

# Contents

12.1 Introduction	12.6 Pulse Noise Reduction
12.1.1 Characterizing Receivers	12.6.1 The Noise Limiter
12.1.2 Elements of Early Receivers	12.6.2 The Noise Blanker
12.2 Basics of Heterodyne Receivers	12.6.3 Operating Noise Limiters and Blankers
12.2.1 The Direct Conversion Receiver	12.6.4 DSP Noise Reduction
<i>Project: A Rock-Bending Receiver For 7 MHz</i>	
12.3 The Superheterodyne Receiver	12.7 VHF and UHF Receivers
12.3.1 Superhet Bandwidth	12.7.1 FM Receivers
12.3.2 Selection of IF Frequency for a Superhet	12.7.2 FM Receiver Weak-Signal Performance
<i>Project: The MicroR2 — An Easy to Build SSB or CW Receiver</i>	12.7.3 FM Receiver ICs
12.4 Superhet Receiver Design Details	12.7.4 VHF Receive Converters
12.4.1 Receiver Sensitivity	
12.4.2 Receiver Image Rejection	12.8 UHF Techniques
12.4.3 Receiver Dynamic Range	12.8.1 UHF Construction
12.5 Control and Processing Outside the Primary Signal Path	12.8.2 UHF Design Aids
12.5.1 Automatic Gain Control (AGC)	12.8.3 A 902 to 928 MHz (33-cm) Receiver
12.5.2 Audio-Derived AGC	12.8.4 Microwave Receivers
12.5.3 AGC Circuits	12.9 References and Bibliography

# Chapter 12

# Receivers

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.

One of the largest challenges of the receiver is the processing of that *particular* signal. In a real world environment, the antenna will often deliver a multitude of signals with a wide range of amplitudes. The biggest job of the receiver is often to extract the information content of the desired signal from a within large group of signals, with many perhaps much stronger in amplitude than the one we want.

This chapter was written by Joel Hallas, W1ZR, including updating of existing material from previous editions of this book.

## 12.1 Introduction

This first section defines some important receiver concepts that will be used throughout the chapter.

### 12.1.1 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  $\mu$ V 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.

## **12.1.2 Elements of Early Receivers**

A modern receiver is typically composed of multiple stages. The circuits are typically ones that have been presented in other sections of this book combined in ways to provide improved performance of the various key parameters we have discussed.

## THE DETECTOR

The primary function of a receiver takes place in a *detector*, the circuit that extracts the information from the modulated RF signal that arrives from the antenna. As shown in **Fig 12.1**, the simplest receiver consists of only a detector. The earliest receivers were constructed in just that way, starting with some unusual electromechanical marvels from Marconi's days. The crystal detector, in the form of the crystal set, was the first radio experienced by youngsters from the 1930s to the 1980s and was the mainstay of commercial users until vacuum tubes became available. While even simpler configurations are possible, the circuit of **Fig 12.2** is typical.

The key performance parameters of this receiver are easy to describe. The sensitivity is the signal level required at the antenna input to create an audible signal as a consequence of the diode acting as a square-law detector. The selectivity is that determined by the loaded Q of the single-tuned resonant circuit while the dynamic range is from the sensitivity to the point that the headphone diaphragms hit their limits, or the diode opens due to high current.

While this receiver looks like it would be rather limited in performance (and is), it does work and was even the basis for most microwave radar receivers from WWII and for some years thereafter. That was the origin of the venerable 1N34 diode. An example of a crystal set used as the receiver in an on-off keyed data communications system is shown in **Fig 12.3**. The diode detector is still used as the detector for full carrier amplitude modulated signals in more complex modern receivers.

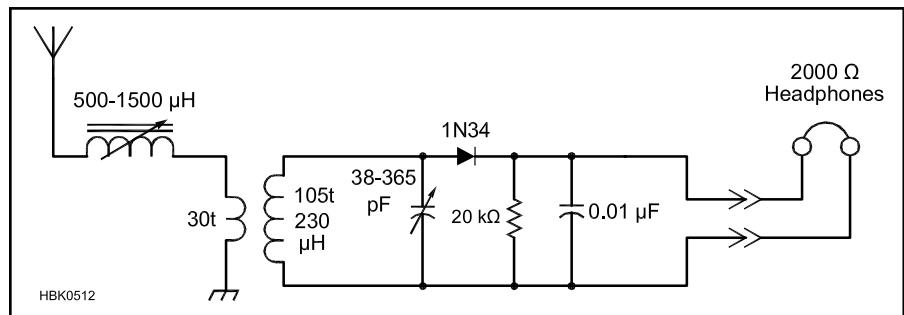
Following early crystal receivers were various forms of detectors that used vacuum tubes, and later transistors, to provide gain

in the signal path. An oscillating detector, called a *regenerative* detector, provided high sensitivity and improved selectivity. It could also be used for the reception of CW signals by increasing the gain to the point of oscillation and tuning it near the frequency of the

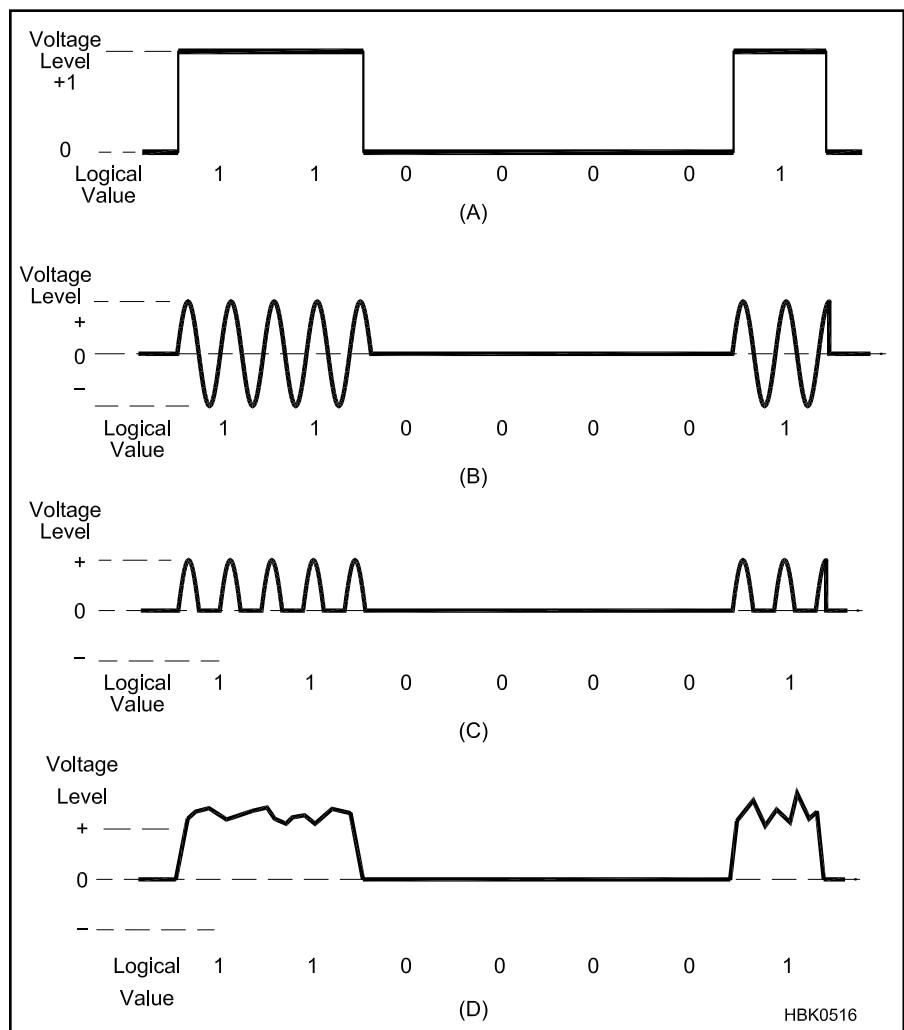
received signal. This is the precursor to the modern product detector to be discussed in a following section.

AMPLIFIERS

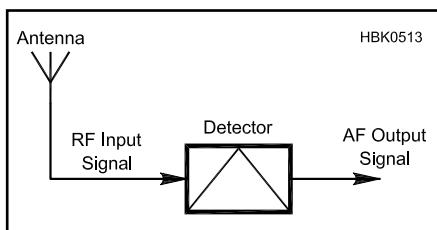
The next element added to the detector



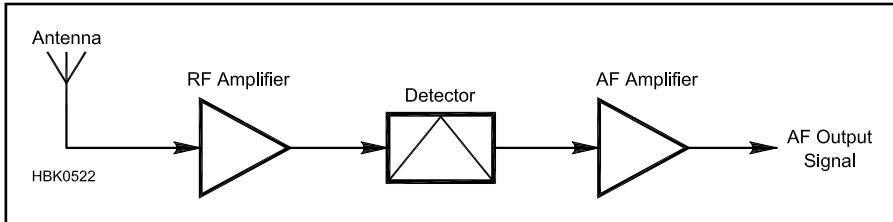
**Fig 12.2 — Circuit of a typical “modern” crystal set with a semiconductor diode as the crystal detector.**



**Fig 12.3 — The signals in a basic on-off keyed data system using a crystal receiver.** At (A), voltage pulses corresponding to the ASCII representation of the lower case letter "a." At (B), The signals at the output of an on-off keyed radio transmitter. At (C), received pulses at output of detector in the absence of any smoothing. At (D) pulse output to processor following smoothing by shunt capacitor.



**Fig 12.1 — The minimal receiver is just a detector.**



**Fig 12.4 — Detector with RF and AF amplifiers to improve sensitivity, selectivity and power output.**

was the amplifier. The first amplifiers only provided bandwidth capable of amplifying audio frequency signals and thus were inserted following the detector. In that position, they offered no improvement in sensitivity, which was still limited by the diode; however, they appeared to, because signals were now louder. It was now possible to have more than one person listen to a received signal through the use of a loudspeaker.

As shown in **Fig 12.4**, it wasn't long before vacuum tube and circuit design technology improved to the point that amplifiers could be used at RF as well as AF. The addition of RF amplifier(s) provided two performance improvements. First, the level of signals into the detector could be increased, providing an improvement in sensitivity. Second, if each amplifier were coupled using tuned resonant circuits, as was the practice, the selectivity

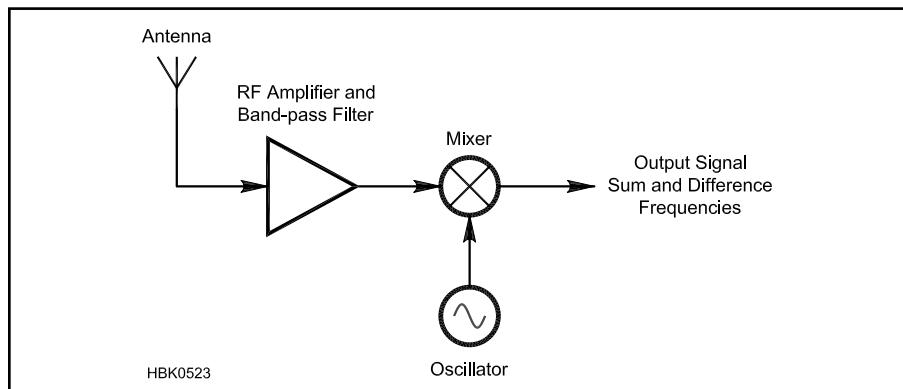
could be significantly improved.

A receiver with one or more tuned RF amplifier stages was called a *tuned radio frequency*, or TRF receiver. Some had as many as three or four stages, initially with separate controls, and later with up to a five-gang variable capacitor for station selection. Needless to say, a certain amount of skill was needed to adjust each stage so it would properly *track* as it was tuned over the frequency range. There were other challenges with the TRF. With the high gain of the multiple stages all tuned to the same frequency, it was not trivial to avoid oscillation. Another concern was the selectivity provided over the tuning range. If the selectivity of the five tuned circuits could provide a bandwidth of 2% of the tuned frequency, that would result in 30 kHz at the top of the broadcast band (around 1500 kHz) but only 10 kHz at the bottom (500 kHz). Top-end receivers of the day used various methods to automatically maintain similar bandwidth across the range.

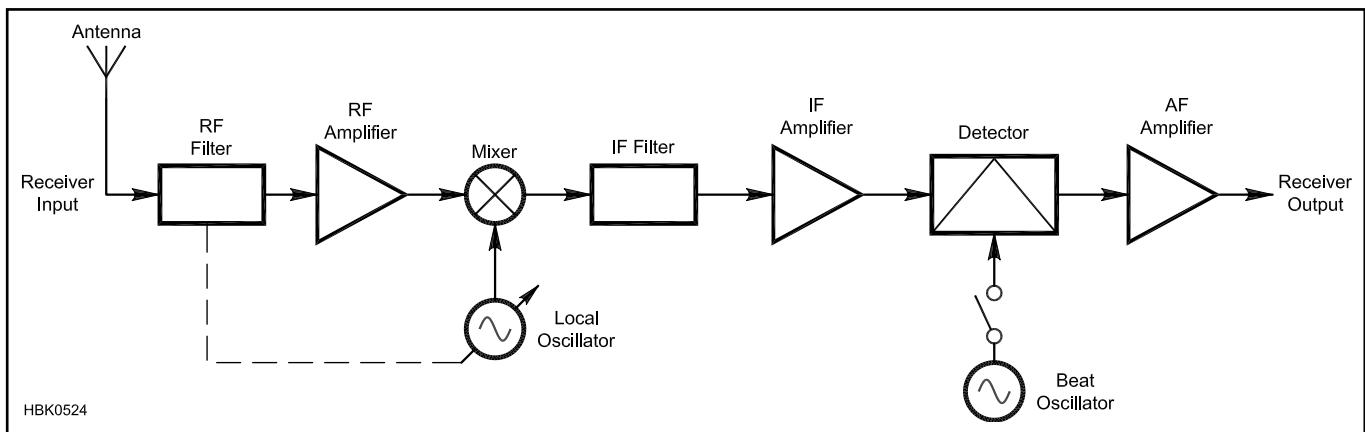
## 12.2 Basics of Heterodyne Receivers

A receiver design that avoids many of the issues inherent in a TRF receiver is called a *heterodyne*, or often *superheterodyne* (*superhet* for short) receiver. The superhet combines the input signal with a locally generated signal in a nonlinear device called a *mixer* to result in the sum and difference frequencies as shown in **Fig 12.5**. 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 local oscillator, the operator shifts the input fre-



**Fig 12.5 — Basic architecture of a heterodyne direct conversion receiver.**



**Fig 12.6 — Elements of a traditional superheterodyne radio receiver**

quency 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 — much easier to process than the variable arrangement of a TRF set.

A block diagram of a typical superhet is shown in **Fig 12.6**. 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, the input filter is often tracked along with the local oscillator. While this is similar to one of the issues about TRF receivers, note that generally there are fewer tuned circuits involved, making it easier to accomplish. The IF filter is traditionally used to establish operating selectivity — that required by the information bandwidth.

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.

In many instances, it is not possible to achieve all the receiver design goals with a single-conversion receiver and multiple conversion steps are taken. 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.

Modern receivers using digital signal processing for operating bandwidth and information detection often have an additional conversion step to a final IF at a frequency in the tens of kHz. This is established by the maximum sampling rate of a finite frequency response analog-to-digital converter (ADC). Advances in the art have resulted in ADC and processor speeds sufficient that the ADC has been moving closer and closer to the RF frequency resulting in fewer, and in some cases, no required conversions.

### 12.2.1 The Direct Conversion Receiver

As noted in the introduction, the heterodyne process can occur at a number of different points in the receiver. The simplest form of heterodyne receiver is called a *direct*

## 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:

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

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

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

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:

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

Using the trigonometric identity (see Reference 1):

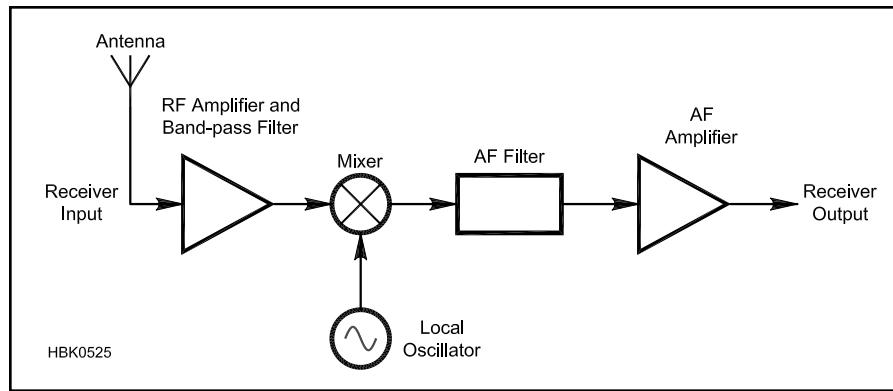
$$\sin \alpha \sin \beta = \frac{1}{2} \{ \cos(\alpha - \beta) - \cos(\alpha + \beta) \}$$

the product term becomes:

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

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.

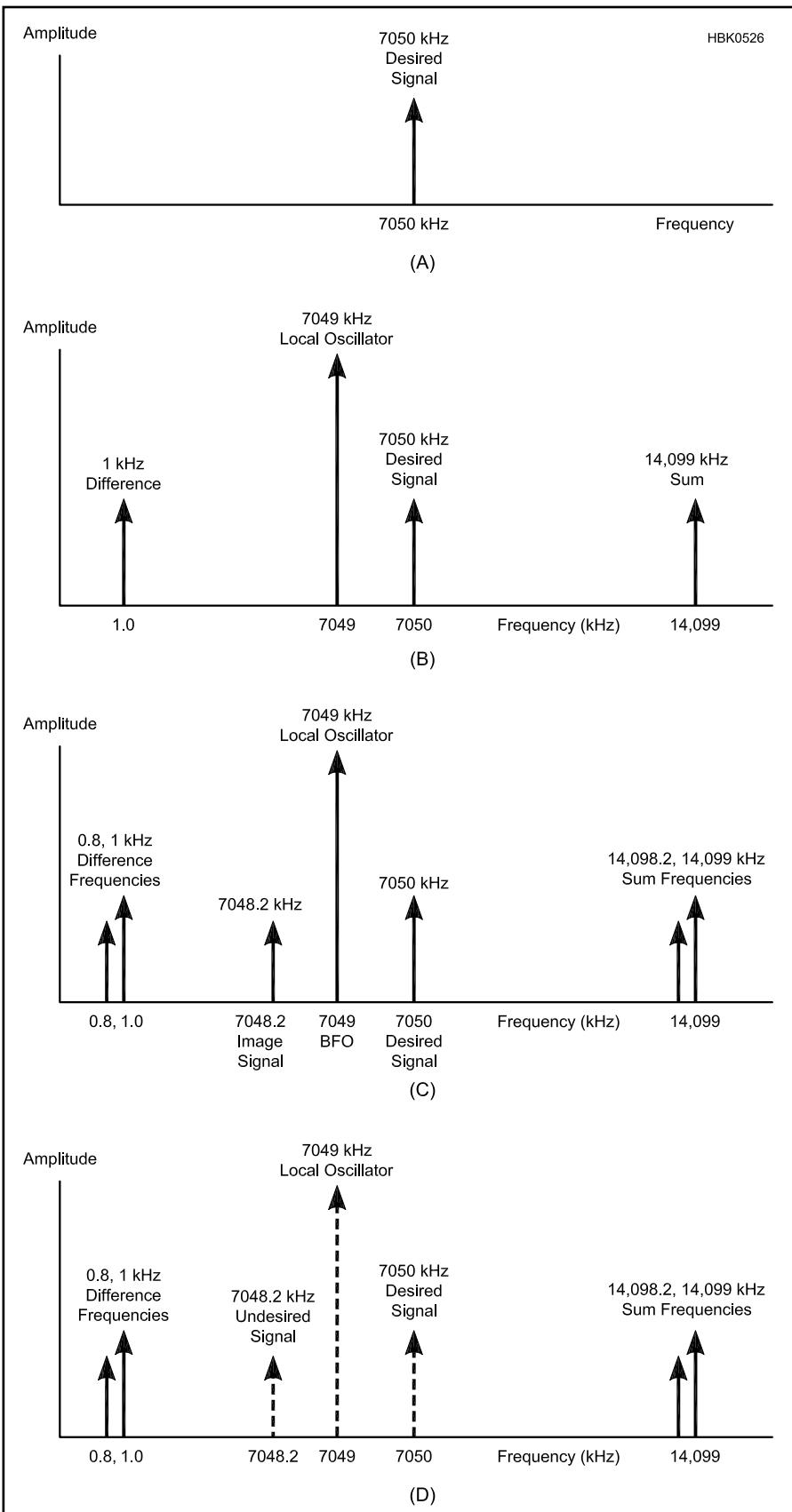


**Fig 12.7 — Block diagram of a direct conversion receiver.**

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 12.6. In this case, the detector is often preceded by an RF amplifier with a typical complete receiver shown in **Fig 12.7**. Such

a receiver can be very simple to construct, yet can be quite effective — especially for an ultra-compact low-power consumption oriented portable station.

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 12.8 shows the progression of the spectrum of an on-off keyed CW signal through such a receiver based on the relationships described above. In 12.8C, 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. A *balanced* mixer is designed 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 12.8D. We will see these again in transmitters in the next chapter.

Products generated by nonlinearities in the mixing process (see the sidebar on Nonlinear Mixing) 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.

### 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.<sup>2</sup>

This DC receiver design by Randy 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.<sup>3</sup> 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

**Fig 12.8 — 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).

## Digital Processing and Software Defined Radios

In a day in which even a toaster might include a processor (and perhaps its own IP address!), most receivers and transceivers have processors — so what's the big deal? I think we need to make a distinction between a software-*controlled* radio and a software-*defined* radio.

### Software-Controlled Radio

Most radios — or toasters for that matter — fall into the controlled category. In this case the configuration is pretty much fixed. No matter how much you change the software, you can't make that toaster into a radio — or anything besides a toaster. The toaster is limited or constrained by its physical configuration to make toast. The processor, along with associated sensors and control features, may be able to make perfect toast — just don't try to get it to do too much else.

In a similar manner, the processor in a software-controlled radio is used to sense control positions and cause logic elements to react in predictable ways. The processor may be used to count and display frequency, or form code characters at the speed you like. No matter what you do to the software controlling it, the radio is limited by its physical architecture to do what the designers thought you'd want on the day they built it.

### Software-Defined Radio

As we'll discuss, there may be a range of definitions — subject to some controversy — on what constitutes a software-defined radio (SDR) in the Amateur Radio world. The FCC has defined the SDR concept in terms of their commercial certification process as:

“...a radio that includes a transmitter in which the operating parameters of the transmitter, including the frequency range, modulation type or conducted output power can be altered by making a change in software without making any hardware changes.”

The FCC expects this to yield streamlined equipment authorization procedures by allowing “manufacturers to develop reconfigurable transceivers that can be multi-service, multi-standard, multi-mode and multi-band....”<sup>1</sup> In this context, they are envisioning radios that can be modified *at the factory* by using different software to meet different requirements. While they allow for field changes, their focus is different from ours.

### SDR in the Amateur World

In the amateur environment, we are particularly interested in radios that can be changed through software by

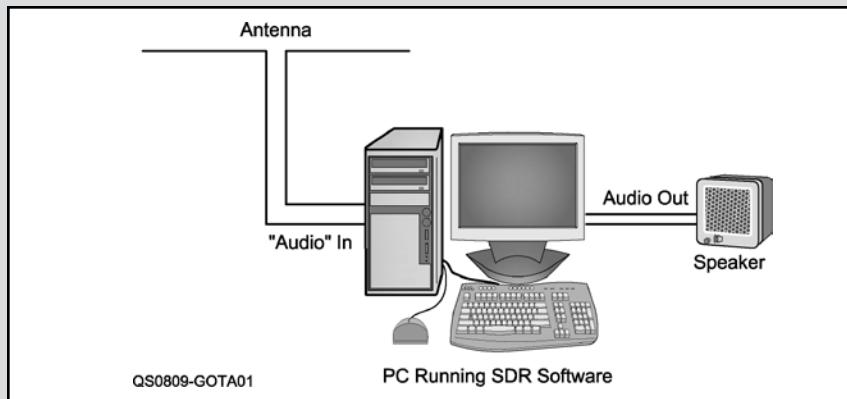


Fig 12.A1 — Conceptual block diagram of an ideal software defined radio (SDR) receiver.



Fig 12.A2 — The FlexRadio Flex-5000A SDR transceiver. A blank front panel SDR.

the end user or operator to meet their needs or to take advantages of newly developed capabilities.

The *ideal* SDR would thus have a minimum of physical constraints. On the receive side, the antenna would be connected to an analog-to-digital converter that would sample the entire radio spectrum. The digitized signal would enter a processor that could be programmed to analyze and decode any form of modulation or encoding and present the result as sights and sounds on the output side of the processor.

### So What's Wrong With This Picture?

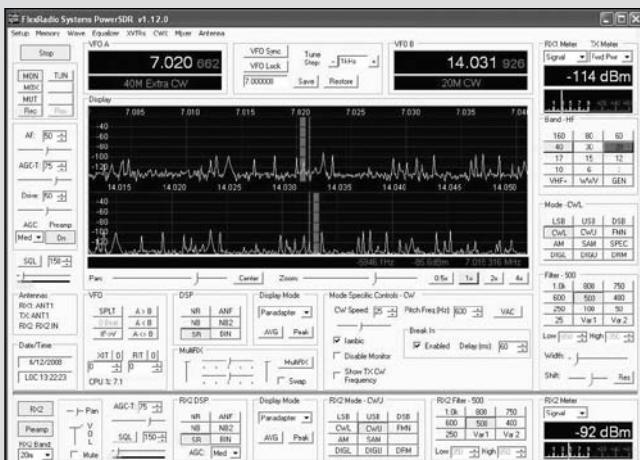
Not surprisingly, our utopian SDR is much easier to imagine than to construct. As a practical matter, the digital signal processor for most amateur work is a PC, which has some constraints that don't allow us to do quite what we want since it was not intended for use as a radio signal processor. Still, for a few hundred dollars, it is

possible to purchase a PC that gets us fairly close.

The key to amateur SDR operation with a PC is the sound card. This card, or sometimes an external interface device, can accept an analog signal and convert it to digital data samples for processing. Advanced SDRs such as the FlexRadio FLEX-5000 have the analog-to-digital conversion function built in, usually providing higher performance than a sound card ADC. The software will determine the type of processing and the nature of the signals we can deal with. It can also take the results of processing and convert them into an analog signal. This sounds like just what we are looking for to make an SDR — and it is. Such an SDR in receive mode would consist of the blocks in Fig 12.A1. We do have a few significant limitations, however.

First, the analog-to-digital converter has a specific rate at which it samples the analog signal.<sup>2</sup> The signal frequency response that it can deal with, without going to extraordinary means, is limited to no higher than half the sampling rate.<sup>3</sup> For most sound cards, the sampling rate is 192 kHz or less, limiting the received analog signal to a frequency of 96 kHz (and sometimes only 48 kHz). Some kinds of dual-channel processing allow a response as high as the sampling rate.

Second, the available signal levels are not quite what we would want. And of course, we don't have amateur bands in the dc-to-96 kHz region of the spectrum. Thus we are faced with the need to insert some external processing functions outside the PC. These will be used, at a minimum, to translate the frequency range we wish to use to one that the sound card can deal with on the receive side, and make sure the signal levels are compatible.



**Fig 12.A3 — One version of the main operating screen of PowerSDR operating software.**

SDRs usually require a baseband signal with two versions of the input signal separated by 90° of phase. We have previously discussed how such signals can be used to process a signal to remove one sideband. By having the two versions of the signal 90° apart, they are in *phase quadrature* and referred to as the I (in-phase) and Q (in-quadrature) channels. Signal processing algorithms then combine the signals in various ways to demodulate it and recover the information. (If you'd like to hear what the I and Q outputs sound like for yourself, you can build "The Binaural Receiver" by Rick Campbell, KK7B, from the instructions on the CD-ROM included with this book.)

These topics are covered in detail in the **DSP and Software Radio Design** chapter.

#### So How Does it All Come Together?

The SDR designer, as with all designers, is faced with a trade-off. The equipment external to the PC required to make it do what we want may also limit the choices we can make by software change in the PC. The more hardware features we build in, the fewer choices we may have. In addition to PC software, there is often *firmware*, hard-wired instruction in the box outside the PC. This has resulted in two general approaches in SDR.

#### The "Blank Front Panel" Architecture

Radios marketed as SDRs tend to be of this type. The classic is the FlexRadio Flex-5000A, reviewed in *QST* for July 2008.<sup>4</sup> As shown in **Fig 12.A2**, the front panel has no controls. All control functions are accessible via the soft buttons on the *PowerSDR* software's computer screen (**Fig 12.A3**), or

via computer-connected pointing and knob devices.

There are a number of other radios using a similar configuration, and useable with the same open source *PowerSDR* software freely available from FlexRadio under the GNU Public License. These include the modular High Performance SDR system available from Tucson Amateur Packet Radio (TAPR), and some other very low cost, but no longer available SDRs such as the Firefly and SoftRock 40 (although derivatives of the SoftRock 40 are still available).<sup>5</sup> Some firmware for these radios is generally so-called "open source," meaning that if you are a programmer, you can view and modify the source code and thus not only upgrade, but also make the radio do what you want.

#### The "Looks Like a Radio" Approach

Many current radios are actually built as SDRs. Some, transceivers such as the Elecraft K3 (see **Fig 12.A4**), ICOM IC-7800, Ten-Tec Orion and Yaesu FT-2000, for example, are provided with a mechanism to allow an easy end-user upgrade to new firmware revisions. These radios look like most any other pre-SDR radio in that they have front panels with knobs and dials. Unless you looked at all the revisions to the operating instructions you wouldn't know that they were field-reconfigurable.

Another distinction between the groups is that most of the firmware for the radios in this group is proprietary with revisions available only from the manufacturer, at least as of this writing. That isn't to say that a solid programmer couldn't and perhaps hasn't developed custom software for one of these, but if it's happened, it hasn't happened often.



**Fig 12.A4 — The Elecraft K3. It looks like a traditional amateur transceiver, but is actually an SDR.**

While all radios in this group are primarily designed to operate without an external computer, they all can be computer-controlled using aftermarket software, available from multiple developers. While this software can make them feel a bit like the radios in the other group, the operating parameter ranges are all set by the radio's internal operating firmware.

#### So What's it All Mean?

The blank front panel type generally has the most flexibility in operation, since they are not constrained by the physical buttons and knobs on the front panel. The more traditional looking SDR versions may take advantage of the hardware constraints that limit some operating choices to gain improved performance, but a look at the specs will indicate that it isn't always the case. Some blank panel SDRs offer top shelf performance. — *Joel R. Hallas, W1ZR*

#### Notes

<sup>1</sup>FCC Report and Order 01-264, released Sep 14, 2001.

<sup>2</sup>J. Taylor, K1RFD, "Product Review — Computer Sound Cards for Amateur Radio," *QST*, May 2007, pp 63-70. Available at [www.arrl.org/product-review](http://www.arrl.org/product-review).

<sup>3</sup>H. Nyquist, "Certain Topics in Telegraph Transmission Theory," *Transactions of the AIEE*, Vol 47, pp 617-644, Apr 1928. Reprinted as classic paper in *Proceedings of the IEEE*, Vol 90, No. 2, Feb 2002.

<sup>4</sup>R. Lindquist, N1RL, "Product Review — The FlexRadio Flex-5000A," *QST*, Jul 2008, pp 39-45. Available on the ARRLWeb at [www.arrl.org/product-review](http://www.arrl.org/product-review).

<sup>5</sup>Contact Softrock developer Tony Parks, KB9YIG, at [raparks@ctcisp.com](mailto:raparks@ctcisp.com) to see if a new version is available.

at Fig 12.8B, 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.

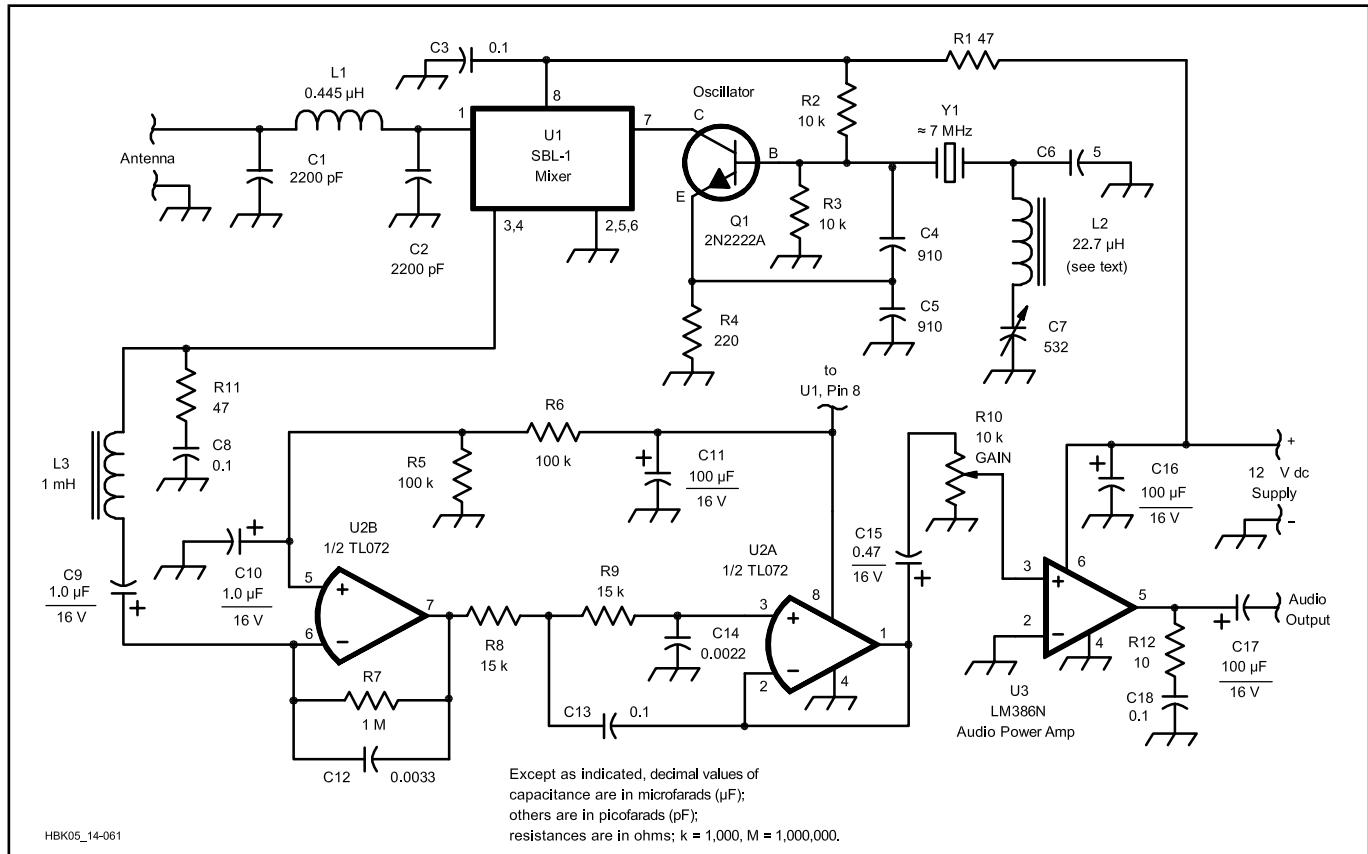
Building a stable oscillator is often the most challenging part of a simple receiver. This one uses a tunable crystal-controlled oscillator that is both stable and easy to reproduce. All

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 12.9, mixer U1 puts out sums and differ-

ences 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.



**Fig 12.9 —**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 (VCO), 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  $\frac{1}{4}$  W, 5% tolerance types; the circuit's polarized capacitors are 16 V electrolytics, except C10, which can be rated as low as 10 V. The 0.1  $\mu$ F 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),  $\frac{1}{4}$  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  $\mu$ F.  
U2 — TL072CN or TL082CN dual JFET op amp.

L1 — Four turns of AWG #18 wire on  $\frac{1}{4}$  inch PVC pipe form. Actual pipe OD is 0.85 inch. The coil's length is about 0.65 inch; adjust turns spacing for maximum signal strength. Tack the turns in place with cyanoacrylic adhesive, coil dope or 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  $\mu$ H; consists of one or more encapsulated RF chokes in series (two 10- $\mu$ H chokes [Mouser #43HH105 suitable] and one 2.7- $\mu$ H

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 k $\Omega$  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.

## 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 rock-bound receiver's tuning range with crystal pulling made *rock bending* seem just as appropriate!

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

Inductor L2 and the crystal, Y1, have more effect on the oscillator than any other components. Breaking up L2 into two or three series-connected components often works better than using one RF choke. (The author used three molded RF chokes in series — two 10  $\mu\text{H}$  chokes and one 2.7  $\mu\text{H}$  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  $\mu\text{H}$  to eke out a mere 15 kHz of tuning range.

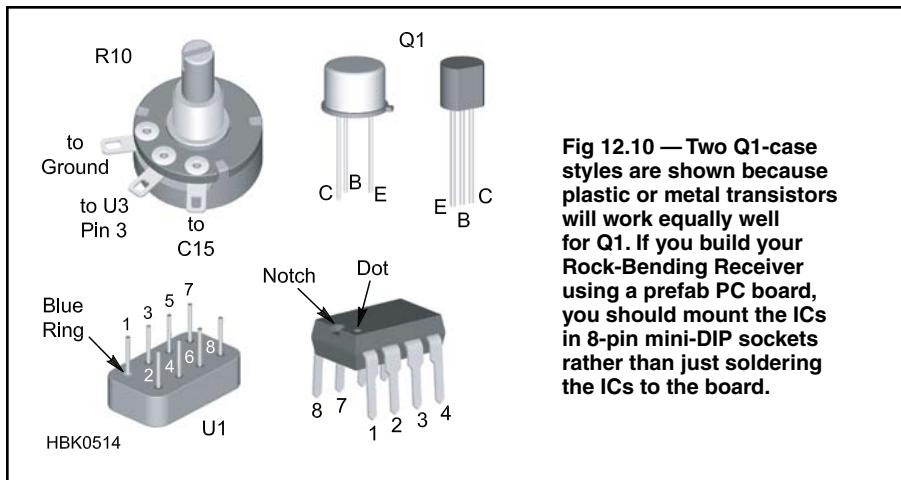
## Input Filter and Mixer

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.

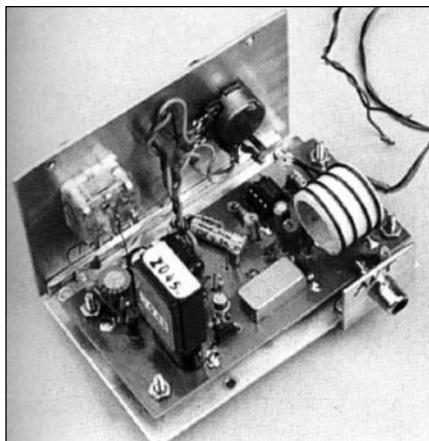
## Audio Amplifier and Filter

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 12.9.

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



**Fig 12.10 —** 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 12.11 —** Ground-plane construction, PC-board construction — either approach can produce the same good Rock Bending Receiver performance.

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 12.10** 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 12.11** shows a photo of the receiver built 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  $\frac{1}{4}$ -inch diameter drill bit works well for this. Attach a control knob to the bit's shank and you can safely use the bit as a manual countersinking tool. If you countersink your board in a drill press, set it to about 300 rpm or less, and use very light pressure on the feed handle.

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

## CHECKOUT

Before connecting the receiver to a power source, thoroughly inspect your work to spot obvious problems like solder bridges, incor-

rectly 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  $1\ \Omega$  of resistance. The resistance between the supply lead and pin 8 of U1 should be about  $47\ \Omega$  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  $0.3\ \mu\text{V}$ . The author could easily copy  $1\ \mu\text{V}$  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

The *superheterodyne*, or *superhet* for short, uses the principles of the direct conversion heterodyne receiver above — at least twice. 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 12.6.

In a typical configuration, the local oscillator (LO) and RF amplifier stages are adjusted so that as the LO is changed in frequency, the RF amplifier is also tuned to the appropriate frequency to receive the desired station. An example may help. Let's pick a common IF frequency used in an AM broadcast radio, 455 kHz. Now if we want to listen to a 600 kHz broadcast station, the RF stage should be set to amplify the 600 kHz signal and the LO should be set to  $600 + 455$  kHz or

1055 kHz.<sup>4</sup> The 600 kHz signal, along with any audio information it contains, is translated to the IF frequency and is amplified. It is then detected, just as if it were a TRF DC receiver at 455 kHz.

Note that to detect standard AM signals, the second oscillator, usually called a *beat frequency oscillator* or BFO, is turned off since the AM station provides its own carrier signal over the air. Receivers designed only for standard AM reception, the typical "table or kitchen radio," 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. With the appropriate ganged and tracked tuning capacitors, the RF stage is tuned to 1560 kHz, and the LO is set to 1560 + 455 or 2010 kHz and now that station is

translated to our 455 kHz IF. Note that the bulk of our amplification can take place at the 455 kHz IF frequency, so not as many stages must be tuned each time we change to a new frequency. Note also that with the superheterodyne configuration the 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. The superhet design has thus eliminated the major limitations of the TRF at the cost of two additional building blocks.

It should be no surprise then that the superhet, in various flavors that we will discuss, has become the primary receiver architecture in use today. The concept was introduced by Major Edwin Armstrong,<sup>5</sup> a US Army artillery officer, just as WW I was coming to a close. The superhet quickly gained popularity and, following the typical patent battles of the times, became the standard of a generation of vacuum-tube broadcast receivers. These

were found in virtually every US home from the late 1920s through the 1960s, when they were slowly replaced by transistor sets, but still of superhet design.

While the superheterodyne receiver is still the standard in many applications — Amateur Radio, broadcast radio, televisions, microwave, and radar, to name but a few — it may be surprising to amateurs to learn that in terms of sheer numbers, the direct-conversion receiver is far and away the most widely used technology. More than one billion wireless telephone handsets — “cell phones” — are manufactured every year and more than 90% of them employ direct-conversion receivers!

### 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 12.8, 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 by 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 (telephone company “toll quality”) human speech of 300 to 3300 Hz, we would have a pair of

bands of information carrying waveforms extending from 596.7 to 603.3 kHz, as shown in Fig 12.12. 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 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. 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 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 car-

rier 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.

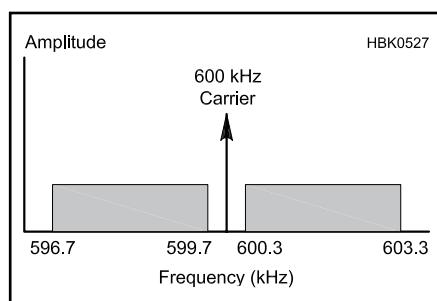
#### SINGLE-SIDEBAND SUPPRESSED-CARRIER MODULATION (SSB)

The standard commercial AM format is very convenient for receiver design, since the carrier needed to demodulate the received sidebands is sent along with the sidebands. Applying the total received signal to a detector or multiplier, allows the audio to be recovered without having to worry about any of the finer points we will discuss later. This is very cost effective in a broadcast environment in which there are many inexpensive receivers and only a relatively few expensive transmitters.

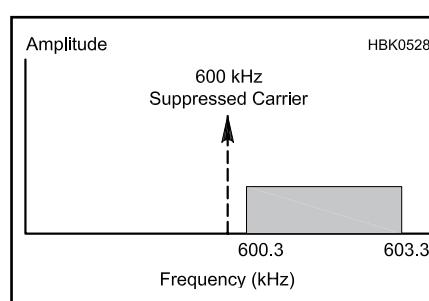
In looking at Fig 12.12, you might have noticed that both sidebands carry the same information, and are thus redundant. In addition, the carrier itself conveys no information. It is thus possible to transmit a *single sideband and no carrier*, as shown in Fig 12.13, relying on the BFO (beat frequency 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, even by a few Hz, a baritone can come out sounding like a soprano and vice versa!

This makes 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 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 area around the AM carrier need not be received. Thus the toll quality spectrum (a term carried over from long-distance telephone systems) 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



**Fig 12.12 — Spectrum of sidebands of an AM voice signal sent on a 600 kHz carrier.**



**Fig 12.13 — Spectrum of single sideband AM voice signal sent adjacent to a suppressed 600 kHz carrier.**

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 telegraphy case, it will be turned on and off quite rapidly. The rise and fall of the carrier results in sidebands extending out from the carrier for some distance, and they must be received in order to reconstruct the signal in the receiver.

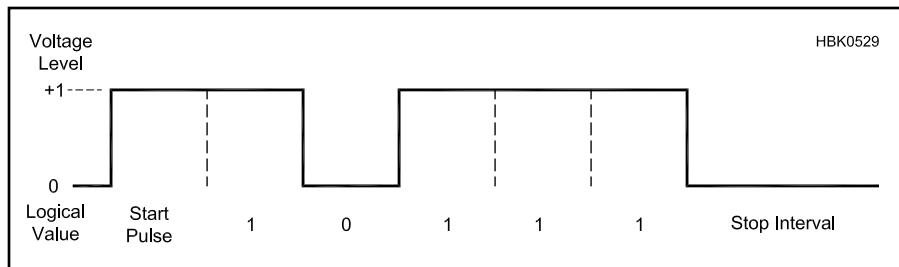
A rule of thumb is to consider the rise and fall time as about 10% of the pulse width and the bandwidth as the reciprocal of the quickest of rise or fall time. This results in a bandwidth requirement of about 50 to 200 Hz for the usual radiotelegraph transmission rates. Another way to visualize this 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.

## DATA COMMUNICATION

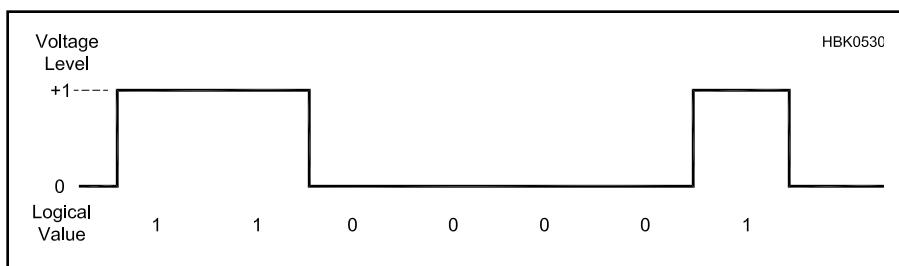
(This is a short overview of receivers and data communications. See the **Modulation**, and **Digital Modes** chapters and the **Digital Communications** supplement on the *Handbook* CD for more in-depth treatment.)

The Baudot code (used for teletype communications) and ASCII code — two popular digital communications codes used by amateurs — are constructed with sequences of ON-OFF elements or bits as shown in **Figs 12.14** and **12.15**. 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*. (These states are named after the early Morse tape readers that placed a pen *mark* on a paper tape when the key was down and made a *space* when the key was up.) 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.<sup>6</sup> 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.



**Fig 12.14 — Voltage pulses of the letter “A” in Baudot code, with start and stop pulses.**



**Fig 12.15 — Voltage pulses corresponding to the lower case letter “a” in ASCII code.**

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 the facilities of 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 bps 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.

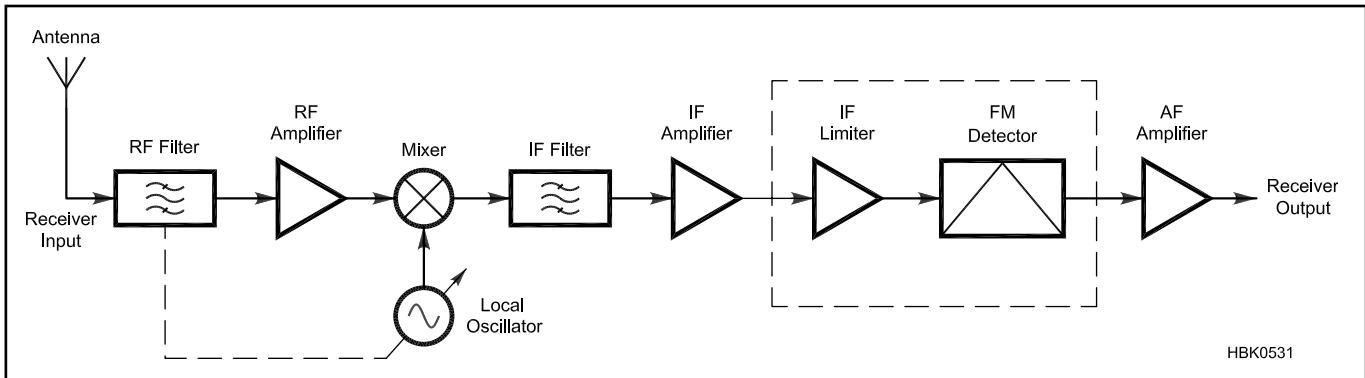
At HF, the signal is subjected to phase and amplitude distortion as it travels. Noise is also substantially higher on the HF bands. Under these conditions, modulation and de-modulation techniques designed for “wire-line” connections become unusable at bit rates of more than a few hundred bps. As

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. (Spread-spectrum techniques are also being adopted on the 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. 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  $2 \times (D + M)$ , where D is the deviation and M is the maximum modulating frequency.



**Fig 12.16 — Block diagram of an FM superhet. Changes are in dashed box.**

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 *wide-band 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 wide-band 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 12.16**. The fundamental change is that the detector must recover information from the frequency variations of the input signal. The most common type is called a *discriminator*. The discriminator does not require the 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.

#### 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.

For many years, the operating bandwidth of a receiver was established by multiple tuned interstage transformers in the IF portion of the receiver. All things being equal, the lower the IF frequency the narrower the selectivity; the higher the IF frequency the wider the passband. The actual bandwidth was determined by the details of the design. The number and quality of the tuned circuits, and the amount of coupling between them had a major impact, not only on the nominal bandwidth, but also the slope of the response curve. **Table 12.2** shows the selectivity of some typical early IF arrangements based on tuned circuits. The typical single IF stage had two tuned transformers, one

on each side. Adding each additional stage typically added two tuned circuits, with advanced models sometimes having additional tuned circuits between stages.

The usual AM broadcast receiver, and low end “communications” receiver of the vacuum tube day, has a single IF stage at 455 kHz, and as can be seen, this provided reasonable selectivity for use in areas that did not have signals on adjacent channels.

It was relatively easy for a designer to increase the bandwidth for wider bandwidth modes, either by stagger tuning the transformers to slightly different frequencies across the band, or even by using resistive loading to reduce the circuit Q. Decreasing the bandwidth for narrower modes was a bit trickier and early communications receivers often provided a single crystal filter that could be used to provide a narrow band pass for CW operation.

### 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*.

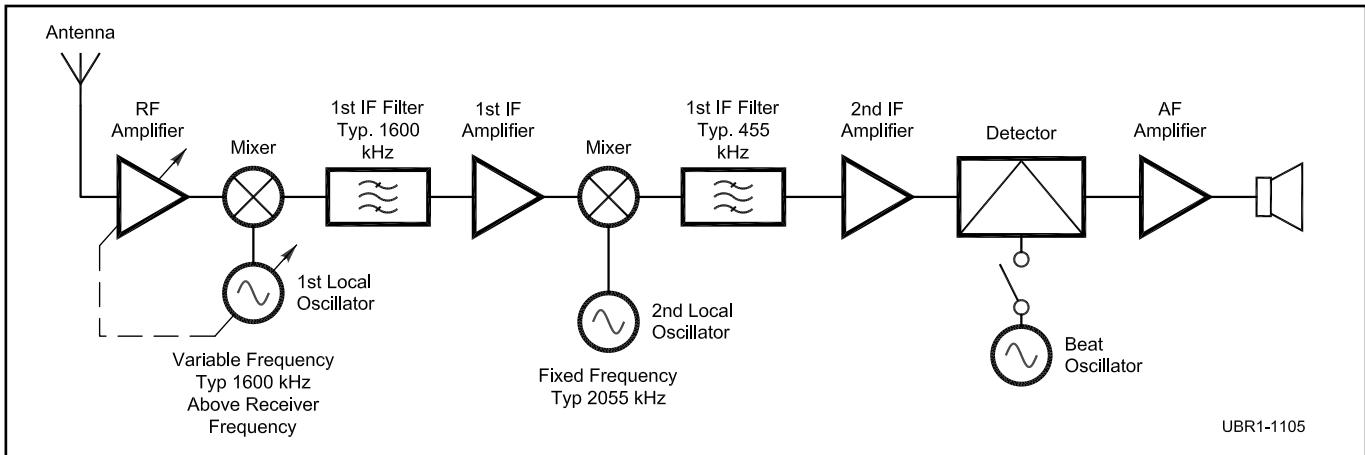
Images will be separated from the desired frequency by twice the IF and must be filtered

**Table 12.1**  
Typical Communications Bandwidths for Various Operating Modes

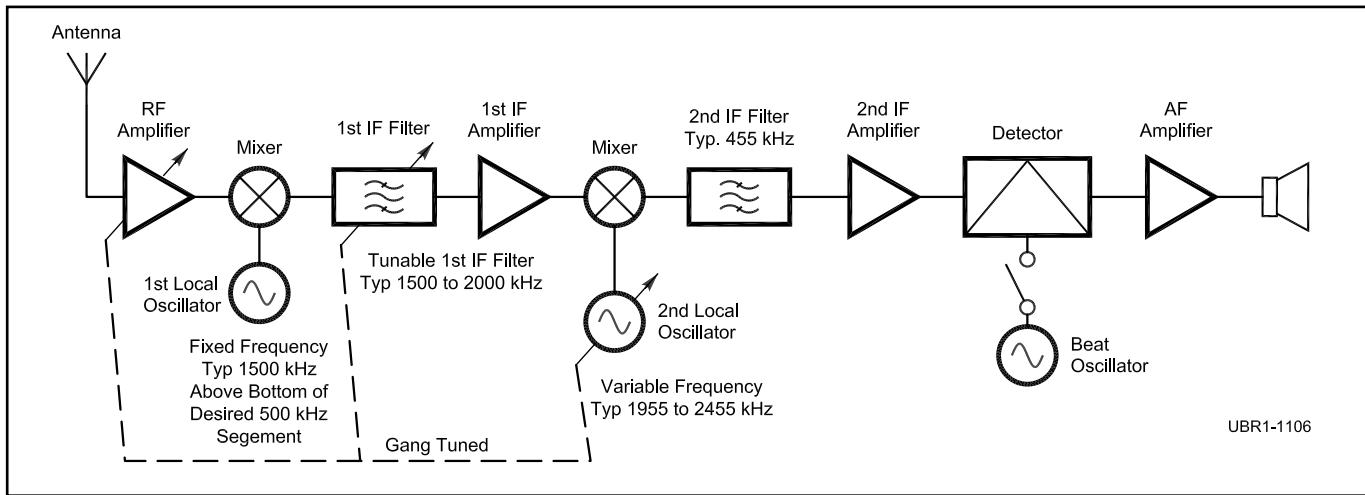
Mode	Bandwidth (kHz)
FM Voice	15
AM Broadcast	10
AM Voice	4-6.6
SSB Voice	1.8-3
RTTY (850 Hz shift)	0.3-1.0
CW	0.1-0.5

**Table 12.2**  
Selectivity of Early Superhet IF Amplifier Arrangements

IF Amplifier	-6 dB	-20 dB	-40 dB
One stage 50 kHz (iron core)	0.8	1.4	2.8
One stage 455 kHz (air core)	8.7	17.8	32.3
One stage 455 kHz (iron core)	4.3	10.3	20.4
Two stages 455 kHz (iron core)	2.9	6.4	10.8
Two stages 1600 kHz (iron core)	11.0	16.6	27.4



**Fig 12.17 — Double-conversion superhet receiver. Early type.**



**Fig 12.18 — Double-conversion superhet receiver. Collins system.**

out ahead of the associated mixer by the ratio of desired 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, a much more difficult filtering problem

The popular 455 kHz IF frequency generally started to have serious image problems at receiving frequencies around 10 MHz and higher. Early top-of-the-line receivers extended this range somewhat by having multiple gang-tuned RF stages ahead of the mixer to reduce the image, but the nonlinear responses of the additional gain stages ahead of the mixer often resulted in additional unwanted mixer products that could be as bad as the image.

While an image that fell on an occupied channel was obviously troubling—it's rarely desirable to receive two signals at the same

time — it is a problem 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

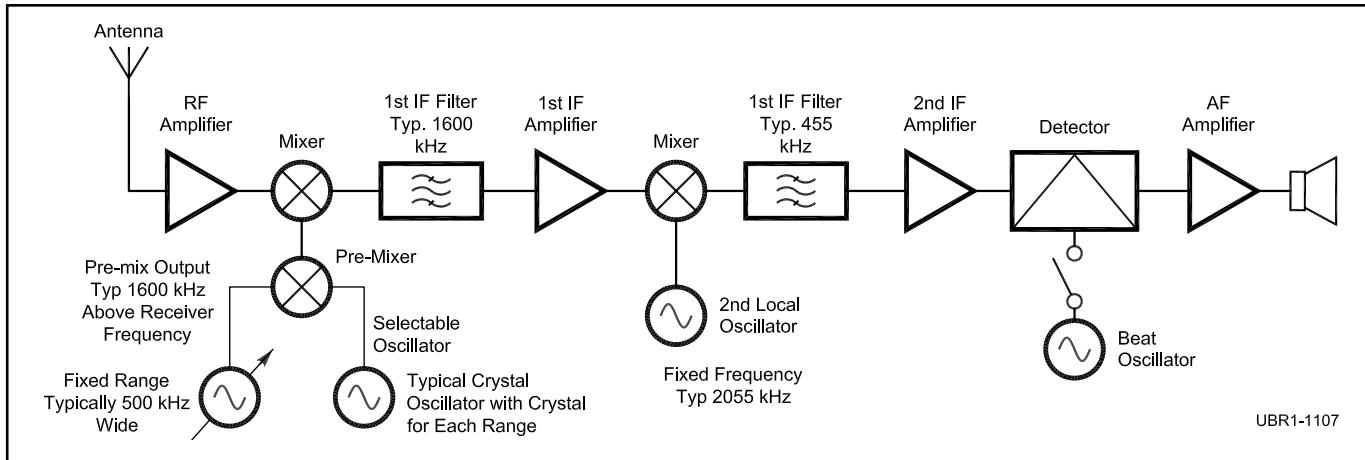
An obvious solution to the RF image response is to raise the IF frequency high enough so that signals at twice the IF frequency from the desired signal are sufficiently attenuated by the tuned circuit(s) ahead of the mixer. This can easily be done, with most selections in the area of 5 to 10% of the 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 circuits 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.

#### Double Conversion

In the 1950s, a new technique called *double conversion* was defined to solve the image problem. Rather than having to decide between a high IF frequency for good image rejection, or a low IF frequency for narrow channel selectivity, some bright soul decided to do *both!* As shown in **Fig 12.17**, a conversion of the desired signal to a relatively high IF is followed by a second conversion to a lower IF to set the selectivity. This arrangement solved the image problem nicely,



**Fig 12.19 — Double-conversion superhet receiver. Pre-mixed local oscillator configuration.**

while the rest of the receiver could be pretty much kept the same as before. A number of manufacturers offered revised receivers in the fifties that added an additional conversion stage, often just above 10 or 15 MHz, regions that had poor image rejection in the original design. Note that major changes were not required in the receiver, and some looked just like their single conversion predecessors from the outside. The extra stage was shoehorned in, with just a retuning of the first local oscillator required.

Some manufacturers decided that if two conversions were good, three must be even better. These designs often started in the same place, but converted the 455 kHz second IF down to a third IF, often in the 50 to 100 kHz range, for even sharper selectivity. This approach lasted until crystal lattice filters became available that outperformed the low frequency LC circuits and avoided the need for the additional conversion stage.

In the 1940s, a visionary radio pioneer, Arthur Collins, W0CXX, founder of Collins Radio in the 1930s, came up with another approach to double conversion. One problem with earlier MF and HF receivers was that they used a standard LC (first) oscillator using a variable capacitor of the type used in broadcast receivers. As noted earlier, these capacitors typically had a 9:1 capacitance range resulting in a 3:1 frequency range. The typical tuning arrangement for multi-band receivers was .5 to 1.5, 1.5 to 4.5, 4.5 to 13.5, and 13.5 to 30 MHz. This covered from the broadcast band to the top of the HF range in four bands.<sup>7</sup> With this arrangement, the tuning rate and dial calibration marks got less and less precise as higher bands were selected making tuning difficult.

The Collins system (see Fig 12.18) switched the variable oscillator to the second position and used crystal-controlled oscillators as the first oscillator. Although many more bands were required (30 for the famous

51J series that covered 0.5 to 30.5 MHz) each tuned with exactly the same tuning rate. Collins went a step further and designed an inductance-tuned oscillator (usually called a permeability-tuned oscillator or PTO) that was linear throughout its range. These oscillators could be tuned to the nearest 1 kHz from beginning to end, avoiding the tuning uncertainty of other receiver types. For this to work properly, synchronized or gang tuning is required between the oscillator, first IF variable filter and RF stages. This was no problem for the engineers at Collins Radio, but their equipment brought a premium price.

A third approach to double conversion was a mix of the first two (no pun intended). The *pre-mixed* arrangement, see Fig 12.19, uses a single variable oscillator range, as with the Collins, but does the mixing outside of the signal path. This avoids the need for variable tuning at the first IF, allowing a tighter *roofing filter* but trades it for the need for filtering at the output of the pre-mixer (not shown).

### High Frequency Crystal Lattice Filters

Just as double conversion receivers were settling into becoming the “way things were done,” the use of piezoelectric crystals as elements of band-pass filters became feasible. While crystal lattice filters at 455 kHz were popular following WW2 because of availability of large stocks of surplus crystals at that range, they didn’t address the image response issue, just provided better shape factors at the same IF frequency. Starting in the 1950s, crystal lattice filter technology moved higher in the MF region and into HF.

Commercial filters at CW and SSB appropriate bandwidths became available at center frequencies up to perhaps 10 MHz. Now a single-conversion receiver with an IF in the HF range could provide both high image rejection and needed channel selectivity. Fig 12.20 is an example of a single-

conversion superhet using simple filters at 1500 kHz. While the filters shown are actually buildable at low cost, multiple-section filters with much better performance can be purchased or constructed to make this a top performing receiver. Still, the circuit shown will demonstrate the concepts and can be reproduced at low cost. Other IF frequencies can be used, perhaps depending on crystal or filter availability.

### The Image-Rejecting Mixer

The third 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-canceling, or more accurately *image-rejecting* mixer requires phase-shift networks, as shown in Fig 12.21. Frequency  $F_1$  represents the input frequency while  $F_2$  is that of the local oscillator. Note that the two  $90^\circ$  phase shifts are at different frequencies. While conceptually straightforward, the phase shift network following mixer one is at a fixed center frequency corresponding to the IF, while the phase shift network at frequency two must provide the required phase shift as the local oscillator tunes across the band.

As usual, the devil is in the details. 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  $90^\circ$  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 employed toward that end. This method can be found in a number of places in some receivers.

- 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 per-

former, although the added complexity is not always warranted in many 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.

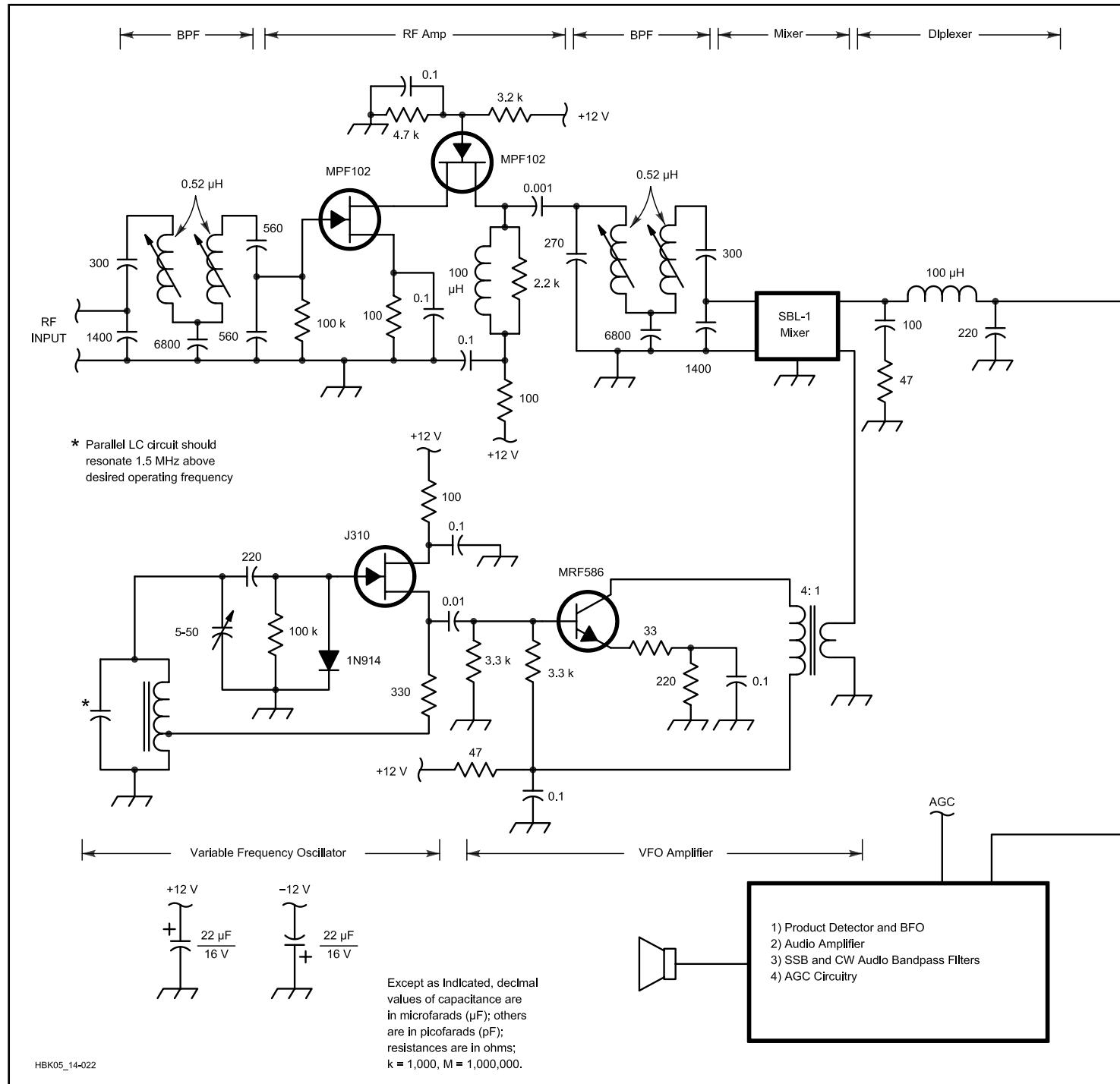
Even with a high first IF frequency, additional image rejection can be provided.

- In transmitters, the same system is called the *phasing method* of SSB generation. The same blocks run “backwards,” with one of the phase shift networks applied to the speech band, can be used to cancel one sideband. More about this technique will be presented in the **Transmitters** chapter. The following is the description of a direct conversion receiver that uses an image reject-

ing mixer as a product detector for single sideband or single signal CW reception.

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

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



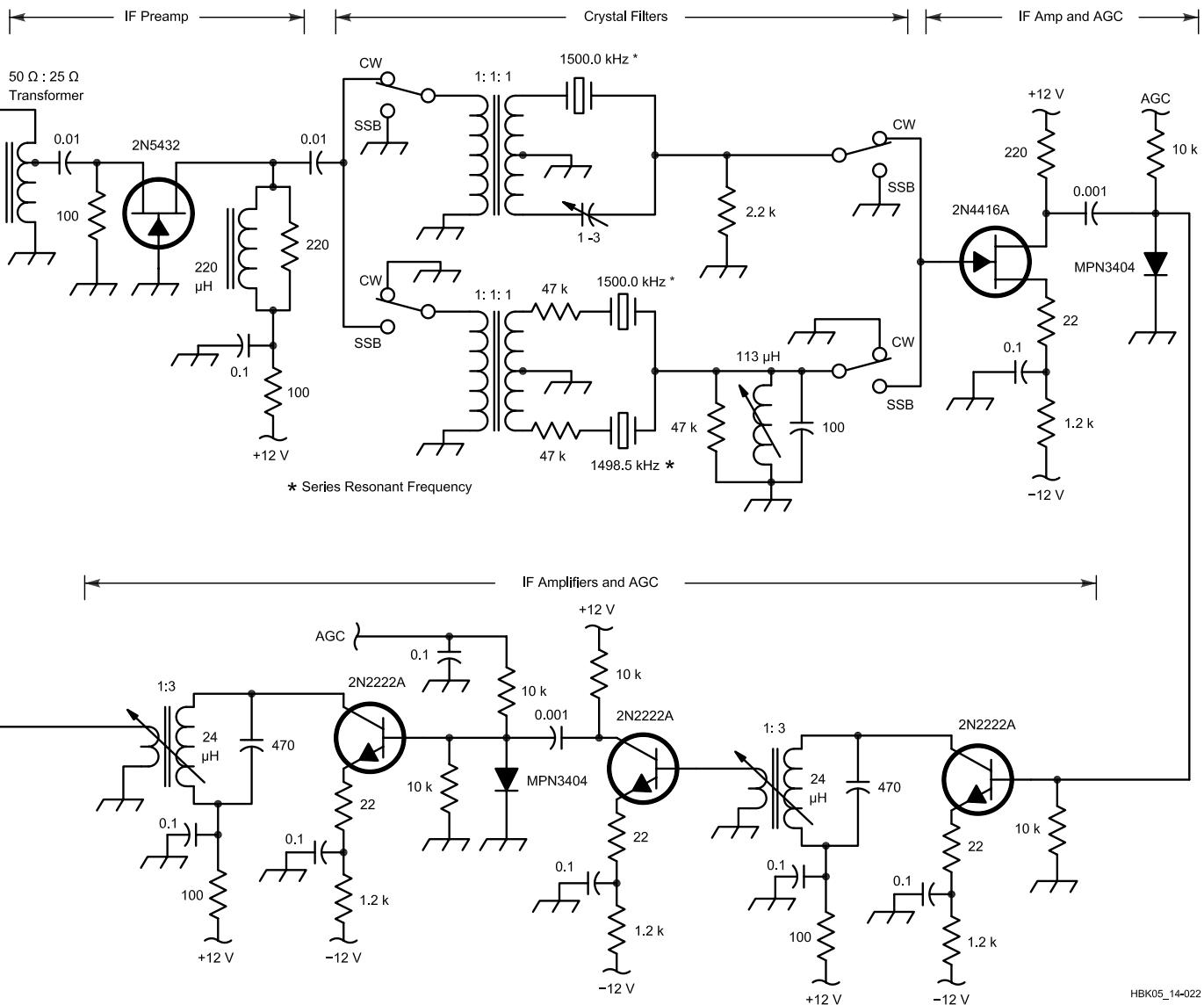
Rick Campbell, KK7B, captures some of the elegance of the classic all-analog projects of the past while achieving the performance needed for operation in 21st-century band conditions across a selected segment of a single band.

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

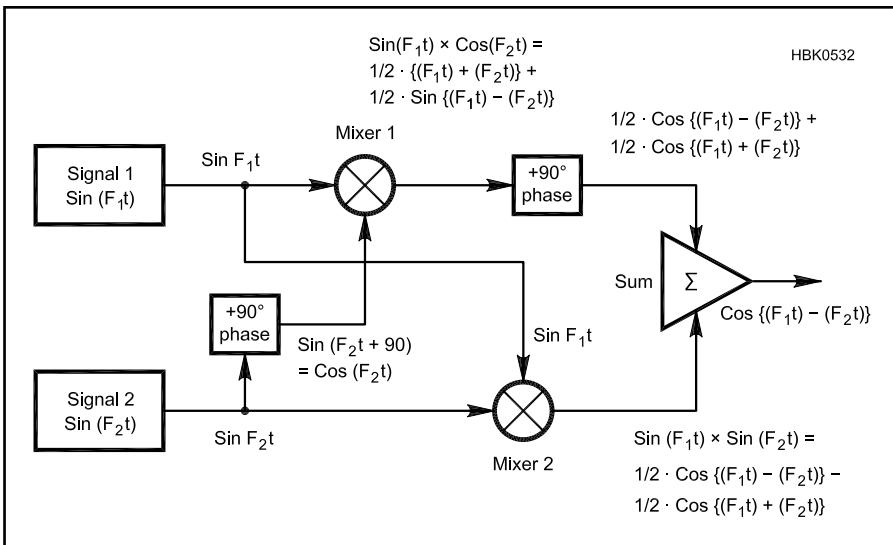
and simple receiver. The receiver doesn't need a well-calibrated dial, because the transmitter is used to spot the frequency. The transmitter doesn't need sidetone, VFO offset or other circuitry because the crystal determines the transmitted frequency and the receiver is used to monitor the transmit 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 12.22**) 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 12.23** 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

**Fig 12.20 — Schematic of representative superhet using crystal lattice filters to achieve selectivity with a high IF frequency.**



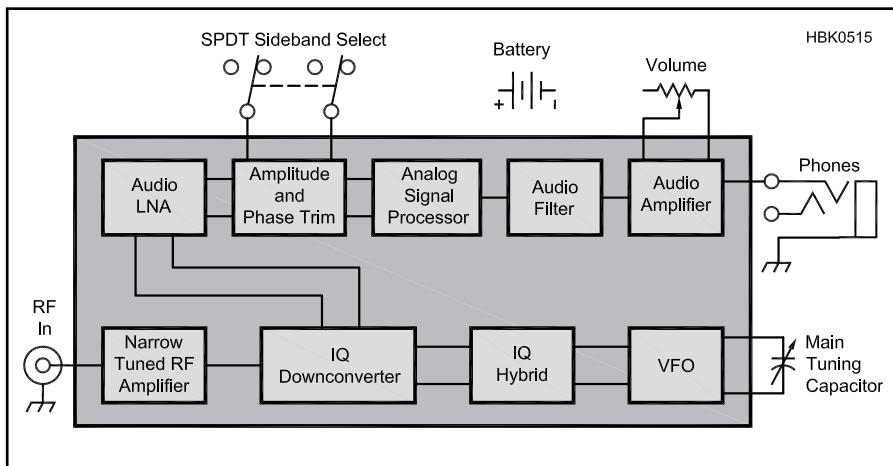
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**Fig 12.21 — Image rejecting mixer, block diagram and signal relationships.**



**Fig 12.22 — Two versions of the MicroR2. Hard to beat a big vernier dial!**



**Fig 12.23 — Block diagram of the MicroR2 SSB/CW receiver.**

by some very talented designers.

### “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 12.23 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 12.23 and the schematic in **Fig 12.24** follows. (Additional detail and rationale for the circuits is provided in *EMRFD*.)

The 50- $\Omega$  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

Decimal values of capacitance are in microfarads ( $\mu\text{F}$ ); others are in picofarads ( $\text{pF}$ ); Resistances are in ohms;  $k=1,000$ ,  $M=1,000,000$ .

JFETs are type J310

NPN Bipolar Transistors are type 2N3904

Op-amps are type NE5532

\* Jumper or Sideband Select Switch

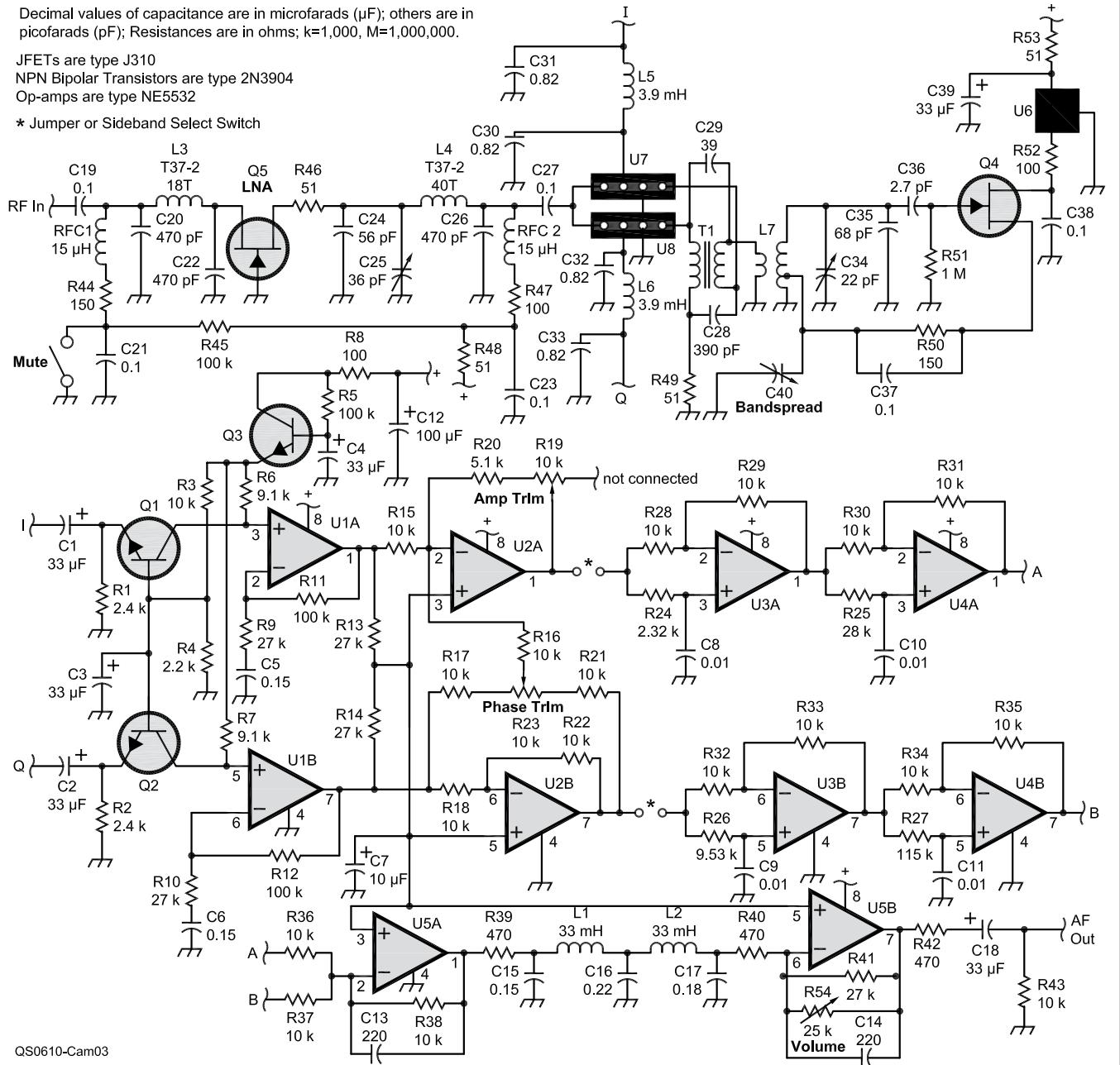


Fig 12.24 — Schematic diagram of the MicroR2 SSB/CW receiver. The parts are listed in Table 12.3.

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 receivers 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 frequency and all voltages except the LO are less than a few millivolts.<sup>8</sup>

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  $50 \Omega$  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

**Table 12.3****MicroR2 Parts List**

Parts for CW and SSB Version

C1-C4, C18, C39 — 33 $\mu$ F, 16 V electrolytic.	C34 — 2.2-22 pF, poly trimmer capacitor.
C5, C6 — 0.15 $\mu$ F, 5% polyester.	C35 — 68 pF, NP0 ceramic.
C7 — 10 $\mu$ F, 16 V electrolytic.	C36 — 2.7 pF 5% NP0 1000V ceramic.
C8-C11 — 0.01 $\mu$ F, polyester matched to 1%.	C40 — Variable capacitor, see Note D.
C12 — 100 $\mu$ F, 16 V electrolytic.	L1, L2 — 33 mH inductor, see Note A.
C13, C14 — 220 pF, NP0 ceramic.	L3 — 18t #28 T37-2, see Note E.
C15 — 0.15 $\mu$ F, 5% polyester, see Note A.	L4 — 40t #30 T37-2, see Note E.
C16 — 0.22 $\mu$ F, 5% polyester, see Note A.	L5, L6 — 3.9 mH inductor, see Note C.
C17 — 0.18 $\mu$ F, 5% polyester, see Note A.	L7 — 36t #28 T50-6, tap at 8 turns with a 2 turn link, see Note E.
C19, C21, C23, C27, C37, C38 — 0.1 $\mu$ F, 5% polyester.	Q1-Q3 — 2N3904.
C20, C22, C26 — 470 pF, NP0 ceramic.	Q4, Q5 — J310.
C24 — 56 pF, NP0 ceramic.	R1, R2 — 2.4 k $\Omega$ .
C25 — 3-36 pF, poly trimmer capacitor.	R3, R15-R18, R21, R22, R38, R43 — 10 k $\Omega$ .
C28 — 390 pF, NP0 ceramic.	R4 — 2.2 k $\Omega$ .
C29 — 39 pF, NP0 ceramic on back of board, see Note B.	R5, R11, R12, R45 — 100 k $\Omega$ .
C30-C33 — 0.82 $\mu$ F, 5% polyester, see Note C.	R6, R7 — 9.1 k $\Omega$ .
	R8, R47, R52 — 100 $\Omega$ .
	R9, R10, R13, R14, R41 — 27 k $\Omega$ .
	R19, R23 — 10 k $\Omega$ trimpot.
	R20 — 5.1 k $\Omega$ .
	R24 — 2.32 k $\Omega$ , 1% see Note A.
	R25 — 28.0 k $\Omega$ , 1% see Note A.
	R26 — 9.53 k $\Omega$ , 1% see Note A.
	R27 — 115 k $\Omega$ 1% see Note A.
	R28-R37 — 10.0 k $\Omega$ 1%.
	R39, R40, R42 — 470 $\Omega$ .
	R44, R50 — 150 $\Omega$ .
	R46, R48, R49, R53 — 51 $\Omega$ .
	R51 — 1 M $\Omega$ .
	R54 — 25 k $\Omega$ volume control.
	RFC1,RFC2 — 15 $\mu$ H 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 k $\Omega$ , 1%.  
 R25 — 41.2 k $\Omega$ , 1%.  
 R26 — 16.9 k $\Omega$ , 1%.  
 R27 — 147 k $\Omega$ , 1%.  
 C15 — 0.47  $\mu$ F, 5% polyester.  
 C16 — 0.68  $\mu$ F, 5% polyester.  
 C17 — 0.56  $\mu$ F, 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  $-j50 \Omega$  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 12.26. This modification is recommended if the MicroR2 must be used near 60 Hz power transformers.  
 C30, C32 — 0.10  $\mu$ F 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  $\pm 50$  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  $+j100 \Omega$  at about 8 MHz (not particularly critical). L4 and L7 should be about  $+j250 \Omega$  at mid band. Each winding of T1 should be  $+j50 \Omega$  at mid band.

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 construc-

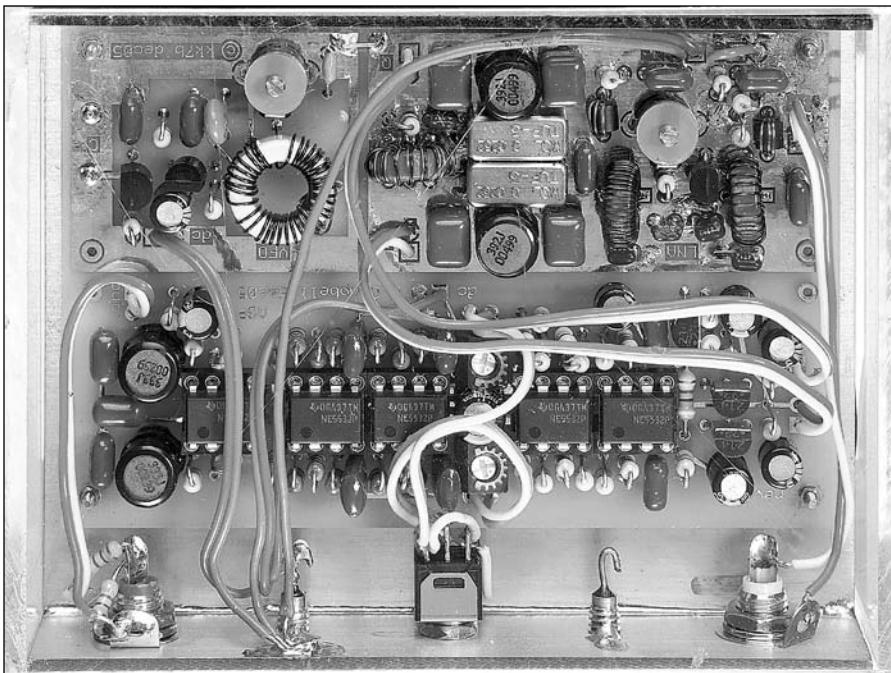
tion and alignment time by a factor of 10.

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 12.25.

### 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



**Fig 12.25 — Photo of the inside of the receiver. The PC board and parts are available from Kanga US ([www.kangaus.com](http://www.kangaus.com)).**

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 '60s. 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 elimi-

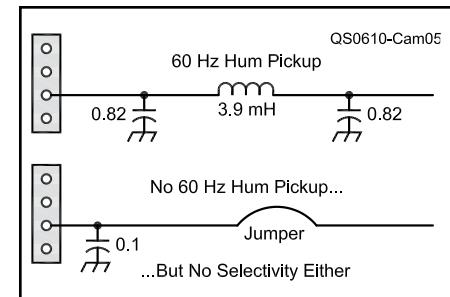
nating the film trimmer capacitor that sets the VFO tuning range. It may be replaced by an air trimmer, a piston trimmer, a small compression trimmer, or a combination of small fixed NP0 ceramic capacitors experimentally selected to obtain the correct tuning range. The holes in the PC board are large 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 ab-

**Table 12.4  
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 20 kHz spacing, -5 dBm.
Third-order intercept:	
Third-order intercept points were determined using equivalent S5 reference.	



**Fig 12.26 — Modified circuit to minimize hum.**

solutely 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 12.4**.

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 12.26**. 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.

### 12.3.3 Multiple-Conversion in the Digital Age

Each of the design choices described above are still found in current receiver architectures, but some recent advances have added a few new twists. 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**. An overview is presented here.

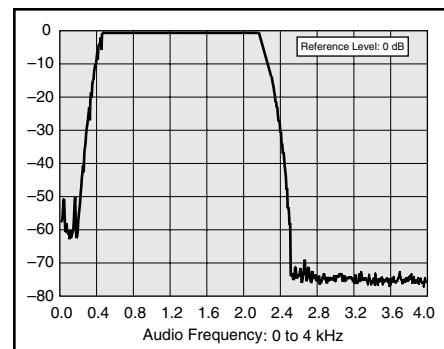
Digital signal processing provides a level of bandwidth setting filter performance not practical with other technologies. While much better than most low frequency IF LC bandwidth filters, the very good crystal or mechanical bandwidth filters in amateur gear are not very close to the rectangular shaped 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  $2400 \times 1.4 = 3360$  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 12.27** 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.

Early DSP filters operated in the audio frequency range, based on the frequency response limitations of the analog-to-digital converter (ADC) that provides the digitization for further processing. The limitation was based on the Nyquist sampling theorem (AT&T, 1924) that established that a periodic signal could be sampled and then reconstructed from its samples if sampled at least twice during each cycle of its highest frequency component. Early ADCs were limited to sampling rates less than about 50 kHz. Such a DSP filter could be added to the audio signal at the output of any receiver and work its magic on the signals in the audio stream. Unfortunately, at that time ADC and DSP that worked at typical receiver IF frequencies were either unavailable, or cost prohibitive for amateur use.

Receiver designers quickly resurrected the earlier triple conversion receiver architecture—now with IF frequencies even lower, in the 12-20 kHz range, so that DSP could be employed in the IF rather than audio section of receivers. This allowed the receiver



**Fig 12.27 — ARRL Lab measured response of an aftermarket 2400 Hz DSP bandwidth filter.**

to eliminate interfering signals before the application of automatic gain control, avoiding gain reduction in the presence of off-channel signals. As time and technology have advanced, the clocking speed and input bandwidth of ADCs continues to increase while prices continue to fall, allowing the third IF frequency to move toward the second. Soon the additional conversion step will no longer be required, and we will soon see high-performance HF receivers that apply signals at the input frequency directly to the ADC.

## 12.4 Superhet Receiver Design Details

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

### 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. Before beginning

this section, review 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 in-

crease 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, acting very much like an early spark transmitter and sending 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 lo-

**Table 12.5**  
**Typical Noise Levels (Into the Receiver) and Their Source, by Frequency**

Frequency range	Dominant noise sources	Typical level ( $\mu\text{V/m}$ ) <sup>*</sup>
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.

cal 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 rotating and other kinds of electrical machinery as well as gasoline engine ignition systems and some types of lighting. 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 higher frequency 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 particles 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 do not produce sinusoids that are as pure as we might like. A receiver local oscillator, depending on the design, will have phase-noise sidebands that extend out on either side of the nominal carrier frequency at hopefully low amplitudes. Any such noise will be transferred to the received signal and create mixing 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.

### 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:

$$F = 1 + \frac{T_E}{T_0} \quad (1)$$

where

F is the noise factor,  
 $T_E$  is the equivalent noise temperature and  
 $T_0$  is the standard noise temperature, usually 290 K.

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

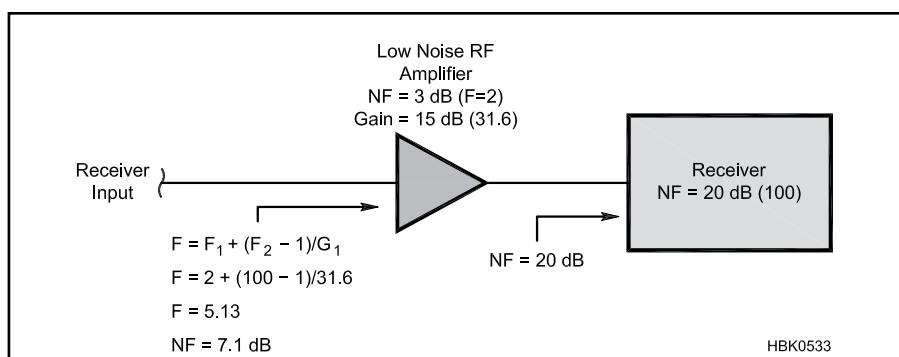
where

$N_i$  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,  $1.38 \times 10^{-23}$  Joules/kelvin, and  
B is the system noise bandwidth in Hertz.

Expressing  $N_i$  in dBm is more convenient and is calculated as:

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

If  $N_i$  is greater than the noise generated



**Fig 12.28 — The effect of adding a low noise preamplifier in front of a noisy receiver system.**

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

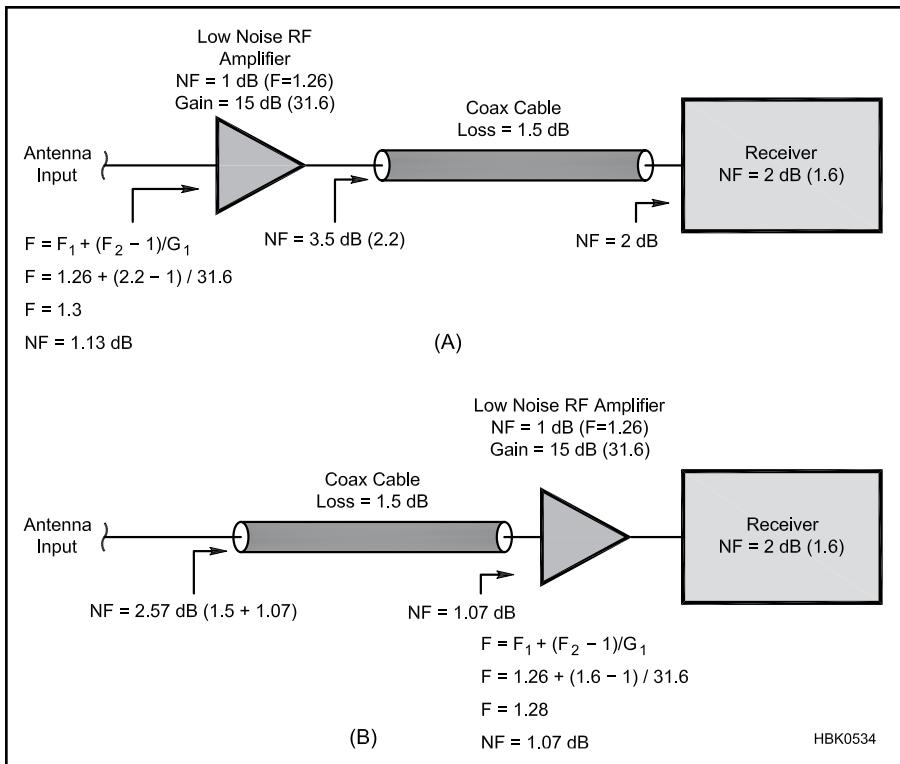
$G_1$  is the gain of the stage closest to the input.

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 12.28**. 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 12.29** shows a typical example.

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



**Fig 12.29 — Illustration of adding a VHF low noise preamplifier at the antenna (A) compared to one at the receiver system input (B).**

$$\begin{aligned} F &= F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots \\ &\quad + \frac{F_N - 1}{G_1 G_2 \dots G_{N-1}} \end{aligned} \quad (5)$$

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 to 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.

### 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, crystal filter technology has allowed down-conversion HF receivers to use a higher single IF than had early receivers, typically 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 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.

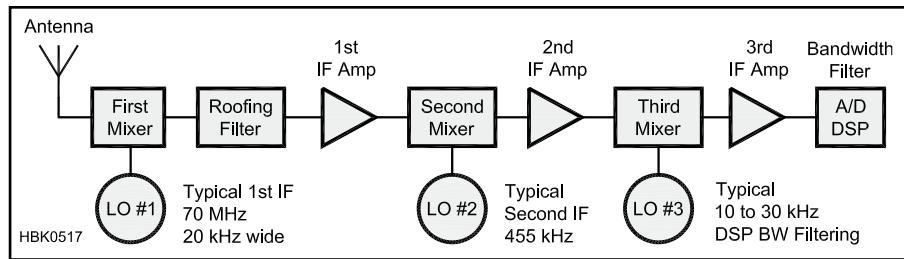
Many current HF receivers (or receiver sections of transceivers) have elected to employ an upconverting architecture. They typically have an IF in the VHF range, perhaps 60 to 70 MHz. This makes HF image rejection easy. A 30 MHz signal with a 60 MHz IF will have an image at 150 MHz. Not only is it five 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 feasible 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 12.30** 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 12.3 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.



**Fig 12.30 — Block diagram of a typical upconverting receiver using DSP for operating bandwidth (BW) determination.**

### ON-CHANNEL DYNAMIC RANGE

The signal you wish to listen to can range 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, or neighbor, 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.

While listening to a weak signal, all stages are at maximum gain. If the weak signal were at a level of S-0, a strong signal could be at S-9 + 60 dB. Using the Collins standard of S-9 equal to 50  $\mu$ V, and each S-unit a change of 6 dB, the receiver’s front-end stages would be receiving a 0.1  $\mu$ V signal and a 50,000  $\mu$ V into the front end at the same time. After a few stages of full gain amplification, the stronger signal causes amplifier clipping, which reduces the gain and thus the amplitude of the weaker signal. Of course a perfectly linear

receiver would amplify each equally until the undesired signal is eliminated at the operating bandwidth setting stage.

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.

### 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.

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

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

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 (see Note 1) reveal individual intermodulation (IM) products, recognizing that the signals have cross-modulated each other due to the nonlinear action of the circuit. 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 signal have frequencies of  $f_0 + \Delta$ , and  $f_0 + 2\Delta$ , where  $\Delta$  is some offset frequency, we have for the third-order term:

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

A good example would be interfering signals with offsets of 2 kHz and  $2 \times 2$  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 12.31 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 limiting parameter 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  $\Delta$  away from our operating frequency and another at twice  $\Delta$ . In real life, we may be trying to copy a weak signal at  $f_0$ , and have other signals at  $f_0 \pm 500, 750, 1000, 1250 \dots 5000$  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.

## 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 distor-

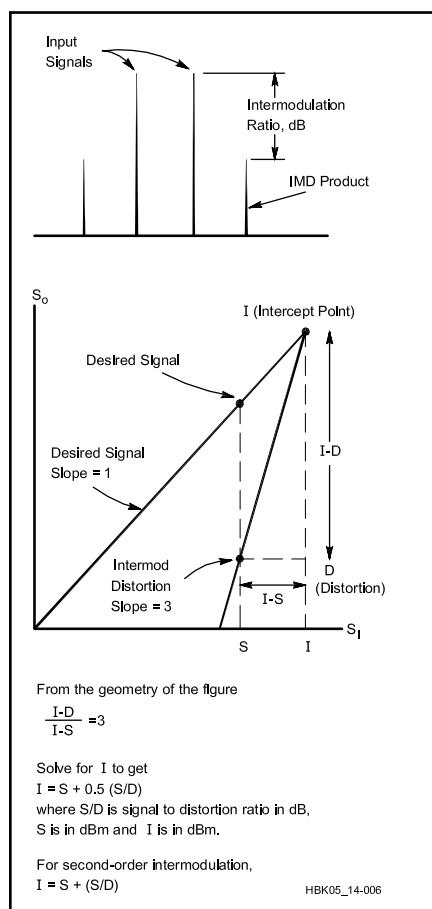
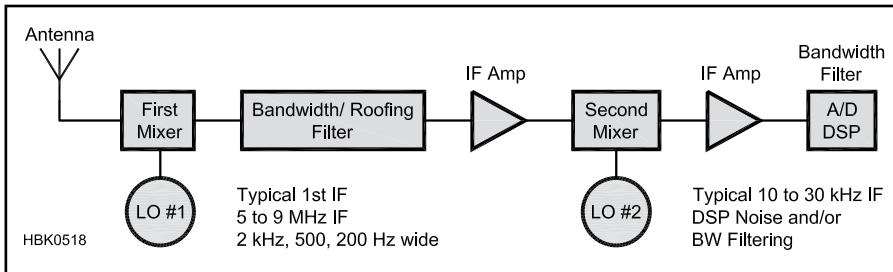


Fig 12.31 — Graphical representation of the third-order intercept concept.



**Fig 12.32 — Block diagram of a downconverting amateur band receiver with HF IF filtering to eliminate near channel interference.**

tion 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 12.32**, 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 12.32, 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.<sup>9</sup>

Now look at a typical modern general coverage receiver as shown previously in Fig 12.30. 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 70 MHz, and so far they have much wider skirts than the crystal filters used in Fig 12.32.

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.<sup>10</sup> 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.

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 12.33**. 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.

## 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.

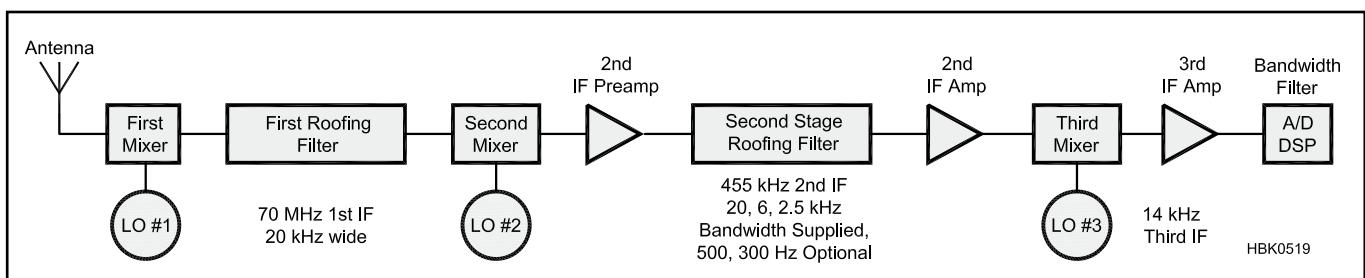
A look at recent receiver measurements indicates that receivers have IMD dynamic range with 2 kHz spacing results in the following ranges:

Upconverting with VHF IF (Fig 12.30): 60 to 80 dB

Downconverting with HF IF (Fig 12.32): 75 to 102 dB

Hybrid distributed architecture (Fig 12.33), with the Omni VII the only sample, as this is written: 82 dB

Let's take an example of what this would mean. If we are listening to a signal at S-3, 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 S-9 +27 dB, or each at S-9 +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 be at S-9 +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 how close is the proximity 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 there.



**Fig 12.33 — Block diagram of a upconverting multiple conversion receiver with distributed roofing filter architecture.**

# 12.5 Control and Processing Outside the Primary Signal Path

Discussion to date 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 AGC, 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 have  $-40$  dB quality. A correct AGC design ensures that each stage gets the right input level. It is often necessary to redesign some stages in order to accomplish this.

### THE AGC LOOP

**Fig 12.34A** shows a typical AGC loop that is often used in amateur receivers. The AGC is applied to the stages through RF decoupling circuits that prevent the stages from interacting with each other. The AGC amplifier helps to provide enough AGC loop gain so that the gain-control characteristic of Fig 12.34B is achieved. The AGC action does not begin until a certain level, called the AGC *threshold*, is reached. The THRESHOLD VOLTS input in Fig 12.34A serves this purpose. After that level is exceeded, the audio level slowly increases. The audio rise beyond the threshold value is usually in the 5 to 10 dB range. Too much or too little audio rise are both undesirable for most operators.

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

**Fig 12.35** is a typical plot of the signal levels at the various stages of a certain ham band receiver. Each stage has the proper level and a 115 dB change in input level produces a 10 dB change in audio level. A manual gain control could produce the same effect.

### AGC TIME CONSTANTS

In Fig 12.34, following the precision recti-

fier, R1 and C1 set an *attack* time, to prevent excessively fast application of AGC. One or two milliseconds is a good value for the  $R1 \times C1$  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  $C1 \times R2$  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  $R2 \times C1$  much longer, say 3 seconds or more, the AGC voltage remains almost constant until the  $R5, C2$  circuit decays with a switch selectable time constant of 100 to 1000 ms. At that time R3 quickly discharges C1 and full receiver gain is quickly restored. This type of control is appreciated by many operators because of the lack of AGC *pumping* due to modulation, rapid fading and other sudden signal level changes.

### AGC 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 adding external audio DSP filtering to a receiver. 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 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 good transient response when the

loop goes from open to closed state.]

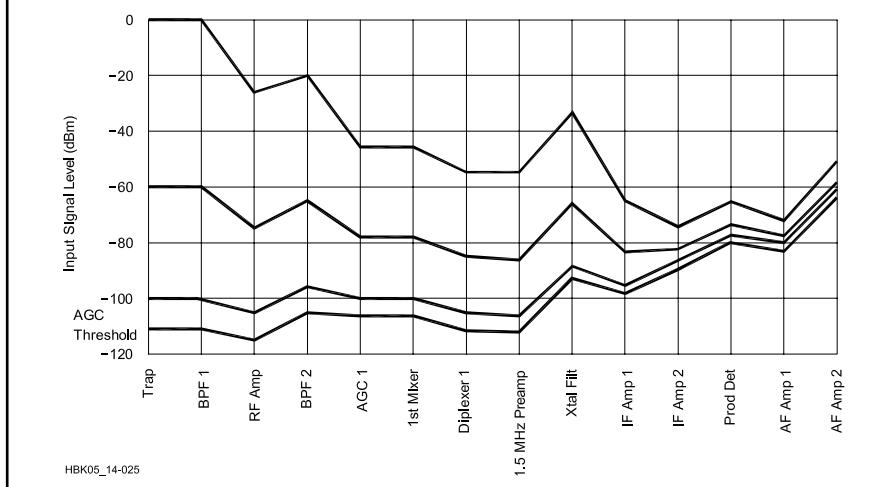
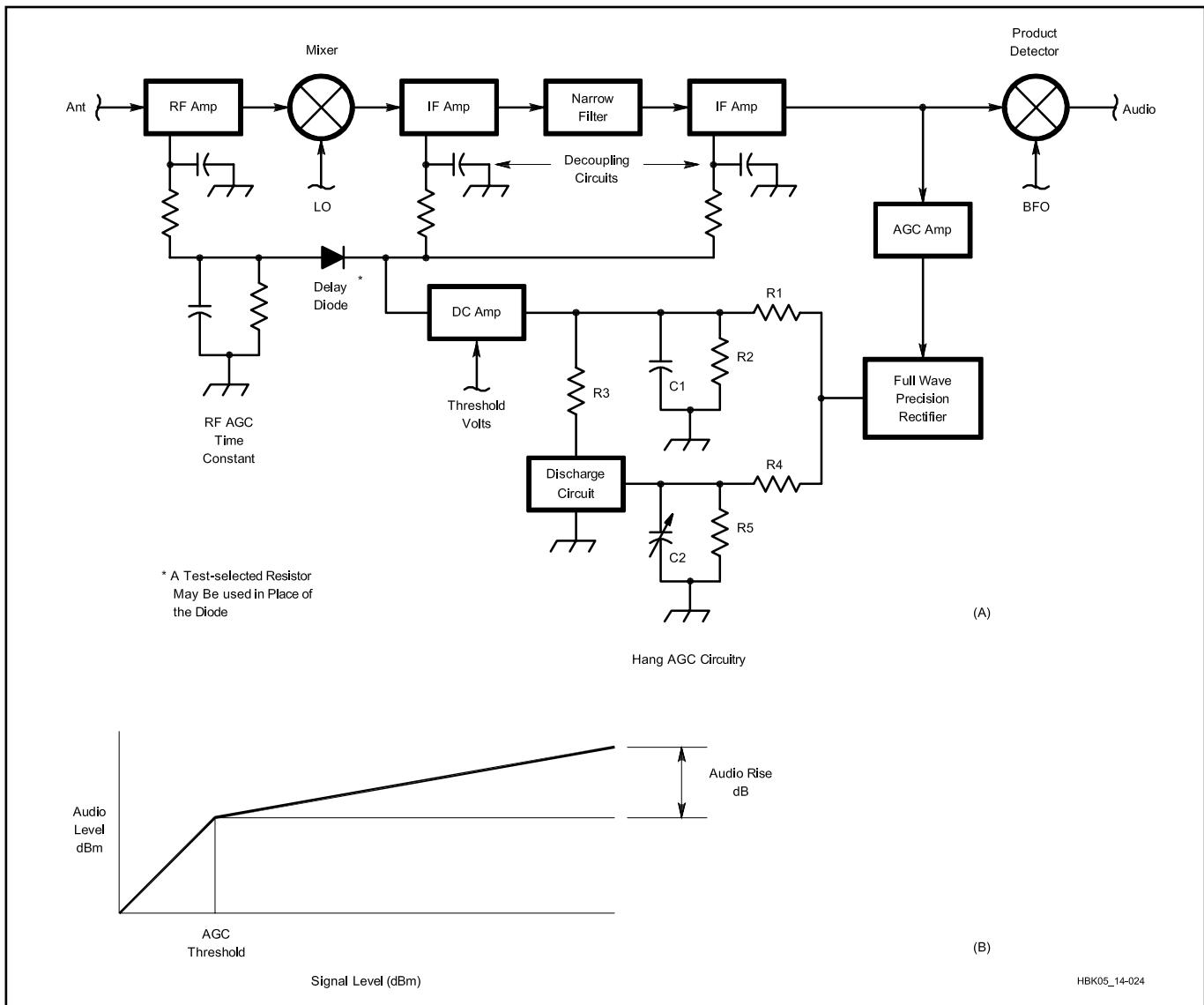
Another problem involves the narrow band-pass IF filter. The group delay of this filter constitutes a time lag in the loop that can make loop stabilization difficult. Moreover, these filters nearly always have much greater group delay at the edges of the passband, so that loop problems are aggravated at these frequencies. Overshoots and undershoots, called *gulping*, are very common. Compensation networks that advance the phase of the feedback help to offset these group delays. The design problem arises because some of the AGC is applied before the filter and some after the filter. It is a good idea to put as much fast AGC as possible after the filter and use a slower decaying AGC ahead of the filter. The delay diode and RC in Fig 12.34A 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. There are variable gain op amps and other ICs available that are intended for gain control loops.

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

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

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



### 12.5.2 Audio-Derived AGC

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

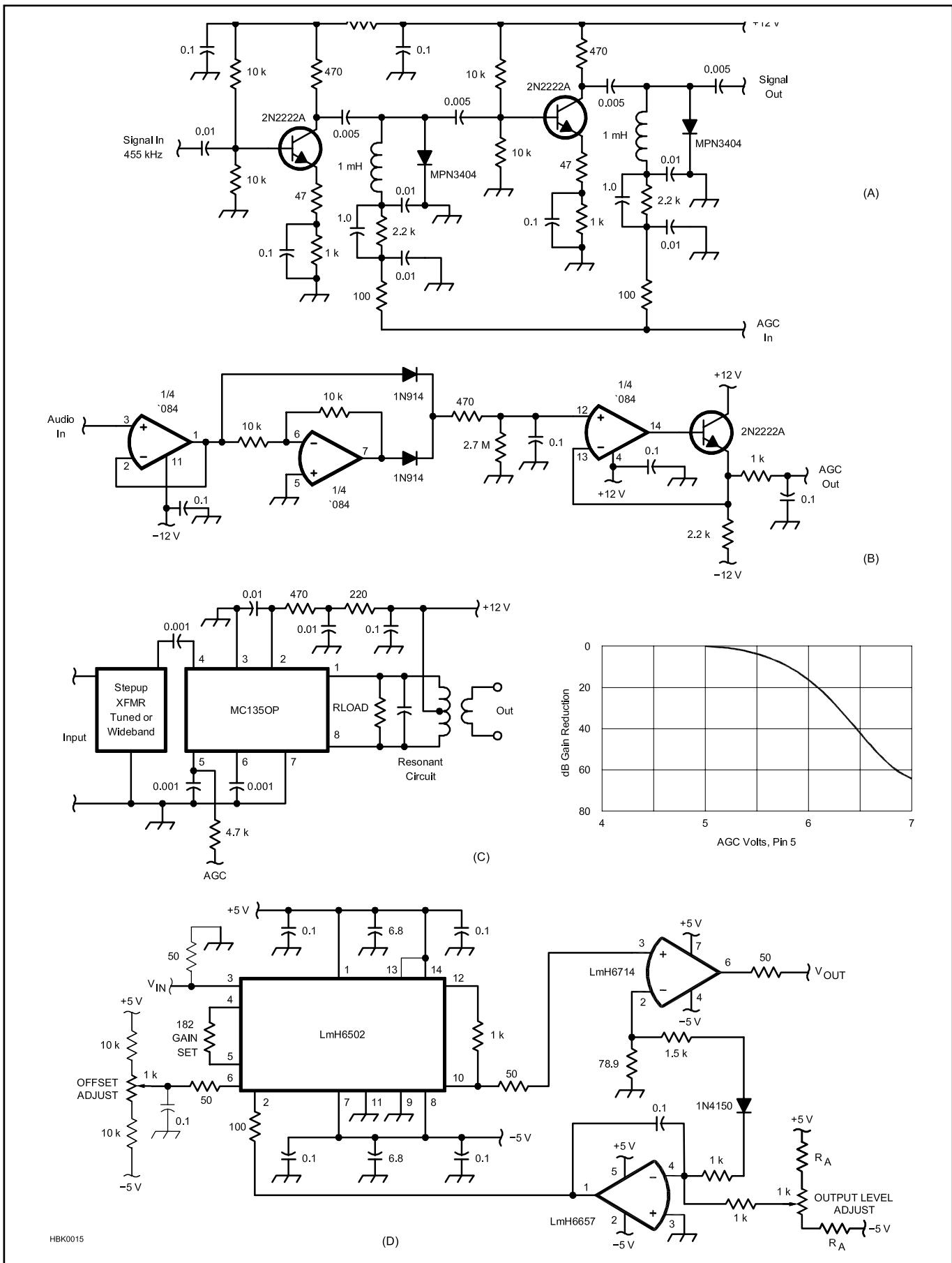


Fig 12.36 — Some gain controllable amplifiers and a rectifier suitable for audio derived AGC.

### 12.5.3 AGC Circuits

Fig 12.36 shows some gain controllable circuits. Fig 12.36A 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  $\mu\text{F}$  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 a such a low-frequency IF. Look for a device with

$\tau > 10 / (2\pi f)$  where  $\tau$  is the minority carrier lifetime in  $\mu\text{s}$  and  $f$  is the frequency in MHz.

Fig 12.36B 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 12.36C 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 12.36D shows the high-end 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  $\pm 3$  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](http://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 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 attempt to limit impulse noise was called a *noise limiter* or *clipper* circuit.<sup>11</sup> 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 loudness; second, it also reduced the loudness of loud non-noise signals and in the process distorted them considerably.

The second problem was fixed fairly early, with the advent of the *automatic noise limiter* or ANL.<sup>12</sup> 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.

### 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.<sup>13</sup> The noise limiters described so far 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 gets its input from an early stage in the receiver before much selectivity or AGC happens. It amplifies the wideband signal and detects the narrow noise pulses. 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.

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 shuts 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. In some cases the noise blanker is an extra cost option. 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.

#### OTHER TECHNIQUES

The previous techniques have been around for quite a few years and represented 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 we want to listen to. 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 can be very effective at eliminating either interfering stations or noise (both if they happen to be in the same direction).<sup>14</sup>

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

The previous techniques have been around for quite a few years and represented about all we could do about impulse noise prior to the current "digital age." Now processing capability seems to be in everything from toasters to artillery shells — why wouldn't we expect to find processing in our radios, too? Modern radios use digital processing in many ways — setting frequency, display functions, memories and, import in this discussion, digital signal processing or DSP. (The **DSP and Software Radio Design** chapter covers these topics thoroughly.)

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. It can't quite eliminate the noise, and it needs enough of the signal to figure out what's happening, so it 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. It's always worth experimenting with the radio's features to find out.

## 12.7 VHF and UHF Receivers

Most of the basic ideas presented in previous section 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 12.37A 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 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 12.37A 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 chip will be presented later.

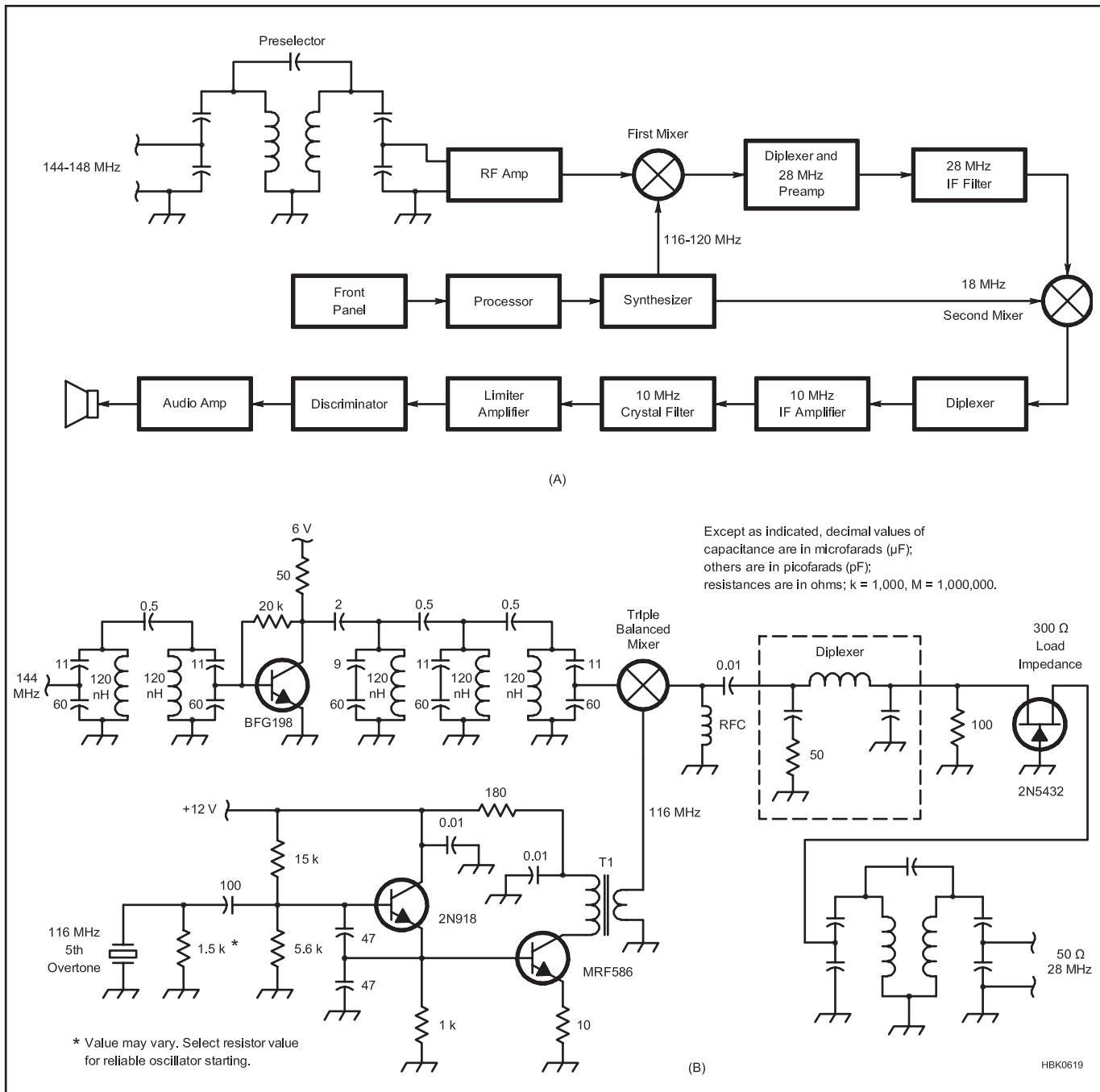


Fig 12.37 — At A, block diagram of a typical VHF FM receiver. At B, a 2 meter to 10 meter receive converter (partial schematic; some power supply connections omitted.) See this chapter's supplemental files on this book's CD-ROM for a complete 2 meter receiving converter design by Dr Ulrich Rohde, N1UL, using the *OrCAD* schematic capture and PCB layout software.

### 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 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.<sup>15</sup> 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.<sup>16</sup>

### 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 cellular telephone applications and are widely used. Fig 12.38 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.

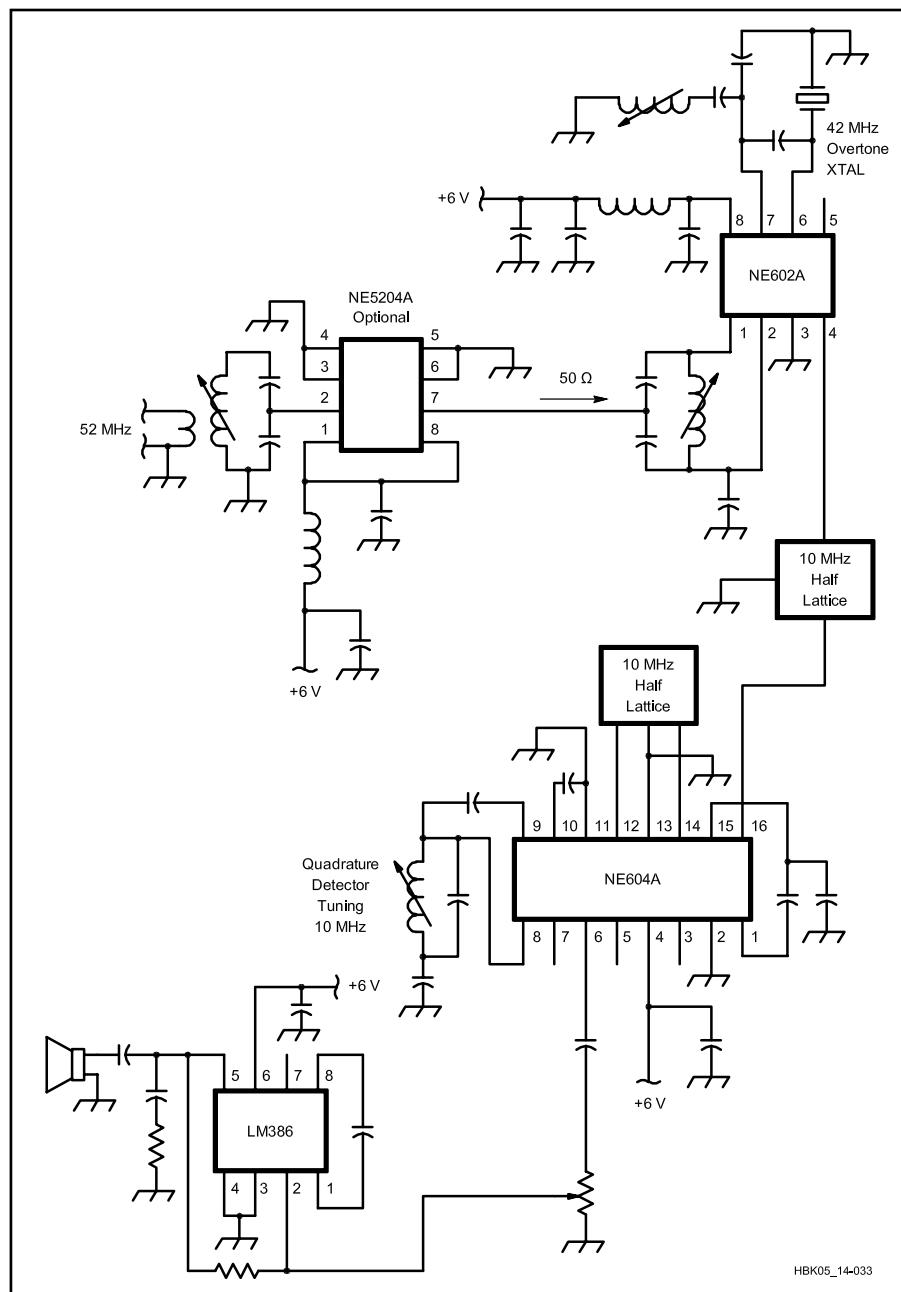


Fig 12.38 — The NE/SA5204A, NE/SA602A, NE/SA604A NBFM ICs and the LM386 audio amplifier in a typical amateur application for 50 MHz.

The diagram in Fig 12.38 is (intentionally) only a general outline that shows how chips can be combined to build complete equipment. The design details and specific parts values can be learned from a careful study of the data sheets and application notes provided by the IC vendors. Amateur 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, mul-

timode 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 12.37B shows the schematic for a high-performance converter. The front-end design was contributed by Dr 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 Chapter 10 folder of the CD-ROM that accompanies this *Handbook*.

Based on the Philips BFG198 8 GHz transistor, the 20 dB gain front-end amplifier is optimized for noise figure (NF is approxi-

mately 2.6 dB), not for input impedance. The output circuit is optimized for best selectivity. The transistor bias is designed for dc stability at  $I_C = 30 \text{ mA}$  and  $V_C = 6 \text{ V}$ . 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  $50 \Omega$ .

## 12.8 UHF Techniques

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

### 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 12.39** shows the schematic diagram and the physical construction of a typical RF circuit at 430 MHz. It is a GaAsFET preamplifier intended for low noise 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 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. 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), but very few amateurs can afford this.

Another very desirable thing would be

a circuit design and analysis program for the personal computer. Software packages created especially for UHF and microwave circuit design are available. They tend to be somewhat expensive, but worthwhile for a serious designer. Inexpensive SPICE programs are a good compromise. 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 12.40A** 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

**Fig 12.39 — 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  $\mu$ F 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,  $\frac{3}{16}$ -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,  $\frac{1}{8}$ -inch ID.

L5 — 4t #24 tinned wire,  $\frac{1}{8}$ -inch ID, spaced 1 wire dia.

Q1 — Mitsubishi MGF1402.

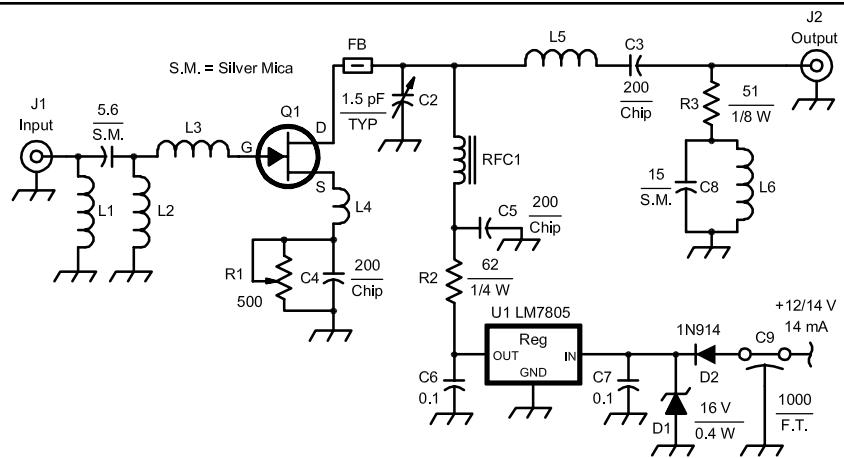
R1 — 200 or 500- $\Omega$  Cermet potentiometer (initially set to midrange).

R2 — 62  $\Omega$ ,  $\frac{1}{4}$ -W.

R3 — 51  $\Omega$ ,  $\frac{1}{8}$ -W carbon composition resistor, 5% tolerance.

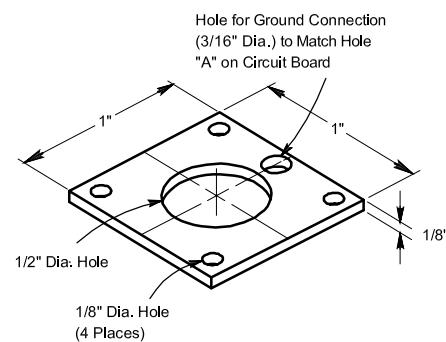
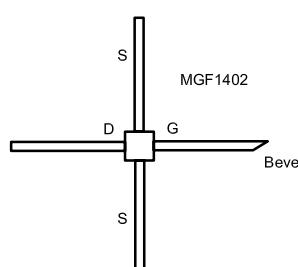
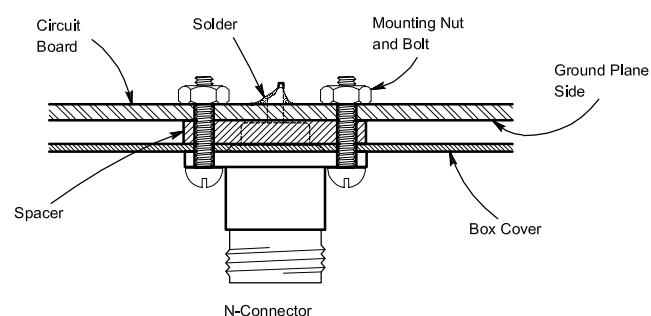
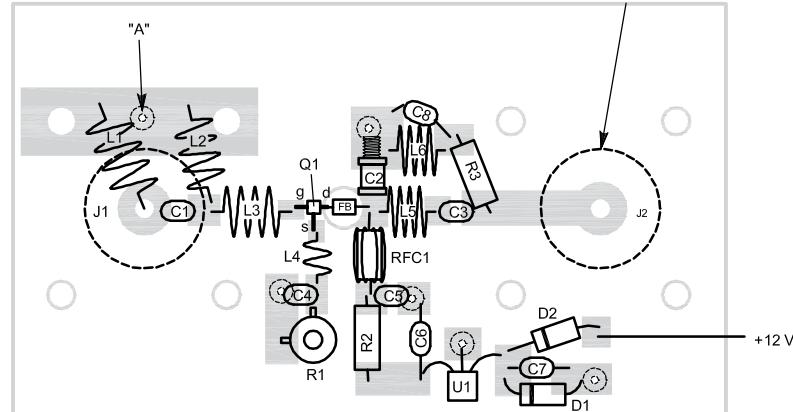
RFC1 — 5t #26 enameled wire on a ferrite bead.

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



(○) = Eyelet Soldered Both Sides

Remove Copper 1/2" Dia.  
Around Connector on  
Ground Plane Side of  
Board (2 Planes)



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 12.40B shows the cumulative noise figure (NF) of the signal path, including the 28 MHz receiver. The 1.5 dB cumulative NF of the input cavity and first RF-amplifier combination, considered by itself, is degraded to 1.9 dB by the rest of the system following the first RF amplifier. The NF values of the various components for this example are reasonable, but may vary somewhat for actual hardware. Also, losses prior to the input such as transmission line losses (very important) are not included. They would be part of the complete receive system analysis, however. It is common practice to place a low noise preamp outdoors, right at the antenna, to overcome coax loss (and to permit use of less expensive coax).

### LOCAL OSCILLATOR (LO) DESIGN

The +7-dBm LO at 874 to 900 MHz is derived from a set of crystal oscillators and frequency multipliers, separated by band-pass filters. These filters prevent a wide assortment of spurious frequencies from appearing at the mixer LO port. They also enhance the ability of the doubler stage to generate the second harmonic. That is, they have a

very low impedance at the input frequency, thereby causing a large current to flow at the fundamental frequency. This increases the nonlinearity of the circuit, which increases the second-harmonic component. The higher filter impedance at the second harmonic produces a large harmonic output.

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

A phase-locked loop using GHz range prescalers (as shown in Fig 12.40C) is an alternative to the multiplier chain. The divide-by-N block is a simplification; in practice, an auxiliary dual-modulus divider in a “swallow count” loop would be involved in this segment. The cascaded 902 MHz band-pass filters in the signal path should attenuate any image frequency noise (at RF–56 MHz) that might degrade the mixer noise figure.

### 12.8.4 Microwave Receivers

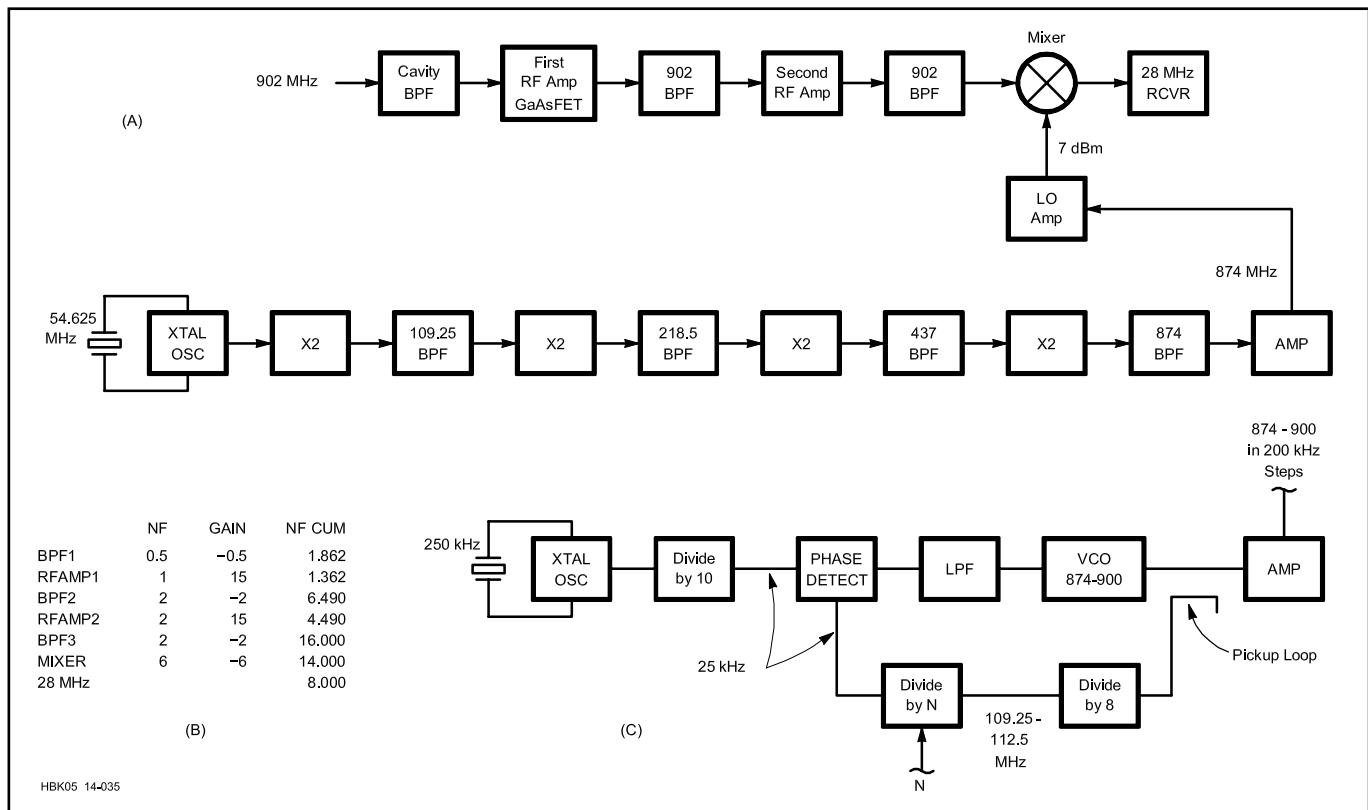
The world above 3 GHz is a vast territory with a special and complex technology and an “art form” that are well beyond the scope of this chapter. We will scratch the surface by describing a specific receiver for the 10 GHz

frequency range and point out some of the important special features that are unique to this frequency range.

### A 10 GHZ PREAMPLIFIER

**Fig 12.41B** is a schematic and parts list, **Fig 12.41C** is a PC board parts layout and **Fig 12.41A** 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- $\Omega$  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



**Fig 12.40 — A down converter 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.**

microwave amateur technology.

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

## System Performance

At microwaves, an estimation of system performance can often be performed using known data about the signal path terrain, atmosphere, transmitter and receivers systems. 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 12.42** 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

$$T_E = 290(F - 1) \quad (8)$$

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.

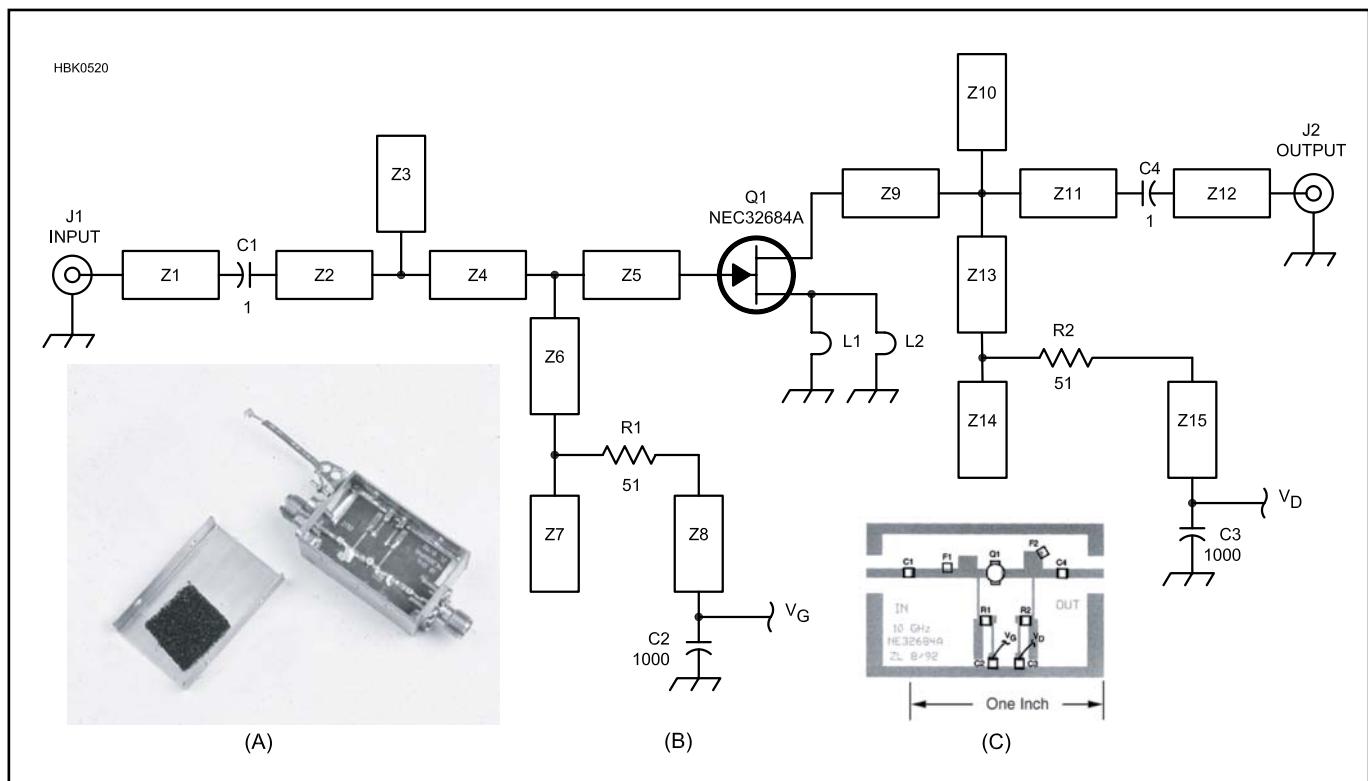
$$T_S = T_G + TE_1 + \frac{TE_1}{G_1} + \frac{TE_3}{G_1 G_2} + \frac{TE_4}{G_1 G_2 G_3} + \dots \quad (9)$$

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 290 number 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 12.42 shows, the value of this system noise factor  $F_S$  is just the ratio of the total system temperature to the antenna temperature.

Having calculated a system noise temperature, the receive system noise floor (that is, the antenna input level of a signal that would exactly equal system noise, both observed



**Fig 12.41 — 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 Durod, 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 =  $P_{XMIT}$  + GA = -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:

$$T_S = T_A + T_L + (L_L)(T_R)$$

$$T_S = 200 + 288.6 + (2.0)(864.5) = 2217.6 \text{ K}$$

$$N_S = kT_S B = 1.38 \times 10^{-23} \times 2217.6 \times 2400 = 7.34 \times 10^{-17} \text{ W} = -131.3 \text{ dBm}$$

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":

$$F_S = T_S / T_A = 2217.6 / 200 = 11.09 = 10.45 \text{ dB.}$$

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

at the receiver output) associated with that temperature is:

$$N = k T_S B_N \quad (10)$$

where

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

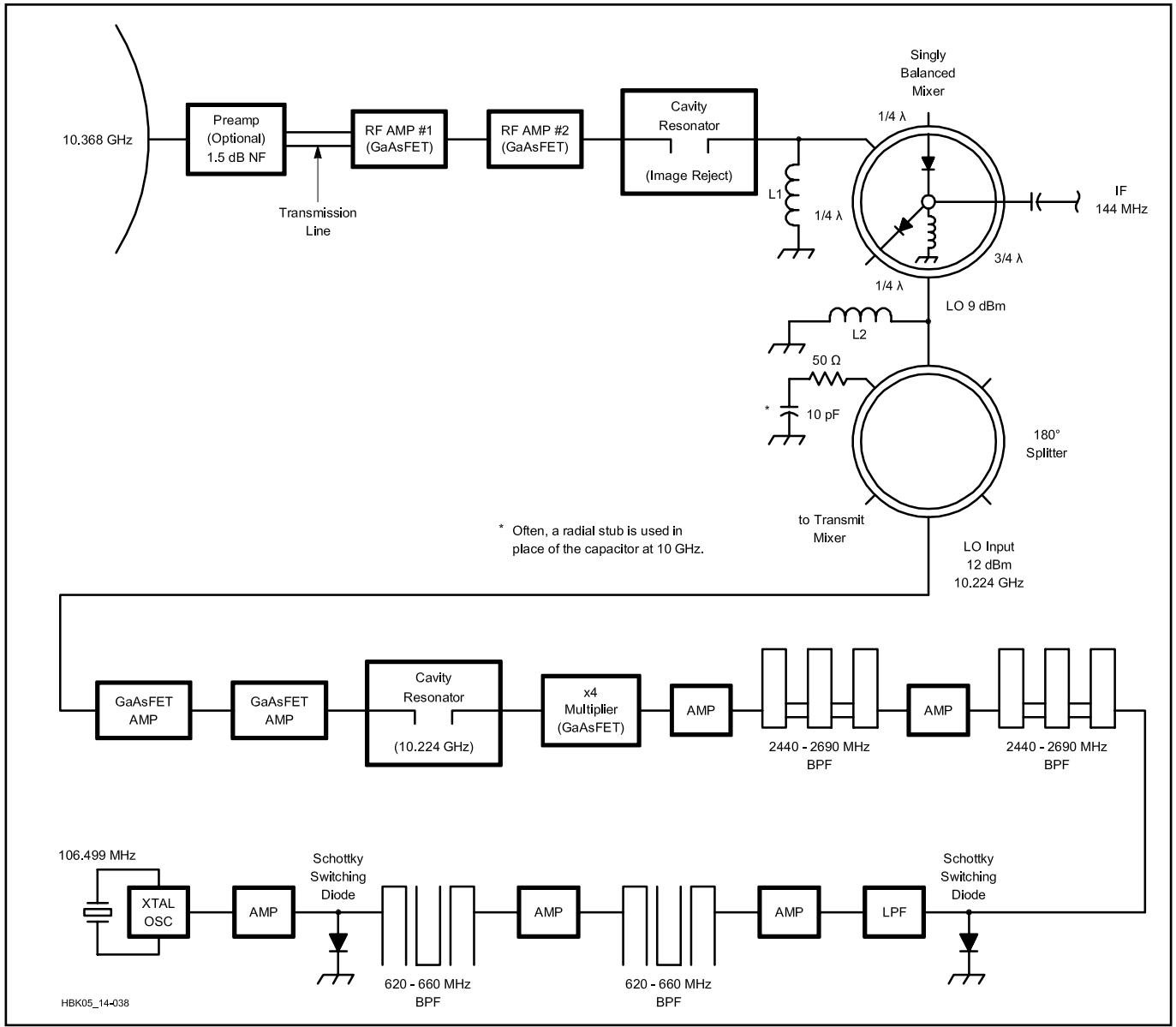
Here is a good example of amateur techniques for the 10 GHz band. The intended use for the radio is narrowband CW and SSB work, which requires extremely good frequency stability in the LO. Here, we will discuss the receiver circuit.

### Block Diagram

Fig 12.43 is a block diagram of the receiver. Here are some important facets of the design.

- 1) The antenna should have sufficient gain. At 10 GHz, gains of 30 dBi are not difficult to get, as the example of Fig 12.42 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
- 3) After two stages of RF amplification using GaAsFETs, a probe-coupled cavity resonator attenuates noise at the mixer's image frequency, which is  $10.368 - 0.288 = 10.080$  GHz. An image reduction of 15 to 20 dB is enough to prevent image frequency noise generated by the RF amplifiers from affecting the mixer's noise figure.
- 4) The single-balanced diode mixer uses a "rat-race" 180° hybrid. Each terminal of the ring is  $\frac{1}{4}$  wavelength (90°) from its closest neighbors. So the anodes of the two diodes are 180° ( $\frac{1}{2}$  wavelength) apart with respect to the LO port, but in-phase with respect to the RF port. The inductors (L1, L2) connected to ground present a low impedance at the IF frequency. The mixer microstrip circuit is carefully "tweaked" to improve system performance. Use the better mixer in the transmitter.
- 5) The crystal oscillator is a fifth-overtone Butler circuit that is capable of high stability. The crystal frequency error and drift are multiplied 96 times ( $10.224/0.1065$ ), so for

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.



**Fig 12.43 — A block diagram of the microwave receiver discussed in the text.**

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- $\Omega$  resistor improves isolation between the two output ports. A two-part *QST* article is recommended reading for this very interesting project, which provides a fairly straightforward (but not extremely simple) way to get started on 10 GHz.<sup>17</sup>

## 12.9 References and Bibliography

<sup>1</sup>CRC Standard Mathematical Tables, CRC Press, Cleveland, OH, 1974, p 231.

<sup>2</sup>W. Hayward, W7ZOI, R. Campbell, KK7B, and B. Larkin, W7PUA, *Experimental Methods in RF Design*. (ARRL, Newington, 2009)

<sup>3</sup>R. Henderson, WI5W, "A Rock Bending Receiver for 7 MHz," *QST*, Aug 1995, pp 22-25.

<sup>4</sup>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 1500 kHz with the difference, our LO would have to cover from 45 to 1045, a 23:1 range. Using the sum requires LO coverage from 955 to 1955 kHz, a range of about 2:1 — much easier to implement.

<sup>5</sup>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.

<sup>6</sup>The Baudot code is still the standard employed in wire-line teletype (TTY) keyboard terminals used by hearing-impaired people.

<sup>7</sup>A notable exception was the Hammarlund Manufacturing Corporation whose

receivers used a 2:1 tuning range. The typical bands were .5 to 1, 1 to 2, 2 to 4, 4 to 8, 8 to 16 and 16 to 32 MHz. By covering the range in six rather than four bands, the tuning was noticeably more precise, especially across the higher bands.

<sup>8</sup>R. Fisher, W2CQH, "Twisted-Wire Quadrature Hybrid Directional Couplers," *QST*, Jan 1978, pp 21-24.

<sup>9</sup>For additional discussion of the issues, see J. Hallas, "International Radio Roofing Filters for the Yaesu FT-1000 MP Series Transceivers," Product Review, *QST*, Feb 2005, pp 80-81.

<sup>10</sup>Exceptions with general coverage receivers and switchable VHF roofing filters include the ICOM IC-7800 and Yaesu FTdx9000 and FT-2000 transceivers.

<sup>11</sup>H. Robinson, W3LW, "Audio Output Limiters for Improving the Signal-to-Noise Ratio in Reception", *QST*, Feb 1936, pp 27-29.

<sup>12</sup>J. Dickert, "A New Automatic Noise Limiter", *QST*, Nov 1938, pp 19-21.

<sup>13</sup>You can demonstrate this in your receiver if it has a narrow filter. Find a band with heavy impulse noise and switch between

wide and narrow filters. If your narrow filter is 500 Hz or less, I'll bet the noise is more prominent with the narrow filter.

<sup>14</sup>R. D. Straw, Ed., *The ARRL Antenna Book*, 20th Edition (Newington 2003), pp 5-24 to 5-26.

<sup>15</sup>J. Grebenkemper, KI6WX, "Phase Noise and its Effects on Amateur Communications," *QST*, Mar and Apr 1988.

<sup>16</sup>W. Sabin, "Envelope Detection and AM Noise-Figure Measurement," *RF Design*, Nov 1988, p 29.

<sup>17</sup>Z. Lau, W1VT, "Home-Brewing a 10-GHz SSB/CW Transverter," *QST*, May and Jun 1993.

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