Contents

Chapter 13 — Downloadable Supplemental Content

Articles
- Clean, Punchy, Competitive Contest Audio Without Splatter by Jim Brown, K9YC
- SDR Simplified — Demystifying PID Control Loops by Ray Mack, W5IFS
- SDR Simplified — Noise Reduction and Adaptive Filters by Ray Mack, W5IFS
- Speech Processing: Some New Ideas by Jim Tonne, W4ENE

*Also see the Downloadable Content for the **Receivers** and **Transceivers** chapters*

HF Transmitter and Transceiver Projects
- A Fast TR Switch by Jack Kuecken, KE2QJ
- Designing and Building Transistor Linear Power Amplifiers Parts 1 and 2 by Rick Campbell, KK7B
- MicroT2 — A Compact Single-Band SSB Transmitter by Rick Campbell, KK7B
- MkII — An Updated Universal QRP Transmitter by Wes Hayward, W7ZOI
- TAK-40 SSB/CW Transceiver by Jim Veatch, WA2EUJ
- The Rockmite — A Simple Single-Band CW Transceiver by Dave Benson, K1SWL (article and folder with HEX file)
- The Tuna Tin 2 Today by Ed Hare, W1RFI
- Transmitter and transceiver projects from previous editions of the *ARRL Handbook.*

VHF/UHF Transmitter and Beacon Projects
- A 2-Meter Transmitter for Fox Hunting by Mark Spencer, WA8SME
- A Microwave Transverter Controller by Hamish Kellock, OH2GAQ
- CW Beacon Exciter for 50 MHz by Michael Sapp, WA3TTS (article and parts list)
- Simple Frequency Doublers with High Performance by Paul Wade, W1GHZ
- VHF and UHF CW Beacons by Michael Sapp, WA3TTS
- VHF Open Sources by Rick Campbell, KK7B (article and parts placement)

Chapter 13
Transmitting

As with receiving, analog heterodyne transmitting is rapidly being displaced by SDR techniques. Nevertheless, many of the same functions are required, regardless of whether they are implemented in analog electronics or digitally by software. As such, each function is discussed from both the traditional analog perspective and from the SDR perspective. Subsequent editions will continue to evolve as practices change and standardize on new approaches.

The discussion on microphone selection and audio optimization is by Jim Brown, K9YC, and a set of analog audio signal processing circuits is provided by Jim Tonne, W4ENE. Sections on SDR implementations of basic functions are taken from SDR: Simplified columns in *QEX* by Ray Mack, W5IFS and from material contributed by Steve Hicks, N5AC. Material for the section on PIN Diode RF Switching was contributed by Hans Summers, GØUPL. Bob Allison, WB1GCM, and Adam Farson, VA7OJ/AB4OJ, contributed material on transmitter testing.

The RF power stages of a transmitter are still firmly in the analog camp although design has greatly simplified through integrated circuits and amplifier modules. Circuits for power levels above 100 W (at HF) are covered in the **RF Power Amplifiers** chapter. The **DSP and SDR Fundamentals** chapter has more information on digital techniques and architectures.

Projects included with previous editions have been collected into a set of projects in the online content accompanying this book.

Transmitter technology has advanced in a parallel process similar to that of the technology of receivers. While transmitters are composed of many of the same named blocks as those used in receivers, it's important to keep in mind that there are significant differences. An RF amplifier in a receiver may deal with amplifying picowatts while one in a transmitter may output up to kilowatts. While the circuits may even look similar, the size of the components, especially cooling systems and power supplies, may differ significantly in scale. Still, many of the same principles apply.

Transmitters also make use of many of the same functional blocks as receivers, such as filters and mixers. You will find additional material on these functions covered in the **Receiving** chapter. Oscillators are covered in the **Oscillators and Synthesizers** chapter. Elements of these functions that are pertinent to transmitter design are covered in this chapter but not duplicated.

Transmitters (and transceivers) using vacuum tubes and solid-state matching circuits are likely to present hazardous voltages. At higher power levels RF exposure issues must be considered — review the **Safety** chapter for more information. Techniques for transmitter measurement are covered in the **Test Equipment and Measurements** chapter.

13.1 Characterizing Transmitters

13.1.1 FCC Rules

A survey of the FCC rules in Part 97 shows that most of them are about transmitted signals! Thus, amateurs should have a clear idea of what required their transmitters are expected to satisfy.

There are two important definitions that FCC Part 97 applies to amateur signals (also known as *emissions*):

97.3(a)(8) *Bandwidth*. The width of a frequency band outside of which the mean power of the transmitted signal is attenuated at least 26 dB below the mean power of the transmitted signal within the band. (See Figure 3.16 later in this chapter.)

97.3(a)(43) *Spurious emission*. An emission, or frequencies outside the necessary bandwidth of a transmission, the level of which may be reduced without affecting the information being transmitted.

There are also some important definitions in Part 2.1031 – 2.1060 that apply to all wireless services:

2.202 *Bandwidths.*

(a) *Occupied bandwidth*. The frequency bandwidth such that, below its lower and above its upper frequency limits, the mean powers radiated are each equal to 0.5 percent of the total mean power radiated by a given emission.

(b) *Necessary bandwidth*. For a given class of emission, the minimum value of the occupied bandwidth sufficient to ensure the transmission of information at the rate and with the quality required for the system employed, under specified conditions. Emissions useful for the good functioning of the receiving equipment as, for example, the emission corresponding to the carrier of reduced carrier systems, shall be included in the necessary bandwidth.

The measurement of bandwidth (and other characteristics) is covered by Part 2.1031-2.1060.

Part 97, Subpart D, sets forth a number of technical standards for signals. Part 97.307 — Emission Standards covers all of the signal characteristics from signal quality through symbol rates. Part 97.313 covers transmitter power limits by band. Parts 97.315 and 97.317 set the standards for certification of external RF power amplifiers.

The FCC rules are easily accessible online via the ARRL website at **www.arrl.org/part-97-amateur-radio** and so are not reproduced here.

13.1.2 Performance Measurements

The ARRL Lab has created a series of standardized tests for product review and compliance testing. The entire test program for receivers and transmitters is described in the book *Amateur Radio Transceiver Performance Testing* by Bob Allison, WB1GCM. Important sections are summarized here. The book also addresses a number of other significant parameters that affect various elements of on-the-air performance. The chapter **Test Measurements and Equipment** also discusses transceiver performance.

EMISSION STANDARDS

A spurious signal is any signal (unwanted) other than the intended (fundamental) transmitted signal. Any oscillator will create unwanted products, such as spurs (spurious emissions) and harmonics (multiples of the fundamental frequency). Some consider harmonics to be spurious emissions but they are addressed separately in this book.

Spurs are usually found near or around the fundamental frequency. They can be reduced with appropriate circuit design and, to some extent, with filtering. Harmonics are suppressed to an acceptable level with the use of band-pass filters. An “acceptable level” is one that will not create interference on two times the fundamental frequency (second harmonic), three times the fundamental frequency (third harmonic), and so on.

The single most important standard any transceiver operating in the amateur bands must meet is the FCC Part 97 rules for emissions. According to Part 97.307(d) and 97.307(e), an HF transceiver’s emissions must have at least 43 dB of spurious emission and harmonic suppression below 30 MHz. For 30 MHz through 225 MHz, the spurious and harmonic suppression must be 60 dB. (Emission designators are discussed in the **Modulation** chapter.)

CARRIER AND SIDEBAND SUPPRESSION

While there are no FCC specifications for suppression of the carrier and unwanted sideband for SSB signals, the ARRL Lab has established 60 dB of suppression as good practice. 50 dB of suppression is considered a minimum acceptable level.

INTERMODULATION DISTORTION (IMD)

Spurious emissions caused by IMD are definitely of concern, even though they might satisfy the rules on signal bandwidth. For today’s transceivers, a third-order product measurement of 30 dB below PEP is typical, 35 dB is considered good, and 25 dB below PEP is mediocre. A clean signal with the lower possible IMD products should be a key consideration for purchasing or operating a transceiver. It is also important to remember that even the best transmitter can be misadjusted to put out an excessively wide signal that creates a lot of interference.

Figures 13.1 and **13.2** show a typical example of a signal with good IMD performance (Figure 13.1) and an unacceptably wide signal (Figure 13.2). The signal in Figure 13.1 has acceptable 3rd-order IMD products and the higher-order products diminish rapidly farther from the main signal. The signal in Figure 13.2, however, has mediocre performance for the 3rd-order products and the higher-order products are far too high out to ±6 kHz from the carrier frequency. This signal would interfere with at least two SSB contacts to each side of it.

Using AFSK to generate multi-tone FSK (MFSK) with multiple simultaneous tones or simultaneous FSK signals in the same transmitter (for example, the fox-and-hound or DXpedition mode of FT8) can result in IMD similar to that produced by the two-tone test method. As with the two-tone method, the IMD can occur from non-linearities in the audio processing path even if the FSK tones are undistorted and not over-driving the audio input.

CW AND FSK KEYING

The shape of a transmitter’s keying waveform impacts the quality and effectiveness of its transmissions. By observing the keying waveform, it is possible to assess whether it has characteristics that will affect the quality of the transmitted signal and its impact on adjacent signals.

There are two primary signatures of problems that are can be detected by observing the shape of the keyed waveform. The first is *overshoot* in which the transmitter sends a short high-power transient at the beginning

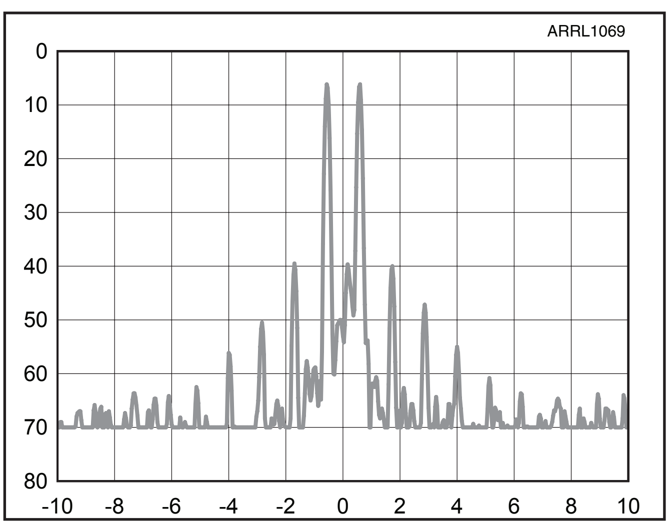


Figure 13.1 — Two-tone IMD test results for a good transmitted signal. The 3rd-order products are close to 40 dB below peak power. Scales are 2 kHz/div (horizontal) and 10 dB/div (vertical).

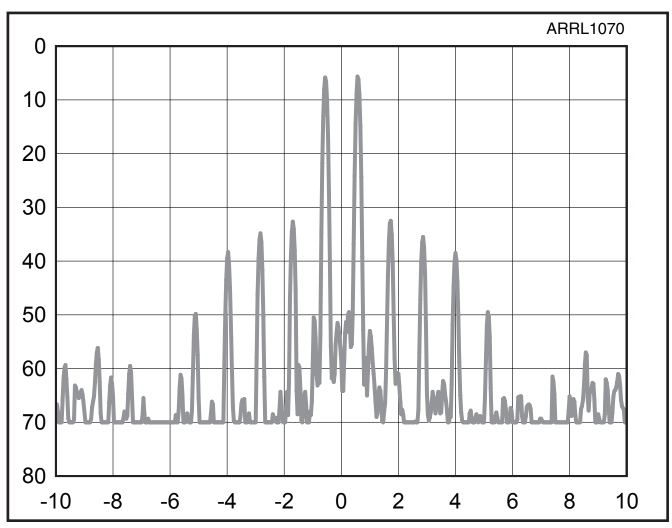


Figure 13.2 — Third-order IMD products are acceptable at 32 dB below PEP but the higher-order products are unacceptably high. Scales are 2 kHz/div (horizontal) and 10 dB/div (vertical).

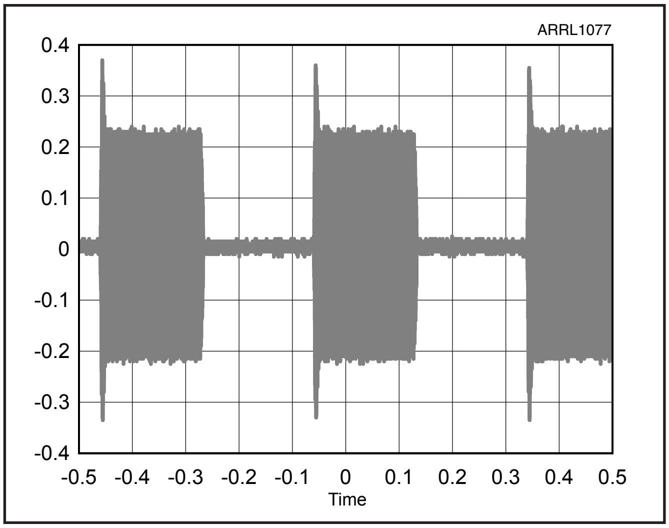


Figure 13.3 — A transient at the leading edge of a keying waveform.

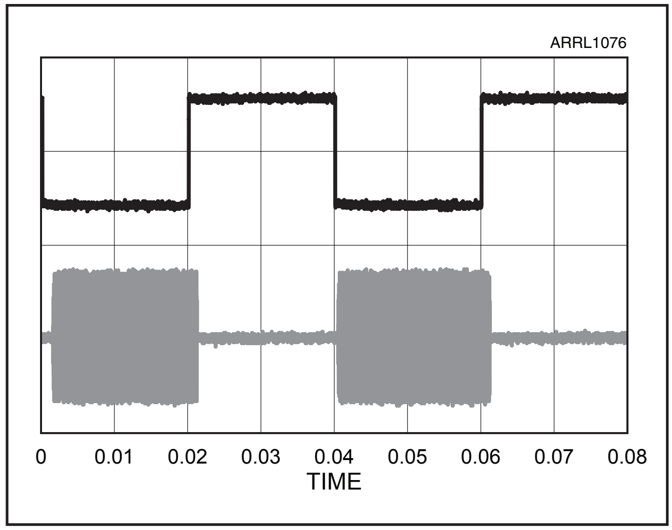


Figure 13.4 — Abrupt, nearly vertical rising and falling edges of "hard" keying.

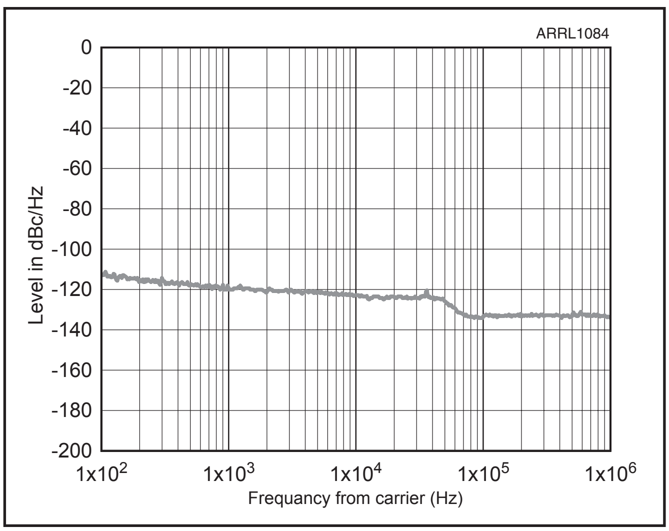


Figure 13.5 — A transmitter with a relatively low composite noise profile. This transmitter is unlikely to cause significant interference on adjacent channels or other bands.

of a transmission (see **Figure 13.3**). This is caused by the power control subsystem not reacting properly to the rapid change in output power. The transient can cause an amplifier's power protection circuitry to activate or trip. It can also cause a key click on adjacent channels. The system at fault can be the ALC system, a software problem, or some other internal power control mechanism.

The second signature is too-rapid rise and fall times or abrupt edges and corners of the envelope as shown in **Figure 13.4**. There may also be discontinuities or artifacts present during the rising and falling edges of the waveform. These create distortion products that extend well beyond the CW signal's necessary bandwidth as key clicks that cause interference to adjacent channels.

FSK keying can also generate spurious products and sidebands outside the intended channel. An abrupt change in tone waveform as the frequency changes creates keying sidebands, just as turning on and off a CW signal does. These artifacts can also disrupt demodulation by the receiver. There are various techniques for shifting between frequencies that minimize these artifacts. (See the **Modulation** chapter.) Direct-FSK which uses a digital signal to shift the transmit frequency directly by changing the frequency of an oscillator can create similar artifacts.

COMPOSITE (PHASE) NOISE

Composite noise is composed of both amplitude and phase noise on the transmitted signal. We are concerned with noise present from 100 Hz to 1 MHz from the carrier frequency of the signal. (See **Figure 13.5**) Composite noise can raise the noise floor of adjacent channels and in severe cases, across an entire band or on multiple bands. (See the discussions on phase noise in this chapter and in the **Transceiver Design Topics** chapter.)

In addition to key clicks, composite noise is also transmitted with CW and FSK signals, so it is appropriate to consider the overall set of keying sidebands output by the transmitter. This parameter is generally measured by keying the transmitter with continuous dits (an approximately square wave keying waveform) at high speed while using a spectrum analyzer to observe the resulting output on frequencies a few kHz above and below the carrier frequency.

ALC OVERSHOOT

The purpose of ALC (Automatic Level Control) is to prevent the various stages in the transmitter from being overdriven. Overdrive can generate out-of-band distortion or cause excessive power dissipation, either in the amplifiers or in the power supply. ALC operates by sampling the peak amplitude of the modulation (the envelope variations) of the output signal and developing a dc gain-control voltage that is applied to an earlier amplifier stage. An ALC voltage can also be developed externally by an RF power amplifier and applied to a transmitter acting as an exciter.

The ALC system will respond to variations in the output signal amplitude. If the response is too slow, output power can be too high until the ALC system regains control. This is called *ALC overshoot*, and the result is distortion of the output signal, usually at the leading or trailing edge of keyed waveforms like CW or of modulations with rapidly changing amplitudes such as speech. The transient at the leading edge of a CW waveform in Figure 13.3 is typical but there can also be artifacts in the rising and falling edges of the waveform. The ALC system may have different time responses on attack and decay as it responds to variations in the signal. Waveform distortion and artifacts created by the ALC system create spurious emissions on adjacent channels.

For digital modes that are not constant-power, ALC action that distorts the waveform can also make it more difficult for a demodulator to recover digital data. For these modes, power output should be reduced to where the ALC system is inactive or the ALC is turned off.

An ancillary function of the ALC system is reduction of drive power (excitation) to a safe level when a load mismatch is detected, typically by an SWR meter at the output of the transmitter or external RF power amplifier.

COMPRESSION AND SPEECH PROCESSING

The purpose of compression and speech processing functions is to increase a voice signal's average power so the output RF signal "sounds louder" to the receiving station. (See the Speech Amplification and Processing section of this chapter.) This can be performed on the baseband audio signal, on an IF signal in the exciter chain, or low-level modulated RF signal before it is amplified. Properly configured, compression should not increase spurious emissions on adjacent channels. Compression should not be used on digital modulation as the resulting distortion products will make demodulation more difficult.

"HOT SWITCHING" AND TURNAROUND

The transmitter's output SEND or KEY control signal operates in conjunction with the RF generation circuits to indicate that the transmitter is in a transmit state and RF output may be present. It is important that the SEND signal becomes active before RF output is generated and stays active until RF output is no longer present. This allows external devices such as preamplifiers and power amplifiers to switch safely between receive and transmit before RF is applied to the antenna system or amplifier input.

If RF output power is present before or after the SEND signal is active, that is called *hot switching*. Hot switching can damage sensitive receiving components or cause switch and relay contacts to be damaged from closing or opening when high voltages or currents are present. Most transmitters have adjustable delay settings to control the time between when a keying signal is asserted and RF is actually generated.

A related metric is *turnaround time* which is the time between when a transmitter stops outputting RF power and a receiver (usually part of an integrated transceiver) can again receive signals. This affects high-speed, full break-in (QSK) CW operation and digital modes that use ARQ protocols in which each transmitted packet is acknowledged by the receiver. (See the **Digital Protocols and Modes** chapter.) The shorter the recovery time, the faster CW speeds and higher packet transmission rates can be.

AUDIO FREQUENCY RESPONSE AND DEVIATION

A transmitter intended for speech modes must be able to transmit audio frequencies from 300 Hz to 3000 Hz although filtering may be applied to reduce this bandwidth under crowded conditions. AFSK digital modes such as FT8 can make use of this entire range. SDR transmitters can have an even wider audio response at the higher end of the range. Low frequency audio components are usually attenuated or "rolled off" below 300 Hz to avoid interfering with the carrier which would cause distortion in the demodulated signal. Many transceivers feature audio equalization on both transmit and receive to accommodate different voices, operating styles, and microphones. This will also affect frequency response and any tests should note the equalization settings. Equalization should be turned off when using digital modulation via the speech processing circuits or functions.

In FM transmitters, deviation is a function of both input audio amplitude and frequency. Insufficient deviation will result in low audio level recovery. Excessive deviation will increase the signal bandwidth and the resulting sidebands can cause interference on adjacent channels. FCC rules (§97.307(f)(1)) also limit angle-modulated signals (both FM and PM) below 30 MHz to a maximum modulation index of 1 to control bandwidth.

LATENCY

Latency of a transmitter is the time interval between an event at the transmitter's modulation input and the corresponding change in RF output at the transmitter's output. The transmitter input can be the CW keying input, the microphone input (used for both voice and AFSK), a USB audio codec, a dedicated data audio input for AFSK, or an FSK keying line. Each is measured by using a dual-channel oscilloscope to observe both the modulation input and RF output. If a USB audio codec is the input, the input to the USB encoding device is the point of input measurement.

Latency in an analog transmitter is generally quite low, limited primarily by delays in any keying sequences that switch filters and amplifiers. An electromechanical relay will create an additional latency of several msec. In an SDR, the processing time in DSP filters dominate the delay.

As with a receiver, whether latency is considered excessive depends on the mode being used. For example, some digital modes that use ARQ protocols (see the **Digital Protocols and Modes** chapter) have a maximum response time before a data frame is assumed to have been lost. The resulting re-transmissions can slow the system or cause the link to fail completely. Excessive latency when monitoring one's own speech or manually sent CW signal can become quite confusing, as well.

13.1.3 CCS, ICAS, and IMS Ratings

Amateur equipment is usually advertised as being rated at some level of ICAS service. There are several related types of operating service. From the RCA Transmitting Tubes manual (TT-4, 1956), the following definitions are obtained:

Continuous Commercial Service (CCS) covers applications involving continuous tube operation in which maximum dependability and long tube life are the primary considerations.

Intermittent Commercial and Amateur Service (1CAS) covers applications in which high tube output is a more important consideration than long tube life. The term "Intermittent Commercial" in this title applies to types of service in which the operating or "on" periods do not exceed 5 minutes each, and are followed by "off" or stand-by periods of the same or greater duration. The term "Amateur Service" covers other applications where operation is of an infrequent or highly intermittent nature, as well as the use of tubes in "amateur" transmitters. ICAS ratings generally are considerably higher than CCS ratings.

Although the ability of a tube to produce greater output power is usually accompanied by a reduction in tube life, the equipment designer may decide that a small tube operated at its ICAS ratings meets the requirements better than a larger tube operated within CCS ratings.

Intermittent Mobile Service (IMS) covers applications in which very high power output for short periods is required from equipment of the smallest practical size and weight. Tube ratings for IMS service are based on the premise that transmitter "on" periods do not exceed 15 seconds each, and are followed by "off" periods of at least 60 seconds duration. In equipment tests, however, maximum "on" periods of not more than 5 minutes each followed by "off" periods of at least 5 minutes are permissible, provided the total "on" time of such test periods does not exceed 10 hours during the life of the tube. Although tubes operated under IMS ratings may have a life of only about 100 hours, the use of these ratings is economically justified where high power must be obtained intermittently from very small tubes.

If equipment is specified as meeting

Continuous Commercial Service requirements, that is exactly how the equipment or device is certified to perform. Very few amateur transmitters or amplifiers, however, are operated that way and if they had to meet the CCS level of service, would be quite expensive.

ICAS, on the other hand, is just a name for the type of service the equipment is expected to provide. The description is intended to apply to use in which the equipment is operated with a 50% duty cycle having equal "on" and "off" periods of 5 minutes or less. Terms like "ICAS 50%" have no standard meaning. That could refer to the original ICAS definition or it could mean a period of 50% duty cycle operation followed by an equal period of no operation at all. The safest choice is to ask the manufacturer directly for their interpretation of ICAS or any similar rating and not assume anything.

Regardless of which rating is used, the intent is to describe the conditions under which the equipment (or devices) may be operated and still be expected to meet the performance specifications. Typically, devices are *de-rated* from their maximum ratings in order to meet the performance specifications but there are many other considerations to take into account when the equipment is designed. For example, equipment designed to operate at a high ambient temperature would have to be de-rated more than equipment operated at a lower temperature.

13.2 Transmitter Architecture

In the following section, "in-band" refers to signal frequencies within the bandwidth of the desired signal. For example, for an upper sideband voice signal with a carrier frequency of 14.200 MHz, frequencies of approximately 14.2003 to 14.203 would be considered in-band. Out-of-band refers to frequencies outside this range, such as on adjacent channels.

13.2.1 Upconverting Heterodyne Architecture

Figure 13.6 shows a traditional superheterodyne architecture for transceivers in which the sideband filter, some amplifiers, and other filters are shared between transmit and receive modes through the use of extensive switching. A limitation of that architecture is that it is difficult to provide operation on frequencies near the first IF. The typical transceiver designer selected a first IF frequency away from the desired operating frequencies and proceeded on that basis.

New amateur bands at 30, 17, and 12 meters were approved at the 1979 ITU World Administrative Radio Conference. The difficulties of managing image rejection on the new bands and the desire for continuous receiver coverage of LF, MF and HF bands (general-coverage receive) required a significant change in the architecture of receivers and transceivers. Thus, the upconverting architecture discussed in the **Receiving** chapter became popular in the 1980s and, with a few notable exceptions, became almost universal in commercial products over the following decade.

The solution was to move to the upconverting architecture shown in **Figure 13.7**. By selecting a first IF well above the highest receive frequency, the first local oscillator can cover the entire receive range without any

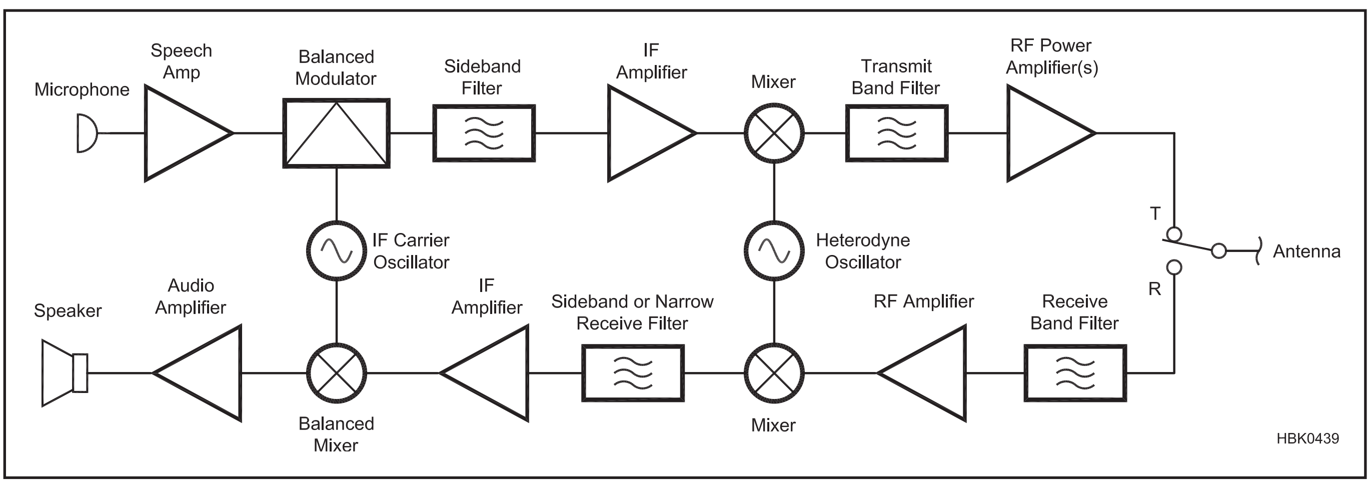


Figure 13.6 — Block diagram of a simple SSB transceiver sharing oscillator frequencies.

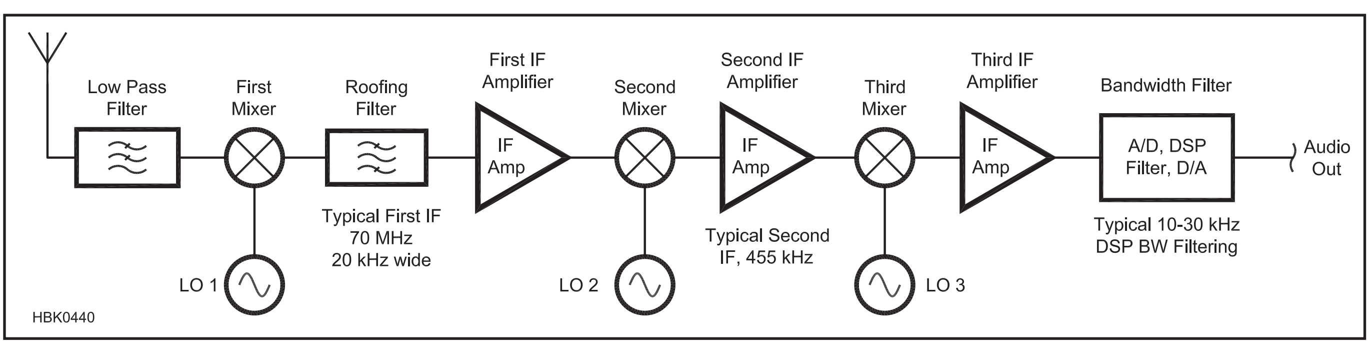


Figure 13.7 — Simplified block diagram of upconverting general coverage transceiver, receiver section shown.

gaps. With the 70 MHz IF shown, the full range from 0 to 30 MHz can be covered by an LO covering 70 to 100 MHz, less than a 1.5:1 range, making it easy to implement with modern PLL or DDS technology. Note that the high IF makes image rejection very easy and, rather than the usual tuned band-pass front end, we can use more universal low-pass filtering. The low-pass filter is generally shared with the transmit side and designed with octave cutoff frequencies to reduce transmitter harmonic content. A typical set of HF transceiver low-pass filter cut-off frequencies would be 1, 2, 4, 8, 16 and 32 MHz.

This architecture offers significant benefits. By merely changing the control system programming, any frequency range or ranges can be provided with no change to the architecture or hardware implementation. Unlike the more traditional transceiver architecture (Figure 13.6), continuous receive frequency coverage over the range is actually easier to provide than to not provide, offering a marketing advantage for those who also like to do shortwave or broadcast listening.

IF FILTERS

The desired IF filter response is shown in **Figure 13.8A**. The reduction of the carrier frequency is augmented by the filter response. It is common to specify that the filter response be down 20 dB at the carrier frequency. Rejection of the opposite sideband should (hopefully) be 60 dB, starting at 300 Hz below the carrier frequency, which is the 300-Hz point on the opposite sideband. The ultimate attenuation should be at least 70 dB. This would represent a very good specification for a high quality transmitter. The filter passband should be as flat as possible (with passband ripple less than 1 dB or so).

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

Filters require special attention to their terminations. The networks that interface the filter with surrounding circuits should be accurate and stable over temperature. They should be easy to adjust. One very good way to adjust them is to build a narrow-band sweep generator and look at the output IF envelope with a logarithmic amplifier, as indicated in Figure 13.8B. There are three goals:

• The driver stage must see the desired load impedance.

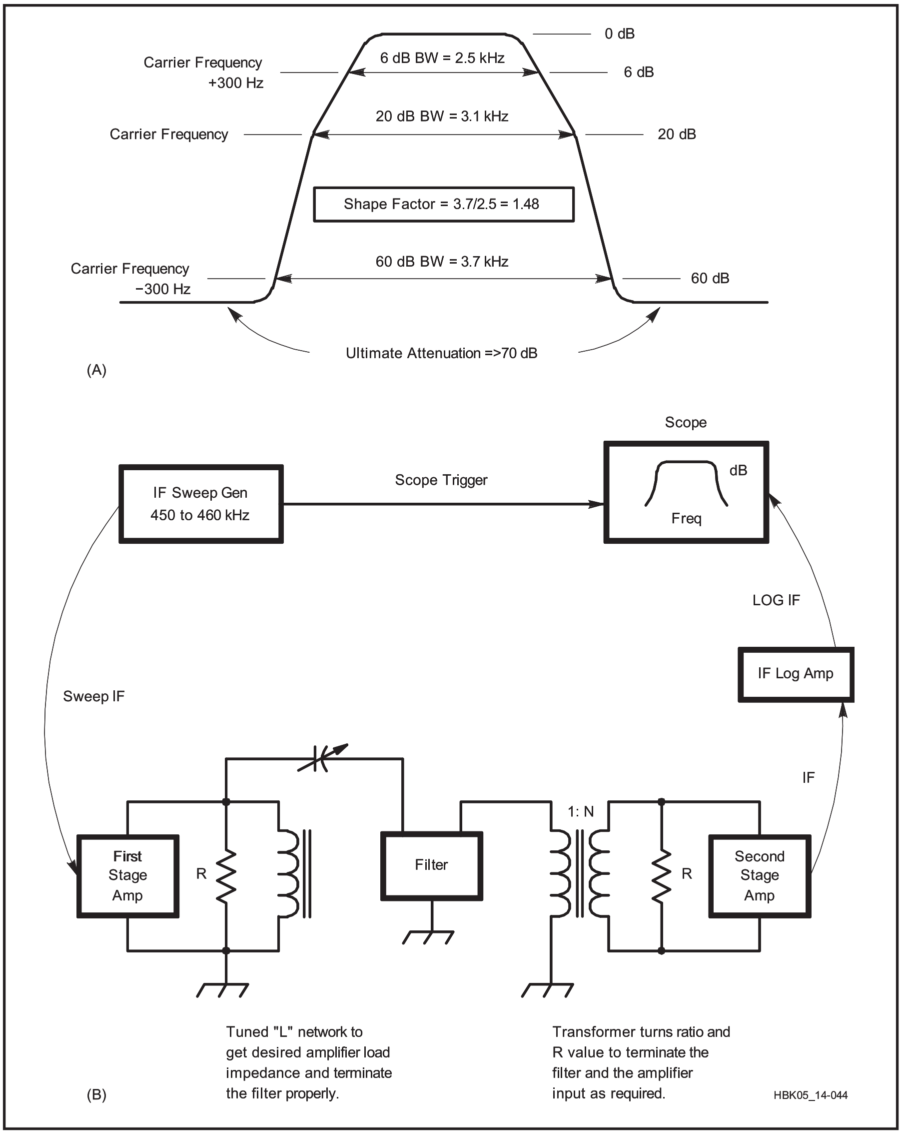


Figure 13.8 — At (A), desired response of a SSB IF filter. At (B), one method of terminating a mechanical filter that allows easy and accurate tuning adjustment and also a possible test setup for performing the adjustments.

• The stage after the filter must see the desired source (generator) impedance.

• The filter must be properly terminated at both ends.

Lack of any of these conditions will result in loss of specified filter response. Figure 13.8B shows two typical approaches. This kind of setup is a very good way to make sure the filters and other circuitry are working properly.

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

IF Linearity and Noise

Figure 13.9 indicates that after the last SSB filter, whether it is just after the SSB modulator or after the IF clipper, subsequent BPFs are considerably wider. For example, the 70 MHz crystal filter may be 15 to 30 kHz wide. This means that there is a “gray region” in the transmitter in which out-of-band IMD that is generated in the IF amplifiers and mixers can cause adjacent-channel interference.

A possible exception, not shown in Figure 13.9, is that there may be an intermediate IF in the 10 MHz region that also contains a narrow filter.

The implication is that special attention must be paid to the linearity of these circuits. It’s the designer’s job to make sure that distortion in this gray area is much less than distor-

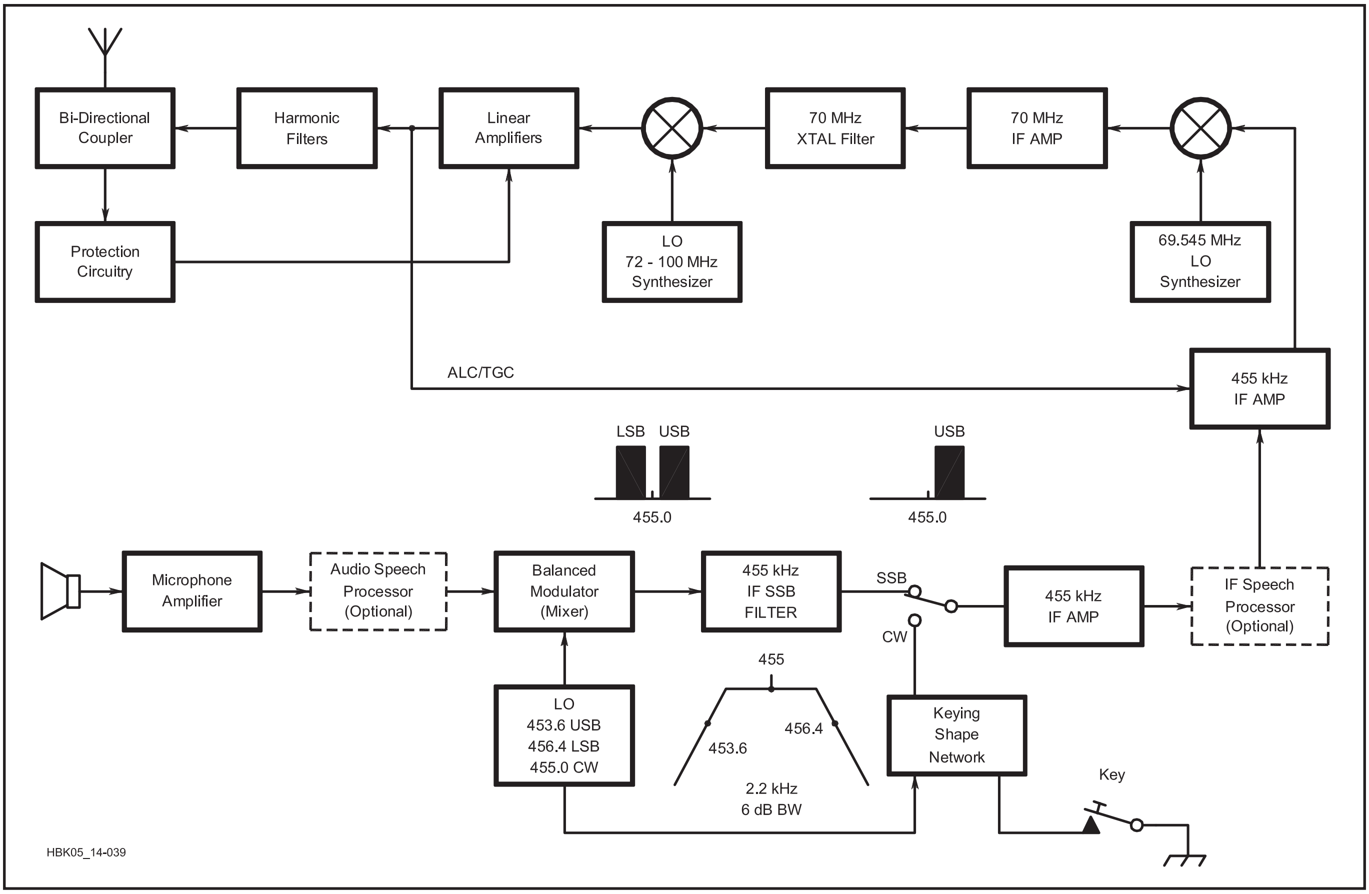


Figure 13.9 — Block diagram of an upconversion SSB/CW transmitter.

tion generated by the PA and also less than the phase noise generated by the final mixer. Recall also that the total IMD generated in the exciter stages is the result of several amplifier and mixer stages in cascade; therefore, each element in the chain must have at least 40 to 50 dB IMD quality. The various drive levels should be chosen to guarantee this. This requirement for multistage linearity is one of the main technical and cost burdens of the SSB mode.

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

TRANSMIT MIXER LINEARITY AND NOISE

The last IF and the last mixer LO in Figure 13.9 are selected so that, as much as possible, harmonic IMD products are far enough away from the operating frequency that they fall outside the passband of the low-pass filters and are highly attenuated. This is difficult to accomplish over the transmitter's entire frequency range. It helps to use a high-level mixer and a low enough signal level to minimize those products that are unavoidable. Low-order crossovers that cannot be sufficiently reduced are unacceptable, however; the designer must go back to the drawing board.

13.2.2 SDR Transmitter Architecture

SDR transmitter architecture looks a lot like the SDR receiver architecture as described in the **DSP and SDR Fundamentals** and **Receiving** chapters "turned around." For example, the FFT can be reused as its own inverse to translate back and forth between the time and frequency domains. I/Q modulation looks very much like I/Q demodulation. Substitute a DAC for the ADC to change digital to analog and vice versa. Digital filters just need a data stream and enough clock speed to handle the throughput.

The main question is one of providing enough computing resources and deciding where to make the jump from analog to digital. As the speed and resolution of data converters increases while the cost plummets, the transition between the analog and digital realms is moving ever closer to the antenna. Commercial mainstream transceiver designs being introduced in 2017 are no longer based on the traditional analog superheterodyne architecture. It is only a matter of time before the superheterodyne becomes a legacy technology.

Superheterodyne techniques are still used at points in the signal path, however. **Figure 13.10** shows the progression of transmitter architectures from all-superheterodyne through direct-sampled. If you think they look like the SDR receiver architectures of Figures 8.3 through 8.6 in the **DSP and SDR Fundamentals** chapter, you're right. They are essentially the same processes used for receiving but converting signals to RF instead of vice versa. (See the **Receiving** chapter for a discussion of the architecture pros and

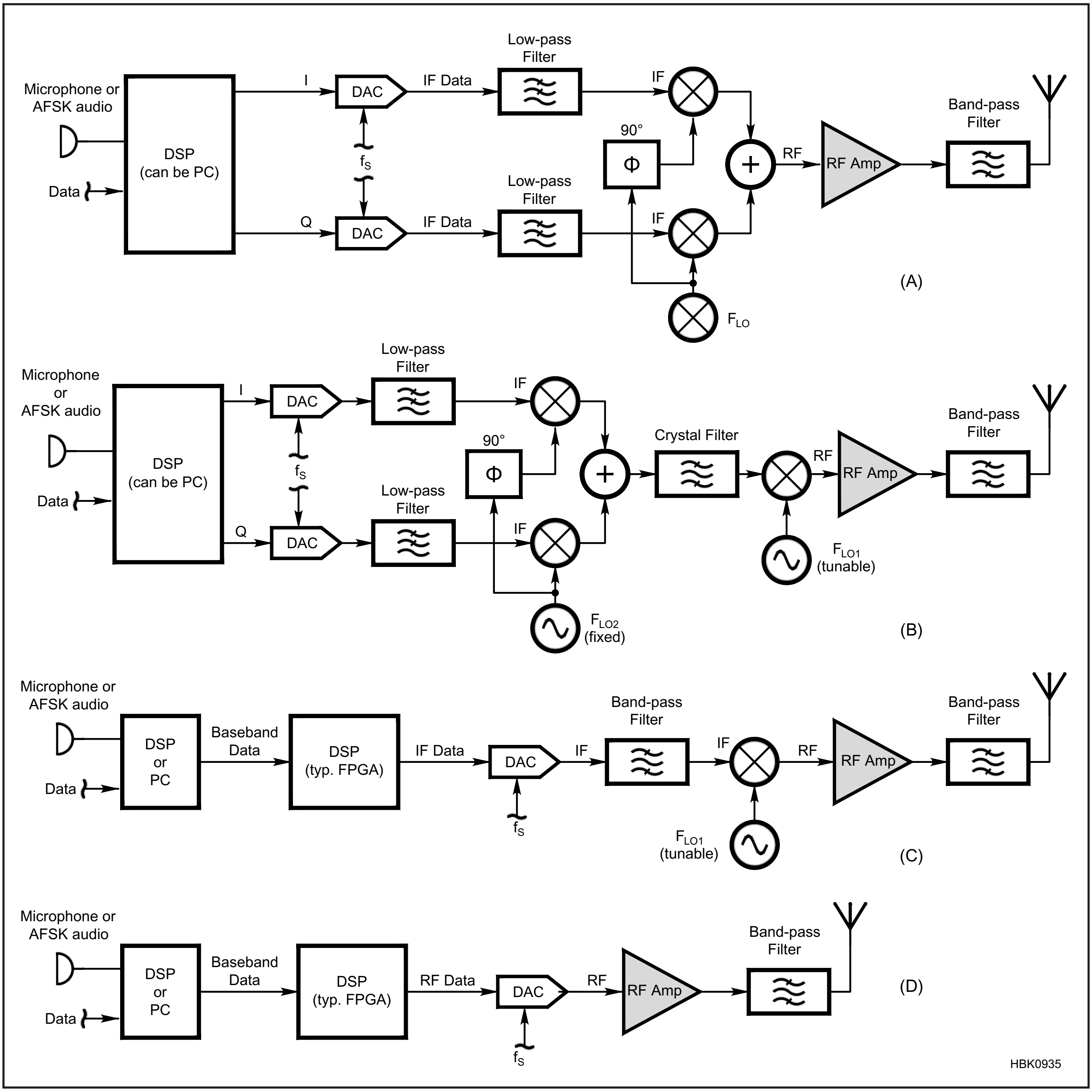


Figure 13.10 — Several SDR transmitter architectures. Direct-from-baseband (A), Hybrid superhet/SDR combination (B), Hybrid DSP-at-IF (C), Direct-sampled (B).

cons.) All of these architectures are in use by amateurs in homebuilt and commercial equipment today.

In any specific transceiver, the receiver and architectures are complementary in that they make use of the same DSP components and convert between analog and digital at roughly the same point in the signal path. The architecture is largely driven by the cost and availability of the data converters and the FPGA or similar DSP computing devices. (Some transceivers are based on generic FPGA parts while others make use of specialized DSP or graphics processors.) Tradeoffs involved with the various architectures are discussed in the **Transceiver Design Topics** chapter.

SDR technology is evolving rapidly so it is premature to make blanket statements about expected levels of performance. The best approach is to read the *QST* Product Reviews, review comparisons such as by Rob Sherwood, NCØB (**www.sherweng.com/table.html**) and Jim Brown, K9YC (**k9yc.com/publish.html**), ask experienced operators for their opinions, and then try the equipment for yourself!

13.3 Modulators

Previous editions of the *Handbook* covered modulators in a general treatment of mixers. Much of that discussion is now included in the **Receiving** chapter where you can find more detailed information on mixer operation.

Mixer Math: Amplitude Modulation

We can easily allow the carrier to be part of the mixer along with the sidebands merely by adding enough *dc level shift* into the information we want to mix so that its waveform never goes negative. In the sidebar "Mixer Math: Mixing as Multiplication" in the **Receiving** chapter, mixer math was kept relatively simple by setting the peak voltage of the input signals directly equal to their sine values. Each input signal's peak voltage therefore varies between +1 and –1, so all we need to do to keep our modulating- signal term (provided with a subscript *m* to reflect its role as the modulating or information waveform) from going negative is add 1 to it. Identifying the carrier term with a subscript c, we can write

$$\text{AM signal} = (1 + m \sin 2\pi f_m t) \sin 2\pi f_c t \qquad \text{(A)}$$

Notice that the modulation ($2\pi f_m t$) term has company in the form of a coefficient, *m*. This variable expresses the modulating signal's varying amplitude — variations that ultimately result in amplitude modulation. Expanding the equation gives us:

$$\text{AM signal} = \sin 2\pi f_c t + \frac{1}{2} m \cos (2\pi f_c - 2\pi f_m)t - \frac{1}{2} m \cos (2\pi f_c - 2\pi f_m)t \qquad \text{(B)}$$

The modulator's output now includes the carrier ($\sin 2\pi f_c t$) in addition to sum and difference products that vary in strength according to m. According to the conventions of talking about modulation, we call the sum product, which comes out at a frequency higher than that of the carrier, the *upper sideband (USB)*, and the difference product, which comes out a frequency lower than that of the carrier, the *lower sideband (LSB)*.

Why We Call It Amplitude Modulation

This process is called *amplitude modulation* because the complex waveform consisting of the sum of the sidebands and carrier varies with the information signal's magnitude (m). Concepts long used to illustrate AM's mechanism may mislead us into thinking that the *carrier* varies in strength with modulation, but careful study of the equation above shows that this doesn't happen. The carrier, $\sin 2\pi f_c t$, goes into the modulator as a sinusoid with an unvarying maximum value of |1|. The modulator multiplies the carrier by the dc level (+1) that we added to the information signal ($m \sin 2\pi f_m t$). Multiplying $\sin 2\pi f_c t$ by 1 merely returns $\sin 2\pi f_c t$. Thus, the carrier's amplitude does not vary as a result of amplitude modulation.

13.3.1 Amplitude Modulators

You can see how an AM signal is constructed as illustrated in **Figure 13.11**. Figure 13.11A shows the carrier, and the sidebands from a modulating tone are shown in 13.11B and 13.11C. The waveform of an AM signal appears to vary the carrier amplitude but this is not the case. The varying envelope of the AM signal results from the signal's three components — the carrier and the two sidebands — adding together. As the components reinforce and cancel each other, their sum (which appears as the envelope) rises and falls.

If you look closely, you can see that the waveforms in Figures 13.11B and 13.11C have slightly different frequencies than the carrier. If the two sidebands are added together, the signal of Figure 13.11D is produced. This is what the two sidebands look like as waveforms without the carrier. This is a *double-sideband, suppressed carrier* (DSBSC).

When the carrier signal is added, the full AM signal is produced in Figure 13.11D. When all of the signals are in-phase, the resulting signal has its maximum amplitude. When all of the signals are out of phase, the resulting signal goes to zero. If the carrier's phase is used as our reference, the phase of each sideband can be viewed as slipping behind (lower sideband) or moving ahead (upper sideband) of the carrier. The sidebands are out of phase with each other at the frequency of the tone so the resulting envelope reproduces the modulating tone's sine wave.

In the transmitter, the SSB modulator must suppress both the carrier and the unwanted sideband. Carrier suppression is normally accomplished with a *balanced modulator*, a

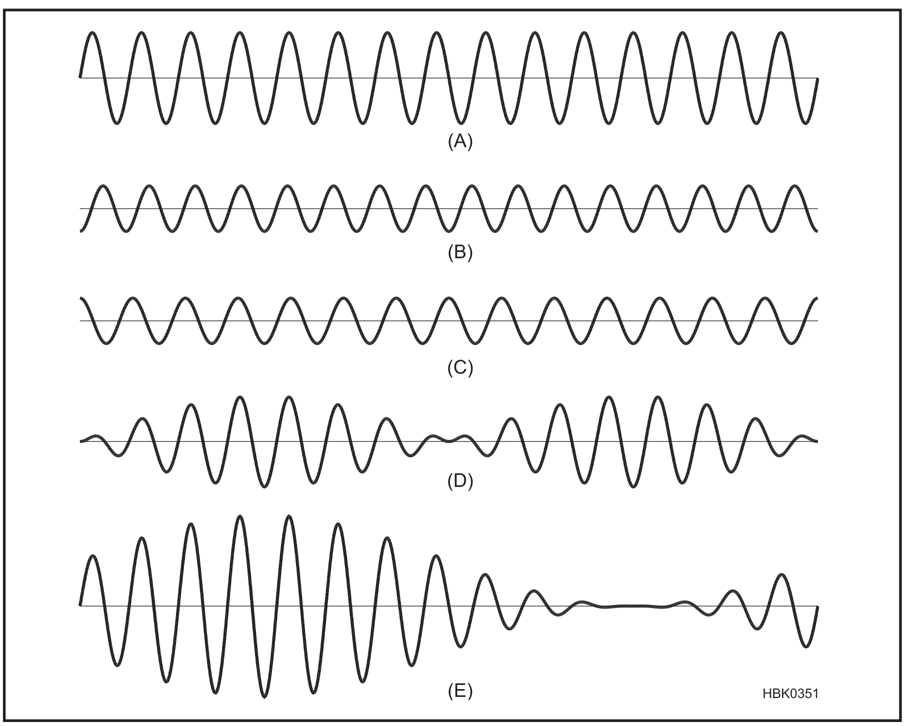


Figure 13.11 — At A is an unmodulated carrier. If the upper (B) and lower (C) sidebands are added together a double-sideband suppressed carrier (DSBSC) signal results (D). If each sideband has half the amplitude of the carrier, then the combination of the carrier with the two sidebands results in a 100%-modulated AM signal (E). Whenever the two sidebands are out of phase with the carrier, the three signals sum to zero. Whenever the two sidebands are in phase with the carrier, the resulting signal has twice the amplitude of the unmodulated carrier.

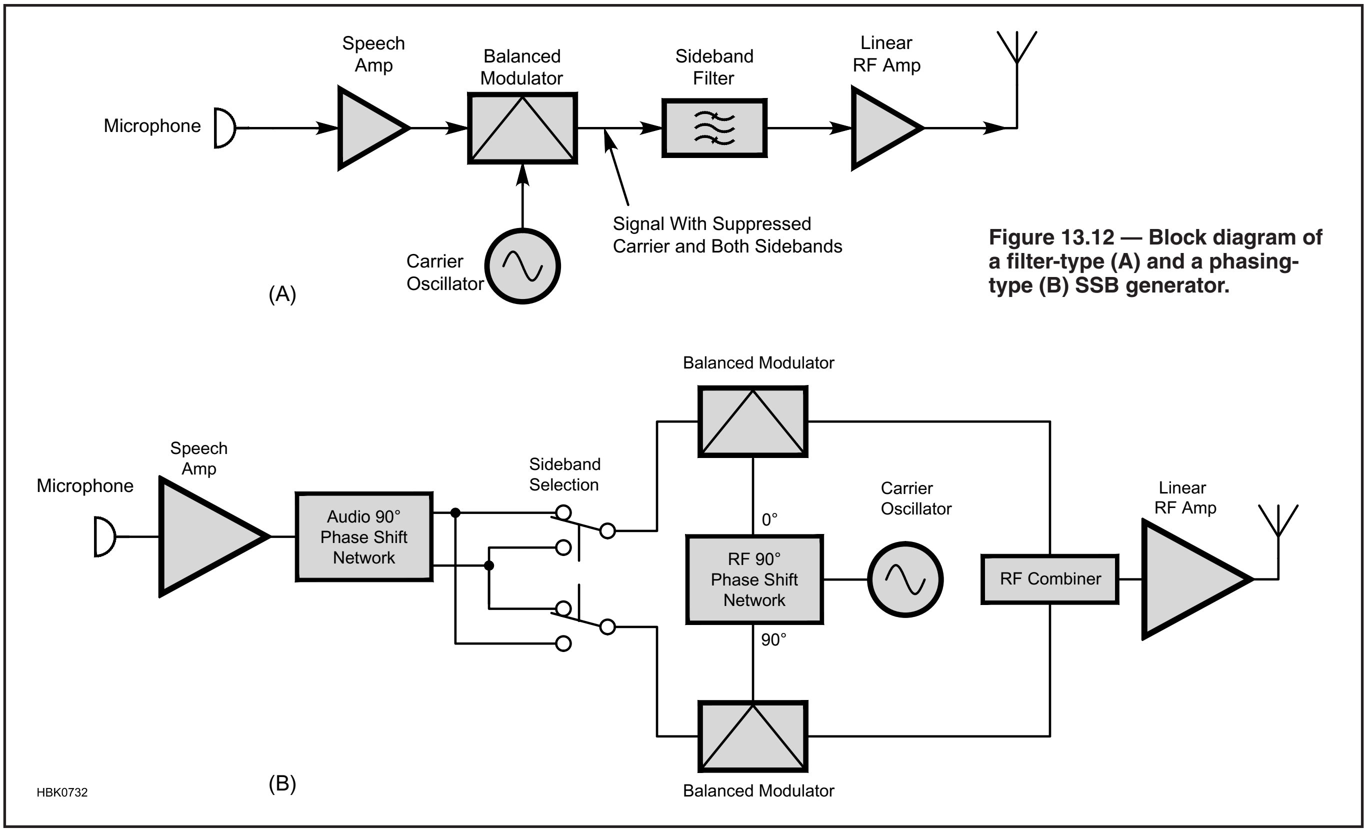


Figure 13.12 — Block diagram of a filter-type (A) and a phasing-type (B) SSB generator.

type of mixer whose output contains the sum and difference frequencies of the two input signals (the modulating signal and the carrier) but not the input signals themselves. There are several ways to eliminate the unwanted sideband, but the most common, shown in **Figure 13.12A**, is to pass the output of the balanced mixer through a crystal filter that passes the wanted sideband while filtering out the unwanted one. This is convenient in a transceiver since the same filter can be used in the receiver by means of a transmit-receive switch.

Another method to generate single sideband is called the *phasing method*. See Figure 13.12B. Using trigonometry, it can be shown mathematically that the sum of the signals from two balanced modulators, each fed with audio signals and RF carriers that are 90° out of phase, consists of one sideband only. The other sideband is suppressed. The output can be switched between LSB and USB simply by reversing the polarity of one of the inputs, which changes the phase by 180°. The phasing method eliminates the need for an expensive crystal filter following the modulator or, in the case of a transceiver, the need to switch the crystal filter between the transmitter and receiver sections. In addition, the audio quality is generally better because it eliminates the poor phase dispersion that is characteristic of most crystal filters. Using analog techniques, designing and building an audio phase-shift network that accurately maintained a 90° differential over a decade-wide frequency band (300 Hz to 3000 Hz) was rather complicated. With modern DSP techniques, the task is much easier.

BALANCED MODULATORS

A balanced modulator is a mixer. Briefly, the IF frequency LO (455 kHz in the example of Figure 13.9) translates the audio frequencies up to a pair of IF frequencies — the LO plus the audio frequency and the LO minus the audio frequency. The balance from the LO port to the IF output causes the LO frequency to be suppressed by 30 to 40 dB. Adjustments are provided to improve the LO null.

The filter method of SSB generation uses an IF band-pass filter to pass one of the sidebands and block the other. In Figure 13.9 the filter is centered at 455.0 kHz. The LO is offset to 453.6 kHz or 456.4 kHz so that the upper sideband or the lower sideband (respectively) can pass through the filter. This creates a problem for the other LOs in the radio, because they must now be properly offset so that the final transmit output's carrier (suppressed) frequency coincides with the frequency readout on the front panel of the radio.

Various schemes have been used to create the necessary LO offsets. One method uses two crystals for the 69.545 MHz LO that can be selected. In synthesized radios, the programming of the microprocessor controls the various LOs. Some synthesized radios use two IF filters at two different frequencies, one for USB and one for LSB, and a 455.0 kHz LO, as shown in Figure 13.9. These radios can be designed to transmit two independent sidebands (ISB) resulting in two separate channels in the spectrum space of the usual AM channel.

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

AMPLITUDE MODULATION WITH A DBM

We can generate DSB, suppressed-carrier AM with a DBM (double balanced mixer) by feeding the carrier to its RF port and the modulating signal to the IF port. (See the **Receiving** chapter for a detailed discussion of the DBM.) This is a classical *balanced modulator*, and the result — sidebands at radio frequencies corresponding to the carrier signal plus audio and the RF signal minus audio — emerges from the DBM's LO port. If we also want to transmit some carrier along with the sidebands, we can dc-bias the IF port (with a current of 10 to 20 mA) to upset the mixer's balance and keep its diodes from turning all the way off. (This technique is sometimes used for generating CW with a balanced modulator otherwise intended to generate DSB as part of an SSB-generation process.) **Figure 13.13** shows a more elegant approach to generating full-carrier AM with a DBM.

As we saw earlier when considering the many faces of AM, two DBMs, used in conjunction with carrier and audio phasing, can be used to generate SSB, suppressed-carrier AM. Likewise, two DBMs can be used with RF and LO phasing as an image-reject mixer.

AN MC1496P BALANCED MODULATOR

Although it predates the SA602/612, Freescale's MC1496 Gilbert cell multiplier remains a viable option for product detection and balanced modulator service. It has been around for decades and is still one of the best and least expensive. **Figure 13.14** is a typical balanced modulator circuit using the MC1496. The circuit of **Figure 13.15** includes an MC1496-based balanced modulator that is capable of carrier suppression greater than 50 dB. Per its description in Hayward, Campbell, and Larkin's *Experimental Methods in RF Design*, its output with audio drive should be kept to about –20 dBm with this circuit. LO drive should be 200 to 500 mV P-P.

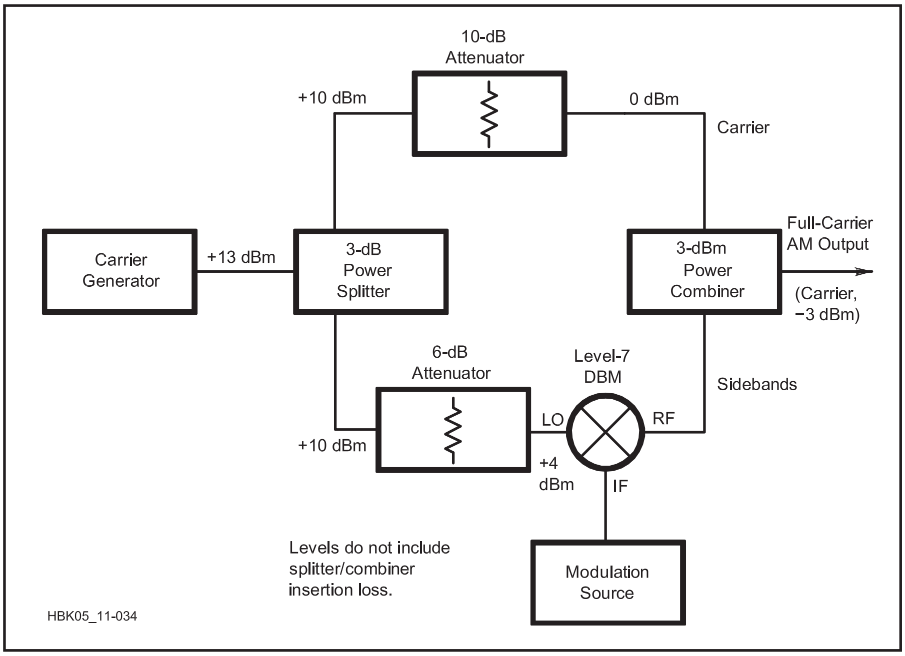


Figure 13.13 — Generating full-carrier AM with a diode DBM. A practical modulator using this technique is described in *Experimental Methods in RF Design*.

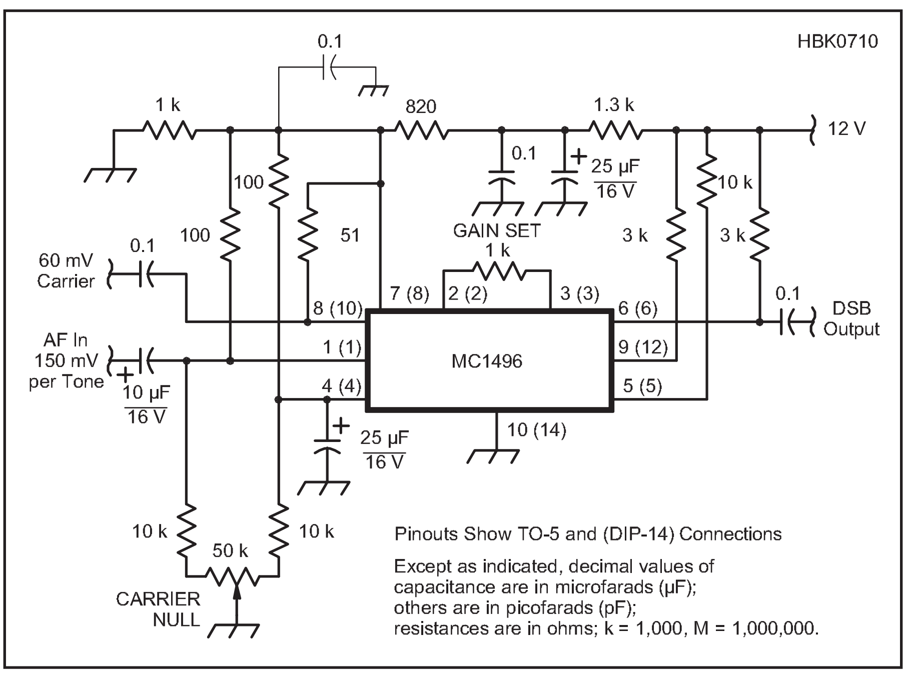


Figure 13.14 — An IC balanced modulator circuit using the MC1496 IC. The resistor between pins 2 and 3 sets the subsystem gain.

SDR SSB GENERATORS

(This section is taken from the SDR: Simplified column by Ray Mack, W5IFS in the September/October 2012 issue of *QEX*.) The structure of this SDR SSB generator software is the same as if it were implemented in analog hardware. **Figure 13.16** shows the block diagram of the system. The program operates in a serial fashion: first an audio baseband filter limits the audio to a band of 300 Hz to 3 kHz. Second, the DDS phase step value to determine the carrier frequency is computed. Following that is the multiplication for the balanced mixer. Finally the undesired sideband is removed. Once the single sideband signal has been created, up-conversion is used to translate it to RF.

The audio band-pass filter response is only useful with 200 taps or more (See **Figure 13.17**). At 100 taps the rejection is only on the order of 12 dB below 100 Hz. Likewise, the opposite sideband filter requires on the order of 700 to 1000 taps to give approximately 60 dB of opposite sideband suppression. The large number of taps also makes the skirts very steep, so that we can use the filter to also further reduce any carrier feed-through. **Figure 13.18** illustrates how steep the skirts can be. If the low-frequency carrier is at 18 kHz, carrier suppression is approximately 52 dB and the unwanted sideband more than that.

One alternative to reduce taps required in the audio filter is to simply use a dc block in the analog portion of the audio chain to set a lower boundary on the frequency. The response will be zero at 0 Hz and rise very rapidly to the frequency we set. This reduces the need for a sharp cutoff in DSP.

The close-in rejection of audio above 3 kHz is 45 dB or more with the 200-tap filter. Additionally, there is almost no energy above 3 kHz in the human voice, so energy in that region will likely be at least 60 dB below the lower frequencies after filtering.

Limiting the higher frequencies allows the use of a 6 kHz wide sideband selection filter instead of the normal 3 kHz filter to get better skirt response. A low-pass or high-pass filter would also work and give approximately the

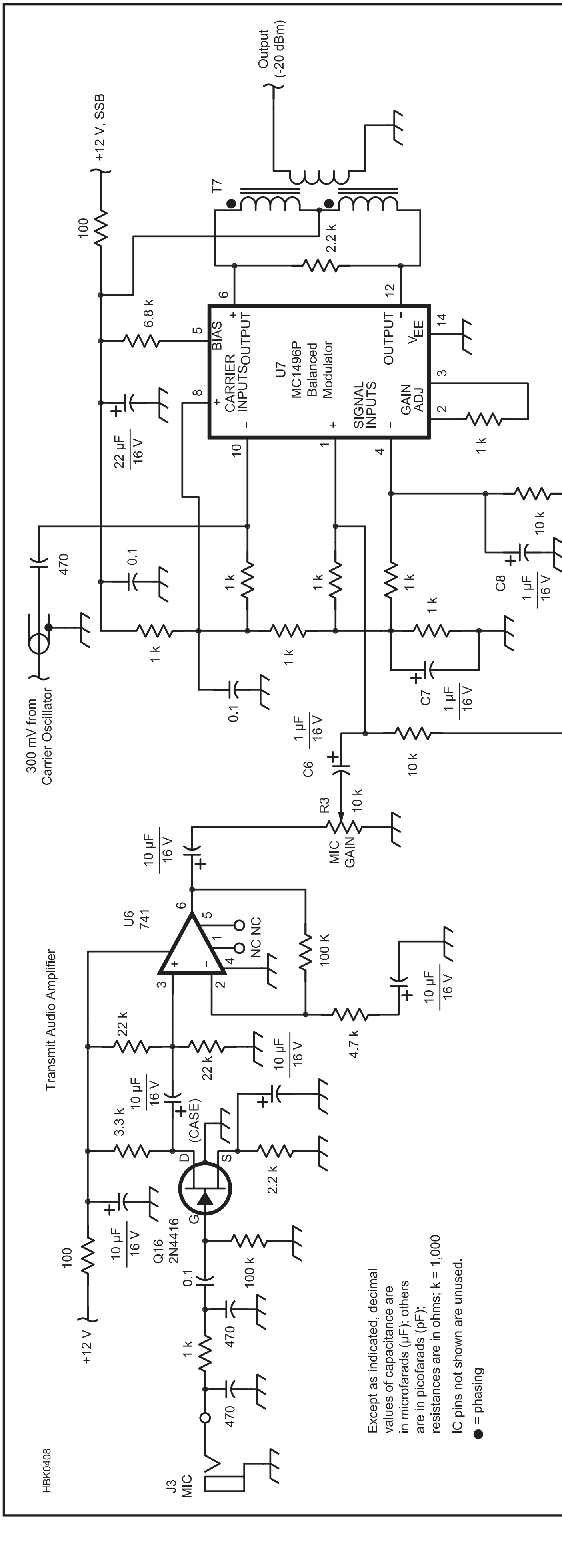


Figure 13.15 — Speech amplifier and 9-MHz balanced modulator using an MC-1496P analog multiplier IC. The transformer consists of 10 bifilar turns of #28 AWG enameled wire on an FT37-43 ferrite toroidal core with a 3-turn output link. The pin numbers shown are those of the DIP version of the 1496; builders using other package variants should consult manufacturer data to obtain the correct pinout.

same skirt response, but we want to be sure to eliminate any residual energy at baseband in the case of a lower sideband transmission. The wider bandwidth limits the lower frequency for our carrier. We want the carrier frequency to be as high as possible in order to limit image response when we up-convert to our final RF signal.

There is a practical limit with respect to the number of taps in the filters. Each tap requires one multiply-accumulate operation, which is a MAC in the DSP world. (MAC also means Media Access Control to a networking hardware person!) The DSP is capable of one MAC for each MHz of clock frequency for each portion of the hardware chain. If voice is digitized at a 48 kHz sample rate, a transmitter filter with 200 taps for audio and 1000 taps for sideband selection will need 57.6 MMACs to do its job. (See the **Receiving** chapter note from KA9Q regarding fast convolution to implement the filter method as a more efficient method if the hardware can support it.) In addition, a large number of taps can create latency. In this case, for a 48 kHz sample stream, 1000 taps represents 20 msec. Depending on how the SDR is implemented, this could add to overall microphone-to-RF latency.

OVERMODULATION

Since the information we transmit using AM shows up entirely as energy in its sidebands, it follows that the more energetic we make the sidebands, the more information energy will be available for an AM receiver to "recover" when it demodulates the signal. Even in an ideal modulator, there's a practical limit to how strong we can make an AM signal's sidebands relative to its carrier, however. Beyond that limit, we severely distort the waveform we want to translate into radio form.

We reach AM's distortion-free modulation limit when the sum of the sidebands and carrier at the modulator output *just reaches zero* at the modulating wave-form's most negative peak (**Figure 13.19**). We call this condition *100% modulation*, and it occurs when m in equation A in the sidebar "Mixer Math: Amplitude Modulation" equals 1. (We enumerate *modulation percentage* in values from 0 to 100%. The lower the number, the less information energy is in the sidebands. You may also see modulation enumerated in terms of a *modulation factor* from 0 to 1, which directly equals m; a modulation factor of 1 is the same as 100% modulation.) Equation B in the sidebar shows that each sideband's voltage is half that of the carrier. Power varies as the square of voltage, so the power in each sideband of a 100%-modulated signal is therefore $(½)^2$ times, or ¼, that of the carrier. A transmitter capable of 100% modulation when operating at a carrier power of 100 W

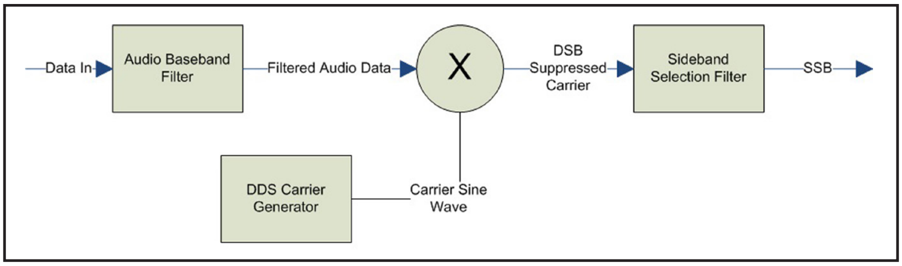


Figure 13.16 — This block diagram shows the software SSB transmit generator using the filter method.

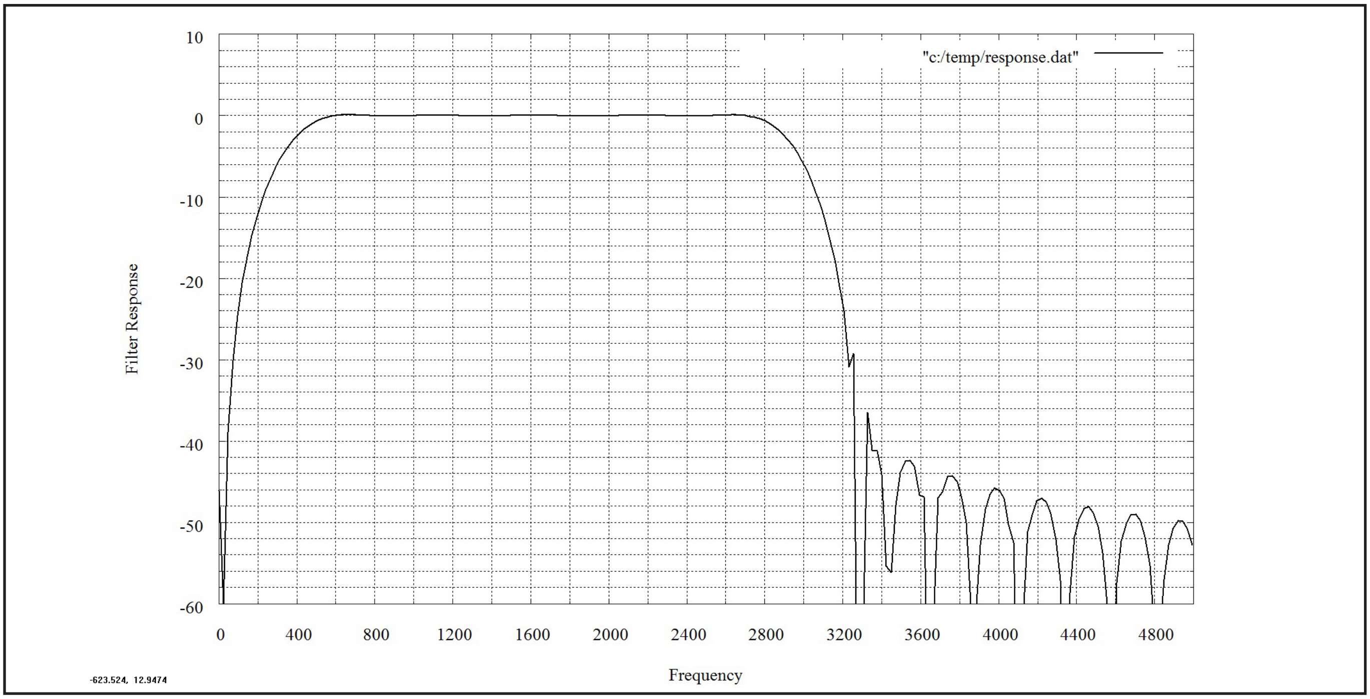


Figure 13.17 — The 200-tap baseband filter response.

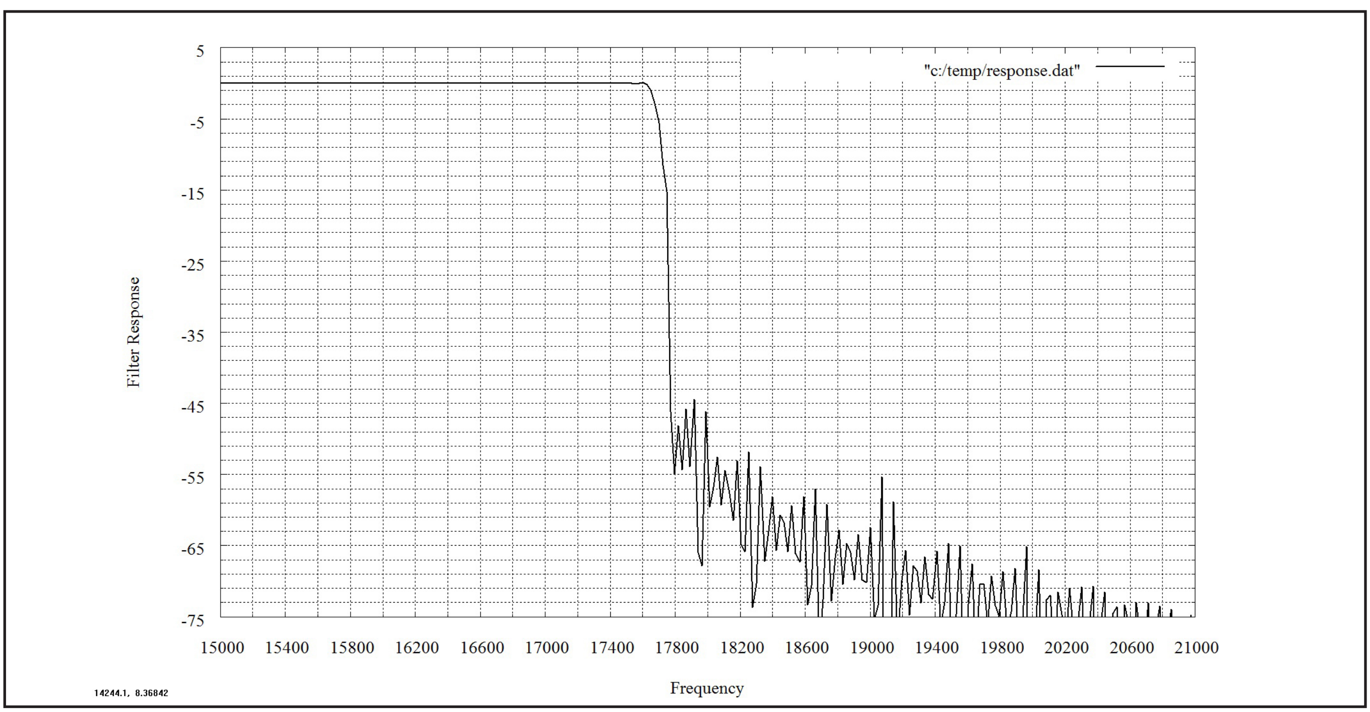


Figure 13.18 — The response of a 700-tap filter, showing a wider frequency view. The filter is 6 kHz wide to allow for a steep skirt on the carrier side. The 6 dB cutoff point is set at 300 Hz away from the low-frequency carrier, which is at 18,000 Hz (see text).

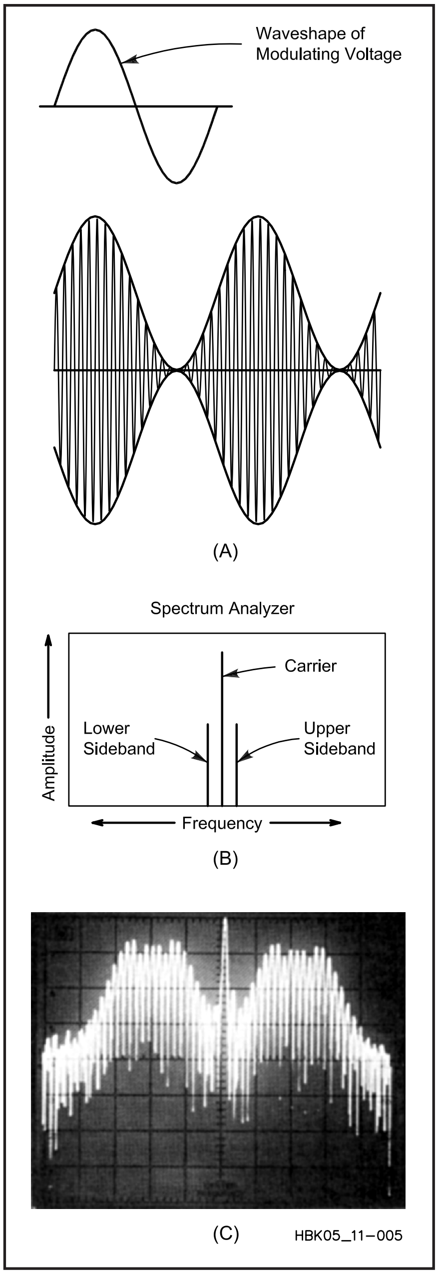


Figure 13.19 — Graphed in terms of amplitude versus time (A), the *envelope* of a properly modulated AM signal exactly mirrors the shape of its modulating waveform, which is a sine wave in this example. This AM signal is modulated as fully as it can be — 100% — because its envelope *just* touches zero on the modulating wave's negative peaks. Graphing the same AM signal in terms of amplitude versus frequency (B) reveals its three spectral components: Carrier, upper sideband and lower sideband. B shows sidebands as single-frequency components because the modulating waveform is a sine wave. With a complex modulating waveform, the modulator's sum and difference products really do show up as bands on either side of the carrier (C).

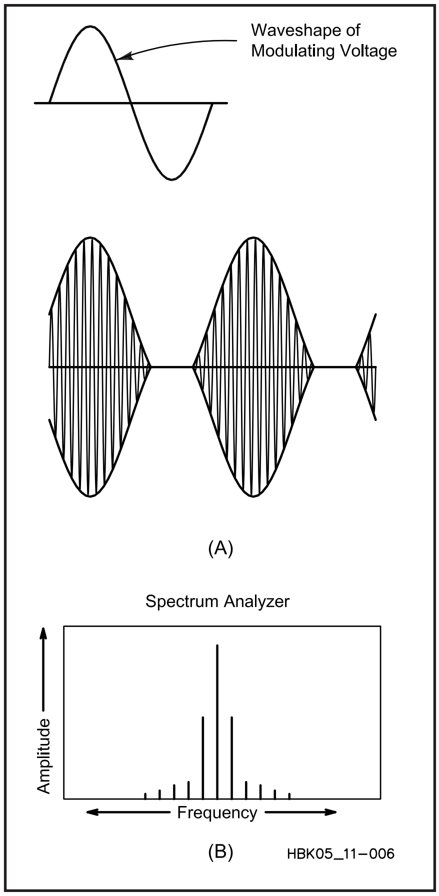


Figure 13.20 — Negative-going overmodulation of an AM transmitter results in a modulation envelope (A) that doesn't faithfully mirror the modulating waveform. This distortion creates additional sideband components that broaden the transmitted signal (B). Positive-going modulation beyond 100% is used by some AM broadcasters in conjunction with negative-peak limiting to increase "talk power" without causing negative overmodulation.

therefore puts out a 150-W signal at 100% modulation, 50 W of which is attributable to the sidebands. (The *peak* envelope power [PEP] output of a double-sideband, full-carrier AM transmitter at 100% modulation is four times its carrier PEP. This is why our solid-state, "100-W" MF/HF transceivers are usually rated for no more than about 25 W carrier output at 100% amplitude modulation.)

One-hundred-percent negative modulation is a brick-wall limit because an amplitude modulator can't reduce its output to less than zero. Trying to increase negative modulation beyond the 100% point results in *overmodulation* (**Figure 13.20**), in which the modulation envelope no longer mirrors the shape of the modulating wave (Figure 13.20A). A negatively overmodulated wave contains more energy than it did at 100% modulation, but some of the added energy now exists as *harmonics of the modulating waveform* (Figure 13.20B). This distortion makes the modulated signal take up more spectrum space than it needs. In voice operation, overmodulation commonly happens only on syllabic peaks, making the distortion products sound like transient noise we refer to as *splatter*.

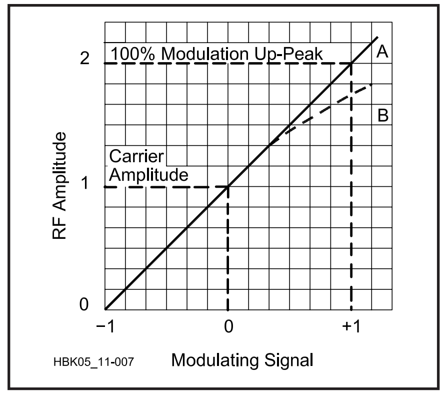


Figure 13.21 — An ideal AM transmitter exhibits a straight-line relationship (A) between its instantaneous envelope amplitude and the instantaneous amplitude of its modulating signal. Distortion, and thus an unnecessarily wide signal, results if the transmitter cannot respond linearly across the modulating signal's full amplitude range (B).

MODULATION LINEARITY

If we increase an amplitude modulator's modulating-signal input by a given percentage, we expect a proportional modulation increase in the modulated signal. We expect good *modulation linearity*. Suboptimal amplitude modulator design may not allow this, however. Above some modulation percentage, a modulator may fail to increase modulation in proportion to an increase in its input signal (**Figure 13.21**). Distortion, and thus an unnecessarily wide signal, results.

13.3.2 Angle Modulators

Amplitude modulation served as our first means of translating information into radio form because it could be implemented as simply as turning an electric noise generator on and off. (A spark transmitter consisted of little more than this.) By the 1930s, we had begun experimenting with translating information into radio form and back again by modulating a radio wave's angular velocity (frequency or phase) instead of its overall amplitude. The result of this process is *frequency modulation (FM)* or *phase modulation (PM)*, both of which are often grouped under the name *angle modulation* because of their underlying principle.

A change in a carrier's frequency or phase

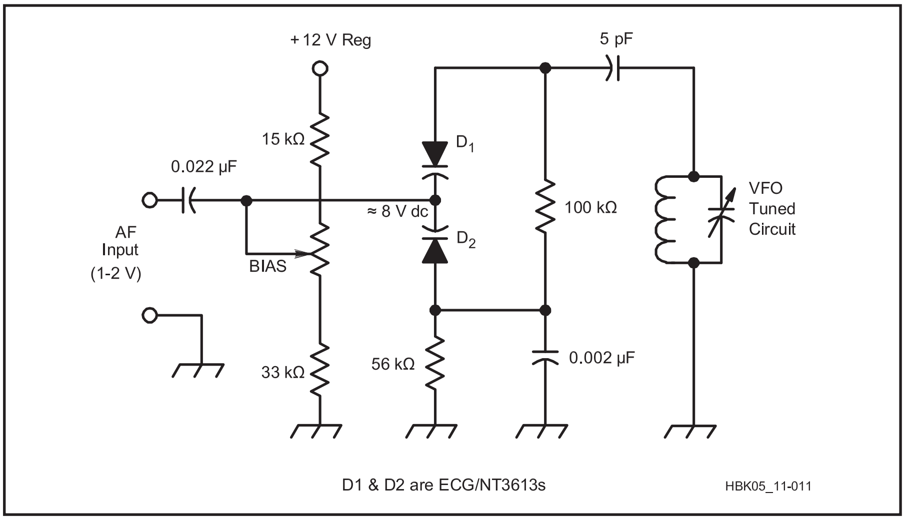


Figure 13.22 — One or more tuning diodes can serve as the variable reactance in a reactance modulator. This HF reactance modulator circuit uses two diodes in series to ensure that the tuned circuit's RF-voltage swing cannot bias the diodes into conduction. D1 and D2 are "30-volt" tuning diodes that exhibit a capacitance of 22 pF at a bias voltage of 4. The BIAS control sets the point on the diode's voltage-versus-capacitance characteristic around which the modulating waveform swings.

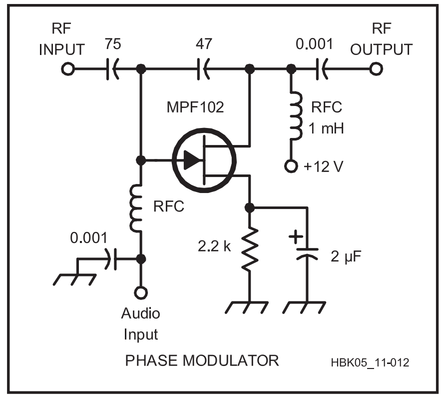


Figure 13.23 — A series reactance modulator acts as a variable shunt around a reactance — in this case, a 47-pF capacitor — through which the carrier passes.

for the purpose of modulation is called *deviation*. An FM signal deviates according to the amplitude of its modulating waveform, independently of the modulating waveform's frequency; the higher the modulating wave's amplitude, the greater the deviation. A PM signal deviates according to the amplitude *and frequency* of its modulating waveform; the higher the modulating wave's amplitude *and/or frequency*, the greater the deviation. See the sidebar, "Mixer Math: Angle Modulation" for a numerical description of these processes.

If you vary a reactance in or associated with an oscillator's frequency-determining element(s), you vary the oscillator's frequency. If you vary the tuning of a tuned circuit through which a signal passes, you vary the signal's phase. A circuit that does this is called a *reactance modulator*, and can be little more than a tuning diode or two connected to a tuned circuit in an oscillator or amplifier (**Figure 13.22**). Varying a reactance through which the signal passes (**Figure 13.23**) is another way of doing the same thing.

The difference between FM and PM depends solely on how, and not how much, deviation occurs. A modulator that causes deviation in proportion to the modulating wave's amplitude and frequency is a phase modulator. A modulator that causes deviation only in proportion to the modulating signal's amplitude is a frequency modulator.

ANGLE MODULATION SIDEBANDS

Although angle modulation produces uncountable sum and difference products, most of them are vanishingly weak in practical systems. They emerge from the modulator spaced from the average ("resting," unmodulated) carrier frequency by integer multiples of the modulating frequency (**Figure 13.24**). The strength of the sidebands relative to the carrier, and the strength and phase of the carrier itself, vary with the degree of modulation — the modulation index. (The *overall* amplitude of an angle-modulated signal does not change with modulation, however; when energy goes out of the carrier, it shows up in the sidebands, and vice versa.) In practice, we operate angle-modulated transmitters at modulation indexes that make all but a few of their infinite sidebands small in amplitude. (A mathematical tool called *Bessel functions* helps determine the relative strength of the carrier and sidebands according to modulation index. The **Modulation** chapter includes a graph to illustrate this relationship.)

Mixer Math: Angle Modulation

An angle-modulated signal can be mathematically represented as

$$f_c(t) = \cos(2\pi f_c t + m \sin(2\pi f_m t))$$

$$= \cos(2\pi f_c t)\cos(m \sin(2\pi f_m t)) - \sin(2\pi f_c t)\sin(m \sin(2\pi f_m t))$$

In this equation, we see the carrier frequency ($2\pi f_c t$) and modulating signal ($\sin 2\pi f_m t$) as in equation A shown in the sidebar Mixer Math: Amplitude Modulation. We again see the modulating signal associated with a coefficient, m, which relates to degree of modulation. (In the AM equation, *m* is the modulation factor; in the angle-modulation equation, *m* is the *modulation index* and, for FM, equals the deviation divided by the modulating frequency.) We see that angle-modulation occurs as the cosine of the sum of the carrier frequency ($2\pi f_c t$) and the modulating signal ($\sin 2\pi f_m t$) times the modulation index (*m*). In its expanded form, we see the appearance of sidebands above and below the carrier frequency.

Angle modulation is a multiplicative process, so, like AM, it creates sidebands on both sides of the carrier. Unlike AM, however, angle modulation creates an *infinite* number of sidebands on either side of the carrier! This occurs as a direct result of modulating the carrier's angular velocity, to which its frequency and phase directly relate. If we continuously vary a wave's angular velocity according to another periodic wave's cyclical amplitude variations, the rate at which the modulated wave repeats *its* cycle — its frequency — passes through an infinite number of values. (How many individual amplitude points are there in one cycle of the modulating wave? An infinite number. How many corresponding discrete frequency or phase values does the corresponding angle-modulated wave pass through as the modulating signal completes a cycle? An infinite number!) In AM, the carrier frequency stays at one value, so AM produces two sidebands — the sum of its carrier's unchanging frequency value and the modulating frequency, and the difference between the carrier's unchanging frequency value and the modulating frequency. In angle modulation, the modulating wave shifts the frequency or phase of the carrier through an infinite number of different frequency or phase values, resulting in an infinite number of sum and difference products.

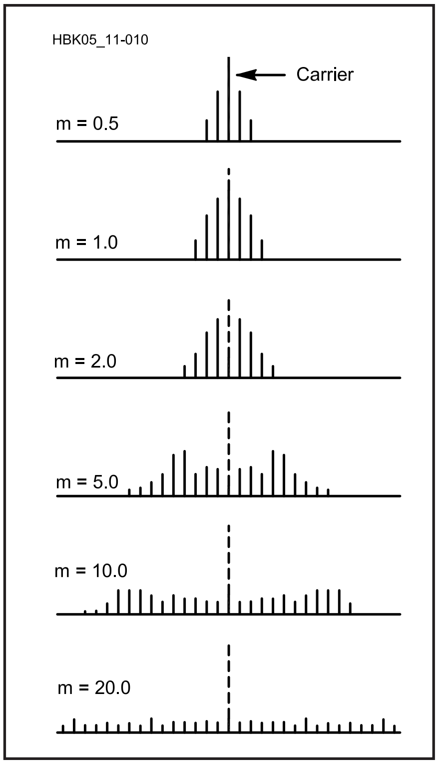


Figure 13.24 — Angle-modulation produces a carrier and an infinite number of upper and lower sidebands spaced from the average ("resting," unmodulated) carrier frequency by integer multiples of the modulating frequency. (This drawing is a simplification because it only shows relatively strong, close-in sideband pairs; space constraints prevent us from extending it to infinity.) The relative amplitudes of the sideband pairs and carrier vary with modulation index, *m*.

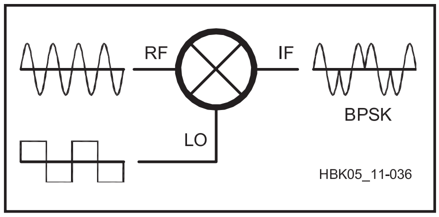


Figure 13.25 — Mixing a carrier with a square wave generates biphase-shift keying (BPSK), in which the carrier phase is shifted 180° for data transmission. In practice, as in this drawing, the carrier and data signals are phase-coherent so the mixer switches only at carrier zero crossings.

Selectivity in transmitter and receiver circuitry further modify this relationship, especially for sidebands far away from the carrier.

BIPHASE-SHIFT KEYING (BPSK) MODULATION WITH A DBM

Back in our discussion of square-wave mixing, we saw how multiplying a switching mixer's linear input with a square wave causes a 180° phase shift during the negative part of the square wave's cycle. As **Figure 13.25** shows, we can use this effect to produce *biphase-shift keying (BPSK)*, a digital system that conveys data by means of carrier phase reversals. A related system, *quadrature phase-shift keying (QPSK)* uses two DBMs and phasing to convey data by phase-shifting a carrier in 90° increments.

DEVIATION AND FREQUENCY MULTIPLICATION

Maintaining modulation linearity is just as important in angle modulation as it is in AM, because unwanted distortion is always our enemy. A given angle-modulator circuit can frequency- or phase-shift a carrier only so much before the shift stops occurring in strict proportion to the amplitude (or, in PM, the amplitude and frequency) of the modulating signal.

If we want more deviation than an angle modulator can linearly achieve, we can operate the modulator at a suitable sub-harmonic — submultiple — of the desired frequency, and process the modulated signal through a series of *frequency multipliers* to bring it up to the desired frequency. The deviation also increases by the overall multiplication factor, relieving the modulator of having to do it all directly. A given FM or PM radio design may achieve its final output frequency through a combination of mixing (frequency shift, no deviation change) and frequency multiplication (frequency shift *and* deviation change).

"TRUE FM"

Something we covered a bit earlier bears closer study: "An FM signal deviates according to the amplitude of its modulating waveform, independently of the modulating waveform's frequency; the higher the modulating wave's amplitude, the greater the deviation. A PM signal deviates according to the amplitude *and frequency* of its modulating waveform; the higher the modulating wave's amplitude *and/or frequency*, the greater the deviation."

The practical upshot of this excerpt is that we can use a phase modulator to generate FM. All we need to do is run a PM transmitter's modulating signal through a low-pass filter that (ideally) halves the signal's amplitude for each doubling of frequency (a reduction of "6 dB per octave," as we sometimes see such responses characterized) to compensate for its phase modulator's "more deviation with higher frequencies" characteristic. The result is an FM, not PM, signal. FM achieved with a phase modulator is sometimes called *indirect FM* as opposed to the *direct FM* we get from a frequency modulator.

We sometimes see claims that one piece of gear is better than another solely because it generates "true FM" as opposed to indirect FM. We can debunk such claims by keeping in mind that direct and indirect FM *sound exactly alike in a receiver* when performed correctly.

CONVEYING DC LEVELS WITH ANGLE MODULATION

Depending on the nature of the modulation source, there *is* a practical difference between a frequency modulator and a phase modulator. Answering two questions can tell us whether this difference matters: Does our modulating signal contain a dc level or not? If so, do we need to accurately preserve that dc level through our radio communication link for successful communication? If both answers are *yes*, we must choose our hardware and/or

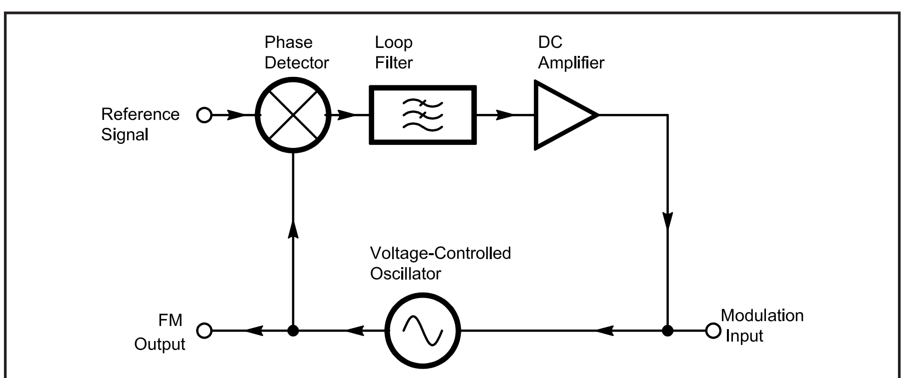


Figure 13.26 — Frequency modulation using a phase-locked loop (PLL).

information-encoding approach carefully, because a frequency modulator can convey dc-level shifts in its modulating waveform, while a phase modulator, which responds only to instantaneous changes in frequency and phase, cannot.

Consider what happens when we want to frequency-modulate a phase-locked-loop-synthesized transmitted signal. **Figure 13.26** shows the block diagram of a PLL frequency modulator. Normally, we modulate a PLL's VCO because it's the easy thing to do. As long as our modulating frequency results in frequency excursions too fast for the PLL to follow and correct — that is, as long as our modulating frequency is outside the PLL's *loop bandwidth* — we achieve the FM we seek. Trying to modulate a dc level by pushing the VCO to a particular frequency and holding it there fails, however, because a PLL's loop response includes dc. The loop, therefore, detects the modulation's dc component as a correctable error and "fixes" it. FMing a PLL's VCO therefore can't buy us the dc response "true FM" is supposed to allow.

We *can* dc-modulate a PLL modulator, but we must do so by modulating the frequency of the loop *reference*. The PLL then adjusts the VCO to adapt to the changed reference, and our dc level gets through. In this case, the modulating frequency must be *within* the loop bandwidth — which dc certainly is — or the VCO won't be corrected to track the shift.

SDR ANGLE MODULATORS

Figure 13.27 shows the block diagram of an FM transmitter using DSP to produce the carrier, and using addition to create true FM from the audio input. An FM signal can be generated directly through the DDS by simply adding or subtracting a small value that corresponds to the audio voltage to the tuning value used for the DDS accumulator. The frequency deviation is adjusted by controlling the gain applied to the audio signal. This method produces true frequency modulation. If the audio signal is used to control the phase accumulation of the DDS, the result is PM.

Angle modulation can also be produced using I/Q modulation as in **Figure 13.28**. A DSB-SC (double-sideband, suppressed-carrier) signal is produced by a balanced modulator or DSP multiplier (see SDR SSB Generators earlier in this chapter). The DSB-SC signal is then added to the carrier signal with a 90º phase difference. The result is a PM signal.

Frequency modulation requires that an integrator (low-pass filter) be applied to the modulating signal. This is because frequency is the time-derivative of phase. By applying the integrated signal to a phase modulator, FM is produced with a deviation that does not depend on signal amplitude — only frequency.

Because of the low-pass filter, speech is usually given a high-frequency boost by a high-pass *pre-emphasis* network. A corresponding *de-emphasis* (low-pass) network must be applied to the recovered modulation in the receiver to restore the original modulating signal's frequency response. This improves intelligibility and signal-to-noise ratio of the received audio.

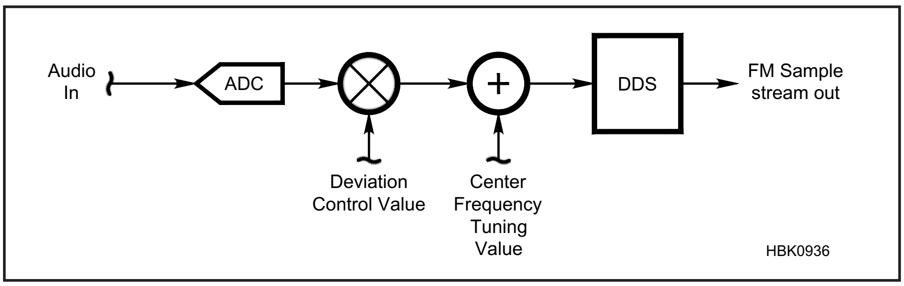


Figure 13.27 — Creating FM by controlling the frequency of a DDS signal source. By controlling the phase step instead of the frequency, PM would be created.

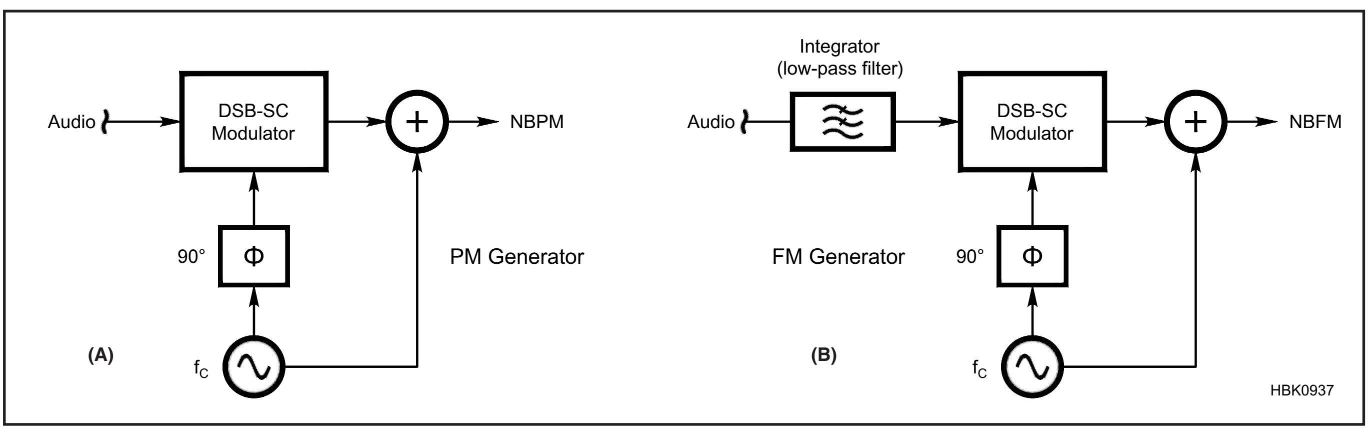


Figure 13.28 — Two methods of using I/Q modulation to produced angle-modulated signals. The block diagram includes a DSB generator (Figure 13.16) with the additional step of mixing (multiplying) the DSB signal with a phase-shifted carrier signal to produce PM (A) or FM (B). FM requires applies an integrator (low-pass filter) to the modulating signal so that output frequency depends only on the amplitude of the modulating signal and not its frequency.

13.4 Transmitting CW

Earlier in this chapter, the importance of shaping the time envelope of the keying pulse of an on-off keyed transmitter is discussed. There are serious ramifications of not paying close attention to this design parameter. The optimum shape of a transmitter envelope should approach the form of a sinusoid raised to a power with a tradeoff between occupied bandwidth and overlap between the successive pulses. This can be accomplished either through filtering of the pulse waveform before modulation in a linear transmitter, or through direct generation of the pulse shape using DSP.

The differences between well-designed and poor pulse shaping can perhaps be best described by looking at some results. The following figures are from recent *QST* product reviews of commercial multimode 100 W HF transceivers. **Figure 13.29** shows the CW keying waveform of a transmitter with good spectrum control. The top trace is the key closure, with the start of the first contact closure on the left edge at 60 WPM using full break-in. Below it is the nicely rounded RF envelope. **Figure 13.30** shows the resultant signal spectrum. Note that the signal amplitude is about 80 dB down at a spacing of ±1 kHz, with a floor of –90 dB over the 10 kHz shown. **Figures 13.31** and **13.32** are similar data taken from a different manufacturer's transceiver. Note the sharp corners of the RF envelope, as well as the time it takes for the first "dit" to be developed. The resulting spectrum is not even down 40 dB at ±1 kHz and shows a floor that doesn't quite make –60 dB over the 10 kHz range. It's easy to see the problems that the latter transmitter will cause to receivers trying to listen to a weak signal near its operating frequency. The unwanted components of the signal are heard on adjacent channels as sharp clicks when the signal is turned on and off, called *key clicks*. Note that even the best-shaped keying waveform in a linear transmitter will become sharp with a wide spectrum if it is used to drive a stage such as an external power amplifier beyond its linear range. This generally results in clipping or limiting with subsequent removal of the rounded corners on the envelope. Trying to get the last few dB of power out of a transmitter can often result in this sort of unintended signal impairment.

13.4.1 CW Operation

Figure 13.33A closely resembles what we see when a properly adjusted CW transmitter sends a string of dots. Keying a carrier on and off produces a wave that varies in amplitude and has double (upper and lower) sidebands that vary in spectral composition according

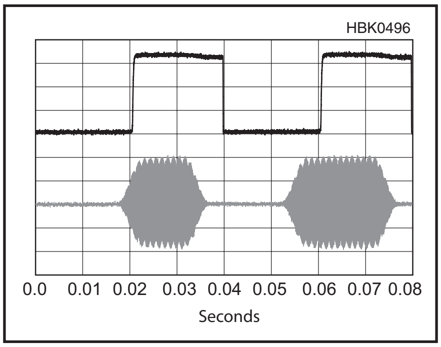


Figure 13.29 — The CW keying waveform of a transmitter with good spectrum control. The top trace is the key closure, with the start of the first contact closure on the left edge at 60 WPM using full break-in. Below that is the nicely rounded RF envelope.

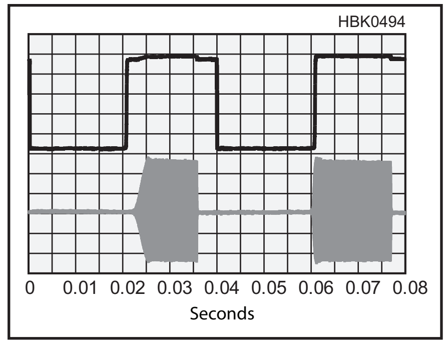


Figure 13.31 — The CW keying waveform of a transmitter with poor spectrum control. The top trace is the key closure, with the start of the first contact closure on the left edge at 60 WPM using full break-in. Note the sharp corners of the RF envelope that result in excessive bandwidth products.

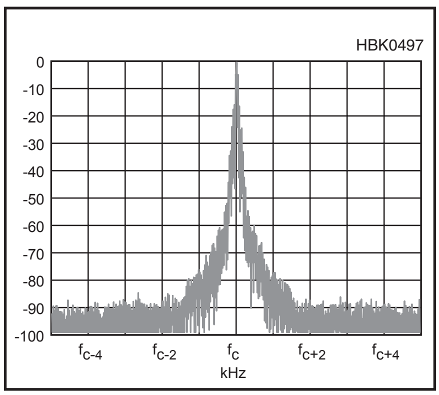


Figure 13.30 — The resultant signal spectrum from the keying shown in Figure 13.29. Note that the signal amplitude is about 80 dB down at a spacing of ±1 kHz, with a floor of –90 dB over the 10 kHz shown.

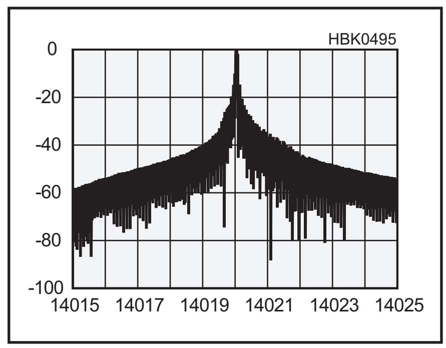


Figure 13.32 — The resultant signal spectrum from the keying shown in Figure 13.31. The resulting spectrum is not even down 40 dB at ±1 kHz and shows a floor that doesn't quite make 60 dB below the carrier across the 10 kHz.

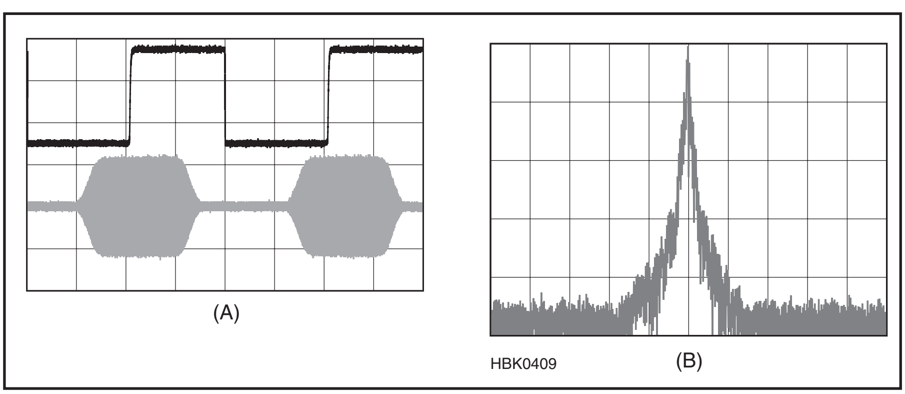


Figure 13.33 — Waveshaping in a CW transmitter often causes a CW signal's RF envelope (lower trace in the amplitude-versus-time display at A) to contain less harmonic energy than the abrupt transitions of its key closure waveform (upper trace in A) suggest should be the case. B, an amplitude-versus-frequency display, shows that even a properly shaped CW signal has many sideband components.

to the duration and envelope shape of the on-off transitions. The emission mode we call CW is therefore a form of AM. The concepts of modulation percentage and overmodulation are usually not applied to generating an on-off-keyed Morse signal, however. This is related to how we copy CW by ear, and the fact that, in CW radio communication, we usually don't translate the received signal all the way back into its original pre-modulator (*baseband*) form, as a closer look at the process reveals.

In CW transmission, we usually open and close a keying line to make dc transitions that turn the transmitted carrier on and off. See Figure 13.33B. CW reception usually does not entirely reverse this process, however. Instead of demodulating a CW signal all the way back to its baseband self — a shifting dc level — we want the presences and absences of its carrier to create long and short audio tones. Because the carrier is RF and not AF, we must mix it with a locally generated RF signal — from a *beat-frequency oscillator (BFO)* — that's close enough in frequency to produce a difference signal at AF (this BFO can, of course, also be inserted at an IF stage). What goes into our transmitter as shifting dc comes out of our receiver as tone bursts of dot and dash duration.

It so happens that we always need to hear one or more harmonics of the fundamental keying waveform for the code to sound sufficiently crisp. If the transmitted signal will be subject to fading caused by varying propagation — a safe assumption for any long-distance radio communication — we can *harden* our keying by making the transmitter's output rise and fall more quickly. This puts more energy into keying sidebands and makes the signal more copyable in the presence of fading — in particular, *selective fading*, which linearly distorts a modulated signal's complex waveform and randomly changes the sidebands' strength and phase relative to the carrier and each other. The appropriate keying hardness also depends on the keying speed. The faster the keying in WPM, the faster the on-off times — the harder the keying — must be for the signal to remain ear- and machine-readable through noise and fading.

Instead of thinking of this process in terms of modulation percentage, we just ensure that a CW transmitter produces sufficient keying-sideband energy for solid reception. Practical CW transmitters usually do not do their keying with a modulator stage as such. Instead, one or more stages are turned on and off to modulate the carrier with Morse, with rise and fall times set by *R* and *C* values associated with the stages' keying and/or power supply lines. A transmitter's CW *waveshaping* is therefore usually hardwired to values appropriate for reasonably high-speed sending (35 to 55 WPM or so) in the presence of fading. However, some transceivers allow the user to vary keying hardness at will as a menu option. Rise and fall times of 1 to 5 ms are common; 5-ms rise and fall times equate to a keying speed of 36 WPM in the presence of fading and 60 WPM if fading is absent.

The faster a CW transmitter's output changes between zero and maximum, the more bandwidth its carrier and sidebands occupy. See Figure 13.33B. Making a CW signal's keying too hard is therefore spectrum-wasteful and inconsiderate of other stations because it makes the signal wider than it needs to be. Keying sidebands that are stronger and wider than necessary are traditionally called *clicks* because of what they sound like on the air.

Radiotelegraph or CW operation can be easily obtained from the transmitter architecture design shown in Figure 13.9. For CW operation, a carrier is generated at the center of the SSB filter passband. There are two ways to make this carrier available. One way is to unbalance the balanced modulator so that the LO can pass through. Each kind of balanced modulator circuit has its own method of doing this. The approach chosen in Figure 13.9 is to go around the modulator and the SSB filter.

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

13.4.2 RF Envelope Shaping

On-off keying (CW) is a special kind of low-level amplitude modulation (a low signal-level stage is turned on and off). It is special because the sideband power is subtracted from the carrier power, and not provided by a separate "modulator" circuit, as in high-level AM. It creates a spectrum around the carrier frequency whose amplitude and bandwidth are influenced by the rates of signal amplitude rise and fall and by the curvature of the keyed waveform.. For additional information see the article by Sabin on IF signal processing in the References section of this chapter.

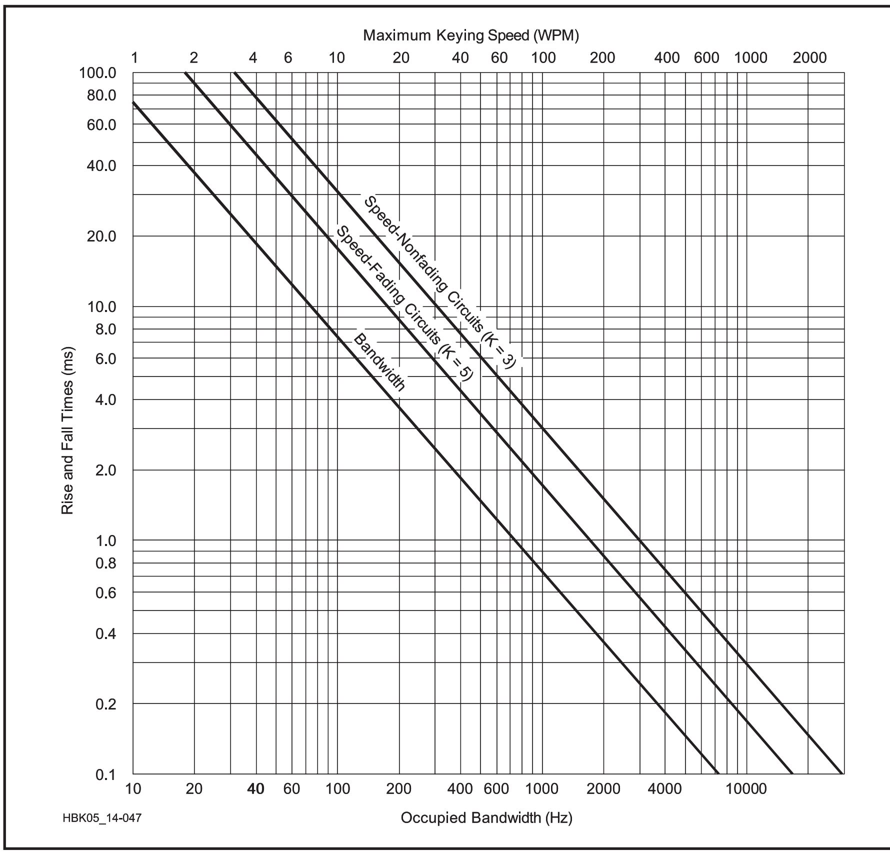


Figure 13.34 — Keying speed versus rise and fall times versus bandwidth for fading and nonfading communications circuits. For example, for transmitter output waveform rise and fall times of approximately 6 ms, draw a horizontal line from 6.0 ms on the rise and fall times scale to the bandwidth line. Then draw a vertical line to the occupied bandwidth scale at the bottom of the graph. In this case the bandwidth is about 130 Hz. Also extend the 6.0 ms horizontal line to the K = 3 line for a nonfading circuit. Finally draw a vertical line from the K = 3 line to the WPM axis. The 6 ms rise and fall time should be suitable for keying speeds up to about 50 WPM in this example.

Now look at **Figure 13.34**. The vertical axis is labeled Rise and Fall Times (ms). For a rise/fall time of 6 ms (between the 10% and 90% values) go horizontally to the line marked Bandwidth. A –20 dB bandwidth of roughly 120 Hz is indicated on the lower horizontal axis. Continuing to the K = 5 and K = 3 lines, the upper horizontal axis suggests code speeds of 30 WPM and 50 WPM respectively.

These code speeds can be accommodated by the rise and fall times displayed on the vertical axis. For code speeds greater than these the Morse code characters become "soft" sounding and difficult to copy, especially under less-than-ideal propagation conditions.

The ITU Classification of Emission Standards for determining necessary bandwidths of signals uses a value of 0.8 for the conversion between baud and WPM and suggests a typical value for K of 5 on an HF channel where the signal is subjected to fading. The bandwidth for a 13 WPM signal would then be:

$$BW = WPM \times 0.8 \times 5 = 10.4 \times 5 = 52 \text{ Hz}$$

For a narrow spectrum and freedom from adjacent channel interference, a further requirement is that the spectrum must fall off very rapidly beyond the –20 dB bandwidth indicated in Figure 13.34. A sensitive narrow-band CW receiver that is tuned to an adjacent channel that is only 1 or 2 kHz away can detect keying sidebands that are 80 to 100 dB below the key-down level of a strong CW signal.

An additional consideration is that during key-up a residual signal, called *backwave*, should not be noticeable in a nearby receiver. A backwave level at least 90 dB below the key-down carrier is a desirable goal.

Microprocessor-controlled transceivers manufactured today control CW keying rise- and fall-time through software. The operator generally accesses the keying shape parameter through a menu selection and adjustment process. Three to four ms is a typical value for most transceivers that balances crisp keying characteristics against excessive off-channel artifacts. See *QST* Product Reviews for waveforms and discussions of rise- and fall-time settings.

Homebrew equipment usually relies on analog circuitry to control keying waveforms. **Figure 13.35** is the schematic of one wave-shaping circuit that has been used successfully. A Sallen-Key third-order op amp low-pass filter (0.1 dB Chebyshev response) shapes the keying waveform, produces the rate of rise and fall and also softens the leading and trailing corners just the right amount. The key closure activates the CMOS switch, U1, which turns on the 455-kHz IF signal. At the key-up time, the input to the wave-shaping filter is turned off, but the IF signal switch remains closed for an additional 12 ms.

The keying waveform is applied to the gain control pin of a CLC5523 amplifier IC. This device, like nearly all gain-control amplifiers, has a *logarithmic* control of gain; therefore some experimental "tweaking" of the capacitor values was used to get the result shown in **Figure 13.36A**. The top trace shows the on/off operation of the IF switch, U1. The signal is turned on shortly before the rise of the keying pulse begins and remains on for

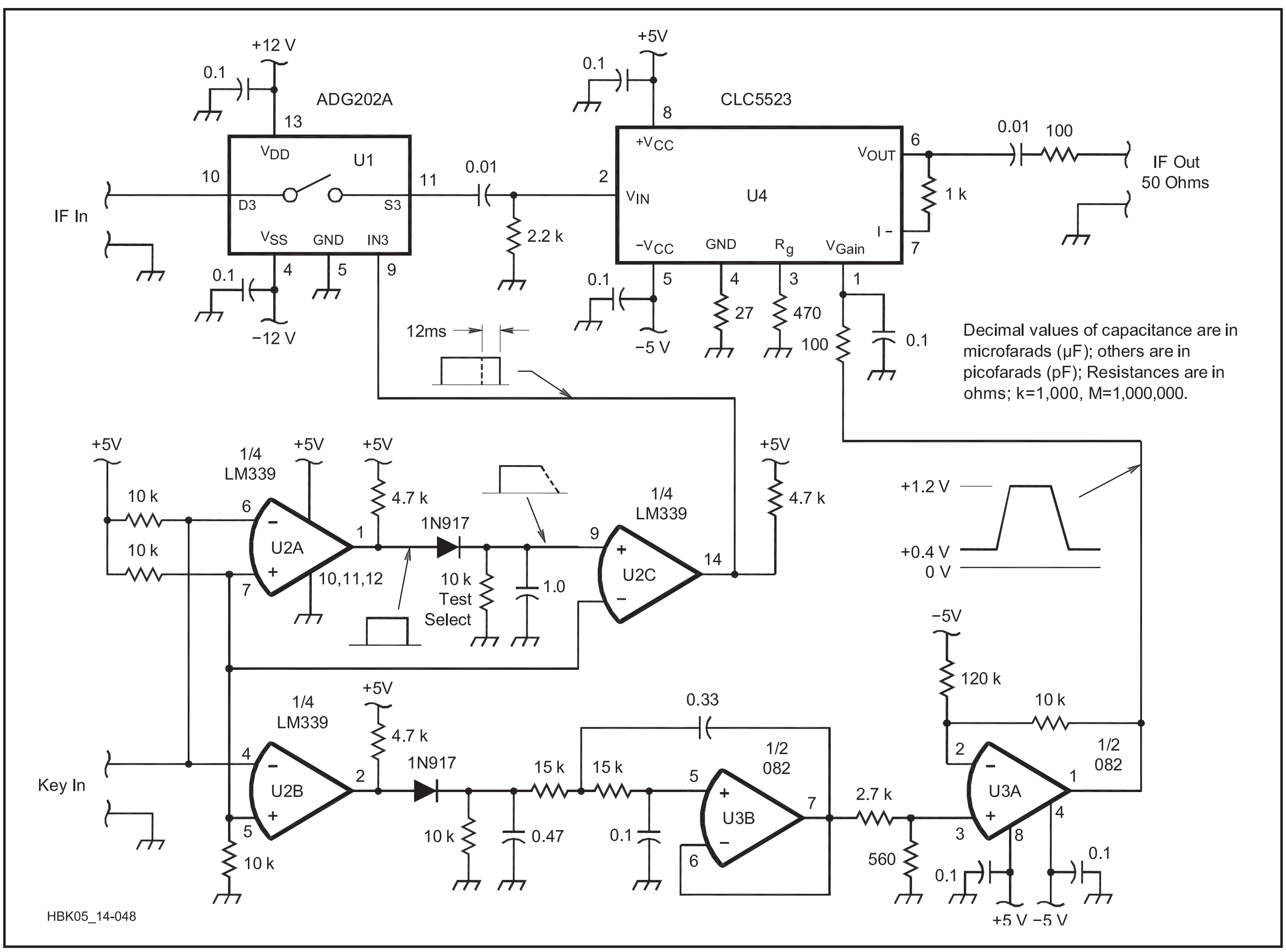


Figure 13.35 — This schematic diagram shows a CW waveshaping and keying circuit suitable for use with an SSB/CW transmitter such as is shown in Figure 13.9.

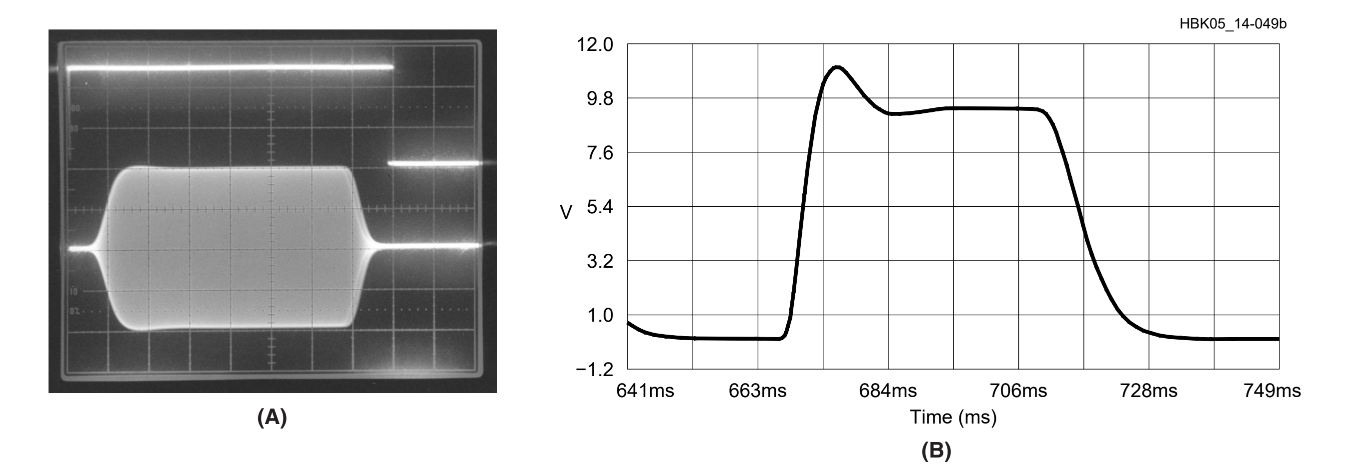


Figure 13.36 — At (A) is the oscilloscope display of the CW waveshaping and keying circuit output. The top trace is the IF keying signal applied to S1 of Figure 13.35. The bottom trace is the transmitter output RF spectrum. At (B) is a *SPICE* simulation of the waveshaping network. When this signal is applied to the logarithmic control characteristic of the CLC5523 amplifier, the RF envelope is modified slightly to the form shown in (A).

about 12 ms after the keying pulse is turned off, so that the waveform falls smoothly to a very low value. The result is an excellent spectrum and an almost complete absence of backwave. Compare this to the factory transmitter waveshapes shown in Figures 13.29 and 13.31. The bottom trace shows the resulting keyed RF output waveshape. It has an excellent spectrum, as verified by critical listening tests. The thumps and clicks that are found in some CW transmitters are virtually absent. The rise and fall intervals have a waveshape that is approximately a cosine. Spread-spectrum frequency-hop waveforms have used this approach to minimize wideband interference.

Figure 13.36B is an accurate *SPICE* simulation of the wave shaping circuit output before the signal is processed by the CLC5523 amplifier. To assist in adjusting the circuit, create a steady stream of 40 ms dots that can be seen on an RF oscilloscope that is looking at the final PA output envelope. It is important to make sure that the excellent waveshape is not degraded on its way to the transmitter output. Single-sideband linear power amplifiers are well suited for a CW transmitter, but they must stay within their linear range, and the backwave problem must be resolved.

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

13.4.3 Break-In CW Operation

Most current 100 W class HF transceivers use high-speed relays (with the relay actually following the CW keying) or solid-state PIN diodes to implement full break-in CW. Some RF power amplifiers use high-speed vacuum relays for the TR switching function. See the section on TR Switching later in this chapter for more information about circuits to perform this function. Two projects for adding QSK switching to linear amplifiers are included in the **Station Accessories** online material.

The term *semi-break-in* is used to designate a CW switching system in which closing the key initiates transmission, but switching back to receive happens between words, not between individual dits. Some operators find this less distracting than full break-in, and it is easier to implement with less-expensive relays for the TR switching.

13.5 Transmitting AM and SSB

13.5.1 Amplitude-Modulated Full-Carrier Voice Transmission

A popular form of voice amplitude modulation is called *high-level amplitude modulation.* It is generated by mixing (or modulating) an RF carrier with an audio signal. **Figure 13.37** shows the conceptual view of this**.** **Figure 13.38** is a more detailed view of how such a voice transmitter would actually be implemented. The upper portion is the RF channel and the lower portion is the audio frequency or AF channel, usually called the *modulator*. The modulator is nothing more than an audio amplifier designed to be fed from a microphone and with an output designed to match the anode or collector impedance of the final RF amplifier stage.

The output power of the modulator is

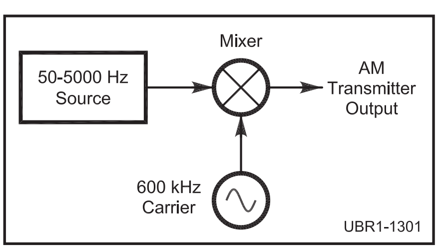


Figure 13.37 — Block diagram of a conceptual AM transmitter.

applied in series with the dc supply of the output stage (only) of the RF channel of the transmitter. The level of the voice peaks needs to be just enough to vary the supply to the RF amplifier collector between zero volts, on negative peaks, and twice the normal supply voltage on positive voice peaks. This usually requires an AF amplifier with about half the average power output as the dc input power (product of dc collector or plate voltage times the current) of the final RF amplifier stage.

The output signal, called *full-carrier double-sideband AM*, occupies a frequency spectrum as shown in **Figure 13.39.** The spectrum shown would be that of a standard broadcast station with an audio passband from 50 Hz

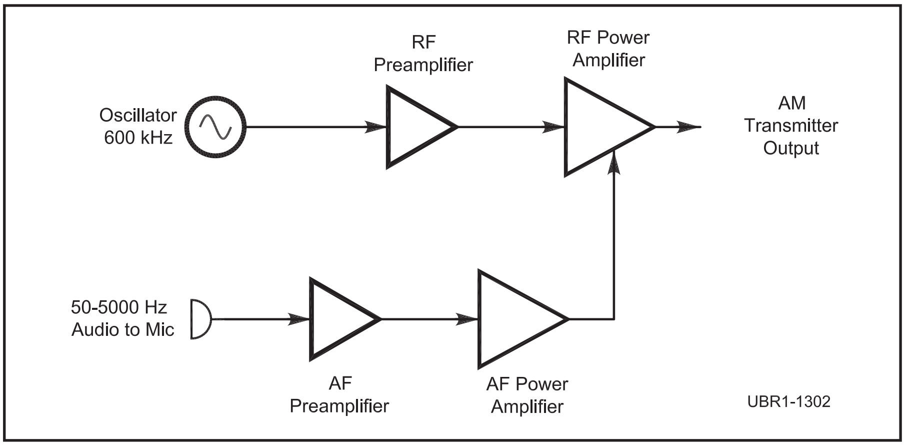


Figure 13.38 — Block diagram of a 600 kHz AM broadcast transmitter.

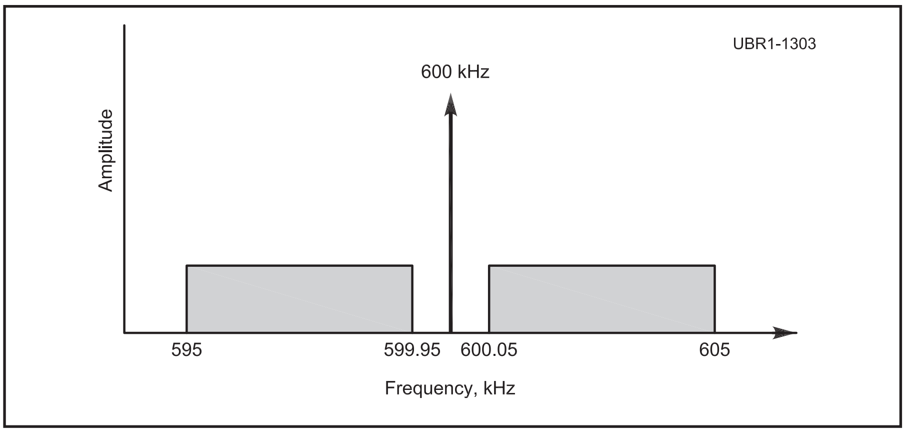


Figure 13.39 — The range of spectrum used by a 600 kHz AM broadcast signal showing sidebands above and below a carrier at 600 kHz.

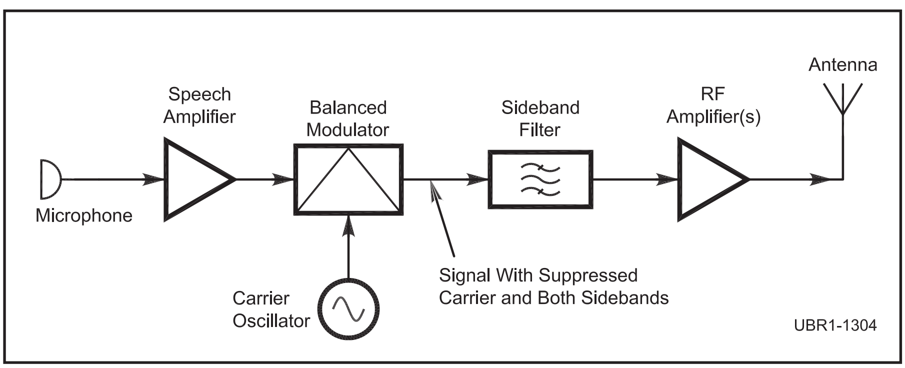


Figure 13.40 — Block diagram of a filter type single-sideband suppressed-carrier (SSB) transmitter.

to 5 kHz. Note that the resulting channel width is twice the highest audio frequency transmitted. If the audio bandwidth were limited to typical "telephone quality speech" of 300 to 3300 Hz, the resulting bandwidth would be reduced to 6.6 kHz. Note also that while a perfect multiplication process would result in just two sidebands and no carrier, this implementation actually provides the sum of the carrier and the sidebands from the product terms. (See the **Receiving** chapter for the mathematical description of signal multiplication.)

Full-carrier double-sideband AM is used in fewer and fewer applications. The spectral and power efficiency are significantly lower than single sideband (SSB), and the equipment becomes quite costly as power is increased. The primary application is in broadcasting — largely because AM transmissions can be received on the simplest and least expensive of receivers. With a single transmitter and thousands of receivers, the overall system cost may be less and the audience larger than for systems that use more efficient modulation techniques. While the PEP output of an AM transmitter is four times the carrier power, none of the carrier power is necessary to carry the information, as we will discuss in the next section.

13.5.2 Single-Sideband Suppressed-Carrier Transmission

The two sidebands of a standard AM transmitter carry (inverted) copies of the same information, and the carrier carries essentially no information. We can more efficiently transmit the information with just one of the sidebands and no carrier. In so doing, we use somewhat less than half the bandwidth, a scarce resource, and also consume much less transmitter power by not transmitting the carrier and the second sideband.

SSB — THE FILTER METHOD

The block diagram of a simple single-sideband suppressed-carrier (SSB) transmitter is shown in **Figure 13.40**. This transmitter uses a balanced mixer as a *balanced modulator* to generate a double sideband suppressed carrier signal without a carrier. That signal is then sent through a filter designed to pass just one (either one, by agreement with the receiving station) of the sidebands.

Depending on whether the sideband above or below the carrier frequency is selected, the signal is called *upper sideband* (*USB*) or *lower sideband* (*LSB*), respectively. The resulting SSB signal is amplified to the desired power level and we have an SSB transmitter. Amateur practice is to use USB above 10 MHz and LSB on lower frequencies. The exception is 60 meter channels, on which amateurs are required to use USB for data and voice signals.

While a transmitter of the type in Figure 13.40 with all processing at the desired transmit frequency will work, the configuration is not often used. Instead, the carrier oscillator and sideband filter are often at an intermediate frequency that is heterodyned to the operating frequency as shown in **Figure 13.41**. The reason is that the sideband filter is a complex narrow-band filter and most manufacturers would rather not have to supply a new filter design every time a transmitter is ordered for a new frequency. Many SSB transmitters can operate on different bands as well,

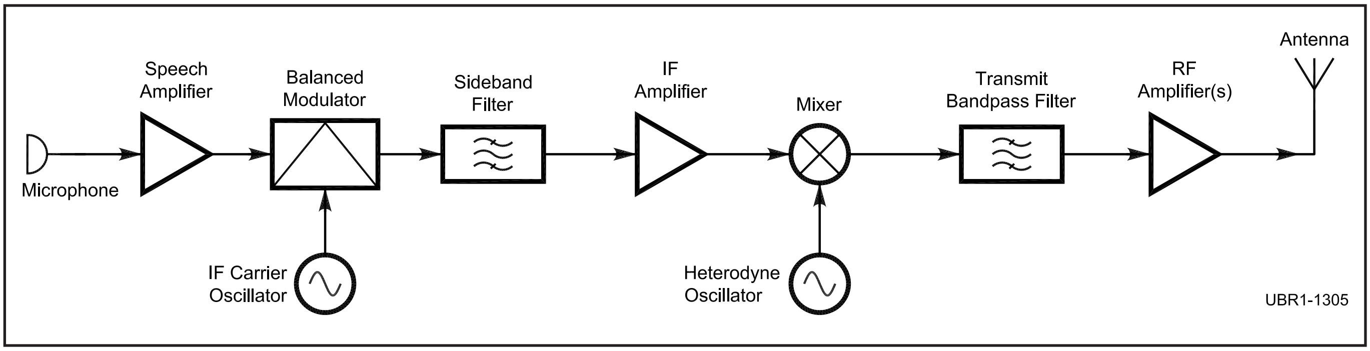


Figure 13.41 — Block diagram of a heterodyne filter-type SSB transmitter for multiple frequency operation.

so this avoids the cost of additional mixers, oscillators and expensive filters.

Note that the block diagram of our SSB transmitter bears a striking resemblance to the diagram of a superheterodyne receiver except that the signal path is reversed to begin with information and produce an RF signal. The same kind of image rejection requirements for intermediate frequency selection that were design constraints for the superhet receiver applies here as well.

The filter method can be implemented by analog circuitry, using high-quality crystal filters to remove the unwanted sideband and a carefully adjusted balanced modulator to eliminate the carrier. An SDR can also implement the filter method by using very sharp DSP filters at a low frequency (carrier frequency less than 100 kHz) and then up-converting the signal to the desired RF frequency. As DSP hardware continues to improve in speed, this method should continue to grow in popularity

SSB — THE PHASING METHOD

Most current transmitters use the method of SSB generation shown in Figure 13.40 to generate the SSB signal. That is the *filter method*, but really occurs in two steps — first a balanced modulator is used to generate sidebands and eliminate the carrier, then a filter is used to eliminate the undesired sideband, and often to improve carrier suppression as well.

The *phasing method* of SSB generation is exactly the same as the image-rejecting mixer described in the **Receiving** chapter. This uses two balanced modulators and a phase-shift network for both the audio and RF carrier signals to produce the upper sideband signal as shown in **Figure 13.42A**. By a shift in the sign of either of the phase-shift networks, the opposite sideband can be generated. This method trades a few phase-shift networks and an extra balanced modulator for the sharp sideband filter of the filter method.

While it looks deceptively simple, a limitation is in the construction of an analog phase-shift network that will have a constant 90° phase shift over the whole audio range. Errors in phase shift result in less than full carrier and sideband suppression. Nonetheless, there have been some successful examples offered over the years.

The 90º phase shift network is also known as a Hilbert transformer after the mathematical operation it performs. Difficult to implement well in analog form, the phase shift is straightforward to implement digitally. That

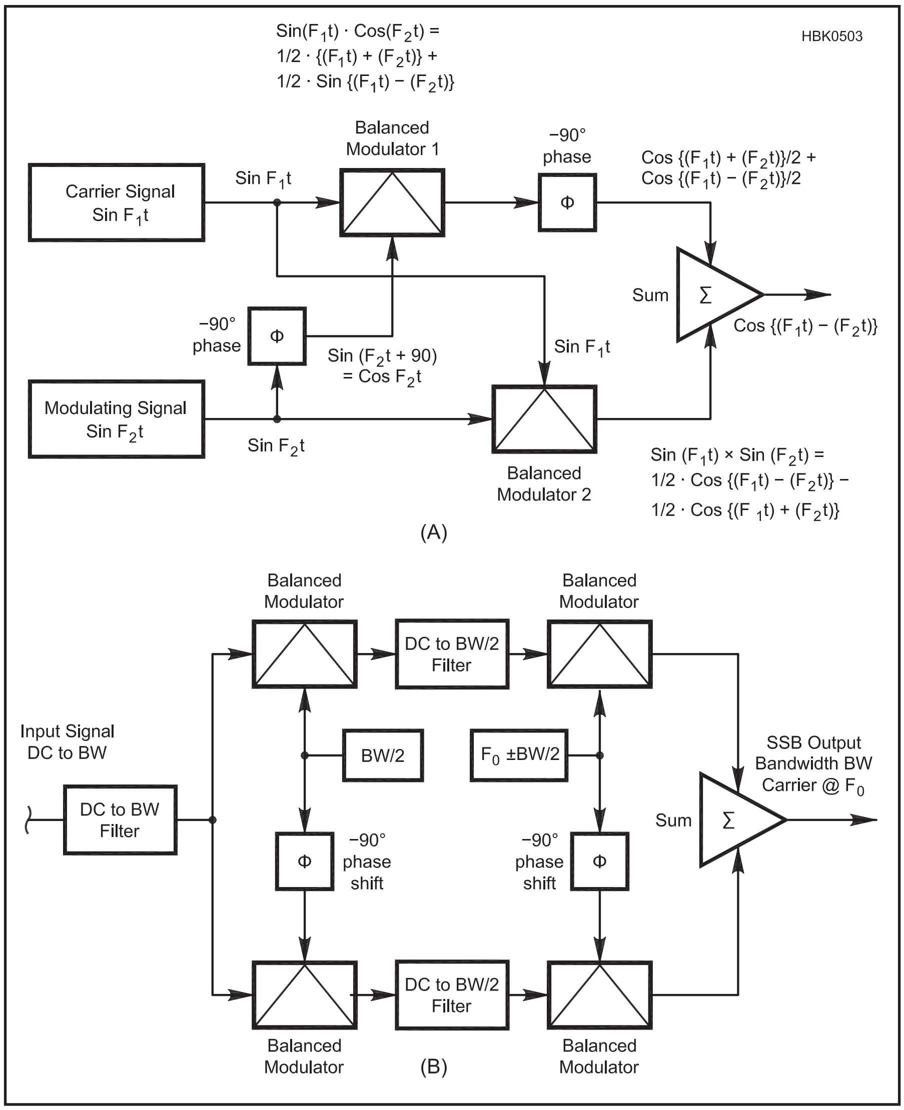


Figure 13.42 — Block diagram of (A) phasing type and (B) Weaver method SSB generators for single-frequency operation.

makes the phasing method a good choice for SSB generation by an SDR transmitter.

SSB — THE WEAVER METHOD

Taking the phasing method one step further, the Weaver method solves the problem of requiring phase-shift networks that must be aligned across the entire audio range. Instead, the Weaver method, shown in Figure 13.42B, first mixes one copy of the message (shown with a bandwidth of dc to BW Hz) with an in-band signal at BW/2 Hz and another copy with a signal at BW/2 Hz that is phase-shifted by –90°. Instead of phase-shifting the message, only the signal at BW/2 Hz must be phase-shifted — a much simpler task!

The output of each balanced modulator is filtered, leaving only components from dc to BW/2. These signals are then input to a second pair of balanced modulators with a more conventional LO signal at the carrier frequency, f_0, offset by +BW/2 for USB and –BW/2 for LSB. The output of the balanced modulators is summed to produce the final SSB signal.

The Weaver method is difficult to implement in analog circuitry, but is well-suited to digital signal processing systems. The Weaver method has become common in DSP-based equipment that generates the SSB signal digitally.

13.6 Transmitting Angle Modulation

13.6.1 Angle-Modulated Transmitters

Transmitters using frequency modulation (FM) or phase modulation (PM) are generally grouped into the category of *angle modulation* since the resulting signals are often indistinguishable. An instantaneous change in either frequency or phase can create identical signals, even though the method of modulating the signal is somewhat different. To generate an FM signal, we need an oscillator whose frequency can be changed by the modulating signal.

We can make use of an oscillator whose frequency can be changed by a "tuning voltage." If we apply a voice signal to the TUNING VOLTAGE connection point, we will change the frequency with the amplitude and frequency of the applied modulating signal, resulting in an FM signal.

The phase of a signal can be varied by changing the values of an R-C phase-shift network. One way to accomplish phase modulation is to have an active element shift the phase and generate a PM signal. In **Figure 13.43**, the drain current through the field-effect transistor is varied with the applied modulating signal, varying the phase shift at the stage's output. Because the effective load on the stage is changed, the carrier is also amplitude-modulated and must be run through an FM receiver-type limiter in order to remove the amplitude variations.

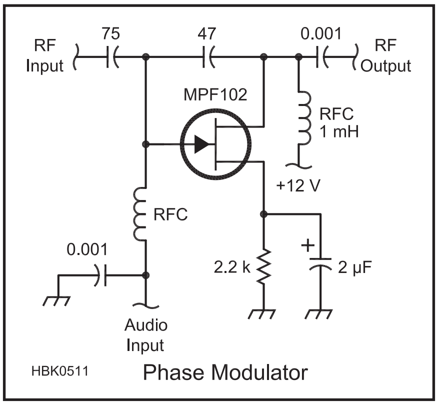


Figure 13.43 — Simple FET-based phase modulator circuit.

FREQUENCY MODULATION TRANSMITTER DESIGN

Frequency modulation is widely used as the voice mode on VHF for repeater and other point-to-point communications. **Figure 13.44** shows the phase-modulation method, also known as *indirect FM*, as used in many FM transmitters. It is the most widely used approach to FM. Phase modulation is performed at a low frequency, say 455 kHz. Prior to the phase modulator, speech filtering and processing perform four functions:

1. Convert phase modulation to frequency modulation (see below).

2. Apply pre-emphasis (high-pass filtering) to the speech audio higher speech frequencies for improved signal-to-noise ratio after de-emphasis (low-pass filtering) of the received audio.

3. Perform speech processing to emphasize the weaker speech components.

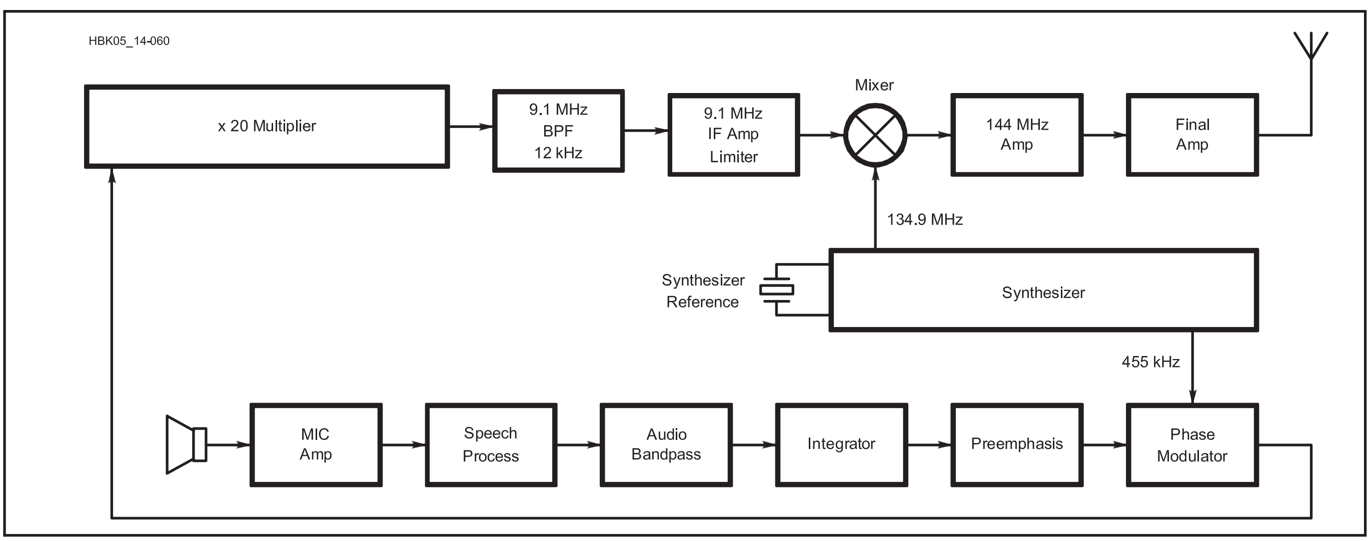


Figure 13.44 — Block diagram of a VHF/UHF NBFM transmitter using the indirect FM (phase modulation) method.

4. Compensate for the microphone's frequency response and possibly also the operator's voice characteristics.

Multiplier stages then move the signal to some desired higher IF and also multiply the frequency deviation to the desired final value. If the FM deviation generated in the 455 kHz modulator is 250 Hz, the deviation at 9.1 MHz is 20 × 250, or 5 kHz.

13.6.2 Frequency Multipliers

Frequency multipliers are frequently used in FM transmitters as a way to increase the deviation along with the carrier frequency. They are composed of devices that exhibit high levels of harmonic distortion, usually an undesired output product. In this case the desired harmonic is selected and enhanced through filtering. The following examples show the way this can be done, both with amplifiers and with passive diode circuits. (Additional discussion of frequency multipliers can be found in Chapter 5.4 of *Experimental Methods in RF Design* and in the "VHF Signal Sources" article by Rick Campbell, KK7B which is included in this book's downloadable supplemental content.)

A passive multiplier using diodes is shown in **Figure 13.45A**. The full-wave rectifier circuit can be recognized, except that the dc component is shorted to ground. If the fundamental frequency ac input is 1.0 V_{RMS}, the second harmonic is 0.42 V_{RMS} or 8 dB below the input, including some small diode losses. This value is found by calculating the Fourier series coefficients for the full-wave-rectified sine wave, as shown in many textbooks.

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

An FET, biased at a certain point, is very nearly a *square-law* device as described in the **Electrical Fundamentals** chapter. That is, the drain-current change is proportional to the square of the gate-voltage change. It is then an efficient frequency doubler that also de-emphasizes the fundamental.

A push-push doubler is shown in Figure 13.45B. The FETs are biased in the square-law region and the BALANCE potentiometer minimizes the fundamental frequency. Note

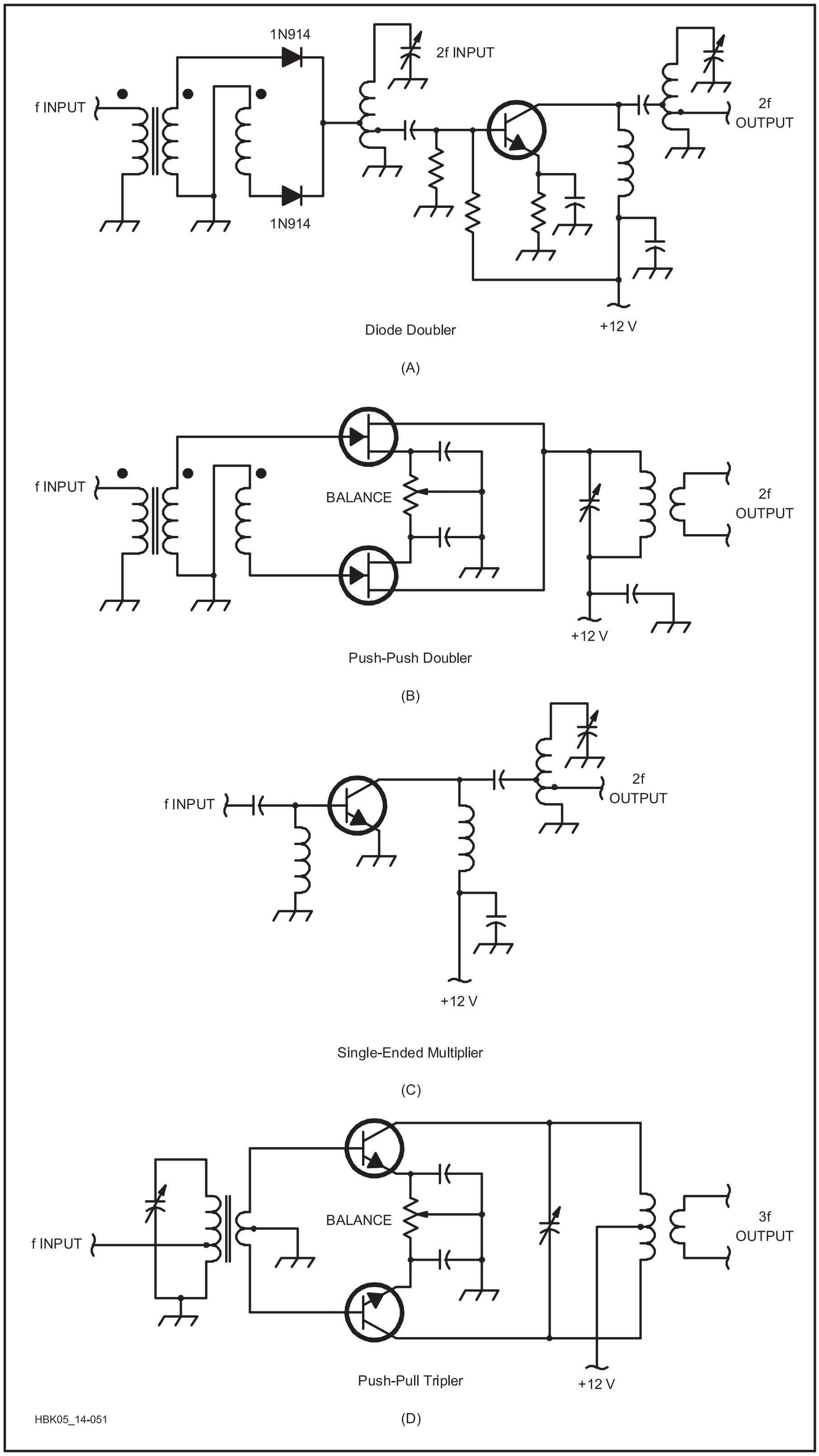


Figure 13.45 — Frequency multipliers. A: diode doubler. B: push-push doubler using JFETS. C: single-ended multiplier using a BJT. D: push-pull tripler using BJTs.

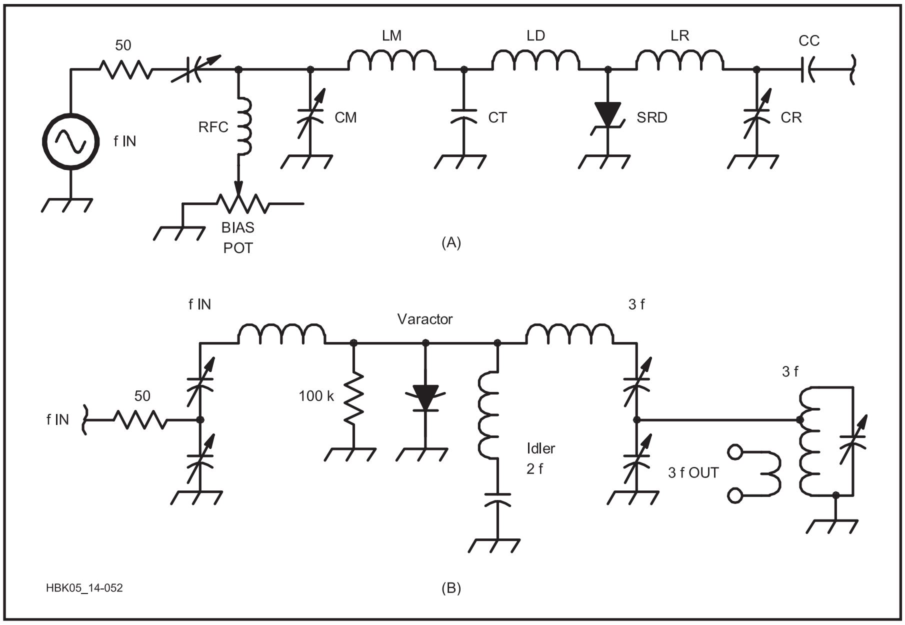


Figure 13.46 — Specialized diode frequency multipliers. A: step-recovery diode multiplier. B: varactor diode multiplier.

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

Figure 13.45C shows an example of a single-ended doubler using a bipolar transistor. The efficiency of a doubler of this type is typically 50%, that of a tripler 33%, and of a quadrupler 25%. Harmonics other than the one to which the output tank is tuned will appear in the output unless effective band-pass filtering is applied. The collector tap on L1 is placed at the point that offers the best compromise between power output and spectral purity.

A push-pull tripler is shown in Figure 13.45D. The input and output are both push-pull. The balance potentiometer minimizes even harmonics. Note that the transistors have no bias voltage in the base circuit; this places the transistors in Class C for efficient third-harmonic production. Choose an input drive level that maximizes harmonic output.

The step recovery diode (SRD) shown in **Figure 13.46A** is an excellent device for harmonic generation, especially at microwave frequencies. The basic idea of the SRD is as follows: When the diode is forward conducting, a charge is stored in the diode's diffusion capacitance, and if the diode is quickly reverse-biased, the stored charge is very suddenly released into an LC harmonic-tuned circuit. The circuit is also called a "comb generator" because of the large number of harmonics that are generated. (The spectral display looks like a comb.) A phase-locked loop (PLL) can then lock onto the desired harmonic.

A varactor diode can also be used as a multiplier. Figure 13.46B shows an example. This circuit depends on the fact that the capacitance of a varactor changes with the instantaneous value of the voltage across it, in this case the RF excitation voltage. This is a nonlinear process that generates harmonic currents through the diode. Power levels up to 25 W can be generated in this manner.

Following frequency multiplication, a second conversion to the final output frequency is performed. Prior to this final translation, IF band-pass filtering is performed in order to minimize adjacent channel interference that might be caused by excessive frequency deviation. This filter needs good phase linearity to assure that the FM sidebands maintain the correct phase relationships. If this is not done, an AM component is introduced to the signal, which can cause nonlinear distortion problems in the PA stages. The final frequency translation retains a constant value of FM deviation for any value of the output signal frequency.

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

In phase modulation, the frequency deviation is directly proportional to the frequency of the audio signal. (In FM, the deviation is proportional to the audio signal's amplitude.) To make deviation independent of the audio frequency, an audio-frequency response that rolls off at 6 dB per octave is needed. An op-amp low-pass circuit in the audio amplifier accomplishes this function. This process converts phase modulation to frequency modulation.

In addition, audio speech processing helps to maintain a constant value of speech amplitude, and therefore constant IF deviation, with respect to audio speech levels. Pre-emphasis of speech frequencies (a 6 dB per octave high-pass response from 300 to 3000 Hz) is commonly used to improve the signal-to-noise ratio at the receive end. Analysis shows that this is especially effective in FM systems when the corresponding de-emphasis (complementary low-pass response) is used at the receiver. (See reference for Schwartz.) By increasing the amplitude of the higher audio frequencies before transmission and then reducing them in the receiver, high-frequency audio noise from the demodulation process is also reduced, resulting in a "flat" audio response with lower hiss and high-frequency noise.

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

FM is more tolerant of frequency misalignments between the transmitter and receiver than is SSB. In commercial SSB communication systems, this problem is solved by transmitting a *pilot carrier* with an amplitude 10 or 12 dB below the full PEP output level. The receiver is then phase-locked to this pilot carrier. The pilot carrier is also used for squelch and AGC purposes. A short-duration "memory" feature in the receiver bridges across brief pilot-carrier dropouts, caused by multipath nulls.

In a "direct FM" transmitter, a high-frequency (say, 9 MHz or so) crystal oscillator is frequency-modulated by varying the voltage on a varactor diode. The audio is pre-emphasized and processed ahead of the frequency modulator as for indirect-FM.

13.7 Effects of Transmitted Noise

With receiver sensitivity, selectivity, and linearity having reached extraordinary levels of performance, a reduction in transmitted spurious emissions is clearly in the best interests of all amateurs. It does us no good to spend time and effort creating an exceptional receiver if the channel is filled with transmitted noise and distortion products! (See the article by Grebenkemper in the Reference section.)

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

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

SDR transmitters may not have analog mixers but they certainly have data converters and many processes that generate phase noise through clock jitter, non-linear data conversion, and other causes, even rounding errors! The result — heard on the air as transmitted noise — causes the same problems as noise generated by an analog transmitter. **Figure 13.47** is a collection of composite noise (phase noise) and CW keying sideband noise from several current transceivers evaluated by the ARRL Lab during product review assessments. The data is from a single band (14 MHz) and does not represent either the best or worst performance. This chart was initially created in 2014 from that data by Jim Brown, K9YC, to compare the noise performance of several transceivers. (The FT1000MP Mark V Field legacy spectrum shows the improved performance available today.) The data points on the graph represent noise peaks from the measured noise spectra. For the CW sideband noise data, a comparison of averaged and smoothed data is available in K9YC's online paper referenced at the end of this section. The paper also includes data for several older transceiver models.

There is quite a bit of difference between the various models as described in the *QST* Product Reviews. The variation shows why it is important for amateurs to pay attention to noise performance specifications, especially if they plan on operating in the crowded HF bands. With many transceivers now offering upgradable firmware, amateurs are strongly encouraged to install the latest version. This will improve performance on both receive and transmit, as well as help all of us transmit cleaner signals on the air.

Transmitted noise plays the same role in interference as receiver phase noise. The two are ultimately additive and interference becomes a "weakest link" problem in that poor noise performance of the receiver or the transmitter can be the culprit in an interference

Noise Consideration for Antenna Layout

By Steve Hicks, N5AC

Transmitted noise is a point to consider when laying out the antennas at a multi-transmitter site like Field Day or a contest station. For every 10 dB stronger any given signal appears in the receiver, another 10 dB improvement in phase noise performance is required to ensure interference-free operation. A poor antenna layout that produces very strong signals in receivers when local transmitters are operating can render even the best phase noise performance receivers inoperable. George Cutsogeorge, W2VJN's book *Managing Interstation Interference* (available from Inrad (**www.inrad.net**) is an excellent reference for station planning and interference mitigation techniques.

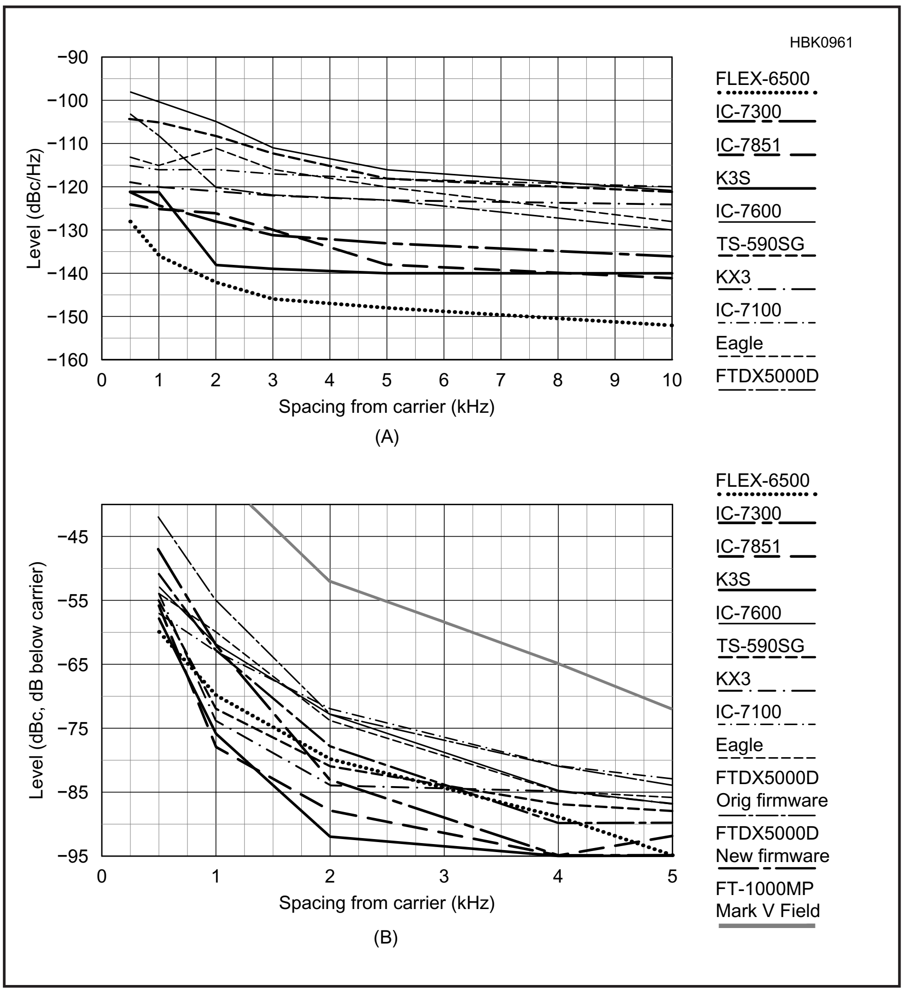


Figure 13.47 — Transmitted composite noise (A) and keying sidebands (B) from a selection of representative modern transmitters measured by the ARRL Lab. These graphs are based on ARRL Lab test data published in *QST* Product Reviews, following graphs originally created by Jim Brown, K9YC

problem. While both amplitude and phase noise contribute to transmitted composite noise, most of the noise contribution is phase noise from the various oscillators and clock signals in the transceiver. (The effects of receiver phase noise are addressed in the **Receiving** chapter.)

In a multiple-transmitter operation such as Field Day or a contest station, a single radio with poor transmitted noise performance can render all of the receivers at the site useless. Similarly, a receiver with poor phase noise performance can suffer from interference issues regardless of how good the performance of co-located transmitters is. In situations where multiple radios are brought to a single location for a joint operation, transmit phase noise is typically more strongly scrutinized since a single poor performer in the transmit phase noise arena can render the whole operation a failure.

Transmitted noise is not just a problem when multiple stations are at one site. Poor noise performance at full power can have an adverse effect on other stations for miles around. In competitive environments, noise from a closely-spaced strong signal can render adjacent channels nearly unusable. Under normal circumstances, transmitted noise raises the noise floor everywhere there is propagation. After decades of receiver improvement, it is important for amateurs to pay closer attention to transmitted noise performance.

The ARRL Lab measures transmitted noise as part of Product Reviews for *QST* magazine. These tests are described in the **Test Measurements and Equipment** chapter and in the book *Amateur Radio Transceiver Performance Testing* by Bob Allison, WB1GCM. Individual transceiver performance is documented in the *QST* Product Review. In addition to creating Figure 13.47, Jim Brown, K9YC has compiled the information from these reviews in an online paper, "Comparison of ARRL Lab Data for Selected Transceivers" which is available at **k9yc.com/publish.html**. Figure 13.47 is typical of the many figures in the paper comparing various noise measurements.

13.8 Microphones and Speech Processing

13.8.1 Frequency Content of Speech

Human speech has content from about 100 Hz to 8 kHz, but only the energy between about 400 Hz and 4 kHz contributes to speech intelligibility. Vocal content below 400 Hz provides "body" to the voice (great for singers and radio announcers), but that low frequency output of the microphone also contains breath pops, room noise, microphone handling noise, wind noise, and reverberation. This low frequency energy can easily be as much as half of the power picked up by the microphone, but it contributes nothing to communications, so it wastes transmitter power.

Likewise, speech content above 3 kHz provides "presence" and helps communications a bit, but the added bandwidth adds noise (and QRM to and from other stations). Most SSB transmitting filters are 2.7 kHz wide, so a well-adjusted radio will align those filters so that they pass audio between 400 Hz and 3.1 kHz. A few radios allow the user to adjust this setting via a configuration menu.

These bandwidth limits for speech communications were established in the earliest days of long distance telephony — they allow what's necessary, but nothing extra. Over more than a century, they have allowed more and more conversations to be packed into less and less bandwidth.

Thus, our first rule is to minimize any part of the audio signal below about 400 Hz, and to not waste bandwidth transmitting sound above 3 kHz. We have several controls over this. First, we can choose a microphone without excess low frequency response. See "Choosing a Microphone" later in this section. Many transceivers provide menu settings to tailor the audio frequency response. Study the manual to understand and choose settings for your radio.

Some newer transceivers make it even easier to tailor the frequency response — they have a built-in octave-band equalizer covering the speech range. (An octave is a 2:1 frequency step). Each band can be set for up to 18 dB of boost or cut in 1 dB steps. A good starting point for most microphones and voices is maximum cut of the three lowest bands (50, 100, 200 Hz), and 3–6 dB cut of the fourth band (400 Hz) leaving all other bands set flat (no boost or cut). Some microphone or voices may benefit from a bit more cut at 400 Hz or from 3-6 dB of cut or boost in the two highest bands. Save these adjustments for when you have a trained listener to advise you.

13.8.2 Dynamics Processing

The loudness of speech varies over a wide range as we speak, sometimes by as much as 20 dB. The audio section of our transmitter must be adjusted so that we never overmodulate, which causes interference on adjacent frequencies (splatter), but keeping audio level as close as practical to 100% modulation makes our signal louder at the other end. Modern transceivers include peak limiters to prevent overmodulation on loudness peaks and compressors to increase the audio gain for quieter parts of speech. When well-designed and carefully adjusted, these circuits work well, increasing our "talk power," but when badly adjusted, cause transmitted audio to be distorted, mushy, dull, and hard to understand.

The strength of sound falls off with distance from the source by inverse square law, just as RF field strength falls off with distance from the antenna. Doubling (or halving) the distance between microphone and mouth causes loudness to change by 6 dB. A boom microphone attached to a headset helps maintain more constant level by keeping the microphone element at a fixed distance.

To adjust these circuits, first set equalization as in the previous section, then set microphone gain so that normal speech causes near 100% modulation with compression disabled. Then set the transceiver display to show compression and adjust compression (or processing) to provide about 10 dB gain reduction on voice peaks. While radios can be set for more compression, few sound very good with more than 10 dB, and too much compression can make speech hard to copy. (On some transceivers, compression may be called "processing.") The combined effect of equalization to eliminate speech content below 400 Hz and using 10 dB of compression is about 13 dB, which is equivalent to multiplying transmitter power by a factor 20!

13.8.3 Types of Microphones

Most common microphones in the amateur station are one of two basic types:

Dynamic microphones operate on the same principle as a loudspeaker, (a coil mounted to a diaphragm is moved by varying magnetic field that surrounds it makes air vibrate) but in reverse (vibration moves a diaphragm in the microphone, generating a voltage in the coil as it moves within the magnetic field). A loudspeaker works pretty well as a microphone, and has been used that way for more than half a century in intercom systems.

Electret condenser microphones are very different — the diaphragm is one plate of a capacitor; a voltage is applied between the two plates (the other being fixed). The source impedance is quite high (megohms), and must be transformed to a lower impedance by a FET follower built into the microphone (so that what it feeds doesn't load down the microphone). The electret capsule is pre-polarized, but the FET follower needs a small positive voltage fed through a load resistor to operate. This is referred to as *bias voltage* and it is applied to the output of the microphone: 8 V dc through a 5.6 kΩ resistor is typical. (Note that if bias voltage is applied to a dynamic microphone element, the result will be muffled, low-volume audio. If the radio's microphone connection supplies bias voltage, a series blocking capacitor of 0.1 to 1 µF should be used to remove the bias voltage.)

Both electrets and dynamic microphones are available with an *omnidirectional* pattern (picks up equally in all directions) or a *cardioid* pattern (picks up better in front of the microphone with a null to the rear). Cardioids can be thought of as "half space" microphones, meaning they pick up sounds from anywhere in front of the microphone but reject sound from all directions to the rear.

Cardioid microphones have an important characteristic called *proximity effect*, which is a very strong bass boost for sound sources very close to the microphone. In addition to making voices "bass heavy," proximity effect magnifies breath pops, wind noise, and handling noise. Virtually all microphones used in live sound are cardioids, and those intended for use by singers have a strong low frequency rolloff that partially compensates proximity effect. Although cardioids reduce room noise pickup, proximity effect generally makes them a poor choice in the ham shack. A microphone designed for radio communications is a better choice.

Cardioids work on the principle of acoustic cancellation between sound reaching the element via front and rear openings of the microphone housing. (Omnidirectional microphones have a single opening). Proximity effect is the result of that process and there being a single front opening and a single rear opening.

An important variation of the cardioid microphone is built with extra openings spaced along the length of the handle, which greatly reduces proximity effect. The ElectroVoice (EV) 664 and 666 were the first popular microphones of this type, which are called "variable-D" (for the variable distant openings), as opposed to "single-D" cardioids with a single rear opening. If you're looking for a good used pro-quality microphone for your ham station, the variable-D EV RE10, 11, 15, 16, 18, 20, and 27, and the Shure SM53 and SM54 are great choices. All but the RE16, 20, and 27 are long discontinued, but dynamic microphones last indefinitely as long as they are not badly mistreated, so buying used from a trustworthy source is a good option.

An omnidirectional microphone, whether dynamic or an electret, or one of the variable-D models listed above, are the best choice for ham radio. They have no proximity effect, so can be used close to the mouth. This minimizes breath pops, while still being close enough to minimize room noise. The soft foam supplied with many microphones is intended to reduce breath pops.

MICROPHONE EQUALIZATION

Beginning in the late 1950s, Shure introduced the model 440, the first microphone designed specifically for SSB transmission. The modern version is the 444D. These are omni-directional mics with the recommended equalization built in — low frequency response rolling off below 400 Hz and a pronounced peak around 3 kHz that compensates for some of the loss in the SSB transmit filter. These are excellent sounding microphones and are primarily desktop models. Most microphones in the Heil line use the same concepts.

13.8.4 Using a Professional or PC Microphone

Pro microphones are balanced and are designed to feed balanced inputs using shielded, twisted pair cables, while ham microphones and rigs use unbalanced wiring. Pro electret microphones cannot easily be used with ham gear (because of the method used to power the balanced microphone's FET follower output stage).

Pro dynamic microphones work well and are easy to connect with ham gear. Their 3-pin XL-connector comes wired for balanced circuits — Pin 1 is the shield, Pins 2 and 3 carry the signal. To connect them to your ham rig with shielded twisted pair, wire the shield to the shell of the microphone connector and connect the signal pair to MIC and MIC RETURN (or MIC GND).

Alternatively, with coaxial cable between the microphone and the radio, wire the cable shield to pin 1 and 3 at the microphone and the center conductor to pin 2. At the radio's microphone connector, wire the center conductor to the radio's MIC input and wire the cable shield both to MIC RETURN and to the connector shell.

Pro microphones generally have uniform ("flat") frequency response. They have more bass response than communications microphones and lack the boost around 3 kHz, so they generally require more equalization. The low frequency rolloff in audio circuits is set by the time constant of interstage coupling capacitors and their resistive load. Using a smaller value capacitor raises the –3dB frequency ($X_C = R$) producing a gentle low frequency rolloff (6 dB/octave). In transmitters that lack an equalizer, raising the –3dB frequency to 1-2 kHz (by reducing those time constants) can provide much of the equalization needed to make a pro microphone produce good communications audio.

Headsets with a microphone attached made for use with computers work well with ham transceivers. **Figure 13.48** illustrates good placement that minimizes breath pops and allows an occasional sip of water or coffee. Most of these microphones are electrets with both headphones and microphone wired to 3.5-mm (⅛-inch) phone plugs (also known as TRS or tip-ring-sleeve plugs). Almost any of these headsets will work with ham transceivers, but some are far more comfortable than others. The Yamaha CM500 and Koss SB-45 and CS-100 are popular with contesters and DXers. All that is needed is a cable adapter to a Foster plug (the round 8-pin connector used on most transceivers) to mate with the 8-pin microphone connector on transceivers that lack a 3.5 mm microphone jack. Adapters for RJ-45 microphone connectors are also available. **Figure 13.49** shows a typical

Figure 13.48 — Using a computer-style headset electret boom microphone. Place the microphone far enough from your mouth to avoid picking up excessive noise from breathing and pops from speech. Balance the placement for normal speech levels while minimizing pickup of room and fan noise. On-the-air testing is important to account for your station circumstances and speaking style.

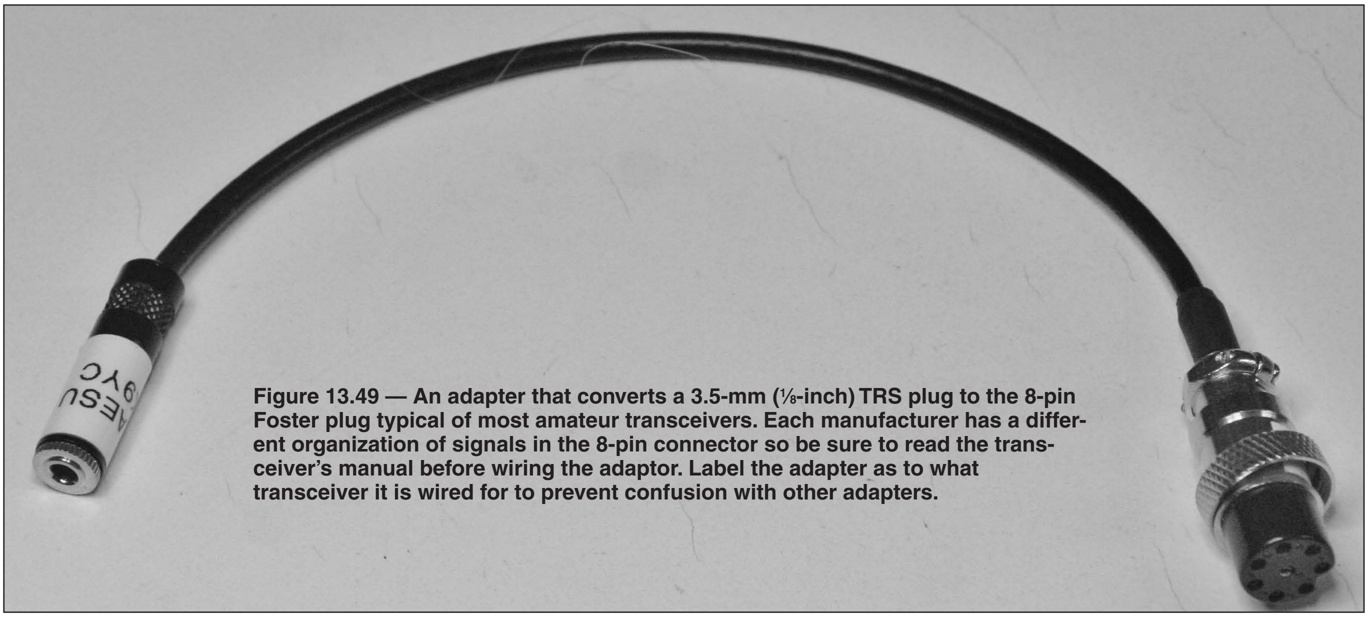


Figure 13.49 — An adapter that converts a 3.5-mm (⅛-inch) TRS plug to the 8-pin Foster plug typical of most amateur transceivers. Each manufacturer has a different organization of signals in the 8-pin connector so be sure to read the transceiver's manual before wiring the adaptor. Label the adapter as to what transceiver it is wired for to prevent confusion with other adapters.

adapter you can build yourself.

To make the adapter, you'll need a cable-mount plug to mate with your radio and a female ⅛-in TRS jack to mate with the TRS plug on the headset. Check the manual for your transceiver for wiring of the microphone connector and label the adapter.

To connect the microphone to the radio, run a single-conductor shielded cable (such as mini-coax or braid-shielded audio cable) from the tip of the TRS jack to the microphone input pin of the Foster plug, connecting the cable shield to the sleeve of the TRS jack and the shell of the Foster plug.

Nearly all modern radios supply bias voltage in the range of 8 V on one pin of the microphone connector; wire a 5.6 kΩ resistor between that pin and the MIC pin. This resistor can have a very low power rating, so it's usually possible to fit it inside the Foster plug. Buy Foster plugs from ham vendors; female TRS jacks can be bought from pro audio vendors such as Full Compass and Sweetwater. (Neutrik part number NYS240BG is typical.)

13.8.5 Optimizing Your Microphone Audio

Summarizing the steps to optimize audio for communications:

1) Set your radio to minimize audio content below 400 Hz and above 3.2 kHz.

2) Keep the distance between microphone and mouth as constant as possible. A boom microphone attached to a headset solves this problem.

3) Get audio gains set right: adjust the mic input of the radio (or of the computer), the output gain of the computer sound card, and/or the line input of the radio.

4) Set processing for an indicated 10 dB on voice peaks.

5) Resist the urge to turn up mic gain or compression — once levels have been set as described here, turning it up louder makes your voice sound *worse*, not better.

You should also make sure the transmitter RF controls are set properly:

6) Tune the RF power amp carefully.

7) Don't overdrive the RF amp, and don't use ALC to set TX power.

8) If your radio requires a nominal 13.8 V dc power supply, make sure the supply is as close to 14 V as possible.

13.8.6 Microphone Pads

(The following material was contributed by Ethan Winer. More design information is available on his website: **ethanwiner.com/gadgets.html**.)

The balanced microphone attenuator or *pad* shown in **Figure 13.50**. is about as basic a circuit as you're likely to encounter. But finding the optimum resistor values can be a challenge for the beginner, and a time consuming nuisance for the more advanced.

Figure 13.51.shows the same general configuration, except this circuit yields a low-end rolloff beginning at a frequency dictated by the resistor and capacitor values. Notice that the designation "R1" does not appear in this diagram. The same component numbering is used throughout the figures to make the

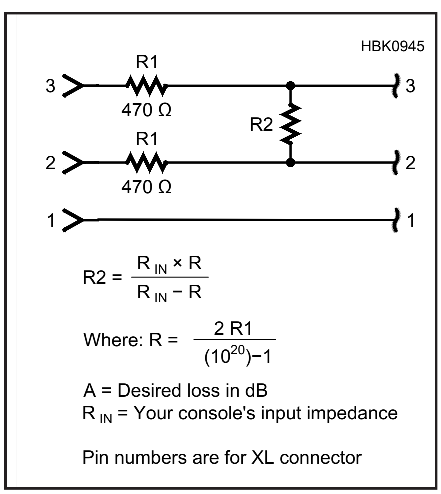


Figure 13.50 — Basic microphone attenuation pad.

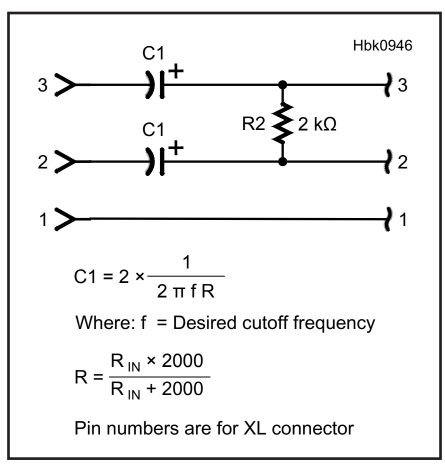


Figure 13.51— Low-frequency rolloff pad.

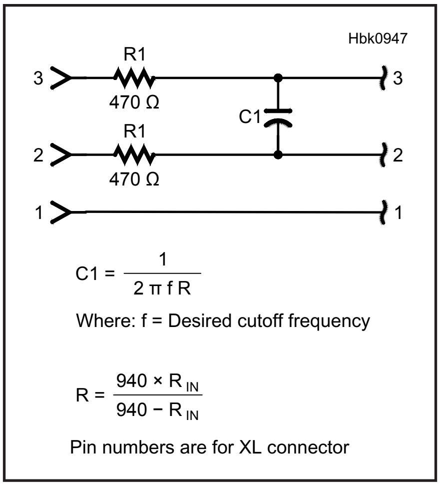


Figure 13.52— High-frequency rolloff pad.

formulas easier to understand. Also note that the low-cut circuit in Figure 51 will not pass bias voltage (also known as *phantom power* in professional audio).

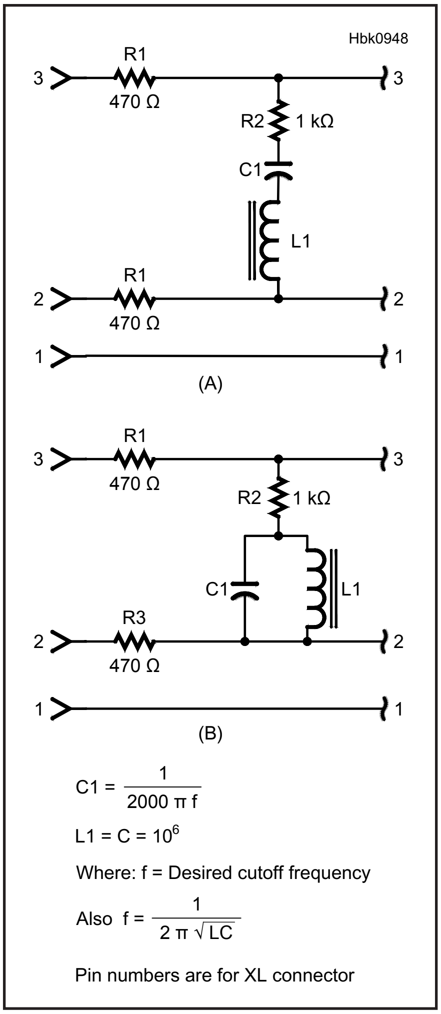
Exchanging the capacitors and resistor in Figure 13.51 results in a high-frequency loss (**Figure 13.52**), while adding an inductor allows adjusting the midrange response for boosting presence or controlling sibilance **(Figures 13.53A** and **13.53B).**

One problem with inductors is their inherent series resistance, which is inevitable due to the wire they're made from. The smaller (physically) a given inductor is built, the higher the resistance since the wire must be a smaller diameter. The Mouser 43LH and 43LJ series inductors minimize this by winding the wire around a ferrite core. At low currents the ferrite core does a great job of increasing the inductance without requiring as many turns of wire, which keeps the series resistance relatively low. Also, these coils are encapsulated in a rugged phenolic case which eliminates the danger of damaging the hair-thin wires during assembly.

For the midrange networks shown in this sidebar you can expect to alter the response by about 4 or 5 dB, though this can be varied by adjusting R2 up or down in value. With the high- and low-end rolloffs, the cutoff slope is 6 dB. per octave beyond the chosen frequency. This rolloff rate is hard to change because additional capacitors and resistors would be needed. That not only complicates the design, but also reduces the signal level overall.

CONSTRUCTION NOTES

The pads can be constructed in the usual metal enclosures. Note that the S3FM connector assembly made by Switchcraft — a 3.5-inch tube fitted with a male XL (or XLR) at one end and a female at the other — has enough room in the middle to build a mike pad, low-end rolloff network, polarity revers-



Figures 13.53 — Midrange cut (A) and boost (B) pads.

er, or any number of other useful little gadgets.

With space generally at a premium in any small enclosure, you must use the smallest components you can find. For microphone-level signals you can use ⅛-watt resistors. Another space saver is to use low-voltage capacitors. When the component values are large, such as the low-frequency rolloff, tantalum capacitors are the best choice and you may be able to find ultra-miniature types rated as low as 6 or even 3 V. Tantalums cost more than standard electrolytics, but they are smaller and generally higher quality.

Use shrink-fit sleeving over all of the components or otherwise ensure that the components don't short out to one another. The S3FM comes with a plastic-coated cardboard tube that lines the inside of the case, and this prevents any wires from touching the grounded case. Clear shrink tubing lets you see what parts are inside — a definite advantage if you have to take it apart for repair or just to see what's in there. Avoid black electrical tape for insulation; not only can't you see through it, but after a while it can turn into a sticky mess that's difficult to unwrap. Draw the schematic on a small piece of paper, and roll it up to serve as the outer insulation.

13.8.7 Speech Amplification and Processing

AUDIO SIGNAL LEVELS

The output level of most modern microphones ranges from about 10 mV to a few hundred mV when used close to the mouth. Microphone inputs are designed to accept signals in this range and are likely to be overloaded by stronger signals. In addition to microphone inputs, most ham equipment also features line-in inputs that are designed to accept signals between about 100 mV and 2 V.

IMPEDANCE IN AUDIO CIRCUITS

Since solid-state electronics became standard, audio circuits are no longer impedance-matched. Instead, audio circuits are voltage-matched. *Output* circuits have a low source impedance, typically around 100 Ω for pro line level, 300 Ω for consumer line level, and around 200 Ω for most modern microphones. Audio *input* circuits have a high input impedance, typically 10 kΩ for pro line level inputs, 50 kΩ for consumer line level inputs, and at least 1 MΩ for microphone inputs.

Thanks to these impedance relationships, very little current flows in microphone and line circuits. Output stages are constant voltage sources, so they can easily drive multiple inputs in parallel. While output circuits are usually *rated* for a 600 Ω load, loading them with 600 Ω degrades their performance.

Loudspeaker amplifiers are different, in that they *do* provide power. They are still constant voltage sources, and those that drive loudspeakers typically have output impedance of a few tens of milliohms, designed to drive loads in the range of 4-16 Ω. Since they are constant voltage sources, they will deliver four times more power into a 4-Ω load than into a 16-Ω load. The ratio of load impedance to source impedance is the *damping factor*.

Headphone amplifiers also supply power but a lot less of it. Headphones range in impedance from 8 Ω to more than 600 Ω, and most will be pretty loud with only a few volts drive. Most headphone amplifiers can drive more than one pair of headphones in parallel. Most headphone amplifiers have a resistor in series with their output so that they are not damaged when a plug is not inserted properly.

Computer sound cards have stereo *outputs* designed to drive headphones and line level inputs. The sound cards built into many laptop computers have only a mono *input*, often designed only for microphone levels. Most outboard sound cards, and better built-in sound cards, provide both a mono microphone and a stereo line input, or a single stereo input that can be switched to microphone or line level via the computer's operating system or the software that uses the audio signal.

SPEECH PROCESSING

Transmitters generally require audio signals in the volt region and it will be found advantageous to shape the frequency response prior to transmission. ("Amplified" microphones are outside this discussion.) Before being applied to the modulator circuits in the transmitter the audio signal is applied to a clipper or compressor for amplitude control. Such circuits along with their associated filtering increase the average value of signal level into the transmitter and so make the signal sound louder and define the occupied bandwidth.

PREAMPLIFIER

The output of the microphone is amplified and its frequency response shaped in a preamplifier as shown in **Figure 13.54.** R4 allows coarse gain adjustment and it can be a 100 kΩ potentiometer if various microphones are to be accommodated. The low-pass filter has a flat response out to about 3 kHz, including a peak of several dB at 2500 Hz (discussed below). The filters work in concert with the following treble peaking circuit to produce a response very suitable for communications as shown in **Figure 13.55.** The magnitude of the peak at 2500 Hz may be adjusted by simply changing the value of R12. This treble peaking circuit only affects speech signals in the 2500 Hz region.

The response as shown has a rise of 15 dB at 2500 Hz relative to 400 Hz but is nevertheless down 30 dB at 10 kHz. Very low frequencies have been rolled off. The output of the preamplifier is applied to a level control and then to a clipper or compressor for level control.

CLIPPER

After the speech signal has been increased in level and the basic response has been shaped, it may be applied to a clipper to increase its average volume level. At this point it must be pointed out that clipping must be used with great care if the transmitter is operating in the single-sideband mode. SSB transmitters cannot handle clipped waveforms gracefully. Clipping can always be used to catch occasional overshoots resulting from sluggish AGC systems, as an example, but in SSB systems should not be used as a routine method of volume maximization. However, clipping can be used in AM systems with great effect. A clipper followed by a low-pass filter sets the occupied bandwidth of the transmitter, assuming the following stages are operating cleanly.

Figure 13.56 shows the schematic of a speech clipper with its associated filtering. The "preclip" filters are optional but recommended. The "postclip" low-pass filter is a mandatory requirement to limit the transmitted bandwidth.

The 200 Hz high-pass filter ahead of the clipper greatly reduces low-frequency intermodulation distortion. The 3 kHz low-pass filter ahead of the clipper prevents high-frequency audio (sibilants) from being clipped and similarly causing high-frequency intermodulation distortion.

This circuit uses an over-driven op-amp as a clipper. Its sensitivity may be changed by changing the value of R8. The output is reduced by a factor of two (R11 and R12) so the following low-pass filter can handle the signal in a linear mode.

This filter has a sharp cutoff which would ordinarily have overshoots in its transient response when a clipped waveform is applied to its input. This can result in overmodulation, but it has been modified to have a step in its frequency response with the result that the overshoots have been turned into "undershoots" and so are rendered harmless. The overall response of the filtering in this block including the high-pass and both low-pass filters is shown in **Figure 13.57.**

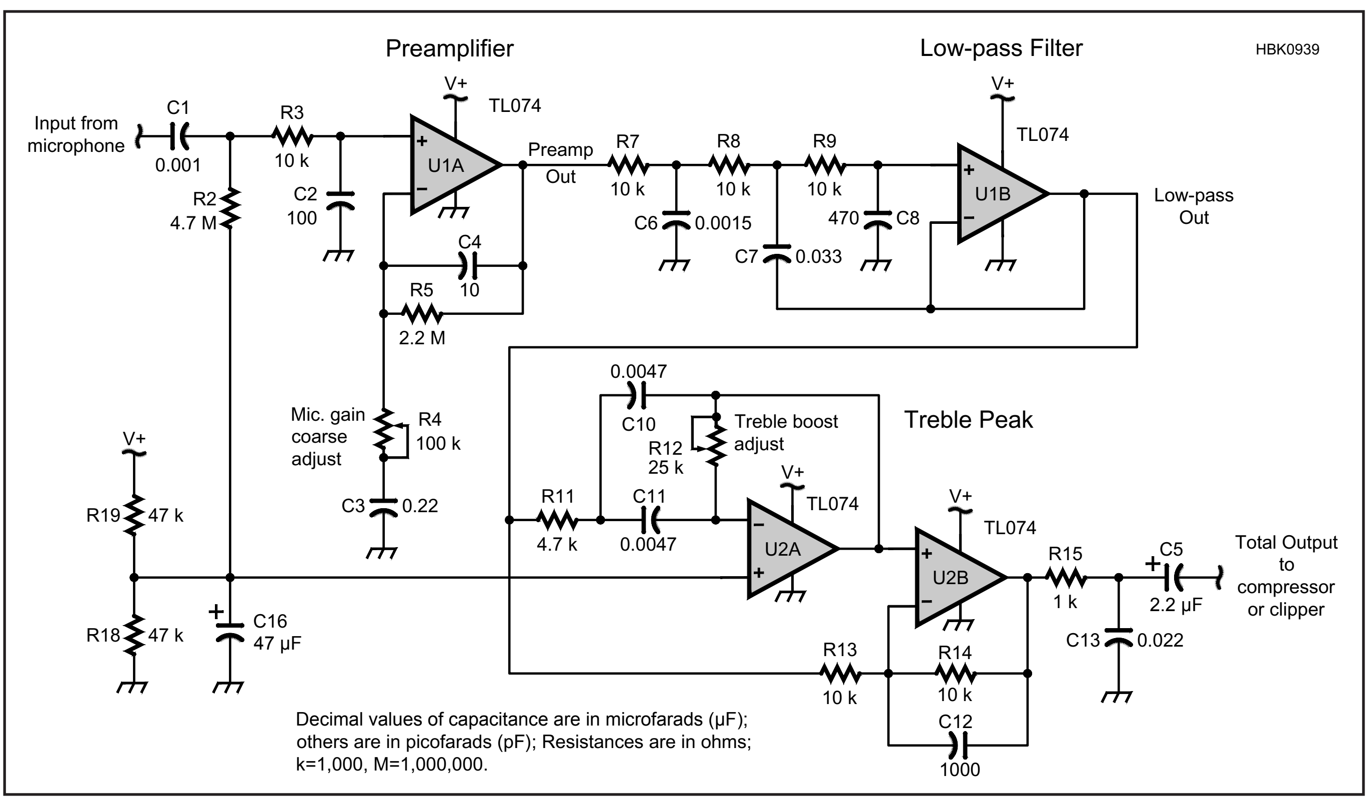


Figure 13.54 — Schematic of a microphone preamplifier with a gain stage, a low-pass filter and a treble peaking stage.

AUDIO COMPRESSOR

Another way to increase the average modulation level is to use automatic gain control (AGC) or a compressor. Such a circuit is shown in **Figure 13.58.** Compressors can be used for any mode of speech transmission. Compression should *never* be used with digital modes using AFSK modulation.

On the output of the op-amp are two LEDs in parallel but with opposite polarities. Audio voltages of either polarity illuminate a photoconductor causing its resistance to decrease. The photoconductor will then adjust the gain of the associated op-amp circuit. As audio levels increase, the LEDs illuminate the photoconductor more, reducing its resistance. This causes the photoconductors to reduce the circuit gain, maintaining a constant audio output level. A second photoconductor monitors the illumination and so shows the degree of compression. The photoconductors (Luna PDV-P8101 or equivalent) are mounted directly on top of the white LEDs (Cree C503D-WAN-CCBEB152 or equivalent). This is shown in the inset for Figure 13.58.

This circuit responds to signal level increases within 2 or 3 milliseconds. After a transient, it increases the gain back to normal in less than 100 milliseconds. The result is a very high average modulation level with far lower distortion than a clipper. Some signal overshoots will escape while gain reduction is underway but time-wise, on a percentage basis, they are quite small. If this block of circuitry is followed by a clipper to catch the overshoots then even those small overshoots

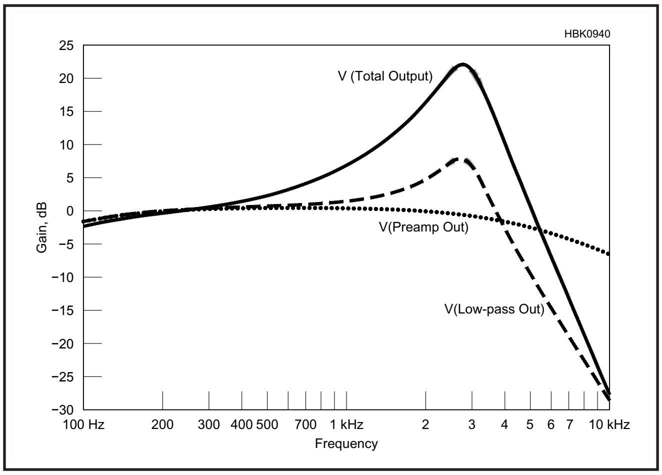


Figure 13.55 — Frequency response at various points within the speech amplifier.

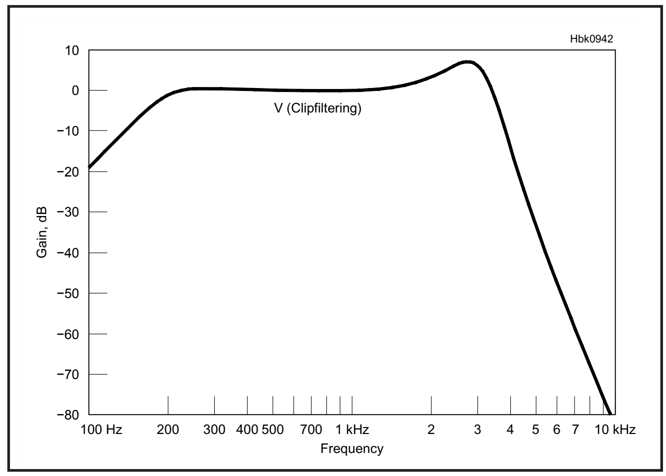


Figure 13.57 — Overall response of the clipper-filtering block.

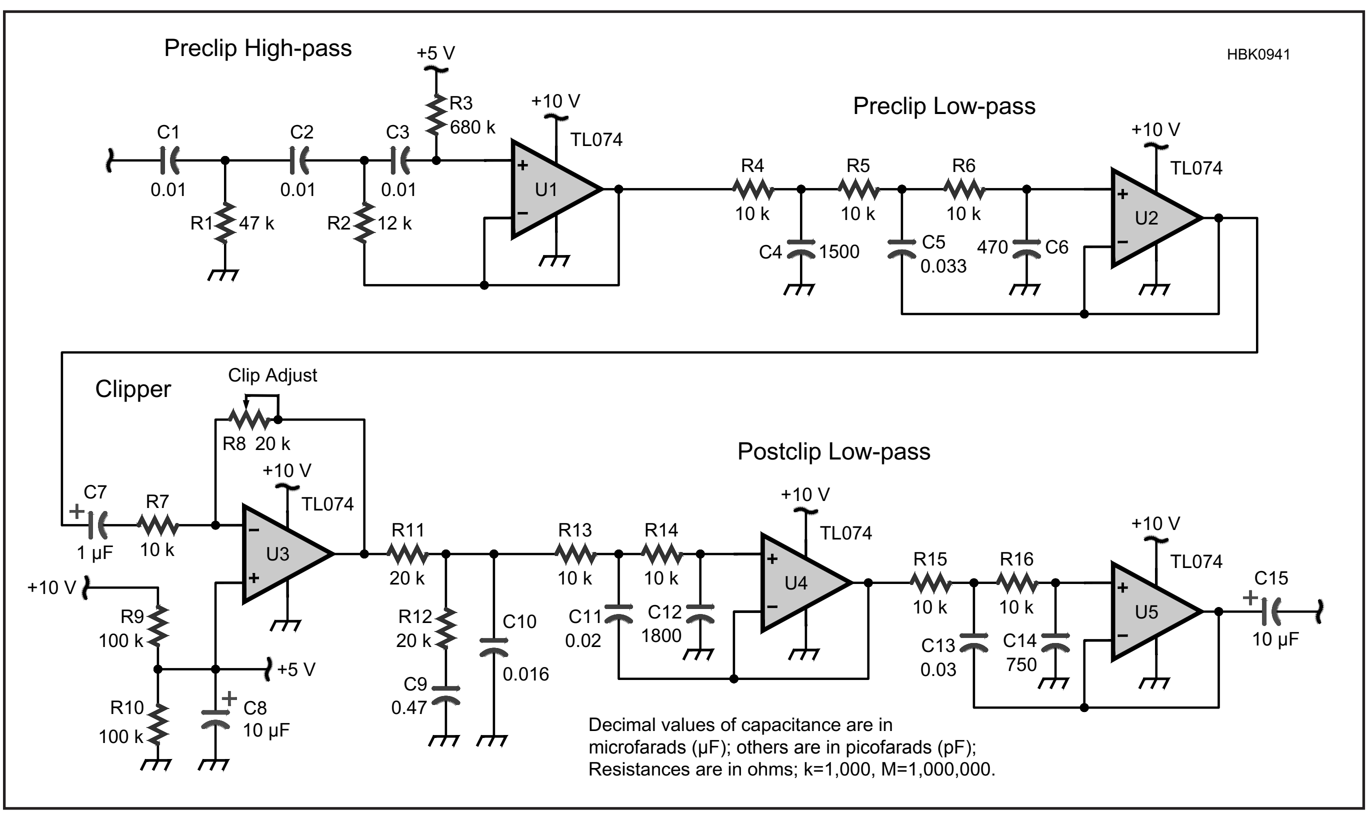


Figure 13.56 — Schematic of the speech clipper and its associated filtering.

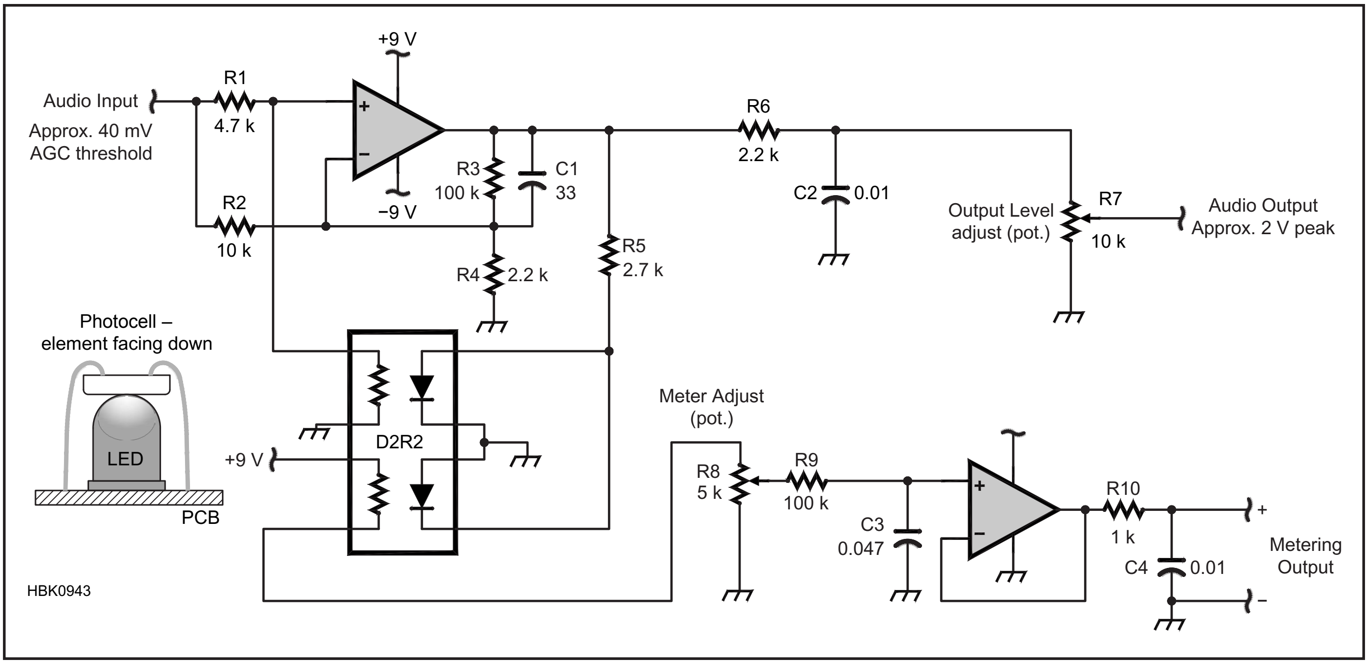


Figure 13.58 — Compressor circuit based on LEDs shining on photoconductors (cadmium sulfide photocells). See the text for a description of how to fabricate D2R2.

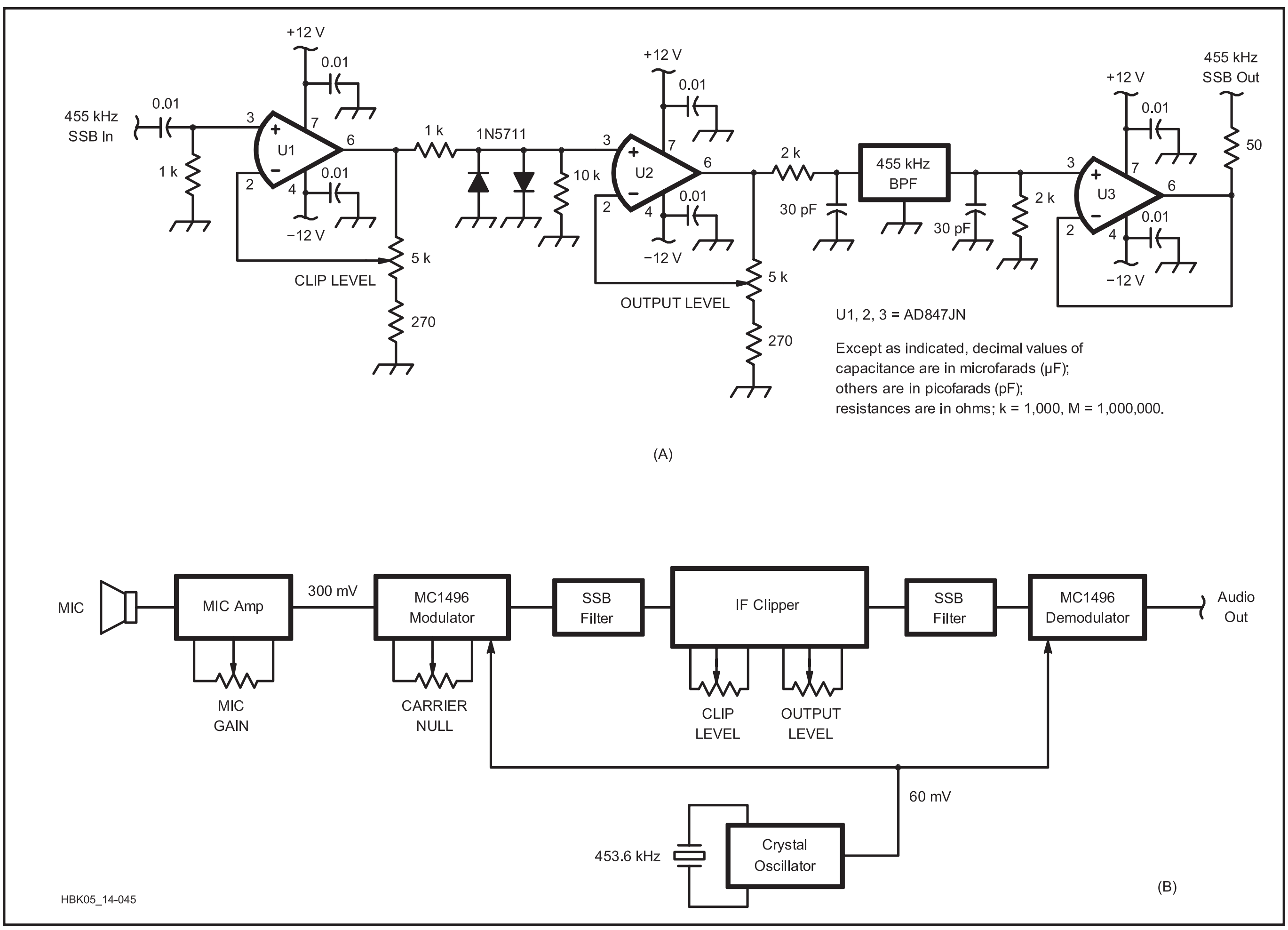


Figure 13.59 — IF speech clipping. At (A), schematic diagram of a 455 kHz IF clipper using high-frequency op amps. At (B) block diagram of an adaptation of the above system to an audio in-audio out configuration.

will be of no concern.

The pre-clip high-pass and pre-clip low-pass filters shown in the clipper block may be used in the compressor; they were not shown here. If used, they must be on the input to the compressor.

IF SPEECH CLIPPER

Audio clipper speech processors can generate a considerable amount of in-band harmonics and IMD (involving different simultaneously occurring speech frequencies). The total distortion detracts somewhat from speech intelligibility. IF clippers (also known as RF speech processors) overcome most of these problems, especially the Hilbert Transform problem. (See Sabin and Schoenike in the References section.)

Figure 13.59A is a schematic diagram of a 455 kHz IF clipper using high-frequency op-amps. 20 dB of gain precedes the diode clippers. A second amplifier establishes the desired output level. The clipping produces a wide band of IMD products close to the IF frequency. Harmonics of the IF frequency are easily rejected by subsequent selectivity. "Close-in" IMD distortion products are band-limited by the 2.5 kHz wide IF filter so that out-of-band splatter is eliminated. The in-band IMD products are at least 10 dB below the speech tones.

Figure 13.59B shows a block diagram of an adaptation of the above system to an audio in-audio out configuration that can be inserted into the mic input of any transmitter to provide the benefits of RF speech processing. These are sometimes offered as aftermarket accessories.

Figure 13.60 shows oscilloscope pictures of an IF clipped two-tone signal at various levels of clipping. The level of clipping in a radio can be estimated by comparing with these photos. Listening tests verify that the IMD does not sound nearly as bad as harmonic distortion. In fact, processed speech sounds relatively clean and crisp. Tests also verify that speech intelligibility in a noise background is improved by 8 dB. (See the article on RF clippers by Sabin in the References section.)

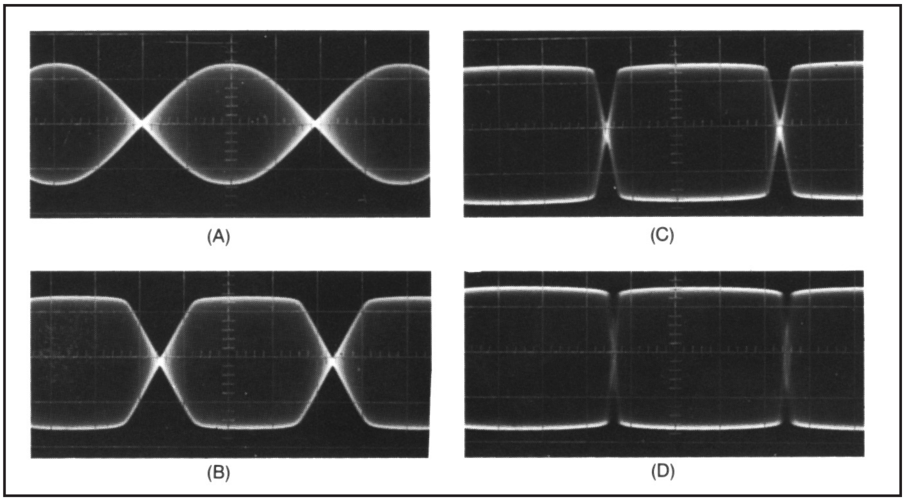


Figure 13.60 — Two-tone envelope patterns with various degrees of RF clipping. All envelope patterns are formed using tones of 600 and 1000 Hz. At A, clipping threshold; B, 5 dB of clipping; C, 10 dB of clipping; D, 15 dB of clipping.

The repeaking effect from band-pass filtering the clipped IF signal occurs, and must be accounted for when adjusting the output level. A two-tone audio test signal or a speech signal should be used. The ALC circuitry (discussed later) will reduce the IF gain to prevent splattering in the power amplifiers. If the IF filter is of high quality and if subsequent amplifiers are clean, the transmitted signal is of very high quality and is very effective in noisy situations and often also in pile-ups.

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

An electrically identical approach to the IF clipper can be achieved at audio frequencies. If the audio signal is translated to, say 455 kHz, processed as described and translated back to audio, all the desirable effects of IF clipping are retained. This output then plugs into the transmitter's microphone jack. Figure 13.59B shows the basic method. The mic amplifier and the MC1496 circuits have been previously shown and the clipper circuit can be the same as in Figure 13.59A.

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

13.9 Managing Computer Audio

13.9.1 Computer Audio

DIGITAL DYNAMIC RANGE

The number of bits in a digital system (called the *bit depth*) sets the limit for its maximum dynamic range. An ideal 16-bit system would provide 96 dB of dynamic range (the difference between the highest and lowest instantaneous voltage that it can accurately reproduce). The dynamic range of real products is a few dB less. Each additional bit doubles the range of voltage that can be reproduced, and double (or half) the voltage is a change of 6 dB. Thus, a 12-bit system should be capable of a few dB less than 72 dB dynamic range.

DATA RATES

The *bit rate* of a digital system, expressed in kilobits/sec (kbps), sets the limit for its audio bandwidth, which is 90% of half the numerical value, in kHz. Thus, a 48 kHz system can provide about 22 kHz audio bandwidth. When a sound card is used as part of a spectrum display, the maximum displayed bandwidth in kHz is equal to the maximum bit rate in kbps. Most computer sound cards and sound recording software can operate at standard sample rates and bit depth less than their maximum value, and most software that uses a computer sound system can vary these settings either automatically or as desired by the user.

AUDIO DATA COMPRESSION

The size of a sound recording file depends both on its bit depth and bit rate. A 12-bit sound file recorded at 12 kbps provides audio bandwidth of about 5 kHz with about 70 dB of dynamic range and is a good choice for use in amateur radio. Uncompressed sound files (such as AIFF and WAV) require about 16 kB on a hard drive for each second of recording time for each channel of audio recorded. WAV files are the standard for CD-ROM recordings at 16-bit sample width and 44.1 kHz sample rate.

Digital audio signals can be compressed to reduce both the bandwidth needed to transmit them and the size of files needed to contain them. *Lossless compression* (such as ALAC, FLAC, and ZIP) reduces file size somewhat without losing any of the sound information. Lossless systems provide an exact copy of the original. ZIP provides very little compression of uncompressed sound files, but ALAC and FLAC can reduce file size by about one half.

Lossy compression systems like AAC, MP3, and OGG use algorithms to approximate the audio by leaving out parts of the waveform that the ear/brain is unlikely to miss. On playback, the system guesses at what parts were left out and adds them back in. Such systems are referred to as lossy because they do not provide a perfect copy of the original digital signal. These algorithms may compromise digital mode modulation and demodulation.

While file compression can save space on a hard drive, it takes time for the compression and de-compression algorithms to operate. MP3 compression is a great choice for recording a QSO, but a poor choice for recording messages for playback during a contest due to the delay. MP3 compression can reduce file size (and transmission bandwidth) by 75-95% (that is, files and bandwidth between 5% and 25% of their uncompressed size). Since contest messages are short, a lossless format is recommended.

USB AUDIO

It is increasingly common for transceivers to have a USB port for a PC CAT interface and audio input and output. A radio's USB interface is usually a Silicon Labs or FTDI IC. Driver software for the PC to use the radio interface will be available from the radio manufacturer, from the IC supplier, or from the developer of the PC software.

Audio can be transferred over the USB interface as a digital data stream called a USB Audio Codec. This digital interface is recognized by computers and other devices that support the USB standard. Your computer software will treat the USB Audio Codec as an audio playback or recording device for which you select and adjust levels the same as speakers or microphones.

A transceiver's USB interface will use the same data rates (up to 115 kbaud or higher) for PC control and audio transfer so use the highest reliable data rate available to minimize audio latency or dropouts. The radio will also have adjustable input and output levels for USB audio.

USB sound cards are standalone devices that have a standard analog audio interface and a USB port that connects to a host PC. USB sound cards designed for semi-pro audio users and even for gamers are usually good performers. They often perform significantly better than the sound cards built into computers, especially laptops. An inferior quality sound card can degrade the decoding capability of software for digital modes. Some higher-end units have LEDs to indicate safe signal levels — for example, green for good levels, red for overload (also called "digital clip").

Using a USB audio interface can eliminate audio cables between a PC and transceiver. This also reduces the possibility of RFI although USB interfaces can be susceptible to RFI. Use high-quality, shielded USB cables to minimize RFI.

13.9.2 Setting Digital Mode Levels

Many digital modes use PC software to perform audio encoding and decoding. If a USB audio codec is not used, the analog audio is then routed through a sound card to the transceiver's microphone or data audio input on an accessory connector. For USB audio, follow the transceiver and software level adjust instructions. The rest of this section assumes that analog audio is being used.

The transceiver uses the input audio to perform AFSK modulation of a SSB or FM signal. To avoid distorting the modulated audio and generating spurious products in the audio and on adjacent RF channels, it is important to use the right levels for the transceiver's audio input. Similarly, the decoding software in the PC will work best with undistorted audio from the transceiver that is well above the noise floor.

Remember that sound card inputs and outputs will be relatively low-level analog signals and shielded cables are required. Be sure that the shields are properly connected to the outside of the transceiver's metal enclosure. Do not connect the shield to a microphone input ground pin or directly to a circuit board inside the radio. RF current on the shield of these cables will cause RFI if given a path to the internal circuitry. In addition, PC and transceiver enclosures should be bonded to minimize any voltage between them that could result in RFI and noise in the audio circuits. If you do experience RFI, use a ferrite core of the right mix for the frequency range involved. See the **RFI and EMC** chapter for more about managing RF in your station.

SETTING SOUND CARD INPUT LEVEL

This is important to achieving maximum signal-to-noise performance. For sound cards with a digital clip indicator, turn up the input gain until the red light flashes with the loudest signals, then reduce it slightly so that it never flashes. Then set the digital gain in the encode/decode software as directed by the manual.

SETTING SOUND CARD OUTPUT LEVEL

It is just as important to avoid transmitting a distorted or overmodulated signal. There are three steps, all of which are important: (1) making the output of the computer clean, (2) not overdriving the radio's audio input stage, and (3) setting the audio input gain in the radio. There are (at least) four good ways to set output level from the computer, depending on what test equipment is available.

The first method uses an oscilloscope connected to the output of the sound card that feeds the radio. Set the sweep so that you see clearly defined sine waves that make up the signal from the digital program (RTTY, PSK, WSJT modes, and so on) in transmit mode. Because these modulation schemes include multiple frequencies, you probably won't get the display to sync. Increase the output level until you see squaring ("flat-topping") at the top of the sine waves or "spiky" digital distortion, then reduce the output level by half the voltage.

The second method uses an audio spectrum analyzer connected to the output of the sound card. With the digital program in transmit mode and output level set fairly low, note the spectrum lines. With JT65, you should see only the tones that you would see on the WSJT-X or JT65-HF display, roughly 200 Hz wide. With RTTY, you should see only two tones spaced by 170 Hz. With JT9 or PSK, you should only see a single tone. Now, increase levels until you begin to see additional lines spread out from the normal tones. These additional lines are distortion products and will cause interference. Note the *difference* in strength (in dB) between the signal tones and the distortion, then gradually increase output level until you see the difference become smaller (which indicates that the percentage distortion is rising). Now, reduce the output level until the difference is larger (the distortion is much less). With proper adjustment, the distortion products should be at least 40 dB below the tones.

The third method uses an audio voltmeter connected to the output of the sound card. With the digital program in transmit mode, start with very low output, and gradually increase it until you no longer see voltage increasing, then reduce the output level to one-half the maximum measured voltage.

The fourth method uses your ears and headphones connected to the output of the sound card. Start with the output of the sound card set low, gradually increasing it until you hear harshness or sharpness in the tones. That harshness is the distortion products. Now, reduce the output of the sound card until the tones sound half as loud. This works because a change of 6 – 10 dB is perceived as half (or twice) as loud.

With all of these methods, adjust the input gain control of your radio according to the user manual for the radio for transmitting digital modes. Always make sure that any processing in the radio is turned off.

As a final check, ask another amateur who is receiving your signal fairly well to look carefully for sidebands in the received waterfall display (first making sure that the noise blanker is turned off, and that you are not overloading the receiver).

CONNECTING COMPUTER AUDIO TO THE RADIO

To avoid overdriving the radio's audio input, feed the computer sound card to the radio's LINE IN input (if there is one). In older equipment, this may be labeled as a phone patch input, or for use with a hardware RTTY interface. It may be available on an accessory connector.

If the radio has no line input, you'll need to feed the microphone input through a simple voltage divider (often called a "pad" — see the previous section in this chapter) so that you don't overload the microphone preamp. All it takes is two resistors, one in series with the audio path, and one in parallel with the input of the radio. 20 dB (a 10:1 voltage divider) of attenuation should be enough for most radios and calls for a 10:1 ratio between the two resistors. The values are not critical, but 10 kΩ for the series resistor and 1 kΩ for the parallel resistor, or 4.7 kΩ and 470 Ω are good choices. Low-wattage resistors may be used here and can fit within the connectors of the audio cable between the computer and the radio if you use the right connector. RCA and ⅛-inch connectors made by Switchcraft have the most space inside them. Neutrik is also a good brand.

13.10 Voice Operation

13.10.1 Push-To-Talk for Voice

Another advance in amateur station switching followed longstanding practices of aircraft and mobile voice operators who had other things to contend with besides radio switches. Microphones in those services included built-in switches to activate TR switching. Called push-to-talk (PTT), this function is perhaps the most self-explanatory description in our acronym studded environment.

Figure 13.61 — A classic Astatic D-104 mic with PTT stand.

Relays controlled the various switching functions when the operator pressed the PTT switch. Some top-of-the-line transmitters of the period included at least some of the relays internally and had a socket designed for PTT microphones. **Figure 13.61** is a view of the ubiquitous Astatic D-104 microphone with PTT stand, produced from the 1930s to 2004, and still popular at flea markets and auction sites. PTT operation allowed the operator to be out of reach of the radio equipment while operating, permitting "easy chair" operation for the first time.

Modern transceivers include some form of PTT (or "one switch operation"). Relays, diodes, transistors and other components seamlessly handle myriad transmit-receive changeover functions inside the transceiver. Most transceivers have additional provisions for manually activating PTT via a front-panel switch. And many have one or more jacks for external PTT control via foot switches, computer interfaces or other devices.

13.10.2 Voice-Operated Transmit-Receive Switching (VOX)

How about break-in for voice operators? SSB operation enabled the development of voice operated transmit/receive switching, or VOX. During VOX operation, speaking into the microphone causes the station to switch from receive to transmit; a pause in speaking results in switching back to receive mode. Although VOX technology can work with AM or FM, rapidly turning the carrier signal on and off to follow speech does not provide the smooth operation possible with SSB. (During SSB transmission, no carrier or signal is sent while the operator is silent.)

VOX OPERATION

VOX is built into current HF SSB transceivers. In most, but not all, cases they also provide for PTT operation, with switches or menu settings to switch among the various control methods. Some operators prefer VOX, some prefer PTT and some switch back and forth depending on the operating environment.

VOX controls are often considered to be in the "set and forget" category and thus may be controlled by a software menu or by controls on the rear panel, under the top lid or behind an access panel. The following sections discuss the operation and adjustment of radio controls associated with VOX operation. Check your transceiver's operating manual for the specifics for your radio.

Before adjusting your radio's VOX controls, it's important to understand how your particular mic operates. If it has no PTT switch, you can go on to the next section! Some mics with PTT switches turn off the audio signal if the PTT switch is released, while some just open the control contacts. If your mic does the former, you will need to lock the PTT switch closed, have a different mic for VOX, or possibly modify the internal mic connections to make it operate with the VOX. If no audio is provided to the VOX control circuit, it will never activate. If the mic came with your radio, or from its manufacturer, you can probably find out in the radio or mic manual.

VOX Gain

Figure 13.62 shows some typical transceiver VOX controls. The VOX gain setting determines how loud speech must be to initiate switchover, called "tripping the VOX." With a dummy load on the radio, experiment with the setting and see what happens. You should be able to advance it so far that it

Figure 13.62 — The function of VOX controls is described in the text. They require adjustment for different types of operating, so front-panel knobs make the most convenient control arrangement. In some radios, VOX settings are adjusted through the menu system.

switches with your breathing. That is obviously too sensitive or you will have to hold your breath while receiving! If not sensitive enough, it may cause the transmitter to switch off during softly spoken syllables. Notice that the setting depends on how close you are to the microphone, as well as how loud you talk. A headset-type microphone (a "boom set") has an advantage here in that you can set the microphone distance the same every time you use it.

The optimum setting is one that switches to transmit whenever you start talking, but isn't so sensitive that it switches when the microphone picks up other sounds, such as a cooling fan turning on or normal household noises.

VOX Delay

As soon as you stop talking, the radio can switch back to receive. Generally, if that happens too quickly, it will switch back and forth between syllables, causing a lot of extra and distracting relay clatter. The VOX delay control determines how long the radio stays in the transmit position once you stop talking. If set too short, it can be annoying. If set too long, you may find that you miss a response to a question because the other station started talking while you were still waiting to switch over.

You may find that different delay settings work well for different types of operation. For example, in a contest the responses come quickly and a short delay is good. For casual conversation, longer delays may be appropriate. Again, experiment with these settings with your radio connected to a dummy load.

Anti-VOX

This is a control with a name that may mystify you at first glance! While you are receiving, your loudspeaker is also talking to your microphone — and tripping your VOX — even if you aren't! Early VOX users often needed to use headphones to avoid this problem. Someone finally figured out that if a sample of the speaker's audio signal were fed back to the mic input, out-of-phase and at the appropriate amplitude, the signal from the speaker could be cancelled out and would not cause the VOX circuit to activate the transmitter. The ANTI-VOX (called ANTI-TRIP in the photo) controls the amplitude of the sampled speaker audio, while the phase is set by the transceiver design.

As you tune in signals on your receiver with the audio output going to the speaker, you may find that the VOX triggers from time to time. This will depend on how far you turn up the volume, which way the speaker is pointed and how far it is from the mic. You should be able to set the anti-VOX so that the speaker doesn't trip the VOX during normal operation.

Generally, setting anti-VOX to higher values allows the speaker audio to be louder without activating the VOX circuit. Keep in mind that once you find a good setting, it may need to be changed if you relocate your microphone or speaker. With most radios, you should find a spot to set the speaker, microphone and anti-VOX so that the speaker can be used without difficulty.

13.11 Transmitter Power Stages

The functions described so far that process input data and information and result in a signal on the desired output radio frequency generally occur at a low level. The one exception is full-carrier AM, in which the modulation is classically applied to the final amplification stage. More modern linear transmitter systems generate AM in the same way as SSB at low levels, typically between 1 mW and 1 W.

13.11.1 Types of Power Amplifiers

The **RF Power Amplifiers** chapter provides a detailed view of power amplifiers; however, we will take a quick peek here to set the stage for the following discussions. Amplifiers use dc power applied to active devices in order to increase the power or level of signals. As will all real devices, they introduce some distortion in the process, and are generally limited by the level of distortion products. Power amplifiers can be constructed using either solid-state devices or vacuum tubes as the active device. At higher powers, typically above a few hundred watts, vacuum tubes are more frequently found, although there is a clear trend toward solid state at all amateur power levels.

Independent of the device, amplifiers are divided into classes based on the fraction of the input cycle over which they conduct. A sinusoidal output signal is provided either by the *flywheel* action of a resonant circuit or by other devices contributing in turn. The usual amplifier classes are summarized in **Table 13.1**. Moving from Class A toward Class C, the amplifiers become progressively less linear but more efficient. The amplifiers with a YES in the LINEAR column thus are not all equally linear however A, AB or Class B amplifiers can be suitable for operation in a linear transmitter chain. Class C amplifiers can be used only for amplification of signals that do not have modulation information contained in the amplitude, other than on-off keyed signals. Thus class C amplifiers are useful for amplification of sinusoids, CW, FM, or as the nonlinear stage at which high-level AM modulation is employed.

Recent developments in switching-type amplifiers and in single-band matching network design have created several additional classes. Class D is a switchmode amplifier most often used for high-efficiency audio amplification. Class E and F use tuned output networks that let the amplifying device act like a switch but prevent high voltage and high current at the same time. Class G is similar to a Class B amplifier, but switches between two voltage levels to reduce power dissipation at low signal levels. Class I uses two devices driven with complementary pulse duty cycles to cancel harmonics and follow the input waveform. Class S is a variation on Class D, and Class T uses DSP to optimize pulse widths in a Class D amplifier for better performance. See the references at the end of this chapter for Rosu and for Silver.

Table 13.1
Characteristics of Transceiver Power Amplifier Classes
Values are typical

Class	*Conduction*	*Linear*	*Efficiency*
A	360°	Yes	30%
AB	270°	Yes	55%
B	180°	Yes	65%
C	90°	No	74%

13.11.2 Linear Amplifiers

While transmitters at power levels of 1 mW to 1 W have been successfully used for communication across many portions of the spectrum, most communications systems operate with more success at higher powers. The low level stage is usually referred to as an exciter, while higher power is provided by one or more linear amplifier stages as shown in **Figure 13.63**.

The power levels shown at the various points in Figure 13.63 are fairly typical for a high powered amateur station. The 1500 W PEP output represents the legal limit for US amateurs in most bands (200 W PEP on 30 meters and 100 W ERPD on the 60 meter channels are notable exceptions). The first amplifier block may contain more than one stage, while the final output amplifier is often composed of multiple parallel active devices.

Typical power supply requirements for the amplifier stages are noted for a number of reasons. First, while power is rarely an issue at the exciter level, often it is a significant issue at the power levels shown for the amplifiers. The power supplies represent a large portion of the cost and weight of the system as the power increases. Some manufacturers are beginning to use switching-type power supplies for high-power amplifiers, resulting in a major reduction in size and weight.

Note also that a gross amplifier efficiency of about 50% is assumed for the amplifiers, taking into account ancillary subsystems as well as the inefficiency of the active devices in linear mode. The 50% that doesn't result in actual RF output is radiated as heat from the amplifier and must be removed from the amplifier as it is generated to avoid component damage. This represents another cost and weight factor that increases rapidly with power level.

The voltages shown for the supplies are those typical of modern solid state amplifiers. While virtually all commercial equipment now includes solid state amplifiers at the 100 W level, vacuum tube active devices are frequently found at higher levels, although the trend is clearly moving toward solid state. Vacuum tube amplifiers typically operate at voltages in the 2 to 4 kV range, requiring stringent measures be taken to avoid arcing across components. In addition, vacuum tube amplifiers typically dissipate up to 100 W of filament power that must be added to the power supply and heat dissipation planning.

13.11.3 Nonlinear Amplifiers

Nonlinear transmitters are somewhat different in architecture than the linear systems discussed previously. The configuration of a high-level AM modulated transmitter is shown in **Figure 13.64**. Note that none of the upper RF stages (the "RF chain") need to be particularly linear. The final stage must be nonlinear to have the modulation applied. Thus the RF stages can be the more power-efficient Class C amplifiers if desired.

There are some observations to be made here. Note that the RF chain is putting out the full carrier power whenever in transmit mode, requiring a 100% duty cycle for power and amplifier components, unlike the SSB systems discussed previously. This imposes a considerable weight and cost burden on the power supply system. Note also that the PEP output of a 100% modulated AM system is equal to four times the carrier power.

The typical arrangement to increase the power of such a system is to add not only an RF amplifier stage capable of handling the desired power, but also to add additional audio power amplification to fully modulate the final RF stage. For 100% high-level plate modulation, an audio power equal to half the dc input power (plate voltage times plate current of a vacuum tube amplifier) needs to be provided. This arrangement is shown in **Figure 13.65**. In the example shown, the lower level audio stages are provided by those of the previous 50 W transmitter, now serving as an exciter for the power amplifier and as a driver for the modulating stage. This was frequently provided for in some transmitters of the AM era, notably the popular E. F. Johnson Ranger series, which provided special taps on its modulation transformer for use as a driver for higher-power systems.

It is worth mentioning that in those days the FCC US amateur power limit was expressed in terms of dc *input* to the final stage and was limited to 1000 W, rather than the 1500 W PEP *output* now specified. A fully modulated 1000 W dc input AM transmitter would likely have a carrier output of 750 W or 3000 W PEP — 3 dB above our current limit. If you end up with that classic Collins KW-1 transmitter, throttle it back to make it last and stay out of trouble!

13.11.4 Hybrid Amplifiers

Another alternative that is convenient with current equipment is to use an AM transmitter

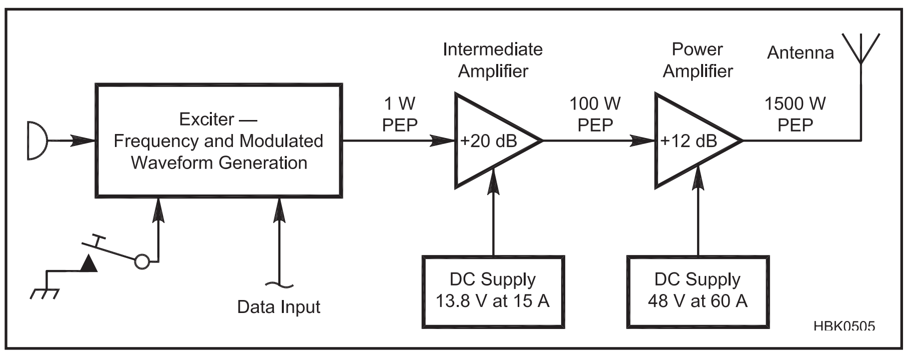


Figure 13.63 — Block diagram of a solid-state linear transmitter chain with multiple amplifier stages.

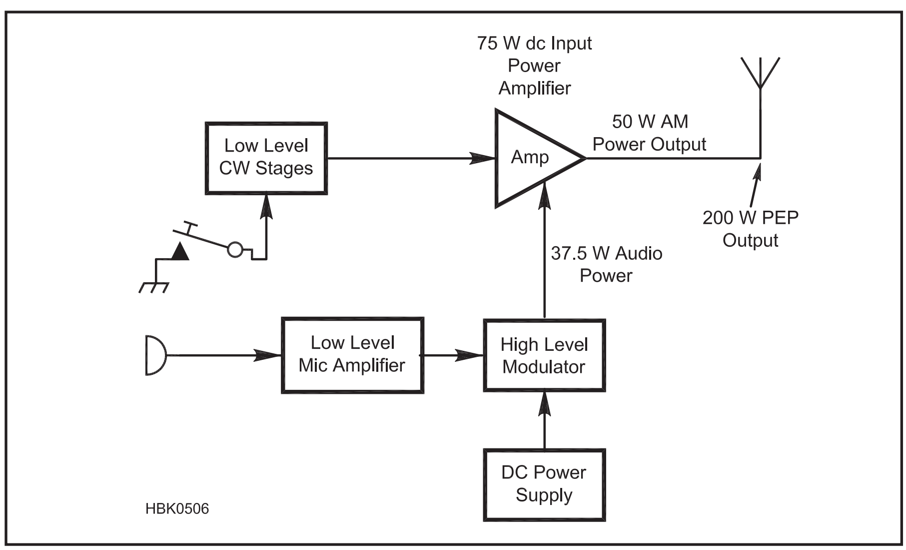


Figure 13.64 — Block diagram of a high level AM modulated transmitter.

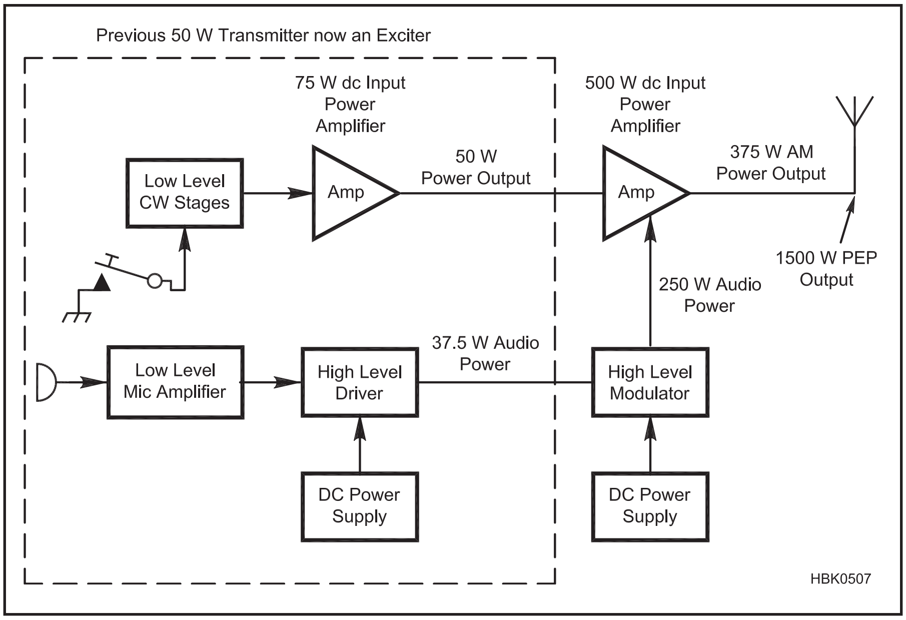


Figure 13.65 — Block diagram of a high level AM modulated transmitter with added output stage.

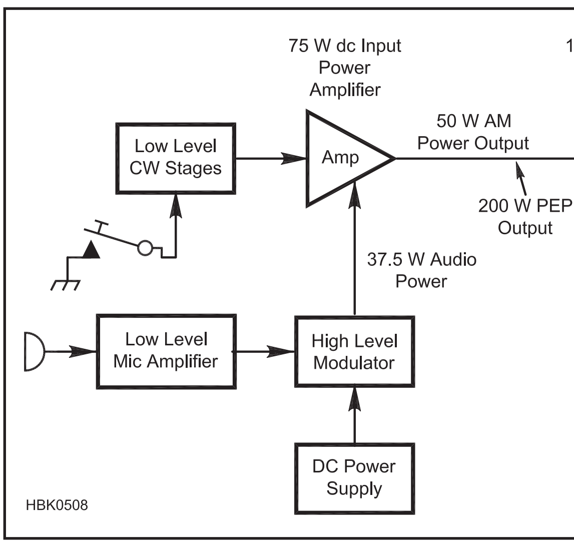


Figure 13.66 — Block diagram of a hybrid non-linear/linear AM transmission system.

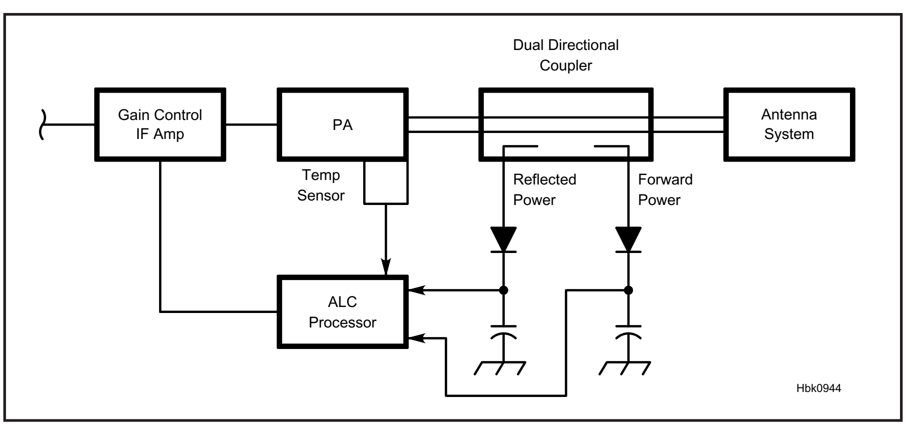


Figure 13.67 — An ALC protection method for a solid-state transmitter using a directional coupler to sense power level.

with a linear amplifier. This can be successful if the relationship that PEP = 4 × Carrier Power is maintained. **Figure 13.66** shows a 1500 W PEP output linear amplifier following a typical 50 W AM transmitter. In this example, t he amplifier would be adjusted to provide a 375 W carrier output with no modulation applied to the exciter. During voice peaks the output seen on a special PEP meter, or using an oscilloscope, should be 1500 W PEP.

Note that during AM operation, the amplifier is producing a higher average power than it would without the carrier being present, as in SSB mode. The duty cycle specification of the amplifier should be checked to be sure it can handle the heavier load. If the amplifier has an RTTY rating, it should be safe to run an AM carrier at 66% of the RTTY output, following the required on and off time intervals.

13.11.5 Automatic Level Control (ALC)

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

Note that ALC should be minimally active with most digital mode transmissions. The modulation of these signals requires linear amplification to preserve the waveform shape and minimize distortion products. ALC action creates distortion as it alters the power level of the signal. Adjust the radio drive levels so that the ALC is at its minimum level of activity — usually shown as the lower bar of a multi-segment LCD meter or a needle position just above zero. (The same caution applies to any form of audio or speech processing if the digital signal is generated by audio tones applied to the transmitter's microphone input.)

Figure 13.67 shows how a dual directional coupler can be used to provide ALC for a solid-state power amplifier (PA). The basic idea is to protect the PA transistors from excessive SWR and dissipation by monitoring both the forward power and the reflected power.

13.11.6 Transmit-Receive (TR) Switching

As the complexity of a transceiver increases, switching between receive and transmit becomes quite complex. In commercially built equipment, this function is usually controlled by a microprocessor that manages any necessary sequencing and interlock functions that would require an excessive amount of circuitry to implement with discrete components. For an example of just how complex TR switching could be in an advanced transceiver, look at the schematic for any modern mid-level or top-of-the-line solid-state transceiver.

Nevertheless, the basic functions of TR switching are well within scope for the amateur building a transceiver. Understanding TR switching will also assist in troubleshooting a more complex commercial radio.

QRP TR SWITCHING

Numerous schemes are popular for switching an antenna between transmitter and receiver functions. But these schemes tend to get in the way when one is developing both simple receivers and low-power transmitters, perhaps as separate projects. A simple relay-based TR scheme is then preferred and is presented here. In this system, used in the MkII Updated Universal QRP Transmitter by Wes Hayward, W7ZOI (see the full article in this book's online supplemental information), the TR relay not only switches the antenna from the receiver to the transmitter, but disconnects the headphones from the receiver and attaches them to a sidetone oscillator that is keyed with the transmitter.

The circuitry that does most of the switching is shown in **Figure 13.68**. A key closure discharges capacitor C1. R2, the 1 kΩ resistor in series with C1, limits the discharge current. Key closure causes Q6 to saturate, causing Q7 to also saturate, turning the relay on. The relay picked for this example has a 700 Ω, 12 V coil with a measured 4 ms pull-in time.

If full break-in TR switching is required, a high-speed reed relay can be used. KE2QJ provided circuits that can be adapted for internal use in a home-built transceiver, although the original purpose was to integrate a stand-alone receiver with a transceiver and linear amplifier. His full article is available in this book's downloadable supplemental information. The References also include an excellent online paper by W8ZR on adding a high-speed vacuum relay QSK switch for full legal-limit QSK operation.

If you already have a receiver and transmitter and want to integrate them under the control of a separate TR switch, the K8IQY "Magic Box" (**www.4sqrp.com/MagicBox.php**) incorporates a number of useful features. This is a microprocessor-controlled design that can handle up to 10 W of transmitter power, switches at up to 50 WPM, and includes an audio sidetone output, as well. Although the kit is no longer offered, complete documentation is available online, including schematics and design information. The design could be extended to handle more transmit power with heavier components and the appropriate circuit changes.

AMPLIFIER-TRANSCEIVER TR SWITCHING

Amateur transceivers intended for use with external amplifiers have a KEY OUT output. This is usually a contact closure while in transmit mode intended to connect to a corresponding KEY IN input on the external amplifier. Check the transceiver and amplifier manuals to find out what they are called on your equipment. A diagram of the proper cabling to connect the transceiver and amplifier will be provided in the manual. Amplifiers may

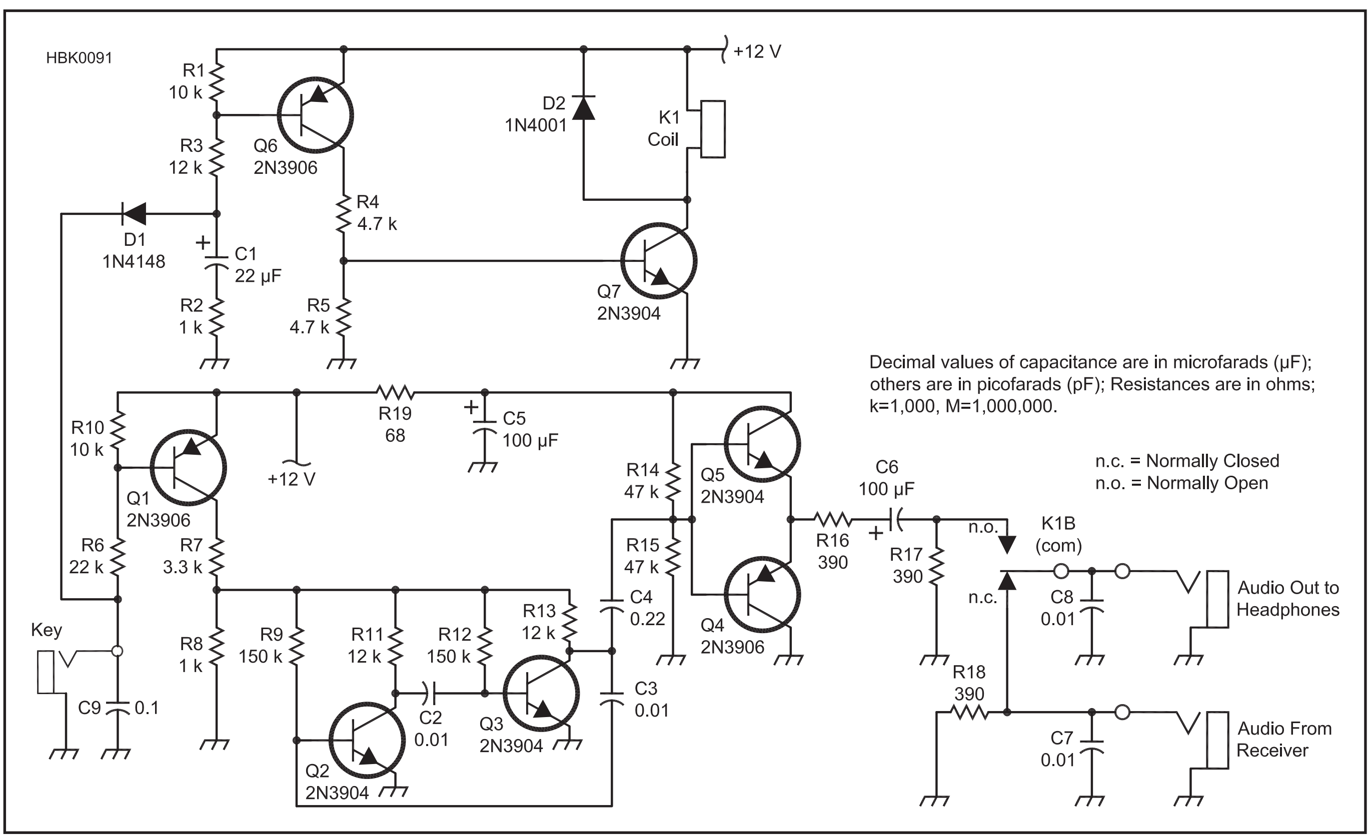


Figure 13.68 — Detailed schematic diagram and parts list for transmit-receive control section and sidetone generator of the universal QRP transmitter. Resistors are ¼ W, 5% carbon film.

C1 — 22 µF, 25 V electrolytic
C2, C3, C7, C8 — 0.01 µF, 50 V ceramic
C4 — 0.22 µF, 50 V ceramic
C5, C6 — 100 µF, 25 V electrolytic
C9 — 0.1 µF, 50 V ceramic
K1 — DPDT 12 V coil relay. An NAIS DS2Y-S-DC12, 700 Ω, 4 ms relay was used in this example.
Q1, Q4, Q6 — 2N3906, PNP silicon small signal transistor
Q2, Q3, Q5, Q7 — 2N3904, NPN silicon small signal transistor

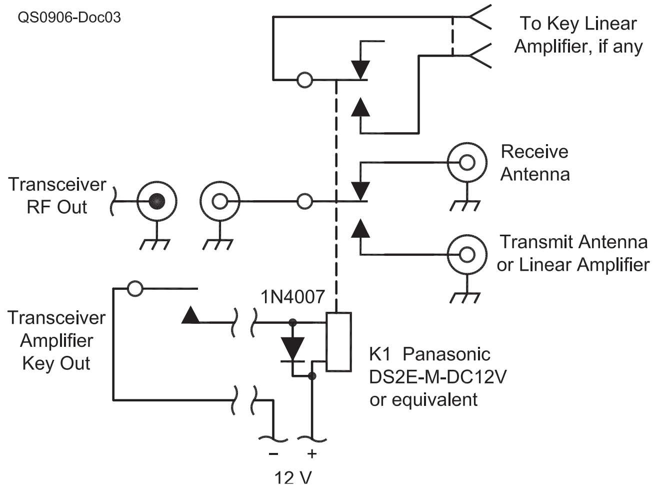


Figure 13.69 — Schematic of an external circuit that allows a modern transceiver to key a linear amplifier with TR switch voltage or current requirements that exceed the transceiver's ratings. As a bonus, it can also be used to allow reception from a low-noise receiving antenna.

also have an output signal that can be used to inhibit RF output from the transceiver until TR switching is complete and the amplifier is ready to operate. This avoids "hot switching" when the transceiver output is active while the amplifier's TR switch is still changing state.

Check the ratings to find out how much voltage and current the transceiver can safely switch, whether by a relay or a solid-state device. Many transceivers have both with the solid-state output used for high-speed switching or full break-in (QSK). Although most amplifiers are compatible with the switching capabilities of current transceivers, the voltage and/or current required to switch the relays in an older linear amplifier may exceed the ratings.

If your amplifier manual doesn't say what the switching voltage is, you can find out with a multimeter or DMM. Set the meter to read voltage of 250 V dc or higher. Connect the positive meter probe to amplifier key jack's center conductor, and connect the negative meter probe to the chassis ground (or the other key jack terminal if it's not grounded). This will tell you what the open circuit voltage is on the amplifier key jack. You may need to try a lower voltage range or switch to read ac voltage.

Now set the meter to read current. Start with a range that can read 1 A dc, and with the leads connected as before, you should hear the amplifier relay close and observe the current needed to operate the TR relay or circuit. Adjust the meter range, if needed, to get an accurate reading.

These two levels, voltage and current, are what the transceiver will be asked to switch. If *either* reading is higher than the transceiver specification, do not connect the transceiver and amplifier together. Doing so will likely damage your transceiver. You will need a simple interface circuit to handle the amplifier's switching voltage and current.

The simple, low-cost relay circuit shown in **Figure 13.69** can be used to key an older amplifier with a modern transceiver. It offers an added benefit: Another potential use of the transceiver KEY OUT jack is to switch to a separate low-noise receive antenna on the lower bands. While most high-end transceivers have a separate receive-only antenna connection built in, many transceivers don't. If you don't need one of the extra functions, just omit that connection.

13.11.7 PIN Diode RF Switching

Many current transceivers and amplifiers use PIN diodes for RF switching. They switch very quickly (as fast as nanoseconds) and do not wear out. They can be destroyed through excess reverse voltage or average forward current, requiring more protection than a simple pair of relay contacts. If high-speed, transparent break-in or digital mode operation is required, the PIN diode is the practical choice. Two practical papers available on-line with a lot of information about PIN diode use are recommended to learn more about switch design — see the References for Huff and Summers. Two industry papers on using PIN diodes are also referenced from Microsemi and Skyworks as well as a general paper by Rosu on RF switches using PIN diodes.

BASIC PIN DIODE OPERATION

A PIN diode is a diode with a wide undoped "intrinsic semiconductor" region between the usual P-type and N-type doped semiconductor layers of a regular diode (See the discussion of PIN diodes in the **Circuits and Components** chapter.) The PIN diode's wide intrinsic layer acts as a storage reservoir for charge. When forward bias current is flowing, the intrinsic layer fills up with electrons. When the diode is reverse biased, it takes time for the stored electrons to empty out of the intrinsic layer. At high enough frequencies, the time available is too short and the diode never "turns off". It remains conducting through the whole RF cycle.

PIN diodes behave like variable resistances. The resistance is determined by the amount of dc forward bias current. They are near-perfect variable resistors which add very little distortion to the signal. For switching, all we want is either a very low resistance (ON), or a very high resistance (OFF).

• To switch the PIN diode OFF, apply a large reverse bias voltage. The voltage should be higher than the peak voltage of the RF being switched. That means high voltage is required for the switch to withstand high power RF voltages when it is turned off.

• To switch the PIN diode ON, apply a forward bias current; the more current, the lower the insertion loss.

The basic element of the PIN diode switch is shown in **Figure 13.70**. The input RF signal passes through C2, then D1, and finally C3 to the output. (RF can flow in either direction — the left-to-right convention is followed here for convenience.) The dc blocking capacitors C2 and C3 are necessary to isolate the dc bias for D1. For the HF range, a 0.1 µF capacitor is satisfactory, with a reactance of approximately 1 Ω at 1.8 MHz. Remember that the capacitor will have to carry all of the RF current so a low-loss, low-inductance type is required for transmitting applications. L1 and L2 are RF chokes that prevent RF from getting into the bias source and switching

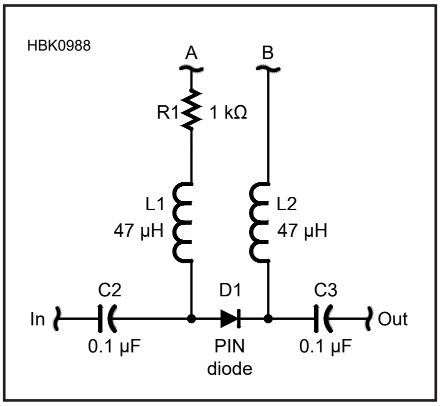


Figure 13.70 — The basic PIN diode RF switch. See text.

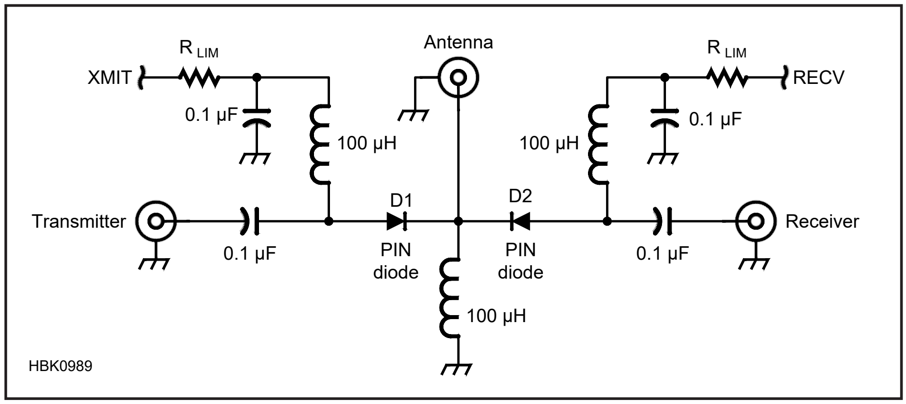


Figure 13.71 — A simple SPDT PIN diode RF TR switch to connect either a transmitter or receiver to an antenna. See text for design information.

circuits. At 1.8 MHz, the reactance of 47 µH is over 500 Ω and should be sufficient in a 50 Ω system.

To switch on D1, we apply a forward bias current through D1. For this, we set A to the supply voltage (assumed to be +12 V here) and B to ground. The current flows through the diode, limited by R1. Neglecting the diode's forward voltage, a 1 kΩ resistor results in 12 / 1000 = 12 mA. This is typical for low-power designs. Higher-power PIN diodes require more bias current to minimize insertion loss.

To switch off D1, we apply a reverse bias voltage by setting A to 0 V and B to the supply voltage. The reverse bias voltage must be higher than the peak voltage of the applied RF with some margin. The reverse voltage must also be high enough that when the diode is off there is enough isolation for the circuit requirements. Complex PIN diode switches can achieve several tens of dB of isolation. 30 dB is typical of single-diode switches like this one.

BASIC SPDT PIN DIODE SWITCH

The switch in Figure 13.70 is a SPST switch and RF current can flow through it in either direction when the PIN diode is forward biased. The switch in **Figure 13.71** is a SPDT switch with the antenna port connected to the switch common point. The diodes are turned on alternately to connect either the transmitter or the receiver to the antenna, but not both at the same time. This is a typical PIN diode TR switch.

When the XMIT signal is ON (a positive voltage), PIN diode D1 is forward biased and conducting RF while D2 is reverse biased and non-conducting. The transmitter is connected to the antenna and the receiver input path is open-circuited. When the RECV signal is ON, the situation is reversed and the receiver is connected to the antenna and the transmitter path open-circuited. The circuit's user must ensure that both XMIT and RECV cannot be on at the same time.

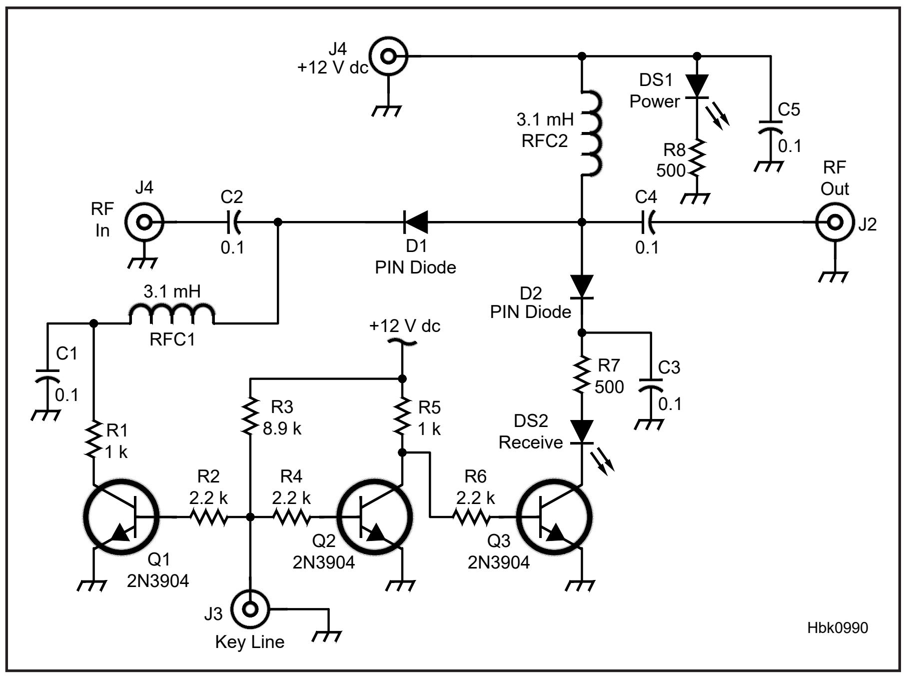


Figure 13.72 — Schematic and parts list for the PIN diode TR switch. Resistors are ¼ W.

C1-C5 — 0.1 µF ceramic capacitor
D1, D2 — PIN diode
D3, D4 — Switching diode, 1N914 or equivalent
DS1, DS2 — LED
J1, J2 — Chassis mount BNC connector
J3, J4 — Chassis mount phono connector
Q1-Q3 — Small signal PNP transistor, 2N3904 or equivalent
RFC1, RFC2 — 3.1 mH RF choke, 225 turns #30 AWG enameled wire wound on a 5⁄16 inch diameter, 5⁄8 inch long plastic tube

Depending on the type of diode being used, isolation between the transmitter and receiver ports will be about 30 dB at 28 MHz and as much as 60 dB at 1.8 MHz if careful construction techniques are followed. If additional isolation is required, multiple diode sections can be added in series or a shunt diode added as in the next section.

The value of R_{LIM} and the dc resistance of the 100 µH RF chokes limit diode current. Typical PIN diodes require 10 to 20 mA of forward bias to reach insertion loss of 0.1 dB at HF. The choice of diode will determine the amount of forward bias current required.

To perform PIN diode switching at high power levels, considerably more details are involved in the design. The online paper by Garland should be studied to learn more about use of PIN diodes at full legal limit power levels.

SERIES-SHUNT PIN DIODE SWITCH

If the transmitter is connected to the antenna at all times while the receiver is switched in and out, the series-shunt circuit of **Figure 13.72** can be used. This circuit was designed by KE2QJ and its complete operation is described in the article included with the downloadable supplemental information. When the receiver is to be connected to the antenna (RF In), series diode D1 is forward biased on and bias is removed from shunt diode D2. (It is not reverse biased.) When the transmitter is keyed (Key Line shorted to ground), forward bias is removed from series diode D1 and shunt diode D2 is forward biased on, connecting the receiver input to ground through C3. The designer reports isolation between the transmitter/antenna and receiver ports to be between 43 and 53 dB depending on bias current level.

REVERSE BIAS

To switch a PIN diode completely off, the reverse bias must be larger than the peak voltage of the applied RF. For a given power level, P, and impedance, Z:

$$V_{PK} = 1.414\ V_{RMS} = 1.414\sqrt{P \times Z}$$

At P = 10 W, V_{PK} = 32 V; at P = 100 W, V_{PK} = 100 V; and at P = 1.5 kW, V_{PK} = 387 V. This assumes SWR = 1:1. At higher SWR, the voltage at the output of a transmitter or amplifier will be higher. (See the **Transmission Lines** chapter.)

The current requirements for reverse bias voltage are very low, consisting of the diode reverse leakage current and any leakage through switching transistors. For two examples of how reverse bias voltage is obtained, see the Reference entries for Summers and Huff. A small dc-to-dc converter is another possible choice.

USING 1N4007 RECTIFIERS AS PIN DIODES

PIN diodes tend to be expensive items. However, the 1N4007 rectifier diode has a peak inverse voltage (PIV) rating of 1000 V. In order to achieve this very high PIV rating, its internal construction is very similar to a "real" PIN diode. Hence the 1N4007 is often referred to as the "Poor man's PIN diode."

Since the 1N4007 is sold as a power rectifier, its RF characteristics are not highly repeatable and its resistance is not linearly controllable. Nevertheless, to act like an ON/OFF switch across the HF range, the 1N4007 costs just a few cents and functions as well as a PIN diode designed for the same application and ratings. See the References for Huff and Summers for more information about 1N4007 performance in this application.

A properly used 1N4007 should not impact the dynamic range or third-order intercept point, IP3 in the receiver. An excellent set of measurements by IN3OTD confirms the 1N4007 is suitable for use at HF as a switching diode. (**www.qsl.net/in3otd/electronics/PIN_diodes/1N4007.html**). His broad conclusions are that a properly biased 1N4007 at HF (1.8 to 30 MHz) has the following characteristics:

- "ON" insertion loss less of than 0.1 dB for 10 mA forward bias current
- "OFF" isolation of at least 30 dB (at 30 MHz)
- IP3 > +50 dBm

Thus, the use of 1N4007 PIN diode switches should result in very little loss of transmitter output power, very little loss of sensitivity on receive, and will not degrade the excellent IP3 (and dynamic range) modern receivers.

13.12 References and Bibliography

See also the list of Books in the **Receiving** chapter references along with a list of articles about Mixers.

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