Contents

Chapter 13

Transmitters

Transmitters are the companion to the receivers discussed in the previous chapter. Many of the concepts presented here build on other sections of this book that detail modulation, mixing, filters and other related concepts. Amplifiers for power levels above 100 W are covered in the **RF Power Amplifiers** chapter. **DSP and Software Radio Design** has more information on digital techniques. Transmitters may contain hazardous voltages, and 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 Introduction

Transmitters have what would appear to be a rather straightforward job — generate an ac signal and modify it in some way to carry information. In the early days, it was found that making an electrical spark in a circuit that was connected to an antenna could be received by various devices connected to another antenna at some distance. It didn't take folks very long to figure out that if the sparks were sent in patterns corresponding to Morse or other telegraphy codes, information could be sent distances without connecting wires. Wireless communication was born!

Technology quickly advanced through a number of variations of spark and arc transmitters and a wide variety of ingenious receiving devices until wireless telegraphy became a serious business involved with monitoring the safety of ships at sea in ways that hadn't been possible before. See **Fig 13.1** for a view of an early shipboard wireless station. The systems of the

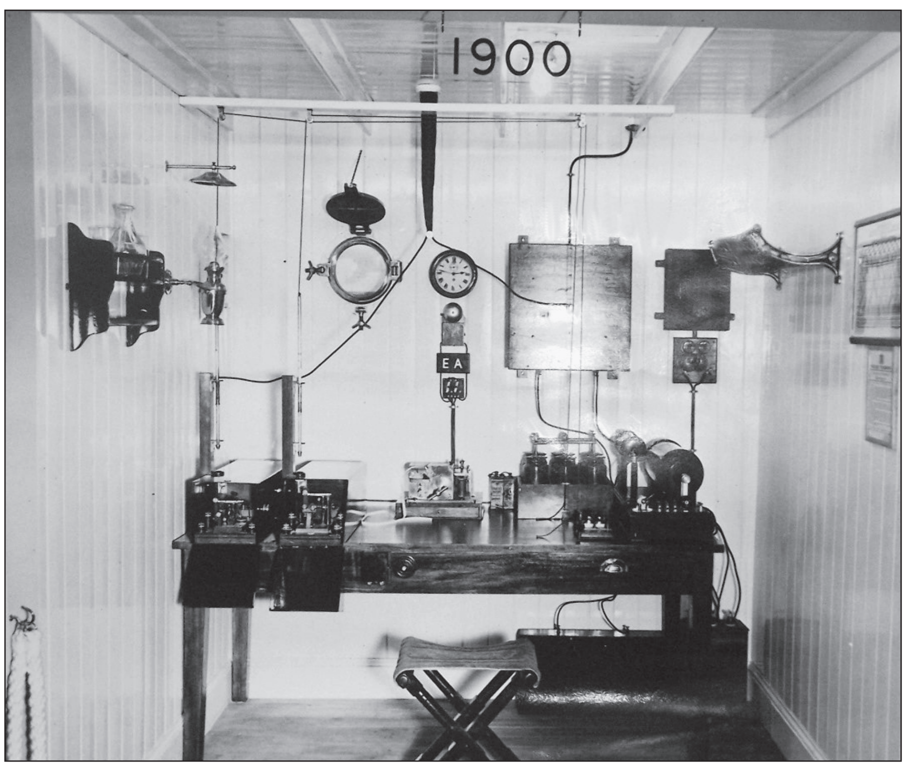


Fig 13.1 — Marconi shipboard radiotelegraph station from 1900. (*Photo courtesy of Marconi Company, PLC*)

day were more electromechanical than what we would consider electronic today, culminating in very high speed rotary alternators that could generate high power ac directly. An RF alternator was an electromechanical alternating current generator that produced ac at radio frequencies.

Transmitter technology 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 they may not be the same size. An RF amplifier in a receiver may deal with amplifying picowatts while one in a transmitter may output up to megawatts. 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.

13.2 Early Transmitter Architectures

13.2.1 Alternator Transmitters

A very basic alternator might consist of a coil with a core of *soft iron* (pure molecular iron that will magnetize and demagnetize easily) rotating past two poles of a magnet inside a round iron mounting as in **Fig 13.2**. As one end of the rotating iron cored coil passes the north magnetic pole, a magnetic field is built up in it. As a result of the magnetic field increasing and then decreasing in the rotor coil's core, one half of an ac cycle is induced in its coil. As it continues to rotate and passes the south pole, an opposite half cycle of ac is induced in it. To produce 60 Hz ac, the coil would have to rotate 60 times per second, or 3600 r/min, a very fast rotation. The two ends of the rotating coil are connected to two *slip rings* on the shaft that rotates the coil. There are two brushes of fixed carbon or other material that make a constant contact with the rotating slip rings. Any ac voltage that is generated is available at these two brushes. By using multiple rings and brushes, multiple cycles of ac will be produced with each rotation.

Better systems were used to generate lower frequency RF ac. Ernst Alexanderson, a prodigious Swedish/American electrical engineer and inventor, and Canadian inventor Reginald Fessenden first produced a lower power 60 kHz RF alternator in 1906 with which they could transmit telegraphy as well as voice and music. Later the higher powered *Alexanderson alternators* used frequencies closer to 20 kHz. Other popular alternators of the period were the German Joly-Arco and the Goldschmidt, which was somewhat similar to the Alexanderson.[1]

The final form of the Alexanderson alternator uses a rotary toothed disc, called an *inductor*. This inductor is a large round, flat, soft-iron disk often with teeth ground out at its edge as shown in **Fig 13.3**. The many teeth of the inductor are rotated rapidly by a constant speed ac motor between the N and S poles of double wound field pole electromagnets mounted one tooth-width apart all around the inside of the machine. As an iron tooth is passing between the N and S ends of the dc-excited electromagnet field poles it provides a better path for the N to S magnetic lines of force. As a result the magnetism in the field pole builds up, but collapses as the tooth moves on.

As the next inductor tooth passes the field pole it develops another magnetic field build up and collapse. These varying magnetic fields induce ac cycles in all of the secondary RF output field-pole coils. Since there are as many field poles around the inside of the machine as teeth on the inductor, ac voltages are developed in the field pole ac pick-up coils all at the same time and are all inductively coupled to a coil/antenna/ground circuit that radiates the RF waves. The speed of the motor rotating the inductor and the number of teeth determines the output frequency.[2]

13.2.2 Vacuum Tube Transmitters

By the early 1920s, large vacuum tubes were developed that could be used in high powered transmitters. Most of the big alternators were replaced by less massive high power tube transmitters. In 1920, KDKA in Pittsburgh, Pennsylvania became the first federally licensed commercial AM broadcast station in the country, operating on 1020 kHz, initially using a 100 W vacuum tube transmitter.

OSCILLATOR TRANSMITTERS

Amateur transmitters also quickly moved from spark and arc to vacuum tube sets.

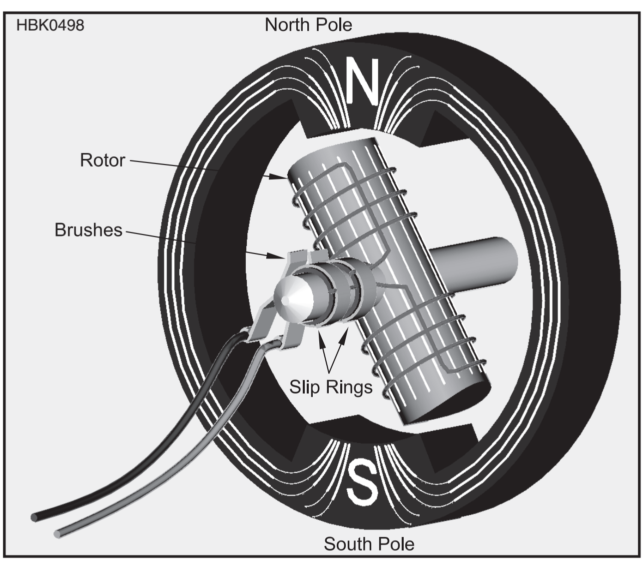


Fig 13.2 — Basic alternator consisting of a soft-iron cored coil rotating past two poles of a magnet.

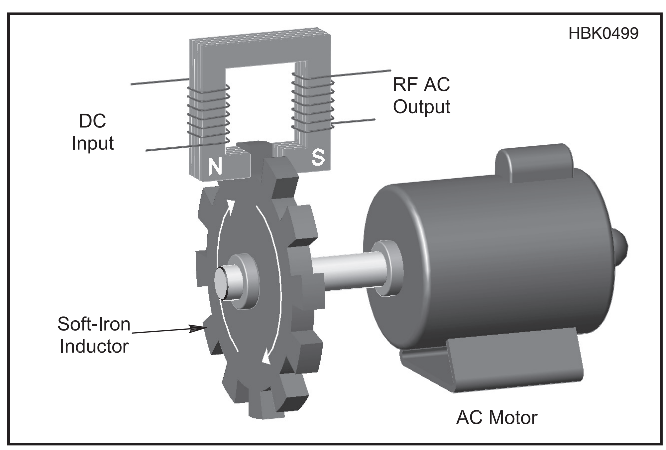


Fig 13.3 — High frequency alternator using an inductor disc.

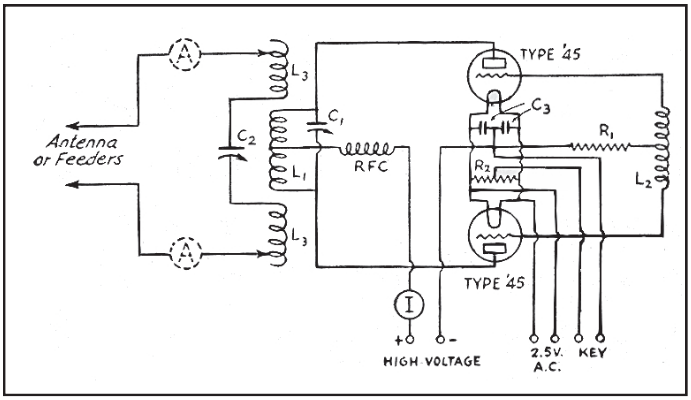


Fig 13.4 — Transmitter schematic as shown in an early *QST* magazine.

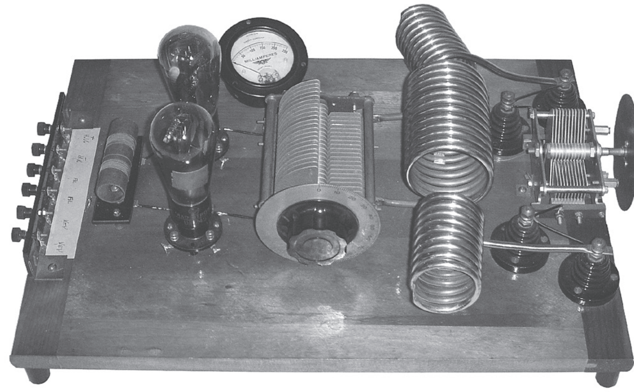


Fig 13.5 — Photo of a transmitter built from the schematic in Fig 13.4.

Because the vacuum tube oscillators generated a single-frequency RF signal directly and continuously, transmitters using them were called *continuous wave* or *CW* transmitters. While generally having less nominal power than the spark transmitters of the day, more power was actually provided to the antenna by CW transmitters. Signals were of a much narrower bandwidth, allowing more usable channels once receiver capabilities caught up. The early tube transmitters were generally single-stage self-excited oscillators coupled directly to an antenna as shown in **Figs 13.4** and **13.5**.

While a vast improvement over early technology, such transmitters had their limitations. Note that the antenna was connected directly to the tuned circuit that controlled the frequency. This meant that every time the antenna moved in the wind, the resultant change in loading shifted the transmit frequency. If the cat walked by and was lucky enough not to be electrocuted by the exposed high voltage wiring, the frequency moved even on a calm day. The frequency would often shift with each key closure as the power supply voltage sagged with the additional current drain, resulting in signals with characteristic chirps and whoops.

MASTER OSCILLATOR-POWER AMPLIFIER TRANSMITTERS

All of these issues were at least partly addressed by adding a power amplifier between the oscillator and the antenna. Morse code transmission was accomplished by using the key to turn the power amplifier stage on and off. If properly designed and adjusted and powered by a solid power supply, the oscillator could be left on between the transmitted dots and dashes. This stabilized the transmitted signal's frequency and the antenna was well-isolated from the frequency determining components. This architecture was termed a *master oscillator-power amplifier* (MOPA) configuration and remained popular in WWII military radios and amateur equipment into the 1950s. The popular ARC-5 series military aircraft HF transmitters were of MOPA design using a single-triode variable-tuned oscillator and a parallel pair of tetrode amplifier tubes. While the military ran these at more conservative ratings, amateurs used them to provide up to 100 W output, all from a box the size of a small loaf of bread. Two examples are shown in **Fig 13.6**.

CRYSTAL-CONTROLLED TRANSMITTERS

The transmitters discussed so far operated on frequencies under the control of variable-tuned LC circuits. This provided a certain amount of flexibility, but had the down side of making the operating frequency uncertain. Stability over time, supply voltage and temperature was not as good as might be desired.

Properties of crystal structures were studied at least as far back as the 18th century, but the first major application of the piezoelectric effect that converts crystal motion to voltage and back occurred during WWI. The French developed a piezoelectric ultrasonic transducer that was used as the transmitter in an acoustical submarine detection system. Between WWI and WWII, the use of resonant wafers of quartz crystal as frequency determining elements in oscillators became feasible and packaged crystals became readily available. Crystal frequency control was particularly suitable for radios that operated on assigned fixed channels, common to most services other than Amateur Radio. Still, crystal-controlled transmitters were popular in amateur service, with most stations having a large collection of crystals that could be selected to change frequency. The now-discontinued entry-level Novice class license for many years

Fig 13.6 — A pair of 7-9.1 MHz ARC-5 transmitters from WW2. On the left the US Army Air Forces version (BC-459), on the right a Navy equivalent (T-22, ARC-5). These were available for typically $5 in "new in box" condition after WW2.

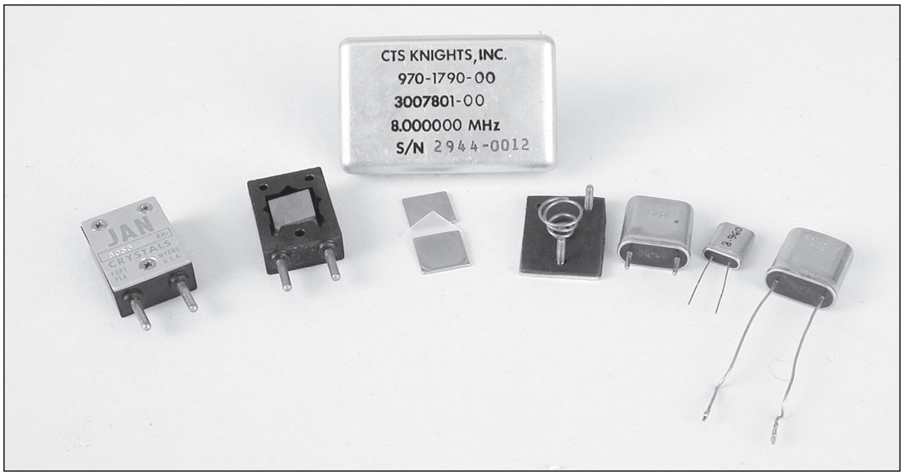


Fig 13.7 — Sample of typical frequency control crystals for amateur use. The units on the right are modern sealed types. On the far left is a WWII era crystal with a disassembled holder in the center.

required crystal control of the transmitter. **Fig 13.7** shows a selection of popular crystal types.

MULTIBAND TRANSMITTERS

The transmitters discussed so far generally operated over a single band or frequency range. In order to operate on multiple bands, a number of approaches were used. The early amateur bands were selected to be harmonically related in order to avoid interference to other services from spurious harmonic signals. The amateur bands of the early years were 160, 80, 40, 20, 10, 5 and 2½ meters.

Frequency Multiplier Transmitters

A popular approach to early AM and CW transmitters designed for multiple bands was to use frequency multiplier stages to raise the frequency to a desired harmonic. A frequency multiplier was just an amplifier operated in class C with the output tuned circuit tuned to a multiple of the input frequency. The plate current pulses would cause the output tank circuit to ring at the selected harmonic, filling in the cycles between the pulses. If the Q of the tank circuit was high enough, harmonics through the third or fourth could be generated with acceptable distortion. Additional tuned stages would reinforce the desired harmonic and reject the other signals.

A block diagram of a typical period transmitter for 80, 40 and 20 meters is shown in **Fig 13.8**. Note that a particular advantage of this configuration is that a single crystal at say 3.505 MHz can be used at 7.010 and 14.020 MHz. This arrangement was quite common in transmitters designed for frequency modulation in which not only the frequency, but also the deviation, is multiplied at each step.

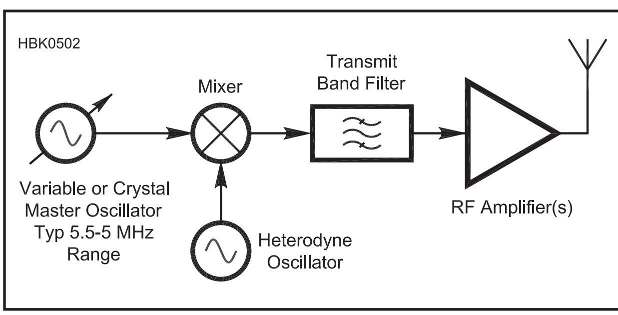
While the figure shows cascaded doublers, some used circuits switchable between doublers, triplers and quadruplers to avoid the need for additional stages. Note that a limitation of this arrangement is that any drift or instability is also multiplied, making it necessary that particular attention is paid to oscillator stability. The tuning rate is also different on each band.

Heterodyne Transmitters

Another transmitter configuration looks and sounds a lot like the architecture of a modern receiver, but backwards — the heterodyne transmitter. A block diagram of a representative transmitter of that type is shown in **Fig 13.9**. This is the basis for the transmitter section of most SSB transmitters as well as modern SSB transceivers described in the **Transceivers** chapter. There are some clear advantages to this approach. In the early days, it was possible to have both oscillators running continuously with CW keying applied to the mixer stage. This provided the highest frequency stability of any configuration discussed so far.

Another plus compared to the multiplying transmitter was that the tuning rate of the variable frequency oscillator (VFO) was the same on each band. This allows more precise tuning adjustment on the higher frequency ranges. As with heterodyne receivers, multiple conversions are encountered, particularly in radios working into the VHF and UHF regions and those with DSP waveform generation. Any images resulting from the conversion process must be eliminated by the filtering between the mixer and power amplifier stages, with similar constraints to the image issues of receivers.

Some early heterodyne SSB transmitters



Heterodyne Osc Freq MHz	Output MHz
9	3.5-4.0
12.5	7-7.5
19.5	14-14.5
26.5	21-21.5
33.5	28-28.5

Fig 13.9 — Block diagram of typical heterodyne type transmitter.

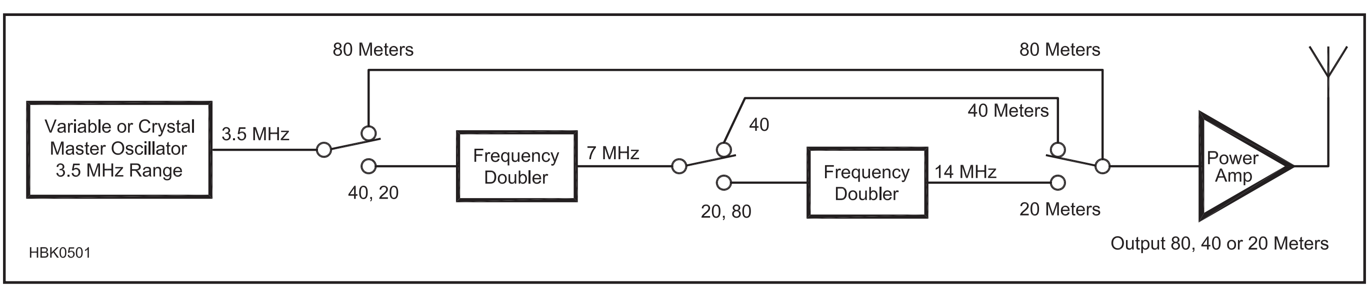


Fig 13.8 — Block diagram of typical frequency multiplier transmitter.

took advantage of the fact that the 9 MHz heterodyne oscillator frequency could actually be used on both 80/75 and 20 meters with the same 5.0 to 5.5 MHz frequency range in the VFO. The difference product at 9 MHz — the VFO frequency (4 to 3.5 MHz) was used for the lower band while the sum product at 9 MHz + the VFO frequency (14 to 14.5 MHz) was used for 20 meters. This was a popular dual-band scheme in the early days of SSB voice. One drawback was that the VFO tuned in the opposite direction on each band. For many hams of the day, it was a small price for the simplicity.

13.3 Modulation Types and Methods Applied to Transmitter Design

13.3.1 Data, Morse Code or Pulse Transmitters

The simplest transmitter consists of an oscillator generating a signal at the frequency we want to transmit. If the oscillator is connected to an antenna, the signal will propagate outward and be picked up by any receivers within range. Such a transmitter will carry little information, except perhaps for its location — it could serve as a rudimentary beacon for direction finding or radiolocation, although real beacons generally transmit identification data. It also indicates whether or not it is turned on, perhaps useful as part of an alarm system.

To actually transmit information, we must *modulate* the transmitter. The modulation process, covered in detail in the **Modulation** chapter, involves changing one or more of the signal parameters to apply the information content. This must be done in such a way that the information can be extracted at the receiver. As previously noted, the parameters available for modulation are:

Frequency — this is the number of cycles the signal makes per second.

Amplitude — although the amplitude, or strength, of a sinusoid is constantly changing with time, we can express the amplitude by the maximum value that it reaches.

Phase — the phase of a sinusoid is a measure of when a sinusoid starts compared to another sinusoid of the same frequency.

We could use any of the above parameters to modulate a simple transmitter with pulse type information, but the easiest to visualize is probably amplitude modulation. If we were to just turn the transmitter on and off, with it on for binary "one" and off for a "zero" we could surely send Morse code or other types of pulse-coded data. This type of modulation is called "On-Off-Keying" or "OOK" for short.

Some care is needed in how we implement such a function. Note that if we performed the obvious step of just removing and turning on the power supply, we might be surprised to find that it takes too much time for the voltage to rise sufficiently at the oscillator to actually turn it on at the time we make the connection. Similarly, we might be surprised to find that when we turn off the power we would still be transmitting for some time after the switch is turned. These finite intervals are referred to as *rise* and *fall times* and generally depend on the time constants of filter and switching circuits in the transmitter.

There is an easy-to-visualize analogy with using light to transmit code signals. If you've ever seen a WWII movie showing a Navy Morse operator sending data between ships with a lantern, you may have wondered why they used a special mechanical shutter device instead of turning the light bulb on and off.[3] One reason is that the bulb continues to radiate for a period while the filament is still hot, even after the power is off.

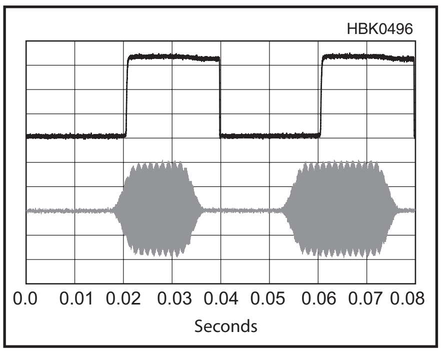


Fig 13.10 — The CW keying waveform of a transmitter with good spectrum control. The top trace is the key closure, with the start of the first contact closure on the left edge at 60 WPM using full break-in. Below that is the nicely rounded RF envelope.

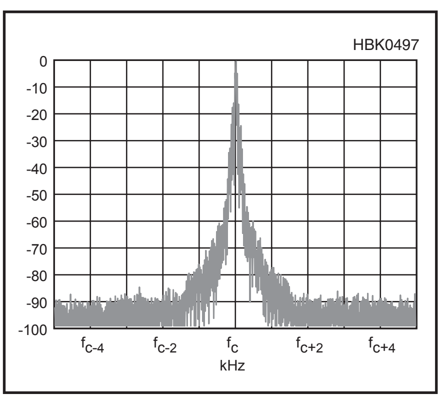


Fig 13.11 — The resultant signal spectrum from the keying shown in Fig 13.10. Note that the signal amplitude is about 80 dB down at a spacing of ±1 kHz, with a floor of –90 dB over the 10 kHz shown.

REAL WORLD CW KEYING

In the **Modulation** 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. **Fig 13.10** 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. **Fig 13.11** 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. **Figs 13.12** and **13.13** 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

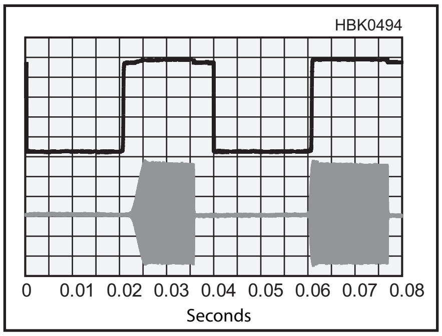


Fig 13.12 — The CW keying waveform of a transmitter with poor spectrum control. The top trace is the key closure, with the start of the first contact closure on the left edge at 60 WPM using full break-in. Note the sharp corners of the RF envelope that result in excessive bandwidth products.

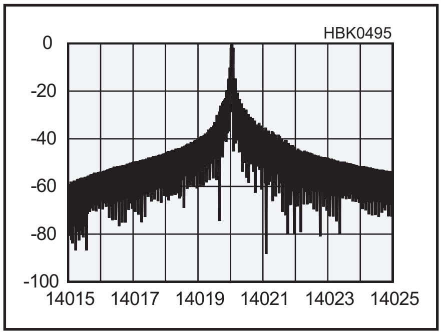


Fig 13.13 — The resultant signal spectrum from the keying shown in Fig 13.12. The resulting spectrum is not even down 40 dB at ±1 kHz and shows a floor that doesn't quite make 60 dB below the carrier across the 10 kHz.

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.

13.3.2 *Project*: Simple CW Transmitters

The schematic for a very basic low power HF CW oscillator transmitter is shown in **Fig 13.14.** Using a crystal-controlled oscillator gains frequency accuracy and stability but gives up frequency agility.

In this design, the key essentially just turns power on and off. The value of the keying line 0.1 µF bypass capacitor was chosen

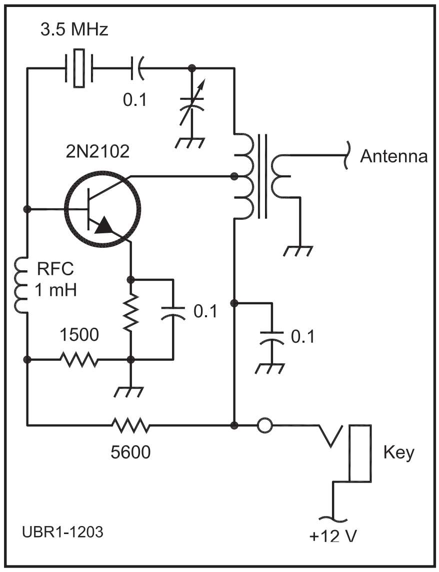


Fig 13.14 — A simple solid-state low-power HF crystal controlled oscillator-transmitter. The unspecified tuned circuit values are resonant at the crystal frequency.

so it would not create an excessive rise or fall time at reasonable keying speeds. This transmitter will generate a few mW of RF power at the crystal frequency, but it has a number of limitations in common with early vacuum tube transmitters that were such an improvement over the spark transmitters that preceded them.

First, the oscillator is dependent on the environment. Changes in the antenna from wind, for example, change the load and can cause the frequency to shift. Second, the oscillator generates signals at harmonics of the fundamental frequency. Third, every time the key is closed the oscillator must start up, taking a short but audible time for the output frequency to stabilize. This creates a distinctive change in output frequency referred to as "chirp," since the signal sounds like a bird's chirp in the receiver.

By adding an amplifier stage to the oscillator transmitter, the oscillator can be isolated from changes in the environment to improve stability. Adding filtering at the output addresses the problem of harmonics.

The following project by Rev George Dobbs, G3RJV, was published in the Spring 2012 issue of *QRP Quarterly* and is reprinted courtesy of the QRP Amateur Radio Club (**www.qrparci.org**) and *Practical Wireless* (**www.pwpublishing.ltd.uk**). An article by Ed Hare, W1RFI describing the Tuna-Tin 2 transmitter that was included in previous editions of the *ARRL Handbook* is available on this book's CD-ROM of supplemental material.[4]

Fig 13.15 is the schematic of a much better, but still simple, CW transmitter. It is an update to the "Pebble Crusher" originally designed by Doug DeMaw, W1FB (SK) and described in his *QRP Notebook* (out of print). The transmitter uses an oscillator followed by a stage of amplification to about ½ W of output. Note that the output circuit includes a filter to reduce any harmonic output below the levels required by FCC rules.

The circuit is a two transistor transmitter using 2N2222A devices. These are small-signal devices so the transmitter output is only about ½ W. What might surprise the reader is the number of inductors (coils) in the design, but the circuit has good performance in impedance matching and harmonic reduction. A careful design ensures a clean output waveform, low in harmonic and other spurious content.

The oscillator is a variable frequency crystal oscillator (VXO) in which the 100 pF variable capacitor between the crystal and ground enables the oscillator frequency to be shifted. Depending upon individual crystals, that movement can be on the order of 5 kHz. Increasing the variable capacitance too much may cause the transistor to cease oscillating. The 150 pF capacitor in the emitter of the oscillator controls the oscillator feedback. This value is a compromise that appears to work well — but if the transistor fails to oscillate try increasing this value a little.

The 4.7 µF electrolytic capacitor across the KEY connection minimizes key clicks by softening the keying waveform. Lowering this value gives a harder keying waveform and increasing it will further soften the waveform. A 10 Ω resistor is added to the base of the transistor to reduce the risk of high frequency parasitic oscillation.

A 22 µH RF choke (RFC1) provides the RF collector dc supply for the oscillator. A molded inductor was used in the original design. If this isn't available, about six turns of thin enameled wire wound through a small ferrite bead (Type 43 mix) would give roughly the same inductance value.

The output from the collector goes to a single-element harmonic filter in a pi configuration. A 100 pF variable capacitor tunes the filter to resonance at the oscillator frequency. It should be adjusted for maximum output consistent with a clean CW signal.

If the intent is to use the oscillator stage as a transmitter without additional amplification, the filter is designed to have a 50 Ω output impedance for a matched 7 MHz antenna. The antenna can be connected in place of T1 in the schematic.

A 4:1 impedance ratio transformer (T1) matches the output of the oscillator to the input of the class C amplifier. This enables good power transfer and isolation from the oscillator stage. A ferrite bead (Type 43 mix) is added to the base lead of the amplifier to

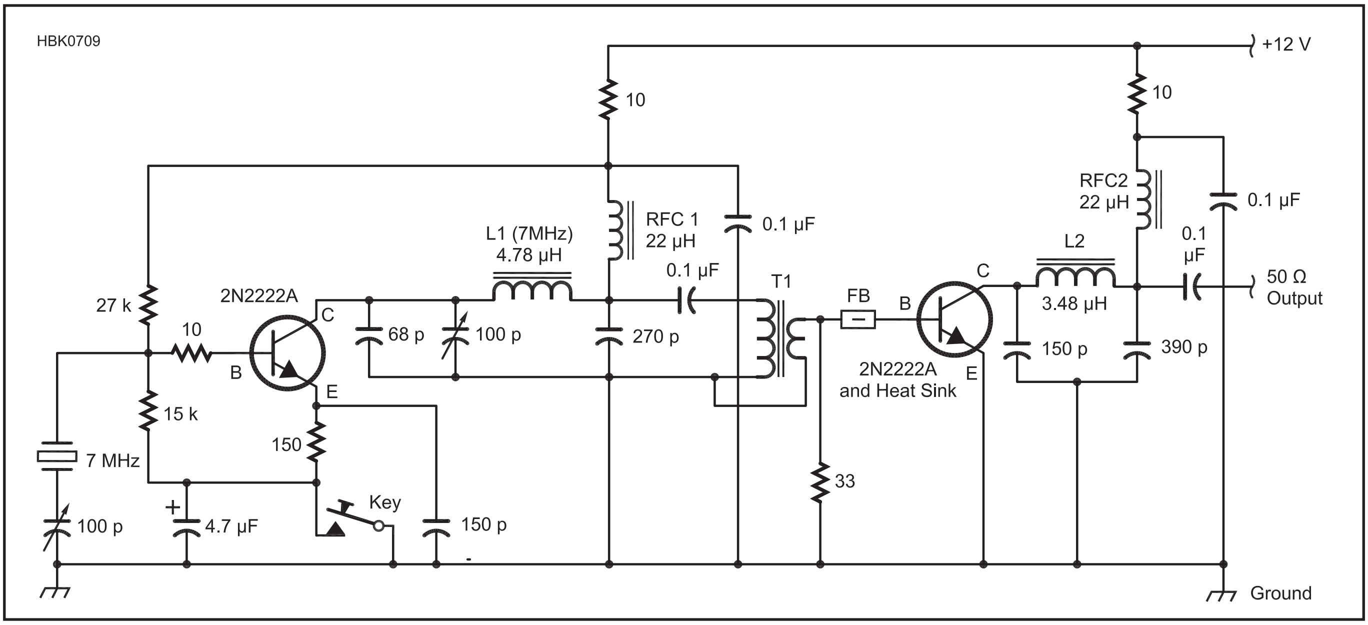


Fig 13.15 — The circuit of the Pebble Crusher transmitter. The oscillator section at left can be used as a standalone transmitter by replacing T1 with an antenna. The output of the amplifier circuit is approximately ½ W.

L1 — 32t #30 AWG on T50-6 toroid core
L2 — 28t #24 AWG on T50-6 toroid core
RFC1, RFC2 — 22 µH or 6t #30 AWG on ferrite bead (Type 43 mix)
T1 — 12t #24 AWG (primary) and 6t #24 AWG (secondary) on FT37-43 toroid core (see text)
FB — ferrite bead (Type 43 mix)

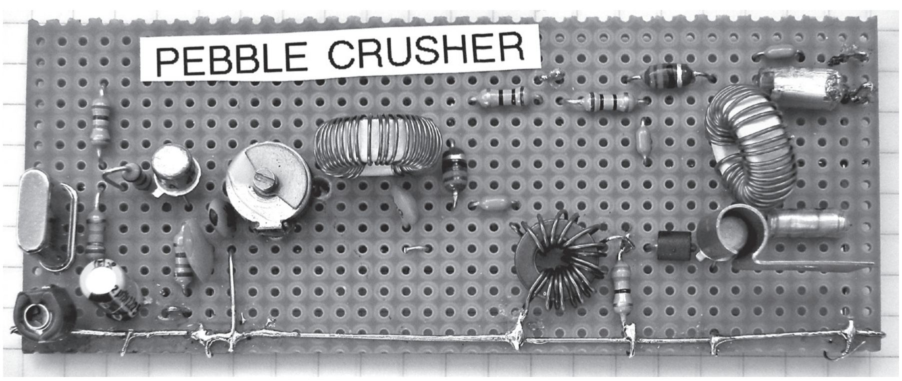


Fig 13.16 — G3RJV's "Anglicised" version of the Pebble Crusher is built on 0.1 inch spacing perfboard. The heat sink for the output transistor at lower right can be a commercial unit or created from metal strip or tubing (see text).

prevent possible parasitic oscillations in the amplifier transistor.

In the amplifier output circuit another 22 µH RF choke supplies the collector and a pi-type low-pass filter matches the collector to a 50 Ω output. The filter is designed with a cutoff frequency of around 7.3 MHz. Some experienced constructors of simple transmitters may consider two harmonic filters in such a simple project to be overkill. However, it does mean that the driving power to the amplifier has little harmonic content so unwanted signal energy is not amplified or radiated.

G3RJV's version of the Pebble Crusher shown in **Fig 13.16** was built on 0.1 inch spacing insulating perfboard. (See the **Construction Techniques** chapter for information on building circuits in this way.) The layout roughly follows the component placement in the circuit diagram. This version in the photo does not have the VXO adjustment capacitor.

A trimmer capacitor can be used for the VXO control although some builders may prefer a variable capacitor with a control shaft. The variable capacitor in the oscillator low-pass filter is also a trimmer capacitor.

The inductors L1 and L2 should be wound to occupy about three-quarters of the circumference of the toroid core. L1 has 32 turns of #30 AWG enameled copper wire on a T50-6 core and L2 has 28 turns of #24 AWG enameled copper wire also on a T50-6 core. The transformer (T1) is wound on a ferrite FT37-43 core. Wind the primary first with 12 turns of #24 AWG spread out over three-quarters of the core circumference. Then add the secondary winding which is made up of 6 turns of #24 AWG wire between turns of the primary winding at the grounded end.

2N2222A transistors in a metal TO-18 case are shown, but the plastic TO-92 version would also work. An advantage of using the metal TO-18 is that a small heat sink can be attached to the output transistor to dissipate surplus heat. If a commercial heat sink is not available, a small piece of aluminum or copper strip can be formed into a heat sink. The strip is wrapped around a drill bit the same diameter as the transistor case, one side overlapping where the ends of the metal meet. This is then squeezed to make a tight fit on the TO-18 case. A small piece of ¼ inch OD brass tubing slightly flattened and cemented to the transistor casing with epoxy will work, as well.

13.3.3 Amplitude-Modulated Full-Carrier Voice Transmission

While the telegraph key in the transmitters of the previous section can be considered a *modulator* of sorts, we usually reserve that term for a somewhat more sophisticated system that adds information to the transmitted signal. As noted earlier, there are three signal parameters that can be used to modulate a radio signal and they all can be used in various ways to add voice (or other information) to a transmitted signal.

One way to add voice to a radio signal is to first convert the analog signal to digital data and then transmit it as "ones" and "zeros". This can be done even using the simple telegraph transmitters of the last section. This is a technique frequently employed for some applications using data applied to our "pulse" transmitter described earlier. Here we will talk about the more direct application of the analog voice signal to a radio signal.

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. **Fig 13.17** shows the conceptual view of this**. Fig 13.18** is a more detailed view of how such a voice transmitter would actually be implemented. The upper portion is the RF channel, and you can think of the previously described Tuna Tin Two transmitter as a transmitter that we could use, after shifting the frequency into the voice portion of the band. The lower portion is the audio frequency or AF channel, usually called the *modulator*, and 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 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 **Fig 13.19.** The spectrum shown would be that of a standard broadcast station with an audio passband from 50 Hz 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 **Receivers** 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.

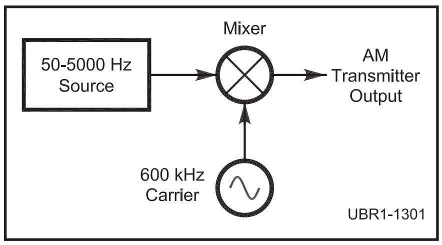


Fig 13.17 — Block diagram of a conceptual AM transmitter

13.3.4 Single-Sideband Suppressed-Carrier Transmission

The two sidebands of a standard AM transmitter carry (reversed) 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.

SINGLE-SIDEBAND — THE FILTER METHOD

The block diagram of a simple single-sideband suppressed-carrier (SSB) transmitter is shown in **Fig 13.20**. This transmitter uses a balanced mixer as a *balanced modulator* to generate a double sideband suppressed carrier signal without a carrier. (See the **Mixers, Modulators, and Demodulators** chapter for more information on these circuits.) That signal is then sent through a filter designed to pass just one (either one, by agreement with the receiving station) of the sidebands.[5]

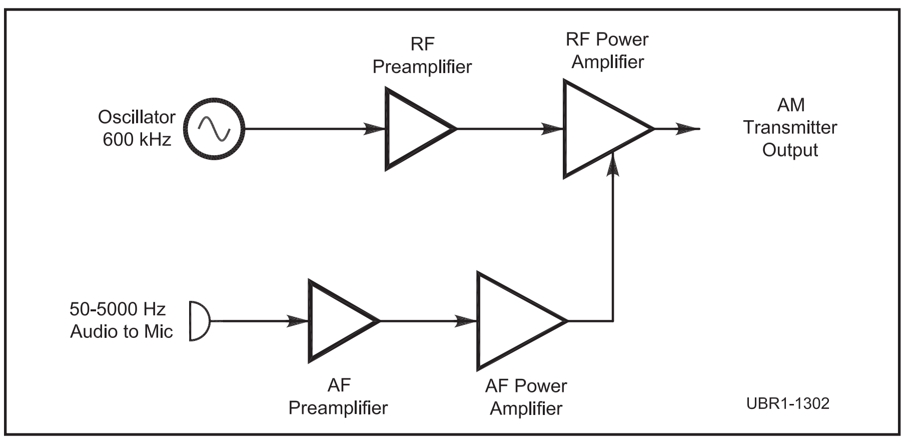


Fig 13.18 — Block diagram of a 600 kHz AM broadcast transmitter.

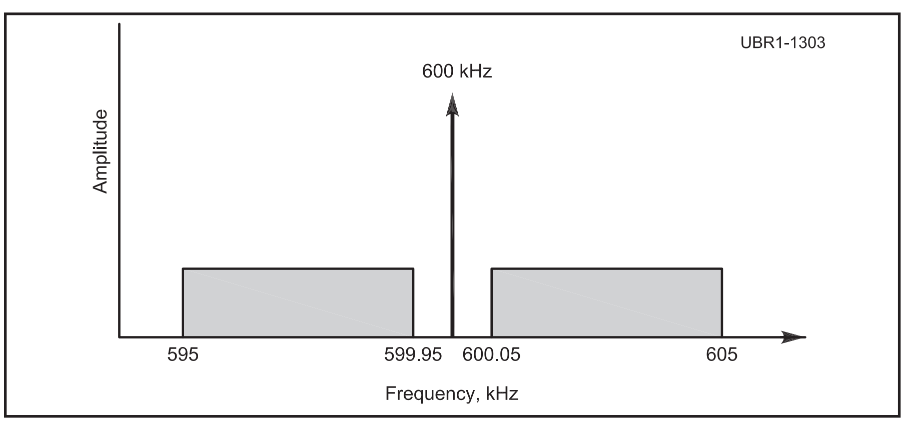


Fig 13.19 — Spectrum mask of a 600 kHz AM broadcast signal.

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.

While a transmitter of the type in Fig 13.20 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 **Fig 13.21**. 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, 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 as shown in the **Receivers** chapter, 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.

SINGLE-SIDEBAND — THE PHASING METHOD

Most current transmitters use the method of SSB generation shown in Fig 13.20 and discussed in the previous section 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 **Receivers** 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 **Fig 13.22A**. 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 a 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.

SINGLE-SIDEBAND — 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 Fig 13.22B, 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 outputs 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. As more transmitters incorporate DSP technology, the Weaver method will become common.

13.3.5 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.

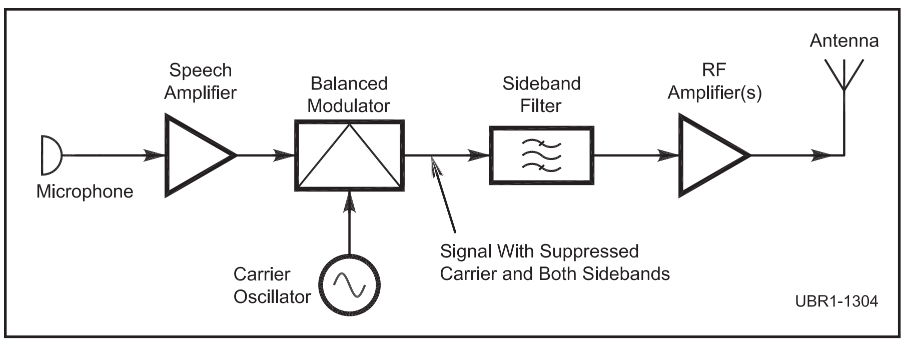


Fig 13.20 — Block diagram of a filter type single sideband suppressed carrier (SSB) transmitter.

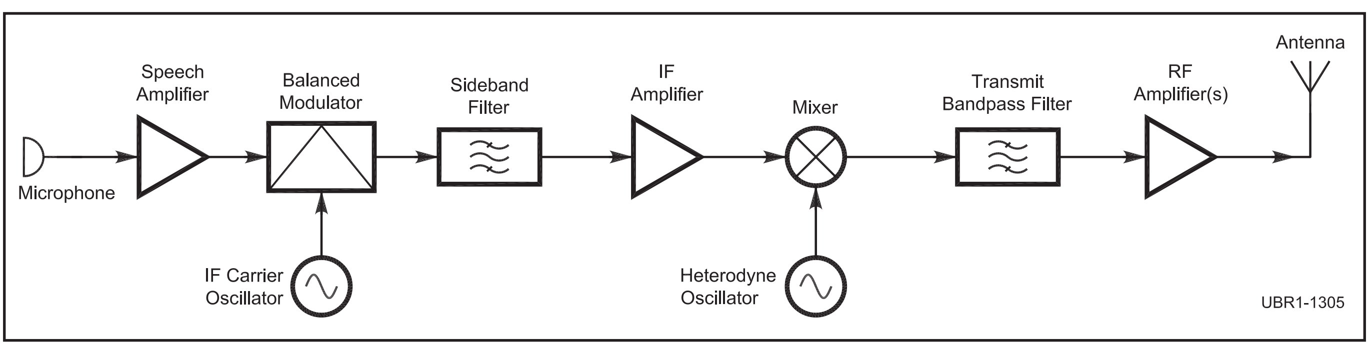


Fig 13.21 — Block diagram of a heterodyne filter-type SSB transmitter for multiple frequency operation.

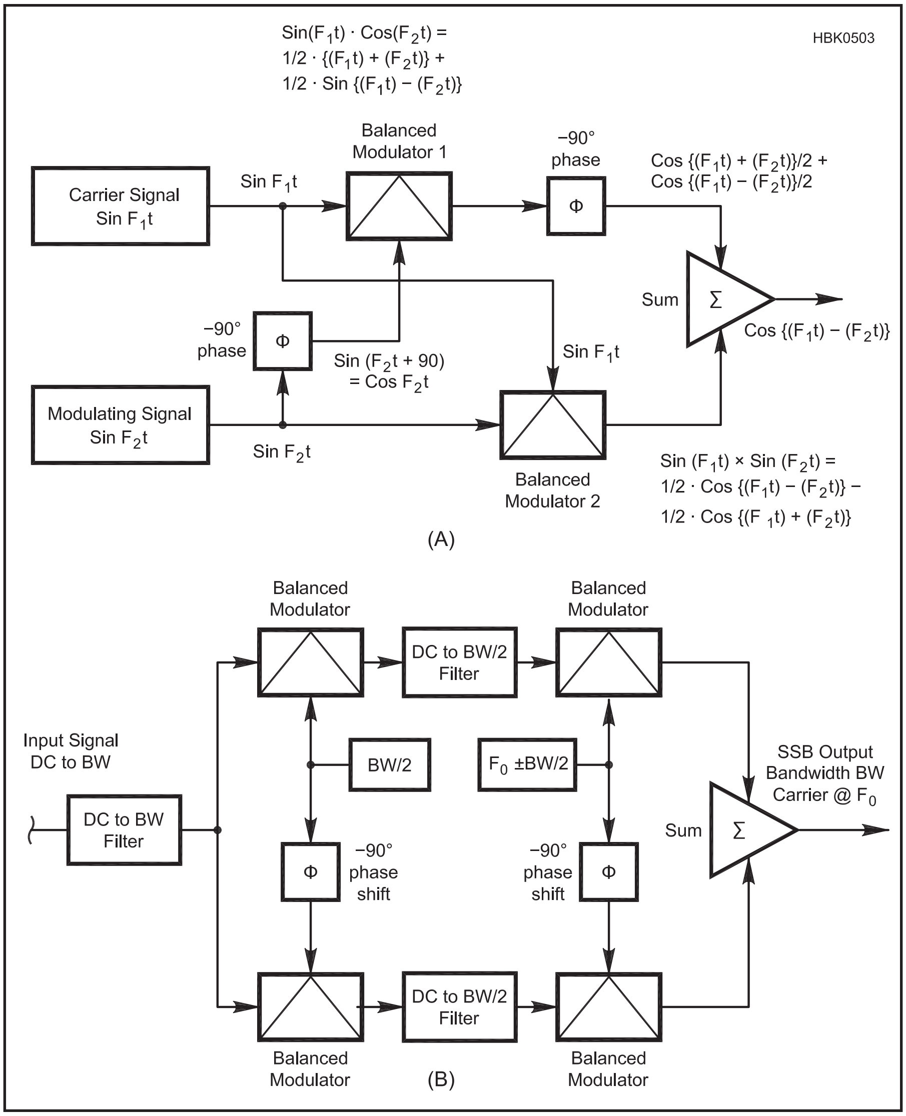


Fig 13.22 — Block diagrams of phasing type SSB transmitters for single frequency operation.

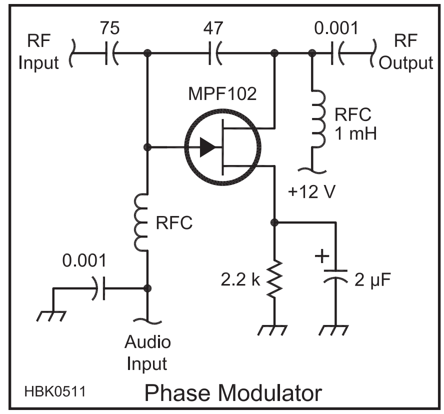


Fig 13.23 — Simple FET phase modulator circuit.

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 **Fig 13.23**, the 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.

FREQUENCY MODULATION TRANSMITTER DESIGN

Frequency modulation is widely used as the voice mode on VHF for repeater and other point-to-point communications. **Fig 13.24** 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

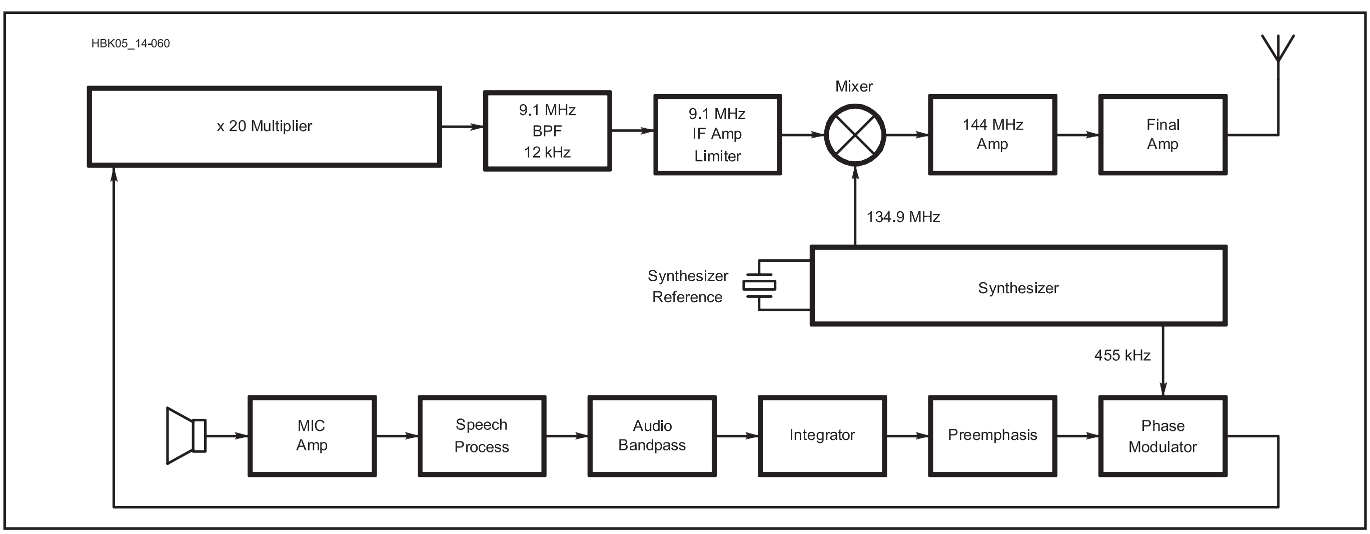


Fig 13.24 — Block diagram of a VHF/UHF NBFM transmitter using the indirect FM (phase modulation) method.

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.

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; 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.

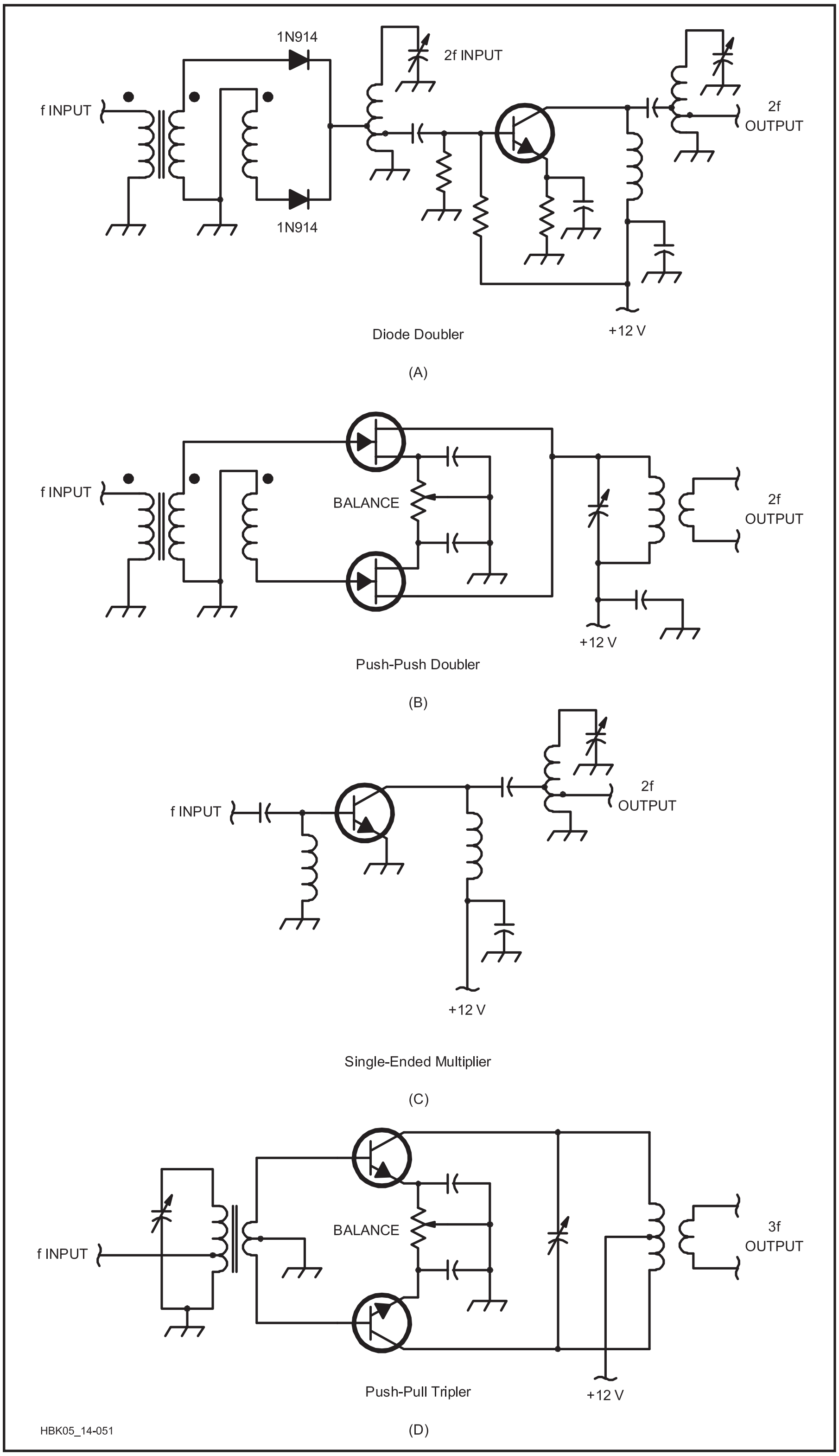


Fig 13.25 — A: diode doubler. B: push-push doubler using JFETS. C: single-ended multiplier using a BJT. D: push-pull tripler using BJTs.

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.

A passive multiplier using diodes is shown in **Fig 13.25A**. 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 **Analog Basics** 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 Fig 13.25B. The FETs are biased in the square-law region and the BALANCE potentiometer minimizes the fundamental frequency. Note that the gates are in push-pull and the drains are in parallel. This causes second harmonics to add in-phase at the output and fundamental components to cancel.

Fig 13.25C shows an example of a bipolartransistor doubler. 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 Fig 13.25D. 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 **Fig 13.26A** 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. Fig 13.26B 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. Preemphasis 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.[6] 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.

13.3.6 A Superhet SSB/CW Transmitter

The modern linear transmitting chain makes use of the concepts presented previously. We will now go through them in more detail. The same kind of mixing schemes, IF frequencies and IF filters that are used for superhet receivers can be, and very often are, used for a transmitter. In the following discussion, "in-band" refers to signal frequencies within the bandwidth of the desired signal. For example, for an upper sideband voice sig-

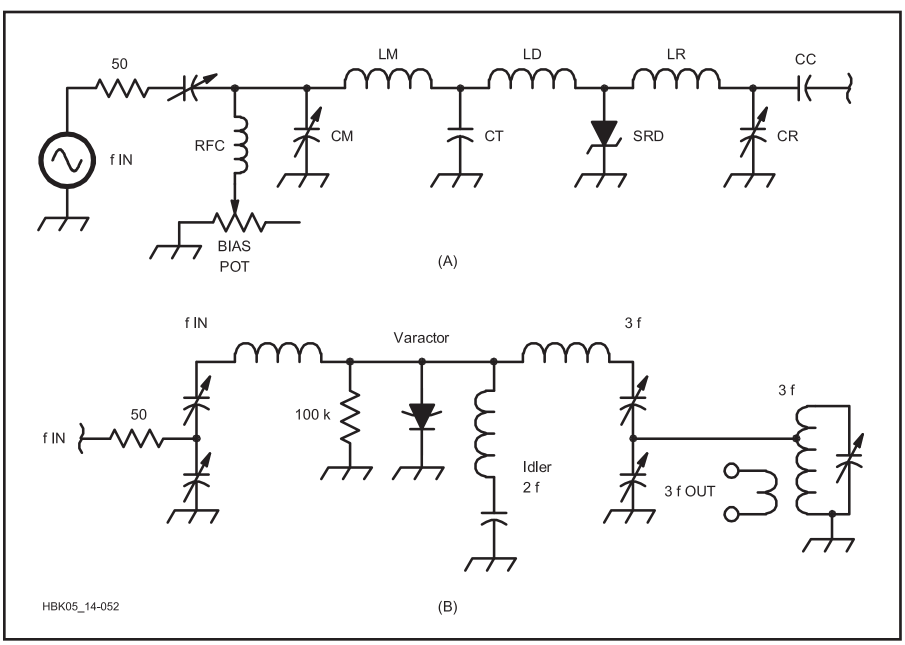


Fig 13.26 — Diode frequency multipliers. A: step-recovery diode multiplier. B: varactor diode multiplier.

nal 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.

For this example we will use the commonly-encountered dual-conversion scheme with the SSB generation at 455 kHz, a conversion to a 70 MHz IF and then a final conversion to the HF operating frequency. **Fig 13.27** is a block diagram of the system under consideration. Let's discuss the various elements in detail, starting at the microphone.

MICROPHONES

A microphone (mic) is a transducer that converts sound waves into electrical signals. For communications quality speech, its frequency response should be as flat as possible from around 200 to above 3500 Hz. Response peaks in the microphone can increase the peak to average ratio of speech, which then degrades (increases) the peak to average ratio of the transmitted signal. Some transmitters use "speech processing", which is essentially a specialized form of speech amplification either at audio or IF/RF. Since most microphones pick up a lot of background ambient noises, the output of the transmitter due to background noise pickup in the absence of speech may be as much as 20 dB greater than without speech processing. A *noise canceling* microphone is recommended to reduce this background pickup if there is much background noise. Microphone output levels vary, depending on the microphone type. Typical amateur microphones produce about 10 to 100 mV output levels.

Ceramic

Ceramic mics have high output impedances but low level outputs. They require a high-resistance load (usually about 50 kΩ) for flat frequency response and lose low-frequency response as this resistance is reduced (electrically, the mic "looks like" a small capacitor). These mics vary widely in quality, so a cheap mic is not a good bargain because of its effect on the transmitted power level and generally poor speech quality.

Dynamic

A dynamic mic resembles a small loudspeaker, with an impedance of about 600 Ω and an output of about 12 mV on voice peaks. Early dynamic mics (designed for vacuum tube transmitters) included a built-in transformer to transform the impedance to 100 kΩ suitable for high input impedance speech amplifiers. Currently available dynamic mics provide the output directly, although the transformers are available for connection outside the mic. Dynamic mics are widely used by amateurs.

Electret

Electret mics use a piece of special insulator material, such as Teflon, that contains a "trapped" polarization charge (Q) at its surfaces to create a capacitance (C). Sound waves modulate the capacitance of the material and cause a voltage change according to $\Delta V/V = -\Delta C/C$. For small changes in capacitance the change in voltage is almost linear. These mics have been greatly improved in recent years and are used in most cellular handsets.

The output level of the electret is fairly low and an integrated preamp is generally included in the mic cartridge. A voltage of about 4 V dc is required to power the preamp, and some commercial transceivers provide this voltage at the mic connector. The dc voltage must be blocked by a coupling capacitor if a dynamic mic element is to be used with a transceiver that supplies power to the microphone. The dynamic mic is unlikely to be damaged by the applied voltage, but the usual symptom is very low audio output with a muffled sound.

In recent years, micro electro-mechanical systems (MEMS) technology has been ap-

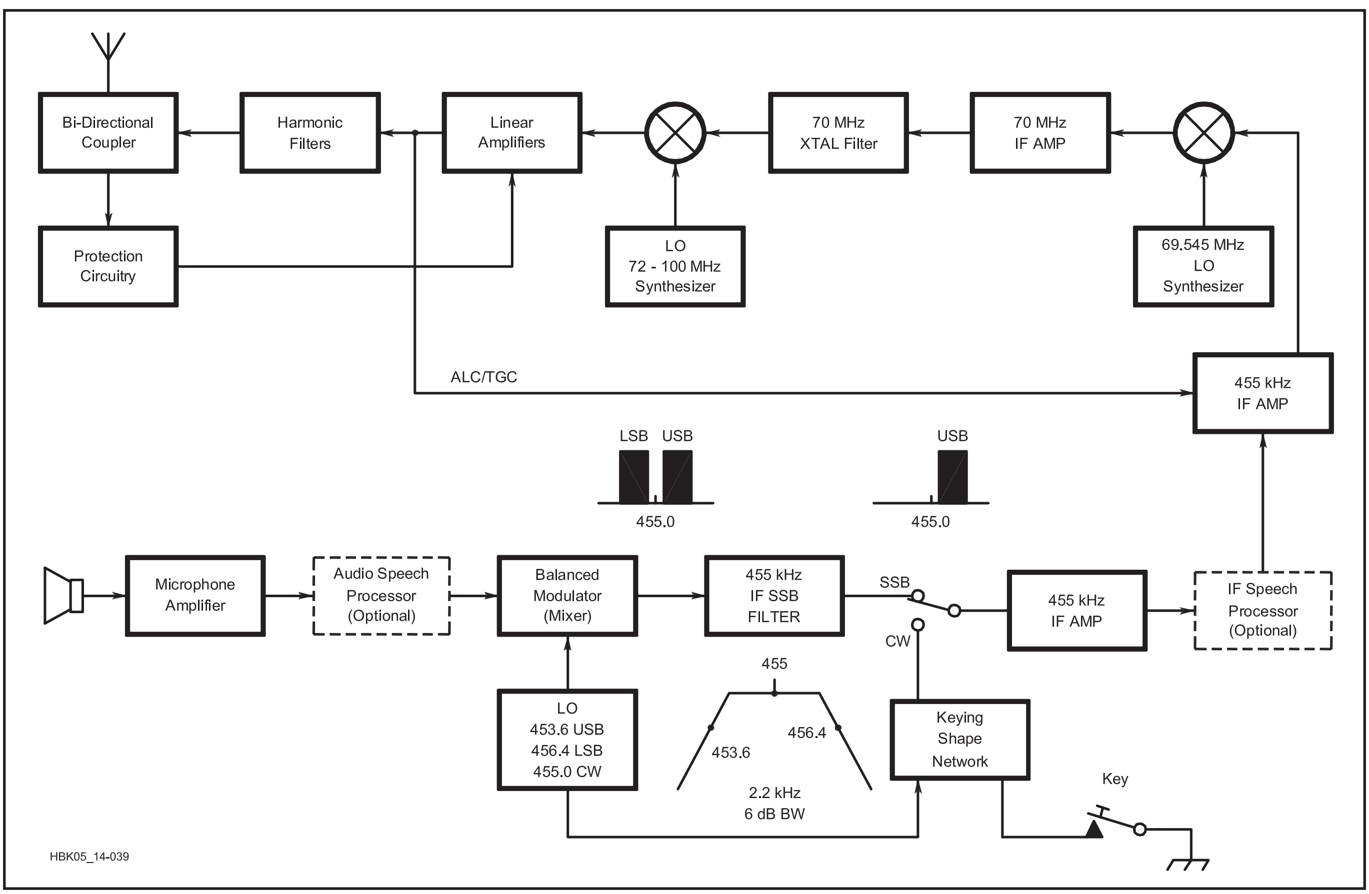


Fig 13.27 — Block diagram of an upconversion SSB/CW transmitter.

plied to microphones. The sound pressure is applied to a small transducer integrated on a silicon chip, which creates a mechanical motion in response to the force of the sound waves. This motion is converted into an electrical signal, usually by the varying capacitance of the transducer. Mass production of MEMS microphones is being driven by the cellular handset market.

MICROPHONE AMPLIFIERS

The balanced modulator and (or) the audio speech processor need a certain optimum level, which can be in the range of 0.3 to 0.6 V ac into perhaps 1 kΩ to 10 kΩ. Excess noise generated within the microphone amplifier should be minimized, especially if speech processing is used. The circuit in **Fig 13.28** uses a low-noise BiFET op amp. The 620 Ω resistor is selected for a low impedance microphone, and switched out of the circuit for high-impedance mics. The amplifier gain is set by the 100 kΩ potentiometer.

It is also a good idea to experiment with the low-and high-frequency responses of the mic amplifier to compensate for the frequency response of the mic and the voice of the operator.

SPEECH PROCESSING

The output of the speech amplifier can be applied directly to the balanced modulator. The resulting signal will be reproducible with the maximum fidelity and dynamic range available for the bandwidth provided. The communications efficiency of the system will depend on the characteristics of the particular voice used. The usual peak to average ratio is such that the average transmitted power is on the order of 5% of the peak power.

For communications use, it is generally beneficial to increase the average-to-peak ratio by distorting the speech waveform in a measured way. If one were to merely increase the amplifier gain, a higher average power would be obtained. The problem is that clipping, heavy distortion and likely spurious signals beyond legal limits would be generated. By carefully modifying the speech waveform before application to the balanced modulator, a synthetic waveform with a higher average-to-peak waveform can be generated that will retain most of the individual voice characteristics and avoid spurious signals. We will discuss two such techniques of *speech processing* below and another as we move into the RF circuitry.

Audio Speech Clipping

If the audio signal from the microphone amplifier is further amplified, say by as much as 12 dB, and then if the peaks are clipped (sometimes called *slicing* or *limiting*) by 12 dB by a speech clipper, the output peak value is the same as before the clipper, but the average value is increased considerably. The resulting signal contains harmonics and IMD but the speech intelligibility, especially in a white-noise background, is improved by 5 or 6 dB.

The clipped waveform frequently tends to have a square-wave appearance, especially on voice peaks. It is then band-pass filtered to remove frequencies below 300 and above 3000 Hz. The filtering of this signal can create a "repeaking" effect. That is, the peak value tends to increase noticeably above its clipped value.

An SSB generator responds poorly to a square-wave audio signal, creating significant peaks in the RF envelope. (This is described mathematically as the Hilbert Transform effect.) These peaks cause out-of-band splatter in transmitter's linear output power amplifier unless Automatic Level Control (ALC, to be discussed later) cuts back on the RF gain. The peaks increase the peak-to-average ratio and the ALC reduces the average SSB power output, thereby reducing some of the benefit of the speech processing. The square-wave effect is also reduced by band-pass filtering (300 to 3000 Hz) the input to the clipper as well as the output.

Fig 13.29 is a circuit for a simple audio speech clipper. A CLIP LEVEL potentiometer before the clipper controls the amount of clipping and an OUTPUT LEVEL potentiometer controls the drive level to the balanced modulator. The correct adjustment of these potentiometers is done with a two-

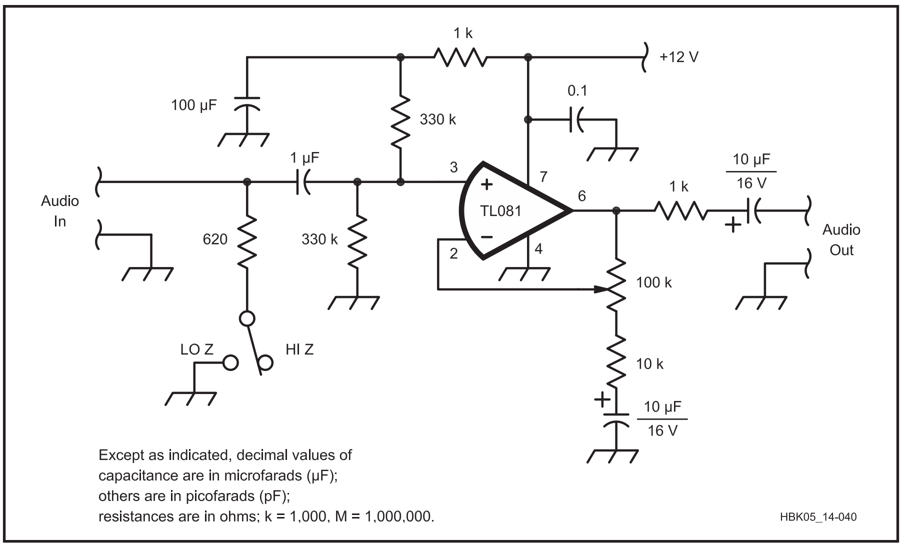


Fig 13.28 — Schematic diagram of a microphone amplifier suitable for high or low impedance microphones.

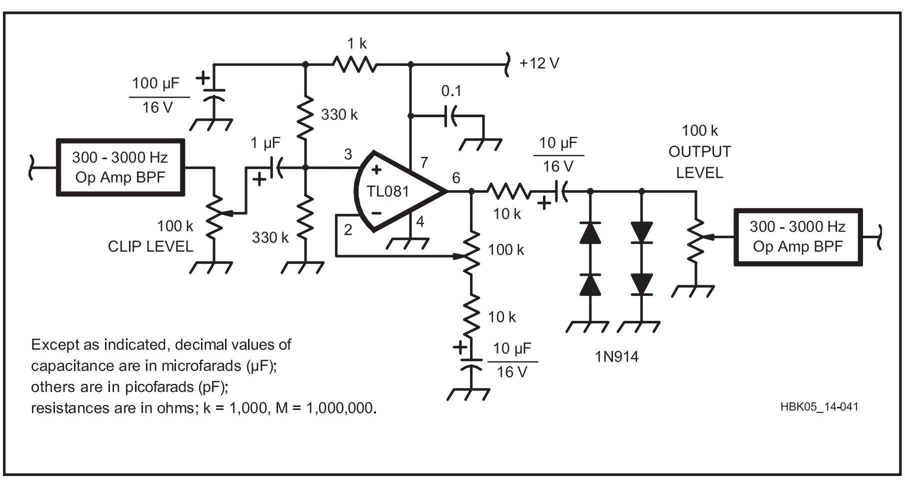


Fig 13.29 — Schematic diagram a simple audio speech clipper.

tone audio input or by talking into the microphone, rather than driving with a single tone, because single tones don't exhibit the repeaking effect.

Audio Speech Compression

Although it is desirable to keep the voice level as high as possible, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use an automatic gain control that follows the average variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the resultant dc to a control terminal in an early stage of the amplifier.

If an audio AGC circuit derives control voltage from the output signal, the system is a closed loop. If short attack time is necessary, the rectifier-filter bandwidth must be opened up to allow syllabic modulation of the control voltage. This allows some of the voice frequency signal to enter the control terminal, causing distortion and instability. Because the syllabic frequency and speech-tone frequencies have relatively small separation, the simpler feedback AGC systems compromise fidelity for fast response.

Problems with loop dynamics in audio AGC can be side-stepped by eliminating the loop and using a forward-acting system. The control voltage is derived from the input of the amplifier, rather than from the output. Eliminating the feedback loop allows unconditional stability, but the trade-off between response time and fidelity remains. Care must be taken to avoid excessive gain between the signal input and the control voltage output. Otherwise the transfer characteristic can reverse; that is, an increase in input level can cause a decrease in output. A simple forward-acting compressor is shown in **Fig 13.30**.

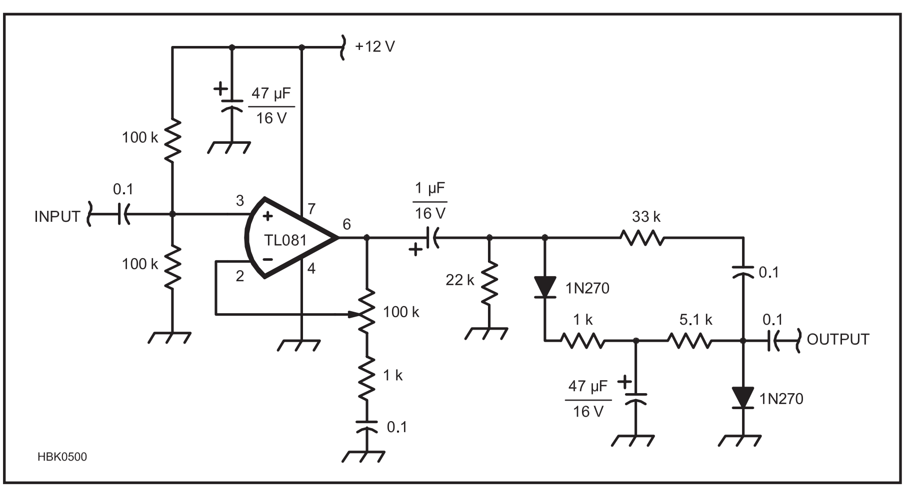


Fig 13.30 — A solid state forward acting speech compressor circuit.

BALANCED MODULATORS

A balanced modulator is a mixer. A more complete discussion of balanced modulator design was provided in the **Mixers, Modulators, and Demodulators** chapter. Briefly, the IF frequency LO (455 kHz in the example of Fig 13.25) 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 Fig 13.27 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 Fig 13.27. 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.

In times past, balanced modulators using diodes, balancing potentiometers and numerous components were used. These days it doesn't make sense to use this approach. ICs and packaged diode mixers do a much better job and are less expensive. The most widely known modulator IC, the MC1496, has been around for more than 25 years and is still one of the best and least expensive. **Fig 13.31** is a typical balanced modulator circuit using the MC1496.

The data sheets for balanced modulators and mixers specify the maximum level of audio for a given LO level. Higher audio levels create excessive IMD. The IF filter 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.

IF FILTERS

The desired IF filter response is shown in **Fig 13.32A**. 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 **RF and AF Filters** chapter) are frequently called "single-sideband" filters because they also have this property. The steep skirt can be on the low side or the high side, depending on whether the crystals are across the signal path or in series with the signal path, respectively.

Filters require special attention to their terminations. The networks that interface the filter with surrounding circuits should be accurate and stable over temperature. They should be easy to adjust. One very good way to adjust them is to build a narrow-band sweep generator and look at the output IF envelope with a logarithmic amplifier, as indicated in Fig 13.32B. There are three goals:

- The driver stage must see the desired load impedance.
- The stage after the filter must see the desired source (generator) impedance.
- The filter must be properly terminated at both ends.

Lack of any of these conditions will result in loss of specified filter response. Fig 13.32B shows two typical approaches. This kind of setup is a very good way to make sure the fil-

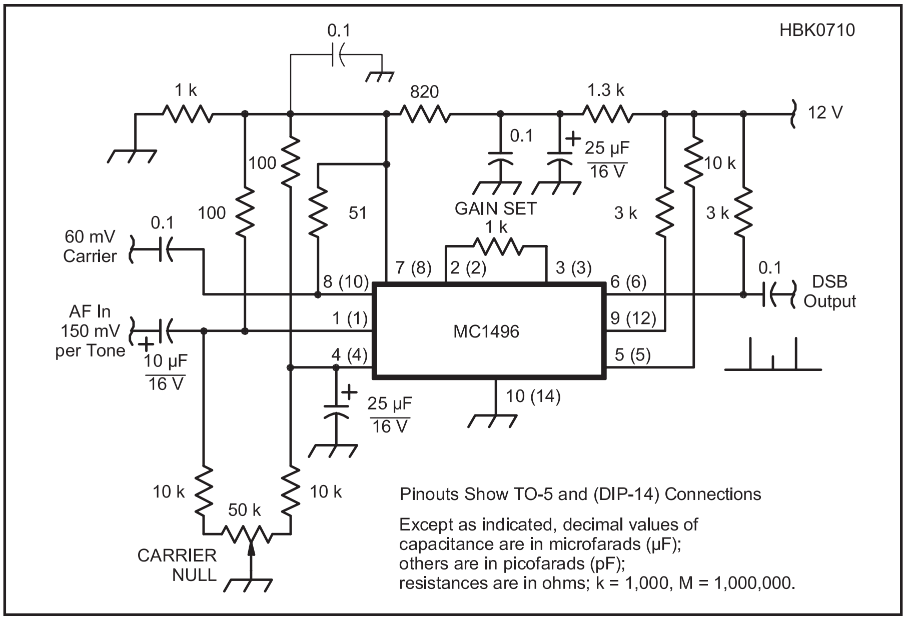


Fig 13.31 — An IC balanced modulator circuit using the MC1496 IC. The resistor between pins 2 and 3 sets the subsystem gain.

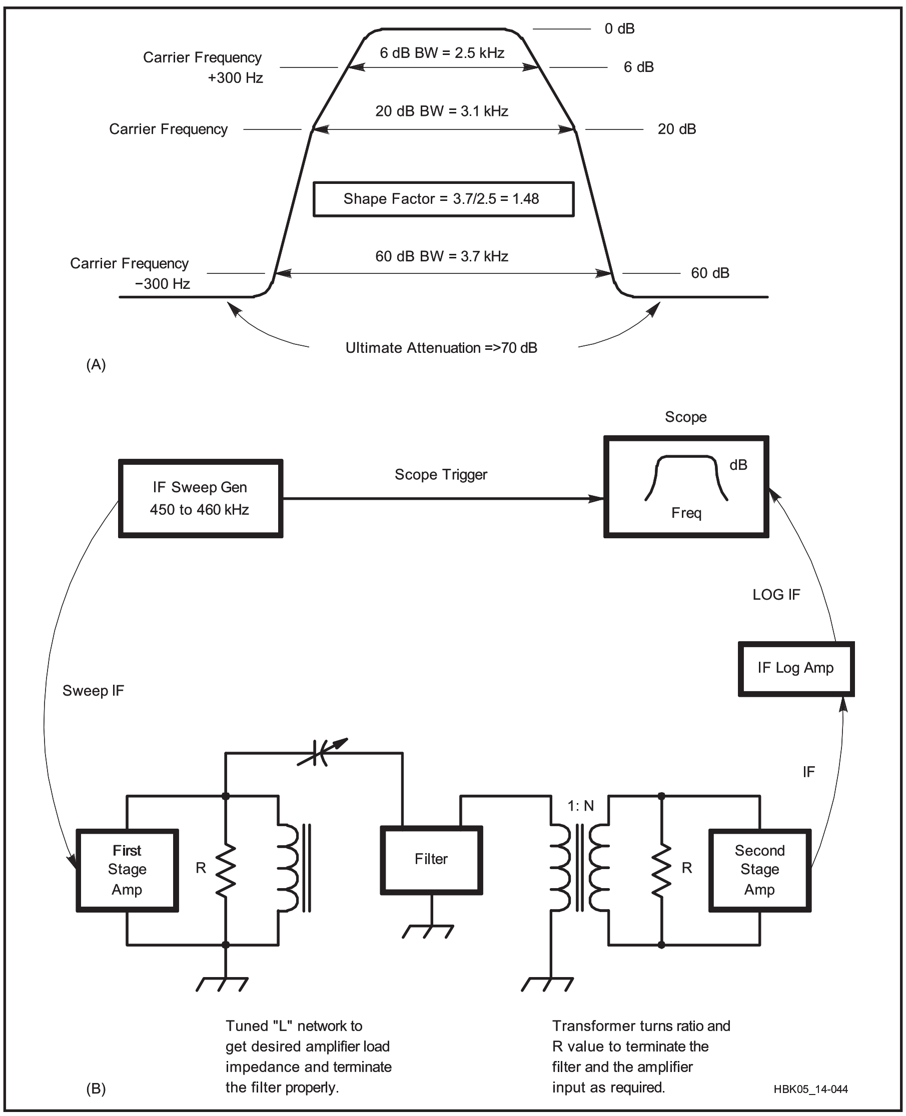


Fig 13.32 — At (A), desired response of a SSB IF filter. At (B), one method of terminating a mechanical filter that allows easy and accurate tuning adjustment and also a possible test setup for performing the adjustments.

ters 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 SPEECH CLIPPER

Audio clipper speech processors generate a considerable amount of in-band harmonics and IMD (involving different simultaneously occurring speech frequencies). The total distortion detracts somewhat from speech intelligibility. Other problems were mentioned in the earlier section on speech processing. IF clippers overcome most of these problems, especially the Hilbert Transform problem.[7]

Fig 13.33A 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.

Fig 13.33B 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.

Fig 13.34 shows oscilloscope pictures of an IF clipped two-tone signal at various levels of clipping.[8] The level of clipping in a radio can be estimated by comparing with these photos. Listening tests verify that the IMD does not sound nearly as bad as harmonic distortion. In fact, processed speech sounds relatively clean and crisp. Tests also verify that speech intelligibility in a noise background is improved by 8 dB.

The repeaking effect from band-pass filtering the clipped IF signal occurs, and must be accounted for when adjusting the output level. A two-tone audio test signal or a speech signal should be used. The ALC circuitry (discussed later) will 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. Fig 13.33B shows the basic method. The mic amplifier and the MC1496 circuits have been previously shown and the clipper circuit

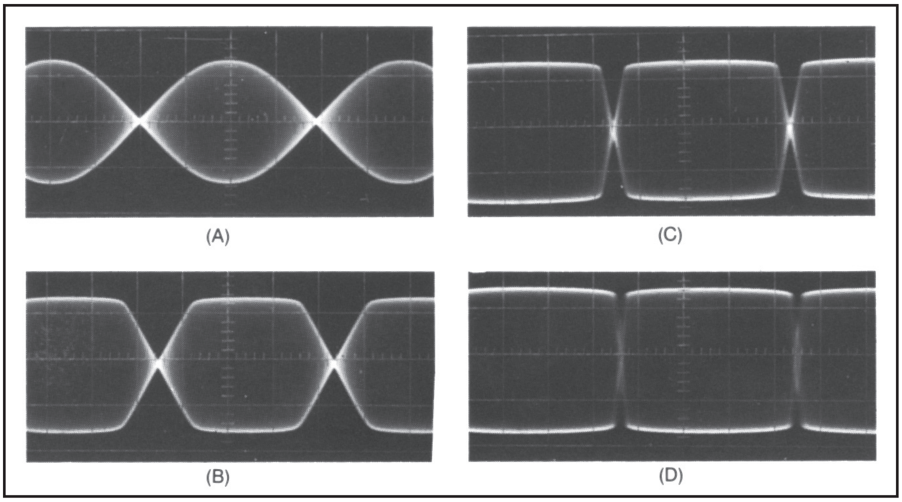


Fig 13.34 — Two-tone envelope patterns with various degrees of RF clipping. All envelope patterns are formed using tones of 600 and 1000 Hz. At A, clipping threshold; B, 5 dB of clipping; C, 10 dB of clipping; D, 15 dB of clipping (from).

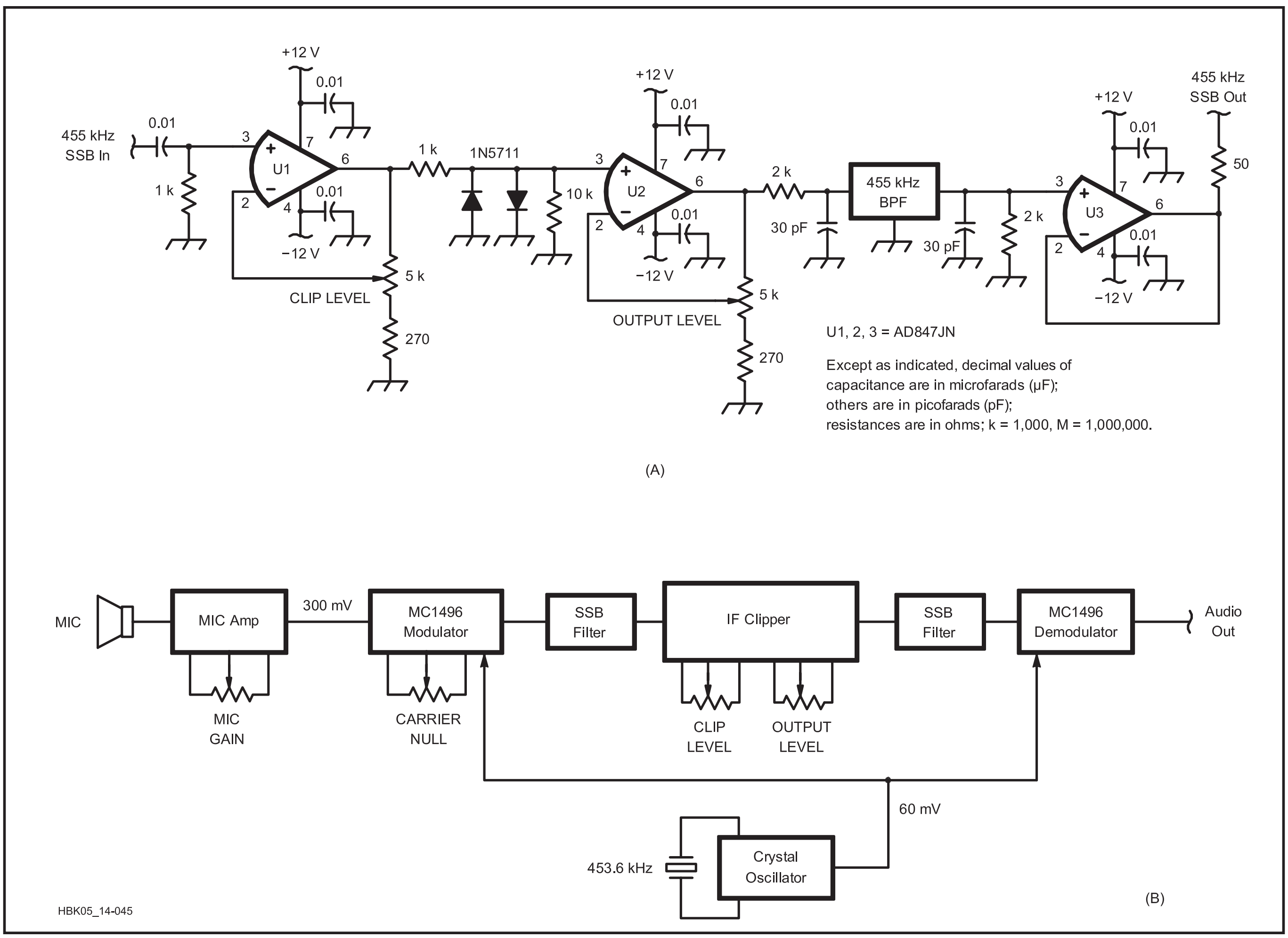


Fig 13.33 — IF speech clipping. At (A), schematic diagram of a 455 kHz IF clipper using high-frequency op amps. At (B) block diagram of an adaptation of the above system to an audio in-audio out configuration.

can be the same as in Fig 13.33A.

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.

IF Linearity and Noise

Fig 13.27 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 Fig 13.27, 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 distortion generated by the PA and also less than the phase noise generated by the final mixer. Recall also that the total IMD generated in the exciter stages is the resultant of several amplifier and mixer stages in cascade; therefore, each element in the chain must have at least 40 to 50 dB IMD quality. The various drive levels should be chosen to guarantee this. This requirement for multistage linearity is one of the main technical and cost burdens of the SSB mode.

Of interest also in the gray region are 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.

13.3.7 CW Mode Operation

Radiotelegraph or CW operation can be easily obtained from the transmitter architecture design shown in Fig 13.27. 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 Fig 13.27 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.

RF ENVELOPE SHAPING

On-off keying (CW) is a special kind of low-level amplitude modulation (a low signal-level stage is turned on and off). It is special because the sideband power is subtracted from the carrier power, and not provided by a separate "modulator" circuit, as in high-level AM. It creates a spectrum around the carrier frequency whose amplitude and bandwidth are influenced by the rates of signal amplitude rise and fall and by the curvature of the keyed waveform. Refer to the discussion of keying speed, rise and fall times, and bandwidth in the **Modulation** chapter and the earlier discussion in this chapter for some information about this issue.

Now look at **Fig 13.35**. 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.

For a narrow spectrum and freedom from adjacent channel interference, a further requirement is that the spectrum must fall off very rapidly beyond the –20 dB bandwidth indicated in Fig 13.35. A sensitive narrowband CW receiver that is tuned to an adjacent channel that is only 1 or 2 kHz away can detect keying sidebands that are 80 to 100 dB below the key-down level of a strong CW signal. An additional consideration is that during key-up a residual signal, called "backwave" should not be noticeable in a nearby receiver. A backwave level at least 90 dB below the key-down carrier is a desirable goal.

Fig 13.36 is the schematic of one wave-

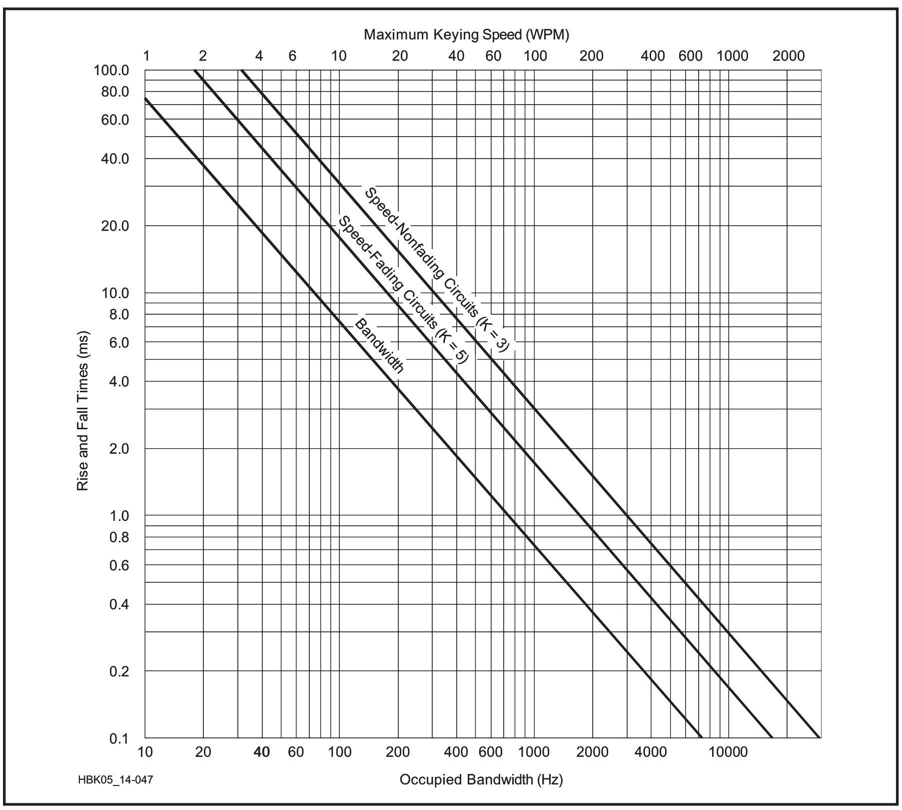


Fig 13.35 — Keying speed versus rise and fall times versus bandwidth for fading and nonfading communications circuits. For example, for transmitter output waveform rise and fall times of approximately 6 ms, draw a horizontal line from 6.0 ms on the rise and fall times scale to the bandwidth line. Then 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.

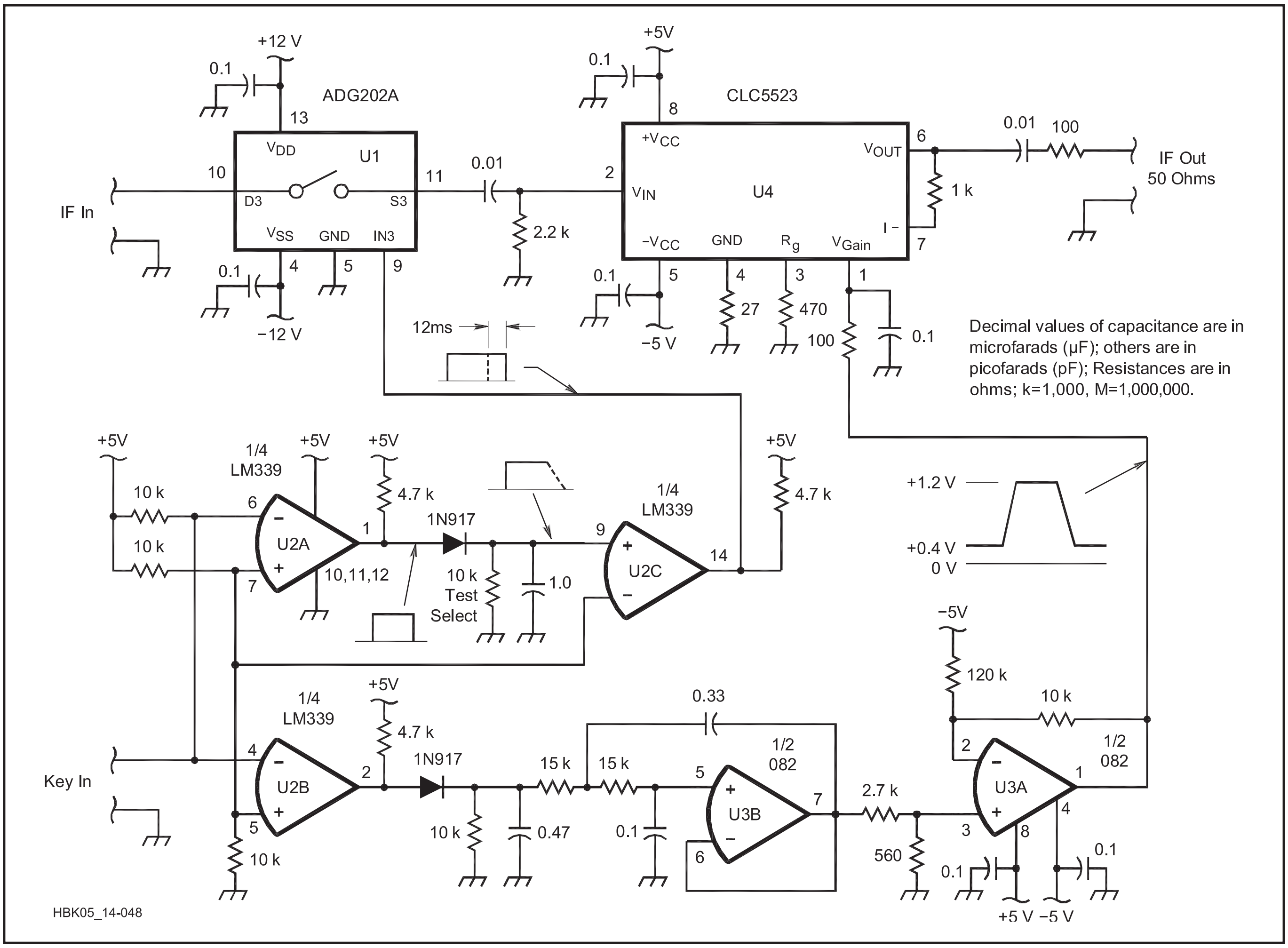


Fig 13.36 — This schematic diagram shows a CW waveshaping and keying circuit suitable for use with an SSB/CW transmitter such as is shown in Fig 13.27.

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

The keying waveform is applied to the gain control pin of a CLC5523 amplifier IC. This device, like nearly all gain-control amplifiers, has a *logarithmic* control of gain; therefore some experimental "tweaking" of the capacitor values was used to get the result shown in **Fig 13.37A**. The top trace shows the on/off operation of the IF switch, U1. The signal is turned on shortly before the rise of the keying pulse begins and remains on for about 12 ms after the keying pulse is turned off, so that the waveform falls smoothly to a very low value. The result is an excellent spectrum and an almost complete absence of backwave. Compare this to the factory transmitter waveshapes shown in Fig 13.13 and 13.15. The bottom trace shows the resulting keyed RF output waveshape. It has an excellent spectrum, as verified by critical listening tests. The thumps and clicks that are found in some CW transmitters are virtually absent. The rise and fall intervals have a waveshape that is approximately a cosine. Spread-spectrum frequency-hop waveforms have used this approach to minimize wideband interference.

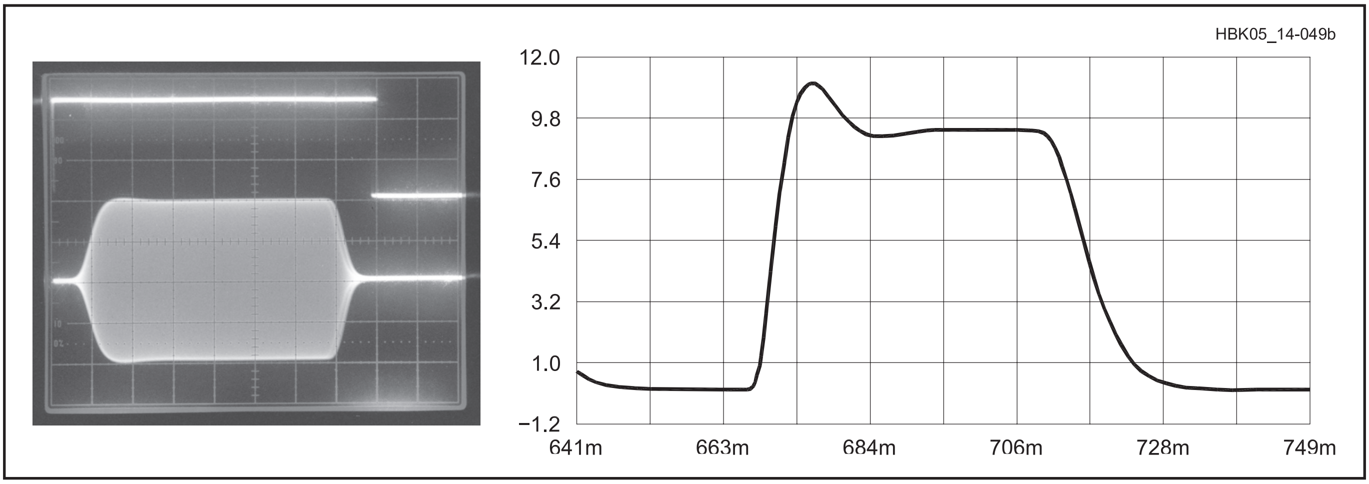
Fig 13.37B is an accurate *SPICE* simulation of the wave shaping circuit output before the signal is processed by the CLC5523 amplifier. To assist in adjusting the circuit, create a steady stream of 40 ms dots that can be seen on an RF oscilloscope that is looking at the final PA output envelope. It is important to make sure that the excellent waveshape is not degraded on its way to the transmitter output. Single-sideband linear power amplifiers are well suited for a CW transmitter, but they must stay within their linear range, and the backwave problem must be resolved.

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

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

13.3.8 Wideband Noise

In the block diagram of Fig 13.27 the last mixer and the amplifiers after it are wideband circuits that are limited only by the harmonic filters and by any selectivity that may be in the antenna system. Wide-band phase noise transferred onto the transmitted modulation by the last LO can extend over a wide fre-



(A) **(B)**

Fig 13.37 — 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 Fig 13.36. 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.

quency range, therefore LO (almost always a synthesizer of some kind) cleanliness is always a matter of great concern.[9]

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.

Transmitters that are used in close proximity to receivers, such as on ships, are always designed to control wideband emissions of both additive noise and phase noise, referred to as "composite" noise.

TRANSMIT MIXER SPURIOUS SIGNALS

The last IF and the last mixer LO in Fig 13.27 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.

13.3.9 Automatic Level Control (ALC)

The purpose of ALC is to prevent the various stages in the transmitter from being over-driven. Over-drive can generate too much out-of-band distortion or cause excessive power dissipation, either in the amplifiers or in the power supply. ALC does this by sampling the peak amplitude of the modulation (the envelope variations) of the output signal and then developing a dc gain-control voltage that is applied to an early amplifier stage, as suggested in Fig 13.27.

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

SPEECH PROCESSING WITH ALC

A fast response to the leading edge of the modulation is needed to prevent a transient overload. After a strong peak, the control voltage is "remembered" for some time as the voltage in a capacitor. This voltage then decays partially through a resistor between peaks. An effective practice provides two capacitors and two time constants. One capacitor decays quickly with a time constant of, say 100 ms, the other with a time constant of several seconds. With this arrangement a small amount of speech processing, about 1 or 2 dB, can be obtained. (Explanation: The dB of improvement mentioned has to do with the improvement in speech intelligibility in a random noise background. This improvement is equivalent to what could be achieved if the transmit power were increased that same number of dB).

The gain rises a little between peaks so that weaker speech components are enhanced. But immediately after a peak it takes a while for the enhancement to take place, so weak components right after a strong peak are not enhanced very much. **Fig 13.38A** shows a complete ALC circuit that performs speech processing.

ALC IN SOLID-STATE POWER AMPLIFIERS

Fig 13.38B shows how a dual directional coupler can be used to provide ALC for a solid-state power amplifier (PA). The basic idea is to protect the PA transistors from excessive SWR and dissipation by monitoring both the forward power and the reflected power.

TRANSMIT GAIN CONTROL (TGC)

This is a widely used feature in commercial and military equipment. A calibrated "tune-up" test carrier of a certain known level is applied to the transmitter. The output carrier level is sampled, using a diode detector. The resulting dc voltage is used to set the gain of a low-level stage. This control voltage is digitized and stored in memory so that it is semi-permanent. A new voltage may be generated and stored after each frequency change, or the stored value may be used. A test signal is also used to do automatic antenna tuning. A dummy load is used to set the level and a low-level signal (a few mW) is used for the antenna tune-up.

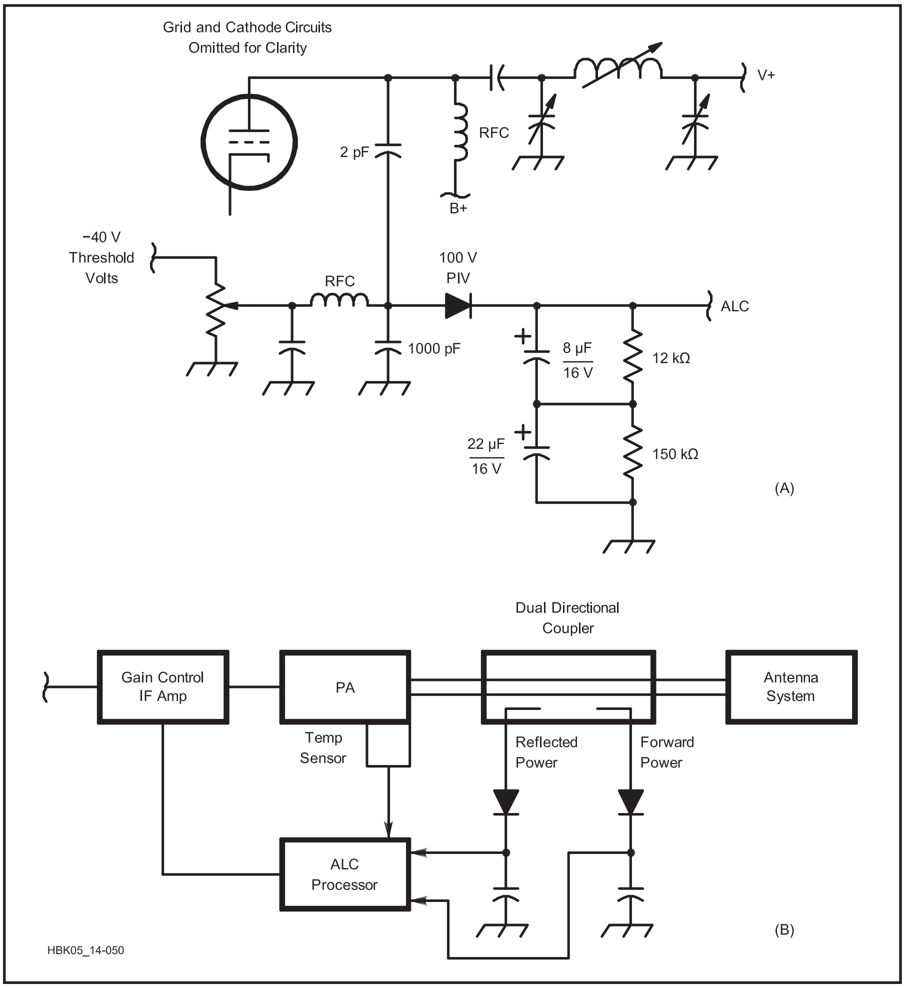


Fig 13.38 — At (A), an ALC circuit with speech processing capability. At (B), protection method for a solid-state transmitter.

— a simple battery-operated transmitter intended to keep in touch with the folks back home on an HF net frequency. The crystal oscillator in the transmitter functions as a calibrator for the receiver, so that the receiver may be tuned around the band and then easily reset to the net frequency. The current drain during receive is only 60 mA, because the entire transmitter is turned off except when actually transmitting.

EXCITER BLOCK DIAGRAM

Fig 13.40 is the block diagram of the circuitry on the PC board and **Fig 13.41** is a detailed schematic. It is a complete VXO SSB exciter on a 2.5 × 3.8 inch circuit board with 1 mW (0 dBm) peak output. The exciter uses the phasing method of SSB generation, which makes it easy to operate on different frequencies. In a phasing SSB exciter, two identical signals with a 90° phase difference are generated and then combined so that one sideband adds and the other subtracts. The signal quality from this exciter is not merely adequate for the HF amateur bands — it is exceptional.

RF Circuitry

The frequency generator consists of a VXO, buffer amplifier and quadrature hybrid. It has a lot of parts: three transistors; a voltage regulator IC, a Zener diode, four toroids and many resistors and capacitors. There are no adjustments. You just build it and it works. The frequency stability is better than that of most commercial radios, even when portable.

There are simpler VXO circuits, but this one is excellent. It provides 0 dBm sine wave

Project: The MicroT2 — A Compact Single-Band SSB Transmitter

As an example of an SSB transmitter including many aspects of design covered heretofore, we present the MicroT2, a simple SSB transmitter that generates a high-quality USB or LSB signal on any single band from 1.8 MHz to 50 MHz. Rick Campbell, KK7B, developed the MicroT2 as a companion to the MicroR2 receiver project described in the **Receivers** chapter. While it is a bit more involved to generate an SSB signal than a CW signal, we greatly simplify the task if all the necessary circuitry is on a single PC board exciter module. Once we have a high-quality low-level SSB signal, a 5 or 500 W SSB transmitter is as easy to build as a 5 or 500 W CW transmitter. Simple transmitters are delightful, but relaxed standards are not. The MicroT2 is designed to be clean, stable and reliable, exceed FCC Part 97 requirements, and sound good, too. A more complete description of the circuitry in this transmitter can be found in *Experimental Methods in RF Design*[10] *(EMRFD)* of which the project's designer was a co-author.

THE CASE FOR CRYSTAL CONTROL

A jumper on the PC board makes it easy to use an external VFO for frequency control if you wish. For some applications the narrow tuning range offered by the onboard VXO is sufficient, and the secure knowledge that the transmitter is actually on frequency and stable is a real virtue. **Fig 13.39** shows just one example

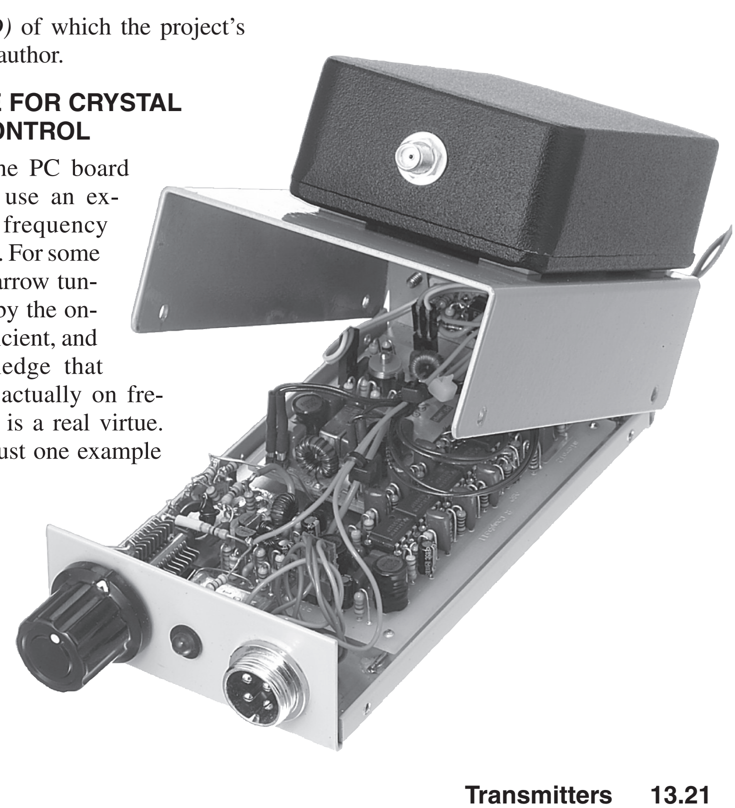


Fig 13.39 — This 40 meter version of the MicroT2 uses the onboard VXO. The black box on top is the 0.5 W amplifier.

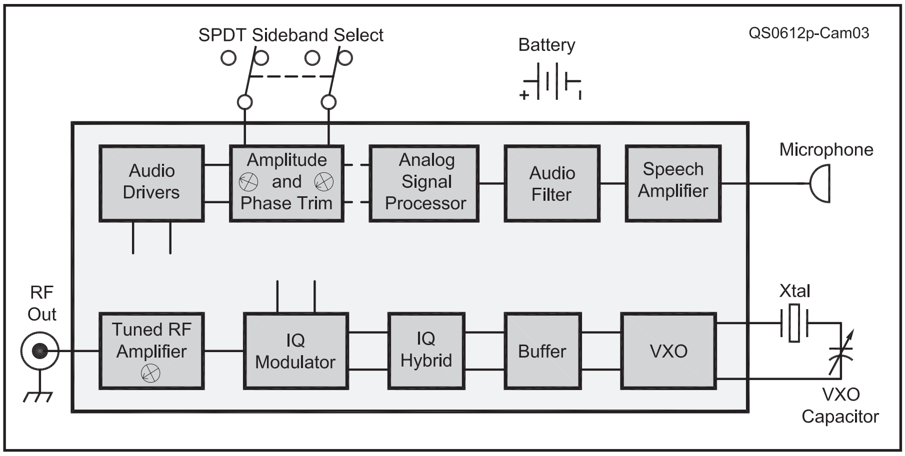


Fig 13.40 — Block diagram of the circuitry on the PC board.

output, draws 4 mA, and has virtually no start-up drift — about 2 Hz at 7 MHz. (It makes a lovely keyed CW generator too.) The three-resistor attenuator between the output of the VXO and input of the buffer amplifier is a convenient place to insert a frequency multiplier or externally generated VFO. Just leave off the top resistor to break the path. Note that the crystal and variable capacitor are mounted off the PC board. Frequency stability is determined by the temperature of the crystal. Slip a foam packing bead over the

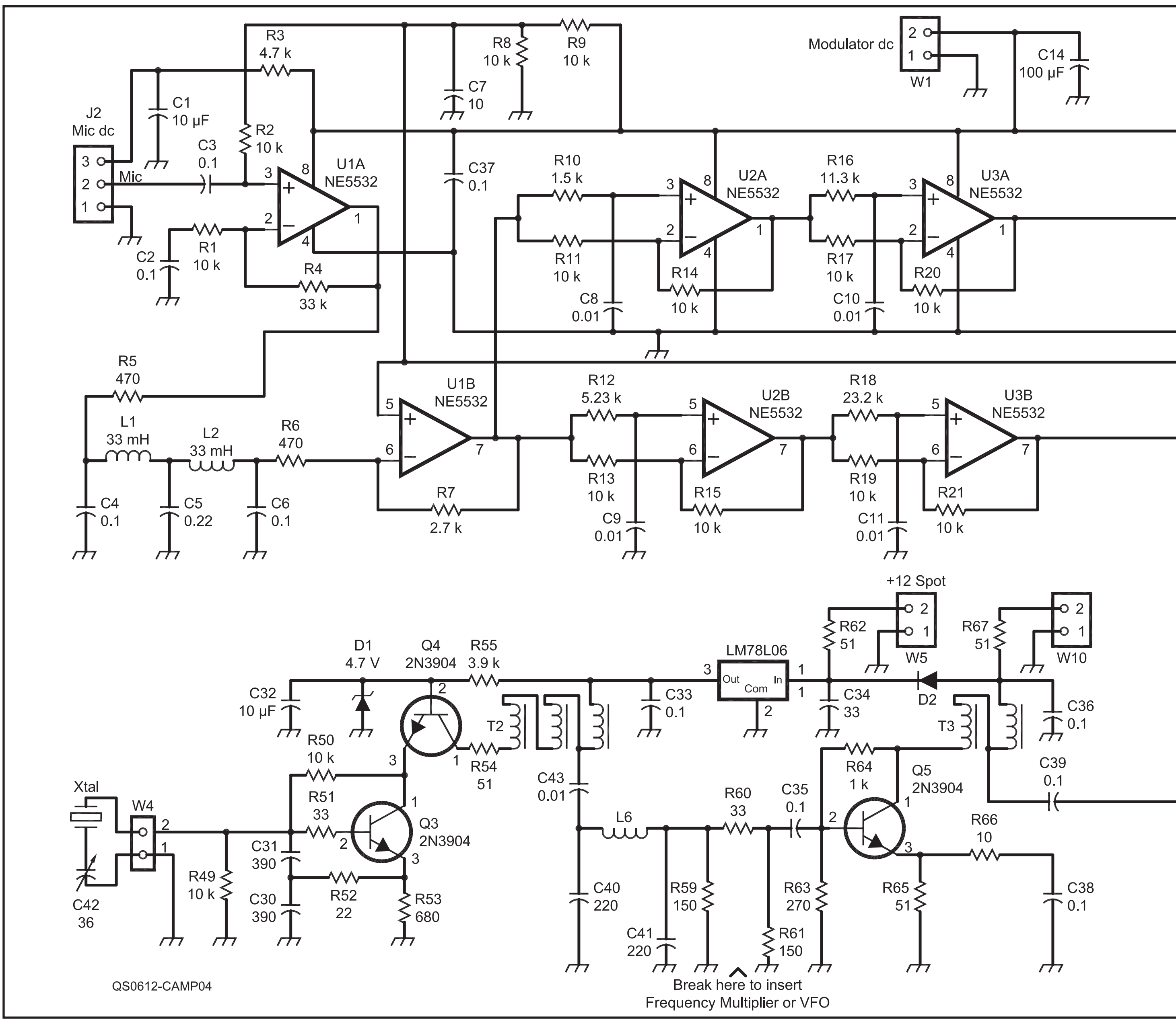


Fig 13.41 — Schematic diagram of the 40 meter version of the MicroT2. See Table 13.1 for parts values. A kit version including parts and PC board is available from Kanga US (www.kangaus.com).

crystal and support it by its leads between the PC board and VXO capacitor. A front-panel-mounted crystal socket looks very classy, but noise picked up by the crystal body gets into the oscillator, and the signal radiated by the crystal sounds like one with poor carrier suppression. If you use a crystal socket, put it inside the case or behind a shield door.

The buffer amplifier provides a 50 Ω source of broadband drive to the quadrature hybrid and isolates the frequency generator from impedance variations at the mixer local oscillator (LO) ports. The simple arrangement used in the MicroR2 receiver doesn't work here, because high-level audio into the diode ring mixers modulates not just the signal at the RF port, but the impedance at the LO and IF ports as well. Experiments confirm that a directly connected VFO experiences severe frequency pulling on voice peaks. The gain of the buffer is set to provide +7 dBm drive to each mixer with an input level of 0 dBm. It draws some current and the transistor gets a little warm — but only while transmitting.

The quadrature hybrid is a venerable circuit first described by Reed Fisher in *QST*.[11] It is the lumped element equivalent of a pair of tightly coupled quarter-wavelength transmission lines. The total capacitance does not need to be symmetrical between the two ends of the inductor, as is commonly shown. It may all be at one end or divided unequally. When driven from a 50 Ω source and terminated in a 51 Ω resistor and two mixer LO ports, the 90° phase difference is nearly perfect across a wide band and the amplitude difference between the two outputs is within 0.5 dB across a 10% bandwidth — more than enough bandwidth to cover the usual SSB portion of any amateur band.

A pair of Mini-Circuits TUF-3 mixers serves as the I and Q balanced modulators. These provide good carrier suppression without adjustment, and reasonable output at a modest distortion level. The carrier

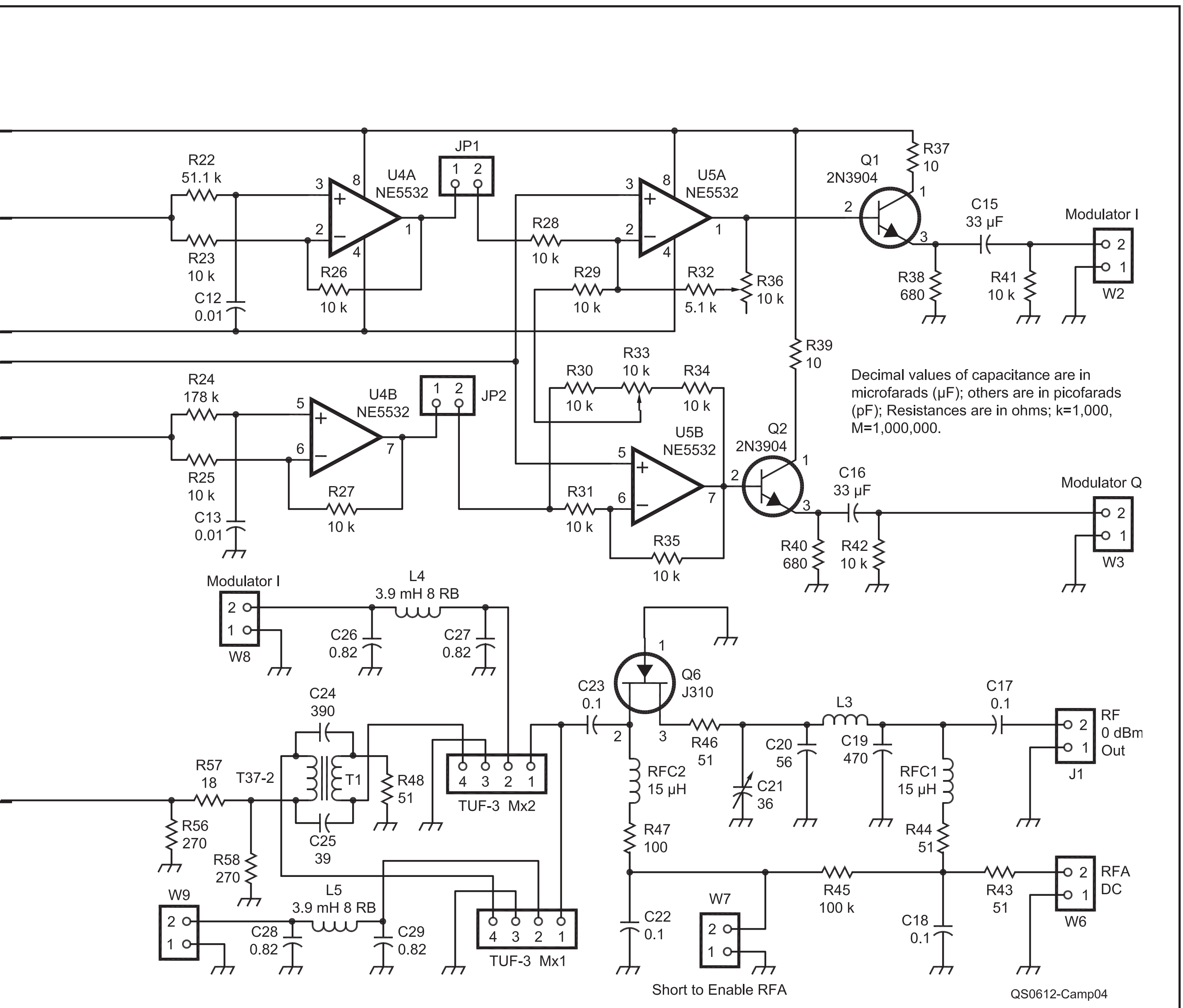


Table 13.1
MicroT2 Parts List

Parts for 40 meter version. See Note A.

C1, C7, C32, C34 — 10 µF, 16 V electrolytic capacitor.
C2-C4, C6, C17, C18, C22, C23, C33, C35-C39 — 0.1 µF, 5% polyester capacitor.
C5 — 0.22 µF, 5% polyester capacitor.
C8-C13 — 0.01 µF polyester capacitor, matched to 1%.
C14 — 100 µF, 16 V electrolytic capacitor.
C15, C16 — 33 µF, 16 V electrolytic capacitor.
C19 — 470 pF NP0 ceramic capacitor.
C20 — 56 pF NP0 ceramic capacitor.
C21 — 3-36 pF poly trimmer capacitor.
C24, C30, C31 — 390 pF, NP0 ceramic capacitor.
C25 — 39 pF, NP0 ceramic on back of board capacitor.
C26-C29 — 0.82 µF, 5% polyester capacitor.
C40, C41 — 220 pF, NP0 ceramic capacitor.
C42 — 36 pF, VXO variable off board capacitor; see Note C.
C43 — 0.01 µF ceramic capacitor.
D1 — 4.7 V Zener.
D2 — 1N4148 diode.
L1, L2 — 33 mH inductor.
L3 — 40 turns #30 enameled wire on T37-2 toroid core; see Note D.
L4, L5 — 3.9 mH inductor.
L6 — 22 turns #28 enameled wire on T30-2 toroid core; see Note D.
Mx1, Mx2 — Mini-Circuits TUF-3 diode ring mixer.
Q1-Q5 — 2N3904 transistor.
Q6 — J310 field effect transistor.
RFC1, RFC2 — 15 µH molded RF choke.
R1, R2, R8, R9, R28-R31, R34, R35, R41, R42, R49, R50 — 10 kΩ resistor.
R3 — 4.7 kΩ resistor.
R4 — 33 kΩ resistor.
R5, R6 — 470 Ω resistor.
R7 — 2.7 kΩ; audio gain select resistor; see Note E.
R10 — 1.5 kΩ, 1% resistor.
R11, R13-R15, R17, R19-R21, R23, R25-R27 — 10.0 kΩ, 1% resistor.
R12 — 5.23 kΩ, 1% resistor.
R16 — 11.3 kΩ, 1% resistor.
R18 — 23.2 kΩ, 1% resistor.
R22 — 51.1 kΩ, 1% resistor.
R24 — 178 kΩ, 1% resistor.
R32 — 5.1 kΩ resistor.
R33, R36 — 10 kΩ trimpot resistor.
R38, R40, R53 — 680 Ω resistor.
R37, R39, R66 — 10 Ω resistor.
R43, R44, R46, R48, R54, R62, R65, R67 — 51 Ω resistor.
R45 — 100 kΩ resistor.
R47 — 100 Ω resistor.
R51, R60 — 33 Ω resistor.
R52 — 22 Ω resistor.
R55 — 3.9 kΩ resistor.
R56, R58, R63 — 270 Ω resistor.
R57 — 18 Ω resistor.
R59, R61 — 150 Ω resistor.
R64 — 1 kΩ resistor.
T1 — 17 turns two colors #28 enameled wire bifilar wound on T37-2 toroid core; see Note D.
T2 — 5 turns #28 enameled wire trifilar wound on FT23-43 toroid core; see Note D.
T3 — 7 turns #28 enameled wire bifilar wound on FT23-43 toroid core; see Note D.
U1-U5 — NE5532 or equivalent dual low-noise high-output op-amp.
U6 — LM7806 or equivalent 6 V three terminal regulator.

Note A: C19, 20, 24, 25, 30, 31, 40, 41; L3, L6 and T1 values are for operation in the 40 meter band.
Note B: The total reactance of the parallel combination of C24 and C25 plus the capacitance between the windings of T1 is $-j50$ Ω at the center of the tuning range. Placing most of the capacitance at one end is a different but equivalent arrangement of the quadrature hybrid we often use with equal capacitors. C25 is only needed if there is no standard value for C24 within a few percent of the required value. C25 is tack soldered to the pads provided on the back of the PC board, and may be a surface mount component if desired.
Note C: Capacitor C42 is the VXO tuning capacitor for the exciter.
Note D: L3, L6, and T1, T2 and T3 are listed as number of turns on the specified core rather than a specific inductance. For those who wish to study the design with a calculator, simulator and inductance meter, L3 should be about $+j300$ Ω at 7.2 MHz, L6 should be $+j100$ Ω at 7.2 MHz and each winding of T1 should be $j50$ Ω at 7.2 MHz. T2 and T3 are noncritical broadband transformers with about 40 µH total inductance using the specified number of turns.
Note E: Resistor R7 sets the audio signal processor gain. If the gain is set too high, intermodulation distortion products generated in the diode ring modulators will be objectionable. Select R7 for a peak exciter output level no greater than 0 dBm.

suppression may be improved by soldering the metal cans of the TUF-3 mixers directly to the PC board ground. IM products in the opposite sideband are more than 30 dB down at the exciter output. The aggressive low-pass filtering right at the mixer IF ports prevents wideband noise and harmonic distortion in the audio stages from contributing to the I and Q modulation. Energy outside the modulation bandwidth is then just intermodulation distortion in the mixers and linear amplifiers. Sideband suppression will be more than 40 dB if the four 0.82 µF capacitors and two 3.9 mH inductors are matched to within 1%. The resultant carrier suppression is greater than 50 dB on any of the HF bands.

The exciter's output RF amplifier uses a common-gate JFET. This stage provides a broadband resistive termination to the mixer RF port IQ summing junction to isolate it from variations in impedance at the exciter output. The RF amplifier has relatively low gain, good harmonic suppression and a very clean 0 dBm SSB signal at the output. This is an appropriate level to drive a linear amplifier, balanced mixer or transverter. It is also a low enough level that it is easy to adjust the exciter with simple equipment. The exciter output signal meets FCC regulations for direct connection to an antenna for flea power experiments.

The transmitter shown here adds the simple two transistor circuit in **Fig 13.42** for 0.5 W PEP output. The seventh order Chebyshev low-pass filter on the output is noncritical and assures a clean signal that easily meets FCC regulations.

Audio Section

The top half of the PC board contains all of the audio circuitry. There are no PC board traces connecting the two halves, and it is okay to cut the board into separate functional blocks for packaging flexibility, or to use the audio and RF portions of the circuitry in other projects.

The speech amplifier drives a passive low-pass filter using two series inductors and three shunt capacitors. The combination of this low-pass filter and the mixer IF port filters limits speech frequencies to just over 3 kHz for natural sounding speech and good spectral purity. There is no ripple in the audio passband.

The position of the sideband select switch in the signal path allows switching without readjusting the amplitude and phase trimmers. For most applications, one sideband will be used exclusively, and the sideband switch may be replaced by a pair of jumpers on the PC board. If that results in the wrong sideband, reverse the connections between the audio driver transistors and mixers.

The I and Q Class A audio drivers are emitter followers directly driven by the amplitude and phase trim op-amps. Emitter followers were used successfully in the original T2 and work well.[12] An emitter follower can only source current, so it must be biased with more than the negative peak current required by the mixer IF ports.

Q1 and Q2 each have a quiescent current of more than 10 mA. For a receive application, a different approach would save operating

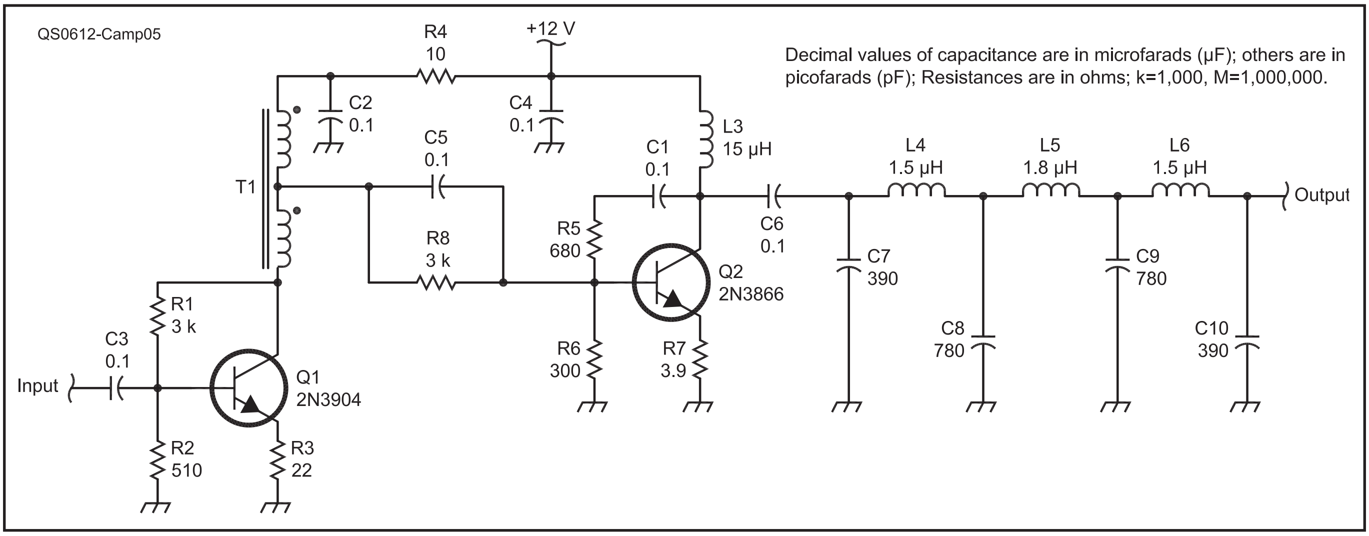


Fig 13.42 — Schematic diagram of 0.5 W power amplifier with low-pass filter shown in the photographs. T1 is 10 turns of #28 enameled wire bifilar wound on an FT37-43 toroid core.

current. Since the exciter draws only a fraction of the total transmitter current in most applications and is turned completely off during receive, use of clean Class A emitter followers to drive the mixer IF ports is a good trade-off. Another option is connecting the feedback resistor to the emitter rather than the base of each follower. A circuit simulator shows a tiny bit less distortion — but also some potential high frequency instability with that connection. This design sticks with the proven, conservative approach, and drops the distortion still further by burning a little more current in each of the transistors.

ADJUSTING THE SSB EXCITER

There are only three adjustments on the SSB generator PC board: RF AMPLIFIER TUNING, AMPLITUDE TRIM and PHASE TRIM. Each adjustment may be set once and then left alone. Adjust the SSB exciter by ear using a wideband audio noise source — such as a spare SSB receiver tuned to noise. Plug a cable into the "noise receiver" headphone jack and connect to the exciter microphone input, starting with the volume all the way down. With about 60 dB of attenuation on the RF output of the exciter, feed it directly into another receiver with selectable sidebands. Turn off the receiver AGC, if possible, and reduce the RF GAIN so that the peak exciter signal does not move the S-meter. Tune the receiver to zero beat on carrier leakage and then slowly turn up the volume on the noise source until the exciter noise output is a strong signal in the receiver.[13]

Switch back and forth between upper and lower sideband on the receiver and confirm that one sideband is much stronger than the other.[14] Switch to the desired sideband (if it is the wrong one, reverse the connections between the I and Q modulators and audio drivers) and peak the RF amplifier capacitor. Then switch to the opposite sideband and adjust the amplitude and phase trimpots for zero noise output. Alternate between the two trimpots, as these two adjustments become increasingly critical as each one approaches zero. Once adjusted, the energy in the opposite sideband will be more than 40 dB below the energy in the desired sideband. At that level, intermodulation products in the opposite sideband will dominate, and all you will hear are the familiar unintelligible pops and clicks on voice peaks common to any clean SSB signal driving a practical linear amplifier.

OTHER MODES WITH THE MICROT2

This exciter is very similar to the one in a 1958 Central Electronics 20A SSB exciter. If you look at the front panel of the 20A you see a mode switch marked USB, LSB, DSBsc, AM, NBPM, CW. The back of the switch has just a bunch of wires and resistors used to unbalance a modulator, turn off the I or Q audio channel, etc. We could get all of those modes out of this exciter, too, by inserting a switch in the wires connecting the I and Q modulator outputs to the I and Q mixer IF inputs.

Connecting either the I or Q signal path (but not both) will result in DSB. Connecting only the I (or Q) path and inserting a little carrier by connecting a 4.7 kΩ resistor from that side to the +12 V supply will generate AM. Connecting only the I path and inserting a 4.7 kΩ resistor from the Q mixer IF port to +12 V will generate narrowband phase modulation (NBPM). Disconnecting both the I and Q paths, or simply turning off the power to the AF section, and connecting 4.7 kΩ resistors from either or both mixer IF ports will generate CW. For CW, key both the voltage on the mixer IF ports and the RF amplifier enable line. **Fig 13.43** is a good circuit to accomplish this.

SO, HOW DOES IT SOUND?

Try evaluating SSB exciters this way: First, translate the frequency to some very quiet

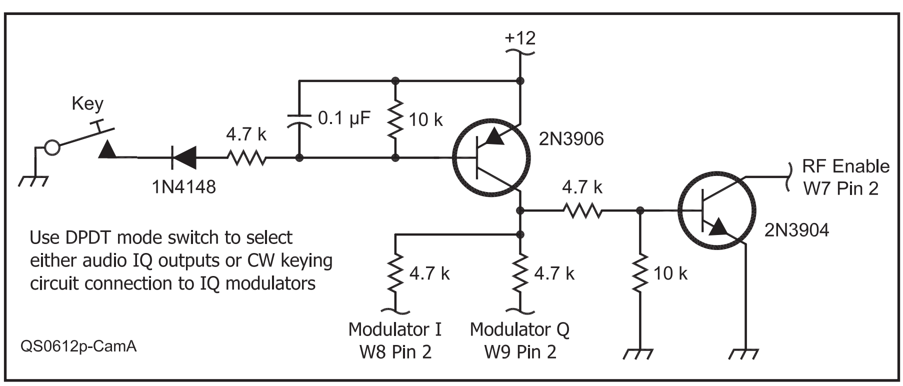


Fig 13.43 — Schematic diagram of suggested keying circuit.

spot in the HF spectrum using a crystal oscillator and mixer. Then put enough attenuators before and after the mixer to obtain a peak SSB output level of around –60 dBm — weak enough to not overload the receiver but strong enough to hear off-channel garbage another 60 or 70 dB below the peak transmitter output. Connect a CD player with some good acoustic folk music — lots of voice, guitar transients, perhaps a mandolin but no drums — to the exciter microphone input. Then connect the frequency converted exciter output directly into the input of a good receiver and tune it in. Don't connect it to an antenna — amateurs are not permitted to transmit music!

Any receiver with selectable sidebands and a manual RF gain control will do. Turn the receiver AGC off and manually reduce the gain so the receiver noise floor is well below the peak signal level. If your receiver is lacking in audio fidelity, run the receiver "line out" into a stereo amplifier. Play the CD through the SSB exciter, through the attenuators and frequency translator, into the receiver, and back into the stereo and out the speakers. It's an acid test, and this exciter sounds pretty good — better than most AM broadcast stations, and even some badly adjusted FM stations. Friends who hear you on the air will say, "Wow, it sounds exactly like you!"

Project: The MkII — An Updated Universal QRP Transmitter

A frequently duplicated project in the now out-of-print book *Solid State Design for the Radio Amateur*[15] was a universal QRP transmitter. This was a simple two-stage, crystal-controlled, single-band circuit with an output of about 1.5 W. The no frills design used manual transmit-receive (TR) switching. It operated on a single frequency with no provision for frequency shift. The simplicity prompted many builders to pick this QRP rig as a first solid state project.

The design simplicity compromised performance. A keyed crystal controlled oscillator often produces chirps, clicks or even delayed starting. The single pi-section output network allowed too much harmonic energy to reach the antenna, and the relatively low output of 1.5 W may seem inadequate to a first time builder.

A THREE-STAGE TRANSMITTER

Wes Hayward, W7ZOI, updated the design to the MKII (**Fig 13.44**). The circuit, shown in **Fig 13.45**, develops an output of 4 W on any single band within the HF spectrum, if provided with 12 V dc. Q1 is a crystal controlled oscillator that functions with either fundamental or overtone mode crystals. It operates at relatively low power to minimize stress to some of the miniature crystals now available. The stage has a measured output at point X of +12 dBm (16 mW) on all bands. This is applied to drive control R17 to set final transmitter output.

A three stage design provides an easy way to obtain very clean keying. Shaped dc is applied to driver Q2 through a keying switch and integrator, Q4.[16] A secondary keying switch, Q5, applies dc to the oscillator Q1. This is a time-sequence scheme in which the oscillator remains on for a short period (about 100 ms) after the key is released. The keyed waveform is shown in **Fig 13.46**.

The semiconductor basis for this transmitter is an inexpensive Panasonic 2SC5739. This part, with typical F_T of 180 MHz, is specified for switching applications, making it ideal as a class C amplifier. The transistor is conveniently housed in a plastic TO-220 package with no exposed metal. This allows it to be bolted to a heat sink with none of the insulating hardware required with many power transistors. A 2 × 4 inch scrap of circuit board served as both a heat sink and as a ground plane for the circuitry.

Another 2SC5739 serves as the driver, Q2. This circuit is a feedback amplifier with RF feedback resistors that double to bias the transistor.[17] Driver output up to 300 mW is available at point Y. Ferrite transformer T2 moves the 200 Ω output impedance seen looking into the Q2 collector to 50 Ω. The maximum output power of this stage can be changed with different R20 values. Higher stage current, obtained with lower R20 values, is needed on the higher bands. The 2SC5739 needs only to be bolted to the circuit board for heat sinking.

The Q3 power amplifier input is matched with transformer T3. The nominal 50 Ω of the driver is transformed to 12 Ω by T3.

The original design started with a simple L network output circuit at the Q3 collector followed by a third-order elliptic low-pass section to enhance harmonic suppression.[18] C5 is a moderately high reactance capacitor at the collector to bypass VHF components. This L network presented a load resistance of 18 Ω to the Q3 collector, the value needed for the desired 4 W output. But this circuit displayed instabilities when either the drive power or the supply voltage was varied. The output amplifier sometimes even showed a divide-by-two characteristic. The original L network was modified with the original inductor replaced with an LC combination, C4 and L1. The new series element has the same reactance at the operating frequency as the original L network inductor. This narrow band modification provided stability on all bands. The components for the various bands are listed in **Table 13.2**.

The inductance values shown in Table 13.2 are those calculated for the networks, but the number of turns is slightly lower than the calculated value. After the inductors were wound, they were measured with a digital LC meter.[19] Turns were compressed to obtain the desired L value. Eliminate this step if an instrument is not available.

The divide-by-two oscillations mentioned above could be observed with either an oscilloscope or a spectrum analyzer and were one of the more interesting subtleties of this project. The oscilloscope waveform looked like amplitude modulation. In the more extreme cases, every other RF cycle had a different amplitude that showed up as a half frequency component in the spectrum analyzer. The amplitude modulation appeared as unwanted sidebands in the spectrum display for the "moderately robust" instabilities. (Never assume that designing even a casual QRP rig will offer no development excitement!)

The output spectrum of this transmitter was examined with V_{CC} set to 12.0 V and the drive control set for an output of 4 W. The third harmonic output is –58 dBc and the others >70 dB down.

The author breadboarded the oscillator

Fig 13.44 — The MKII QRP transmitter includes VXO frequency control, TR switching and a sidetone generator.

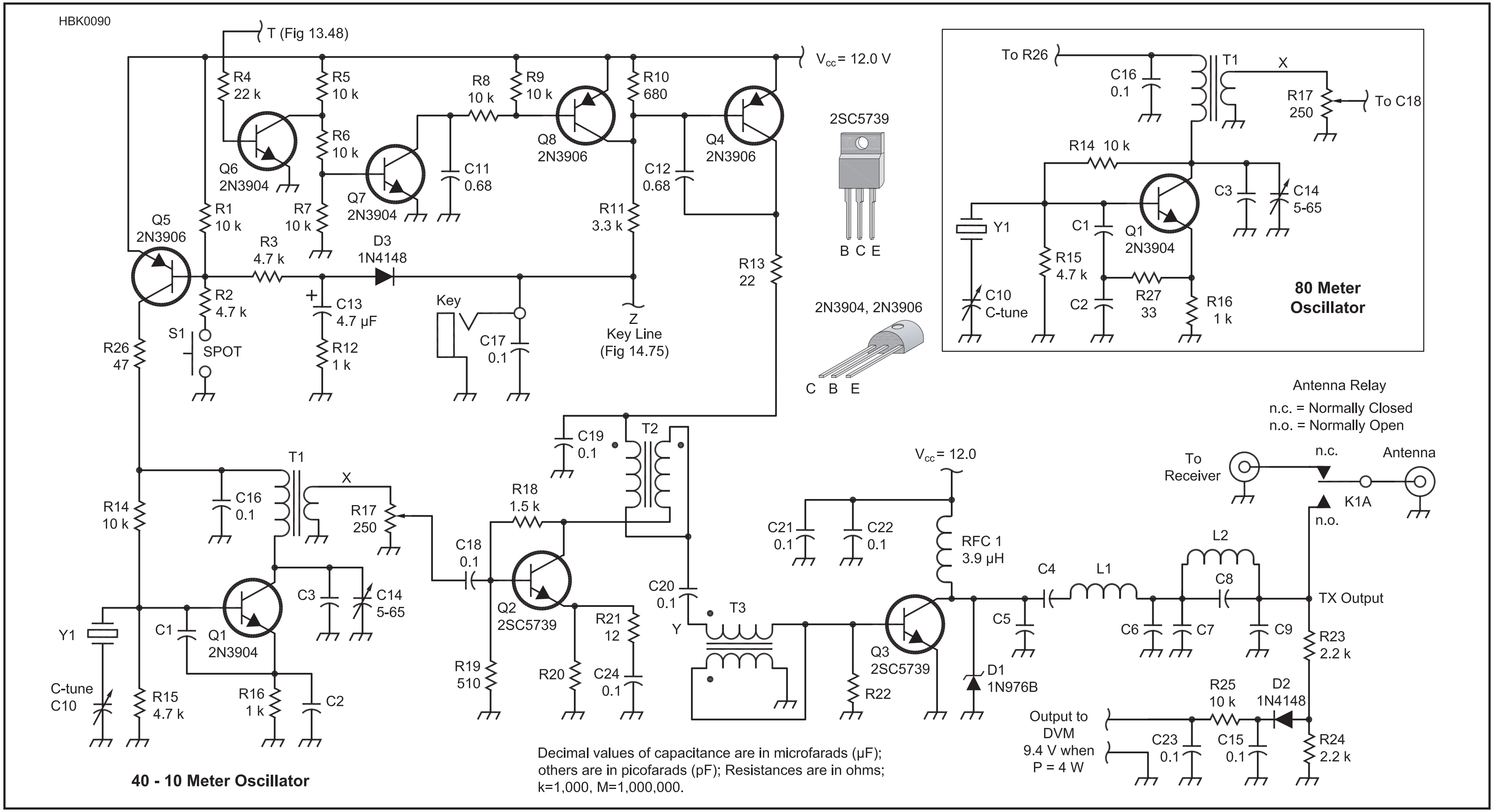


Fig 13.45 — Schematic diagram and parts list for the RF portion of the MKII transmitter. The oscillator (Q1 and associated components) in the main drawing is for the 40-10 m bands, while a modified version for 80 m is shown in the inset (see text). Fixed resistors are ¼ W, 5% carbon film unless otherwise noted. A kit of component parts is available from Kanga US (www.kangaus.com). TR switching is performed with a relay and additional circuitry (Fig 13.48).

C1-C9 — See Table 13.2, all 50 V ceramic or mica.

C10 — VXO control to provide some frequency adjustment around the crystal frequency. Use what you have in your junk box, although smaller capacitance values provide a wider tuning range. The prototype uses a small 2 to 19 pF trimmer. See text.

C11, 12 — 0.68 μF, 50 V metal film or Mylar.

C13 — 4.7 μF, 25 V electrolytic.

C14 — 5-65 pF, compression or plastic dielectric trimmer.

C15-24 — 0.1 μF, 50 V ceramic.

D1 — 1N976B, 43 V Zener diode.

K1A — See Fig 13.48.

L1, L2 — See Table 13.2.

Q1, Q6, Q7 — 2N3904, NPN silicon small signal transistor.

Q2, Q3 — 2SC5739 NPN silicon switching power transistor.

Q4, Q5, Q8 — 2N3906, PNP silicon small signal transistor.

R17 — 250 Ω, potentiometer (a 500 Ω potentiometer in parallel with 270 Ω fixed resistor can be substituted).

R20, R22 — See Table 13.2, carbon film.

RFC1 — 3.9 μH, 0.5 A molded RF choke. In place of a manufactured product, a T68-2 toroid wound with 26 turns of #22 enameled wire can be used.

T1 — See Table 13.2

T2 — 10 bifilar turns #28 enameled wire on FT-37-43 or FB-43-2401 ferrite toroid core.

T3 — 7 bifilar turns #22 enameled wire on FT-37-43 or FB-43-2401 ferrite toroid core.

Table 13.2
Band Specific Components of the MKII Transmitter

Band MHz	*T1 turns-turns*	*C1 pF*	*C2 pF*	*C3 pF*	*R20 Ω*	*R22 Ω*	*L1 nH, turns wire, core*	*L2 nH, turns wire, core*	*C4 pF*	*C5 pF*	*C6 pF*	*C7 pF*	*C8 pF*	*C9 pF*
3.5	51t-3t #26, T68-2	270	270	82	33	18	3000, 26t #28,T37-2	1750, 20t #28, T37-2	1000	390	1000	1000	300	1000
7	32t-4t #28, T50-6	390	100	82	33	33	1750, 19t #26, T37-2	890, 14t #22, T37-2	470	200	560	470	150	470
10.1	32t-4t #28, T50-6	390	100	0	33	33	1213, 19t #28, T37-6	617, 13t #28, T37-6	330	120	390	330	100	330
14	32t-4t #28, T50-6	390	100	0	33	33	875, 16t #28, T37-6	445, 11t #28, T37-6	220	100	270	220	75	220
18.1	20t-3t #28, T37-6	100	33	0	33	33	680, 14t #28, T37-6	346, 9t #28, T37-6	180	75	220	180	56	180
21	20t-3t #28, T37-6	100	33	0	18	33	583, 12t #28, T37-6	297, 9t #28, T37-6	150	62	180	150	50	150
24.9	20t-3t #28, T37-6	33	18	0	18	33	490, 11t #28, T37-6	249, 8t #28, T37-6	133	56	150	133	43	133
28	20t-3t #28, T37-6	33	18	0	18	33	438, 10t #28, T37-6	223, 7t #28, T37-6	120	47	140	120	39	120

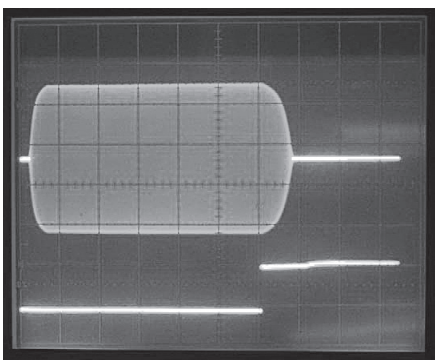


Fig 13.46 — Keyed waveform. The lower trace is the keyer input, which triggered the oscilloscope in this measurement. The horizontal time scale is 5 ms/div.

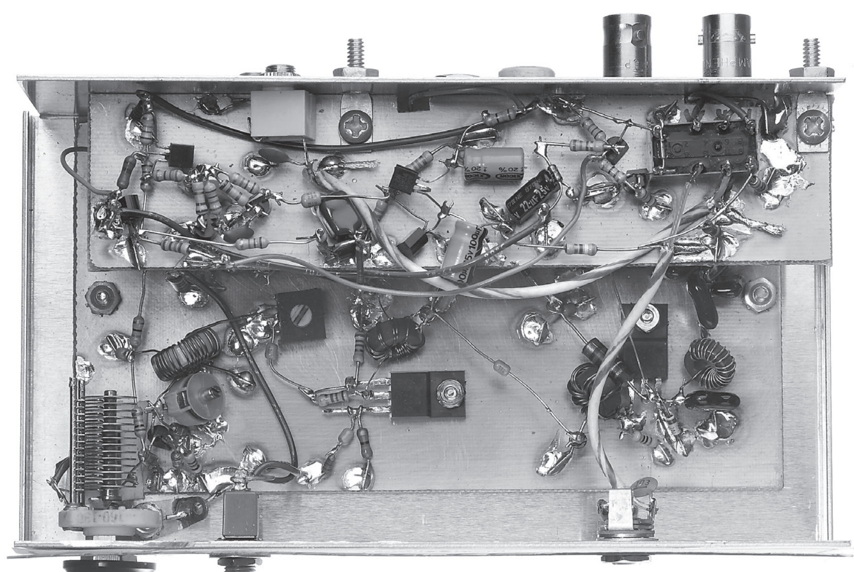


Fig 13.47 — The transmitter packaged in a 2 × 3 × 6 inch LMB #138 box. The basic RF circuitry is on the larger board. TR control is on the smaller board along the top.

and buffer section for all HF amateur bands from 3.5 to 28 MHz.[20] The power amplifier circuit has been built at 3.5, 7, 14 and 21 MHz. The crystals, obtained from Kanga US (**www.kangaus.com**), were fundamental mode units through 21 MHz, and third overtone above. The breadboard was built on two scraps of circuit board. Q1 and Q2 were on one with Q2 bolted to the board to serve as a heat sink. The second board had Q3 bolted to it, also serving as a heat sink.

After the breadboarding work was done, the circuits were moved to an available 2 × 3 × 6 inch box, an LMB #138. A new circuit board scrap was used, but most of the circuitry was moved intact from the breadboard. A diode detector was added to aid tune-up. The final RF board is shown in **Fig 13.47**.

CONSTRUCTION NOTES

Since publication in *QST*, some builders have encountered questions or difficulties. This section addresses those difficulties and further information may be found in *QST*.[21]

VXO Capacitor Grounding

The VXO capacitor, C10 of Fig 13.45, is mounted on the front panel of the transmitter rather than the circuit board. Grounding of the capacitor has been reported to be critical. A lead, ideally a short one, should go from the variable capacitor to the ground foil near the oscillator stage, Q1. In one of the transmitters built, the builder had merely attached the variable capacitor to the panel and relied on the ground connection that held the board to the box. This was, unfortunately, close to the power amplifier. The result was that the crystal oscillator would not always come on when the SPOT button was pushed. Adding a cleaner grounding wire solved the problem. The prototype uses a ground lug on the chassis very close to variable capacitor C10 and soldered directly to the PC foil right next to Q1.

Oscillator Changes

Some builders of the 40-meter version reported difficulty with tuning C14, the variable capacitor that tunes the collector circuit of the oscillator. The variable capacitor was too close to minimum C and a well defined peak was not always found. Of greater significance, tuning C14 to some values could allow the circuit to oscillate without crystal control of the frequency, producing oscillation in the 6.4-6.9 MHz region. Solutions to both problems are simple. First, change C3 from 100 to either 82 pF to remove the tuning ambiguity. (Removing a turn or two from the high L winding on T1 will accomplish the same end.) Second, adding C1 at 390 pF to the 40 meter circuit produces an oscillator that is always crystal controlled for any tuning of C14. Further experimentation with higher frequency versions revealed that the oscillators were generally well behaved but undesired modes could be found with extreme tuning of C14. Adding C1 to the circuit when it was initially absent always fixed this problem. The corrected component values are shown in Table 13.2.

Oscillation without crystal control was also observed with a misadjusted 80 meter circuit. Increasing the value of C1 helped but did not completely remove the problem. Analysis showed that the 80 meter oscillator starting gain was higher than was available on the higher-frequency bands. The gain was high enough that an instability was observed when the crystal was removed and the circuit was driven as an amplifier with a signal generator. Amplifier output jumped as the generator frequency was tuned.

Common "fixes" for amplifier instability include loading and the application of negative feedback. Increased loading through adjustment of R17 helped, but this eliminated the ability to adjust overall transmitter output with this control. So, negative feedback was tried. The resulting oscillator circuit is shown in the inset of Fig 13.45. Emitter degeneration is added in the form of a 33 Ω resistor (R27), while parallel feedback is realized by moving the 10 kΩ base bias resistor (R14) from the bypass capacitor to the Q1 collector. The circuit, as shown, would not oscillate without crystal control. Care is still required for initial adjustment (with a receiver or spectrum analyzer) to avoid crystal controlled oscillation on a crystal spurious resonance. For example, one oscillator tested achieved crystal controlled operation at 3.8 MHz with a crystal built for operation at 3.56 MHz. Crystal spurious modes of this sort are found in virtually all crystals and should not be regarded as a crystal problem.

TRANSMIT-RECEIVE (TR) SWITCHING

Numerous schemes, generally part of a transceiver, are popular for switching an antenna between transmitter and receiver functions. When carefully refined, full-break-in keying becomes possible, an interesting option for transceivers. But these schemes tend to get in the way when one is developing both simple receivers and transmitters, per-

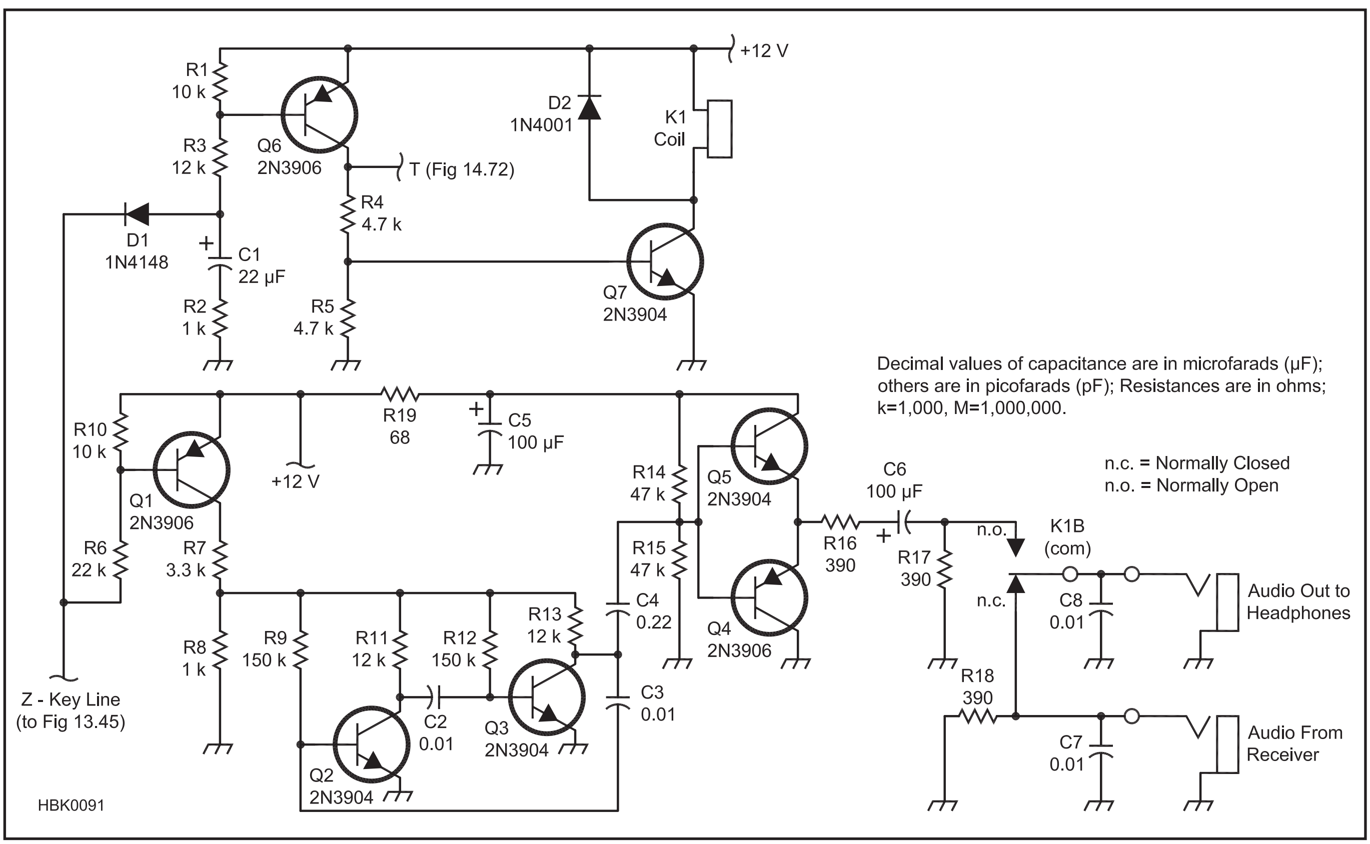


Fig 13.48 — Detailed schematic diagram and parts list for transmit-receive control section and sidetone generator of the universal QRP transmitter. Resistors are ¼ W, 5% carbon film. A kit of component parts is available from KangaUS (www.kangaus.com).

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.
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.

haps as separate projects. A simple relay based TR scheme is then preferred and is presented here. In this system, 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 **Fig 13.48**. Line Z connects to the key. A key closure discharges capacitor C1. R2, the 1 kΩ resistor in series with C1, prevents a spark at the key. Of greater import, it also does not allow us to "ask" that the capacitor be discharged instantaneously, a common request in similar published circuits. 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.

Relay contacts B switch, the audio line. R17 and 18 suppress clicks related to switching. A depressed key turns on PNP switch Q1, which then turns on the sidetone multivibrator, Q2 and Q3. The resulting audio is routed to switching amplifier Q4 and Q5. Although the common bases are biased to half of the supply voltage, emitter bias does not allow any static dc current to flow. The only current that flows is that related to the sidetone signal during key down intervals. Changing the value of R16 allows the audio volume to be adjusted, to compensate for the particular low-impedance headphones used.

There is an additional interface between Figs 13.48 and 13.45. Recall that Q4 of Fig 13.45 keys buffer Q2 while Q5 provides a time sequence control to oscillator Q1. Additional circuitry uses Q6, Q7 and Q8, and related parts. Under static key up conditions, Q7 is saturated, which keeps C11 discharged. Saturated Q7 also keeps PNP transistor Q8 saturated. This closed switch is across the emitter-base junction of Q4. Hence, pressing the key will start relay timing and will allow the oscillator to come on, but will not allow immediate keying of Q2 through Q4. Key closure causes Q6 in Fig 13.48 to saturate causing point T to become positive. This saturates Q6 of Fig 13.45 which turns Q7 off, allowing C11 to charge. When C11 has charged high enough, Q8 is no longer saturated and Q4 can begin its integrator action to key Q2.

This hold-off addition has solved a problem of a loud click, yielding a transmitter that is a pleasure to use. There is still a flaw resulting in the initial CW character being shortened. The result is that an I sent at 40 WPM and faster comes out as an E. Further refinement of timing component values should resolve this. The TR system circuitry is built on a narrow scrap of circuit board that is then bolted to the transmitter rear panel.

What's Next?

This has been an interesting project from many viewpoints. The resulting transmitter, which is usually used with the S7C receiver from *Experimental Methods in RF Design*,[22] is a lot of fun to use and surprisingly effective in spite of its crystal control. Primitive simplicity continues to have its place in Amateur Radio. Also, the development was more exciting than expected. The observed instabilities were interesting, as were the subtleties of the control system. Perhaps we should not approach simple CW systems with a completely casual attitude, for they continue to offer education and enlightenment.

There are clearly numerous refinements available for this transmitter. The addition of an adjustable reactance in series with the crystal will allow its frequency to move more. Try just a small variable capacitor. Two or more similar crystals in parallel form a "super VXO" topology for even greater tuning range. Higher power supply voltage will produce greater output power — over 10 W on the test bench. The transmitter could certainly be moved down to 160 meters for the top band DXer looking for QRP sport. It is not certain that the 2SC5739 will allow operation as high the 6 meter band. The transmitter could easily be converted to a modest power direct conversion transceiver using, for example, the Micromountaineer scheme offered in *QST*.[23]

13.4 Modern Baseband Processing

The term *baseband* refers to the signal or signals that comprise the information content at their natural frequency. For a communications audio signal, it would be a spectrum typically extending from 300 to 3300 Hz. Many transmitter architectures are designed to process and transmit a spectral range rather than any particular type of information. For example, the typical transmitter that shifts the modulating spectrum to occupy a single sideband adjacent to a suppressed carrier — our usual SSB transmitter, is just as happy to handle voice, modem tones or the two tones from an RTTY converter. The transmitter performs a linear operation to shift the input spectrum, the baseband signal, to the output frequency independent of the form or information content of the input spectrum.

This approach has a number of advantages for the transmitter designer and manufacturer. The baseband spectrum width, amplitude and dynamic range are inputs to the design process. The designer can thus focus on establishing the system between the baseband and the antenna port. This leaves the design of the baseband processing subsystem to perhaps another department or another company. Similarly the baseband processing equipment design may have multiple applications. Its output can be plugged into cable systems, HF transmitters or microwave systems as long as they support the required bandwidth, amplitude and dynamic range.

13.4.1 Digital Signal Processing for Signal Generation

The transmitter architecture that was described in Fig 13.27 was based on the classical analog approach to waveform generation and modulation with information content. Many current transmitters have replaced the early analog signal processing stages with a digital

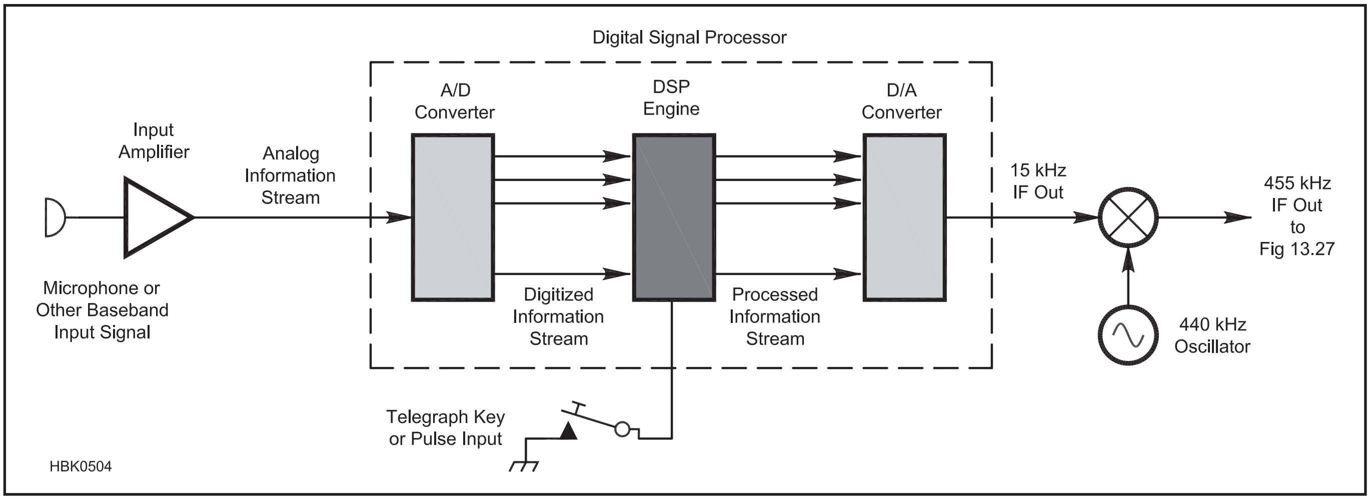


Fig 13.49 — Block diagram of a 15 kHz DSP-based baseband signal processor and IF generator.

signal processor (DSP). (Additional discussion is provided in the **DSP and Software Radio Design** chapter.)

Fig 13.49 provides a simplified block diagram of such a DSP based subsystem. The 455 kHz output would connect up to the 455 kHz amplifier stage in Fig 13.25, in place of the SSB generation and low IF subsystem that occupies the lower half of Fig 13.27.

Note that a DSP engine can generate virtually any desired waveform or modulating envelope desired by using different firmware. Thus the transmitted bandwidth, voice processing, carrier rejection or insertion, sideband selection pulse shaping and control functionality can be established with different firmware. This is a great benefit to manufacturers who can provide a single hardware platform that is usable for multiple services.

13.5 Increasing Transmitter Power

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.

13.5.1 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 13.3**. 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.

Table 13.3
Summary of Characteristics of 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%

In handheld digital cellular transceivers and base stations, the power amplifier must operate in a linear fashion to meet spectral purity requirements for the complex digital modulation schemes used. Since linear amplifiers are generally not very efficient, this is a major contributor to the energy consumption. In an effort to extend battery life in portable units, and reduce wasted energy in fixed equipment, considerable research is underway to use nonlinear amplifiers to provide linear amplification. Such techniques include pre-distortion of the low-level signal, polar modulation, envelope elimination and restoration, Cartesian-loop feedback and others.

13.5.2 Linear Amplifiers

While transmitters at power levels of 1 mW to 1 W have been successfully used for communication across many portions of

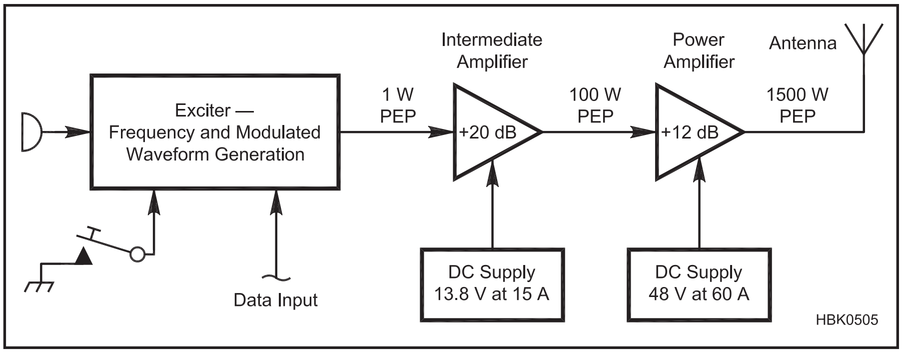


Fig 13.50 — Block diagram of a linear transmitter chain with multiple amplifier stages.

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 **Fig 13.50**.

The power levels shown at the various points in Fig 13.50 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 50 W ERP_D 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, it often 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 tubes 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.

13.5.3 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 **Fig 13.51**. Note that none of the upper RF stages (the "RF chain") needs 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 **Fig 13.52**. 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!

13.5.4 Hybrid Amplifiers

Another alternative that is convenient with current equipment is to use an AM transmitter with a linear amplifier. This can be successful if the relationship that PEP = 4 × Carrier Power is maintained.

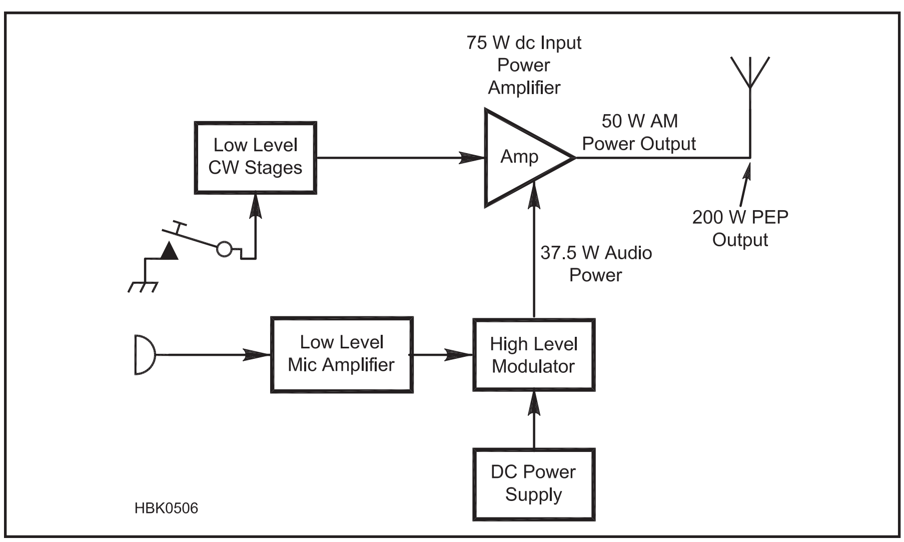


Fig 13.51 — Block diagram of a high level AM modulated transmitter.

Fig 13.52 — Block diagram of a high level AM modulated transmitter with added output stage.

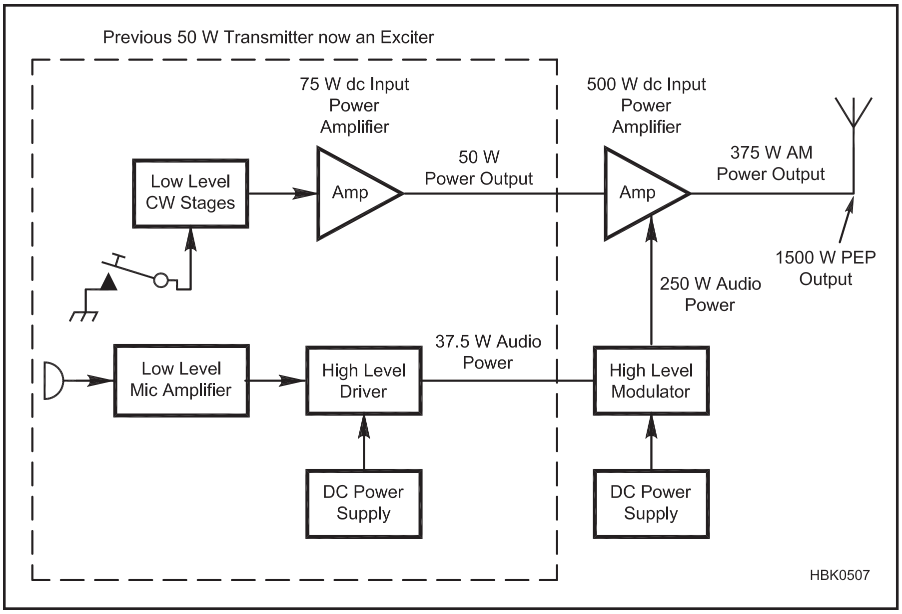


Fig 13.53 shows a 1500 W PEP output linear amplifier following a typical 50 W AM transmitter. In this example, the amplifier would be adjusted to provide a 375 W carrier output with no modulation applied to the exciter. During voice peaks the output seen on a special PEP meter, or using an oscilloscope 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.

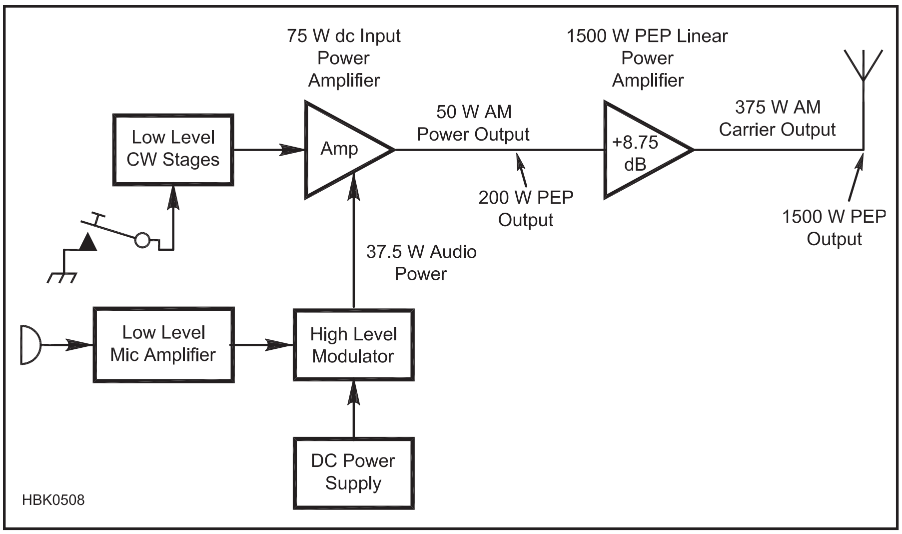


Fig 13.53 — Block diagram of a hybrid nonlinear/linear AM transmission system.

13.6 References and Bibliography

[1]R. Shrader, W6BNB, "When Radio Transmitters Were Machines," *QST*, Jan 2009, pp 36-38.

[2]The only remaining Alexanderson alternator transmitter is the Grimeton VLF transmitter (**en.wikipedia.org/wiki/Grimeton_VLF_transmitter**) that operates on 17.2 kHz.

[3]The Aldis lamp (**en.wikipedia.org/wiki/Signal_lamp**) was used on British navy vessels in the late 1800s and is still in use on naval vessels as it allows signaling between ships and from ship to shore without breaking radio silence.

[4]E. Hare, "The Tuna Tin Two Today," *QST*, Mar 2000, pp 37-40.

[5]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.

[6]M. Schwartz, *Information Transmission, Modulation and Noise,* third edition, McGraw-Hill, 1980.

[7]W. Sabin and E. Schoenike, Editors, Single-Sideband Systems and Circuits, McGraw-Hill, 1987.

[8]W. Sabin, "RF Clippers for SSB," *QST*, Jul 1967, pp 13-18.

[9]J. Grebenkemper, KI6WX, "Phase Noise and its Effect on Amateur Communications," *QST*, Mar, Apr 1988.

[10]W. Hayward, W7ZOI, R. Campbell, KK7B, and B. Larkin, W7PUA, *Experimental Methods in RF Design*, ARRL, 2003.

[11]R. Fisher, W2CQH, "Twisted-Wire Quadrature Hybrid Directional Couplers," *QST*, Jan 1978, pp 21-24.

[12]R. Campbell, KK7B, "A Multimode

Phasing Exciter For 1 to 500 MHz," *QST*, Apr 1993, pp 27-32.

[13]Some amateur receivers have inadequate shielding, and may overload on signal leakage from the exciter crystal oscillator — in that case, use a simple frequency converter to translate the exciter output to a different frequency. The converter may also be used to translate the frequency all the way down to 10 kHz as discussed on *EMRFD* page 7.35 to analyze the exciter signal using a computer sound card and audio spectrum analyzer software.

[14]Even a poorly adjusted phasing single sideband generator will have significant opposite sideband rejection. If it has little or none, stop and find the construction error. With over 100 parts on a PC board, it's easy to make an error — a lead left unsoldered or swapped components.

[15]W. Hayward, W7ZOI, and D. DeMaw, W1FB (SK), *Solid State Design for the Radio Amateur*, ARRL, 1977, pp 26-27.

[16]See Note 10, p 6.63.

[17]See Note 10, pp 2.24-2.28, for information on feedback amplifiers.

[18]See Note 10, p 3.6, for low-pass filter designs, 3.29 for matching network designs.

[19]See **www.aade.com** for details. A homebrew alternative is offered in Note 10, Chapter 7. Also see Carver, "The LC Tester," *Communications Quarterly*, Winter 1993, pp 19-27.

[20]See Note 10, p 1.2, and Hayward and Hayward, "The Ugly Weekender," *QST*, Aug 1981, p 18.

[21]W. Hayward, W7ZOI, "Crystal Oscillator Experiments," Technical Correspondence, *QST*, Jul 2006, pp 65-66.

[22]See Note 2, p 12.16.

[23]Hayward and White, "The Micro-mountaineer Revisited," *QST*, Jul 2000, pp 28-33.