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12.9 References and Bibliography

Chapter 12 — CD-ROM Content

Supplemental Files

- "Amateur Radio Equipment Development — An Historical Perspective" by Joel Hallas, W1ZR

Project Files

- Rock Bending Receiver PCB template by Randy Henderson, WI5W
- 10 GHz preamp PCB template by Zack Lau, W1VT
- Binaural Receiver project by Rick Campbell, KK7B

Chapter 12

Receivers

Receivers are discussed in this chapter from the standpoint of their architecture, composed of functional blocks such as mixers and oscillators and amplifiers that are discussed in the other chapters of this book. Other functions such as AGC that control the receiver from outside the primary signal path are dealt with both functionally and with examples at the circuit level. Special techniques associated with VHF, UHF and microwave receivers are reviewed.

The performance of receivers is described by standard terminology and as a set of measurements that allow a wide range of designs to be evaluated in ways that reflect their behavior in actual use.

This chapter, originally written by Joel Hallas, W1ZR, includes an update of existing material from previous editions of this book. Joel's summary of receiver and transmitter development, "Amateur Radio Equipment Development — An Historical Perspective," provides valuable insight as to why radios are constructed the way they are and is included on the CD-ROM accompanying this book.

The major subsystems of a radio receiving system are the antenna, the receiver and the information processor. The antenna's task is to provide a transition from an electromagnetic wave in space to an electrical signal that can be conducted on wires. The receiver has the job of retrieving the information content from a particular ac signal coming from the antenna and presenting it in a useful format to the processor for use.

The processor typically is an operator, but can also be an automated system. When you consider that most "processors" require signals in the range of volts (to drive an operator's speaker or headphones, or even the input of an A/D converter), and the particular signal of interest arrives from the antenna at a level of mere microvolts, the basic function of the receiver is to amplify the desired signal by a factor of a million. It must do this, while in the presence of signals many orders of magnitude greater and of completely different characteristics, without distortion of the desired signal or loss of the information it carries.

12.1 Characterizing Receivers

As we discuss receivers we will need to characterize their performance, and often their performance limitations, using certain key parameters. The most commonly encountered are as follows:

Sensitivity — This parameter is a measure of how weak a signal the receiver can extract information from. This generally is expressed at a particular signal-to-noise ratio (SNR) since noise is generally the limiting factor. A typical specification might be: "Sensitivity: 1 µV for 10 dB SNR with 3 kHz bandwidth." The bandwidth is stated because the amount (or power) of the noise, the denominator of the SNR fraction, increases directly with bandwidth. Generally the noise parameter refers to the noise generated within the receiver, often less than the noise that arrives with the signal from the antenna.

Selectivity — Selectivity is just the bandwidth discussed above. This is important because to a first order it identifies the receiver's ability to separate stations. With a perfectly sharp filter in an ideal receiver, stations within the bandwidth will be heard, while those outside it won't be detected. The selectivity thus describes how closely spaced adjacent channels can be. With a perfect 3 kHz bandwidth selectivity, and signals restricted to a 3 kHz bandwidth at the transmitter, a different station can be assigned every 3 kHz across the spectrum. In a less than ideal situation, it is usually necessary to include a *guard band* between channels.

Note that the word *channel* is used here in its generic form, meaning the amount of spectrum occupied by a signal, and not defining a fixed frequency such as an AM broadcast channel. A CW channel is about 300 Hz wide, a SSB channel about 2.5-3 kHz wide, and so forth. "Adjacent channel" refers to spectrum immediately higher or lower in frequency.

Dynamic Range — In the case of all real receivers, there is a range of signals that a receiver can respond to. This is referred to as dynamic range and, as will be discussed in more detail, can be established based on a number of different criteria. In its most basic form, dynamic range is the range in amplitude of signals that can be usefully received, typically from as low as the receiver's noise level, or *noise floor*, to a level at which stages overload in some way.

The type and severity of the overload is often part of the specification. A straightforward example might be a 130 dB dynamic range with less than 3% distortion. The nature of the distortion will determine the observed phenomenon. If the weakest and strongest signals are both on the same channel, for example, we would not expect to be able to process the weaker of the two. However, the more interesting case would be with the strong signal in an adjacent channel. In an ideal receiver, we would never notice that the adjacent signal was there. In a real receiver with a finite dynamic range or nonideal selectivity, there will be some level of adjacent channel signal or signals that will interfere with reception of the weaker on-channel signal.

The parameters described above are often the key performance parameters, but in many cases there are others that are important to specify. Examples are audio output power, power consumption, size, weight, control capabilities and so forth.

12.2 Basics of Heterodyne Receivers

The *heterodyne* receiver combines the input signal with a signal from a *local oscillator* (LO) in a nonlinear device called a *mixer* to result in the sum and difference frequencies as shown in **Fig 12.1**. The receiver may be designed so the output signal is anything from dc (a so-called *direct conversion* receiver) to any frequency above or below either of the two frequencies. The major benefit is that most of the gain, bandwidth setting and processing are performed at a single frequency.

By changing the frequency of the LO, the operator shifts the input frequency that is *translated* to the output, along with all its modulated information. In most receivers the mixer output frequency is designed to be an RF signal, either the sum or difference — the other being filtered out at this point. This output frequency is called an *intermediate frequency* or *IF*. The IF amplifier system can be designed to provide the selectivity and other desired characteristics centered at a single fixed frequency.

When more than one frequency conversion process is used, the receiver becomes a *superheterodyne* or *superhet*. A block diagram of a typical superhet receiver is shown in **Fig 12.2**. In traditional form, the RF filter is used to limit the input frequency range to those frequencies that include only the desired sum or difference but not the other — the so-called *image* frequency. The dotted line represents the fact that in receivers with a wide tuning range, such as a simple AM broadcast receiver that tunes from 500 to 1700 kHz, a more than 3:1 range, the input RF amplifier and filter is often tracked along with the local oscillator. The IF filter is used to establish operating selectivity — that required by the information bandwidth. Circuits from the antenna input through and including the mixer (the first mixer if more than one mixing stage is used) are generally referred to as the receiver's *front-end*.

For reception of suppressed carrier single-sideband voice (SSB) or on-off or frequency-shift keyed (FSK) signals, a second *beat frequency oscillator* or *BFO* is employed to provide an audible voice, an audio tone or tones at the output for operator or FSK processing. This is the same as a heterodyne mixer with an output centered at dc, although the IF filter is usually designed to remove one of the output products.

Modern receivers using digital signal processing for operating bandwidth and information detection convert the incoming signal to digital form at one of the intermediate frequencies. Advancing speed and declining costs of analog-to-digital converters (ADC) and processors are moving the conversion closer and closer to the incoming signal's frequency, in some cases converting directly at RF, which is called *direct sampling*. See the **DSP and Software Radio Design** chapter for more information on these techniques.

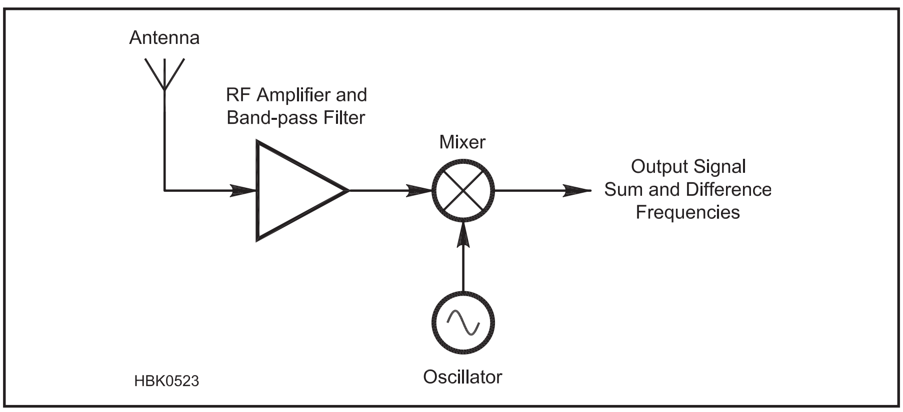


Fig 12.1 — Basic architecture of a heterodyne direct conversion receiver.

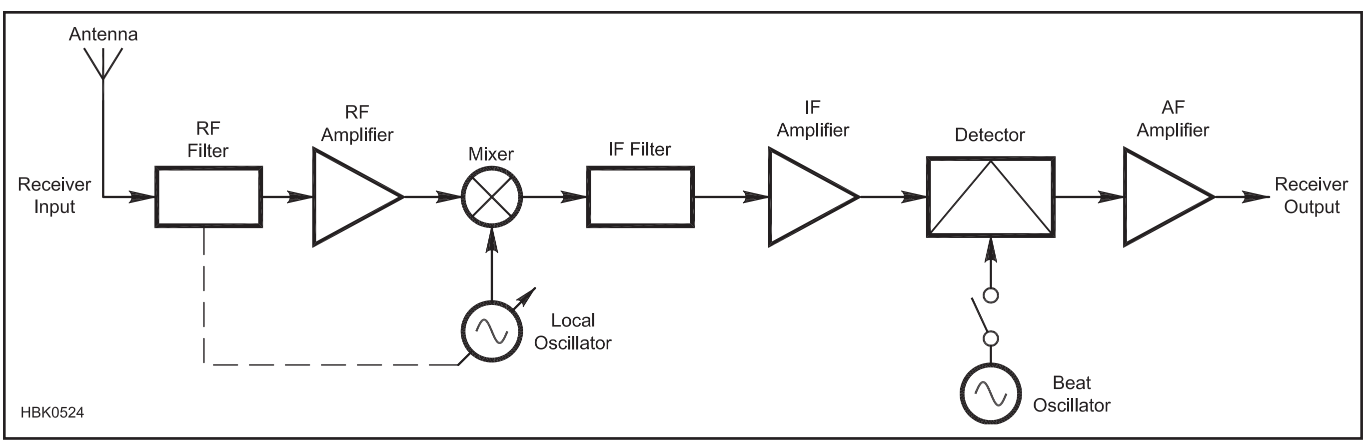


Fig 12.2 — Elements of a traditional superheterodyne radio receiver

12.2.1 The Direct Conversion Receiver

The heterodyne process can occur at a number of different points in the receiver. The simplest form of heterodyne receiver is called a *direct conversion* receiver because it performs the translation directly from the signal frequency to the audio output. It is, in effect, just the BFO and detector of the general superhet shown in Fig 12.2. In this case, the detector is often preceded by an RF amplifier with a typical complete receiver shown in **Fig 12.3**. Such a receiver can be very simple to construct, yet can be quite effective — especially for the ultra-compact low-power consumer-oriented portable station known as the mobile telephone! In fact, given the worldwide presence of the mobile telephone, the direct conversion receiver is the most widely used of all receivers.

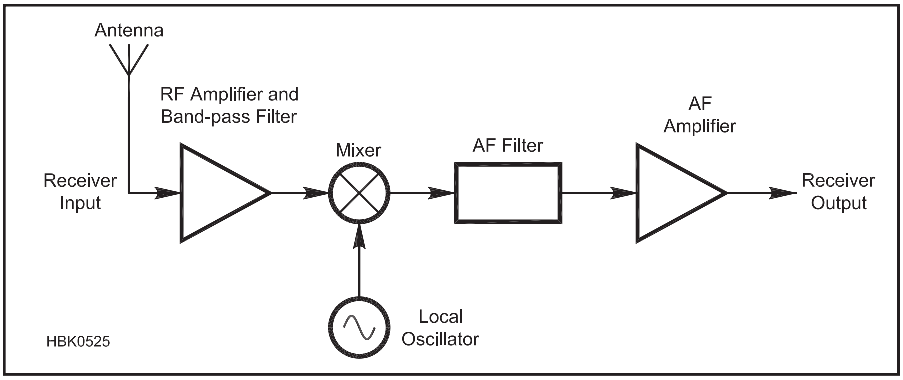


Fig 12.3 — Block diagram of a direct conversion.

The basic function of a mixer is to multiply two sinusoidal signals and generate two new signals with frequencies that are the sum and difference of the two original signals. This function can be performed by a linear multiplier, a switch that turns one input signal on and off at the frequency of the other input signal, or a nonlinear circuit such as a diode. (The output of a nonlinear circuit is made up of an infinite series of products, all different combinations of the two input signals, as described in the sidebar on Nonlinear Signal Combinations.) Much more information about the theory, operation and application of mixers may be found in the **Mixers, Modulators and Demodulators** chapter.

Fig 12.4 shows the progression of the spectrum of an on-off keyed CW signal through such a receiver based on the relationships described above. In 12.4C, we include an undesired image signal on the other side of the local oscillator that also shows up in the output of the receiver. Note each of the desired and undesired responses that occur as outputs of the mixer.

Some mixers are designed to be *balanced* in order to cancel one of the input signals at the output while a *double-balanced* mixer cancels both. A double-balanced mixer simplifies the output filtering job as shown in Fig 12.4D.

Products generated by nonlinearities in the mixing process (see the sidebar on Nonlinear Signal Combinations) are heard as intermodulation distortion signals that we will discuss later. Note that the nonlinearities also allow mixing with unwanted signals near multiples of local oscillator frequency. These signals, such as those from TV or FM broadcast stations, must be eliminated in the filtering before the mixer since their audio output will be right in the desired passband on the output of the mixer.

Nonlinear Signal Combinations

Although a mixer is often thought of as nonlinear, it is neither necessary nor desirable for a mixer to be nonlinear. An ideal mixer is one that linearly multiplies the LO voltage by the signal voltage, creating two products at the sum and difference frequencies and only those two products. From the signal's perspective, it is a perfectly linear but time-varying device. Ideally a mixer should be as linear as possible.

If a signal is applied to a nonlinear device, however, the output will not be just a copy of the input, but can be described as the following infinite series of output signal products:

$$V_{OUT} = K_0 + K_1 \times V_{IN} + K_2 \times V_{IN}^2 + K_3 \times V_{IN}^3 + \ldots + K_N \times V_{IN}^N \qquad \text{(A)}$$

What happens if the input V_{IN} consists of two sinusoids at F_1 and F_2, or A × [sin $(2\pi_1) \times t$] and B × [sin $(2\pi F_2) \times t$]? Begin by simplifying the notation to use angular frequency in radians/second ($2\pi F = \omega$). Thus V_{IN} becomes $A\sin\omega_1 t$ and $B\sin\omega_2 t$ and equation A becomes:

$$V_{OUT} = K_0 + K_1 \times (A\sin\omega_1 t + B\sin\omega_2 t) + K_2 \times (A\sin\omega_1 t + B\sin\omega_2 t)^2 + K_3 \times (A\sin\omega_1 t + B\sin\omega_2 t)^3 + \ldots + K_N \times (A\sin\omega_1 t + B\sin\omega_2 t)^N \qquad \text{(B)}$$

The zero-order term, K_0, represents a dc component and the first-order term, $K_1 \times (A\sin\omega_1 t + B\sin\omega_2 t)$, is just a constant times the input signals. The second-order term is the most interesting for our purposes. Performing the squaring operation, we end up with:

$$\text{Second order term} = [K_2 A^2 \sin^2\omega_1 t + 2K_2 AB(\sin\omega_1 t \times \sin\omega_2 t) + K_2 B^2 \sin^2\omega_2 t] \qquad \text{(C)}$$

Using the trigonometric identity (see Reference 1):

sin á sin â = ½ {cos (á – â) – cos (á + â)}

the product term becomes:

$$K_2 AB \times [\cos(\omega_1 - \omega_2)t - \cos(\omega_1 + \omega_2)t] \qquad \text{(D)}$$

Here are the products at the sum and difference frequency of the input signals! The signals, originally sinusoids are now cosinusoids, signifying a phase shift. These signals, however, are just two of the many products created by the nonlinear action of the circuit, represented by the higher-order terms in the original series.

In the output of a mixer or amplifier, those unwanted signals create noise and interference and must be minimized or filtered out. This nonlinear process is responsible for the distortion and intermodulation products generated by amplifiers operated nonlinearly in receivers and transmitters.

Project: A Rock-Bending Receiver for 7 MHz

There are many direct conversion receiver (DC) construction articles in the amateur literature. Home builders considering construction of a DC receiver should read Chapter 8 in *Experimental Methods in RF Design*, which outlines many of the pitfalls and design limitations of such receivers, as well as a providing suggestions for making a successful DC receiver.

This DC receiver design by Randy

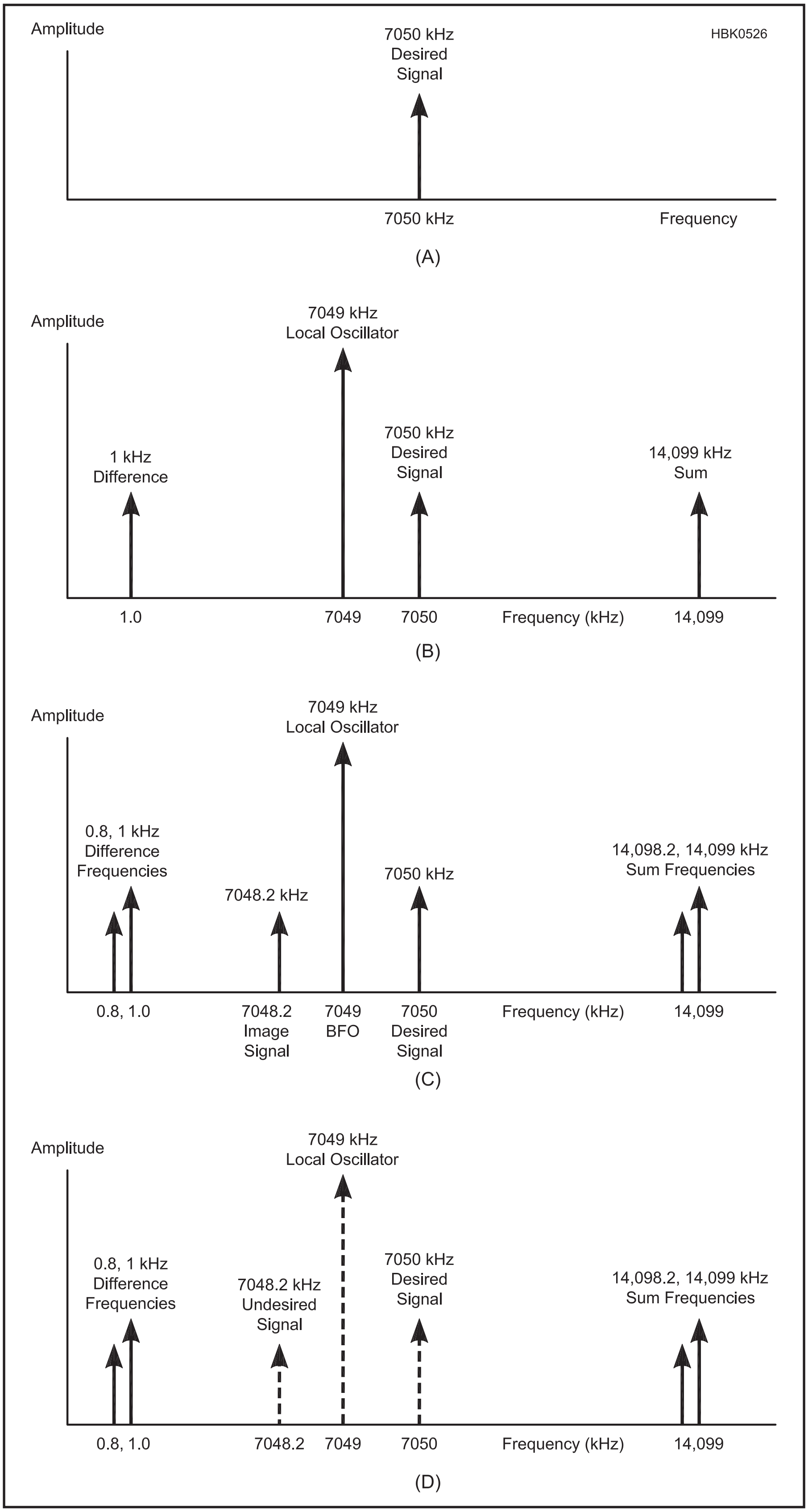


Fig 12.4 — Frequency relationships in DC receiver. At (A), desired receive signal from antenna, 7050 kHz on-off keyed carrier. At (B), internal local oscillator and receive frequency relationships. At (C), frequency relationships of mixer/detector products (not to scale).At (D), sum and difference outputs from double balanced mixer (not to scale). Note the balanced out mixer inputs that cancel at the output (dashed lines).

Henderson, WI5W, is presented here as an example. This receiver represents about as simple a receiver as can be constructed that will offer reasonable performance (see the References entry). The user of such a DC receiver will appreciate its simplicity and the nice sounding response of CW and SSB signals that it receives, but only if there aren't many signals sharing the band.

A major shortcoming of the DC receiver architecture will soon become evident — it has no rejection of *image* signals. Looking again at Fig 12.4B, while the 7049 kHz oscillator will translate the desired 7050 kHz signal to an audio tone of 1 kHz, it will equally well translate a signal at 7048 kHz to the same frequency. This is called an *image* and results in interference. The same kind of thing happens while receiving SSB; the entire channel on the other side of the BFO is received as well.

Building a stable oscillator is often the most challenging part of a simple receiver. This one uses a tunable crystal-controlled oscillator that is both stable and easy to reproduce. All of its parts are readily available from multiple sources and the fixed value capacitors and resistors are common components available from many electronics parts suppliers.

THE CIRCUIT

This receiver works by mixing two radio-frequency signals together. One of them is the signal you want to hear, and the other is generated by an oscillator circuit (Q1 and associated components) in the receiver. In **Fig 12.5**, mixer U1 puts out sums and differences of these signals and their harmonics. We don't use the sum of the original frequencies, which comes out of the mixer in the vicinity of 14 MHz. Instead, we use the frequency *difference* between the incoming signal and the receiver's oscillator — a signal in the audio range if the incoming signal and oscillator frequencies are close enough to each other. This signal is filtered in U2, and amplified in U2 and U3. An audio transducer (a speaker or headphones) converts U3's electrical output to audio.

How the Rock Bender Bends Rocks

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

L2's value determines the degree of pulling available. Using FT-243 style crystals and larger L2 values, the oscillator reliably tunes

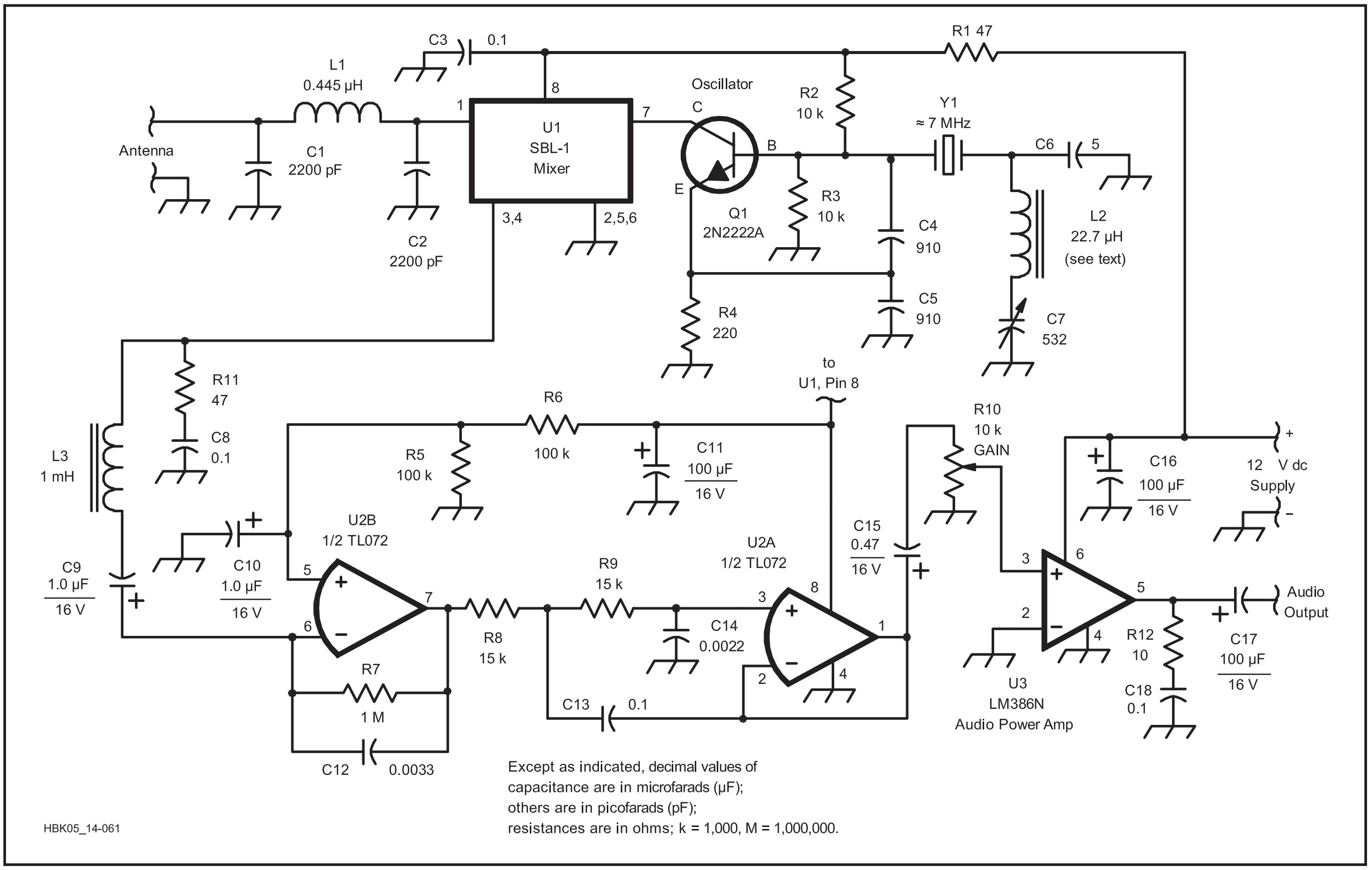


Fig 12.5 — An SBL-1 mixer (U1, which contains two small RF transformers and a Schottky-diode quad), a TL072 dual op-amp IC (U2) and an LM386 low-voltage audio power amplifier IC (U3) do much of the Rock-Bending Receiver's magic. Q1, a variable crystal oscillator (VXO), generates a low-power radio signal that shifts incoming signals down to the audio range for amplification in U2 and U3. All of the circuit's resistors are ¼ W, 5% tolerance types; the circuit's polarized capacitors are 16 V electrolytics, except C10, which can be rated as low as 10 V. The 0.1 µF capacitors are monolithic or disc ceramics rated at 16 V or higher.

C1, C2 — Ceramic or mica, 10% tolerance.
C4, C5, and C6 — Polystyrene, dipped silver mica, or C0G (formerly NP0) ceramic, 10% tolerance.
C7 — Dual gang broadcast variable capacitor (14-380 pF per section), ¼ inch dia shaft, available as #BC13380 from Ocean State Electronics. A rubber equipment foot serves as a knob. (Any variable capacitor with a maximum capacitance of 350 to 600 pF can be substituted; the wider the capacitance range, the better.)
C12, C13, C14 — 10% tolerance. For SSB, change C12, C13 and C14 to 0.001 µF.
U2 — TL072CN or TL082CN dual JFET op amp.
L1 — Four turns of AWG #18 wire on ¾ inch PVC pipe form. Actual pipe OD is 0.85 inch. The coil's length is about 0.65 inch; adjust turns spacing for maximum signal strength. Tack the turns in place with cyanoacrylic adhesive, coil dope or RTV sealant. (As a substitute, wind 8 turns of #18 wire around 75% of the circumference of a T-50-2 powdered-iron core. Once you've soldered the coil in place and have the receiver working, expand and compress the coil's turns to peak incoming signals, and then cement the winding in place.)
L2 — Approximately 22.7 µH; consists of one or more encapsulated RF chokes in series (two 10-µH chokes [Mouser #43HH105 suitable] and one 2.7-µH choke [Mouser #43HH276 suitable] used by author). See text.
L3 — 1 mH RF choke. As a substitute, wind 34 turns of #30 enameled wire around an FT-37-72 ferrite core.
Q1 — 2N2222, PN2222 or similar small-signal, silicon NPN transistor.
R10 — 5 or 10 kΩ audio-taper control (RadioShack No. 271-215 or 271-1721 suitable).
U1 — Mini-Circuits SBL-1 mixer.
Y1 — 7 MHz fundamental-mode quartz crystal. Ocean State Electronics carries 7030, 7035, 7040, 7045, 7110 and 7125 kHz units.
PC boards for this project are available from FAR Circuits.

from the frequency marked on the holder to about 50 kHz below that point with larger L2 values. (In the author's receiver a 25 kHz tuning range was achieved.) The oscillator's frequency stability is very good.

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

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

Input Filter and Mixer

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

Audio Amplifier and Filter

U2B amplifies the audio signal from U1. U2A serves as an active low-pass filter. The values of C12, C13 and C14 are appropriate for listening to CW signals. If you want SSB stations to sound better, make the changes shown in the parts list for Fig 12.5.

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

CONSTRUCTION

If you're already an accomplished builder, you know that this project can be built using a number of construction techniques, so have at it! If you're new to building, you should consider building the Rock-Bending Receiver on a printed circuit (PC) board. (The parts list tells where you can buy one ready-made.) See **Fig 12.6** for details on the physical layout of several important components used in the receiver. The receiver can be constructed either using a PC board or by using ground-plane (a.k.a "ugly") construction. **Fig 12.7** shows a photo of the receiver built on a PC board, but you can also build one on a ground-plane formed by an unetched piece of PC board material.

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

Another important thing to do if you use a homemade double-sided PC board is to countersink the ground plane to clear all ungrounded holes. (Countersinking clears copper away from the holes so components won't short-circuit to the ground plane.) A ¼-inch diameter drill bit works well for this. Attach a control knob to the bit's shank and you can safely use the bit as a manual countersinking tool. If you countersink your board in a drill press, set it to about 300 rpm or less, and use very light pressure on the feed handle.

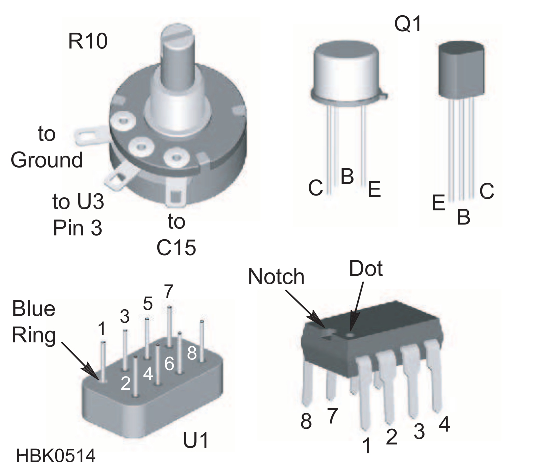


Fig 12.6 — Two Q1-case styles are shown because plastic or metal transistors will work equally well for Q1. If you build your Rock-Bending Receiver using a prefab PC board, you should mount the ICs in 8-pin mini-DIP sockets rather than just soldering the ICs to the board.

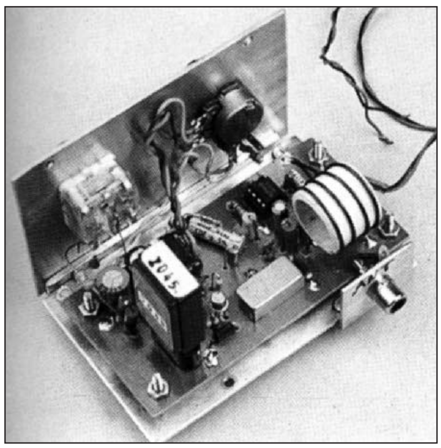


Fig 12.7 — Ground-plane construction or PC-board construction as shown here — either approach can produce the same good Rock Bending Receiver performance.

Mounting the receiver in a metal box or cabinet is a good idea. Plastic enclosures can't shield the TUNING capacitor from the presence of your hand, which may slightly affect the receiver tuning. You don't have to completely enclose the receiver — a flat aluminum panel screwed to a wooden base is an acceptable alternative. The panel supports the tuning capacitor, GAIN control and your choice of audio connector. The base can support the circuit board and antenna connector.

CHECKOUT

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

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

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

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

TROUBLE?

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

You can use this technique at U2 (pin 3, then pin 6) and all the way to the antenna. If you hear loud pops when touching either end of L3 but not the antenna connector, the oscillator is probably not working. You can check for oscillator activity by putting the receiver near a friend's transceiver (both must be in the same room) and listening for the VXO. Be sure to adjust the TUNING control through its range when checking the oscillator.

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

OPERATION

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

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

12.3 The Superheterodyne Receiver

In many instances, it is not possible to achieve all the receiver design goals with a single-conversion receiver and multiple conversion steps are used, creating the superheterodyne architecture. Traditionally, the first conversion is tasked with removing the RF image signals, while the second allows processing of the IF signal to provide the information based IF processing.

The superheterodyne receiver applies the principles of the heterodyne receiver at least twice. The concept was introduced by Major Edwin Armstrong, a US Army artillery officer, just as WW I was coming to a close. He is the same Armstrong who invented frequency modulation (FM) some years later and who held many radio patents between WW I and WW II.

In a superhet, a local oscillator and mixer are used to translate the received signal to an intermediate frequency or IF rather than directly to audio. This provides an opportunity for additional amplification and processing. Then a second mixer is used as in the DC receiver to detect the IF signal, translating it to audio. The configuration was shown previously in Fig 12.2.

An example will illustrate how this works. Let's pick a common IF frequency used in a simple AM broadcast radio, 455 kHz. If we want to listen to a 600 kHz broadcast station, the RF stage would be set to amplify the 600 kHz signal and the LO should be set to 600 + 455 kHz or 1055 kHz. The 600 kHz signal, along with any audio information it contains, is translated to the IF frequency and is amplified and then detected.

Note that we could have also set the local oscillator to 600 – 455 kHz or 145 kHz. By setting it to the sum, we reduce the relative range that the oscillator must tune. To cover the 500 to 1700 kHz with the difference, our LO would have to cover from 45 to 1245, a 28:1 range. Using the sum requires LO coverage from 955 to 2155 kHz, a range of about 2.5:1 — much easier to implement.

Note that to detect standard AM signals, the receiver's second oscillator, the beat frequency oscillator (BFO), is turned off since the AM station provides its own carrier signal over the air. Receivers designed only for standard AM reception generally don't have a BFO at all.

It's not clear yet that we've gained anything by doing this; so let's look at another example. If we decide to change from listening to the station at 600 kHz and want to listen to another station at, say, 1560 kHz, we can tune the single dial of our superhet to 1560 kHz. The RF stage is tuned to 1560 kHz, the LO is set to 1560 + 455 or 2010 kHz, and now the desired station is translated to our 455 kHz IF where the bulk of our amplification can take place. Note also that with the superheterodyne configuration, selectivity (the ability to separate stations) occurs primarily in the intermediate-frequency (IF) stages and is thus the same no matter what frequencies we choose to listen to. This simplifies the design of each stage considerably.

12.3.1 Superhet Bandwidth

Now we will discuss the bandwidth requirements of different operating modes and how that affects superhet design. One advantage of a superhet is that the operating bandwidth can be established by the IF stages, and further limited by the audio system. It is thus independent of the RF frequency to which the receiver is tuned. It should not be surprising that the detailed design of a superhet receiver is dependent on the nature of the signal being received. We will briefly discuss the most commonly received modulation types and the bandwidth implications of each below. (Each modulation type is discussed in more detail in the **Modulation** or **Digital Modes** chapters.)

AMPLITUDE MODULATION (AM)

As shown in Fig 12.4, multiplying (in other words, modulating) a carrier with a single tone results in the tone being translated to frequencies of the sum and difference of the two. Thus, if a transmitter were to multiply a 600 Hz tone and a 600 kHz carrier signal, we would generate additional new frequencies at 599.4 and 600.6 kHz. If instead we were to modulate the 600 kHz carrier signal with a band of frequencies corresponding to human speech of 300 to 3300 Hz (the usual range of communication quality voice signals), we would have a pair of bands of information carrying waveforms extending from 596.7 to 603.3 kHz, as shown in **Fig 12.8**. These bands are called *sidebands,* and some form of these is present in any AM signal that is carrying information.

Note that the total bandwidth of this AM voice signal is twice the highest frequency transmitted, or 6600 Hz. If we choose to transmit speech and limited-range music, we might allow modulating frequencies up to 5000 Hz, resulting in a bandwidth of 10,000 Hz or 10 kHz. This is the standard channel spacing that commercial AM broadcasters use in the US. (9 kHz is used in Europe) In actual use, the adjacent channels are generally geographically separated, so broadcasters can extend some energy into the next channels for improved fidelity. We would refer to this as a *narrow-bandwidth* mode.

What does this say about the bandwidth needed for our receiver? If we want to receive the full information content transmitted by a US AM broadcast station, then we need to set the bandwidth to at least 10 kHz. What if our receiver has a narrower bandwidth? Well, we will lose the higher frequency components of

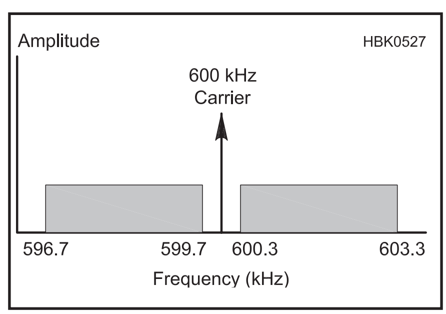


Fig 12.8 — Spectrum of sidebands of an AM voice signal sent on a 600 kHz carrier.

the transmitted signal — perhaps ending up with a radio suitable for voice but not very good at reproducing music.

On the other hand, what is the impact of having too wide a bandwidth in our receiver? In that case, we will be able to receive the full transmitted spectrum but we will also receive some of the adjacent channel information. This will sound like interference and reduce the quality of what we are receiving. If there are no adjacent channel stations, we will get any additional noise from the additional bandwidth and minimal additional information. The general rule is that the received bandwidth should be matched to the bandwidth of the signal we are trying to receive to maximize SNR and minimize interference.

As the receiver bandwidth is reduced, intelligibility suffers, although the SNR is improved. With the carrier centered in the receiver bandwidth, most voices are difficult to understand at bandwidths less than around 4 kHz. In cases of heavy interference, full carrier AM can be received as if it were SSB, as described below, with the carrier inserted at the receiver, and the receiver tuned to whichever sideband has the least interference.

SINGLE-SIDEBAND SUPPRESSED-CARRIER MODULATION (SSB)

In looking at Fig 12.8, you might have noticed that both sidebands carry the same information, and are thus redundant. In addition, the carrier itself conveys no information. It is thus possible to transmit a *single sideband* and *no carrier*, as shown in **Fig 12.9**, relying on the BFO (beat frequency oscillator) in the receiver to provide a signal with which to multiply the sideband in order to provide demodulated audio output. The implications in the receiver are that the bandwidth can be slightly less than half that required for double sideband AM (DSB). There must be an additional mechanism to carefully replace the missing carrier within the receiver. This is the function of the BFO, which must be at just exactly the right frequency. If the frequency is improperly set a baritone can come out sounding like a soprano and vice versa! The effect is quite audible even for small frequency errors of a few tens of Hz.

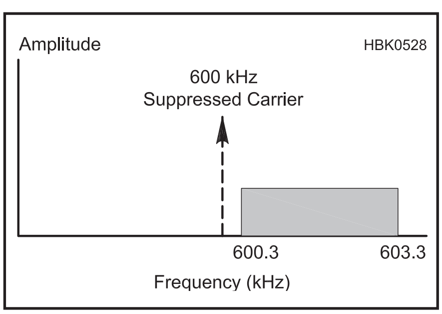


Fig 12.9 — Spectrum of single sideband AM voice signal sent adjacent to a suppressed 600 kHz carrier.

This results in a requirement for a much more stable receiver design with a much finer tuning system — a more expensive proposition. An alternate is to transmit a reduced level carrier and have the receiver lock on to the weak carrier, usually called a *pilot* carrier. Note that the pilot carrier need not be of sufficient amplitude to demodulate the signal, just enough to allow a BFO to lock to it. These alternatives are effective, but tend to make SSB receivers expensive, complex and most appropriate for the case in where a small number of receivers are listening to a single transmitter, as is the case of two-way amateur communication.

Note that the bandwidth required to effectively demodulate an SSB signal is actually less than half that required for the AM signal because the range centered on the AM carrier need not be received. Thus the communications-quality range of 300 to 3300 Hz can be received in a bandwidth of 3000 not 3300 Hz. Early SSB receivers typically used a bandwidth of around 3 kHz, but with the heavy interference frequently found in the amateur bands, it is more common for amateurs to use bandwidths of 1.8 to 2.4 kHz with the corresponding loss of some of the higher consonant sounds.

RADIOTELEGRAPHY (CW)

We have described radiotelegraphy as being transmitted by "on-off keying of a carrier." You might think that since a carrier takes up just a single frequency, the receive bandwidth needed should be almost zero. This is only true if the carrier is never turned on and off. In the case of CW, it will be turned on and off quite rapidly. The rise and fall of the carrier results in sidebands extending on either side of the carrier, and they must be received in order to reconstruct the signal in the receiver.

A rule of thumb is to consider the rise and fall time as about 10% of the pulse width and the bandwidth as the reciprocal of the quickest of rise or fall time. This results in a bandwidth requirement of about 50 to 200 Hz for the usual CW transmission rates. Another way to visualize this is with the bandwidth being set by a high-Q tuned circuit. Such a circuit will continue to "ring" after the input pulse is gone. Thus, too narrow a bandwidth will actually "fill in" between the code elements and act like a "no bandwidth" full period carrier and this is exactly what is heard if a very narrow crystal filter is used when receiving CW.

DATA SIGNALS

(This is a short overview of amateur data signals to establish receiver requirements for the digital modes. See the **Modulation** and **Digital Modes** chapters and the **Digital Communications** supplement on the *Handbook* CD for more in-depth treatment.)

The Baudot code (used for teletype communications) and ASCII code — two popular digital communications codes used by amateurs — are constructed with sequences of elements or bits. The state of each bit — ON or OFF — is represented by a signal at one of two distinct frequencies: one designated *mark* and one designated *space*. This is referred to as *frequency shift keying* (FSK). The transmitter frequency shifts back and forth with each character's individual elements.

Amateur Radio operators typically use a 170 Hz separation between the mark and space frequencies, depending on the data rate and local convention, although 850 Hz is sometimes used. The minimum bandwidth required to recover the data is somewhat greater than twice the spacing between the tones. Note that the tones can be generated by directly shifting the carrier frequency (direct FSK), or by using a pair of 170 Hz spaced audio tones applied to the audio input of an SSB transmitter (audio FSK or AFSK). Direct FSK and AFSK sound the same to a receiver.

Note that if the standard audio tones of 2125 Hz (mark) and 2295 Hz (space) are used, they fit within the bandwidth of a voice channel and thus a voice transmitter and receiver can be employed without any additional processing needed outside the radio equipment. Alternately, the receiver can employ detectors for each frequency and provide an output directly to a computer.

If the receiver can shift its BFO frequency appropriately, the two tones can be received through a filter designed for CW reception with a bandwidth of about 300 Hz or wider. Some receivers provide such a narrow filter with the center frequency shifted midway between the tones (2210 Hz) to avoid the need for retuning. The most advanced receivers provide a separate filter for mark and space frequencies, thus maximizing interference rejection and signal-to-noise ratio (SNR). Using a pair of tones for FSK or AFSK results in a maximum data rate of about 1200 bit/s over a high-quality voice channel.

Phase shift keying (PSK) can also be used to transmit bit sequences, requiring good frequency stability to maintain the required time synchronization to detect shifts in phase. If the channel has a high SNR, as is often the case at VHF and higher, telephone network data-modem techniques can be used.

At HF, the signal is subjected to phase and amplitude distortion as it travels. Noise is also substantially higher on the HF bands. Under these conditions, modulation and demodulation techniques designed for "wire-line" connections become unusable at bit rates of more than a few hundred bps. As a result, amateurs have begun adopting and developing state of the art digital modulation techniques. These include the use of multiple carriers (MFSK, Clover, PACTOR III, etc.), multiple amplitudes and phase shifts (QAM and QPSK techniques), and advanced error detection and correction methods to achieve a data throughput as high as 3600 bits per second (bps) over a voice-bandwidth channel. Newly developed coding methods for digital voice using QPSK modulation result in a signal bandwidth of less than 1200 Hz. (See **http://freedv.org** for more information.) Spread-spectrum techniques are also being adopted on the amateur UHF bands, but are beyond the scope of this discussion.

The bandwidth required for data communications can be as low as 100 Hz for PSK31 to 1 kHz or more for the faster speeds of PACTOR III and Clover. Beyond having sufficient bandwidth for the data signal, the primary requirements for receivers used for data communications are linear amplitude and phase response over the bandwidth of the data signal to avoid distorting these critical signal characteristics. The receiver must also have excellent frequency stability to avoid drift and frequency resolution to enable the receiver filters to be set on frequency.

FREQUENCY MODULATION (FM)

Another popular voice mode is frequency modulation or FM. FM can be found in a number of variations depending on purpose. In Amateur Radio and commercial mobile communication use on the shortwave bands, it is universally *narrow band FM* or *NBFM*. In NBFM, the frequency deviation is limited to around the maximum modulating frequency, typically 3 kHz. The bandwidth requirements at the receiver can be approximated by Carson's Rule of BW = 2 × (D + M), where D is the deviation and M is the maximum modulating frequency. Thus 3 kHz deviation and a maximum voice frequency of 3 kHz results in a bandwidth of 12 kHz, not far beyond the requirements for broadcast AM. (Additional signal components extend beyond this bandwidth, but are not required for voice communications.)

In contrast, broadcast or *wideband FM* or *WBFM* occupies a channel width of 150 kHz. Originally, this provided for a higher modulation index, even with 15 kHz audio that resulted in an improved SNR. However, with multiple channel stereo and sub-channels all in the same allocated bandwidth the deviation is around the maximum transmitted signal bandwidth.

In the US, FCC amateur rules limit wideband FM use to frequencies above 29 MHz. Some, but not all, HF communication receivers provide for FM reception. For proper FM reception, two changes are required in the receiver architecture as shown within the dashed line in **Fig 12.10**. The fundamental change is that the detector must recover information from the frequency variations of the input signal. The most common such detector is called a *discriminator*. The discriminator does not require a BFO, so that is turned off, or eliminated in a dedicated FM receiver. Since amplitude variations convey no information in FM, they are generally eliminated by a *limiter*. The limiter is a high-gain IF amplifier stage that clips the positive and negative peaks of signals above a certain threshold. Since most noise of natural origins is amplitude modulated, the limiting process also strips away noise from the signal.

SUMMARY OF RECEIVER BANDWIDTH REQUIREMENTS

We now have briefly discussed the typical operating modes expected to be encountered by an HF communications receiver. These are tabulated in **Table 12.1** and will be used as design requirements as we develop the various receiver architectures.

Table 12.1
Typical Communications Bandwidths for Various Operating Modes

Mode	*Bandwidth (kHz)*
FM Voice	15
AM Broadcast	10
AM Voice	4-6.6
SSB Voice	1.8-3
Digital Voice	1.0-1.2
RTTY (850 Hz shift)	0.3-1.0
CW	0.1-0.5

12.3.2 Selection of IF Frequency for a Superhet

Now that we have established the range of bandwidths that our receiver will need to pass, we are in a position to discuss the selection of the IF frequency at which those bandwidths will be established.

IF IMAGE RESPONSE

As noted earlier, a superhet with a single local oscillator or LO and specified IF can receive two frequencies, selected by the tuning of the RF stage. For example, using a receiver with an IF of 455 kHz to listen to a desired signal at 7000 kHz can use an LO of 7455 kHz. However, the receiver will also receive a signal at 455 kHz above the LO frequency, or 7910 kHz. This undesired signal frequency, located at twice the IF frequency from the desired signal, is called an *image*.

Images will be separated from the desired frequency by twice the IF and the filter ahead of the associated mixer must reduce the image signal by the amount of the required *image rejection*. For a given IF, this gets more difficult as the received frequency is increased. For example, with a 455 kHz single conversion system tuned to 1 MHz, the image will be at 1.91 MHz, almost a 2:1 frequency ratio and relatively easy to reject with a filter. The same receiver tuned to 30 MHz, would have an image at 30.91 MHz,

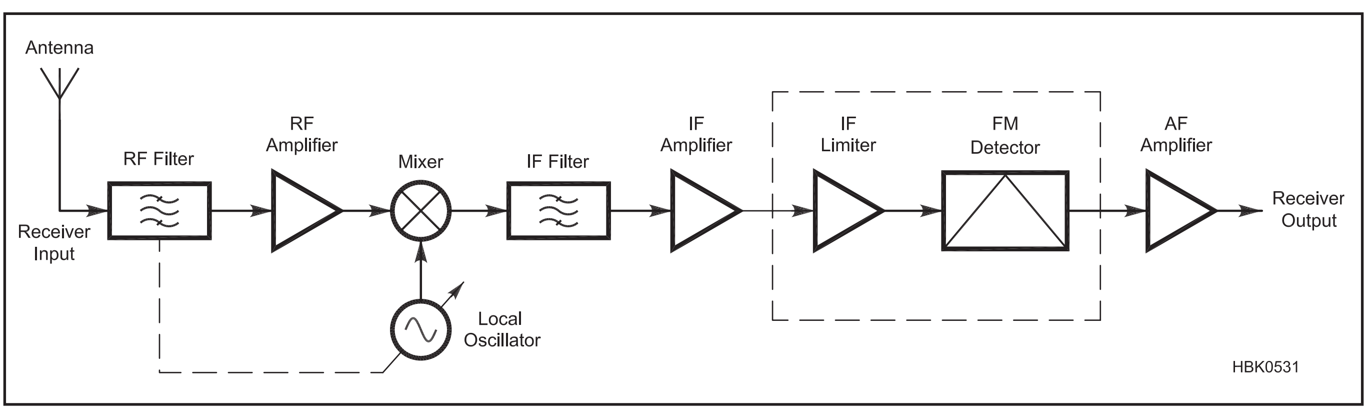


Fig 12.10 — Block diagram of an FM superhet. Changes are in dashed box.

a much more difficult filtering problem

While an image that falls on an occupied channel is obviously a problem — it's rarely desirable to receive two signals at the same time — problems occur even if the image frequency is clear of signals. This is because the atmospheric noise in the image bandwidth is added to the noise of the desired channel, as well as any internally-generated noise in the RF amplifier stage. If the image response is at the same level as the desired signal response, there will be a 3 dB reduction in SNR.

REDUCING IMAGES WITH A HIGHER IF FREQUENCY

An obvious solution to the RF image response is to raise the IF frequency high enough so that signals at twice the IF frequency from the desired signal are sufficiently attenuated by filters ahead of the mixer. This can easily be done with IF stages operating at 5 to 10% of the highest receiving frequency (1.5 to 3 MHz for a receiver that covers the 3-30 MHz HF band). The concept is used at higher frequencies as well. The FM broadcast band (150 kHz wide channels over 87.9 to 108 MHz in the US) is generally received on superhet receivers with an IF of 10.7 MHz, which places all image frequencies outside the FM band, eliminating interference from other FM stations.

The use of higher-frequency tuned cir-

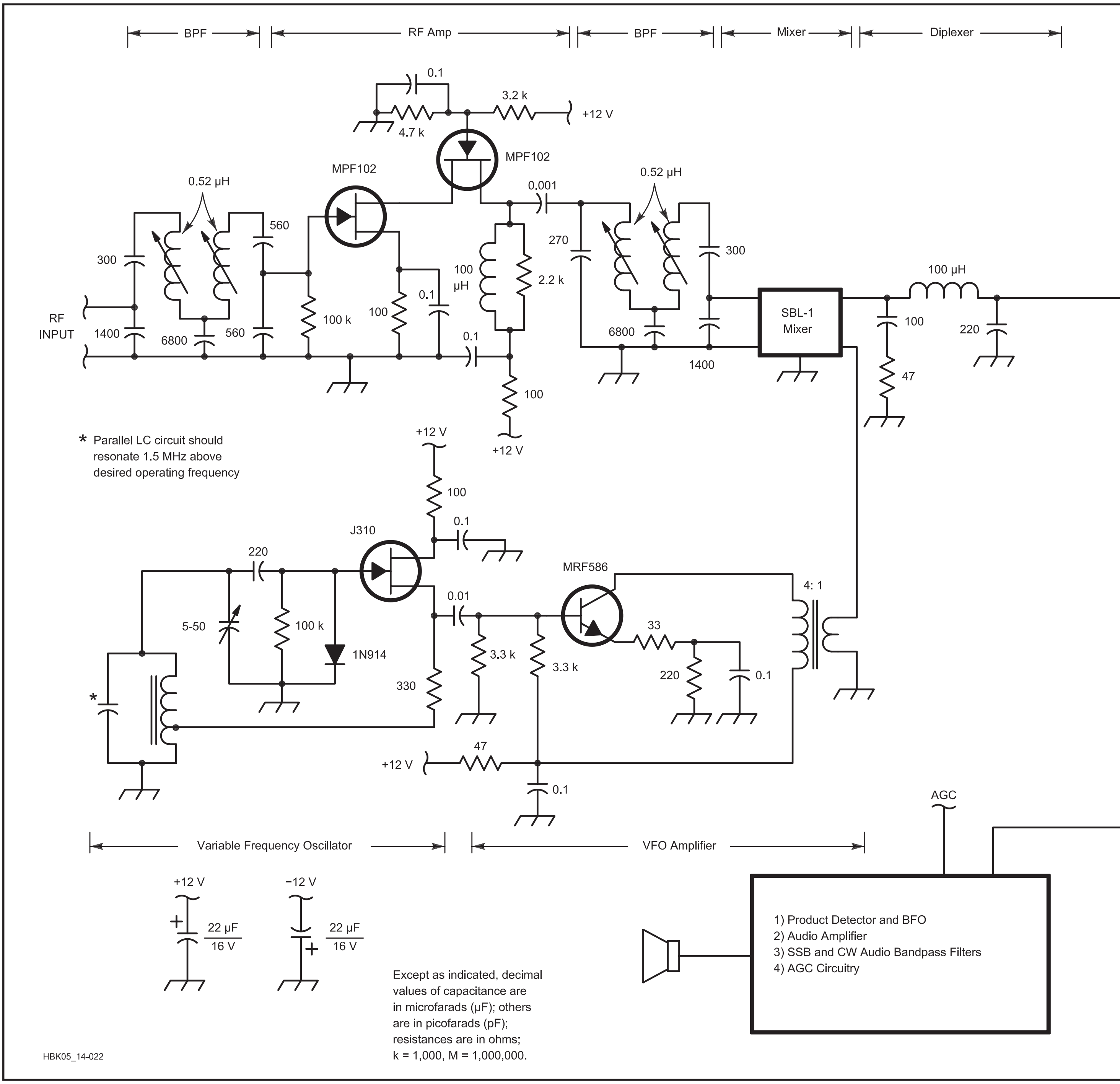


Fig 12.11 — Schematic of representative superhet using crystal lattice filters to achieve selectivity with a high IF frequency.

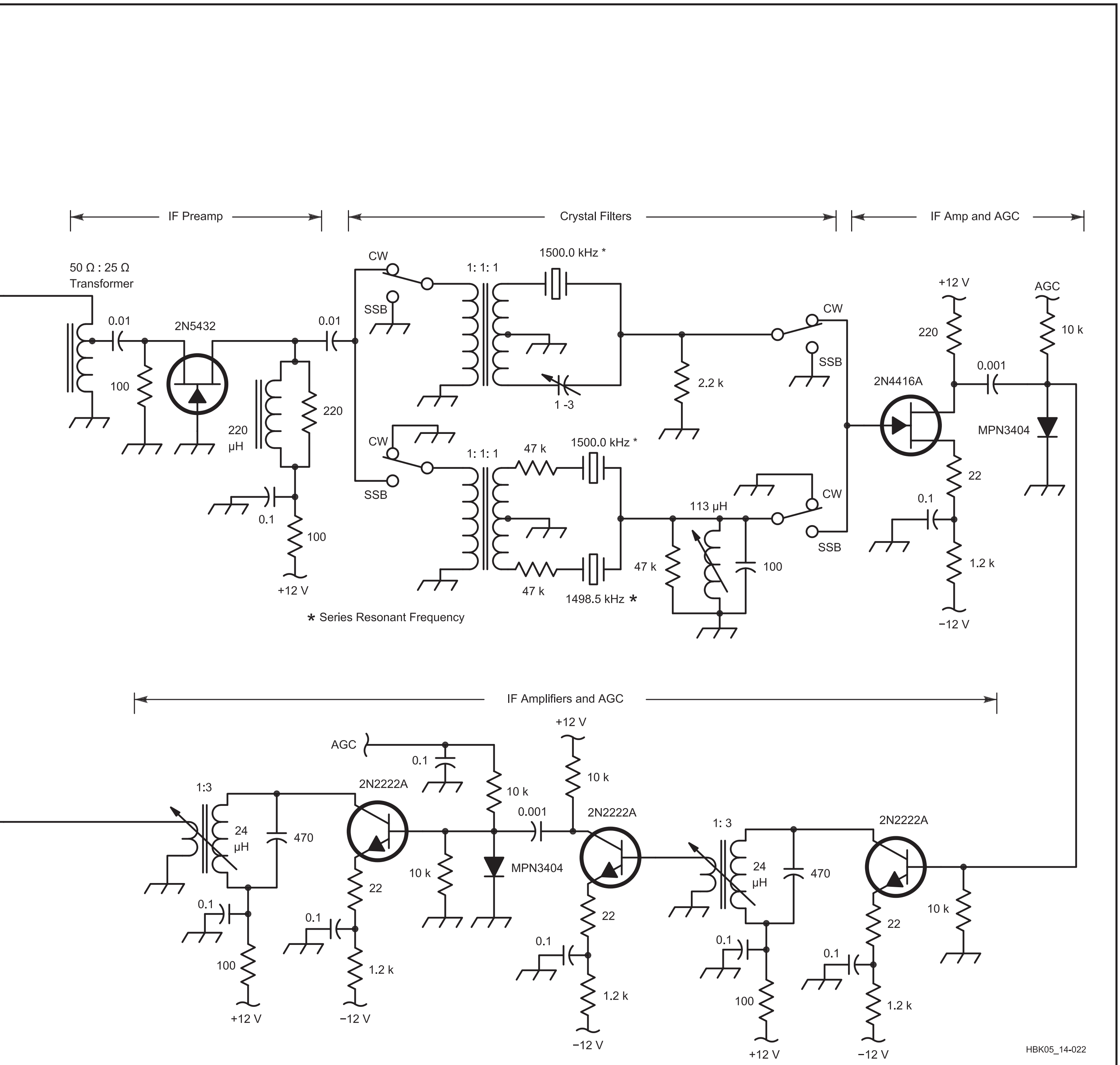
cuits for IF selectivity works well for the 150 kHz wide FM broadcast channels, but not so well for the relatively narrow channels encountered on HF or lower, or even for many V/UHF narrowband services. Fortunately, there are three solutions that were commonly used to resolve this problem.

The first, *double conversion*, converted the desired signal to a relatively high IF followed by a second conversion to a lower IF to set the selectivity. This was a popular technique in the 1950s. Improvements on the double conversion technique led to *triple conversion* with a very low, highly selective third IF. The *Collins system* of moving a single-range VFO to the second mixer and using switchable crystal oscillators for the first mixer also became popular. The *pre-mixed* arrangement, a third approach to double conversion was a combination of the two, used a single variable oscillator range, as with the Collins, but mixes the VFO and LO before applying them to the first mixer – outside of the signal path.

These methods are obsolete for current receivers but are commonly encountered in the vintage equipment popular with many hams. They are discussed in previous editions of the *Handbook*.

High Frequency Crystal Lattice Filters

Commercial quartz-crystal filters with bandwidths appropriate for CW and SSB



became available in the 1970s with center frequencies into the 10 MHz range. This allowed a single-conversion receiver (see Fig 12.2) with an IF in the HF range to provide both high image rejection and needed channel selectivity. This single-conversion architecture remains popular among designers of portable and low-power equipment. Crystal filter design is discussed in the **RF and AF Filters** chapter.

Fig 12.11 shows an example front-end and IF sections of a single-conversion superhet using simple filters centered 1500 kHz. While the filters shown are actually buildable by amateurs at low cost, multiple-section filters with much better performance can be purchased or constructed. Other IF frequencies can be used, depending on crystal or filter availability.

The circuit shown demonstrates the concepts involved and can be reproduced at low cost. Remaining receiver functional blocks such as the AGC circuitry, detectors and BFO, and audio amplifiers and filters can be found elsewhere in the book.

The Image-Rejecting Mixer

Another technique for reduction of image response in receivers is not as commonly encountered in HF receivers as the preceding designs, but it deserves mention because it has some very significant applications. The *image-rejecting* mixer requires phase-shift networks, as shown in **Fig 12.12**. Frequency F_1 represents the input frequency while F_2 is that of the local oscillator. Note that the two 90° phase shifts are applied at different frequencies. The phase shift network following Mixer 1 is at a fixed center frequency corresponding to the IF, while the phase shift network at F_2 must provide the required phase shift as the local oscillator tunes across the band.

If the local oscillator is required to tune over a limited fractional frequency range, this is a very feasible approach. On the other hand, maintaining a 90° phase shift over a wide range can be tricky. The good news is that this approach provides image reduction that is independent of, and in addition to, any other mechanisms such as filters that are employed toward that end.

Additionally, with the ability of DSP components to operate at higher and higher frequencies, the necessary operations seen in Fig 12.12 can be performed in software which does not depend on precision hardware design to maintain nearly exact phase relationships.

The image-rejecting mixing process has several attractive features:

- It is the only way to provide "single signal" reception with a direct conversion receiver, effectively reducing the audio image. This can make the DC receiver a very good performer, although the added complexity is not always warranted in typical amateur DC applications.
- This option is frequently found in microwave receivers in which sufficiently selective RF filtering can be difficult to obtain. Since they often operate on fixed frequencies, maintaining the required phase shift can be straightforward.
- It is found in advanced receivers that are trying to achieve optimum performance. Even with a high first IF frequency, additional image rejection can be provided.
- In transmitters, the same system is called the *phasing method* of SSB generation. The same blocks run "backwards" — one of the phase shift networks can be applied to the speech band and used to cancel one sideband. This is discussed in the **Transmitters and Transceivers** chapter.

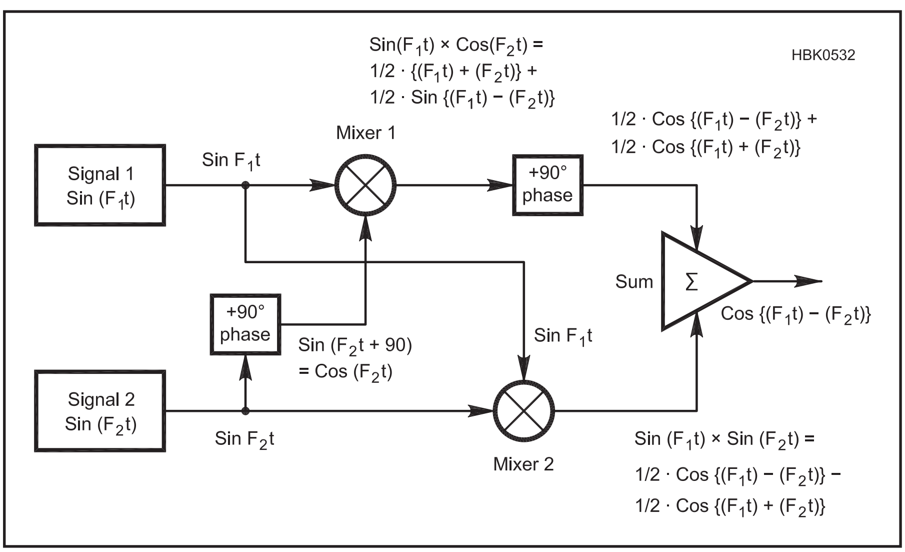


Fig 12.12 — Image rejecting mixer, block diagram and signal relationships.

12.3.3 Multiple-Conversion in the Digital Age

While advances in microminiaturization of all circuit elements have had a radical change in the dimensions of communication equipment, perhaps most significant technology impacts on architecture come in two areas — the application of digital signal processing and direct digital synthesis frequency generation. Both are discussed in detail in the chapter on **DSP and Software Radio Design**.

Digital signal processing provides a level of bandwidth setting filter performance not practical with other technologies. While much better than most low frequency IF LC bandwidth filters, the very good crystal or mechanical bandwidth filters in amateur gear are not very close to the rectangular shaped frequency response of an ideal filter, but rather have skirts with a 6 to 60 dB response of perhaps 1.4 to 1. That means if we select an SSB filter with a nominal (6 dB) bandwidth of 2400 Hz, the width at 60 dB down will typically be 2400 × 1.4 or 3360 Hz. Thus a signal in the next channel that is 60 dB stronger than the signal we are trying to copy (as often happens) will have energy just as strong as our desired signal.

DSP filtering approaches the ideal response. **Fig 12.13** shows the ARRL Lab measured response of a DSP bandwidth filter with a 6 dB bandwidth of 2400 Hz. Note how rapidly the skirts drop to the noise level. In addition, while analog filtering generally requires a separate filter assembly for each desired bandwidth, DSP filtering is adjustable — often in steps as narrow as 50 Hz — in both bandwidth and center frequency. In addition to bandwidth filtering, the same DSP can often provide digital noise reduction and digital notch filtering to remove interference from fixed frequency carriers.

The application of DSP filters at higher and higher frequencies as analog-to-digital conversion samples rates and processing power increase, offers greatly improved per-

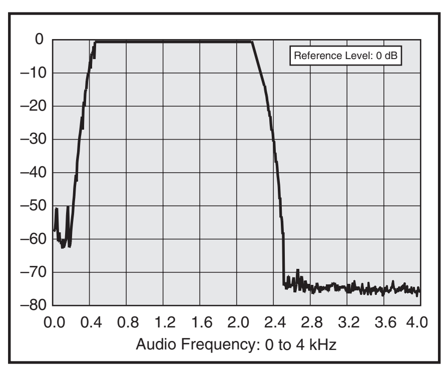


Fig 12.13 — ARRL Lab measured response of an aftermarket 2400 Hz DSP bandwidth filter.

formance for multiple-conversion receivers. The state of the art for amateur receivers is approaching direct sampling (digitization at the signal frequency) which will eliminate the need for separate IF filters, amplifiers, and detectors as all of those functions will be performed inside the DSP processor as software operations. However, it is likely that analog receivers will continue to be popular in various applications, such as portable, low-power, and home-built gear.

Project: The MicroR2 — An Easy to Build SSB or CW Receiver

The following is the description of a direct conversion receiver that uses an image-rejecting mixer as a product detector for single sideband or single-signal CW reception.

Throughout the 100-year history of Amateur Radio, the receiver has been the basic element of the amateur station. This single PC board receiver, originally described by Rick Campbell, KK7B, captures some of the elegance of the classic all-analog projects of the past while achieving the performance needed for operation in 21st-century band conditions across a selected segment of a single band.

A major advantage of a station with a separate receiver and transmitter is that each can be used to test the other. There is a particularly elegant simplicity to the combination of a crystal controlled transmitter and simple receiver. The receiver doesn't need a well-calibrated dial, because the transmitter is used to spot the frequency. The transmitter doesn't need sidetone, VFO offset or other circuitry because the crystal determines the transmitted frequency and the receiver is used to monitor the transmit note off the air. Even the receiver frequency stability may be relaxed, because there is no on-the-air penalty for touching up receiver tuning during a contact.

Let's dive into the design and construction of a simple receiver (**Fig 12.14**) that may be used as a companion for an SSB or CW transmitter on 40 meters, or as a tunable IF for higher bands. **Fig 12.15** is the block diagram of this little receiver, which can be built in a few evenings. The block diagram is similar to the simple direct conversion receivers of the early 1970s, but there the similarity ends. This radio takes advantage of 40 years of evolution of direct conversion HF receivers by some very talented designers.

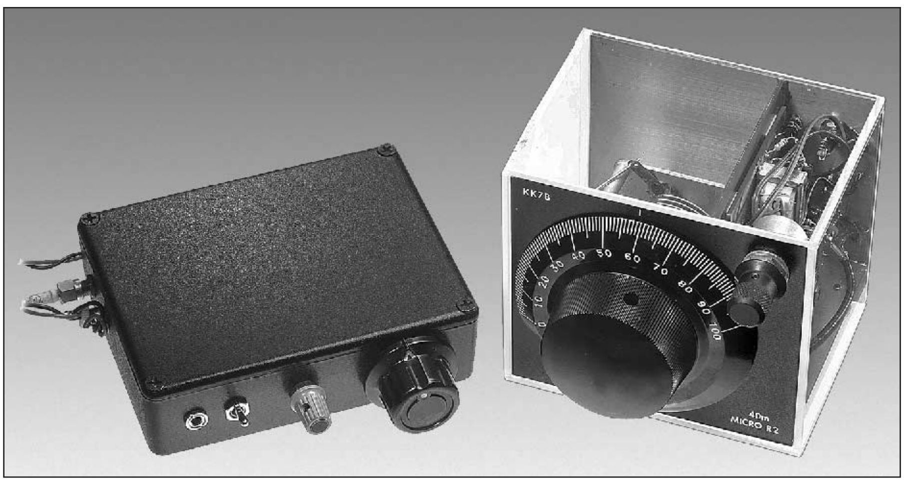


Fig 12.14 — Two versions of the MicroR2. For smooth tuning, it's hard to beat a big vernier dial!

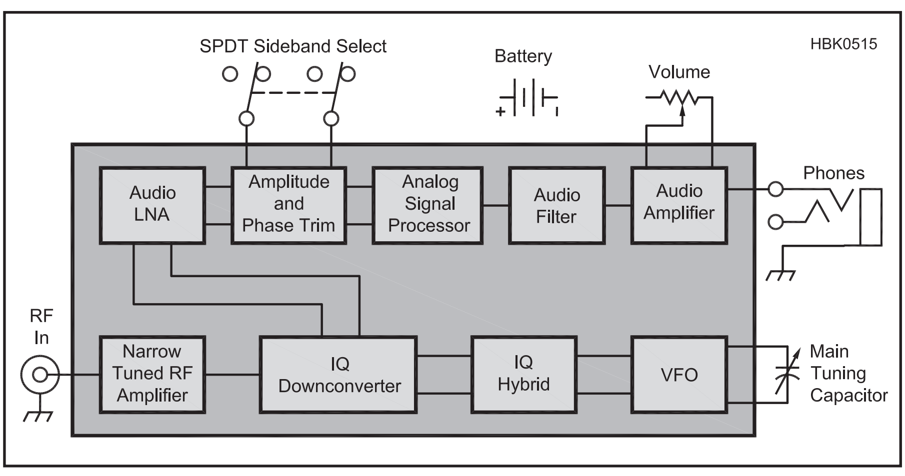


Fig 12.15 — Block diagram of the MicroR2 SSB/CW receiver.

"SINGLE SIGNAL" RECEPTION MAKES ALL THE DIFFERENCE

One major difference from early direct conversion receivers is the use of phasing to eliminate the opposite sideband. The receiver in Fig 12.15 is a "single-signal" receiver that only responds to signals on one side of zero beat. Basic receivers in the '50s and '60s would hear every CW signal at two places on the dial, on either side of zero beat. As you tuned through a CW signal, first you heard a high-pitched tone, and then progressively lower audio frequencies until the individual beats became audible, and then finally all the way down to zero cycles per second: zero beat. As you tuned further you heard the pitch rise back up until the high frequency CW signal became inaudible again at some high pitch.

It's not too terrible to hear the desired signal at two places on the dial — but having twice as much interference and noise is inconvenient at best. Most modern radios, even simple ones, include a simple crystal or mechanical filter for single-signal IF selectivity. Since a direct conversion receiver has no IF, selectivity is obtained by a combination of audio filtering and an image-rejecting mixer. Chapter 9 of *Experimental Methods in RF Design* (*EMRFD)* has a complete discussion of image-reject or "phasing" receivers, and an overview of both conventional superheterodyne and phasing receivers can be found elsewhere in this chapter.

BLOCK BY BLOCK DESCRIPTION

A brief description of each block in Fig 12.15 and the schematic in **Fig 12.16** follows. A parts list is given in **Table 12.2**. (Additional detail and rationale for the circuits is provided in *EMRFD*.)

The 50 Ω RF input connects to a narrow tuned-RF amplifier. The first circuit elements are a low-pass filter to prevent strong signals above the receiver tuning range — primarily local TV and FM broadcast stations — from arriving at the mixers where they would be down-converted to audio by harmonics of the local oscillator. The common-gate JFET low-noise amplifier (LNA) sets the noise figure of the receiver and provides some gain, but even more importantly, it isolates the mixers from the antenna. Isolation prevents the impedance at the RF input from affecting the opposite sideband suppression, and reverse isolation prevents local oscillator leakage from radiating out the antenna and causing interference to others as well as tunable hum.

The tuned drain circuit provides additional selectivity and greatly improves dynamic range for signals more than a few hundred kHz from the desired signal. Lifting the source of the JFET allows a convenient mute control — changing the LNA from a 10 dB amplifier to a 40 dB attenuator.

Following the LNA are the I-Q (short for "in-phase" and "quadrature," the labels we use for the two signal paths in a phasing system) down-converter and local oscillator (LO) I-Q hybrid. Many of the more subtle difficulties of phasing direct conversion reeivers are avoided by a compact, symmetrical layout and direct connection of all the mixer ports to the appropriate circuit elements without the use of transmission lines. The audio low-pass filters ensure that only audio exits the down-converter block, and set the close-in dynamic range of the receiver. The inductors in the low-pass filters will pick up magnetic hum from nearby power transformers — more on this topic later.

To the right of the I-Q hybrid is a basic JFET Hartley VFO. In this application, it is stripped of all the usual accoutrements like buffer amplifiers and receiver incremental tuning (RIT). The drain resistor and diodes in the mixers set the operating level. This simplified configuration, with a link on the VFO inductor directly driving the twisted-wire hybrid, works very well over a 5% bandwidth (350 kHz at 7 MHz) when the quadrature hybrid is optimized at the center

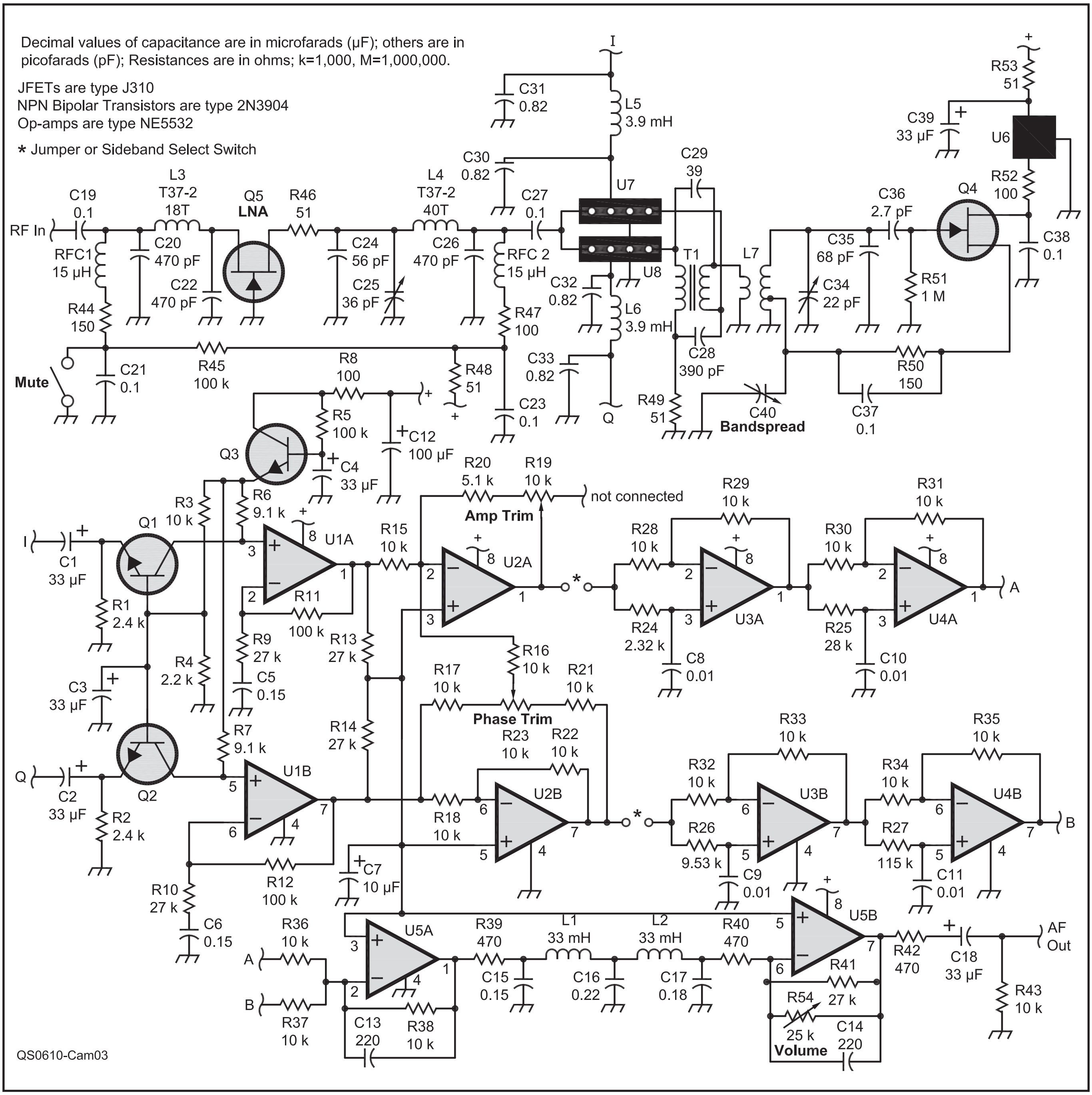


Fig 12.16 — Schematic diagram of the MicroR2 SSB/CW receiver. The parts are listed in Table 12.3.

frequency and all voltages except the LO are less than a few millivolts. (See R. Fisher, W2CQH, "Twisted-Wire Quadrature Hybrid Directional Couplers," in *QST* for Jan 1978.) The top half of the block diagram operates entirely at audio. The left block is a matched pair of audio LNAs with a noise figure of about 5 dB and near 50 Ω input impedance, to properly terminate the low-pass filters at the I-Q mixer IF ports. These are directly coupled into a circuit that performs the mathematical operations needed to remove amplitude and phase errors from the I and Q signal paths. This is a classic use of operational amplifiers to do basic math.

The next block is a pair of second-order all-pass networks. Use of second order networks limits the opposite sideband suppression to 37 dB, which sounds exceptionally clean with multiple CW signals and static crashes in the passband.

It is straightforward to improve the opposite sideband suppression by 10 dB or so, but doing so requires much more circuitry, not just a third-order all-pass network. Relaxing the opposite sideband suppression to a level comparable with the Drake 2B permitted cutting the parts count in half, and the construction and alignment time by a factor of 10.

The final blocks are the summer, audio low-pass filter and headphone amplifier. These are conventional, near duplicates of those used 10 years ago. When the board is finished it is an operating receiver. The completed circuitry is shown in **Fig 12.17**.

Table 12.2
MicroR2 Parts List
Parts for CW and SSB Version

C1-C4, C18, C39 — 33 µF, 16 V electrolytic.
C5, C6 — 0.15 µF, 5% polyester.
C7 — 10 µF, 16 V electrolytic.
C8-C11 — 0.01 µF, polyester matched to 1%.
C12 — 100 µF, 16 V electrolytic.
C13, C14 — 220 pF, NP0 ceramic.
C15 — 0.15 µF, 5% polyester, see Note A.
C16 — 0.22 µF, 5% polyester, see Note A.
C17 — 0.18 µF, 5% polyester, see Note A.
C19, C21, C23, C27, C37, C38 — 0.1 µF, 5% polyester.
C20, C22, C26 — 470 pF, NP0 ceramic.
C24 — 56 pF, NP0 ceramic.
C25 — 3-36 pF, poly trimmer capacitor.
C28 — 390 pF, NP0 ceramic.
C29 — 39 pF, NP0 ceramic on back of board, see Note B.
C30-C33 — 0.82 µF, 5% polyester, see Note C.
C34 — 2.2-22 pF, poly trimmer capacitor.
C35 — 68 pF, NP0 ceramic.
C36 — 2.7 pF 5% NP0 1000V ceramic.
C40 — Variable capacitor, see Note D.
L1, L2 — 33 mH inductor, see Note A.
L3 — 18t #28 T37-2, see Note E.
L4 — 40t #30 T37-2, see Note E.
L5, L6 — 3.9 mH inductor, see Note C.
L7 — 36t #28 T50-6, tap at 8 turns with a 2 turn link, see Note E.
Q1-Q3 — 2N3904.
Q4, Q5 — J310.
R1, R2 — 2.4 kΩ.
R3, R15-R18, R21, R22, R38, R43 — 10 kΩ.
R4 — 2.2 kΩ.
R5, R11, R12, R45 — 100 kΩ.
R6, R7 — 9.1 kΩ.
R8, R47, R52 — 100 Ω.
R9, R10, R13, R14, R41 — 27 kΩ.
R19, R23 — 10 kΩ trimpot.
R20 — 5.1 kΩ.
R24 — 2.32 kΩ, 1% see Note A.
R25 — 28.0 kΩ, 1% see Note A.
R26 — 9.53 kΩ, 1% see Note A.
R27 — 115 kΩ 1% see Note A.
R28-R37 — 10.0 kΩ 1%.
R39, R40, R42 — 470 Ω.
R44, R50 — 150 Ω.
R46, R48, R49, R53 — 51 Ω.
R51 — 1 MΩ.
R54 — 25 kΩ volume control.
RFC1,RFC2 — 15 µH molded RF choke.
T1 — 17 t two colors #28 bifilar T37-2, see Note E.
U1-U5 — NE5532 or equivalent dual low-noise high-output op-amp.
U6 — LM7806 or equivalent 6 V three terminal regulator.
U7, U8 — Mini-Circuits TUF-3 diode ring mixer.

Note A: Make the following parts substitutions for a CW only version:
R24 — 4.75 kΩ, 1%.
R25 — 41.2 kΩ, 1%.
R26 — 16.9 kΩ, 1%.
R27 — 147 kΩ, 1%.
C15 — 0.47 µF, 5% polyester.
C16 — 0.68 µF, 5% polyester.
C17 — 0.56 µF, 5% polyester.
L1, L2 — 100 mH inductor.

Note B: The total reactance of the parallel combination of C28 and C29 plus the capacitance between the windings of T1 is $-j50$ Ω at the center of the tuning range. Placing most of the capacitance at one end is a different but equivalent arrangement to the quadrature hybrid we often use with equal capacitors. C29 is only needed if there is no standard value for C28 within a few % of the required value. C29 is tack soldered to the pads provided on the back of the PC board, and may be a surface mount component if desired.

Note C: To sacrifice close-in dynamic range and selectivity for reduced 60 Hz hum pickup, make the following substitutions, as illustrated in Fig 12.18. This modification is recommended if the MicroR2 must be used near 60 Hz power transformers.
C30, C32 — 0.10 µF 5% polyester.
C31, C33 — Not used.
L5, L6 — Not used. Place wire jumper between pads.

Note D: Capacitor C40 is the tuning capacitor for the receiver. The MicroR2 was specifically designed to use an off-board tuning capacitor to provide mounting flexibility, and to encourage the substitution of whatever high-quality dual bearing capacitor and reduction drive may be in the individual builder's junk box. Any variable capacitor may be used, but values around 100 pF provide a little more than 100 kHz of tuning range on the 40 meter band, which is a practical maximum. The MicroR2 needs to be realigned (RF amplifier peaked and the amplitude and phase trimpots readjusted) when making frequency excursions of more than about ±50 kHz on the 40 meter band. To prevent tunable hum and other common ills of direct conversion receivers, the MicroR2 PC board and tuning capacitor should be in a shielded enclosure. See Chapter 8 of *EMRFD* for a complete discussion.

Note E: L3, L4, L7 and T1 are listed as number of turns on the specified core rather than a specific inductance. For those who wish to study the design with a calculator, simulator and inductance meter, L3 should be about $+j100$ Ω at about 8 MHz (not particularly critical). L4 and L7 should be about $+j250$ Ω at mid band. Each winding of T1 should be $+j50$ Ω at mid band.

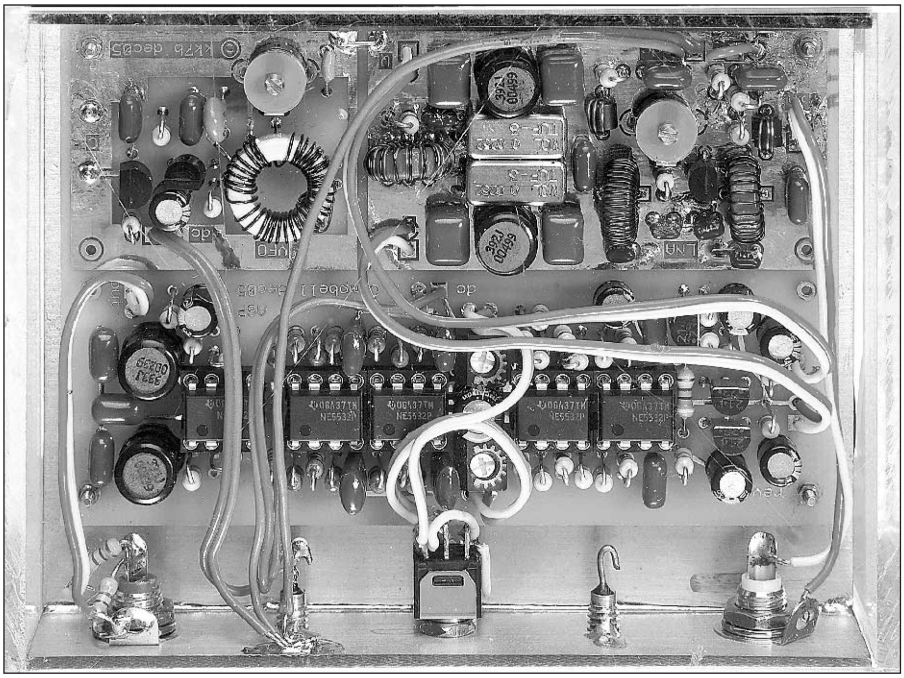


Fig 12.17 — Photo of the inside of the receiver. The PC board and parts are available from Kanga US (www.kangaus.com).

Table 12.3
MicroR2 Receiver Performance

Parameter	*Measured in the ARRL Lab*
Frequency coverage:	7.0-7.1 or 7.2-7.3 MHz.
Power requirement:	50 mA, tested at 13.8 V.
Modes of operation:	SSB, CW.
SSB/CW sensitivity:	–131 dBm.
Opposite sideband rejection:	1000 Hz, 42 dB
Two-tone, 3rd-order IMD dynamic range:	20 kHz spacing, 81 dB
Third-order intercept:	20 kHz spacing, –5 dBm.
Third-order intercept points were determined using equivalent S5 reference.	

ALIGNMENT

The receiver PC board may be aligned on the bench without connecting the bandspread capacitor. There are only four adjustments, and the complete alignment takes just a few minutes with just an antenna, the crystal oscillator in the companion transmitter (or a test oscillator) and headphones. First, center the trimming potentiometers using the dots. Then slowly tune the VFO variable capacitor until the crystal oscillator is heard. The signal should be louder on one side of zero beat. Peak the LNA tuning carefully for the loudest signal. Then tune to the weaker side of zero beat and null the signal with the amplitude and phase trim pots. That's all there is.

If the adjustments don't work as expected or there is no opposite sideband suppression, look for a construction error. Don't forget to jumper the MUTE connection on the LNA — it is a "ground- to-receive" terminal just like simple receivers from the 1960s. Once the receiver PC board is working on the bench, mount it in a box, attach the bandspread capacitor (with two wires — the ground connection needs to be soldered to the PC board too). Reset the VFO tuning capacitor for the desired tuning range, and do a final alignment in the center of the desired frequency range.

The simple on-board VFO is stable enough for casual listening and tuning around the band, but it can be greatly improved by eliminating the film trimmer capacitor that sets the VFO tuning range. It may be replaced by an air trimmer, a piston trimmer, a small compression trimmer, or a combination of small fixed NP0 ceramic capacitors experimentally selected to obtain the correct tuning range. The holes in the PC board are large enough to easily remove the film trimmer capacitor with solder wick. The VFO stability may be improved still further by following the thermal procedure described by W7ZOI in *EMRFD*.

PERFORMANCE

Is it working? I don't hear anything! Most of us are familiar with receivers with a lot of extra gain and noise, and AGC systems that turn the gain all the way up when there are no signals in the passband. Imagine instead a perfect receiver with no internal noise. With no antenna connected, you would hear absolutely nothing, just like a CD player with no disk. The MicroR2 is far from perfect — but it may be much quieter than your usual 40 meter receiver, and it probably has less gain. It is designed to drive good headphones. Weak signals will be weak and clear and strong signals strong and clear over the entire 70 dB in-channel dynamic range — just like a CD player. This receiver rewards listening skill and a good antenna — don't expect to hear much on a couple feet of wire in a noisy room. The measured performance is shown in **Table 12.3**.

Now, about that hum. One reason this receiver sounds so good is that the selectivity is established right at the mixer IF ports, before any audio gain. Then, after much of the system gain, a second low-pass filter provides additional selectivity. Distributed selectivity is a classic technique. Earlier designs included aggressive high-pass filtering to suppress frequencies below 300 Hz. That works well and has other advantages — but the audio sounds a bit thin. K7XNK suggested trying a receiver with the low frequency response wide open. It sounds great, but 60 Hz hum is audible if the receiver is within about a meter of a power transformer. Any closer and the hum gets loud. This makes it difficult to measure receiver performance in the lab, as there is a power transformer in every piece of test equipment.

Hum pickup may be eliminated by modifying the circuit as shown in **Fig 12.18**. This hurts performance in the presence of strong off-channel signals, but in some applications — an instrumentation receiver on the bench — it is a good trade-off. The receiver sounds the same either way. For battery operation in the hills, or on a picnic table in the backyard, hum pickup isn't a problem.

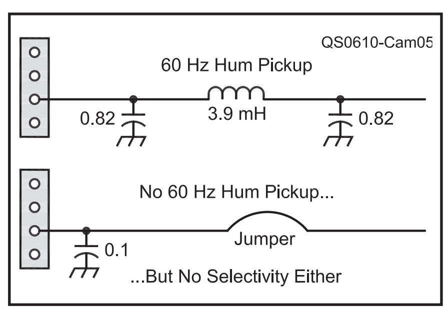


Fig 12.18 — Modified circuit to minimize hum.

12.4 Superhet Receiver Design Details

The previous sections have provided an overview of the major architectural issues involved in superheterodyne receiver design. In this section, we will cover more of the details for the superhet constructed as a hybrid of analog electronics and DSP functions applied in an IF stage, the primary receiver design architecture used in Amateur Radio today.

Table 12.4
Typical Noise Levels (Into the Receiver) and Their Source, by Frequency

Frequency range	*Dominant noise sources*	*Typical level (μv/m)**
LF 30 to 300 kHz	atmospheric	150
MF 300 to 3000 kHz	atmospheric/man-made	70
Low HF 3 to 10 MHz	man-made/atmospheric	20
High HF 10 to 30 MHz	man-made/thermal	10
VHF 30 to 300 MHz	thermal/galactic	0.3
UHF 300 to 3000 MHz	galactic/ thermal	0.2

*The level assumes a 10 kHz bandwidth. Data from *Reference Data for Engineers*, 4th Ed, p 273, Fig 1.

12.4.1 Receiver Sensitivity

The sensitivity of a receiver is a measure of the lowest power input signal that can be received with a specified signal-to-noise ratio. In the early days of radio, this was a very important parameter and designers tried to achieve the maximum practical sensitivity. In recent years, device and design technology have improved to the point that other parameters may be of higher importance, particularly in the HF region and below.

THE RELATIONSHIP OF NOISE TO SENSITIVITY

Noise level is as important as signal level in determining sensitivity. This section builds on the discussion of noise in the **RF Techniques** chapter. The most important noise parameters affecting receiver sensitivity are *noise bandwidth*, *noise figure*, *noise factor*, and *noise temperature*.

Since received noise power is directly proportional to receiver bandwidth, any specification of sensitivity must be made for a particular noise bandwidth. For DSP receivers with extremely steep filter skirts, receiver bandwidth is approximately the same as the filter or operating bandwidth. For other filter types, noise bandwidth is somewhat larger than the filter's 6 dB response bandwidth.

The relationship of noise bandwidth to noise power is one of the reasons that narrow bandwidth modes, such as CW, have a significant signal-to-noise advantage over modes with wider bandwidth, such as voice, assuming the receiver bandwidth is the minimum necessary to receive the signal. For example, compared to a 2400-Hz SSB filter bandwidth, a CW signal received in a 200 Hz bandwidth will have a 2400/200 = 12 = 10.8 dB advantage in received noise power. That is the same difference as an increase in transmitter from 100 to 1200 W.

Sources of Noise

Any electrical component will generate a certain amount of noise due to random electron motion. Any gain stages after the internal noise source will amplify the noise along with the signal. Thus a receiver with no signal input source will have a certain amount of noise generated and amplified within the receiver itself.

Upon connecting an antenna to a receiver, there will be introduction of any noise external to the receiver that is on the received frequency. The usual sources and their properties are described below. **Table 12.5** presents typical levels of external noise in a 10 kHz bandwidth present in the environment from different sources.

Atmospheric noise. This is noise generated within our atmosphere due to natural phenomena. The principal cause is lightning which sends wideband signals great distances. All points on the Earth receive this noise, but it is much stronger in some regions than others depending on the amount of local lightning activity. This source is usually the strongest noise source in the LF range and may dominate well into the HF region, depending on the other noises in the region. The level of atmospheric noise tends to drop off by around 50 dB every time the frequency is increased by a factor of 10. This source usually drops in importance by the top of the HF range (30 MHz).

Man-made noise. This source acts in a similar manner to atmospheric noise, although it is more dependent on local activity rather than geography and weather. The sources tend to be sparks from rotating and other kinds of electrical machinery as well as gasoline engine ignition systems and some types of lighting. In recent years, noise from computing and network equipment, switchmode power supplies, and appliances has increased significantly in urban and suburban environments. All things being equal, this source, on average, drops off by about 20 dB every time the frequency is increased by a factor of ten. The slower decrease at higher frequencies is due to the sparks having faster rise times than lightning. The effect tends to be comparable to atmospheric noise in the broadcast band, less at lower frequencies and a bit more at HF.

Galactic Noise. This is noise generated by the radiation from heavenly bodies outside our atmosphere. Of course, while this is noise to *communicators*, it is the desired signal for *radio-astronomers*. This noise source is a major factor at VHF and UHF and is quite dependent on exactly where you point an antenna (antennas for those ranges tend to be small and are often *pointable*). It also happens that the Earth turns and sometimes moves an antenna into a position where it inadvertently is aimed at a noisy area of the galaxy. If the Sun, not surprisingly the strongest signal in our solar system, appears behind a communications satellite, communications is generally disrupted until the Sun is out of the antenna's receiving pattern. Galactic noise occurs on HF, as well: Noise from the planet Jupiter can be heard on the 15 meter band under quiet conditions, for example.

Thermal Noise. Unlike the previous noise sources, this one comes from our equipment. All atomic structures have electrons that move within their structures. This motion results in very small currents that generate small amounts of wideband signals. While each particle's radiation is small, the cumulative effect of all particles becomes significant as the previous sources roll off with increasing frequency. The reason that this effect is called *thermal* noise is because the electron motion increases with the particle's temperature. In fact the noise strength is directly proportional to the temperature, if measured in terms of absolute zero (0 K). For example, if we increase the temperature from 270 to 280 K, that represents an increase in noise power of 10/270, 0.037, or about 0.16 dB. Some extremely sensitive microwave receivers use cryogenically-cooled front-end amplifiers to provide large reductions in thermal noise.

Oscillator Phase Noise. As noted in the **Oscillators and Synthesizers** chapter, real oscillators will have phase-noise sidebands that extend out on either side of the nominal carrier frequency at low amplitudes. Any such noise will be transferred to the received signal and through *reciprocal mixing* create noise products from signals on adjacent channels through the mixer. A good receiver will be designed to have phase noise that is well below the level of other internally generated noise.

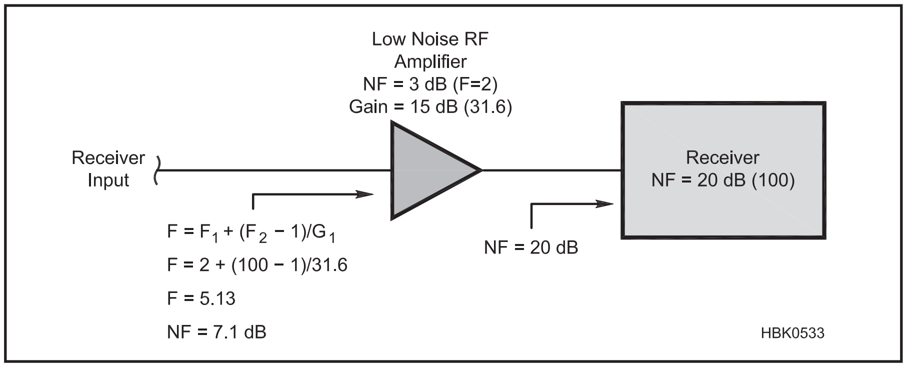


Fig 12.19 — The effect of adding a low noise preamplifier in front of a noisy receiver system.

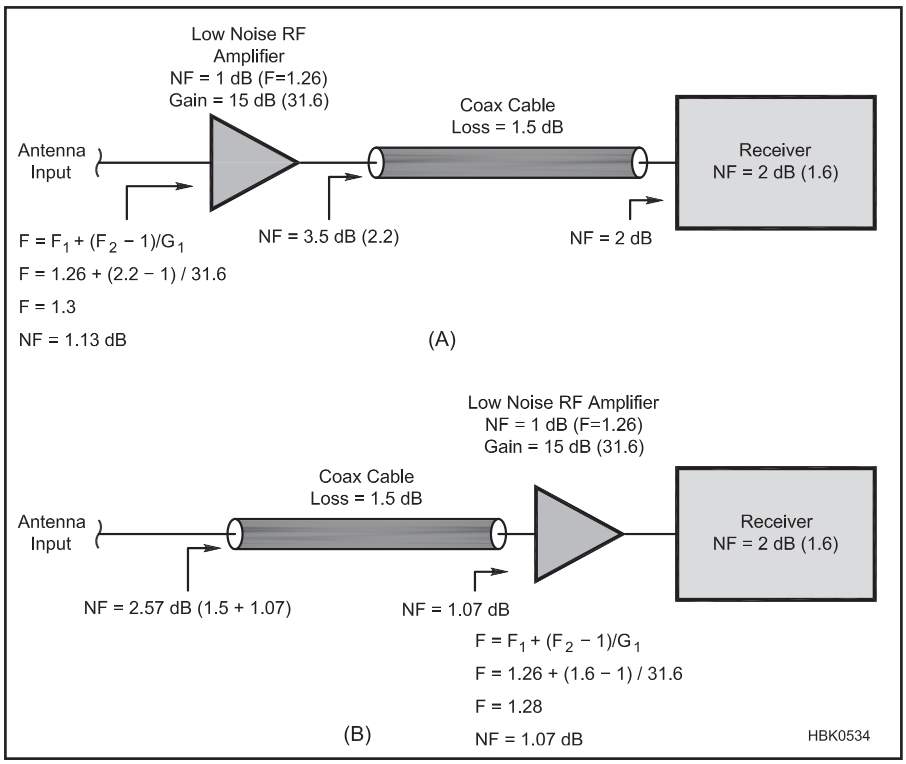


Fig 12.20 — Illustration of adding a VHF low noise preamplifier at the antenna (A) compared to one at the receiver system input (B).

Noise Power and Sensitivity

There are a number of related measures that can be used to specify the amount of noise that is generated within a receiver. If that noise approaches, or is within perhaps 10 dB of the amount of external noise received, then it must be carefully considered and becomes a major design parameter. If the internally generated noise is less than perhaps 10 dB below that expected from the environment, efforts to minimize internal noise are generally not beneficial and can, in some cases, be counterproductive.

While the total noise in a receiving system is, as discussed, proportional to bandwidth, the noise generating elements are generally not. Thus it is useful to be able to specify the internal noise of a system in a way that is independent of bandwidth. It is important to note that even though such a specification is useful, the actual noise is still directly proportional to bandwidth and any bandwidth beyond that needed to receive signal information will result in reduced SNR.

To evaluate the effect of noise power on sensitivity, we will use the following equations from the discussion of noise in the **RF Techniques** chapter:

$$F = 1 + T_E/T_0 \quad (1)$$

where

F is the noise factor,

T_E is the equivalent noise temperature and

T_0 is the standard noise temperature, usually 290 K.

$$N_i = k \times B \times T_E \quad (2)$$

where

N_i is the equivalent noise power in watts at the input of a perfect receiver that would result in the same noise output,

k is Boltzman's constant, 1.38×10^{-23} joules/kelvin, and

B is the system noise bandwidth in hertz.

Expressing Ni in dBm is more convenient and is calculated as:

$$dBm_i = -198.6 + (10 \times \log_{10} B) + (10 \times \log_{10} T_E) \quad (3)$$

If N_i is greater than the noise generated internally by the receiver, the receiver's sensitivity is limited by the external noise. This is usually the case for HF receivers where atmospheric and man-made noise are much stronger than the receiver's internal noise floor. If N_i is within, perhaps, 10 dB greater than the receiver's internal noise, then the effect of the receiver's internal circuits on overall system sensitivity must be taken into account.

Noise Figure of Cascaded Stages

Often we are faced with the requirement of determining the noise figure of a system of multiple stages. In general adding an amplification stage between the antenna and the rest of the system will reduce the equivalent noise figure of the system by the amount of gain of the stage but adds in the noise of the added stage directly. The formula for determining the resultant noise factor is:

$$F = F_1 + \frac{F_2 - 1}{G_1} \quad (4)$$

where

F_1 is the noise factor of the stage closest to the input,

F_2 is the noise factor of the balance of the system, and

G_1 is the gain of the stage closest to the input.

Note that these are noise *factors* not noise *figures*. The change to noise figure is easily computed at the end, if desired.

This is commonly encountered through the addition of a *low noise preamplifier* ahead of a noisy receiver as shown in **Fig 12.19.** As shown in the figure for fairly typical values, while the addition of a low noise preamplifier does reduce the noise figure, as can be observed, the amplifier gain and noise figure of the rest of the receiver can make a big difference.

In many cases elements of a receiving system exhibit loss rather than gain. Since they reduce the desired signal, while not changing the noise of following stages, they increase the noise figure at their input by an amount equal to the loss. This is the reason that VHF low noise preamps are often antenna mounted. If the coax between the antenna and radio has a loss of 1 dB, and the preamp has a noise figure of 1 dB, the resulting noise figure with the preamp at the radio will be 2 dB, while if at the antenna, the noise added by the coax will be reduced by the gain of the preamp, resulting in a significant improvement in received SNR at VHF and above. **Fig 12.20** shows a typical example.

To determine the system noise factor of multiple cascaded stages, Equation 4 can be extended as follows:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + ... + \frac{F_N - 1}{G_1 G_2 ... G_{N-1}} \quad (5)$$

where F_N and G_N are the noise factor and gain of the Nth stage.

This can proceed from the antenna input to the last linear stage, stopping at the point of detection or analog-to-digital conversion, whichever occurs first.

Why Do We Care?

A receiver designer needs to know how strong the signals are to establish the range of signals the receiver will be required to handle. One may compare the equivalent noise power determined in Equation 3 with the expected external noise to determine whether the overall receiver SNR will be determined by external or internal noise. A reasonable design objective is the have the internal noise be less than perhaps 10 to 20 dB below the expected noise. As noted above, this is related closely to the frequency of signals we want to receive. Any additional sensitivity will not provide a noticeable benefit to SNR, and may result in reduced dynamic range, as will be discussed in the next section.

For frequencies at which external noise sources are strongest, the noise power (and signal power) will also be a function of the antenna design. In such cases, the signal-to-noise ratio can be improved by using an antenna that picks up more signal and less noise, such as with a directional antenna that can reject noise from directions other than that of the signal. Some antennas improve SNR simply by rejecting noise, such as the Beverage antenna. Signals from a Beverage antenna are usually much weaker than from a conventional antenna but their ability to reject noise from undesired directions creates a net improvement in the SNR at the receiver output.

12.4.2 Receiver Image Rejection

The major design choice in determining image response is selection of the first and any following IF frequencies. This is important in receiving weak signals on bands where strong signals are present. If the desired signal is near the noise, a signal at an image frequency could easily be 100 dB stronger, and thus to avoid interference, an image rejection of 110 dB would be needed. While some receivers meet that target, the receiver sections of most current amateur transceivers are in the 70 to 100 dB range.

Receiver Architectures for Image Rejection

As noted earlier, improved crystal filter technology allows *down-conversion* HF receivers to use an IF in the 4-10 MHz range. With a 10 MHz IF and an LO above the signal frequency, a 30 MHz signal would have an image at 50 MHz. This makes image-rejection filtering relatively straightforward, although many receiver IF frequencies tend to be at the lower end of the above range. Still, as will be discussed in the next section, they have other advantages.

Many current HF receivers (or receiver sections of transceivers) have elected to employ an *up-converting* architecture. They typically have an IF in the VHF range, perhaps 60 to 70 MHz, making HF image rejection easy. A 30 MHz signal with a 60 MHz IF will have an image at 150 MHz. Not only is it five times the signal frequency, but signals in this range (other than perhaps the occasional taxicab) tend to be weaker than some undesired HF signals. Receivers with this architecture have image responses at the upper end of the above range, often with the image rejected by a relatively simple low-pass filter with a cut off at the top of the receiver range.

Another advantage of this architecture is that the local oscillator can cover a wide continuous range, making it convenient for a general coverage receiver. For example, with a 60 MHz IF, a receiver designed for LF through HF would need an LO covering 60.03 to 90 MHz, a 1.5 to 1 range, easily provided by a number of synthesizer technologies, as described in the **Oscillators and Synthesizers** chapter.

The typical upconverting receiver uses multiple conversions to move signals to frequencies at which operating bandwidth can be established. While crystal filters in the VHF range used by receivers with upconverting IFs have become available with bandwidths as narrow as around 3 kHz, they do not yet achieve the shape factor of similar bandwidth filters at MF and HF. Thus, these are commonly used as *roofing filters*, discussed in the next section, prior to a conversion to one or more lower IF frequencies at which the operating bandwidth is established. **Fig 12.21** is a block diagram of a typical upconverting receiver using DSP for setting the operating bandwidth.

12.4.3 Receiver Dynamic Range

Dynamic range can be defined as the ratio of the strongest to the weakest signal that a system, in this case our receiver, can respond to linearly. Table 12.5 gives us an idea of how small a signal we might want to receive. The designer must create a receiver that will handle signals from below the noise floor to as strong as the closest nearby transmitter can generate. Most receivers have a specified (or sometimes not) highest input power that can be tolerated, representing the other end of the spectrum. Usually the maximum power specified is the power at which the receiver will not be damaged, while a somewhat lower power level is generally the highest that the receiver can operate at without overload and the accompanying degradation of quality of reception of the desired signal.

ON-CHANNEL DYNAMIC RANGE

The signal you wish to listen to can range

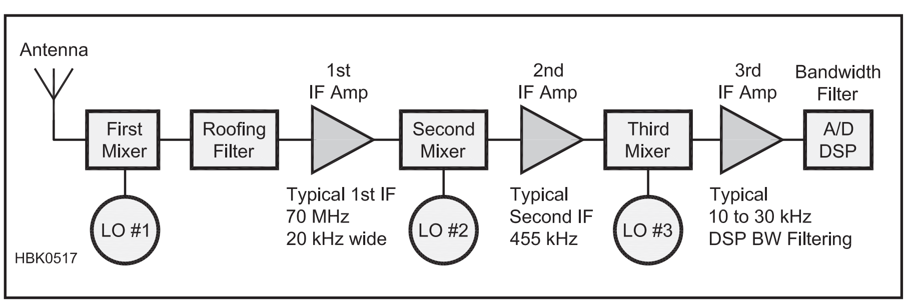


Fig 12.21 — Block diagram of a typical upconverting receiver using DSP for operating bandwidth (BW) determination. Receivers applying DSP filtering at the second IF and at higher frequencies are common.

from the strongest to the weakest, sometimes changing rapidly with conditions, or in a situation with multiple stations such as a net. While a slow change in signal level can be handled with manual gain controls, rapid changes require automatic systems to avoid overload and operator discomfort.

This is a problem that has been long solved with automatic gain control (AGC) systems. These systems are described in a further section of the chapter, but it is worth pointing out that the measurement and gain control points need to be applied carefully to the most appropriate portions of the receiver to maintain optimum performance. If all control is applied to early stages, the SNR for strong stations may suffer, while if applied in later stages, overload of early stages may occur in the presence of strong stations. Thus, gain control has to be designed into the receiver distributed from the input to the detector.

The next two sections illustrate a frequent limitation of receiver performance — dynamic range between the reception of a weak signal in the presence of one or more strong signals outside of the channel.

BLOCKING GAIN COMPRESSION

A very strong signal outside the channel bandwidth can cause a number of problems that limit receiver performance. *Blocking gain compression* (or "blocking") occurs when strong signals overload the receiver's high gain amplifiers and reduce its ability to amplify weak signals. (Note that the term "blocking" is often used outside Amateur Radio when referring to reciprocal mixing of oscillator noise with strong local signals.)

While listening to a weak signal, all stages operate at maximum gain. If the weak signal were at a level of S0, a strong signal could be at S9 + 60 dB. Using the standard of S9 representing a 50 μV input signal, and each S unit reflecting a change of 6 dB, the receiver's front-end stages would be receiving a 0.1 μV signal and a 50,000 μV into the front end at the same time. A perfectly linear receiver would amplify each signal equally until the undesired signal is eliminated at the operating bandwidth setting stage. However, in practical receivers, after a few stages of full gain amplification, the stronger signal causes amplifier clipping, which reduces the gain available to the strong signal. This is seen as a gradual reduction in gain as the input signal amplitude increases. Gain reduction also reduces the amplitude of the weaker signal which is perceived to fade as the strong signal increases in amplitude. Eventually, the weaker signal is no longer receivable and is said to have been "blocked", thus the name for the effect.

The ratio in dB between the strongest signal that a receiver can amplify linearly, with no more than 1 dB of gain reduction, and the receiver's noise floor in a specified bandwidth is called the receiver's *blocking dynamic range* or *BDR* or the *compression-free dynamic range* or *CFDR*. In an analog superhet, BDR is established by the linear regions of the IF amplifiers and mixers. If the receiver employs DSP, the range of the analog-to-digital converter usually establishes the receiver's BDR.

A related term is "near-far interference" which is used primarily in the commercial environment to refer to a strong signal causing a receiver to reduce its gain and along with it the strength of weak received signals.

INTERMODULATION DYNAMIC RANGE

Blocking dynamic range is the straightforward response of a receiver to a single strong interfering signal outside the operating passband. In amateur operation, we often have more than one interferer. While such signals contribute to the blocking gain compression in the same manner as a single signal described above, multiple signals also result in a potentially more serious problem resulting from *intermodulation products*.

If we look again at Equation B in the earlier sidebar on Nonlinear Signal Combinations, we note that there are an infinite number of higher order terms. In general, the coefficients of these terms are progressively lower in amplitude, but they are still greater than zero. Of primary interest is third-order term, $K_3 \times V_{IN}^3$, when considering V_{IN} as the sum of two interfering signals (f_1 and f_2) near our desired signal (f_0) and within the first IF passband, but outside the operating bandwidth.

$$V_{OUT} = K_3 \times [A\sin(f_1)t + B\sin(f_2)t]^3 \qquad (6)$$

$$= K_3 \times \{A^3\sin^3(f_1)t + 3A^2B\,[\sin^2(f_1)t \times \sin(f_2)t]$$

$$+ 3AB^2\,[\sin(f_1)t \times \sin^2(f_2)t] + B^3\sin^3(f_2)t\,\}$$

The cubic terms in Equation 6 (the first and last terms) result in products at three times the frequency and can be ignored in this discussion. Using trigonometric identities to reduce the remaining $\sin^2$ terms and the subsequent cos( )sin( ) products reveal individual intermodulation (IM) products, recognizing that the signals have cross-modulated each other due to the nonlinear action of the circuit. (Math handbooks such as the *CRC Standard Mathematical Tables and Formulae* have all the necessary trigonometry information.)

IM products have frequencies that are linear combinations of the input signal frequencies, written as $n(f_1) \pm m(f_2)$, where n and m are integer values. The entire group of products that result from intermodulation are broadly referred to as *intermodulation distortion* or IMD. The ratio in dB between the amplitude of the interfering signals, f_1 and f_2, and the resulting IM products is called the *intermodulation ratio*.

If *all* of the higher-order terms in the original equation are considered, n and m can take on any integer value. If the sum of n and m is odd, (2 and 1, or 3 and 2, or 3 and 4, etc.) the result is products that have frequencies near our desired signal, for example, $2(f_1) - 1(f_2)$. Those are called *odd-order* products. Odd-order products have frequencies close enough to those of the original signals that they can cause interference to the desired signal. If the sum of n and m is three, those are *third-order IM products* or *third-order IMD*. For fifth-order IMD, the sum of n and m is five, and so forth. The higher the order of the IM products, the smaller their amplitude, so our main concern is with third-order IMD.

If the two interfering signals have frequencies of $f_0 + \Delta$, and $f_0 + 2\Delta$, where Δ is some offset frequency, we have for the third-order term:

$$V_{OUT} = K_3 \times [A\sin(f_0 + \Delta)t + B\sin(f_0 + 2\Delta)t]^3 \qquad (7)$$

A good example would be interfering signals with offsets of 2 kHz and 2 × 2 kHz or 4 kHz from the desired frequency, a common situation on the amateur bands.

Discarding the cubic terms and applying the necessary trigonometric identities shows that a product can be produced from this combination of interfering frequencies that has a frequency of exactly f_0 — the same frequency as the desired signal! (The higher-order terms of Equation B can also produce products at f_0, but their amplitude is usually well below those of the third-order products.)

Thus we have two interfering signals that are not within our operating bandwidth so we don't hear either by themselves. Yet they combine in a nonlinear circuit and produce a signal exactly on top of our desired signal. If the interfering signals are within the passband of our first IF and are strong enough the IM product will be heard.

As the strength of the interfering signals increases, so does that of the resulting intermodulation products. For every dB of increase in the interfering signals, the third-order IM products increase by approximately 3 dB. Fifth-order IM increases by 5 dB for every dB increase in the interfering signals, and so forth. Our primary concern, however, is with the third-order products because they are the strongest and cause the most interference.

INTERCEPT POINT

Intercept point describes the IMD performance of an individual stage or a complete receiver. For example, third-order IM products increase at the rate of 3 dB for every 1-dB increase in the level of each of the interfering input signals (ideally, but not always exactly true). As the input levels increase, the distortion products seen at the output on a spectrum analyzer could catch up to, and equal, the level of the two desired signals if the receiver did not begin to exhibit blocking as discussed earlier.

The input level at which this occurs is the *input intercept point*. **Fig 12.22** shows the concept graphically, and also derives from the geometry an equation that relates signal level, distortion and intercept point. The intercept point of the most interest in receiver evaluation is that for third-order IM products and is called the *third-order intercept point* or IP_3. A similar process is used to get a second-order intercept point for second-order IMD. A higher IP_3 means that third-order IM products will be weaker for specific input signal strengths and the operator will experience less interference from IM products from strong adjacent signals.

These formulas are very useful in designing radio systems and circuits. If the input intercept point (dBm) and the gain of the stage (dB) are added the result is an output intercept point (dBm). Receivers are specified by input intercept point, referring distortion back to the receive antenna input. Intercept point is a major performance limitation of receivers used in high density contest or DX operations. Keep in mind that we have been discussing this as an effect of two signals, one that is Δ away from our operating frequency and another at twice Δ. In real life, we may be trying to copy a weak signal at f_0, and have other signals at $f_0 \pm 500$, 750, 1000, 1250…5000 Hz. There will be many combinations that produce products at or near our weak signal's frequency.

Note that the products don't need to end up exactly on top of the desired signal to cause a problem; they just need to be within the operating bandwidth. So far we have been talking about steady carriers, such as would be encountered during CW operation with interference from nearby CW stations. SSB or other wider bandwidth modes with spectrum distributed across a few kHz will have signal components that go in and out of a relationship that results in on-channel interference from IMD. This manifests itself as a time-varying synthetic noise floor, composed of all the resulting products across the channel. The difference in this low level "noise" can be dramatic between different receivers, especially when added to phase noise received from other stations and reciprocal mixing inside the receiver!

SPURIOUS-FREE DYNAMIC RANGE

IM products increase with the amplitude of the interfering signals that cause them and at some point become detectable above the receiver's noise floor. The ratio of the strength of the interfering signals to the noise floor, in dB, is the receiver's *spurious-free dynamic range* or *SFDR*. This is the range of signal strengths over which the receiver does not produce any detectable spurious products. SFDR can be specified for a specific order of IM products; for example, SFDR3 is the SFDR measured for third-order IM products only. The bandwidth for which the receiver's noise floor is measured must also be specified, since smaller bandwidths will result in a lower noise floor.

RECEIVER DESIGN FOR DYNAMIC RANGE

As noted previously, there are a number of possible architectural choices for an amateur receiver. In the past, the receivers with the best close-in third order intermodulation distortion and maximum blocking dynamic range were amateur-band-only receivers, such as the primary receiver in the TEN-TEC Orion series, the receiver in the Elecraft K2 , the earlier TEN-TEC Omni VI and both receivers in the Elecraft K3. A look at a typical block diagram, as shown in **Fig 12.23**, makes it easy to see why. The problems resulting from strong unwanted signals near a desired one are minimized if the unwanted signals are kept out of the places in the receiver where they can be amplified even more and cause the nonlinear effects that we try to avoid.

Note that in Fig 12.23, the only place where the desired and undesired signals all coexist is before the first mixer. If the first mixer and any RF preamp stages have sufficient strong-signal handling capability, the undesired signals will be eliminated in the filter immediately behind the first mixer. This HF crystal filter is generally switchable to support desired bandwidths as narrow as 200 Hz. The later amplifier, mixer and DSP circuits only have to deal with the signal we want. For additional discussion of these issues, see "International Radio Roofing Filters for the Yaesu FT-1000 MP Series Transceivers," by Joel Hallas, W1ZR, in *QST* Product Review for February 2005.

Now look at a typical modern general coverage receiver as shown previously in Fig 12.21. In this arrangement, a single digital synthesizer, perhaps covering from 70 to 100 MHz, shifts the incoming signal(s) to a VHF IF, often near 70 MHz. A roofing filter at 70 MHz follows the first mixer. This arrangement offers simplified local oscillator (LO) design and the possibility of excellent image rejection. Unfortunately, crystal filter technology has only recently been able to produce narrow filters for

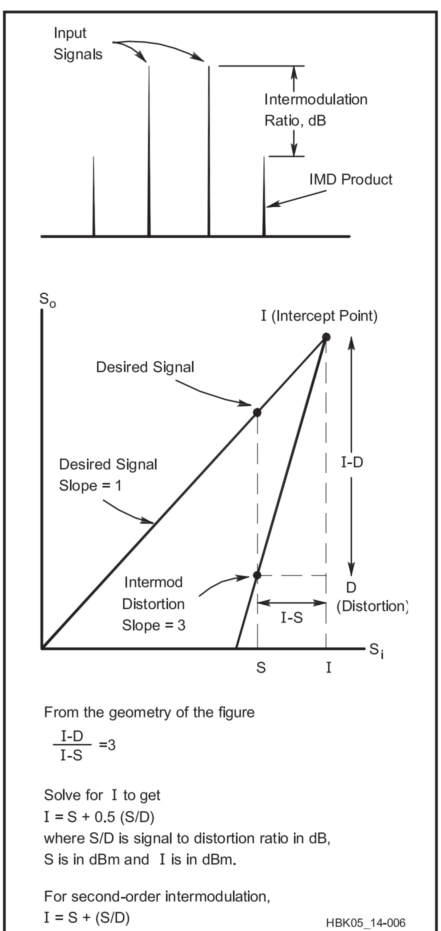


Fig 12.22 — Graphical representation of the third-order intercept concept.

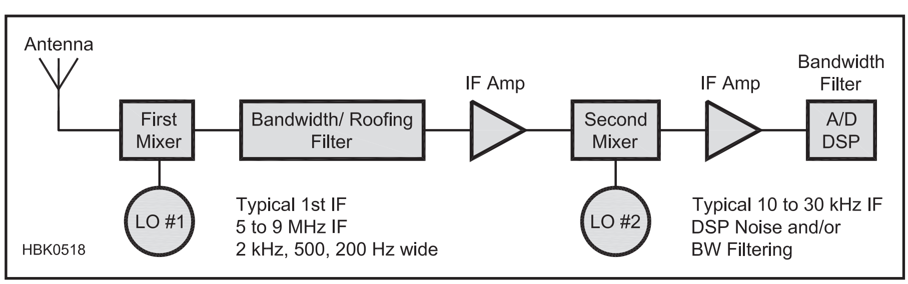


Fig 12.23 — Block diagram of a downconverting amateur band receiver with HF IF filtering to eliminate near channel interference.

70 MHz, and so far they have much wider skirts than the crystal filters used in Fig 12.23.

Many receivers and transceivers set this filter bandwidth wider than any operating bandwidth and use DSP filtering much later in the signal chain to set the final operating bandwidth. For a receiver that will receive FM and AM, as well as SSB and CW, that usually means a roofing filter with a bandwidth of around 20 kHz. With this arrangement, all signals in that 20 kHz bandwidth pass all the way through IF amplifiers and mixers and into the A/D converter before we attempt to eliminate them using DSP filters. By that time they have had an opportunity to generate intermodulation products and cause the blocking and IMD problems that we are trying to eliminate. This situation is changing with the introduction of the top of the line radios that feature both general coverage at HF and VHF roofing filters, such as the ICOM IC-7800 and Yaesu FTdx9000 transceivers. (See the Transceiver Survey by W1ZR on the *Handbook's* CD-ROM for more information on the latest models.)

A hybrid architecture has recently appeared in the TEN-TEC Omni VII transceiver that effectively combines the two technologies. The first IF has a 20 kHz wide roofing filter at 70 MHz, followed by selectable steep skirted 455 kHz Collins mechanical filters at the second IF and then DSP filters at the third IF. The topology is shown in **Fig 12.24**.

Careful attention to gain distribution among the stages between the filters maintains desired sensitivity, but not so high that the undesired products have a chance to become a serious problem. With bandwidths of 20, 6 and 2.5 kHz supplied, and 500 and 300 Hz as accessories, the undesired close-in signals are eliminated before they have an opportunity to cause serious trouble in the DSP stages that follow.

Another variation is found in the Kenwood TS-590S which switches between down-conversion on the more crowded "contest bands" (160, 80, 40, 20 and 15 meters) and up-conversion on the remaining bands. In effect, trading sensitivity for dynamic range.

WHAT KIND OF PERFORMANCE CAN WE EXPECT?

As we would expect, the blocking and IMD dynamic range (IMD DR) performance of a receiver will depend on a combination of the early stage filtering, the linearity of the mixers and amplifiers, and the dynamic range of any ADC used for DSP. Product reviews of receivers and the receiver sections of transceivers published in *QST* now provide the measured dynamic range in the presence of interfering signals with spacings of 2, 5 and 20 kHz. (Details of the test procedures used are given in the **Test Equipment and Measurements** chapter.) At 20 kHz spacing, the interfering signal is usually outside of the roofing filter bandwidth of any of the above architectures. Spacing of 2 and 5 kHz represents likely conditions on a crowded band.

A look at recent receiver measurements indicates that receivers have IMD dynamic range with 2 kHz spacing results in the following ranges:

Upconverting with VHF IF (Fig 12.21): 60 to 80 dB

Downconverting with HF IF (Fig 12.23): 75 to 105 dB

Hybrid distributed architecture (Fig 12.24): Omni VII, 82 dB

Let's take an example of what this would mean. If we are listening to a signal at S3, for signals to generate a third-order IMD product at the same level in a receiver with a dynamic range of 60 dB, the $f_0 + \Delta$ and $f_0 + 2\Delta$ signals would have to have a combined power equal to S9 +27 dB, or each at S9 +24 dB. This is not unusual on today's amateur bands. On the other hand, if we had an IMD dynamic range of 102 dB, the interfering signals would have to at S9 +66 dB, much less likely. How much dynamic range you need depends in large measure on the kind of operating you do, how much gain your receiving antennas have and the closeness of the nearest station that operates on the same bands as you.

Keep in mind also that it is often difficult to tell whether or not you are suffering from IMD — it just sounds like there are many more signals than are really present. A good test to assess the source of interference is first switch off any preamplifiers and noise-blankers or noise-reduction systems that affect the receiver's linearity. Observe the level of the interference (if it's still there) and then switch in some attenuation at the front-end of the receiver. If the level of the interference goes down by *more* than the level of attenuation (estimate 6 dB per S unit), then the interference is being generated (or at least aggravated) by non-linearity inside the receiver. Continue to increase attenuation until the interference either goes away or goes down at the same rate as the attenuation is increased. You might be surprised at how much better the band "sounds" when your receiver is operating in its linear region!

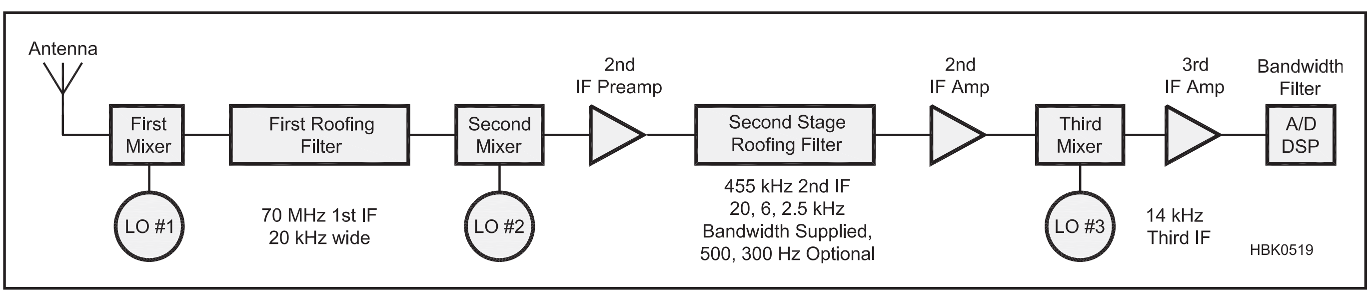


Fig 12.24 — Block diagram of a upconverting multiple conversion receiver with distributed roofing filter architecture.

12.5 Control and Processing Outside the Primary Signal Path

Discussion thus far has been about processes that occur within the primary path for signals between antenna input and transducer or system output. There are a number of control and processing subsystems that occur outside of that path that contribute to the features and performance of receiving systems.

12.5.1 Automatic Gain Control (AGC)

The amplitude of the desired signal at each point in the receiver is generally controlled by the AGC system, although manual control is usually provided as well. Each stage has a distortion versus signal level characteristic that must be known, and the stage input level must not become excessive. The signal being received has a certain signal-to-distortion ratio that must not be degraded too much by the receiver. For example, if an SSB signal has –30 dB distortion products the receiver should create additional distortion no greater than -40 dB with respect to the desired signal. The correct AGC design ensures that each stage gets the right input level. It is often necessary to redesign some stages in order to accomplish this task.

While this chapter deals exclusively with AGC in the guise of analog circuits, the same function is also implemented digitally in DSP and SDR receivers. The goal of both is the same — to maintain a signal level at all stages of the receiver that is neither too large nor too small so that the various processing systems operate properly. Whether or not the AGC offset and time constant are implemented by an analog component or by a microprocessor output is immaterial. The point is to manage the RF amplifier gain so that the overall receiver behavior is satisfactory.

The effects of an improperly operating AGC system can be quite subtle or nearly disabling to a receiver and vary with how

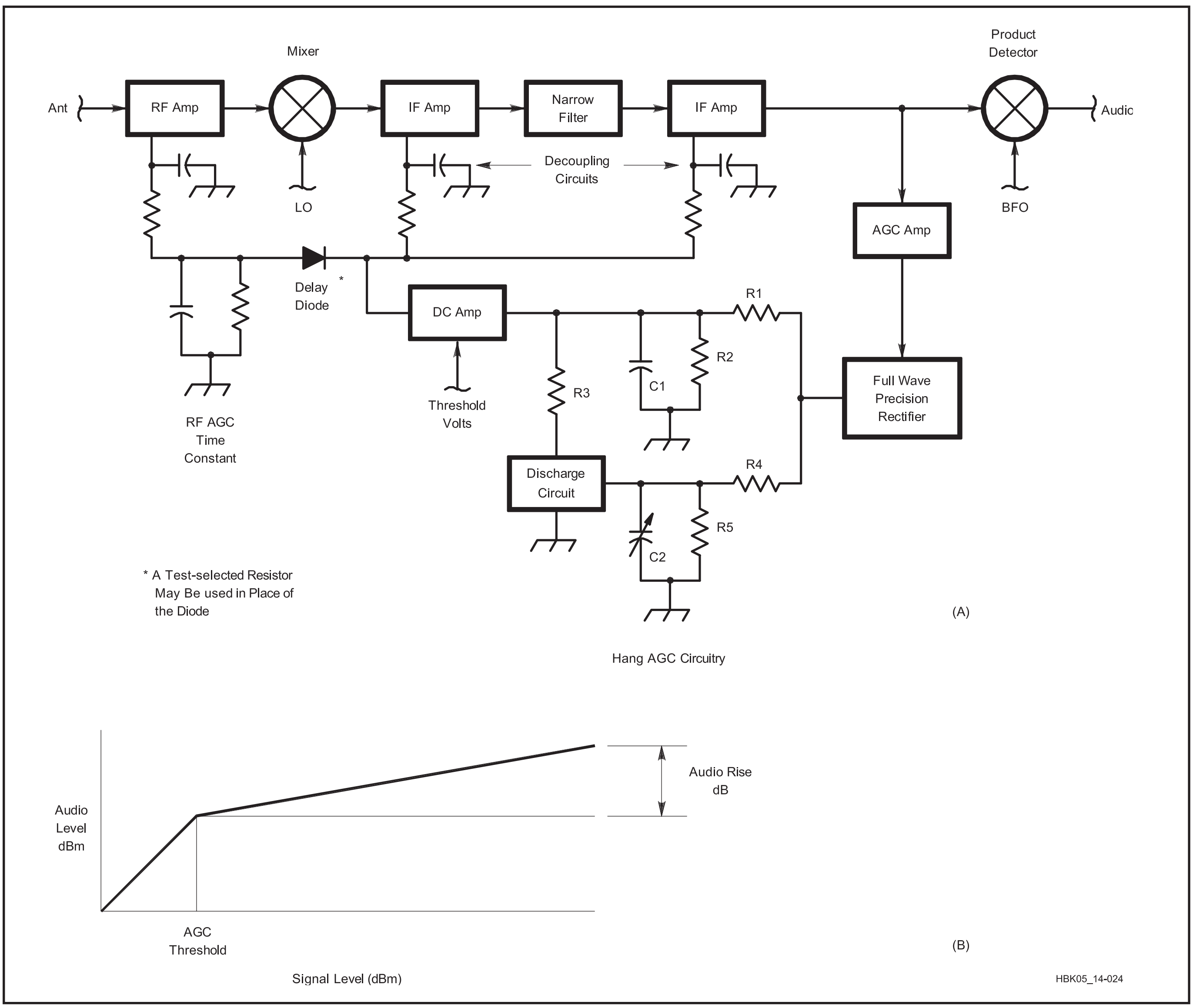


Fig 12.25 — AGC principles. At A: typical superhet receiver with AGC applied to multiple stages of RF and IF. At B: audio output as a function of antenna signal level.

the AGC system is constructed. This chapter attempts to describe the requirements for proper operation and provides some examples of implementation and common AGC failures in terms of analog circuitry which is somewhat easier to describe than software algorithms, noting that similar behaviors exist even in purely software receivers. The interested student should consider studying the AGC systems of commercial receivers to understand how professional design teams deal with the problem of managing so much gain with such stringent requirements for linearity and distortion.

THE AGC LOOP

Fig 12.25A shows a typical AGC loop that is often used in amateur superhet receivers. The AGC is applied to the stages through RF decoupling circuits that prevent the stages from interacting with each other. The AGC amplifier helps to provide enough AGC loop gain so that the gain-control characteristic of Fig 12.25B is achieved. If effect, the AGC system causes the receiver to act as a compression amplifier with lower overall gain for stronger signals.

The AGC action does not begin until a certain level, called the AGC *threshold*, is reached. The THRESHOLD VOLTS input in Fig 12.25A serves this purpose. After that level is exceeded, the audio level increases more slowly than for weaker signals. The audio rise beyond the threshold value is usually in the 5 to 10 dB range. Too much or too little audio rise are both undesirable for most operators.

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

Fig 12.26 is a typical plot of the signal levels at the various stages of a certain ham band receiver using analog circuitry. Each stage has the proper level and a 115 dB change in input level produces a 10 dB change in audio level. A manual gain control could produce the same effect.

AGC TIME CONSTANTS

There are two primary AGC time constants. AGC *attack time* describes the time it takes the AGC system to respond to the presence of a signal. AGC *decay time* describes the response of the AGC system to changes in a signal that is present. The optimum time constants for the AGC system depends on the type of signal being received, the type of operation being conducted, and the operator's preference.

In Fig 12.25, following the precision rectifier, R1 and C1 set an attack time, to prevent excessively fast application of AGC. One or two milliseconds is a good value for the R1 × C1 product. If the antenna signal suddenly disappears, the AGC loop is opened because the precision rectifier stops conducting. C1 then discharges through R2 and the C1 × R2 product can be in the range of 100 to 200 ms. At some point the rectifier again becomes active, and the loop is closed again.

An optional modification of this behavior is the *hang AGC* circuit. If we make R2 × C1 much longer, say 3 seconds or more, the AGC voltage remains almost constant until the R5, C2 circuit decays with a switch selectable time constant of 100 to 1000 ms. At that time R3 quickly discharges C1 and full receiver gain is quickly restored. This type of control is appreciated by many operators because of the lack of AGC *pumping* due to modulation, rapid fading and other sudden signal level changes.

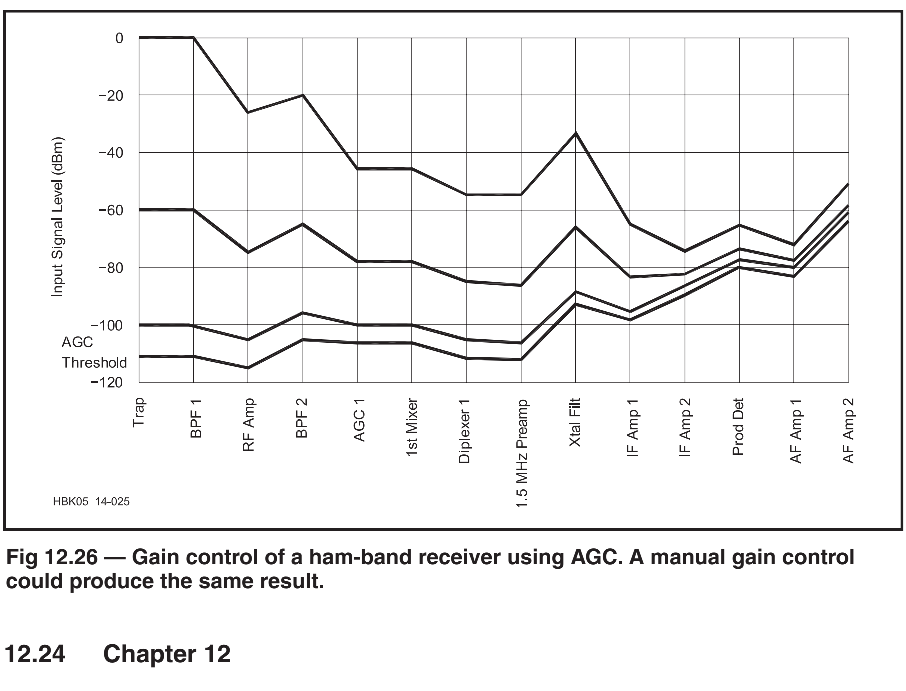


Fig 12.26 — Gain control of a ham-band receiver using AGC. A manual gain control could produce the same result.

AGC PUMPING

AGC pumping can occur in receivers in which the AGC measurement point is located ahead of the stages that determine operating bandwidth, such as when an audio filter is added to a receiver externally and outside the reach of the AGC system. If the weak signal is the only signal within the first IF passband, the AGC will cause the receiver to be at maximum gain and optimum SNR. If an interfering signal is within the first IF passband, but outside the audio DSP filter's passband, we won't hear the interfering signal, but it will enter the AGC system and reduce the gain so we might not hear our desired weak signal. AGC pumping is audible as sudden reductions in signal strength without a strong signal in the passband of the receiver

AGC LOOP RESPONSE PROBLEMS

If the various stages have the property that each 1 V change in AGC voltage changes the gain by a constant amount (in dB), the AGC loop is said to be *log-linear* and regular feedback principles can be used to analyze and design the loop. But there are some difficulties that complicate this textbook model. One has already been mentioned, that when the signal is rapidly decreasing the loop becomes open and the various capacitors discharge in an open loop manner. As the signal is increasing beyond the threshold, or if it is decreasing slowly enough, the feedback theory applies more accurately.

In SSB and CW receivers rapid changes are the rule and not the exception. It is important that the AGC loop not overshoot or ring when the signal level rises past the threshold. The idea is to design the ALC loop to be stable when the loop is closed. It obviously won't oscillate when open (during decay time). But the loop must have smooth and consistent transient response when the loop goes from open to closed state.

Another problem involves the narrow bandpass IF filter. The group delay of these filters constitutes a time lag in the loop that can make loop stabilization difficult. Moreover, these filters nearly always have much greater group delay at the edges of the passband, so that loop problems are aggravated at these frequencies. Overshoots and undershoots, called *gulping*, are very common. Compensation networks that advance the phase of the feedback help to offset these group delays. The design problem arises because some of the AGC is applied before the filter and some after the filter. It is a good idea to put as much fast AGC as possible after the filter and use a slower decaying

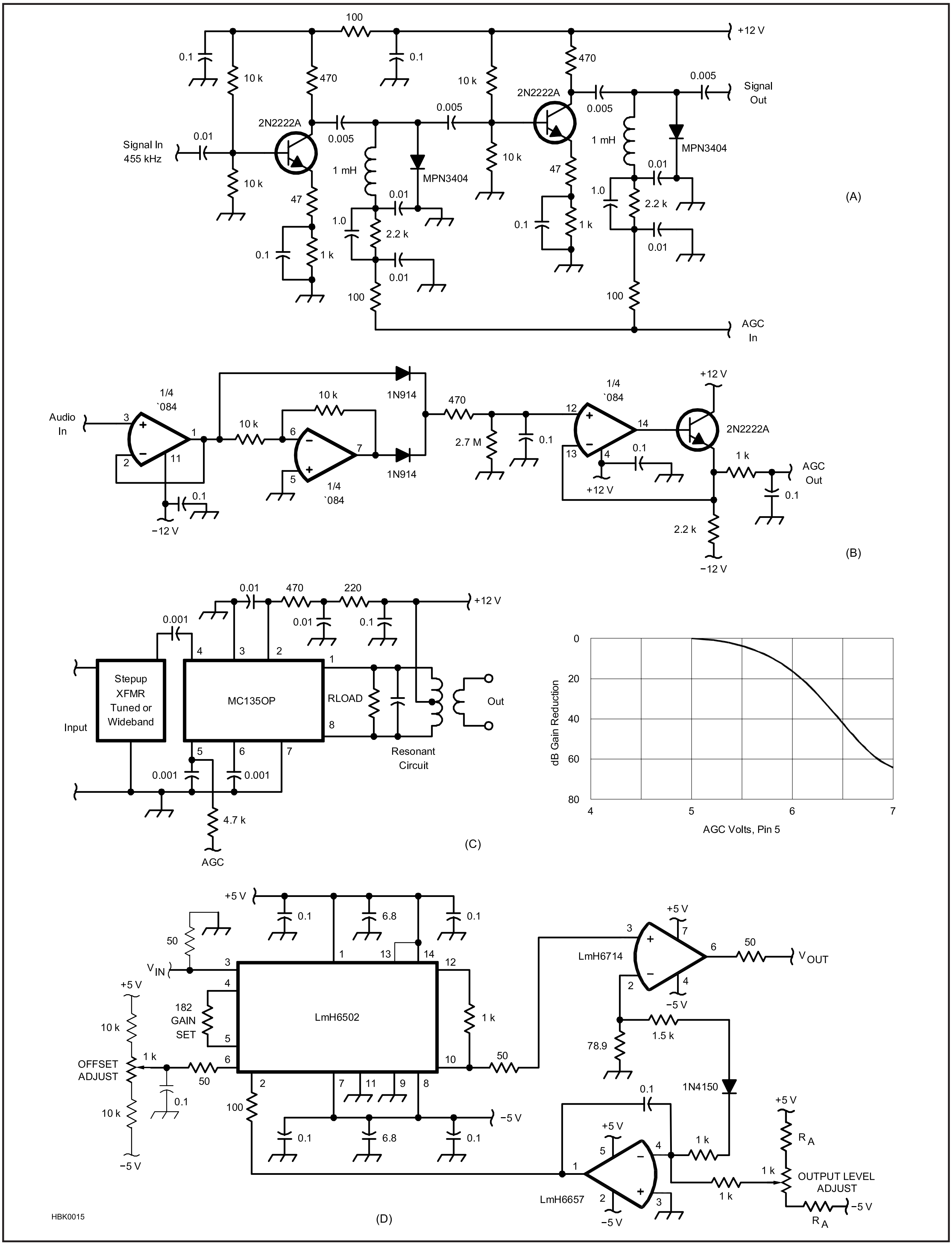


Fig 12.27 — Some gain controllable amplifiers and a rectifier suitable for audio derived AGC.

AGC ahead of the filter. The delay diode and RC in Fig 12.25A are helpful in that respect. Complex AGC designs using two or more compensated loops are also in the literature. If a second cascaded narrow filter is used in the IF it is usually a lot easier to leave the second or *downstream* filter out of the AGC loop at the risk of allowing AGC pumping as described in the preceding section.

Another problem is that the control characteristic is often not log-linear. For example, dual-gate MOSFETs tend to have much larger dB/V at large values of gain reduction. Many IC amplifiers have the same problem. The result is that large signals cause instability because of excessive loop gain. Variable gain op amps and other similar ICs are available that are intended for gain control loops.

Audio frequency components on the AGC bus can cause problems because the amplifier gains are modulated by the audio and distort the desired signal. A hang AGC circuit (essentially a low-pass filter) can reduce or eliminate this problem.

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

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

12.5.2 Audio-Derived AGC

Some receivers, especially direct-conversion types, use audio-derived AGC. There are problems with this approach as well. At low audio frequencies the AGC control action can be slow to develop. That is, low-frequency audio sine waves take longer to reach their peaks than the AGC time constants. During this time the RF/IF/AF stages can be overdriven. If the RF and IF gains are kept at a low level this problem can be reduced. Also, attenuating low audio frequencies prior to the first audio amplifier should help. With audio AGC, it is important to avoid so-called "charge pump" rectifiers or other slow-responding circuits that require multiple cycles of audio to pump up the AGC voltage. Instead, use a peak-detecting circuit that responds accurately on the first positive or negative half-cycle.

12.5.3 AGC Circuits

Fig 12.27 shows some gain-controllable circuits. Fig 12.27A shows a two-stage 455-kHz IF amplifier with PIN diode gain control. This circuit is a simplified adaptation from a production receiver, the Collins 651S. The IF amplifier section shown is preceded and followed by selectivity circuits and additional gain stages with AGC. The 1.0 µF capacitors aid in loop compensation. The favorable thing about this approach is that the transistors remain biased at their optimum operating point. Right at the point at which the diodes start to conduct, a small increase in IMD may be noticed, but that goes away as diode current increases slightly. Two or more diodes can be used in series, if this is a problem (it very seldom is). Another solution is to use a PIN diode that is more suitable for such a low-frequency IF. Look for a device with $\tau > 10 / (2\pi f)$ where τ is the minority carrier lifetime in µs and f is the frequency in MHz.

Fig 12.27B is an audio derived AGC circuit using a full-wave rectifier that responds to positive or negative excursions of the audio signal. The RC circuit follows the audio closely.

Fig 12.27C shows a typical circuit for the MC1350P RF/IF amplifier. The graph of gain control versus AGC voltage shows the change in dB/V. If the control is limited to the first 20 dB of gain reduction this chip should be favorable for good AGC transient response and good IMD performance. Use multiple low-gain stages rather than a single high-gain stage for these reasons. The gain control within the MC1350P is accomplished by diverting signal current from the first amplifier stage into a *current sink*. This is also known as the *Gilbert cell multiplier* architecture. Another chip of this type is the NE/SA5209. This type of approach is simpler to implement than discrete circuit approaches, such as dual-gate MOSFETs that are now being replaced by IC designs.

Fig 12.27D shows the high performance National Semiconductor LMH6502MA (14-pin DIP plastic package) voltage controlled amplifier. It is specially designed for accurate log-linear AGC from 0 to 40 dB with respect to a preset maximum voltage gain from 6 to 40 dB. Its ±3 dB bandwidth is 130 MHz. It is an excellent IF amplifier for high performance receiver or transmitter projects.

Additional info on voltage-controlled amplifier ICs can be found on the Analog Devices web site (**www.analog.com**). Search the site for Tutorial MT-073, which describes the operation of various types of gain-controlled amplifiers with numerous product examples.

12.6 Pulse Noise Reduction

A major problem for those listening to receivers has historically been local impulse noise. For HF and VHF receivers it is often from the sparks of internal combustion engine spark plugs, electric fence chargers, light dimmers, faulty power-line insulators and many other similar devices that put out short duration wide band signals. In the UHF and microwave region, radar systems can cause similar problems. There have been three general methods of attempting to deal with such noise over the years, some more successful than others. We will briefly describe the approaches.

12.6.1 The Noise Limiter

The first device used in an early (1930s) attempt to limit impulse noise was called a *noise limiter* or *clipper* circuit as originally described by H. Robinson, W3LW. (see references) This circuit would *clip* or limit noise (or signal) peaks that exceeded a preset limit. The idea was to have the limit set to about as loud as you wanted to hear anything and nothing louder would get through. This was helpful in eliminating the loudest part of impulse noise or even nearby lightning crashes, but it had two problems. First it didn't eliminate the noise, it just reduced the peak loudness; second, it also reduced the loudness of loud non-noise signals and in the process distorted them considerably.

The second problem was fixed shortly thereafter, with the advent of the *automatic noise limiter* or ANL as described by

J. Dickert (see references). The ANL automatically set the clipping threshold to that of a loud signal. It thus would adjust itself as the loudness of signals you listened to changed with time. An ANL was fairly easy to implement and became standard equipment on amateur receivers from the late 1930s on. While ANL circuits are no longer common, simple receivers used today do sometimes incorporate passive clipping circuits to account for their limited AGC ability.

12.6.2 The Noise Blanker

It turned out that improvements in receiver selectivity over the 1950s and beyond, while improving the ability to reduce random noise, actually made receiver response to impulse noise worse. The reason for this is that a very short duration pulse will actually be lengthened while going through a narrow filter. This is due to the filter's different delay times for the pulse's wide spectrum of components, resulting in the components arriving at the filter output at different times. You can demonstrate this in your superhet receiver if it has a narrow crystal filter. Find a frequency with heavy impulse noise and switch between wide and narrow filters. If your narrow filter is 500 Hz or less, the noise pulses will likely be more prominent with the narrow filter. DSP filters with their superior group delay performance exhibit less smearing than their analog counterparts.

The noise limiters described previously were all connected at the output of the IF amplifiers and thus the noise had passed most of the selectivity before the limiter and had been widened by the receiver filters. In addition, modern receivers include *automatic gain control* (AGC), a system that reduces the receiver gain in the presence of strong signals to avoid overload of both receiver circuits and ears. In SSB receivers, since signals vary in strength as someone talks, the usual AGC responds quickly to reduce the gain of a strong signal and then slowly increases it if the signal is no longer there. This means that a strong noise pulse may reduce the receiver gain for much longer than it lasts.

The solution — a *noise blanker*. A noise blanker is almost a separate wideband receiver. It takes its input from an early stage in the receiver before much selectivity or AGC has been applied. It amplifies the wideband signal and detects the narrow noise pulses without lengthening them. The still-narrow noise pulses are used to shut off the receiver at a point ahead of the selectivity and AGC, thus keeping the noise from getting to the parts of the receiver at which the pulses would be extended. In other words, the receiver is shut off or *gated* during the noise pulse.

A well-designed noise blanker can be very effective. Instead of just keeping the noise at the level of the signal as a noise limiter does, the noise blanker can actually *eliminate* the noise. If the pulses are narrow enough, the loss of desired signal during the time the receiver is disabled is not noticeable and the noise may seem to disappear entirely.

In addition to an ON/OFF switch, many noise blanker designs include a control labeled THRESHOLD. The THRESHOLD control adjusts the level of noise that will blank the receiver. If it is set for too low a level, it will blank on signal peaks as well as noise, resulting in distortion of the signal. The usual approach is to turn on the blanker, then adjust the THRESHOLD control until the noise is just blanked. Don't forget to turn it off when the noise goes away.

Noise blankers can also create problems. The wide-band receiver circuit that detects the noise pulses detects any signals in that bandwidth. If such a signal is strong and has sharp peaks (as voice and CW signals do), the noise blanker will treat them as noise pulses and shut down the receiver accordingly. This causes tremendous distortion and can make it sound as if the strong signal to which the noise blanker is responding is generating spurious signals that cause the distortion. Before you assume that the strong signal is causing problems, turn the noise blanker on and off to check. When the band is full of strong signals, a noise blanker may cause more problems than it solves.

12.6.3 Operating Noise Limiters and Blankers

Many current receivers include both a noise limiter and a noise blanker. If your receiver has both, they will have separate controls and it is worthwhile to try them both. There are times at which one will work better than the other, and other times when it goes the other way, depending on the characteristics of the noise. There are other times when both work better than either. In any case, they can make listening a lot more pleasant — just remember to turn them off when you don't need them since either type can cause some distortion, especially on strong signals that should otherwise be easy to listen to.

Recognizing that it is difficult for a single noise blanker to work properly with the wide variations of noise pulses, it is common for late-model receivers to have two noise blankers with different characteristics that are optimized for the different pulse types. One noise blanker is typically optimized for very short pulses and the other for longer pulses. The operator can switch between the blankers to see which works best on the noise at hand.

OTHER TECHNIQUES

The previous techniques represent the most commonly available techniques to reduce impulse noise. There are a few other solutions as well. Note that we haven't been talking about reducing interference here. By interference, we mean another intended signal encroaching on the channel to which we want to listen. There are a number of techniques to reduce interference, and some also can help with impulse noise.

Many times impulse noise is coming from a particular direction. If so, by using a directional antenna, we can adjust the direction for minimum noise. When we think about directional antennas, the giant HF Yagi springs to mind. For receiving purposes, especially on the lower bands such as 160, 80 and 40 meters (where the impulse noise often seems the worst), a small indoor or outdoor receiving loop antenna as described in the *ARRL Antenna Book* can be very effective at eliminating either interfering stations or noise (both if they happen to be in the same direction).

Another technique that can be used to eliminate either interference or noise is to obtain a copy of the noise (or interference) that is 180° out of phase from the one you are receiving. By adjusting the amplitude to match the incoming signal, the signal can be cancelled at the input to the receiver. Several available commercial units perform this task.

Digital signal processing, described in the next section, is another multifunction system that can help with all kinds of noise.

12.6.4 DSP Noise Reduction

DSP noise reduction can actually look at the statistics of the signal and noise and figure out which is which and then reduce the noise significantly. These *adaptive filters* can't quite eliminate the noise, and need enough of the desired signal to figure out what's happening, so they won't work if the signal is far below the noise. Many DSP systems "color" the resulting audio to a degree. Nonetheless, they do improve the SNR of a signal in random or impulse noise. As with noise blankers, receivers frequently offer two or more noise reduction settings that apply different noise reduction algorithms optimized for different conditions. It's always worth experimenting with the radio's features to find out which work better. The **DSP and Software Radio Design** chapter discusses these features in more detail.

12.7 VHF and UHF Receivers

Most of the basic ideas presented in previous sections apply equally well in receivers that are intended for the VHF and UHF bands. This section will focus on the difference between VHF/UHF and HF receivers.

12.7.1 FM Receivers

Narrow-band frequency modulation (NBFM) is the most common mode used on VHF and UHF. **Fig 12.28A** is a block diagram of an FM receiver for the VHF/UHF amateur bands.

FRONT END

A low-noise front end is desirable because of the decreasing atmospheric noise level at these frequencies and also because portable gear often uses short rod antennas at ground level. Nonetheless, the possibilities for gain compression and harmonic IMD, multi-tone IMD and cross modulation are also substantial. Therefore dynamic range is an important design consideration, especially if large, high-gain antennas are used. FM limiting should not occur until after the crystal filter. Because of the high occupancy of the VHF/UHF spectrum by powerful broadcast transmitters and nearby two-way radio services, front-end preselection is desirable, so that a

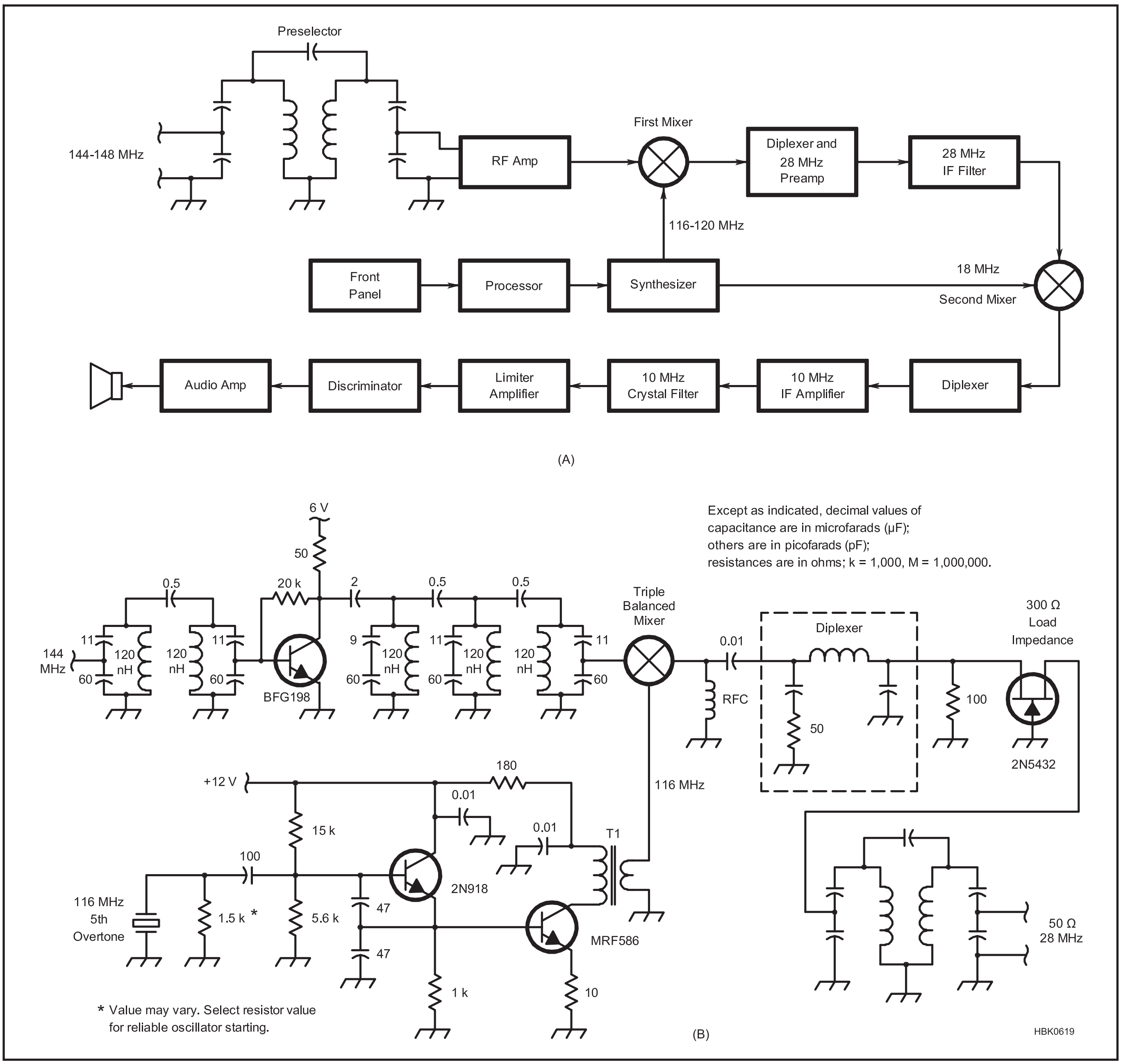


Fig 12.28 — At A, block diagram of a typical VHF FM receiver. At B, a 2 meter to 10 meter receive converter (partial schematic; some power supply connections omitted.)

low noise figure can be achieved economically within the amateur band.

DOWN-CONVERSION

Down-conversion to the final IF can occur in one or two stages. Favorite IFs are in the 5 to 10 MHz region, but at the higher frequencies rejection of the image 10 to 20 MHz away can be difficult, requiring considerable preselection. At the higher frequencies an intermediate IF in the 30 to 50 MHz region is a better choice. Fig 12.28A shows dual down-conversion.

IF FILTERS

The customary peak frequency deviation in amateur FM on frequencies above 29 MHz is about 5 kHz and the audio speech band extends to 3 kHz. This defines a maximum modulation index (defined as the deviation ratio) of 5/3 = 1.67. An inspection of the Bessel functions that describe the resulting FM signal shows that this condition confines most of the 300 to 3000 Hz speech information sidebands within a 15 kHz or so bandwidth. Using filters of this bandwidth, channel separations of 20 or 25 kHz are achievable.

Many amateur FM transceivers are channelized in steps that can vary from 1 to 25 kHz. For low distortion of the audio output (after FM detection), this filter should have good phase linearity across the bandwidth. This would seem to preclude filters with very steep descent outside the passband, which tend to have very nonlinear phase near the band edges. But since the amount of energy in the higher speech frequencies is naturally less, the actual distortion due to this effect may be acceptable for speech purposes. The normal practice is to apply pre-emphasis to the higher speech frequencies at the transmitter and de-emphasis compensates at the receiver.

LIMITING

After the filter, hard limiting of the IF is needed to remove any amplitude modulation components. In a high-quality receiver, special attention is given to any nonlinear phase shift that might result from the limiter circuit design. This is especially important in data receivers in which phase response must be controlled. In amateur receivers for speech it may be less important. Also, the *ratio detector* (see the **Mixers, Modulators and Demodulators** chapter) largely eliminates the need for a limiter stage, although the limiter approach is probably still preferred.

FM DETECTION

The discussion of this subject is deferred to the **Mixers, Modulators and Demodulators** chapter. Quadrature detection is used in some popular FM multistage ICs. An example receiver IC will be presented later.

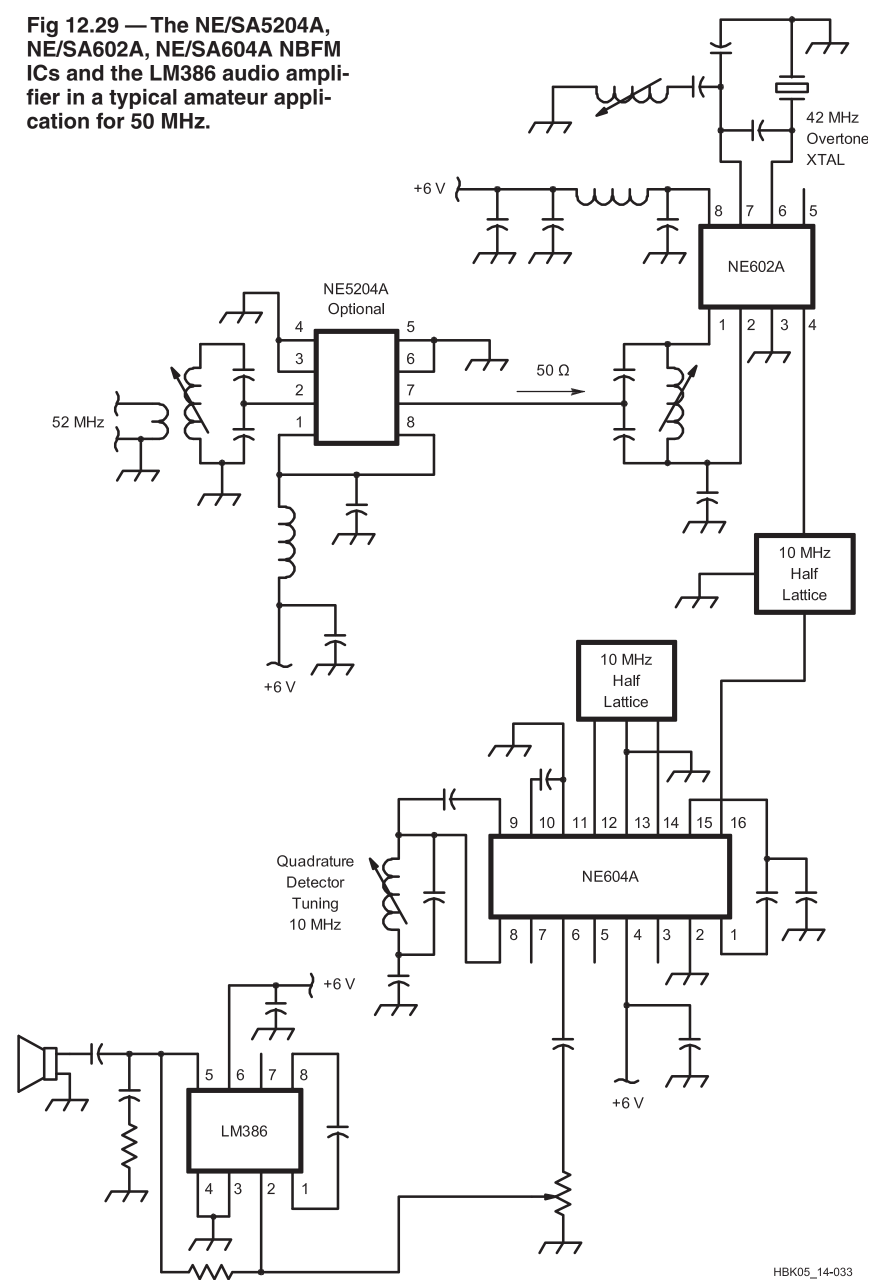
12.7.2 FM Receiver Weak-Signal Performance

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

LO PHASE NOISE

In an FM receiver, LO phase noise superimposes phase modulation, and therefore frequency modulation, onto the desired signal. This reduces the ultimate signal-to-noise ratio within the passband. This effect is called "incidental FM (IFM)." The power density of IFM (W/Hz) is proportional to the phase noise power density (W/Hz) multiplied by the square of the modulating frequency (the familiar parabolic effect in FM). If the receiver uses high-frequency de-emphasis at the audio output (–6 dB per octave from 300 to 3000 Hz, a common practice), the IFM level at higher audio frequencies can be reduced. Ordinarily, as the signal increases the

Fig 12.29 — The NE/SA5204A, NE/SA602A, NE/SA604A NBFM ICs and the LM386 audio amplifier in a typical amateur application for 50 MHz.



noise would be "quieted" (that is, "captured") in an FM receiver, but in this case the signal and the phase noise riding "piggy back" on the signal increase in the same proportion as described in the **Oscillators and Synthesizers** chapter's discuss of reciprocal mixing. IFM is not a significant problem in modern FM radios, but phase noise can become a concern for adjacent-channel interference.

As the signal becomes large the signal-to-noise ratio therefore approaches some final value. A similar ultimate SNR effect occurs in SSB receivers. On the other hand, a perfect AM receiver tends to suppress LO phase noise. (See the reference entry for Sabin.)

12.7.3 FM Receiver ICs

A wide variety of special ICs for communications-bandwidth FM receivers are available. Many of these were designed for "cordless" or mobile telephone applications and are widely used. **Fig 12.29** shows some popular versions for a 50 MHz FM receiver. One is an RF amplifier chip (NE/SA5204A) for 50 Ω input to 50 Ω output with 20 dB of gain. The second chip (NE/SA602A) is a front-end device with an RF amplifier, mixer and LO. The third is an IF amplifier, limiter and quadrature FM detector (NE/SA604A) that also has a very useful RSSI (logarithmic Received Signal Strength Indicator) output and also a "mute" function. The fourth is the LM386, a widely used audio-amplifier chip. Another FM receiver chip, complete in one package, is the MC3371P.

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

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

The diagram in Fig 12.29 is (intentionally) only a general outline that shows how chips can be combined to build complete equipment. The design details and specific parts values can be learned from a careful study of the data sheets and application notes provided by the IC vendors. Amateur designers should learn how to use these data sheets and other information such as application notes available (usually for free) from the manufacturers or on the web.

12.7.4 VHF Receive Converters

Rather than building an entire transceiver for VHF SSB and CW, one approach is to use a receive converter. A receive converter (also called a *downconverter*) takes VHF signals and converts them to an HF band for reception using existing receiver or transceiver as a tunable IF.

Although many commercial transceivers cover the VHF bands (either multiband, multimode VHF/UHF transceivers, or HF+VHF transceivers), receive converters are sometimes preferred for demanding applications because they may be used with high-performance HF transceivers. Receive converters are often packaged with a companion transmit converter and control circuitry to make a *transverter*.

A typical 2 meter downconverter uses an IF of 28-30 MHz. Signals on 2 meters are amplified by a low-noise front-end before mixing with a 116 MHz LO. Fig 12.28B shows the schematic for a high-performance converter. The front-end design was contributed by Ulrich Rohde, N1UL, who recommends a triple-balanced mixer such as the Synergy CVP206 or SLD-K5M.

The diplexer filter at the mixer output selects the difference product: (144 to 146 MHz) – 116 = (28 to 30 MHz). A common-base buffer amplifier (the 2N5432 FET) and tuned filter form the input to the 10 meter receiver. (N1UL suggests that using an IF of 21 MHz and an LO at 165 MHz would avoid interference problems with 222 MHz band signals.) For additional oscillator designs, refer to the papers on oscillators by N1UL in the supplemental CD-ROM files for the **Mixers, Modulators and Demodulators** chapter accompanying this *Handbook*.

Based on the Philips BFG198 8 GHz transistor, the 20 dB gain front-end amplifier is optimized for noise figure (NF is approximately 2.6 dB), not for input impedance. The output circuit is optimized for best selectivity. The transistor bias is designed for dc stability at I_C = 30 mA and V_C = 6 V. Both of the transistor's emitter terminals should be grounded to prevent oscillation. NF might be improved with a higher performance transistor, such as a GaAs FET, but stability problems are often encountered with FET designs in this application.

If a mast-mounted preamplifier is used to improve the system noise figure, an attenuator should be available to prevent overload. Simulation predicts the circuit to have an IP3 figure of at least +25 dBm at 145 MHz with an I_C of 30 mA and a terminating impedance of 50 Ω.

12.8 UHF and Microwave Techniques

The ultra high frequency spectrum comprises the range from 300 MHz to 3 GHz. All of the basic principles of radio system design and circuit design that have been discussed so far apply as well in this range, but the higher frequencies require some special thinking about the methods of circuit design and the devices that are used. Additional material on construction for microwave circuits can be found in the **Construction Techniques** chapter and in the series of *QST* columns, "Microwavelengths" by Paul Wade, W1GHZ.

12.8.1 UHF Construction

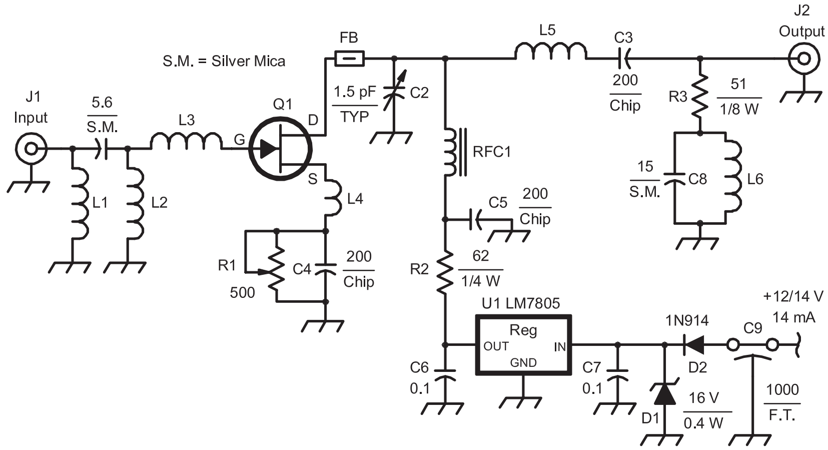
Modern receiver designs make use of highly miniaturized monolithic microwave ICs (MMICs). Among these are the Avago MODAMP and the Mini Circuits MAR and MAV/ERA lines. They come in a wide variety of gains, intercepts and noise figures for frequency ranges from dc to well into the GHz range. (See the **Component Data and References** chapter for information on available parts.)

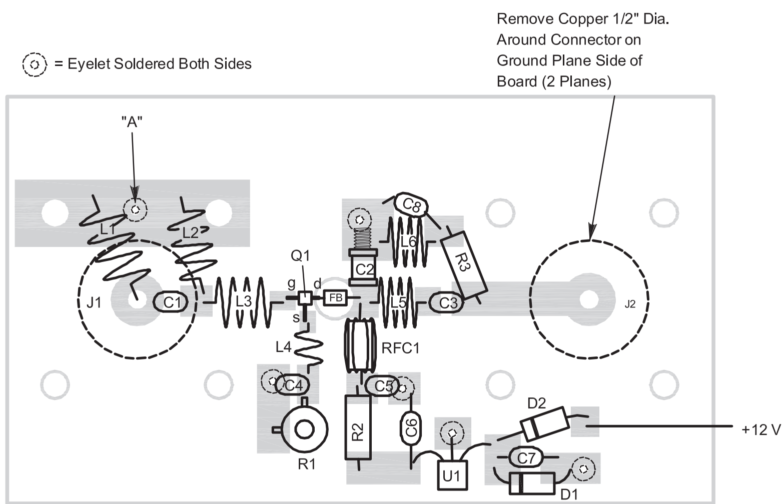
Fig 12.30 shows the schematic diagram and the physical construction of a typical RF circuit at 430 MHz. It is a GaAsFET preamplifier intended for low noise SSB/CW, moonbounce or satellite reception. The construction uses ceramic chip capacitors, small helical inductors and a stripline surface-mount GaAsFET, all mounted on a G10 (two layers of copper) glass-epoxy PC board. The very short length of interconnection leads is typical. The bottom of the PC board is a ground plane. At this frequency, lumped components are still feasible, while microstrip circuitry tends to be rather large.

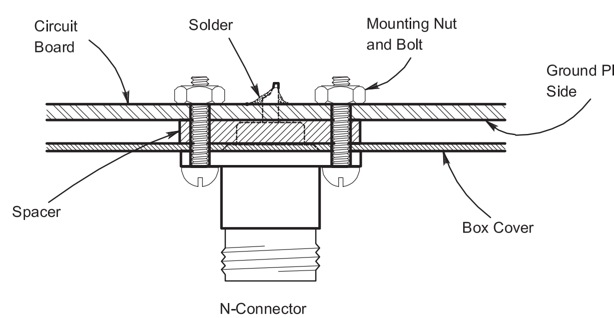
At higher frequencies, microstrip methods become more desirable in most cases because of their smaller dimensions. However, the advent of tiny chip capacitors and chip resistors has extended the frequency range of discrete components. For example, the literature shows methods of building LC filters at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

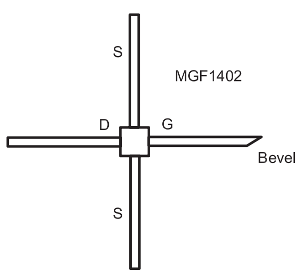
Fig 12.30 — GaAsFET preamplifier schematic and construction details for 430 MHz. Illustrates circuit, parts layout and construction techniques suitable for 430-MHz frequency range.

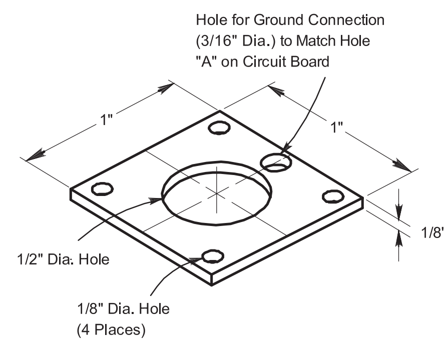
C1 — 5.6 pF silver-mica or same as C2.
C2 — 0.6 to 6 pF ceramic piston trimmer (Johanson 5700 series or equiv).
C3, C4, C5 — 200 pF ceramic chip.
C6, C7 — 0.1 µF disc ceramic, 50 V or greater.
C8 — 15 pF silver-mica.
C9 — 500 to 1000 pF feedthrough.
D1 — 16 to 30 V, 500 mW Zener (1N966B or equiv).
D2 — 1N914, 1N4148 or any diode with ratings of at least 25 PIV at 50 mA or greater.
J1, J2 — Female chassis-mount Type-N connectors, PTFE dielectric (UG-58 or equiv).
L1, L2 — 3t, #24 tinned wire, 0.110-inch ID spaced 1 wire dia.
L3 — 5t, #24 tinned wire, 3/16-inch ID, spaced 1 wire dia. or closer. Slightly larger diameter (0.010 inch) may be required with some FETs.
L4, L6 — 1t #24 tinned wire, ⅛-inch ID.
L5 — 4t #24 tinned wire, ⅛-inch ID, spaced 1 wire dia.
Q1 — Mitsubishi MGF1402.
R1 — 200 or 500-Ω Cermet potentiometer (initially set to midrange).
R2 — 62 Ω, ¼ W.
R3 — 51 Ω, ⅛ W carbon composition resistor, 5% tolerance.
RFC1 — 5t #26 enameled wire on a ferrite bead.
U1 — 5 V, 100-mA 3 terminal regulator (LM78L05 or equiv. TO-92 package).



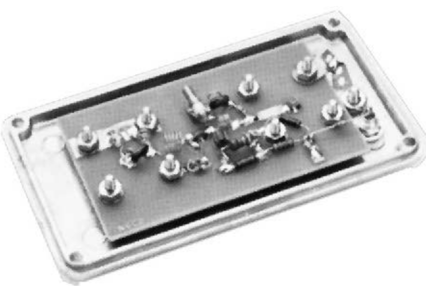








Aluminum Spacer for N-Connector



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mixer circuits operate at well into the GHz range using these types of components on controlled-dielectric PC board material such as Duroid or on ceramic substrates.

Current designs emphasize simplicity of construction and adjustment, leading to "no tune" designs. The use of printed-circuit microstrip filters that require little or no adjustment, along with IC or MMIC devices, or discrete transistors, in precise PC-board layouts that have been carefully worked out, make it much easier to "get going" on the higher frequencies.

12.8.2 UHF Design Aids

Circuit design and evaluation at the higher frequencies usually require some kind of minimal lab facilities, such as a signal generator, a calibrated noise generator and, hopefully, some kind of simple (or surplus) spectrum analyzer. This is true because circuit behavior and stability depend on a number of factors that are difficult to "guess at," and intuition is often unreliable. The ideal instrument is a vector network analyzer with all of the attachments (such as an S parameter measuring setup), an instrument that has become surprisingly affordable in recent years. (See the **Test Equipment and Measurements** chapter.)

Another very desirable thing would be a circuit design and analysis program for the personal computer. Software packages created especially for UHF and microwave circuit design are available. They tend to be somewhat expensive, but worthwhile for a serious designer. Inexpensive *SPICE* programs are a good compromise but have significant limitations at VHF and above. See the chapter on **Computer-Aided Circuit Design** for information on these tools.

12.8.3 A 902 to 928 MHz (33 cm) Receiver

This 902 MHz downconverter is a fairly typical example of receiver design methods for the 500 to 3000 MHz range, in which down-conversion to an existing HF receiver (or 2 meter multimode receiver) is the most convenient and cost-effective approach for most amateurs. At higher frequencies a double down-conversion with a first IF of 200 MHz or so, to improve image rejection, might be necessary. Usually, though, the presence of strong signals at image frequencies is less likely. Image-reducing mixers plus down-conversion to 28 MHz is also coming into use, when strong interfering signals are not likely at the image frequency.

Fig 12.31A is the block diagram of the 902 MHz down-converting receiver. A cavity resonator at the antenna input provides high selectivity with low loss. The first RF amplifier is a GaAsFET. Two additional 902 MHz band-pass microstrip filters and a second RF amplifier transistor provide more gain and image rejection (at RF – 56 MHz) for the mixer. The output is at 28.0 MHz so that an HF receiver can be used as a tunable IF/demodulator stage.

CUMULATIVE NOISE FIGURE

Fig 12.31B shows the cumulative noise figure (NF) of the signal path, including the 28 MHz receiver. The 1.5 dB cumulative NF of the input cavity and first RF-amplifier combination, considered by itself, is degraded to 1.9 dB by the rest of the system following the first RF amplifier. The NF values of the various components for this example are reasonable, but may vary somewhat for actual hardware. Also, losses prior to the input such as transmission line losses (very important) are not included. They would be part of the complete receive system analysis, however. It is common practice to place a low noise preamp outdoors, right at the antenna, to overcome coax loss (and to permit use of less expensive coax).

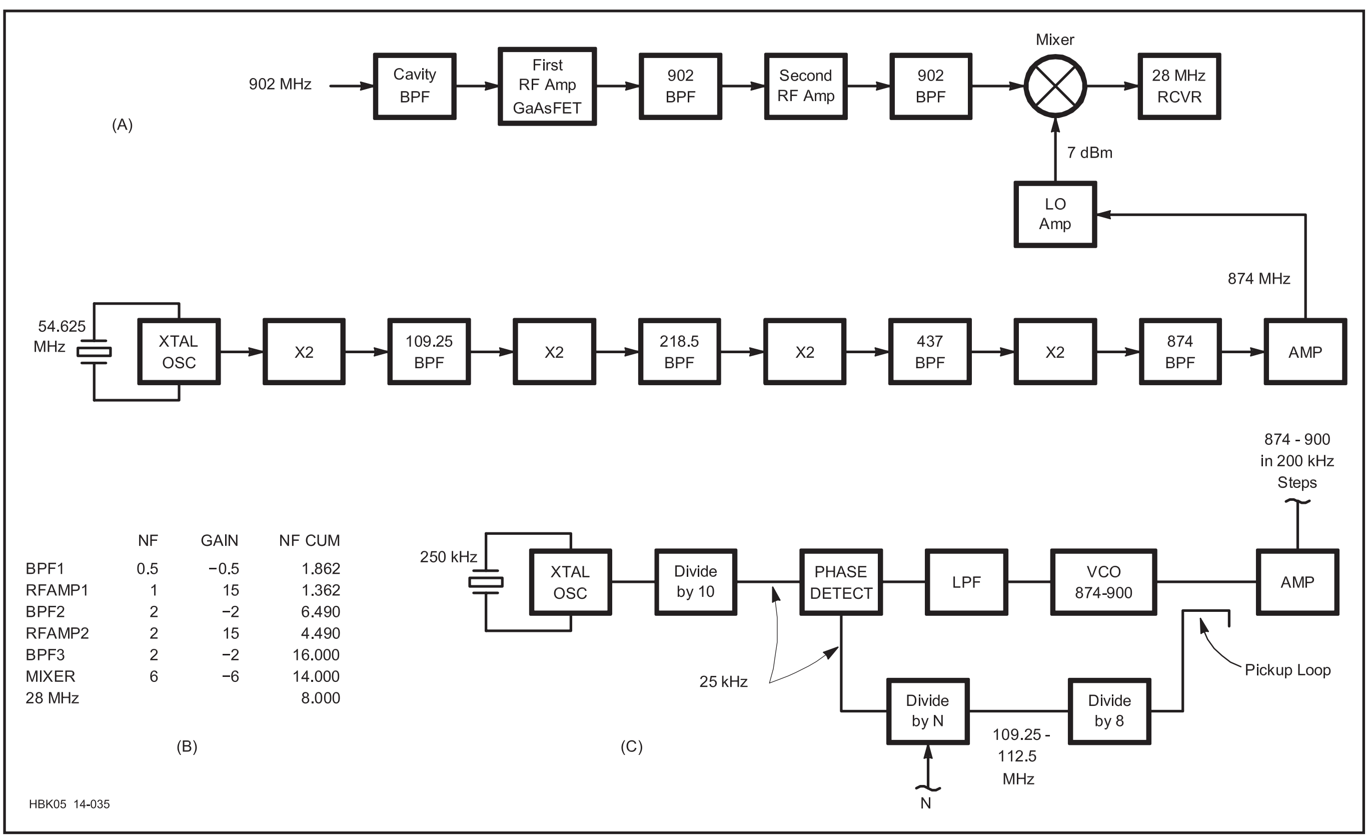

Fig 12.31 — A downconverter for the 902 to 928 MHz band. At A: block diagram; At B: cumulative noise figure of the signal path; At C: alternative LO multiplier using a phase locked loop.

LOCAL OSCILLATOR (LO) DESIGN

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

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

A phase-locked loop using GHz range prescalers (as shown in Fig 12.31C) is an alternative to the multiplier chain. The divide-by-N block is a simplification; in practice, an auxiliary dual-modulus divider (see the **Oscillators and Synthesizers** chapter) would be involved in this segment. The cascaded 902 MHz band-pass filters in the signal path should attenuate any image frequency noise (at RF–56 MHz) that might degrade the mixer noise figure.

12.8.4 Microwave Receivers

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

A 10 GHZ PREAMPLIFIER

Fig 12.32B is a schematic and parts list, Fig 12.32C is a PC board parts layout and Fig 12.32A is a photograph of a 10 GHz preamp, designed by Senior ARRL Lab Engineer Zack Lau, W1VT. With very careful design and packaging techniques a noise figure approaching the 1 to 1.5 dB range was achieved. This depends on an accurate 50-Ω generator impedance and noise matching the input using a microwave circuit-design program such as *Touchstone* or *Harmonica*. Note that microstrip capacitors, inductors and transmission-line segments are used almost exclusively. The circuit is built on a 15-mil Duroid PC board. In general, this kind of performance requires some elegant measurement equipment that few amateurs have. On the other hand, preamp noise figures in the 2 to 4-dB range are much easier to get (with simple test equipment) and are often satisfactory for amateur terrestrial communication.

Articles written by those with expertise and the necessary lab facilities almost always include PC board patterns, parts lists and detailed instructions that are easily duplicated by readers. Microwave ham clubs and their publications are a good way to get started in microwave amateur technology.

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

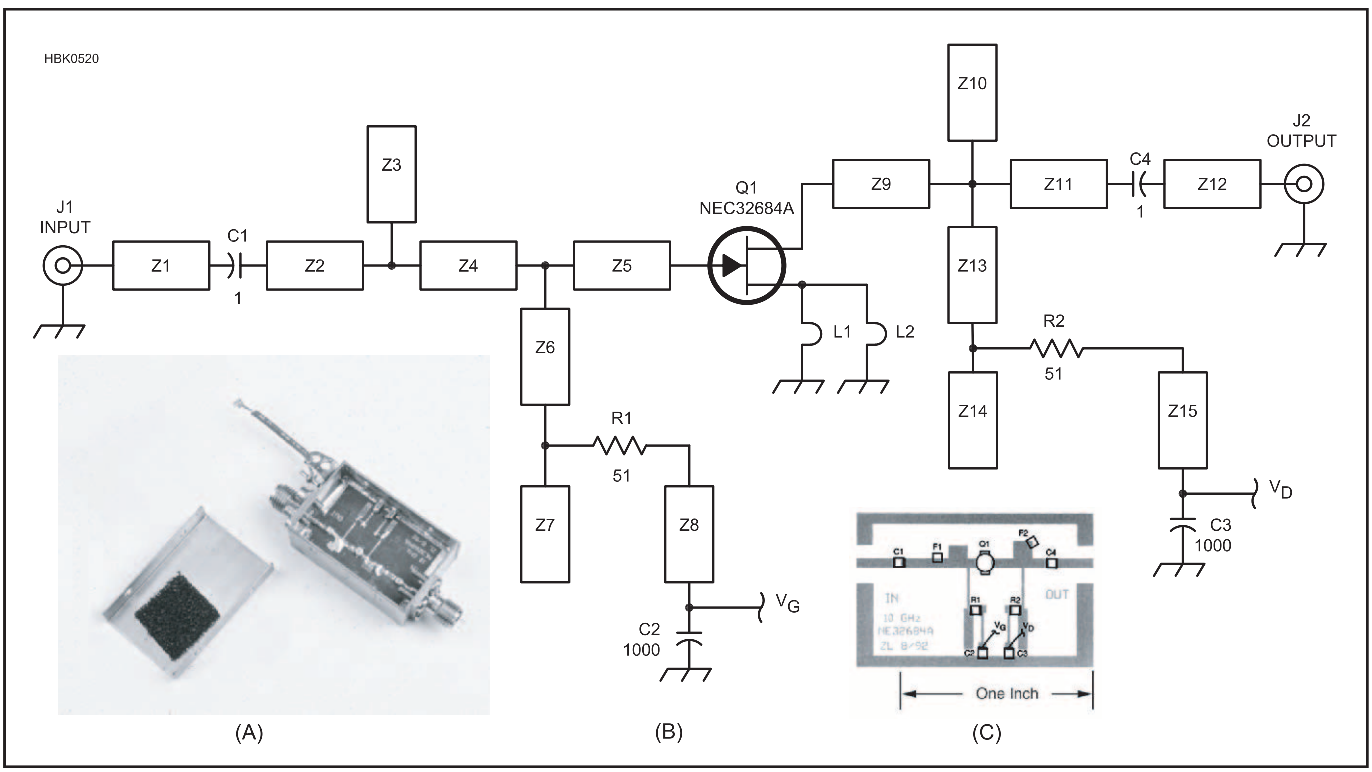

Fig 12.32 — At A, a low-noise preamplifier for 10 GHz, illustrating the methods used at microwaves. At B: schematic. At C: PC board layout. Use 15-mil 5880 Duroid, dielectric constant of 2.2 and a dissipation factor of 0.0011. A template of the PC board is available on the CD-ROM included with this book.

C1, C4 — 1 pF ATC 100 A chip capacitors. C1 must be very low loss.
C2, C3 — 1000 pF chip capacitors. (Not critical.) The ones from Mini Circuits work fine.
F1, F2 — Pieces of copper foil used to tune the preamp.
J1, J2 — SMA jacks. Ideally these should be microstrip launchers. The pin should be flush against the board.
L1, L2 — The 15 mil lead length going through the board to the ground plane.
R1, R2 — 51 Ω chip resistors.
Z1-Z15 — Microstrip lines etched on the PC board.

Analysis of a 10.368 GHz communication link with SSB modulation:

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

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

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

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

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

To find the system noise factor F_S we note that the system noise is proportional to the system temperature T_S and the "generator" (antenna) noise is proportional to the antenna temperature T_A. Using the idea of a "system noise factor":

$F_S = T_S / T_A$ = 2217.6 / 200 = 11.09 = 10.45 dB.

If the antenna temperature were 290 K the system noise figure would be 9.0 dB, which is precisely the sum of receiver and receiver coax noise figures (6.0 + 3.0).

Fig 12.33 — Example of a 10-GHz system performance calculation. Noise temperature and noise factor of the receiver are considered in detail.

System Performance

At microwaves, an estimation of system performance can often be performed using known data about the signal path terrain, atmosphere, transmitter and receivers systems. In the present context of receiver design we wish to establish an approximate goal for the receiver system, including the antenna and transmission line. **Fig 12.33** shows a simplified example of how this works.

A more detailed analysis includes terrain variations, refraction effects, the Earth's curvature, diffraction effects and interactions with the atmosphere's chemical constituents and temperature gradients.

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

$$T_E = 290\,(F - 1) \tag{8}$$

where the noise factor $F = 10^{NF/10}$ and NF is the noise figure in dB.

T_E is a measure, in terms of temperature, of the "excess noise" of a component (such as an amplifier). A resistor at T_E would have the same available noise power as the device (referred to the device's input) specified by T_E. For a passive device (such as a lossy transmission line or filter) that introduces no noise of its own, T_E is zero and G is a number less than one equal to the power loss of the device. The cascade of noise temperatures is similar to the formula for cascaded noise factors.

$$T_S = T_G + TE_1 + \frac{TE_1}{G_1} + \frac{TE_3}{G_1G_2} + \frac{TE_4}{G_1G_2G_3} + \ldots \tag{9}$$

where T_S is the system noise temperature (including the generator, which may be an antenna) and T_G is the noise temperature of the generator or the field of view of the antenna, usually assumed 290 °K for terrestrial communications.

The number 290 in the formulas for T_E is the standard ambient temperature (in kelvins) at which the noise factor of a two-port transducer is defined and measured, according to an IEEE recommendation. So those formulas relate a noise factor F, measured at 290 K, to the temperature T_E. In general, though, it is perfectly correct to say that the ratio $(S_I/N_I)/(S_O/N_O)$ can be thought of as the ratio of total system output noise to that system output noise attributed to the "generator" alone, regardless of the temperature of the equipment or the nature of the generator, which may be an antenna at some arbitrary temperature, for example. This ratio is, in fact, a special "system noise factor (or figure), F_S" that need not be tied to any particular temperature such as 290 K. (Note that regular noise factor (or figure) does depend on reference temperature.) The use of the F_S notation avoids any confusion. As the example of Fig 12.33 shows, the value of this system noise factor F_S is just the ratio of the total system temperature to the antenna temperature.

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

$$N = k\,T_S\,B_N \tag{10}$$

where
k = 1.38×10^{-23} (Stefan-Boltzmann's constant) and
B_N= noise bandwidth

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

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

MICROWAVE RECEIVER FOR 10 GHZ

Here is a good example of amateur techniques for the 10 GHz band. The intended use for the radio is narrowband CW and SSB work, which requires extremely good frequency stability in the LO. Here, we will discuss the receiver circuit.

Block Diagram

Fig 12.34 is a block diagram of the receiver. Here are some important facets of the design.

1) The antenna should have sufficient gain. At 10 GHz, gains of 30 dBi are not difficult to get, as the example of Fig 12.33 demonstrates. A 4-foot dish might be difficult to aim, however.

2) For best results a very low-noise preamp at the antenna reduces loss of system sensitivity when antenna temperature is low. For example, if the antenna temperature at a quiet direction of the sky is 50 K and the receiver noise figure is 4 dB (due in part to transmission-line loss), the system temperature is 488 K for a system noise figure of 4.3 dB. If the receiver noise figure is reduced to 1.5 dB by adding a preamp at the antenna the system temperature is reduced to 170 K for a system noise figure of 2.0 dB, which is a very big improvement.

3) After two stages of RF amplification using GaAsFETs, a probe-coupled cavity resonator attenuates noise at the mixer's image frequency, which is 10.368 – 0.288

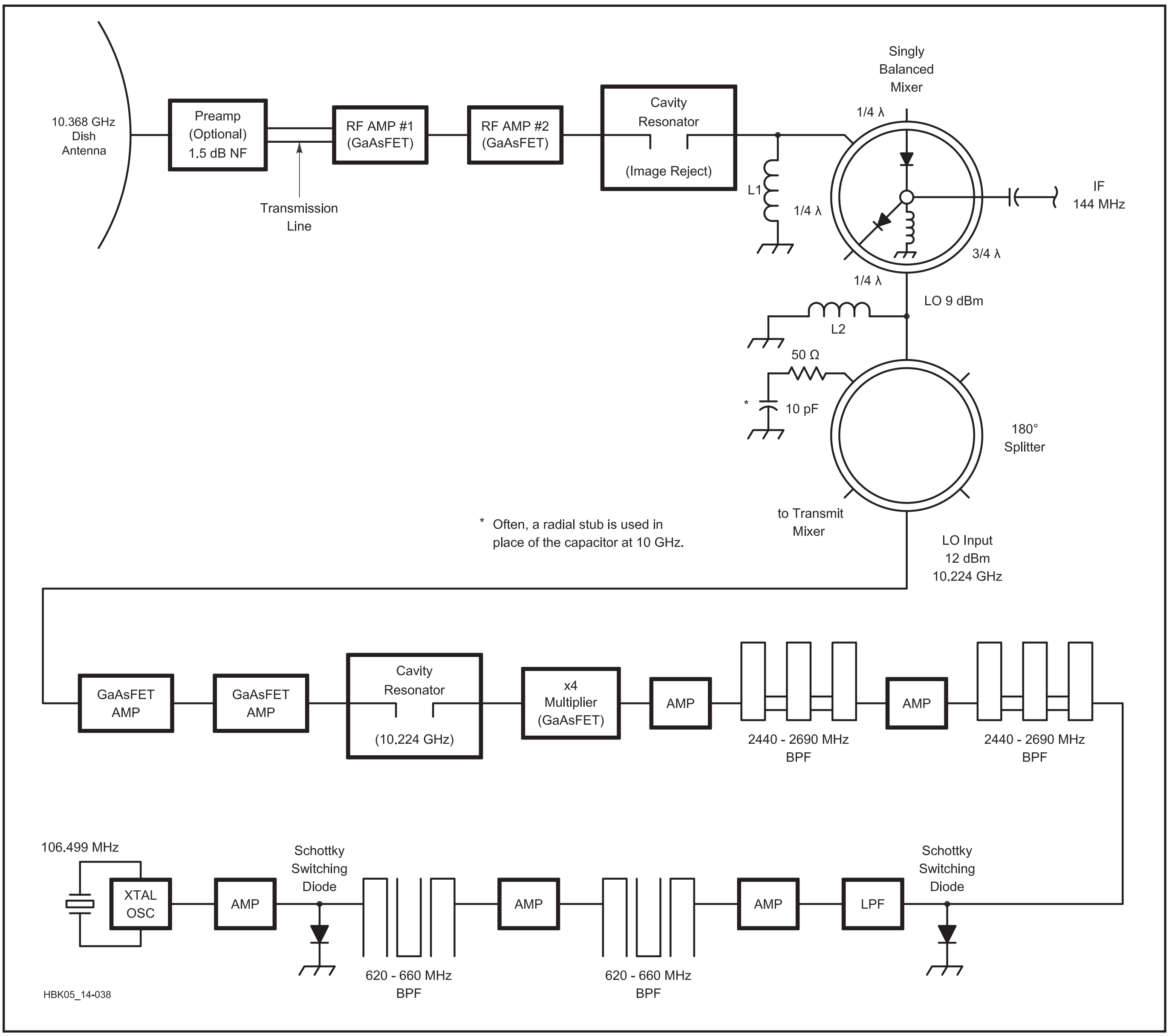

Fig 12.34 — A block diagram of the microwave receiver discussed in the text.

= 10.080 GHz. An image reduction of 15 to 20 dB is enough to prevent image frequency noise generated by the RF amplifiers from affecting the mixer's noise figure.

4) The single-balanced diode mixer uses a "rat-race" 180° hybrid. Each terminal of the ring is ¼ wavelength (90°) from its closest neighbors. So the anodes of the two diodes are 180° (½ wavelength) apart with respect to the LO port, but in-phase with respect to the RF port. The inductors (L1, L2) connected to ground present a low impedance at the IF frequency. The mixer microstrip circuit is carefully "tweaked" to improve system performance. Use the better mixer in the transmitter.

5) The crystal oscillator is a fifth-overtone Butler circuit that is capable of high stability. The crystal frequency error and drift are multiplied 96 times (10.224/0.1065), so for narrowband SSB or CW work it may be difficult to get on (and stay on) the "calling frequency" at 10.368 GHz. One acceptable (not perfect) solution might be to count the 106.5 MHz with a frequency counter whose internal clock is constantly compared with WWV. Adjust to 106.5 MHz as required. At times there may be a small Doppler shift on the WWV signal. It may be necessary to switch to a different WWV frequency, or WWV's signals may not be strong enough. Surplus frequency standards of high quality are sometimes available. Many operators just "tune" over the expected range of uncertainty.

6) The frequency multiplier chain has numerous band-pass filters to "purify" the harmonics by reducing various frequency components that might affect the signal path and cause spurious responses. The final filter is a tuned cavity resonator that reduces spurs from previous stages. Oscillator phase noise amplitude is multiplied by 96 also, so the oscillator must have very good short-term stability to prevent contamination of the desired signal.

7) A second hybrid splitter provides an LO output for the transmitter section of the radio. The 50-Ω resistor improves isolation between the two output ports. The original two-part *QST* article (see the references) is recommended reading for this very interesting project, which provides a fairly straightforward (but not extremely simple) way to get started on 10 GHz.

12.9 References and Bibliography

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