with improved bypassing and power supply decoupling, which is, of course, ineffective.

It is interesting to study receiver LO leakage with a "common-mode hum probe" consisting of an antenna, diode modulator, and modulating signal source. A modulating tone should be chosen that is not harmonically related to 60 Hz. At HF and VHF, a small loop antenna with a diode and a 555 timer works well. At microwaves, a dipole consisting of a diode and its leads serves well. **Fig 8.16** illustrates the circuit. If the probes are small enough, they may be used to find the LO leaks in a direct conversion system.

Eliminating LO Radiation Effects

Understanding common mode hum and

other LO radiation symptoms allows us to eliminate them. If we do not permit any LO signal to leak out into the RF environment around the antenna, then common mode hum cannot occur. There are several primary leaks that we must consider:

1. LO coupling through the mixer to the RF port and through the RF circuitry onto the antenna.

2. LO energy radiating from LO components on the circuit board.

3. LO energy on wires connected to the radio cabinet (**Fig 8.17**).

Reducing the amount of LO energy at the antenna connector involves mixer LO to RF port isolation, eliminating coupling from the LO components into the RF stages, and the reverse isolation of any amplifiers in the system. There are big differences in the LO to RF isolation of various mixers. Some unbalanced mixers have no LO to RF isolation at all. The mixers most suitable for direct conversion receivers are balanced. At 7 MHz, the LO to RF isolation of a TUF-1 mixer is more than 70 dB and the SBL-1 is around 65 dB. This is sufficient for acceptable direct conversion receiver performance with no RF amplifier. At 144 MHz, the TUF-1 LO to RF isolation has dropped to 50 dB and the SBL-1 has dropped to 45 dB. This is low enough to cause problems.

Additional isolation can be obtained by using an RF amplifier ahead of the mixer, as recommended in the excellent papers by Nick Hamilton.[16] This is good practice even at lower HF bands where an RF amplifier may not be needed for noise figure. It is important to note that reverse isolation varies widely between amplifier types. A Mini-Circuits MAR-2 with 12.5-dB gain has only 18-dB reverse isolation at 144 MHz, while a grounded gate U310 with 10-dB gain has 28-dB measured reverse isolation. A cascaded pair of grounded gate U310s on the input to a direct-conversion 2 m receiver can effectively eliminate LO energy coupled through the mixer through the RF amplifiers onto the antenna. At microwaves the differences can be even larger. The 12.5-dB gain MAR-2 has reverse isolation of 17 dB at 1296 MHz, while the 16-dB gain TriQuint 9132 has more than 45-dB reverse isolation.

Even if the mixer has good LO to RF port isolation and the RF amplifier has good reverse isolation, the LO can still couple onto the antenna connector if there is no shielding inside the radio case. The antenna connector should connect to the RF amplifier input with small coax, properly grounded at each end.

All of the components in the LO circuit can radiate LO energy. To gain some intuition for how effective components are as antennas, compare their size in wavelengths to the size of a mobile whip antenna on 80 meters. A typical mobile whip might be two meters tall, 0.025 wavelengths at 80 m. In a 40-m VFO, the individual components are very small in wavelengths, and would therefore make poor radiators. In a 2-m VFO, 0.025 wavelengths is only 0.05 meters, or about two inches. A two-inch long PC board trace could be as effective a radiator as an 80-meter mobile whip. Small magnetic antennas can be very effective. Think about the size in wavelengths of an AM radio ferrite loopstick. Small tuning coils and RF chokes are often the most significant sources of LO energy inside a radio cabinet. The use of shielded coils and toroids is recommended for all direct conversion applications. The most effective way to prevent LO radiation from components is to enclose the entire LO in a shielded enclosure. Small tin cans work well, and can be easily soldered in place. A PC board enclosure with soldered seams is superior to a machined aluminum box held together with screws.

It is meaningless to enclose the LO if there are holes in the enclosure with wires going in and out. The wire will pick up energy inside the box and conduct it outside, where it can be radiated or conducted onto other wiring. The LO signal itself should come out through coax or a coax connector, and dc wiring should use effective feedthrough capacitors and decoupling networks. The most careful VFO compartment shielding can be rendered useless if the VFO capacitor shaft goes through a hole in the compartment wall. Capacitor shafts can be significant radiators if they are not grounded to the wall near the entry hole (**Fig 8.18**). At VHF, a few inches of tuning control shaft through the radio panel can couple LO energy to the outside world. A grounded panel bearing is one option, but the common 1/4-inch sleeve types don't provide reliable grounding, and will result in common mode scratches as the radio is tuned. A better solution is to use a grounded sleeve bearing with a 1/4-inch non-metallic rod for the tuning shaft, and a shaft coupler to the capacitor shaft inside the sealed VFO compartment.

The same rules for keeping LO energy from radiating to the inside of the radio box and being picked up by the RF circuitry apply to keeping LO energy from radiating to the outside world on power supply,

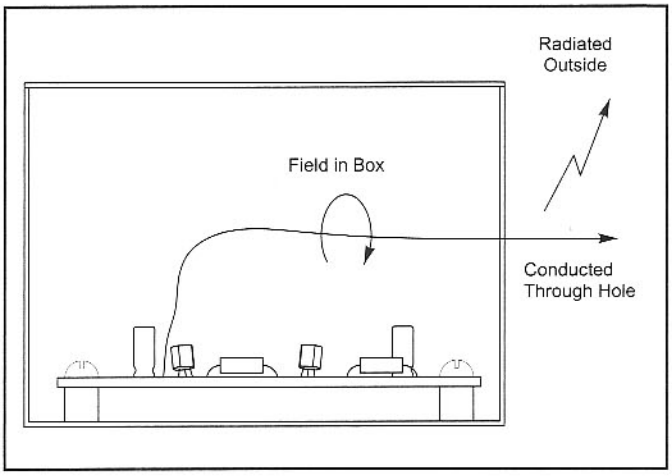


Fig 8.17—A wire pickup in an LO box.

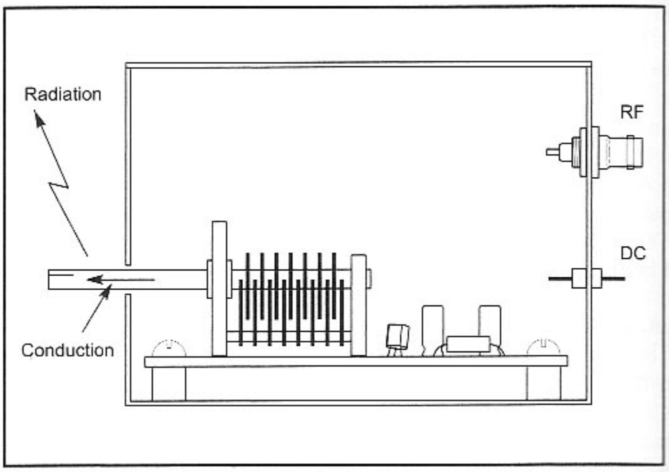


Fig 8.18—Capacitor shaft pickup in an LO box.

speaker, microphone and key leads. All dc and audio leads should be properly decoupled for RF. This can be a problem for speaker leads, since bypassing them to the chassis of a direct conversion receiver with high audio gain will introduce ground loop feedback. One way around the problem is to use a separate powered speaker, preferably with internal batteries, plugged into the headphone jack of the receiver.

A conservatively designed and built direct conversion receiver is double shielded, with internal enclosures around the VFO and RF circuitry, often a small steel or mumetal enclosure to reduce ac hum pickup around the audio preamp inductors, and an outer shielded enclosure. All RF connections are made using shielded connectors, preferably BNC at HF and SMA at VHF and up, and all dc and audio connections to the outside world properly bypassed. Care is also exercised so that mechanical connections like volume controls and the main tuning knob shaft do not conduct signals into or out of the receiver enclosure.

One technique that has been part of the lore for years is using a VFO followed by a frequency doubler. A balanced mixer is insensitive to energy at 1/2 or twice the LO frequency. The expression below shows multiplication of a low level 1/2 frequency signal with the LO. There is no output at dc.

$$a \cos [2\pi(2f_0)t + \phi] \cos 2\pi f_0 t = a/2 \cos [2\pi(3f_0)t + \phi] + a/2 \cos [2\pi f_0 t + \phi] \qquad \textbf{Eq 8.4}$$

Care must be taken to avoid radiating the frequency doubled signal, but a passive doubler right at the mixer port could be used. Then only the actual doubler circuitry must be shielded, and there are not even any dc power leads connected to stages carrying the on-frequency LO signal. In particular, the VFO shaft and capacitor body only have half-frequency energy, and may be left unshielded. The 40-m sleeping bag radio described later was built to test frequency doubling, and there is no separate shielding around the half-frequency VFO. As a fringe benefit, a CW transmitter using a frequency doubled VFO is much less susceptible to chirp than one with the VFO operating directly on frequency.

It might seem that it takes an awful lot of extra effort to build a good direct conversion receiver than to build a good superhet. This is not true. A *good* superhet requires exactly the same construction. Superhet receivers with poor shielding have a different set of problems, like multiple internally generated spurious responses, poor image and IF rejection, and responses to strong out-of-band signals near harmonics of the oscillators. Good mechanical construction, shielding of individual stages, and proper bypassing and decoupling of power supply and audio leads makes a tremendous improvement in performance, whether the receiver is a conventional superhet, direct conversion, or a spectrum analyzer. Good mechanical construction is too expensive for mass produced or even kit radios, but is just a matter of planning, care, some worthwhile mechanical skills, and time for a designer-builder of a single radio. This is one area where a designer-builder can far exceed the mechanical quality and electrical integrity of a mass-produced receiver built under severe time and budget constraints, for example, a Collins 75S3C.

Adaptive Mixer Balance

Some balanced mixer types may be easily adjusted for LO radiation. The familiar "carrier balance" resistor adjustment in Gilbert Cell mixers is an example. It is possible, in concept at least, to measure the instantaneous LO level at the receiver antenna terminal, and vary a set of voltages in the mixer to force the LO leakage to zero. This technique permits eliminating not only stray LO energy from inside the mixer, but energy that arrives via other paths by canceling it with an equal-and-opposite mixer leakage signal. The mixer adjustment may be done once, during alignment or each time the radio is powered up, and then the balance adjustment locked in for normal operation.

There are sobering cautions that need to be mentioned. If the balance adjustment is done continually in real time, it must be recognized that adaptively nulling a signal by adding a sine-wave adjusted for precise amplitude and opposite phase is a form of phase-locked-loop. Since both phase and amplitude are variables, loop stability analysis becomes complicated. Designing an LO suppression loop that offers real benefit and remains stable over a wide range of operating conditions is an ambitious exercise. Another difficulty is that intentionally unbalancing the mixer to obtain a precise amplitude and phase carrier signal will null the LO at the expense of mixer 2nd order distortion performance.

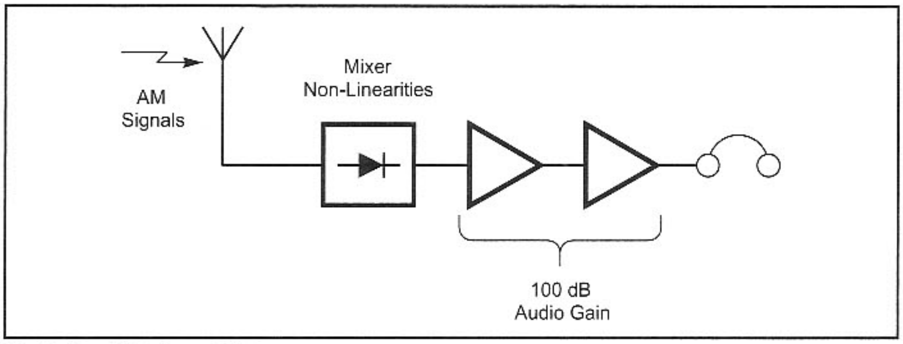


Fig 8.19—AM demodulator.

AM Demodulation

A common problem with direct conversion receivers is demodulation of AM signals anywhere in the RF passband of the receiver. This is most often observed on 40 m when foreign broadcast signals are very strong. **Fig 8.19** illustrates the problem. Any mechanism in the mixer that produces a dc output at the mixer IF port from a signal at the RF port will result in the envelope of an AM signal appearing as weak audio, right at the input to a 100-dB gain audio amplifier. DC outputs occur when a mixer has second order distortion. Second order distortion is common when balanced mixers become unbalanced. Since the usual way that balanced mixers unbalance is the presence of LO signal at the mixer RF port, it is evident that AM demodulation is a symptom of both poor LO to RF isolation and high audio gain. Improving the shielding around the VFO, and LO to RF isolation often improve a receiver's immunity to AM demodulation. Receivers that use VFOs operating at half (or twice) the signal frequency usually have better AM rejection than receivers with fundamental VFOs, due to improved LO to RF isolation.

8.4 MIXERS FOR DIRECT CONVERSION RECEIVERS

The general properties of mixers are covered in a separate chapter, but the front-end of a direct conversion receiver is a unique application that puts some different demands on the mixer. To reduce LO radiation to an acceptable level, LO port to RF port isolation is needed. This usually requires a balanced mixer, but some other topologies are promising. The anti-parallel diode pair driven by a 1/2 frequency LO has been reported to work well, but has limited dynamic range and critical LO drive level requirements. Shunt FETs in switch mode have built-in LO to RF isolation. A number of experimenters have reported good success with different configurations of series FET switches using CMOS parts for several decades. The most common direct conversion mixers are Gilbert Cells like the NE602 and LM1496, and diode rings, both homebrew and commercial. Gilbert Cells have usually been used for low-cost-low-performance applications, but they should not be ruled out for higher performance receivers.

The important specifications for a direct conversion front-end mixer are noise figure (particularly 1/f noise figure when used with an audio IF), two-tone third-order dynamic range, 2nd order dynamic range, and LO to RF port isolation. Conversion gain or loss is less important, as it can be made up with gain elsewhere, and can not make up for poor noise figure.

Mixer recommendations

For the simplest direct conversion receivers, Gilbert Cells offer good performance at low current. The gain of a Gilbert Cell does not enhance receiver performance, since it occurs before any effective channel selectivity, but it does reduce the total receiver parts count. For some applications—carrying a rig into the mountains for a casual non-contest weekend backpacking trip, for example—the receiver is far less likely to fail from overload than from dead batteries. For such applications, "performance" takes on a different meaning, and a receiver that draws 5 mA outperforms one that draws 50 mA. For home station use or any kind of contest environment, a receiver with poor dynamic range can be as useless as one with dead batteries, and far more frustrating. For such applications, diode rings are recommended. For the designer builder, they have the advantage of a wealth of applications information and a published schematic.

Passive FET mixers in various configurations have dynamic range and noise advantages over both Gilbert Cells and diode rings. Considerably less has been published about passive FET mixers, although they are standard in cellular telephone handsets. This is an important area for amateur experimentation. Experiments are encouraged using both integrated quad analog switches and matched FETs on a single die in small multi-pin packages. Since the LO drive to a passive FET mixer goes to the high-impedance FET gate, little LO drive power is needed. The passive FET itself doesn't have a power supply. Thus passive FET mixers for direct conversion receivers offer the potential for the highest performance at the lowest operating current of any mixer type.

Direct Conversion Noise Figure

The noise figure of a direct conversion receiver mixer is generally different than the noise figure of the same mixer used in a superhet application, because of 1/f noise. Mixer noise figure does not have a neat and tidy definition, and mixer 1/f noise is even less well understood. Because of 1/f noise, diode ring mixers have noise figures in direct conversion receiver applications that range from within 1dB of their conversion loss to 15 or 20 dB worse. The increased noise figure is a result of excess noise at the IF port when the mixer is driven by the LO with the RF port terminated in a room temperature 50-Ω load. The noise spectrum is not necessarily a smooth 1/f curve, so merely observing the shape of the noise spectrum across a restricted audio passband is not enough to identify 1/f noise. Mixer noise figure is further complicated by the presence of noise on desired and image frequencies, noise in the bands around the harmonics of the LO, and the fact that the different contributions to mixer noise figure may be partially correlated. Rather than attempting to precisely define direct conversion mixer noise figure, this text will present a few measurements that provide some insight into noise in receiver systems, and will at least allow comparisons between different mixers and direct conversion receiver front-ends.

The first measurement is the noise figure of the audio amplifier itself. We have made this measurement with a hot-cold noise source. The audio amplifier is run at full gain in an environment with no hum or other noise pickup. The input to the audio amplifier is switched between two 50-Ω resistors, one at room temperature and the other at 77K. It is very important to measure the resistance of the cold resistor, to make sure it is still 50 Ω. Most resistors change value when the temperature drops that low. A series or parallel combination can be experimentally determined that provides a cold 50-Ω resistor. The output of the audio amplifier is connected to an averaging true RMS voltmeter reading in dB, and also a speaker or headphones. It is useful to listen while making the measurements, because the difference between hot and cold resistor noise can be heard in the headphones, and the measurements will be corrupted by any extraneous interference pickup, which can also be heard on the headphones. **Fig 8.20** gives noise figure as a function of the difference between the noise output from the hot and cold resistors in dB. The noise figure of the grounded base audio preamplifiers with diplexers in the receiver circuits in this text ranges from 5 to 7 dB.

The second step in the measurement process is to measure the conversion loss of the mixer. This can be done with a known RF signal at the RF port, a low-pass filter and 50-Ω termination on the IF port, and an RMS voltmeter across the 50-Ω resistor.

The last step in the measurement is to measure the excess IF noise when the mixer is connected to the audio amplifier and the LO is turned on. The input to the audio amplifier is switched between a room temperature resistor and the mixer, with LO drive and the RF port terminated in a room temperature 50-Ω load. At 14 MHz, a small sample of TUF-1 mixers produced between 1 and 6 dB more noise output than the 50-Ω room temperature termination. Two homebrew diode ring mixers using hand-wound toroids and 1N4184 diodes had less than 1-dB excess noise. A small sample of TUF-5 mixers operated at 1296 MHz and ADE-35 mixers at 2304 MHz had more than 10-dB excess noise. Special low-1/f noise diodes

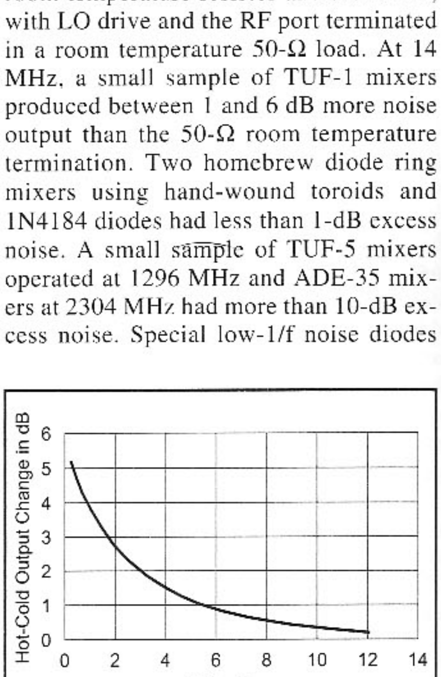


Fig 8.20—Hot-cold resistor noise figure differential.

are used in 10-GHz direct conversion receivers for Doppler Radar applications. This is a very small data set, and it is unwise to draw firm conclusions based on this limited information. More measurements are needed.

When the excess noise is low, a reasonable approximation to direct conversion receiver noise figure is just the baseband amplifier noise figure plus the mixer conversion loss. When excess mixer noise is present, the mixer loss and noise tend to dominate receiver noise figure, and baseband amplifier noise figure is less important. One experiment that may be done on the bench is to add attenuation between the mixer and baseband amplifier while observing receiver sensitivity. A 3-dB 50-Ω attenuator will drop the desired signals by about 3 dB, but it may also drop the receiver noise floor by about 3 dB, leaving the signal-to-noise ratio unchanged. Signals do not drop by precisely 3 dB, because the mixer impedance and the baseband amplifier input impedance are not exactly 50 Ω.

One way around the mixer excess noise uncertainty is to use a low-noise RF amplifier with enough gain to define the system noise figure. In this case it may be beneficial to include a resistive attenuator on the mixer output to optimize mixer dynamic range.

When used ahead of a DSB direct conversion receiver, a low-noise RF amplifier will have equal noise output on the desired and image bands. The image noise will reduce receiver output signal-to-noise ratio by 3 dB. Image noise may be suppressed by a narrow filter after the RF amplifier (practical for fixed-frequency applications), or by phasing, discussed in the following chapter.

Mixers with conversion gain, for example the Gilbert Cells used in LM1496 and NE602 integrated circuits, reduce the need for low-noise audio gain. The NE602 has low noise figure, which makes it attractive for simple receivers without RF amplification. The LM1496, biased for improved mixer linearity, is a better choice when an RF amplifier is used. In DSB direct conversion receiver applications with no provisions for suppressing image noise, each of these has the same 3-dB image noise penalty.

Based on these limited measurements and theory, a few guidelines for direct conversion receivers may be suggested. A homebrew diode ring with common 1N4148 silicon switching diodes, as used in Roy Lewallen's "Optimized QRP Transceiver"[17], with low-loss RF input circuitry and a grounded base audio amplifier, will provide an effective receiver noise figure around 10 dB, which is usually better than is needed at 7 MHz. Because the LO to RF isolation of homebrew mixers may not be as good as commercial packaged mixers using matched quads of Schottky diodes, the use of an RF amplifier ahead of the mixer is recommended. This will tend to negate any 1/f noise advantage of the homebrew switching diode mixer. In our HF designs, we tend to use small commercial packaged mixers, and about 10 dB of high reverse-isolation RF gain. This results in receivers that have noise figures in the 10-dB range, have very low LO radiation, and work well with common commercial packaged diode ring mixers. At VHF, we usually use about 20 dB of RF gain, and phasing to suppress image noise.

8.5 A MODULAR DIRECT CONVERSION RECEIVER

The "High Performance Direct Conversion Receiver" published in August 1992 *QST*[18] is a good benchmark. The ten-year-old design stands up well against more recent work, and the description is recommended reading. The circuitry presented here takes a slightly different approach, and takes advantage of a few improvements in our understanding during the past decade. A basic 40-m circuit is shown, but few changes are needed for operation on other bands.

The block diagram is shown in **Fig 8.21** and the schematic in **Fig 8.22**. The antenna is connected to a grounded-gate FET RF low-noise amplifier. The mixer is a Mini-Circuits TUF-3, with an audio diplexer and low-noise headphone amplifier. The VXO circuit provides clean sine-wave +7 dBm drive to the mixer. For speaker output, a battery powered external speaker from RadioShack, or an amplified computer speaker is recommended.

RF Low-Noise Amplifier

The receiver gain distribution was designed for approximately 10-dB of RF gain ahead of the mixer. RF gain ahead of a diode ring mixer is not normally needed for receiver sensitivity below 10 MHz, but in a direct conversion application, there are other benefits to using an RF preamp. First, with RF gain up front, there is less need to design for low loss through the mixer IF termination and diplexer network. This permits the baseband circuitry to be optimized for selectivity and proper termination of both the mixer and diplexer network. Second, a grounded-gate FET amplifier typically has over 40 dB of reverse isolation, which adds directly to the LO to RF isolation of the mixer, and helps reduce the amount of LO radiation from the antenna. Third, with a buffer amplifier between the antenna connection and the mixer RF port, the mixer environment does not change when the antenna moves in the breeze. Fourth, Direct Conversion Receivers need good low-pass filters on the inputs, and the low-pass matching networks in and out of the FET provide all the attenuation needed. Finally, the simple mute switch turns the RF low-noise-amplifier into a strong 40-dB attenuator, which prevents any strong signals (for example from a companion transmitter) from arriving at the mixer diodes.

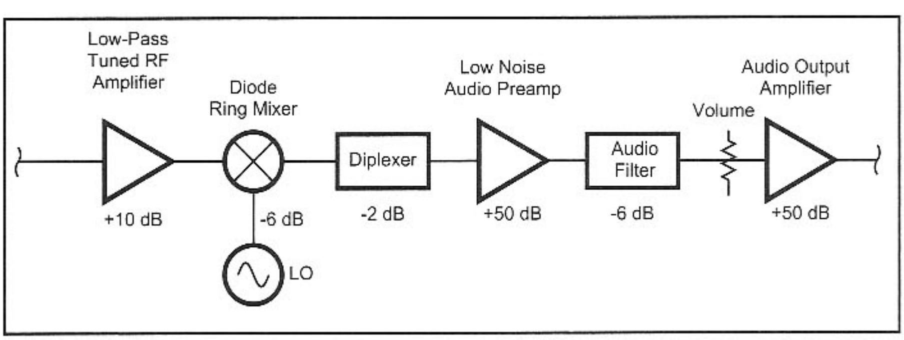


Fig 8.21—A modular receiver block diagram

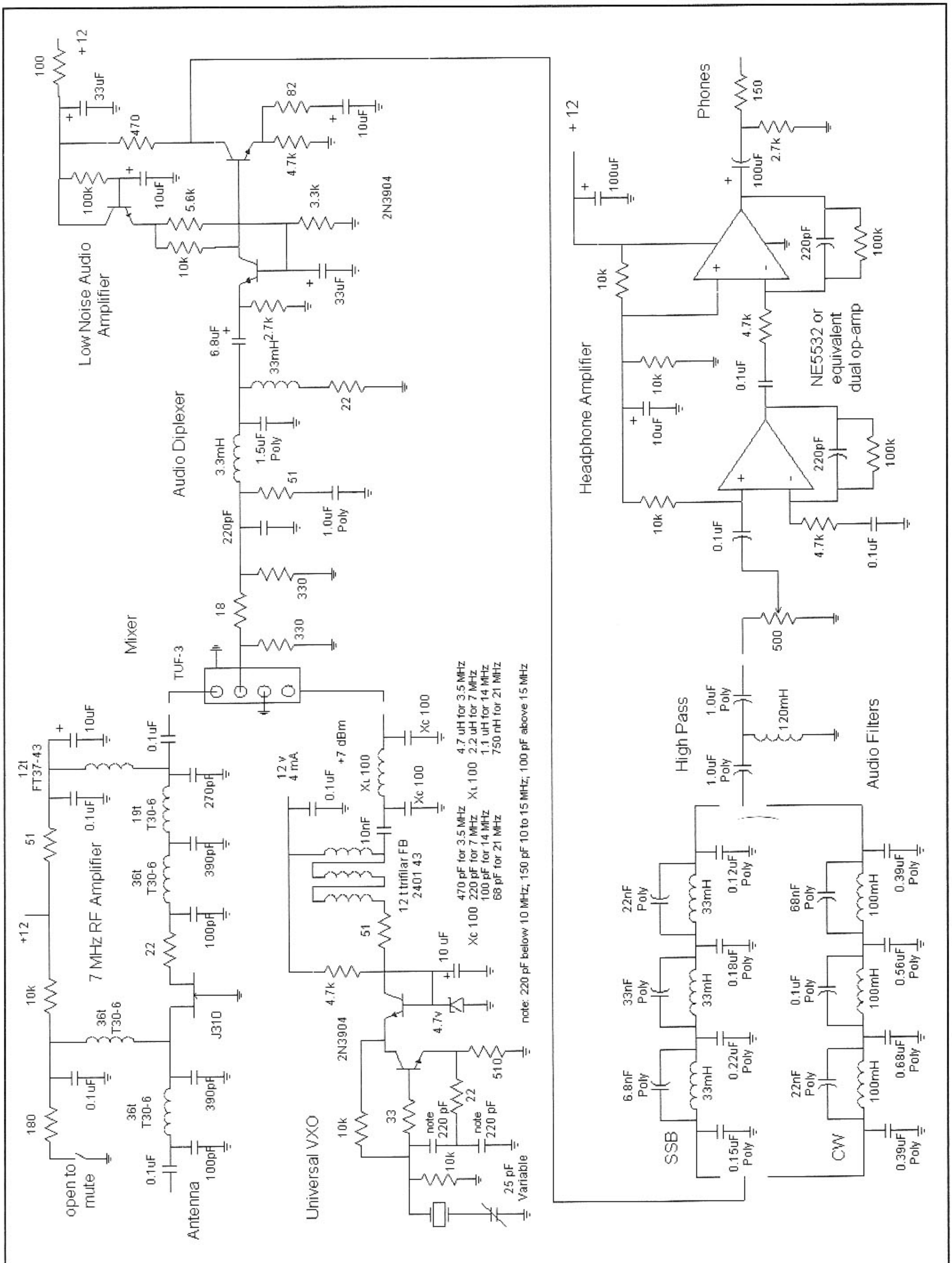


Fig 8.22—Modular receiver schematic.

Audio Diplexer

The diplexer network is designed to provide good selectivity before any wideband audio gain. This greatly improves the receiver close-in dynamic range, and permits the use of a grounded-base audio LNA operating at low current (0.5 mA) to set the impedance to 50 Ω. This audio diplexer is a little more selective than the ones described in the phasing chapter, because there is no need to precisely match amplitude and phase between two channels.

Audio low-noise amplifier

There are many audio low-noise amplifiers that will work in direct conversion receivers, but this one works well and has been widely duplicated for several decades. It has no flaws that impair performance in this application, so the design effort was focused elsewhere.

Filters

Passive audio filters work well, draw no current, and use inexpensive components available from several sources. The SSB and CW bandwidth filters shown are old favorites.

Headphone amplifier

The headphone amplifier provides audio gain to boost the signals from the low-levels in the signal processing components up to comfortable listening volume. This is a standard design, very similar to the headphone amplifier used in the Binaural Receiver[19] published in March '99 *QST*.

VXO

The VXO circuit is another old favorite, evolved over many years from a circuit published by Joe Reisert[20] as a frequency standard. There are a number of subtleties, including stiff regulation of the voltage on all three terminals of the oscillator transistor and the use of a Zener diode operated in the 4.7-V zero-temperature-coefficient sweet-spot. This VXO circuit tunes over about 5 kHz at 7 MHz, provides +7 dBm output and drifts a few Hz at turn-on.

Construction

The receiver was built on separate pieces of unetched copper-clad circuit board. The RF amplifier is on one piece, the VXO on a second piece, and the mixer and audio amplifier on a third piece. The audio filters are on separate pieces. There are a number of reasons for building the receiver on separate boards. The first is entirely practical—each piece is an evening project than can be built and tested as a stand-alone module. The second consideration is equally important: the RF amplifier is good for only one band; the VXO can be easily modified for different HF frequencies; and the mixer-audio board can be used on any frequency from 50 kHz through 250 MHz. By making the pieces separate, any of them may be replaced to put the receiver on a different frequency, or borrowed for a different project.

Generally speaking, receiver circuits built prototype-style on separate pieces of unetched copper-clad circuit board work better than PC board circuits. This is because the unetched copper-clad board permits both the short ground leads required by RF circuitry and the single-point grounding required by low-frequency high gain amplifiers. Receivers that must be mass-produced using PC boards often require many PC layout revisions to overcome the problems that arise when the prototype circuits are transferred to PC board construction.

The more components a receiver module has, the more practical it is to spend time developing a PC board design. For simple circuitry like the modules presented here, it is often more practical to use prototype construction, and avoid the headaches associated with PC board ground faults.

Applications

The modular high-performance direct conversion receiver presented here works equally well connected to an antenna, or as part of a superhet receiver. The well-defined near 50-Ω input impedance to the RF preamp provides a good termination for simple crystal filters, and the VXO circuit is a good BFO with enough tuning range to cover both sidebands.

8.6 DC RECEIVER ADVANTAGES

For much of their history, direct conversion receivers have been viewed as an adequate, simple substitute for more serious receivers. It is time to redefine direct conversion as an alternative architecture that poses a unique set of problems, but also offers significant advantages. Some of the important advantages are:

1. Simplicity
2. Few spurious responses
3. High spurious-free dynamic range
4. Very low distortion of the desired signal
5. Frequency range independence
6. Compatibility with DSP-based receiver architectures
7. Compatibility with adaptive receivers and antennas

Simplicity is best illustrated by the circuit in Fig 8.7. Build it ugly style in a few hours the Thursday evening before Field Day or the November CW Sweepstakes, string up a temporary 40-m dipole Friday evening, and spend a few hours over the weekend listening. Simplicity is appealing. Much of this text is devoted to pushing the performance envelope for designer built radio equipment. Spending two years building a receiver system that offers an incremental performance improvement that must be measured to be perceived is an interesting activity, but with a serious flaw. Suppose the number to be exceeded is the magic "100-dB SSB Bandwidth Two-Tone Third Order Dynamic Range." Magic to whom? Certainly not my teen-age daughter! But she will spend a few minutes politely listening to CW on headphones connected to a hand full of parts with a 9-V battery and some wires going out into the trees in the back yard—and when I ask her if she hears the kind of weak, warbly one and she say yes and then I tell her he's in St. Petersburg, Russia—her eyes light up. Now that's magic!

Superhets for SSB and CW have images, higher order undesired responses, and internally generated birdies. A direct conversion receiver with a low-pass filter between the antenna and mixer hears only signals within a few kHz of the LO. Period.

It is theoretically possible to design superhet receivers for arbitrarily good image and IF rejection, but in practice superhets must be designed to place images and IFs in parts of the spectrum with few strong signals. When image signals are 90-dB stronger than the desired signal, they will find a way into the receiver and cause problems. This severely constrains the choices of IF for frequency bands in heavily used portions of the spectrum. For example, what IF should be used

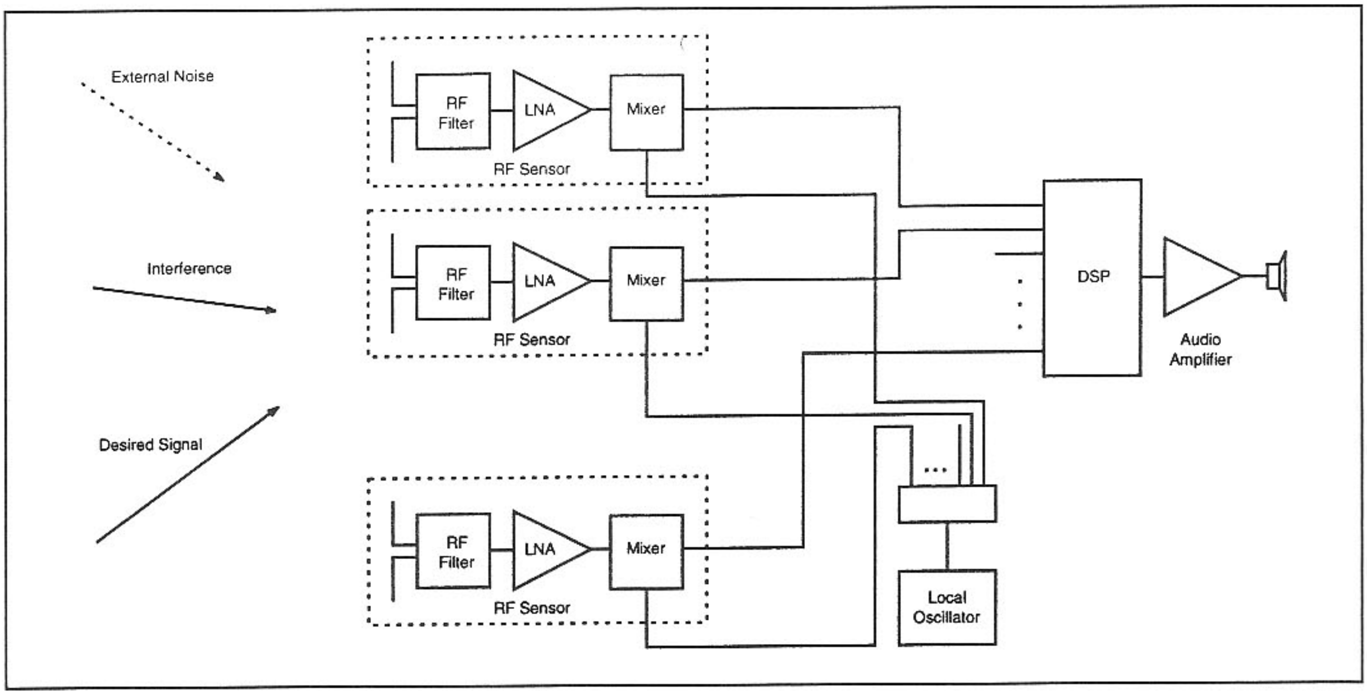


Fig 8.23—Advanced receiver architecture.

for a 144 to 148 MHz receiver? The industry standard IFs at 455 kHz, 10.7 MHz, and 21.4 MHz provide a selection of off-the-shelf filters. 455 kHz is too low for adequate image rejection. 10.7 is useful, but a bit low for providing good image rejection across a 4-MHz wide frequency range without retuning the RF amplifier. 21.4 MHz is attractive, except that with low-side injection, the image falls in the FM broadcast band, and with high side injection, the image is in TV channel 12. Direct conversion offers a technique for tuning across a wide frequency range and recovering 10-nV signals surrounded by 10-mV interfering signals. That is 120 dB of spurious-free dynamic range.

Because direct conversion receivers have only one frequency conversion stage, and it operates before significant receiver gain, mixer distortion does not significantly contribute to in-band intermodulation. The quality of the recovered audio is almost entirely determined by the distortion properties of the audio amplifier chain. Since audio engineers have spent decades reducing the distortion of high-gain audio amplifiers, simply following a diode-ring mixer with a low-noise preamplifier and high-fidelity audio amplifier will produce a receiver with significantly lower in-channel distortion than any commercial superhet. Audio engineers have also developed low-distortion gain control and gain compression techniques that operate strictly at audio, and that "audio AGC" technology is beginning to appear in amateur equipment.

The same block diagram works for direct conversion receivers whether the frequency of interest is 24 kHz or 24 GHz. A superhet designer will draw completely different block diagrams for a SSB receiver for those two frequencies. Furthermore, superhet frequency conversion plans must be designed with an understanding of the levels of all the potential sources of image, higher-order spurious responses, and birdies. A receiver optimized for 10 MHz might have a completely different frequency conversion plan than one optimized specifically for 14 MHz. For the amateur interested in a the entire spectrum, the lessons learned and the time spent optimizing a 10-MHz direct conversion receiver apply just as well to a 2.4-GHz satellite receiver.

As DSP systems improve and become more widely used and understood, it becomes less and less attractive to compromise the signal with multiple frequency conversion, AGC, and crystal filter delay and ripple before it enters the DSP. Direct Conversion offers a way to simply translate a desired radio signal to the frequency range needed by the A to D converters ahead of a DSP engine (**Fig 8.23**). Soft-Radio advocates call this *Direct Sampling* and claim that there is no conventional radio at all—the computer is connected straight to the antenna. Such claims obscure the truth. Direct Sampling is just a different and convenient name for entirely conventional I and Q mixing, in the same sense that the term "wireless" allows people who have no understanding of radio to claim the title Wireless Expert. Such good natured competition between traditional radio designers and digital signal processing artists is a natural part of the evolution. Both camps need to realize that receivers of the future will use both skill sets. There is magic in simple radio circuits, but there is also magic in watching a signal below the noise level appear in a waterfall plot on a computer monitor.

Finally, in it's second hundred years, radio will experience significant changes. For six decades the usual way to collect and process HF and VHF signals has been a Yagi-Uda antenna with a single feed line connected to the back of a complex superhet receiver. Space diversity and adaptive antenna interference cancellation have been impractical because of the amount of hardware required and severe amplitude and phase matching constraints. The hardware problem is solved if each dipole antenna element has its own direct conversion down-converter, all of them driven by a single LO, and each connected to a separate input port of a computer sound card. The actual hardware is very simple, and with more than two dipoles, image-reject techniques can be combined with noise cancellation in the arrival-angle domain and adaptive CW interference cancellation in the frequency domain to produce an output signal-to-noise and interference ratio far better than the best conventional, single feed line system.

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CHAPTER 9

Phasing Receivers and Transmitters

9.1 BLOCK DIAGRAMS

The phasing method of single-sideband generation and reception has been discussed in the literature and incorporated in commercial products for over 50 years. The phasing method fell into disuse in amateur products from the late '60s through the '80s due to the popularity of transceivers built around a single IF crystal filter used for both sideband generation and receive selectivity. During this period, prices of old phasing transmitters dropped until they were only used on the air in modest stations scraped together on a budget, often by folks with no appreciation of the art of maintaining vintage radio gear. Sociology being what it is and amateurs being human, phasing transmitters were soon associated with poor signals, and their unfortunate operators were encouraged to upgrade or get off the air. Even scholarly authors during this period often used a little over-simplified mathematics to show that the phasing method was incapable of generating acceptable signals for the modern amateur bands.

How times have changed. During the '90s the vintage radio craze hit the amateur bands, and amateurs across the US began hearing signals from old Central Electronics transmitters, carefully restored, properly aligned, and conservatively operated. By comparison, the modern transceivers sounded thin and distorted. Modern radios have had to scramble to recapture the lost sound quality of the old rigs. Sociology still being what it is, there is now a market for low-distortion transmitters, and one amateur manufacturer has even introduced a full-sized transceiver with a Class A power amplifier. The lore has changed, and phasing transmitters and receivers now have the reputation for sounding better than conventional systems that use filters for opposite sideband suppression. As usual, careful study reveals that there is an element of truth in conventional wisdom, but that deeper understanding provides freedom from the bonds of lore.

Fig 9.1 is the block diagram of a conventional SSB exciter using a filter to remove the unwanted sideband. Since the filter passband frequency is fixed, the resulting SSB signal must be heterodyned to the desired final output frequency. Since it is difficult to build SSB bandwidth filters for frequencies above 50 MHz, there may need to be multiple frequency conversions to reach a microwave frequency. **Fig 9.2** is the block diagram of a phasing SSB exciter. The signal frequency networks all have considerable bandwidth, so operating the SSB modulator on the final output frequency is an option. Heterodyning the phasing exciter output to the desired output frequency also has merit, and was the method of choice in vintage gear. **Fig 9.3** shows a conventional superhet receiver with a SSB bandwidth IF filter to provide rejection of interference outside the desired bandpass, including rejection of the opposite sideband. **Fig 9.4** shows a superhet receiver with a phasing SSB demodulator at the IF. Note that the phasing system just rejects the opposite sideband—conventional selectivity is still

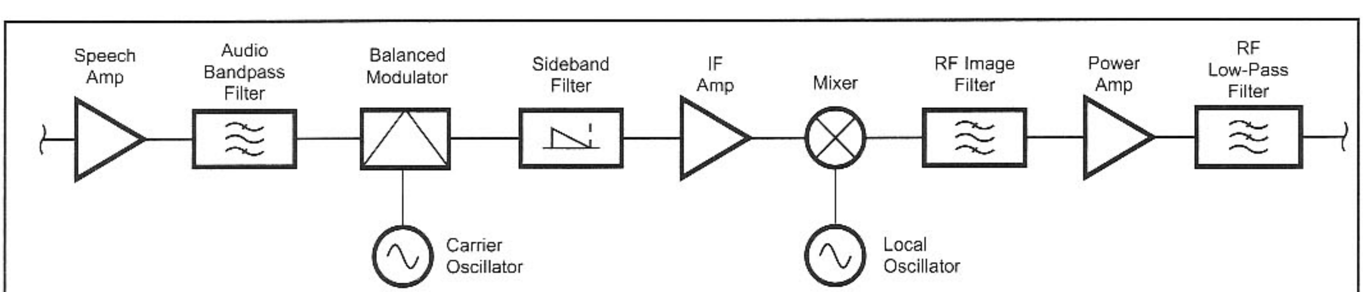

Fig 9.1—A block diagram of a conventional SSB exciter using a filter to remove the unwanted sideband.

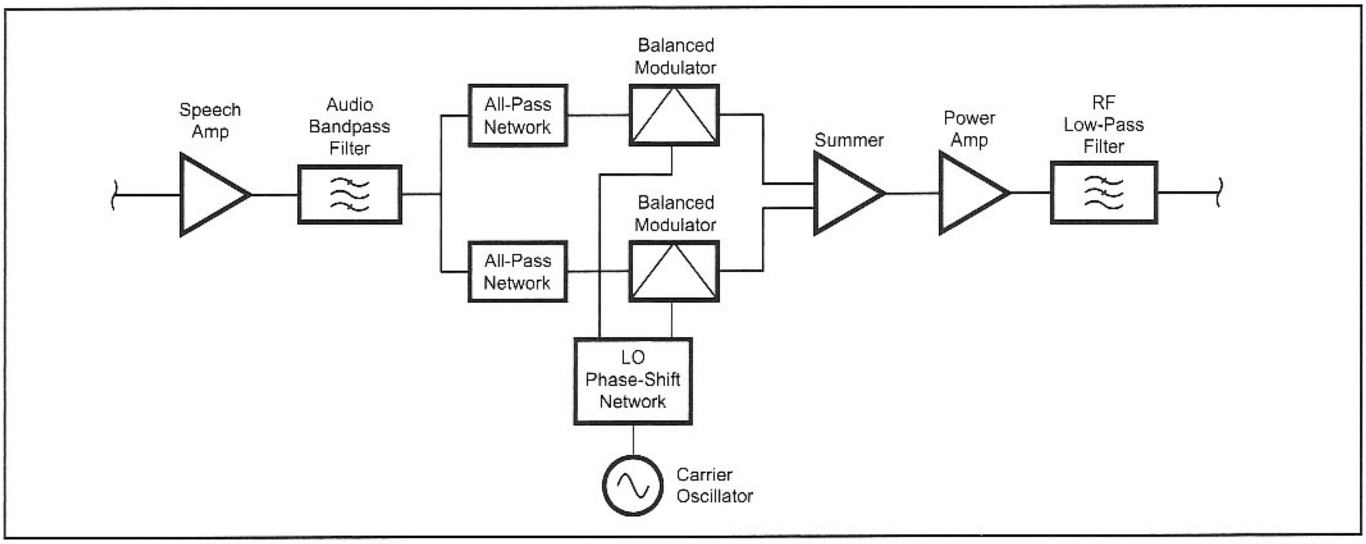

Fig 9.2—Block diagram of a phasing SSB exciter.

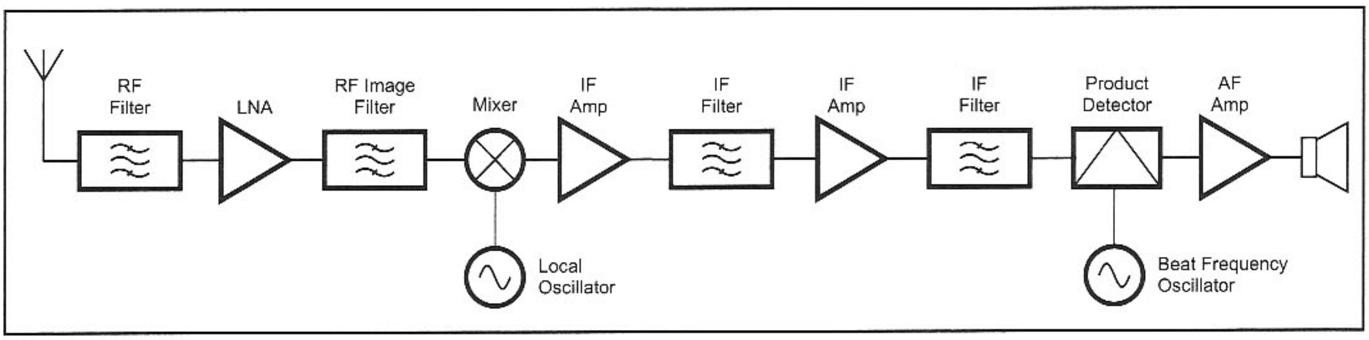

Fig 9.3—A conventional superhet receiver with a SSB bandwidth IF.

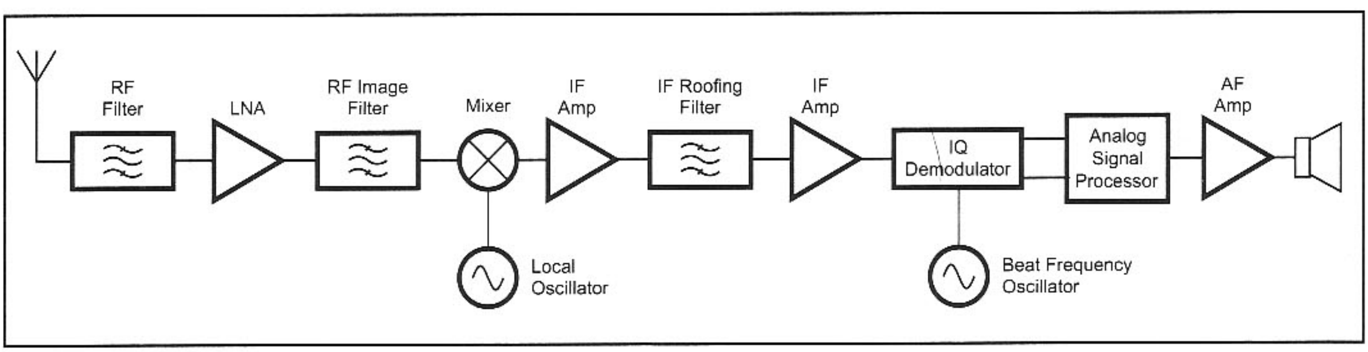

Fig 9.4—A superhet receiver with a phasing SSB demodulator at the IF.

needed to protect the receiver from interference at other frequencies. **Fig 9.5** is the block diagram of a phasing direct conversion receiver (high performance direct conversion receiver techniques are discussed in Chapter 8 of this book.) Phasing is added in **Fig 9.6** with baseband processing functions handled using a pair of analog-to-digital converters and a digital signal processor. Each of the systems shown in the block diagrams is optimum for certain applications, and a designer-builder needs to be familiar with the benefits and limitations of each before concluding that a particular radio architecture is best for a particular application.

Traditionally, phasing is presented as a transmit topic, with receivers tacked on as an "oh by the way, you can also..." This is fine until one wants to actually begin designing and building a receiver using phasing methods, at which point none of the math really makes sense, and signal levels, noise, and distortion terms that don't apply to transmitters become important. The treatment here will take the opposite tack, and discuss phasing direct conversion receivers in detail. There are several justifications for this. The first is that exploration of high performance phasing direct conversion receivers has been a major focus area for the author for over a decade, and many of the observations, much of the analysis, and the mathematical treatment have not been previously published—or at least not for a very

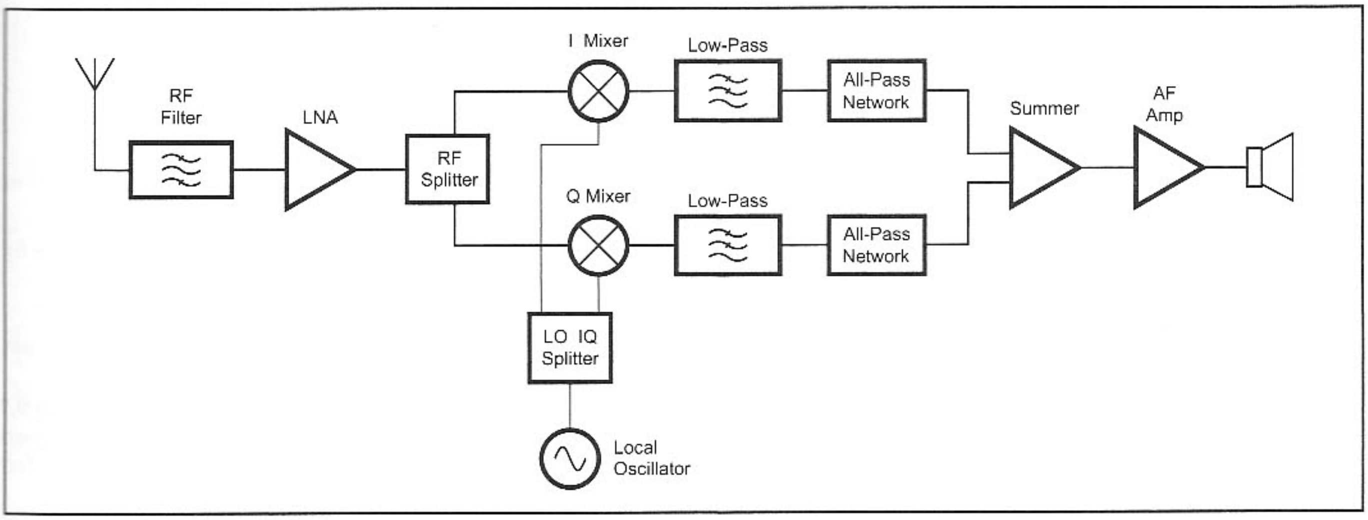

Fig 9.5—A block diagram of a phasing direct conversion receiver.

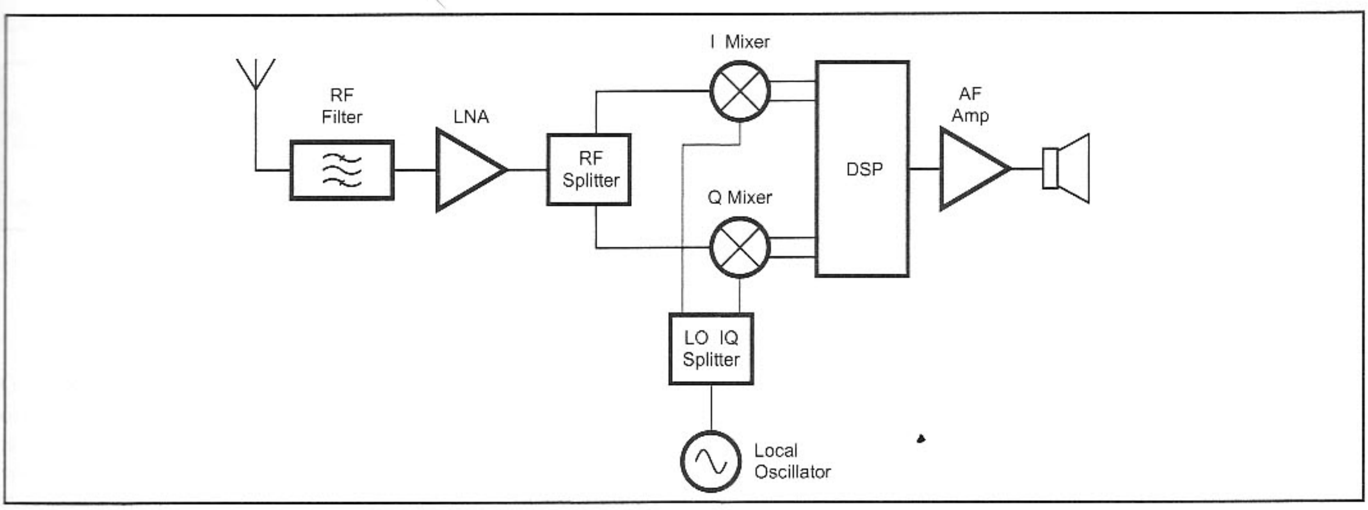

Fig 9.6—High performance direct conversion receiver technique with phasing added—with baseband processing in DSP.

long time. The second is that most of this decade of study has been a purely amateur activity, pursued because listening to that first phasing direct conversion receiver ten years ago was such a profound revelation. Phasing direct conversion receivers are an optimum choice for many applications, amateur and professional, whenever cost, distortion, spurious-free-dynamic range, frequency agility or adaptability to different bandwidths and modulation types are important. Furthermore, they are a rich field for experimentation and contribution to the amateur and professional literature. Finally, by describing the receiver mathematically using a general band-limited input signal $a_s(t)\cos[2\pi f_s t + \phi_s(t)]$, the discussion becomes independent of modulation type, and serious students of communications systems will have no difficulty converting to complex-envelope form, adding correlated and uncorrelated noise terms, and including the effects of various types of distortion.

The emphasis will be on direct conversion phasing receivers, rather than superhet receivers with phasing last-converters, because the direct conversion receiver generally presents a more difficult set of problems. However, it should be mentioned at this point that the ultimate receiver for weak CW and SSB signals in the presence of noise and strong-signal interference is most likely a hybrid superhet that includes a band-limiting filter followed by some IF gain and then a phasing product detector. This is certainly the approach being taken by makers of high-end amateur transceivers, and the technology will trickle down into the low end of the market, as it is less expensive than relying solely on mechanical, quartz crystal and ceramic filters for selectivity. The major difference between using the phasing system at the front-end of a direct conversion receiver or as the product detector for a hybrid superhet is in the gain, selectivity, and noise distributions in the receiver. These considerations will be discussed in detail in the R2pro design exercise.

9.2 INTRODUCTION TO THE MATH

Some mathematics is necessary for understanding how phasing receivers work. Fortunately, all of the necessary functions and identities may be found in a high school algebra and trigonometry textbook. That said, there is nothing trivial about the treatment that follows. It is deliberate and complete. It is also much less interesting than the pictures and schematics of the projects, and many of the subtleties were not appreciated by the author until long after the first signals began pouring out of the working receiver's speaker. Readers with an aversion to math in any form are invited to skip this section. Designer-builders who want to proceed directly to the R2pro design and projects section are encouraged to skim quickly through the math. Electrical Engineering graduate students should work slowly through the material step by step, because this stuff will be on the exam. Refer to Figs **9.7A-G** that appear after the equations.

The Basic Image-Reject Math From Receiver Point Of View

Any band-limited basic signal may be described as:

$$s_i(t) = a_s(t)\cos\left[2\pi f_s t + \varphi_s(t)\right] \qquad \textbf{Eq 9.1}$$

...where f_s is the signal frequency; $a_s(t)$ is the time-varying signal envelope; and $\phi_s(t)$ is the time-varying signal phase

Mixer

In an ideal mixer, a local oscillator multiplies this signal:

$$L_i(t) = 2\cos\left(2\pi f_o t + \varphi_o\right) \qquad \textbf{Eq 9.2}$$

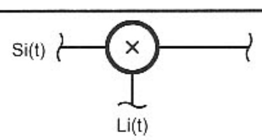

Fig 9.7A

...where the constant 2 simplifies later expressions, f_o is the LO frequency, and ϕ_o is the LO phase.

Multiplying the LO times the signal:

$$2\cos\left(2\pi f_o t + \varphi_o\right) a_s(t)\cos\left[2\pi f_s t + \varphi_s(t)\right]$$
$$= a_s(t)\cos\left[2\pi\left(f_o + f_s\right)t + \varphi_s(t) + \varphi_o\right]$$
$$+ a_s(t)\cos\left[2\pi\left(f_o - f_s\right)t - \varphi_s(t) + \varphi_o\right]$$

Eq 9.3

If the signal frequency f_s is lower than the LO frequency f_o, then the difference expression $(f_o - f_s)$ is a positive number.

Low-Pass Filter

In a receive downconverter application, the difference frequency expression is selected by a low-pass filter following the mixer, and the sum frequency $(f_o + f_s)$ expression is rejected. The downconverter output frequency range may extend from zero Hz up to the cutoff frequency of the low-pass filter, and this frequency range is referred to as "baseband." The baseband output is then:

$$b_i(t)$$
$$= a_s(t)\cos\left[2\pi\left(f_o - f_s\right)t - \varphi_s(t) + \varphi_o\right]$$

Eq 9.4

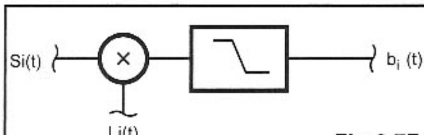

Fig 9.7B

Q Channel

In a phasing system, a second mixer multiplies the identical signal by a LO with $\pi/2$ phase delay. The two mixers and their signals are referred to as I for "in-phase" and Q for "quadrature." Since these expressions represent real signals and electronic components, they are not perfect. In particular, the amplitude of the signal at the Q mixer may not be identical to the I mixer amplitude, and the phase difference between the I and Q mixers may not be exactly $\pi/2$. We can incorporate these differences by introducing error terms into the signal and LO expressions.

$$s_q(t) = (1+\varepsilon)a_s(t)\cos\left[2\pi f_s t + \varphi_s(t)\right]$$

Eq 9.5

and

$$L_q(t) = 2\cos\left(2\pi f_o t + \varphi_o - \pi/2 + \delta\right)$$

Eq 9.6

...where ε is the amplitude difference between the I and Q signals and δ is the error in the $\pi/2$ phase delay. Note that the signal $s_q(t)$ at the input to the Q mixer is the same as $s_i(t)$ at the input to the I mixer except for the error ε.

Multiplying the phase-shifted LO and signal together in the Q mixer:

$$2\cos\left(2\pi f_o t + \varphi_o - \pi/2 + \delta\right)(1+\varepsilon)$$
$$a_s(t)\cos\left[2\pi f_s t + \varphi_s(t)\right]$$
$$= (1+\varepsilon)a_s(t)\cos$$
$$\left[2\pi\left(f_o + f_s\right)t + \varphi_s(t) + \varphi_o - \pi/2 + \delta\right]$$
$$+ (1+\varepsilon)a_s(t)\cos$$
$$\left[2\pi\left(f_o - f_s\right)t - \varphi_s(t) + \varphi_o - \pi/2 + \delta\right]$$

Eq 9.7

Once again, the low-pass filter rejects the sum frequency and passes the difference frequency, so we are left with:

$$b_q(t) = (1+\varepsilon)a_s(t)\cos$$
$$\left[2\pi\left(f_o - f_s\right)t - \varphi_s(t) + \varphi_o - \pi/2 + \delta\right]$$

Eq 9.8

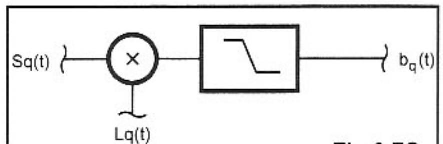

Fig 9.7C

Audio Phase-Shift Networks

In an image-reject receiver, the I and Q outputs of the mixers are then applied to the ports of a pair of all-pass networks that add an additional $\pi/2$ phase delay to the signal at the output of the Q mixer. An ideal all-pass network would introduce no additional amplitude or phase errors, but such errors occur in practice. In addition, the all-pass networks at baseband may have many octaves of bandwidth, and the amplitude and phase errors will vary across the baseband frequency range. We combine all of the amplitude errors into a single baseband frequency dependent error term $\varepsilon(f)$ and all of the phase errors into a single baseband frequency dependent phase error term $\delta(f)$. We also recognize that in practice the IQ all-pass network pair does not simply leave the I channel alone and add a constant $\pi/2$ phase delay to the Q channel, but introduces a frequency dependent phase shift to each channel, chosen so that the phase difference between the I and Q channels remains a (nearly) constant $\pi/2$. We combine this frequency dependent phase shift with the original LO phase ϕ_o and denote the result $\phi_o(f)$. With the additional $\pi/2$ phase delay and all of the modified error and phase terms, the all-pass network Q baseband output becomes:

$$[1+\varepsilon(f)]a_s(t)\cos\begin{bmatrix}2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)\\ -\pi/2-\pi/2+\delta(f)\end{bmatrix}$$
$$=[1+\varepsilon(f)]a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)-\pi+\delta(f)]$$

Eq 9.9

The I baseband output at the output of the all-pass network is:

$$b'_i(t)= a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$

Eq 9.10

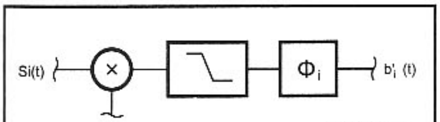

Fig 9.7D

Returning to the baseband Q term, use the trig identity:

$\cos(a-\pi) = -\cos a$ **Eq 9.11**

to obtain

$$b'_q(t)=-[1+\varepsilon(f)]a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)+\delta(f)]$$

Eq 9.12

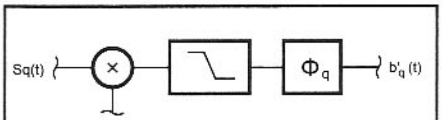

Fig 9.7E

Suppressing the Image

The baseband I and Q all-pass filter outputs are added to implement the image-reject function. To make the addition easier, the Q output may be broken down into separate terms:

$$-a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)+\delta(f)]$$
$$-\varepsilon(f)a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)+\delta(f)]$$

Eq 9.13

We may also separate the phase error $\delta(f)$ out using the trig identity:

$\cos(a+b) = \cos a \cos b - \sin a \sin b$

Eq 9.14

...where $2\pi(f_o - f_s)t - \phi_s(t) + \phi_o(f) = a$

and $\delta(f) = b$

$$b'_q(t)=-a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]\cos[\delta(f)]$$
$$+a_s(t)\sin[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]\sin[\delta(f)]$$
$$-\varepsilon(f)a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]\cos[\delta(f)]$$
$$+\varepsilon(f)a_s(t)\sin[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]\sin[\delta(f)]$$

Eq 9.15

At this point, it is convenient to make our first approximations. For phasing systems with opposite sideband suppression of more than 30 dB, the amplitude and phase error terms ε and δ must both be less than 0.1. Setting $\delta(f)$ to a maximum value of 0.1 and plugging it into the sine and cosine expressions:

$\sin(0.1000) = 0.0998$
$\cos(0.1000) = 0.9950$

...we may then use the "small angle" approximations:

$\sin\varphi \approx \varphi$ **Eq 9.16**

$\cos\varphi \approx 1$ **Eq 9.17**

knowing that the approximation errors are very small in the range of interest. The approximation errors become vanishingly small when we reduce δ still further, to the limits needed for high performance systems.

Using the small angle approximations, the four Q terms at the output of the baseband all-pass network become:

$$b'_q(t)=-a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$
$$+\delta(f)a_s(t)\sin[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$
$$-\varepsilon(f)a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$
$$+\delta(f)\varepsilon(f)a_s(t)\sin[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$

Eq 9.18

We now use a second approximation. If ε and δ are less than 0.1, then their product must be less than $(0.1)^2 = 0.01$. Thus the last term above is always much less than the other three terms. Once again, the approximation error becomes vanishingly small for high performance systems. Discarding the last term, the Q signal at the output of the baseband all-pass network is:

$$b'_q(t)=-a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$
$$+\delta(f)a_s(t)\sin[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$
$$-\varepsilon(f)a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$

Eq 9.19

This signal is added to the I signal at the output of the baseband all-pass network:

$$b'_i(t)=a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$

to obtain:

$$\text{LSB out}=+\delta(f)a_s(t)\sin[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$
$$-\varepsilon(f)a_s(t)\cos[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o(f)]$$

Eq 9.20

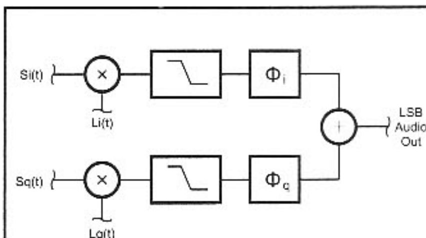

Fig 9.7F

Note that the equal and opposite signal components have added to zero, and only the error terms remain. Also note that the effect of a 0.1-radian phase error is identical to the effect of a 0.1 amplitude error. Finally, note that the two error terms are orthogonal (one is a sine, and the other is a cosine), so that each must be independently reduced to zero—an amplitude error will not cancel a phase error. The two error voltages add to make a resultant error signal, with magnitude:

$$\left\{[\delta(f)]^2+[\varepsilon(f)]^2\right\}^{1/2} \qquad \textbf{Eq 9.21}$$

Recovering the Desired Signal

Now examine the case of a signal frequency f_s greater than the LO frequency f_o.

The expression $(f_o - f_s)$ is now a negative number. The I baseband signal at the output of the I mixer is (as before):

$$b_i(t)=a_s(t)\cos\left[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o\right]$$

...to make the frequency term $(f_o - f_s)$ positive, use:

$(f_o - f_s) = -(f_s - f_o)$ to obtain

$$a_s(t)\cos\left[-2\pi(f_s-f_o)t-\varphi_s(t)+\varphi_o\right] \qquad \textbf{Eq 9.22}$$

Using the trig identity:

$$\cos a = \cos(-a) \qquad \textbf{Eq 9.23}$$

...we obtain the I mixer baseband output:

$$b_i''(t)=a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)-\varphi_o\right] \qquad \textbf{Eq 9.24}$$

The Q mixer baseband output is (as before):

$$b_q(t)=(1+\varepsilon)a_s(t)\cos\left[2\pi(f_o-f_s)t-\varphi_s(t)+\varphi_o-\pi/2+\delta\right]$$

Again using the $(f_o - f_s) = -(f_s - f_o)$ substitution and $\cos a = \cos(-a)$ identity, the Q mixer baseband output is:

$$(1+\varepsilon)a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)-\varphi_o+\pi/2-\delta\right] \qquad \textbf{Eq 9.25}$$

At the output of the all-pass network, which adds $\pi/2$ phase delay and additional errors, the Q signal is:

$$b_q''(t)=(1+\varepsilon(f))a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)-\varphi_o(f)-\delta'(f)\right] \qquad \textbf{Eq 9.26}$$

Note that the combined phase error term $\delta'(f)$ is different than the previous case, because of the sign change on δ. Also note that the minus $\pi/2$ phase shift from the all-pass network has cancelled the plus $\pi/2$ phase shift from the LO.

Performing the same steps as before to reduce the signal at the Q baseband all-pass network output to separate components, we obtain:

$$\begin{aligned}&a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+\delta'(f)a_s(t)\sin\\&\quad\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+\varepsilon(f)a_s(t)\cos\\&\quad\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\end{aligned} \qquad \textbf{Eq 9.27}$$

Adding the I and Q outputs from the baseband all-pass network:

$$\begin{aligned}&a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+\delta'(f)a_s(t)\sin\\&\quad\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+\varepsilon(f)a_s(t)\cos\\&\quad\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&=2a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+\delta'(f)a_s(t)\sin\\&\quad\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\\&+\varepsilon(f)a_s(t)\cos\\&\quad\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right]\end{aligned} \qquad \textbf{Eq 9.28}$$

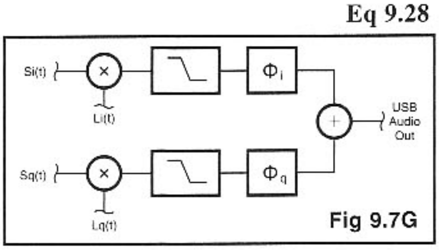

Fig 9.7G

...since $\delta'(f)$ and $\varepsilon(f)$ are both much less than 2, a reasonable approximation for the sum of the I and Q all-pass network outputs for an input signal with a frequency higher than the LO frequency is:

$$\text{USB out}=2a_s(t)\cos\left[2\pi(f_s-f_o)t+\varphi_s(t)+\varphi_o(f)\right] \qquad \textbf{Eq 9.29}$$

Once again, the accuracy of this expression becomes increasingly good as the amplitude and phase errors are reduced.

Sideband Suppression Expressions

In summary, it has been shown that signals at frequencies above the Local Oscillator frequency are downconverted and add at the output of the baseband all-pass network, while signals at frequencies below the Local Oscillator frequency are downconverted and subtract, leaving only the amplitude and phase error terms. It is a straightforward exercise, using the identical steps, to show that reversing the sign of either term, interchanging the LO phase shifts, interchanging the input ports of the all-pass network, or subtracting instead of adding the I and Q signals at the all-pass network output will result in adding the lower frequencies and canceling the higher frequencies.

Since the relative magnitude of the added signal is 2 and the magnitude of the error terms is:

$$\left\{[\delta(f)]^2+[\varepsilon(f)]^2\right\}^{1/2}$$

...the familiar expression for opposite sideband suppression in dB for a given set of amplitude and phase errors is easily obtained:

Opposite sideband suppression in dB

$$=20\log 1/2\left\{[\delta(f)]^2+[\varepsilon(f)]^2\right\}^{1/2} \qquad \textbf{Eq 9.30}$$

For the effect of just an amplitude or phase error, the simpler expressions

$20 \log \varepsilon/2$ (just amplitude error) **Eq 9.31**

$20 \log \delta/2$ (just phase error) **Eq 9.32**

...may be used. The more complete expression above describes the opposite sideband suppression as a function of baseband frequency f for the case where the lower sideband is suppressed. These expressions may be used to obtain the common textbook plot of sideband suppression versus phase and amplitude errors. Plugging in a few numbers: if both the amplitude and phase have the maximum error of 0.1, the opposite sideband suppression is:

$$\left\{20\log\left[(0.1)^2+(0.1)^2\right]^{1/2}/2\right\}=-23\text{ dB} \qquad \textbf{Eq 9.33}$$

Most textbooks quote amplitude and

phase errors in dB and degrees. To convert amplitude error ε to dB, use

$$20 \log [1 + \varepsilon] \qquad \textbf{Eq 9.34}$$

...for $\varepsilon = 0.1$ in the example above, the amplitude error in dB is 20 log (1.1) = 0.83 dB.

To convert phase error in radians δ to error in degrees, multiply δ by 57.3 (degrees per radian). For the example above, the phase error in degrees is 5.73 degrees.

As an example going the opposite direction, suppose a phasing receiver system has 1-degree maximum phase error and 0.1 dB maximum amplitude error. What is the opposite sideband suppression? Converting 0.1 dB to ε:

$$\varepsilon = 10^{[(\text{error in dB})/20]} - 1 \qquad \textbf{Eq 9.35}$$

$$= 10^{0.005} - 1 = 0.0116$$

...and converting the 1-degree phase error to radians

$$\delta = 1/57.3 = 0.0175$$

Using the expression for sideband suppression:

$$20 \log [(0.0116)^2 + (0.0175)^2]^{1/2}/2 = -39.6 \text{ dB} \qquad \textbf{Eq 9.36}$$

This is an easy rule of thumb—to obtain 40 dB of opposite sideband suppression, the amplitude errors must be kept under 0.1 dB and the phase errors under 1 degree.

In the receive case analyzed here, summing the I and Q channel outputs suppresses the lower sideband. The upper sideband may be suppressed by first inverting the Q channel and then summing, which subtracts the I and Q channel outputs. Note that this is the reverse of what happens in a phasing SSB transmitter, where summing the I and Q channel RF outputs suppresses the upper sideband. This interesting result must be considered when designing phasing SSB transceivers.

9.3 FROM MATHEMATICS TO PRACTICE

It is tempting to believe that a good designer draws a perfectly analyzed block diagram, picks the circuit blocks out of a circuit catalog, connects them up, and has an operating receiver on the bench. If the performance is not perfect, then at least the flaws are perfectly understood and predictable. The truth is that the deeper one digs into receiver analysis, the more obvious the omissions and approximations in the mathematical treatment become. A diode ring mixer is not a perfect sine-wave multiplier, and the mathematics for the proper treatment of even simple amplifier distortion is beyond the scope of a practical text.

There are two very different approaches to receiver design and development. The first approach is to design each fundamental circuit block as carefully as possible using whatever analysis and measurement tools are available, and then connect the blocks together in a manner as close as possible to the way they were analyzed and measured. Because RF test and measurement equipment operates in a 50-Ω environment, all circuit blocks are designed and tested to interconnect using 50-Ω transmission lines. The basic rule is that connections between circuit blocks should carry sinusoidal voltages 50 times larger and in-phase with sinusoidal currents. If voltages are not sinusoidal, simple low-pass filters will remove harmonics, and if impedances are different from 50 Ω, transformers may be used. This technique results in receivers with very predictable performance, and many parts. A conservative frequency converter using this approach is shown in **Fig 9.8**.

The second approach is to decide what function needs to be accomplished, and design a circuit with as few components as possible that will perform the task. A minimum-parts-count frequency converter is shown in **Fig 9.9**. Clearly, the second circuit is simpler than the first. From the professional circuit design standpoint, the second circuit might even be called "better" because is uses fewer parts and less operating current to perform the same

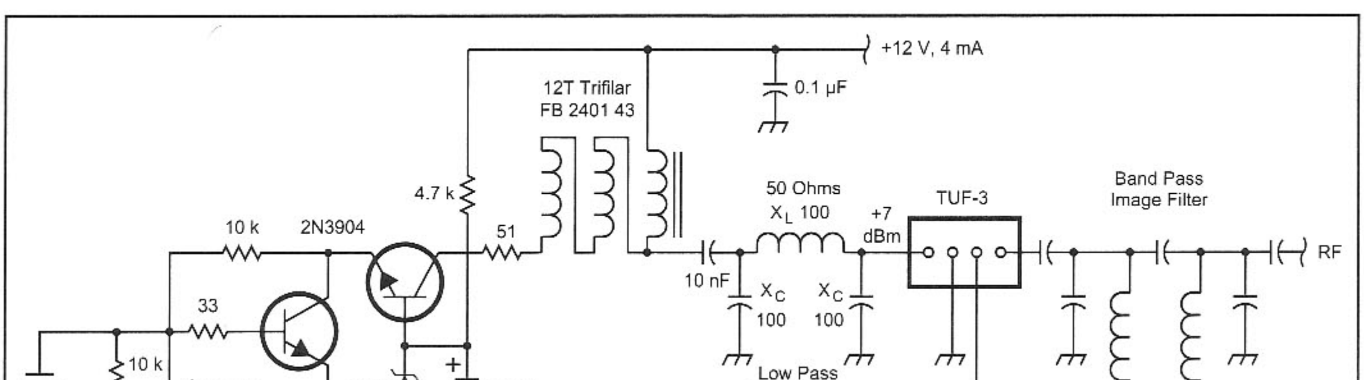

Fig 9.8—If voltages are not sinusoidal, simple low-pass filters will remove harmonics, and if impedances are different from 50 Ω, transformers may be used. This technique results in receivers with very predictable performance, and many parts. A conservative frequency converter using this approach is shown here.

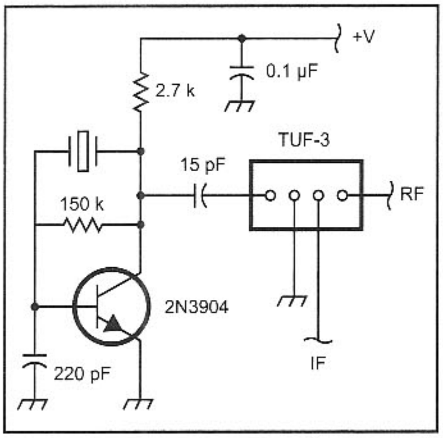

Fig 9.9—A minimum-parts-count frequency converter.

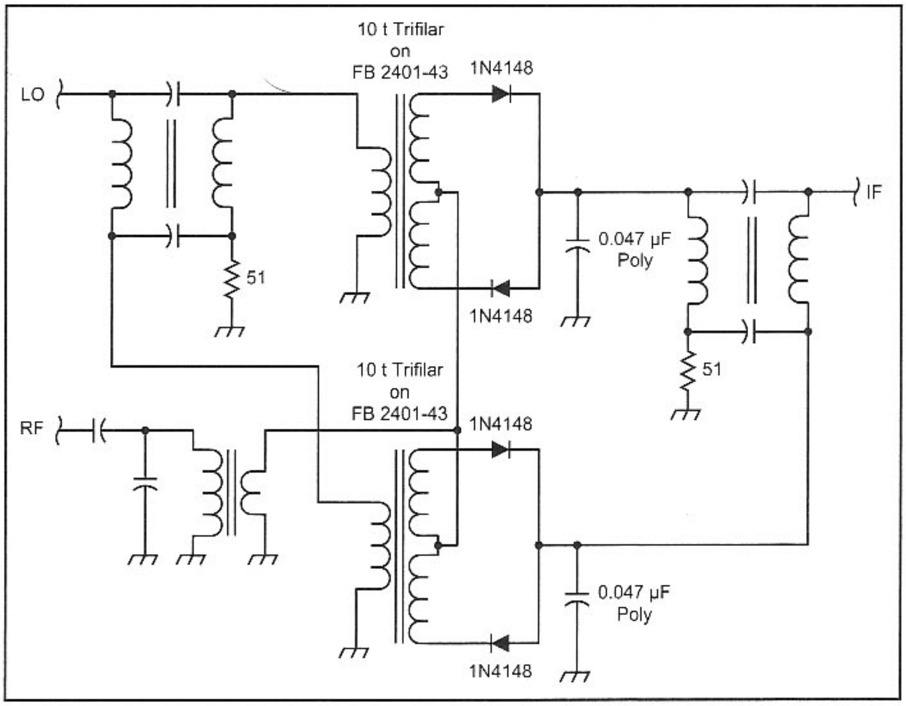

Fig 9.10—A minimum-parts-count image-reject detector that might be used in a simple CW receiver.

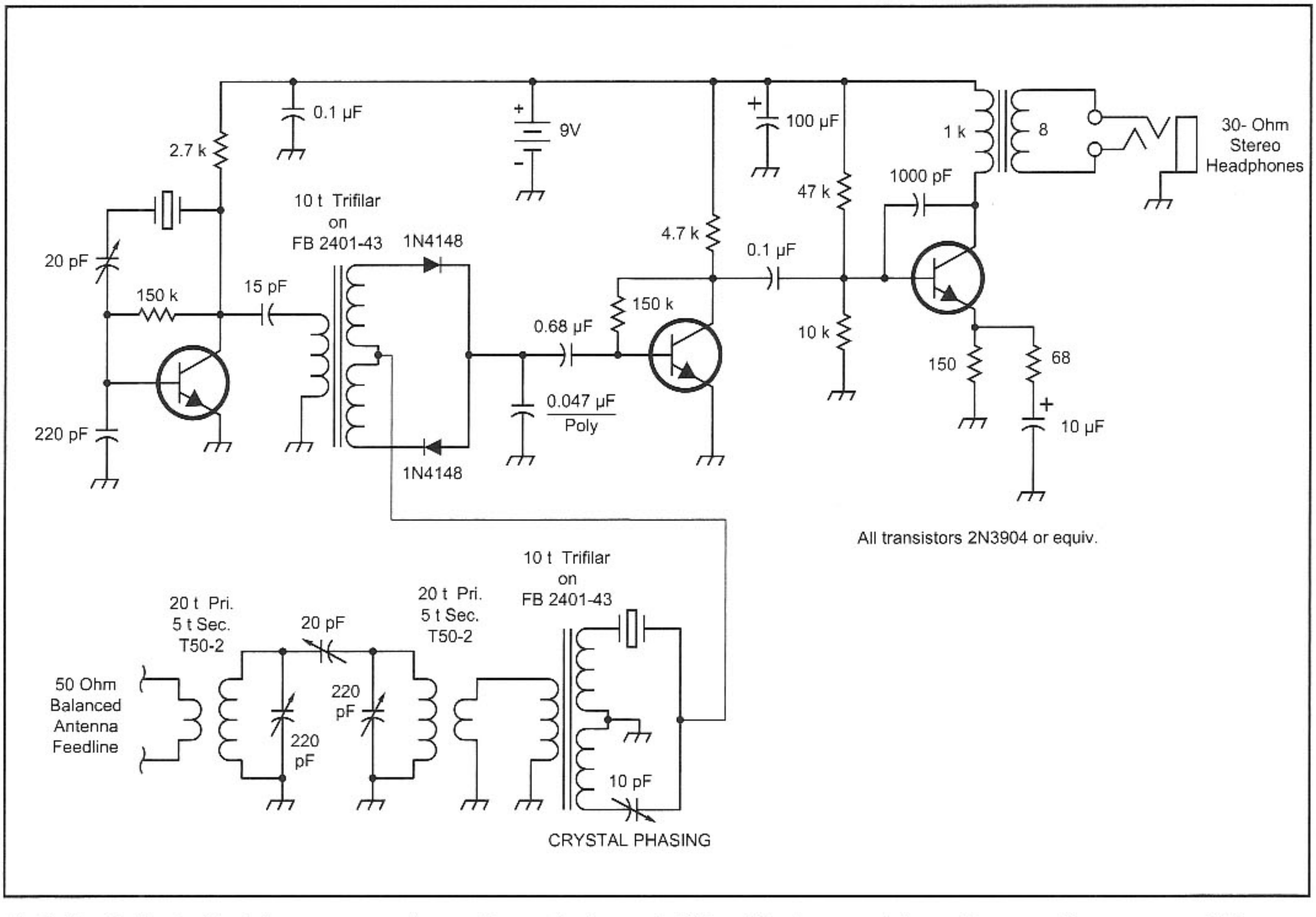

Fig 9.11—A simple fixed-frequency receiver using a single crystal filter. The two crystals are the same frequency, and the input circuit tunes from 3.5 to 7.5 MHz.

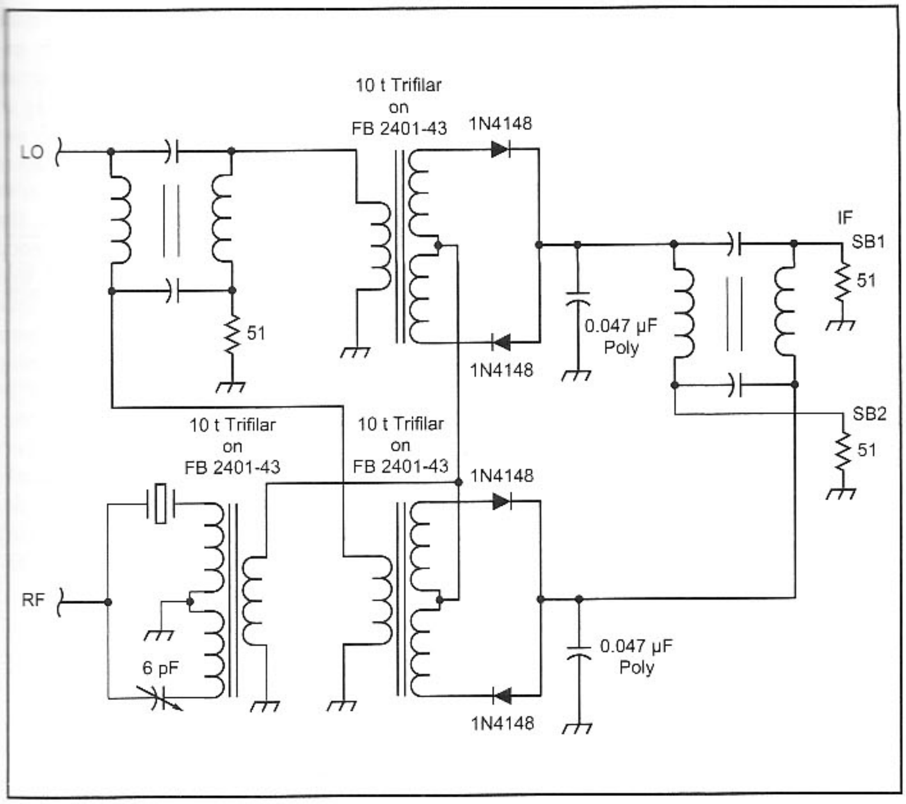

Fig 9.12—If the product detector is operated at a fixed frequency, crystal filter selectivity may be combined with a phasing product detector. This figure shows the basic circuit with a single-crystal CW filter connected directly to the product detector. It doesn't work as expected.

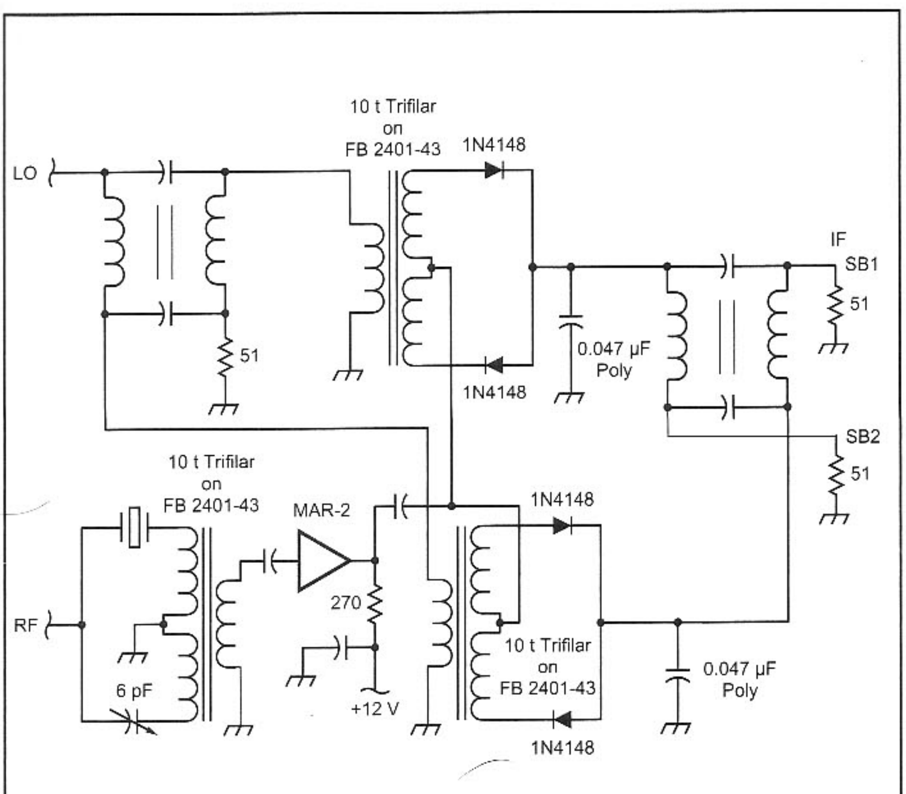

Fig 9.13—This circuit, with a buffer amplifier between the crystal filter and image-reject mixer, works as expected, with more than 40 dB of opposite sideband suppression at 1-kHz offset.

function. The difficulty arises when performance needs to be improved, or the circuit function is interconnected with other circuit blocks in a new and different way.

It is important to recognize that both approaches to RF circuit design are viable—the first offers higher performance from the outset, and a path to constant performance improvement by measuring and analyzing distortion and making incremental changes to the circuit blocks. The second approach involves more creativity and risk taking: attempts at new minimum-parts-count circuits often fail; and without 50 Ω ports, it is difficult to make diagnostic measurements without upsetting circuit behavior. Creative thinking, either in developing original circuits or pondering why they don't work as expected, is the delightful process designers use to solve problems.

There is a valid argument for both approaches to receiver projects—delightful simplicity is always a virtue—but there is a compelling argument for taking the methodical, analytical, 50-Ω approach to developing *phasing* receivers. A phasing receiver is a balanced system that depends on matching both amplitude and phase across significant bandwidths, through at least one frequency conversion, and with significant band limiting needed in both I and Q channels. Any deviation from perfect balance degrades opposite sideband suppression. Since amplitude and phase are both strong functions of termination impedances at mixer and amplifier ports, defining and controlling these impedances is the first step in building successful phasing receivers.

As an example of the problems that arise when impedance matching is neglected, let's look at a minimum-parts-count image-reject detector that might be used in a simple CW receiver. **Fig 9.10** illustrates the circuit. The RF ports of the two balanced diode mixers are simply tied together, and the LO and IF ports are quadrature split and combined using hybrid circuits. This circuit provides a useful reduction in opposite sideband interference. The selectivity curve is very similar to the classic receivers with single crystal filters and phasing controls.

The circuit in Fig 9.10 might be used as the product detector in a simple superhet receiver. For comparison, **Fig 9.11** is a simple fixed-frequency IF receiver using a single crystal filter. The image-reject product detector has a few more parts.

If the product detector is operated at a fixed frequency, crystal filter selectivity may be combined with a phasing product detector. **Fig 9.12** is the basic circuit with a single-crystal CW filter connected directly to the product detector. The crystal filter selectivity should add to the image-reject product detector circuitry, for very respectable performance. It does not work. The opposite sideband suppression is considerably less than expected.

The problem is that image-reject mixer behavior is strongly dependent on the impedances at the various mixer ports. By directly connecting the crystal filter, the mixer RF ports see an impedance that varies rapidly from one sideband to the other. The impedance in the desired band is resistive and reasonably well matched, but the impedance on the undesired sideband is almost perfectly reflective. A reflective mixer termination on one sideband and an absorptive termination on the other severely impacts image-reject mixer performance. In simulations, the opposite sideband suppression of the *filter* is maintained, but almost all of the opposite sideband suppression from the *image-reject mixer* circuitry is lost.

The circuit of **Fig 9.13**, with a buffer amplifier between the crystal filter and image-reject mixer, works as expected, with more than 40 dB of opposite sideband suppression at 1 kHz offset. A casual glance at this circuit would not hint that the added broadband components would significantly improve opposite sideband suppression.

The most termination-sensitive components in a phasing receiver or exciter are usually the mixers. Since providing wideband, resistive terminations to the mixer RF, LO and IF ports improves distortion performance in addition to opposite sideband suppression, it is simply good practice in phasing rigs. Paying attention to termination impedances usually adds components and complexity to circuits.

If adequate performance at minimum cost with few parts is the goal, it is unlikely that a phasing receiver or exciter can compete with a basic superhet. Making an intelligent choice about whether to use phasing techniques in a receiver involves weighing a number of factors. A strict phasing receiver can never achieve the opposite sideband selectivity of a good superhet with multiple crystal filters, and a superhet will always have more spurious responses and internally generated spurs than a direct conversion receiver. Nothing can compare with the sonic clarity of simple wide audio bandwidth direct conversion receivers. The choice of receiver architecture may not be made for purely practical reasons—an Amateur Radio designer-builder has the luxury of working on a technique purely for the joy of exploring new territory.

9.4 SIDEBAND SUPPRESSION DESIGN

The point of adding a phasing system to a receiver or exciter is to suppress one sideband. The first generation of amateur phasing circuits from the late 1940s into the 1950s were literally added on to conventional receivers and transmitters. Later commercial transmitters from Central Electronics, Hallicrafters, and others used conventional heterodyne methods, with phasing sideband selection and conventional tuned circuits at a fixed IF. Many recent improvements in performance have resulted from designing the entire radio system, from headphones and microphone to antenna, with phasing in mind. Before diving into more detailed system discussions, it is useful to discuss the amount of sideband suppression desired.

It is relatively easy to design and reproduce phasing circuitry to achieve undesired sideband suppression of more than 30 dB. With just a little more design care, and well-matched components, just over 40 dB of undesired sideband suppression may be routinely obtained. The receivers and exciter in the *QST* references from 1992 through 1995 all exhibit sideband suppression in the 41 to 43 dB range, when the circuit boards are used as intended. The receiver and exciter circuits shown at the end of this chapter consistently achieve sideband suppression in the mid-50 dB range, using 0.1% matched components and very careful alignment. It is worth emphasizing at this point that the level of sideband suppression depends on circuit design; precision components; and careful alignment. A 60 dB circuit can be designed, but component tolerances are unrealistically tight, alignment is difficult, and performance degrades as components age. A 40 dB circuit design works well with standard 1% components, and has quick and easy alignment that will hold for the life of the radio. 20 dB sideband suppression circuits work with junk box parts and no adjustments at all.

Before continuing with a further exploration of sideband suppression, a discussion of "how much is enough" is in order. As in most engineering questions, the answer begins with "that depends...." First of all, we should note that systems with no sideband suppression at all are entirely functional for some applications. A signal from a DSB transmitter is converted to SSB in the receiver, and once tuned in the operator can't tell the difference. Similarly, DSB receivers have been used for CW and SSB signals since the early days of radio.

DSB is attractive whenever simplicity is more important than spectral efficiency or interference rejection. A DSB transmitter may be paired with a direct conversion DSB receiver to build an ultra-simple rig. A disadvantage of such a radio is that it can not receive DSB very well, and it's transmitted signal must be received on a receiver with some provision for either suppressing one sideband or locking to the missing carrier frequency. Somehow a radio that cannot communicate with an identically equipped station seems incomplete.

Transmitters

A Single Sideband transmitter needs enough carrier suppression that the carrier is not evident when tuning in the signal, and enough opposite sideband suppression that the opposite sideband frequencies may be used for communications by other stations. 40 dB of carrier suppres-

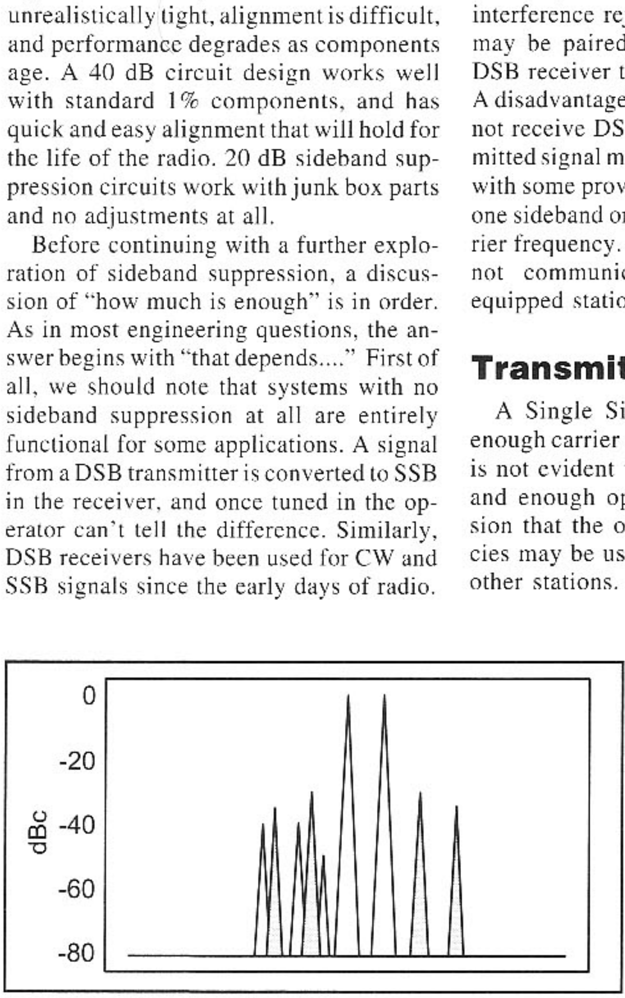

Fig 9.14—The spectrum of a typical SSB transmitter with two-tone modulation, with the carrier suppressed 50 dB, 40 dB opposite sideband suppression, and amplifier intermod products 30 dB (3rd) and 35 dB (5th) below either of the two desired output frequencies.

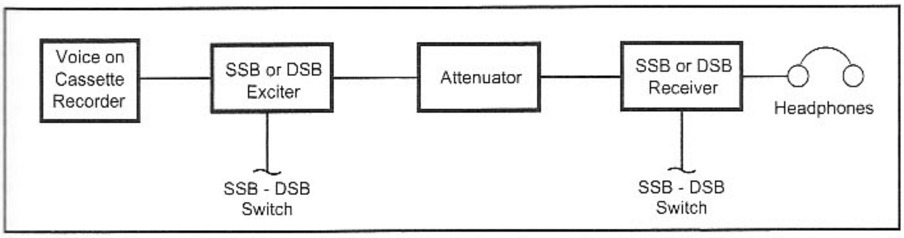

Fig 9.15—This test setup was used for a set of experiments to investigate the minimum sideband suppression needed for good SSB reception.

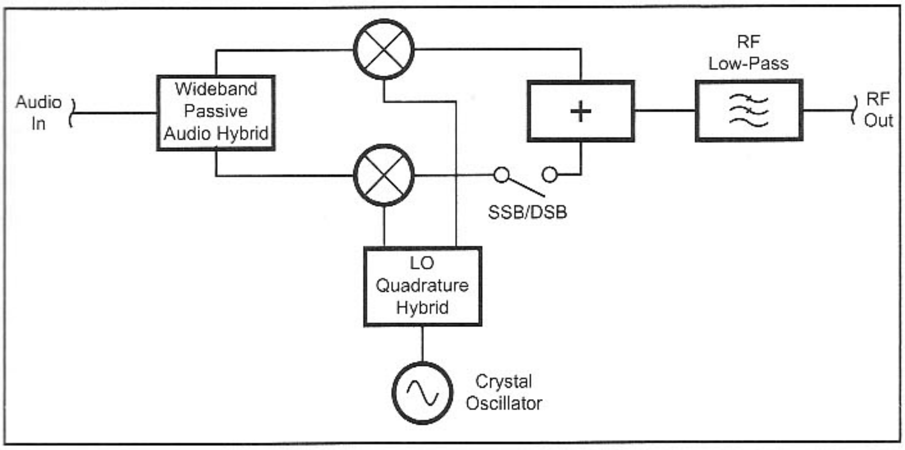

Fig 9.16—The exciter block diagram.

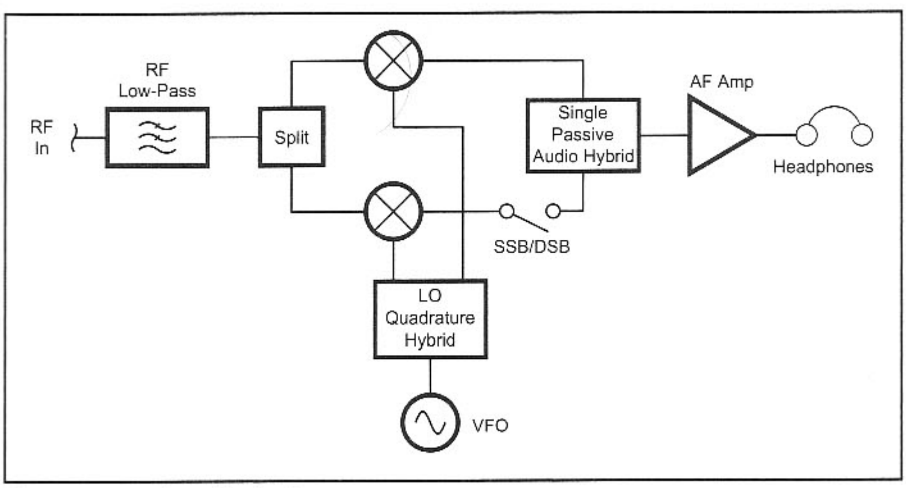

Fig 9.17—Receiver block diagram.

sion is generally considered sufficient, although at this level the carrier will often be noticeable to stations with good receivers and good ears. Opposite sideband suppression should be good enough that interference in the opposite sideband frequency band is dominated by amplifier intermod products, and not intelligible audio. **Fig 9.14** shows the spectrum of a typical SSB transmitter with two-tone modulation, with the carrier suppressed 50 dB, 40 dB opposite sideband suppression, and amplifier intermod products 30 dB (3rd) and 35 dB (5th) below either of the two desired output frequencies. This transmitter would sound very good on the air.

The intermodulation products are highlighted in gray. Clearly it is not necessary to suppress the opposite sideband in a SSB transmitter by much more than 40 dB, because the intermod products occupy the same frequencies and they are only about 30 dB below the desired sideband level in a well-designed transmitter. More carrier suppression is useful, however, because the carrier is present during breaks in speech.

In the past few years, digital modes that use a computer sound card connected to the microphone input of a SSB transmitter have become popular. Transmitters for these modes benefit from having much lower distortion than SSB or keyed-carrier CW transmitters. Combining a phasing exciter with a crystal filter and very low distortion RF amplifier would make it possible to generate a PSK-31 signal that would be stunningly clean. PSK-31 operators display the whole spectrum of received in-channel distortion products on strong signals, so a clean signal is instantly recognizable on the air. Because PSK-31 stations operate in narrow bands, with tuning performed in baseband signal processing, a dedicated PSK-31 exciter and crystal filter can be built at the final output frequency, with no need for heterodyning.

Given that DSB transmitters are functional, and 40 dB of opposite sideband suppression is enough for SSB transmitter applications, are there any benefits to having less than 40 dB of opposite suppression but more than 0 dB? A set of experiments was performed to investigate the minimum sideband suppression needed for good SSB reception. **Fig 9.15** illustrates the test setup. **Fig 9.16** is the exciter block diagram, and **Fig 9.17** is the receiver block diagram. The exciter and receiver each have a switch to enable or disable the sideband suppression circuitry. The approximate sideband suppression available at the exciter, receiver, and the combined sideband suppression are shown in **Fig 9.18**.

Here are the comments copied from the lab notebook:

DSB-DSB Really hard to tune. Very poor sound.

DSB transmit, single-hybrid SSB Receive. Much better With the hybrid zero

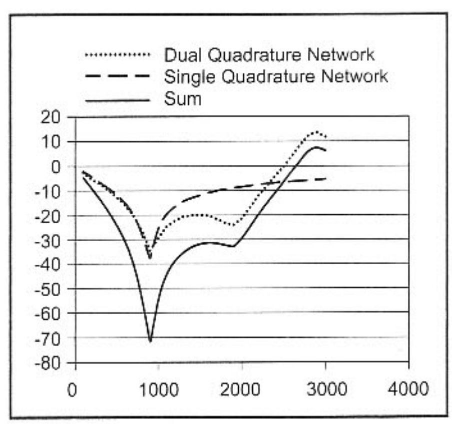

Fig 9.18—The approximate sideband suppression available at the exciter, receiver and the combined sideband suppression.