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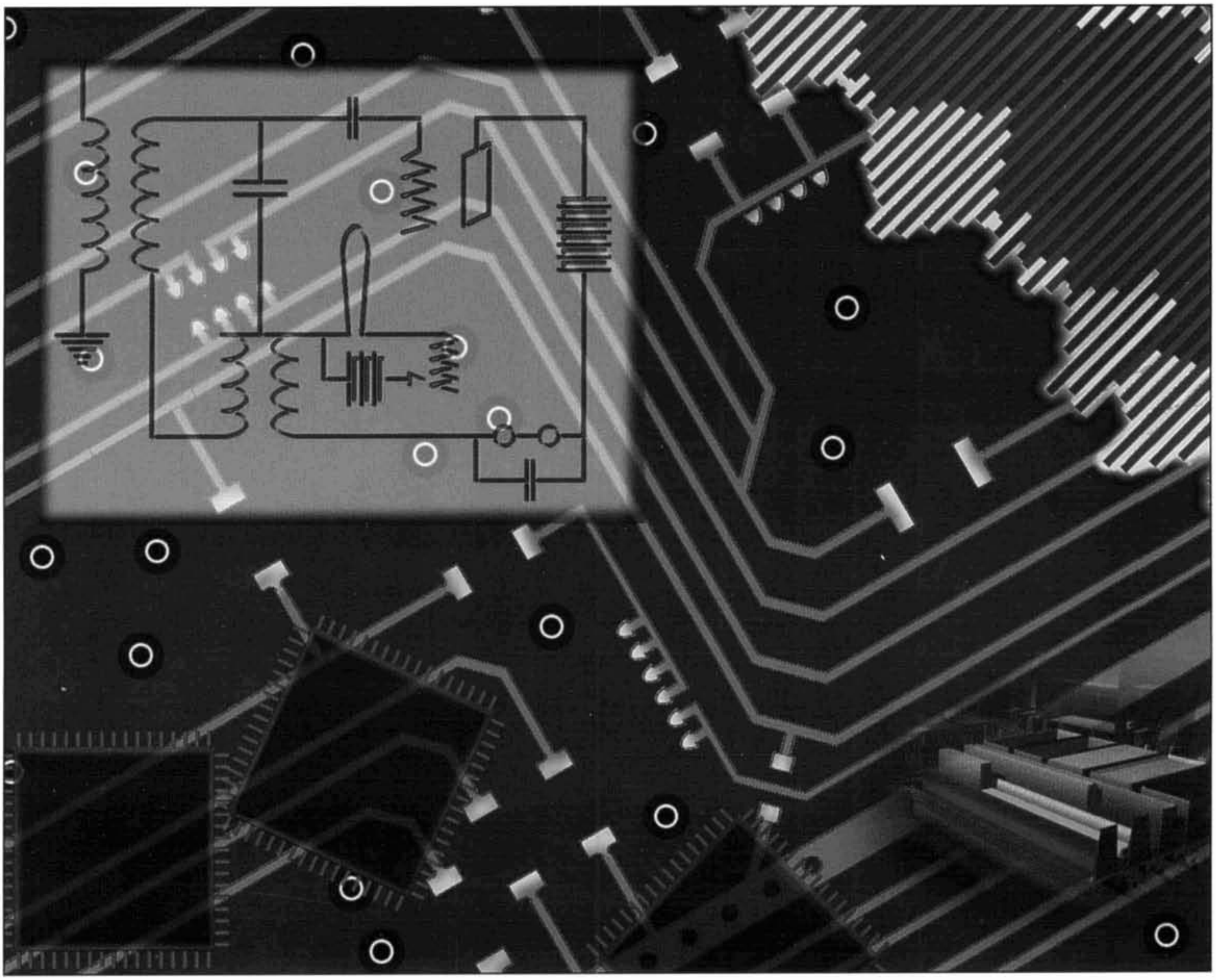

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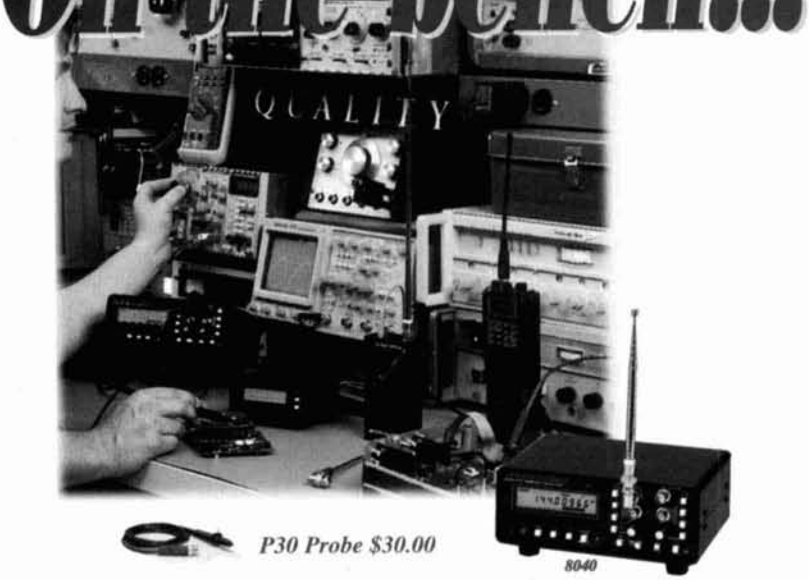

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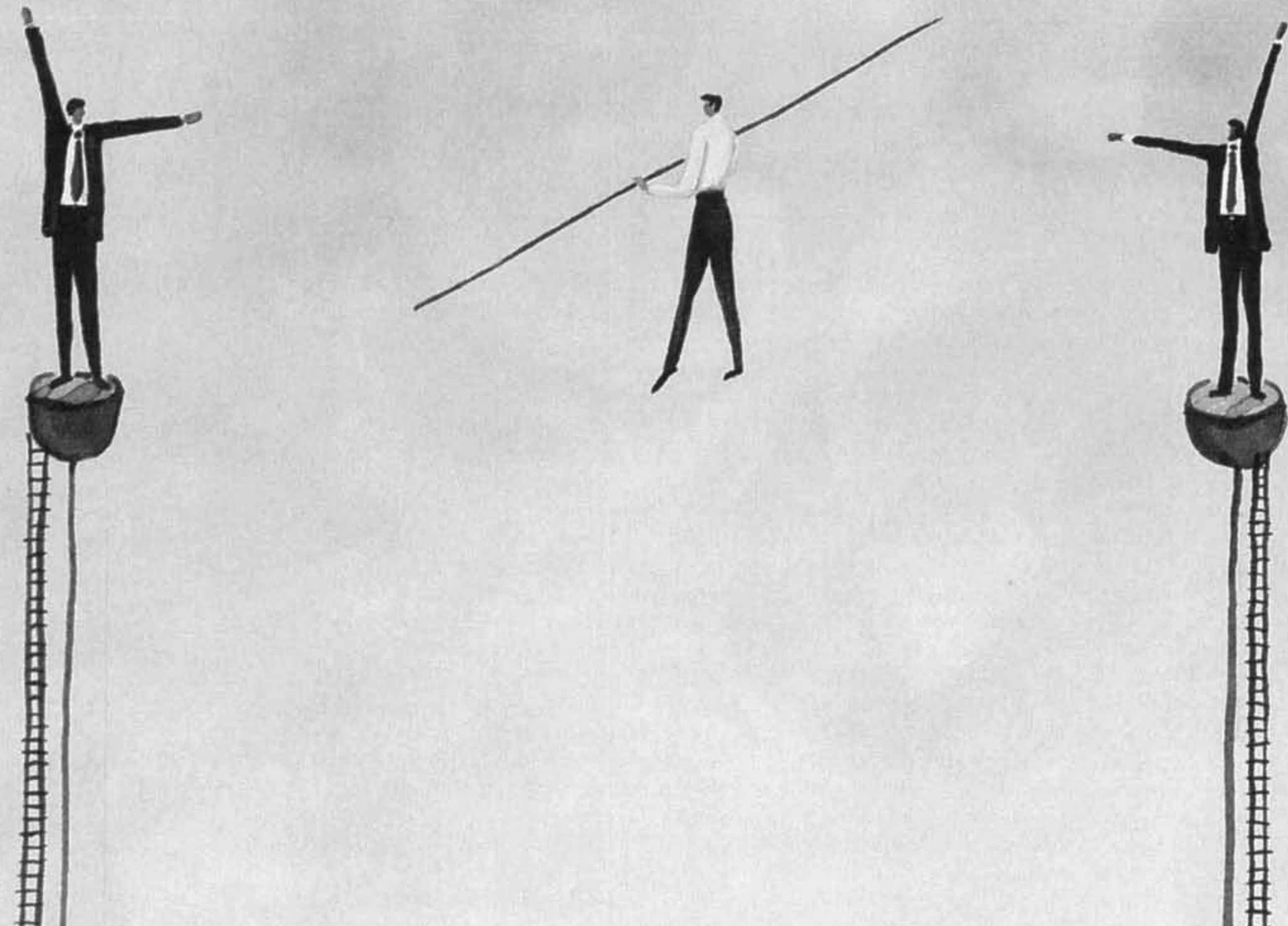
Cover artwork: In his article, "Regenerative Receivers," found on page 7, Charles Kitchin, N1TEV, updates this popular radio. Here, thanks to the computer magic of NU1N, we see Armstrong's original regenerative circuit superimposed on a solid-state design.



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EDITORIAL

Thank You Mr. Morgan!

I grew up in the fifties. Back then, most families in my rural Connecticut neighborhood still relied on radio to keep entertained and in touch with the world. A few families had TVs, but that still didn't stop neighbors from being neighbors. I remember the grownups gathering on each other's front porches most summer evenings and visiting for hours on end, long after the sun had set. As for us kids, anything resembling home computers, Nintendo, or Segas was beyond science fiction. Still, we found ways to entertain ourselves. We seemed better able to use our imaginations and the resources at hand without the need for virtual reality games or computers to have fun and enjoy our free time.

One of my favorite pastimes was scrounging the neighborhood in search of old radios. No corner of a neighbor's garage, barn, or cellar remained unscathed in my quest for these treasures. They were squirrelled away in my basement "laboratory," and later disemboweled to provide goodies for my junkbox. Tubes went in one box, things with dangling wires in another, anything colorful in still another. . . Oblivious to the fact that I had been bitten by the radio bug, I innocently pursued my gift for destroying anything electrical or electronic.

I was a little young to know about ham radio, or what an "Elmer" was, but it was around this time I was introduced to books by Alfred Powe Morgan. (I've never learned if Mr. Morgan was a ham, but I'm sure he must have been!) Mr. Morgan wrote a massive series of "Boys Books" starting in the 1930s. He wrote books about: motors, tools, woodworking, chemistry, electricity and electronics, and—of most interest to me—a whole series of Boys Books about Radio!* Mr. Morgan shared his seemingly endless knowledge with his "boys," showing us how to make Tesla coils, Geiger counters, spark coils, a simple x-ray machine, and even a miniature electric trolley and tracks! No shortcuts—you even had to cut the parts for the little motor out of tin metal and do the windings yourself! I guess Mr. Morgan expected dad to step in and help out at times like this. . .

My first project from his books was a simple regenerative one-tube receiver. I must have been all of about ten or eleven, but I still remember taking the bus alone from the suburbs (something I wouldn't dream of allowing my son to do in this day and age) and traveling to Hartford with my parts list and several weeks allowance in hand. Even in the late '50s, a 1H4 tube was hard to find; but after exploring every parts counter of the radio supply houses and ham stores, I finally secured one at an old-timer's radio repair store. Now I had everything Mr. Morgan said I needed to make his radio. It worked, of course, and that radio was the beginning of my long and enjoyable career in electronics.

Unfortunately, over time, I've lost track of my collection of Morgan books, and I don't know what happened to that old one-tube homemade radio. I was saddened when I learned recently that Mr. Morgan had passed away in 1972. But the knowledge, confidence, and wonder that Mr. Morgan imparted to me remain to this day. "Thank you, Mr. Morgan."

In this issue of *Communications Quarterly*, Mr. Charles Kitchin, N1TEV, reintroduces us to the regenerative detector. Devised by Edwin Armstrong in 1911 at the age of 21, regenerative circuits were immensely popular in early radio—especially since RCA held the superheterodyne patents and vigorously discouraged unlicensed competitors from using this principle. Regenerative receivers remained in vogue in amateur circles through the early 1930s, fading from popularity as the Great Depression eased and more advanced designs became economically feasible.

Mr. Kitchin's circuit reflects what can be done using modern components when applied to vintage techniques. To compliment his article, we decided to include a product review of the Ten Tec 1253 nine-band regenerative receiver kit, for those who'd like to be able to relive the days of the regen without the hassle of hunting for parts. Hopefully, these little kits, gifts from present-day "Elmers," will become the impetus for others who aspire to become hams, engineers, or scientists.

Peter Bertini, K1ZJH
Senior Technical Editor

**Before someone accuses Mr. Morgan of being sexist or politically incorrect, you should know that, while doing research for this editorial on my local library computer system, I discovered Mr. Morgan had penned the "First Chemistry Book for Boys and Girls," and "A Pet Book for Boys and Girls."*

Credit Due

Many thanks to our summer intern, ninth grader Drew Northup, for updating our end of the year index. His sharp eyes found several goofs I'd made during the compilation of last year's version. For next year, Drew hopes to organize all the information on a computer disk, allowing readers to locate articles by subject and other key words. Sounds like a good idea; we'll keep you posted!

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TECHNICAL CONVERSATIONS

Dear Editor:

Kudos to Cohen and his fractal antennas. He's managed to convincingly nibble away at 50 years of antenna dogma and put it in its place. I'm sure there's about a thousand people hitting their heads and saying "why didn't I think of that?" Certainly the coastline problem (large perimeter—small area) is about the first thing you see in any book on fractals. What a coup de grace to apply it to antennas! It's bizarre to know that the efficiency of a small loop is geometry-specific. Now antennas offer some real hope for new developments with this fractal breakthrough.

My only "beef" with Cohen's article is his "measurement" of the ohmic resistance. I'm sure it's low with any type of quad, but how did he find it? SWR analyzers and RX bridges only give *total* resistance, which includes the ohmic and radiative parts. Also, Cohen must have tinier antennas than those he showed. Will these be in Part 2? How about a Landsdorfer antenna article? Let's see more articles like this *CommQuart*!

Steve Kirschner, KA1UTE
Berkeley Lake, Georgia

Dear Editor:

I was interested to see the article on the G2AJV antenna in the Summer edition of *Communications Quarterly*. An earlier article on this antenna was published (as mentioned in the REFERENCES) in the April and May 1994 editions of *Radcom*. At this time I had worked as Technical Editor of *Radcom* at the RSGB for only a month and was asked to sub-edit the article. I was unhappy about publishing it unless the antenna could be validated in some way. When I started this validation work I found that feeding this antenna was not as easy as described in the original article and at one

(Continued on page 102)

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REGENERATIVE RECEIVERS

Past and present

The modern age has provided radio amateurs with lots of fancy gear. Today, virtually all radios are of the multistage superheterodyne type with digital memories and multistage filtering. Unfortunately, this advance in technology has come to us at a very high price. Today, most people are so intimidated by the complexity of modern receivers that they would never even consider building their own equipment from scratch.

Back in the 1920s and '30s, most radio amateurs (even non-hams) built homebrew receivers. The novelty of radio, not to mention the pride hobbyists felt in constructing their own equipment, strongly affected the culture of that time. In addition to having fun and saving money, the knowledge gained from homebrewing also greatly improved builders' technical skills. However, as the great depression came to a close, low-cost commercial receivers became available, and people were no longer required to build their own equipment—so many did not. Naturally, commercial firms chose to market varieties like the superhetero-

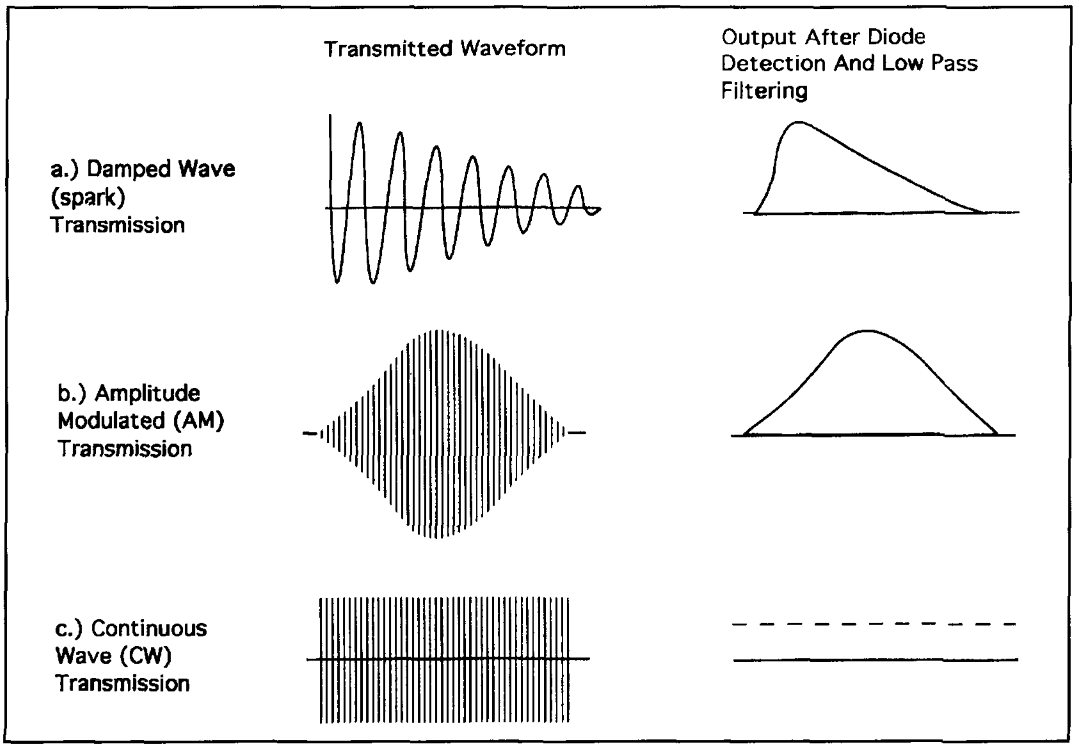


Figure 1. Various types of transmitted waveforms: (A) damped wave (spark) transmission, (B) amplitude modulated (AM) transmission, (C) continuous wave (CW) transmission.

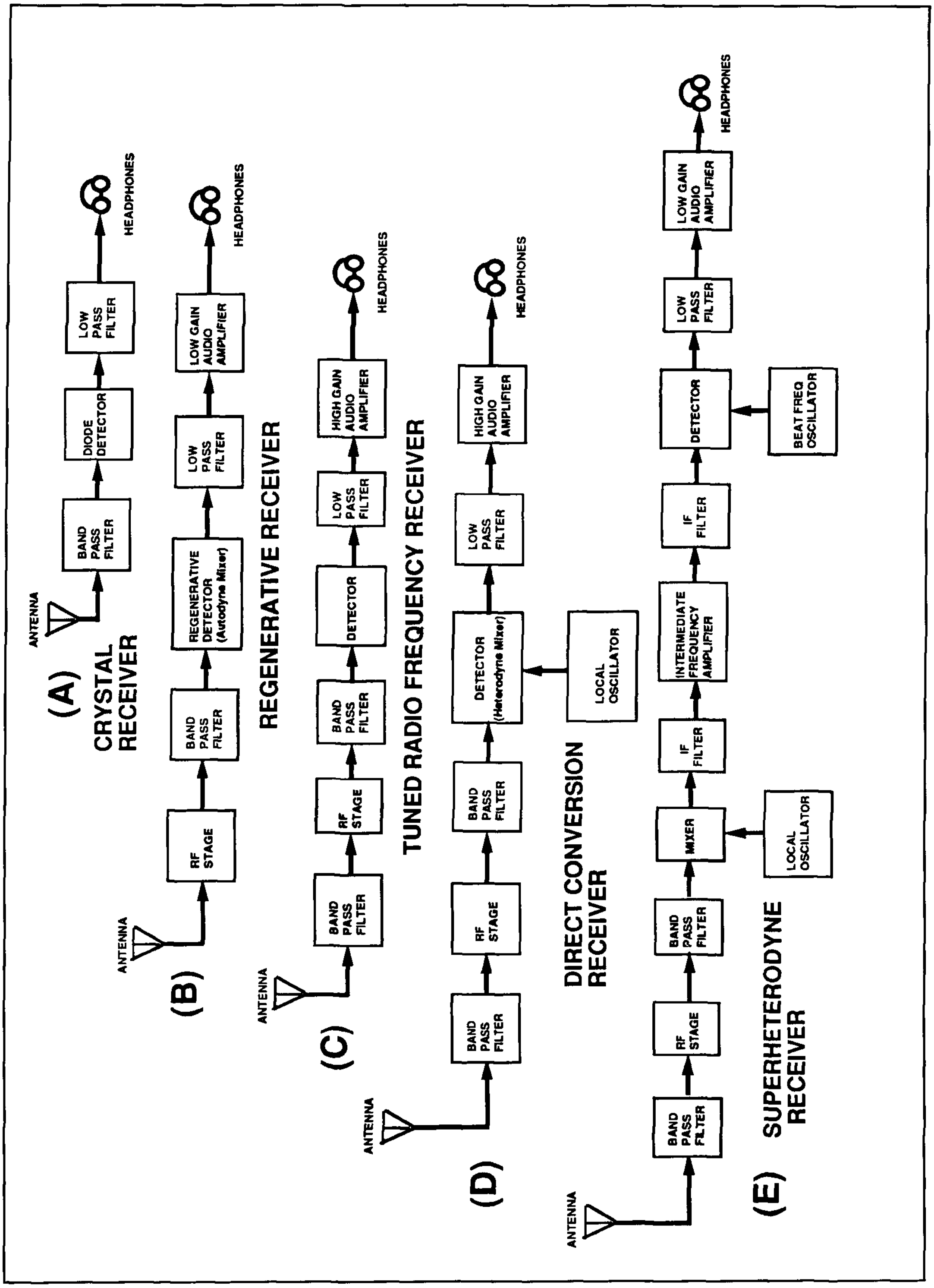


Figure 2. Block diagrams of some common receiver types: (A) crystal receiver, (B) regenerative receiver, (C) tuned radio frequency receiver, (D) direct conversion receiver, (E) superheterodyne receiver.

dyne, which are the easiest to use, but the most difficult to homebrew.

For several decades, we have been living in the "dark ages" of receiver homebrewing. As interest in home construction faded from popular culture, much of the early homebrew circuit technology was also forgotten. This is most unfortunate because many of these early circuits were, and still are, ideal for home construction. Recently, however, a small renais-

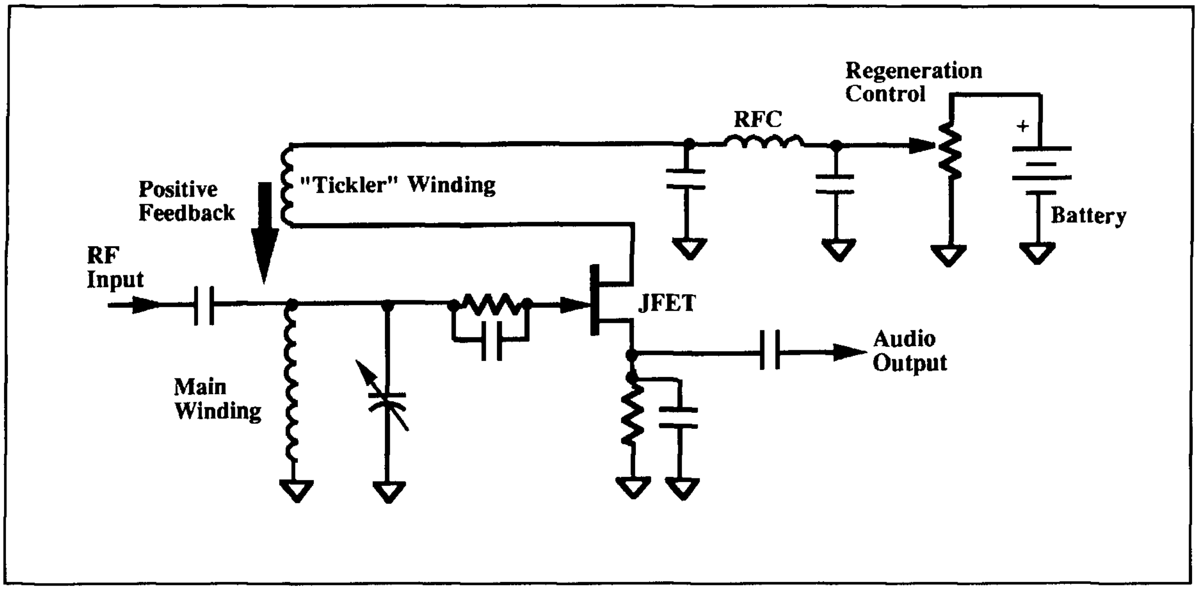


Figure 3. A modern regenerative detector circuit using a JFET.

sance has occurred with the rediscovery of the simple heterodyne receiver. Today it's called the direct conversion receiver.

Another important receiver technology in common use during the 1920s and '30s was the regenerative circuit. This used positive feedback, called "regeneration," to dramatically increase both the sensitivity and selectivity of simple receivers—typically by a thousand times or more. This article provides a comprehensive look into regeneration, reviews its history, and shows how to build high performance regenerative receivers using modern components.

Types of radio signals

Figure 1 shows some of the different types of radio signals that are (or were) used on the shortwave bands. In **Figure 1A**, you see the so-called "damped" waves produced by the spark-gap transmitters used in the early days of radio. When diode detected and low-pass filtered, they provide a crude audio output. This is one of the main reasons why "spark" signals were used: the damped wave provided an easy way to modulate the RF carrier, so a simple crystal diode could detect the signal.

Figure 1B shows an amplitude modulated (AM) signal and its diode detected and filtered output. Like the early spark transmissions, this type of waveform can be detected using a simple diode.

Figure 1C shows a continuous wave, or CW, signal. Here, simple diode detection and filtering produces only a DC voltage. Reception of an audio output from CW requires some type of heterodyne circuit where one frequency, from a local oscillator inside the receiver, is mixed with the incoming signal, to produce sum and difference frequencies.* If the local oscillator's frequency is set just slightly above, or below, the incoming signal's frequency, an audio output frequency is produced. Single sideband signals are AM waveforms without a carrier wave

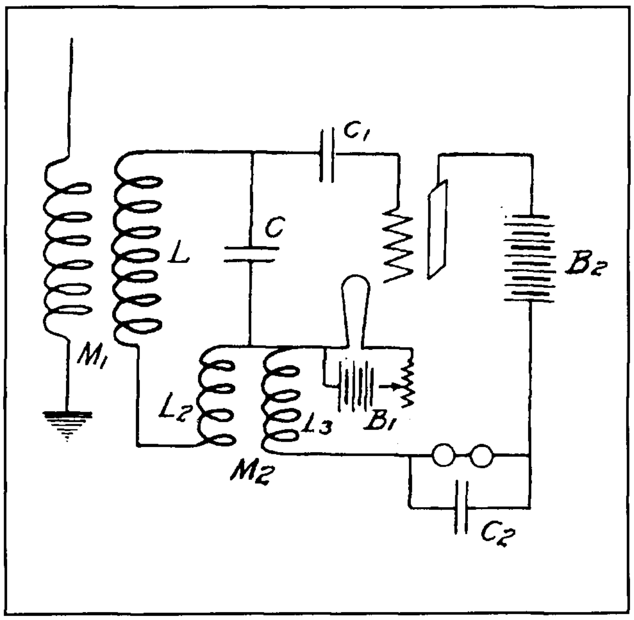


Figure 4. Armstrong's original regenerative circuit. (Courtesy MIT Barker Engineering Library and the IEEE.)

*Note that, strictly speaking, a diode detector isn't just a simple rectifier. Being a nonlinear device, it also operates as a "mixer," producing both sum and difference output frequencies. If used with a local oscillator, it too can produce an audio frequency output for receiving CW.

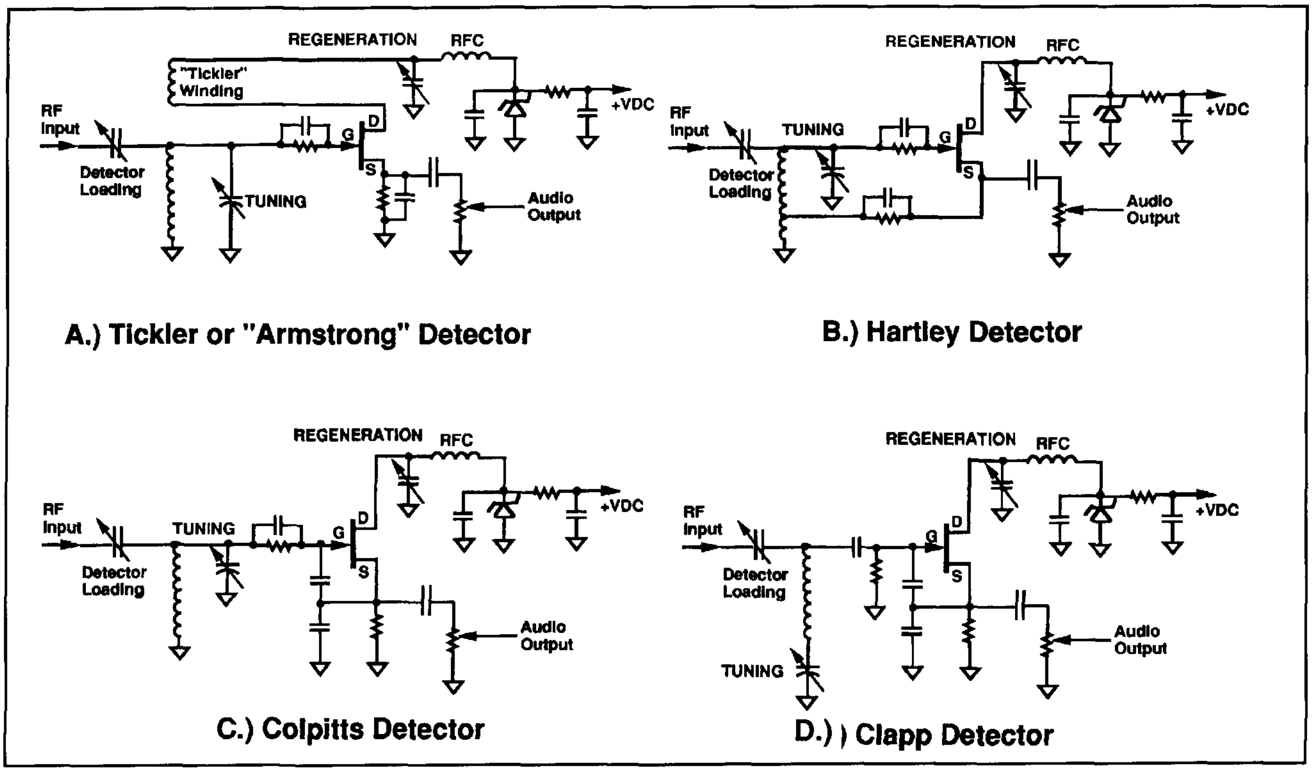


Figure 5. Some of the many possible regenerative detector circuits based on common oscillators: (A) Tickler or "Armstrong" Detector, (B) Hartley Detector, (C) Colpitts Detector, (D) Clapp Detector.

and, like CW, also require some type of heterodyne to provide an audio output.

An overview of some common types of receivers

Figure 2 shows the block diagrams of several types of receivers. It's interesting to note that, except for the crystal receiver, basically all of the fundamental circuits used in the reception of radio frequency signals were developed between 1914 and the mid-1930s by Edwin Howard Armstrong.[1,2]

Crystal receivers

As illustrated in **Figure 2A**, the simple crystal receiver requires only an LC tuned circuit, a diode detector, a low-pass filter, and headphones. The low-pass filter was usually just a single capacitor connected across the headphones. In the early days of radio, crystal detectors were made using an amazing number of different materials including galena, germanium, and zinc oxide. Another type of simple diode detector classified as a "coherer" actually used iron filings inside a sealed chamber. The filings would move in the presence of (strong) radio frequency signals; the detector was shaken periodically to rescramble the filings.[A]

Electrolytic detectors, a close cousin of the electrolytic rectifier, were also used.

Early amateurs used some clever schemes to get the most out of their homemade detectors. Tricks included the addition of a battery and rheostat to forward bias the detector, so it operated at the most sensitive point in its curve.[B] Although they didn't need any active components, all of these passive detectors suffered from poor sensitivity (by modern standards) and required very long outside antennas. Also, there was no buffering between the tuned circuit and the headphones, which meant that the already poor selectivity was further degraded by the loading of the headphones. Finally, they could only detect AM or "spark" signals.

The regenerative receiver

As **Figure 2B** shows, the regenerative receiver heterodynes a local oscillator signal with the incoming RF to produce an audio frequency output. But the regenerative circuit does double duty, also serving as a very high gain amplifier. The regenerative circuit oscillates, heterodynes, and amplifies simultaneously—all within a single stage.

Because of its use of positive feedback, or "regeneration," a regenerative receiver also has very high selectivity, or "Q." A regenerative detector typically provides an audio output

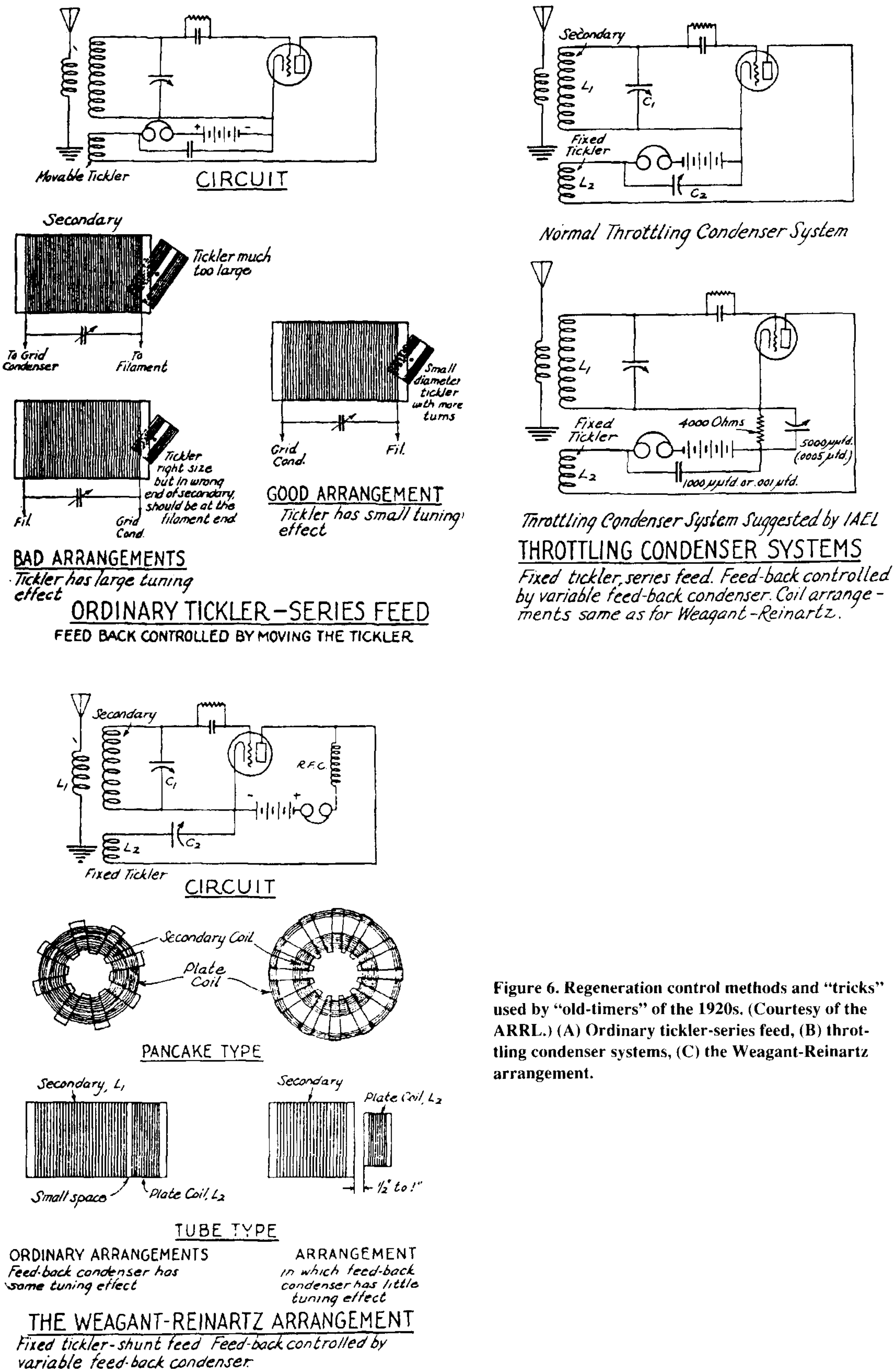


Figure 6. Regeneration control methods and "tricks" used by "old-timers" of the 1920s. (Courtesy of the ARRL.) (A) Ordinary tickler-series feed, (B) throttling condenser systems, (C) the Weagant-Reinartz arrangement.

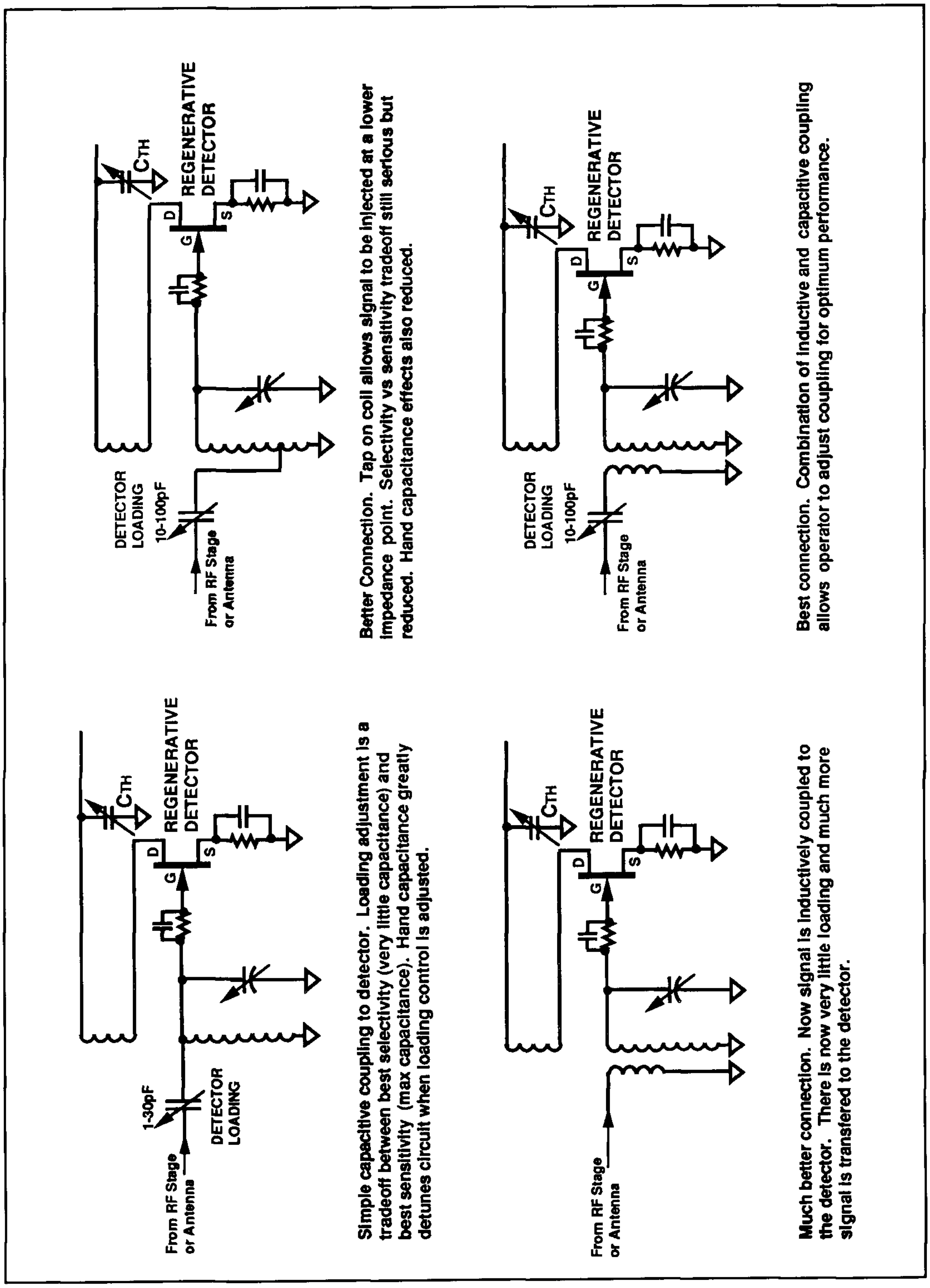


Figure 7. Coupling the RF input signal to the detector.

level of hundreds of millivolts, requiring only a low-gain audio stage. And, due to its inherent high selectivity, high Q band-pass filtering ahead of the detector isn't needed. This greatly simplifies the design.

Another interesting characteristic of regenerative detectors is their ability to detect basically all types of signals including: AM, CW, spark (if there are any left), single sideband (SSB) and frequency modulated signals. Also, a properly designed regenerative receiver characteristically has very high quality audio. Furthermore, because regenerative circuits use far fewer parts, they consume less power, cost less,

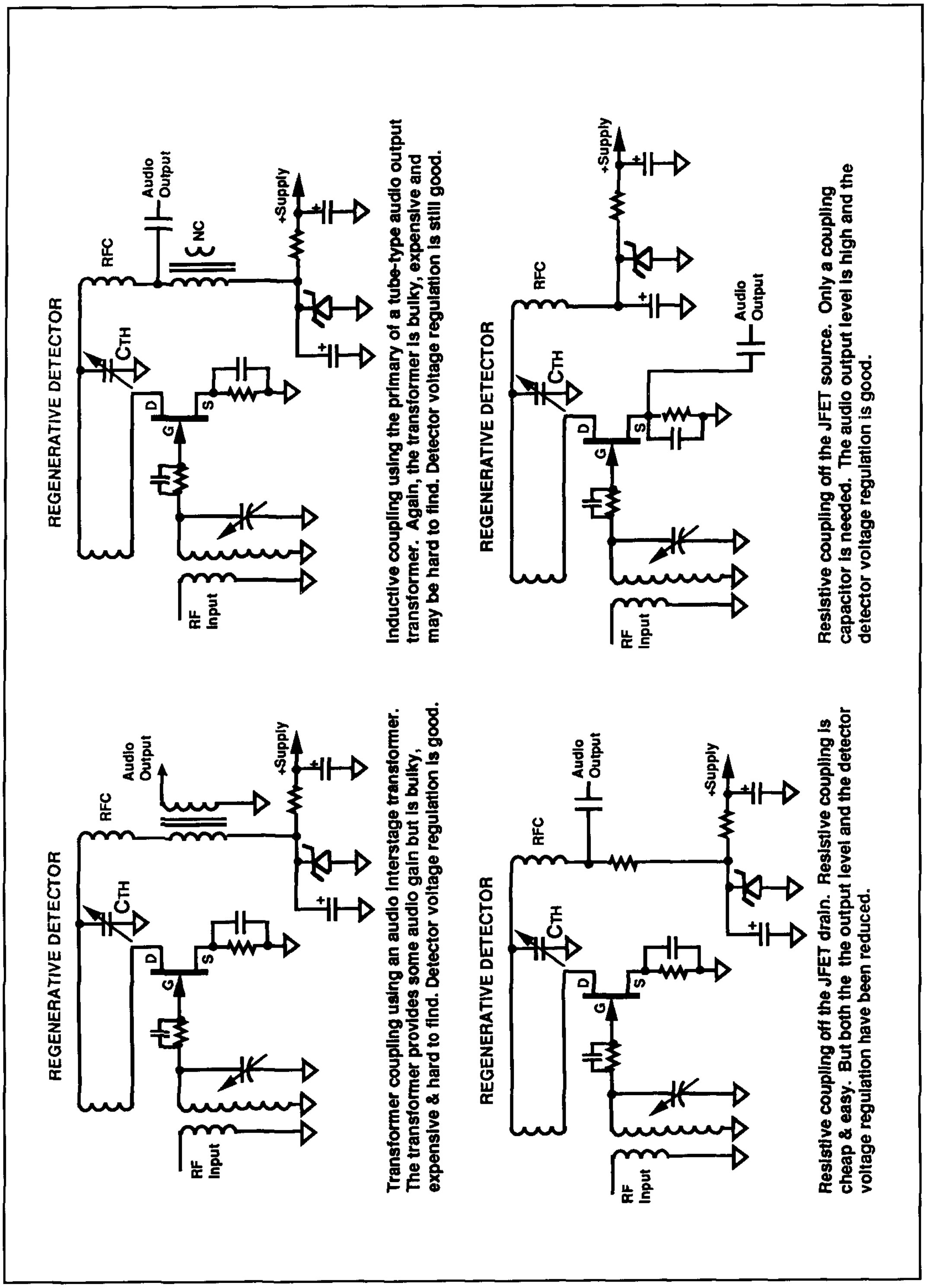
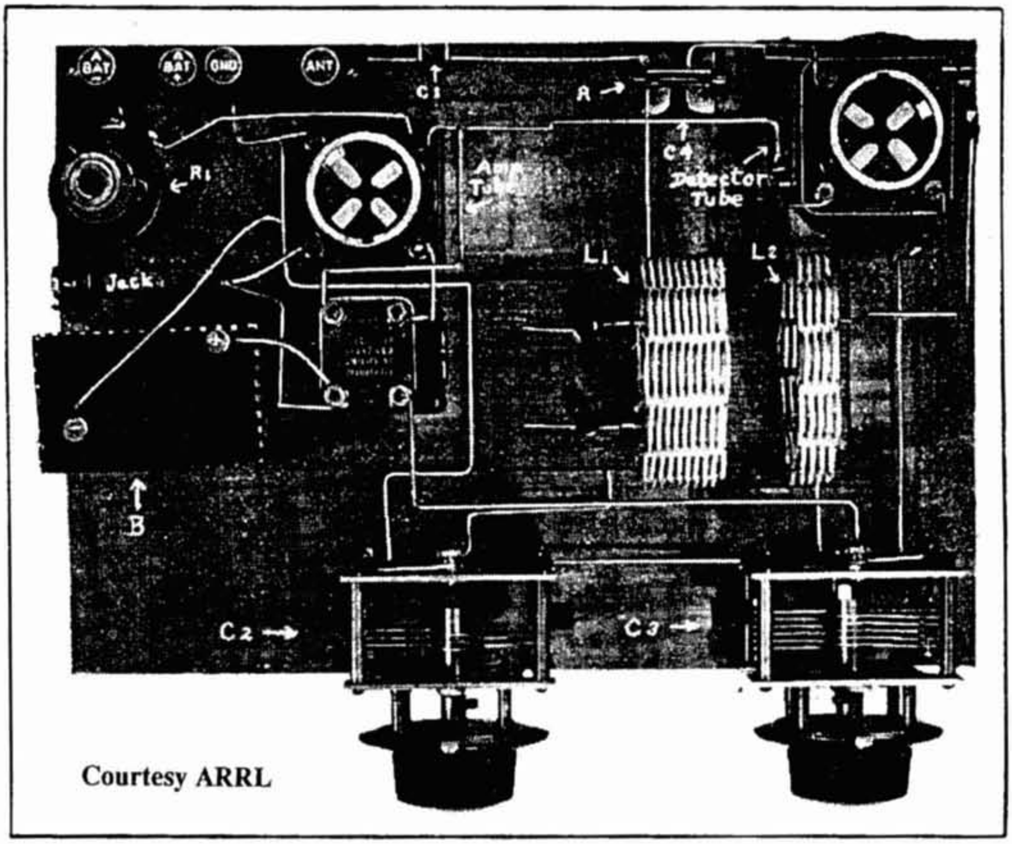


Figure 8. Extracting the audio output from a JFET detector.

and are much easier to homebrew than other receiver types.

Regenerative receiver negatives include detector (oscillator) leakage out to the antenna. This requires some type of isolation: either an RF stage, or a very low power detector combined with effective antenna decoupling. Another inherent problem is the so-called "pulling" effects, where an oscillating detector tends to lock onto the center of strong RF signals requiring some type of variable input attenuator. A regenerative receiver can provide very effective AM (nonheterodyne) reception, but the operator must readjust the receiver frequently.



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2 High-ratio (20:1) dials	3.00
1 audio-frequency amplifying transformer: GE UV712 9:1 ratio (without * beside style number), Thordarson 6:1, General Radio 6:1, or Western Electric 29:1; say	5.00
1 22½ volt block B-battery	2.00
Material for Lorenz coils:	
4 Eby binding posts	
1 pound No. 18 or No. 20 Contenamel wire	1.00
For each secondary coil:	
2 hard rubber strips ¼" x ¾" x 2"	
2 cord tips (unless coil ends are inserted right in binding posts)	
2 6:32 brass machine screws 1" long, and brass washers.	
Coil mounting:	
1 bakelite micarta strip ¼" x 1" x 2¾"	
2 large washers	
2 2" lengths of 3/16" I. D. hard rubber or bakelite sleeving	
2 Carter Imp. jacks (or substitute General Radio plugs and jacks in the set)	
2 2¾" blued round head wood screws	
Terminal Board for Antenna, Ground and A-battery connections:	
1¼" x 1" x 6" bakelite strip	
4 Binding Posts	
Miscellaneous:	
24 wood screws to hold things down, 1 ft. of ⅜" copper strap to fasten down B-battery, 2 copper angles constituting antenna coupling condenser, 4 ft. No. 12 bus wire, 6 brass angles to support condensers, strong light cord to bind Lorenz coils, wooden block ¼" x 4" x 7" for mounting coils (unless *plug* mounting is built). A paper clip or piece of tinned bus wire attached to the frame of the condenser and bent over the top of the dial will make a good indicator lacking a better one. If the particular condensers you get do not permit of mounting the back plate for the vernier dials, some flat 90° angle pieces from the 10c store or the local hardware store will hold them so that the smooth vernier action can be used in tuning. The cost of these miscellaneous items should not greatly exceed	$1.50
Estimated cost of a practical and excellent short wave receiver	$32.00

Figure 9. An amateur "breadboard" receiver of the 1920s and its parts list. (Courtesy of the ARRL.)

Tuned Radio Frequency receivers

A common circuit in early commercial receivers, the tuned radio frequency (TRF) receiver (**Figure 2C**), provided one or more tuned RF gain stages ahead of a diode or low-gain detector and audio amplifying stages after it. Unfortunately, many cascaded tuned RF states were required to provide even modest selectivity, because even a tuned circuit using an air core coil with a Q of 100 would still have a bandwidth of 100 kHz at 10 MHz. In addition, the basic TRF design could only receive AM signals. However, TRF designs that used a regenerative detector had greatly improved performance and CW signals could now be copied.

Heterodyne or "direct conversion" receivers

As illustrated by **Figure 2D**, the heterodyne or "direct conversion" (DC) receiver is similar to the regenerative receiver. It, too, mixes a local oscillator signal with the incoming RF to produce sum and difference frequencies. When the local oscillator is operating just above or below the received frequency, an audio output will result from the reception of CW or SSB signals. If the local oscillator is set to "zero beat" the carrier, makeshift AM reception is also provided. The key difference between the regenerative and DC receivers is that, with regeneration, both the circuit's gain and selectivity are amplified a thousand times or more during the heterodyne process.

Major problems inherent to direct conversion receivers include oscillator leakage out the antenna and a very low-level audio output from the detector, requiring very high audio gain to recover the signal. This makes the DC receiver prone to microphonics. Modern DC homebrewers have come up with some clever designs that overcome these difficulties and provide excellent performance; but, typically, the receiver's parts count is very high. Designs I've seen have required several transistors or other active devices, many toroids, lots of crystals, and large numbers of other components. I believe these complexities have "scared off" many would-be homebrewers.

The superheterodyne receiver

The superheterodyne receiver (**Figure 2E**) mixes the RF signal and that of a local oscillator to produce an intermediate frequency RF signal. The local oscillator tracks the received frequency to ensure that the difference (or the sum) between the two frequencies is always equal to the IF frequency. Most of the amplification is provided at a single, fixed IF frequency using a high gain, high Q, single frequency amplifier (with one or more stages). The superhet, therefore, changes the signal frequency to fit its IF amplifier, rather than trying to create RF amplifiers that can work directly over the entire frequency range of the receiver. Oscillator leakage is minimized in a superhet because the local oscillator doesn't operate at the received frequency and is therefore greatly attenuated by the receiver's input circuitry.

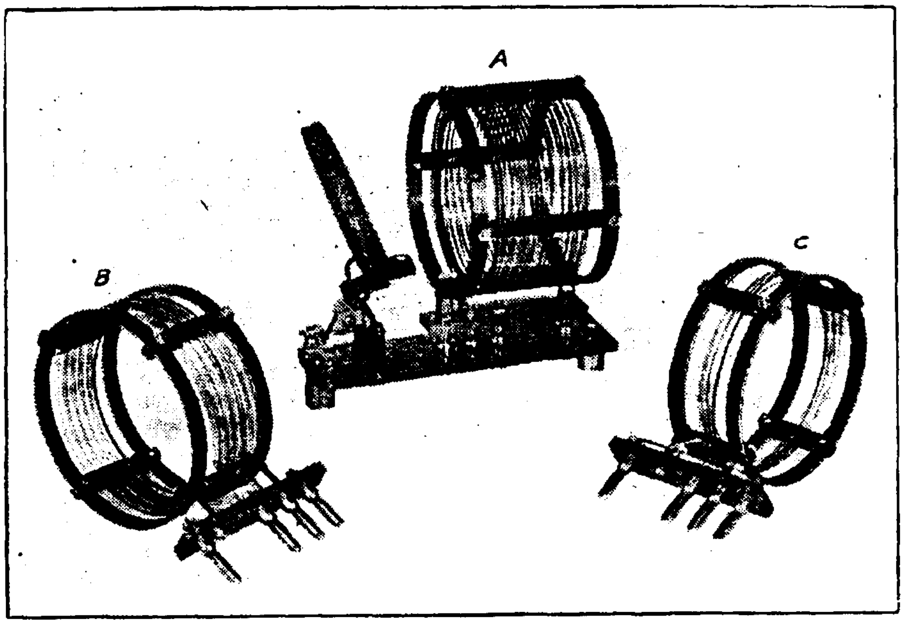
On the negative side, superheterodyne receivers are quite difficult to homebrew unless the range of received frequencies is very small. The reasons for this are: (1) as the receiver is tuned, the local oscillator must closely track mixer tuning over the entire frequency range and (2), very good preselection is needed to prevent the mixer's image from passing through to the IF amplifier. Images occur because the local oscillator can heterodyne with frequencies both above and below that of the IF.[C] Historically, the best superhets have used high Q multistaged preselectors with the RF and mixer stages tracking each other precisely over frequency using multiganged tuning capacitors (each carefully "tweaked"). Unless it's built from a kit, the need for very close tracking and effective image rejection has made the superheterodyne impractical for all but the most skilled (or persistent) homebrewers. Having built two working superhets "from scratch," I can personally attest to these difficulties.

The actual mechanism of regeneration

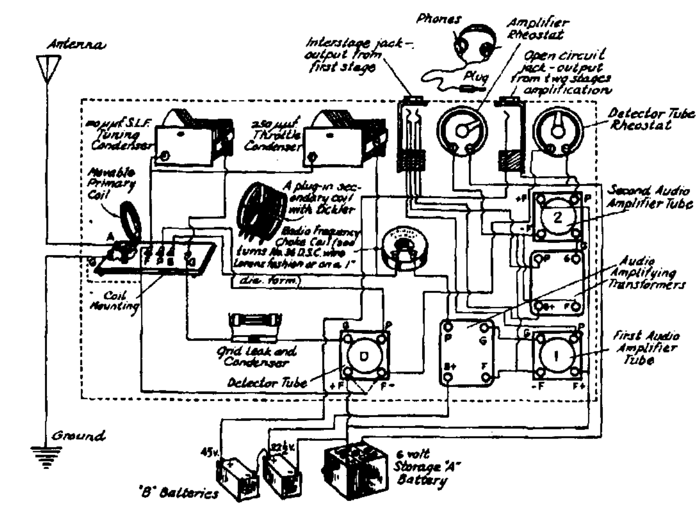
Figure 3 illustrates the basic regenerative circuit. If the output of a radio frequency amplifier is fed back to its input—in phase so the signals add—the input signal will be reamplified over and over, providing a thousand times (or more) increase in gain over a conventional RF amplifying stage. Although the power gain of an active device such as a tube or transistor is fixed, the voltage gain in a regenerative circuit (ideally) approaches infinity as it comes into oscillation. The practical result is that a regenerative detector using a single transistor or JFET can convert microvolt-level RF input signals to hundred-millivolt-level audio output signals.

The actual mechanism of regeneration is complex. It introduces negative resistance into a circuit in such a way that its net positive resistance is reduced. Since the circuit's selectivity or "Q" is equal to its net reactance divided by its net resistance, the circuit's selectivity is also greatly increased when regeneration is introduced. When the regeneration level is below self-oscillation, the circuit's negative resistance (produced by regeneration) is less than its fixed positive resistance. When adjusted to this level, regeneration provides a stable increase in both gain and selectivity.

When more regeneration is applied, the circuit's negative and positive resistances are almost equal. This is a very critical state, just at the threshold of oscillation. The exact "balancing" point where the net circuit resistance is zero is impossible to maintain, as even the smallest random noise source, given time, will



Courtesy ARRL



SHORT WAVE PLUG-IN-COIL RECEIVER – PICTURE DIAGRAM SHOWING WIRING OF DETECTOR AND TWO STEP AUDIO AMPLIFIER

Figure 10. Wiring outline of another amateur receiver of the 1920s, which used commercial plug-in coils. (Courtesy of the ARRL.)

build up to a self-sustained free oscillation.

As regeneration is increased further, the circuit exhibits a net negative resistance and oscillates. As regeneration is increased beyond oscillation, curious secondary oscillations of a lower frequency are introduced that tend to turn off or "quench" the main oscillation under certain conditions of input signal level and degrees of regeneration. Because of the quenching action, RF input signals are amplified tremendously, with circuit gains approaching one million in a single stage. Discovered by Armstrong,[2] the phenomena is called "Super Regeneration" and its discovery led to the development of the first practical VHF receivers (For more information, see "Super Regeneration: The Lost Technology," *Communications Quarterly*, Fall 1994).

In actual use, a regenerative detector performs quite differently depending on whether it's operated above or below the oscillation threshold. For the best sensitivity and selectivi-

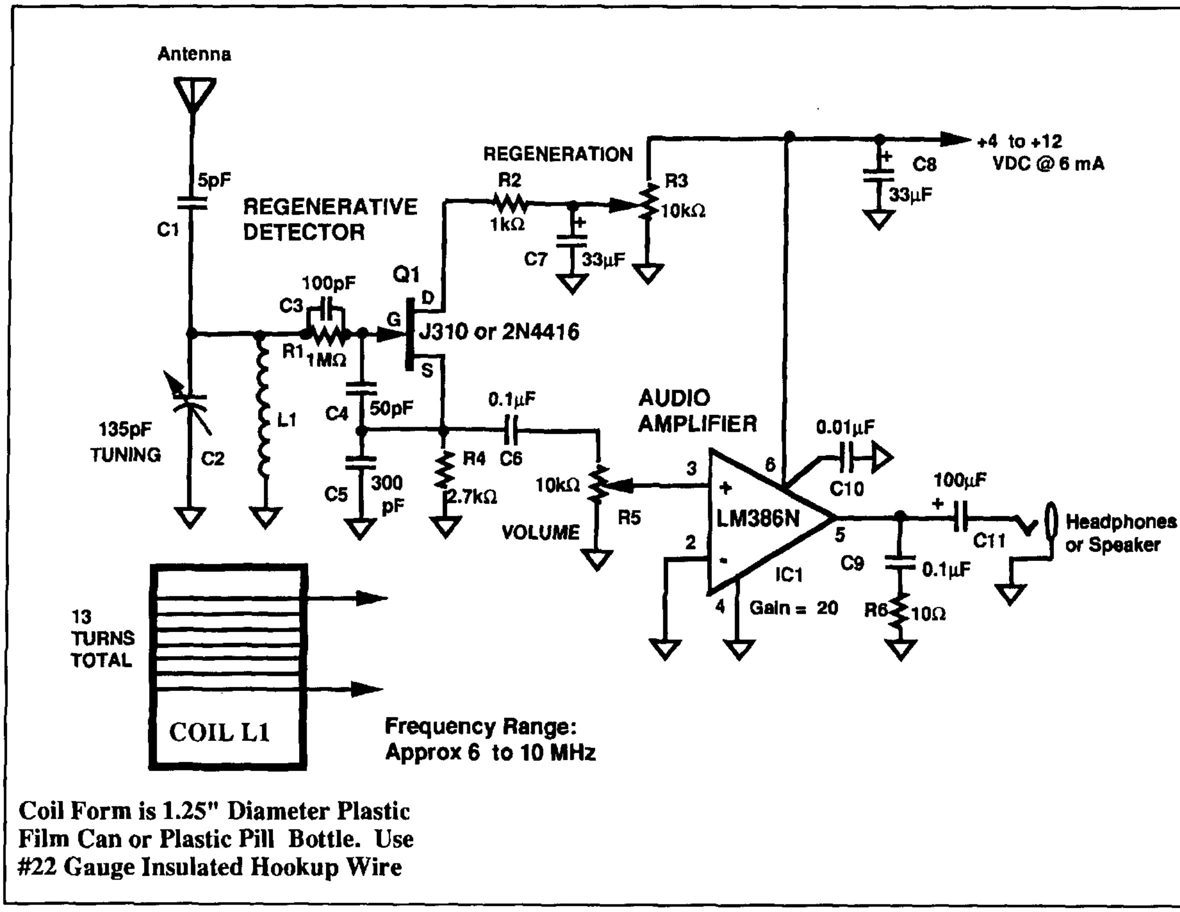


Figure 11. An ultra low-cost shortwave receiver for the beginner.

ty when receiving AM signals, the detector is adjusted just to the threshold of oscillation. Receiver performance can be quite good, but it does require frequent readjustment of the regeneration control and a certain amount of operator skill.

When receiving CW or SSB, the detector is set to oscillate and then detune from the center of the carrier to produce an audio heterodyne or beat note. So-called oscillating or "auto dyne" (self-force) detectors are far more sensitive than any other type. And the "grid leak" biasing normally used in the regenerative detector tends to maintain a constant oscillation amplitude over wide frequency ranges and, therefore, doesn't require frequent readjustment. Furthermore, because the detector is operating beyond the threshold of oscillation, it's less susceptible to static interference than other types.

Key design rules for regenerative detectors

Regenerative detector circuits. Figure 5 shows some of the many possible regenerative detector circuits, all of which are based on standard oscillator circuits. As with an oscillator, a critical design issue is frequency stability over changes in power supply and temperature.

Regeneration control methods. Unlike oscillators, regenerative receivers need a practical method for controlling the amount of regeneration. The "old-timers" of the 1920s used many "tricks" with one common goal: to allow the smooth and gradual adjustment of regeneration up to and beyond the threshold of oscillation, without causing the circuit to detune.

Figures 6A through **6C** from the 1927 *Radio Amateur's Handbook*[3] show some of these tricks. They included winding the tickler coil at the ground end of the main tuning coil; using the smallest possible tickler; winding the tickler on a separate, smaller diameter coil form attached to the bottom of the main form; and using heavier gauge wire for ticklers used in CW receivers than those for AM. One of their most important discoveries was the "throttle condenser" regeneration control method.

"Rotating tickler" control. Figure 6A depicts the rotating tickler method. This technique was commonly used in the 1920s for

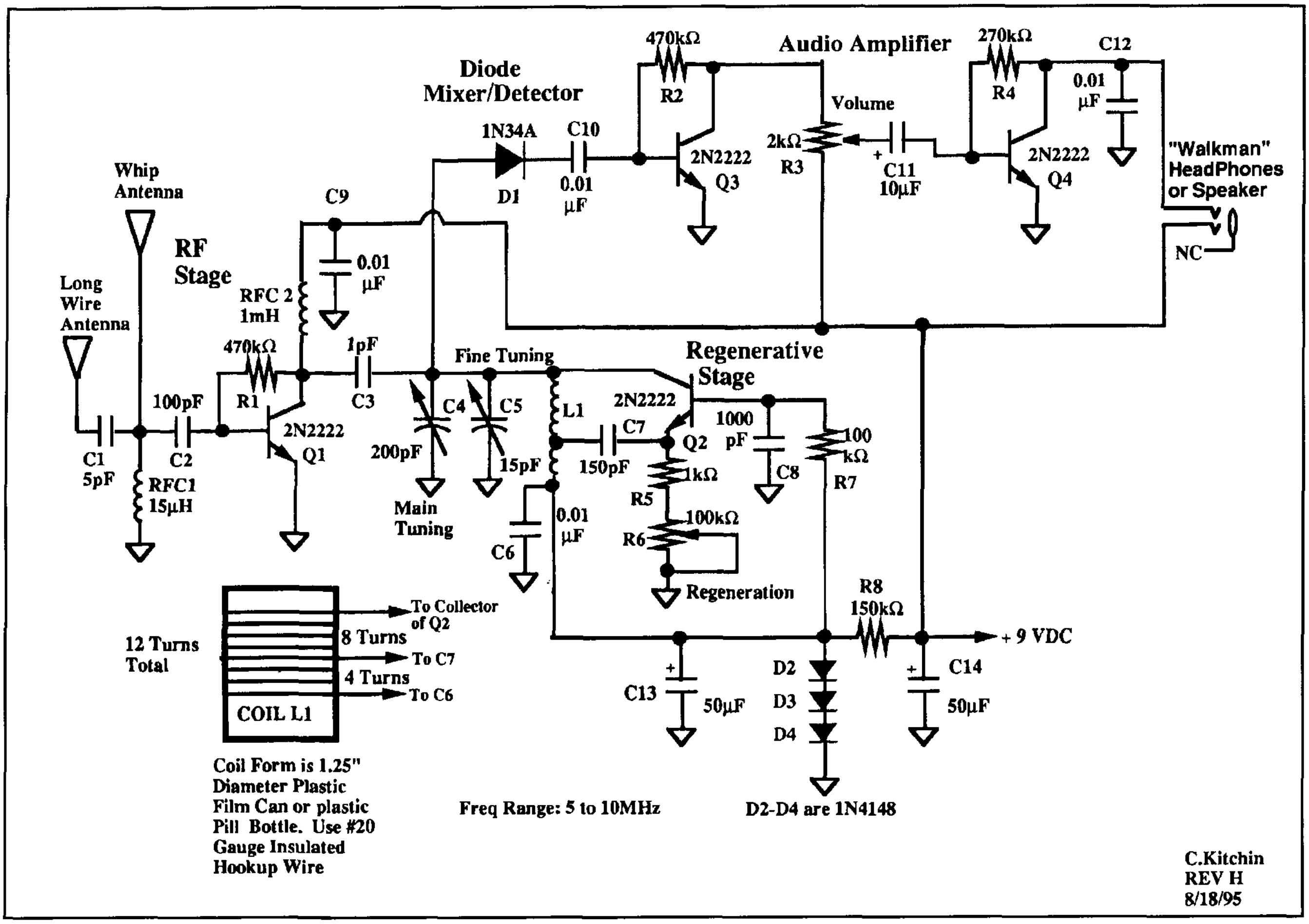


Figure 12. A very sensitive bipolar regenerative shortwave receiver.

broadcast receivers—often using a commercial unit called a "variometer," which consisted of a smaller inner coil rotating inside a larger outer coil.[D] Aside from their cost and complexity, rotating ticklers detuned the signal as regeneration was increased and the detector approached oscillation.

"Throttle capacitor" and resistive regeneration controls. Figures 6B and **6C** illustrate the "throttle capacitor" method of regeneration control. The throttle capacitor method provides by far the smoothest adjustment up to and beyond the threshold of oscillation. Widely used by amateurs in the 1920s, the throttle capacitor method was phased out during the 1930s, and replaced with various resistive control techniques. This was probably due to the fact that higher gain screen grid tubes had replaced the early triode designs and because the simple expedient of using a potentiometer to vary the screen voltage of the detector was cheap and easy to do.

Unfortunately, performance suffered. Resistive controls are noisy and lack the smoothness of a capacitive control. More importantly, the use of a throttle capacitor permits the detector voltage to be regulated—a very desirable feature. A regulated supply allows a much closer adjustment to the oscillation threshold when receiving AM signals and when oscillating, the frequency stability of the detector is also greatly improved. With resistive controls, detector voltage varies widely and there always seems to be some degree of overshoot. The combination of a "throttle" capacitor and a regulated power supply provides the highest sensitivity, selectivity, and stability for a given detector.

JFETs, bipolar transistors, or tubes? JFETs are ideal for use as regenerative detectors. Although they provide much less gain than comparable bipolar devices, their "soft" turn-on characteristics permit much better regeneration control. They are most effective when used with some combination of source and "grid leak" bias. The "grid leak bias" (**Figure 3**) is a parallel RC network in series with the gate that provides an automatic increase in negative bias with increasing RF amplitude. This tends to maintain a near constant output amplitude from the detector when

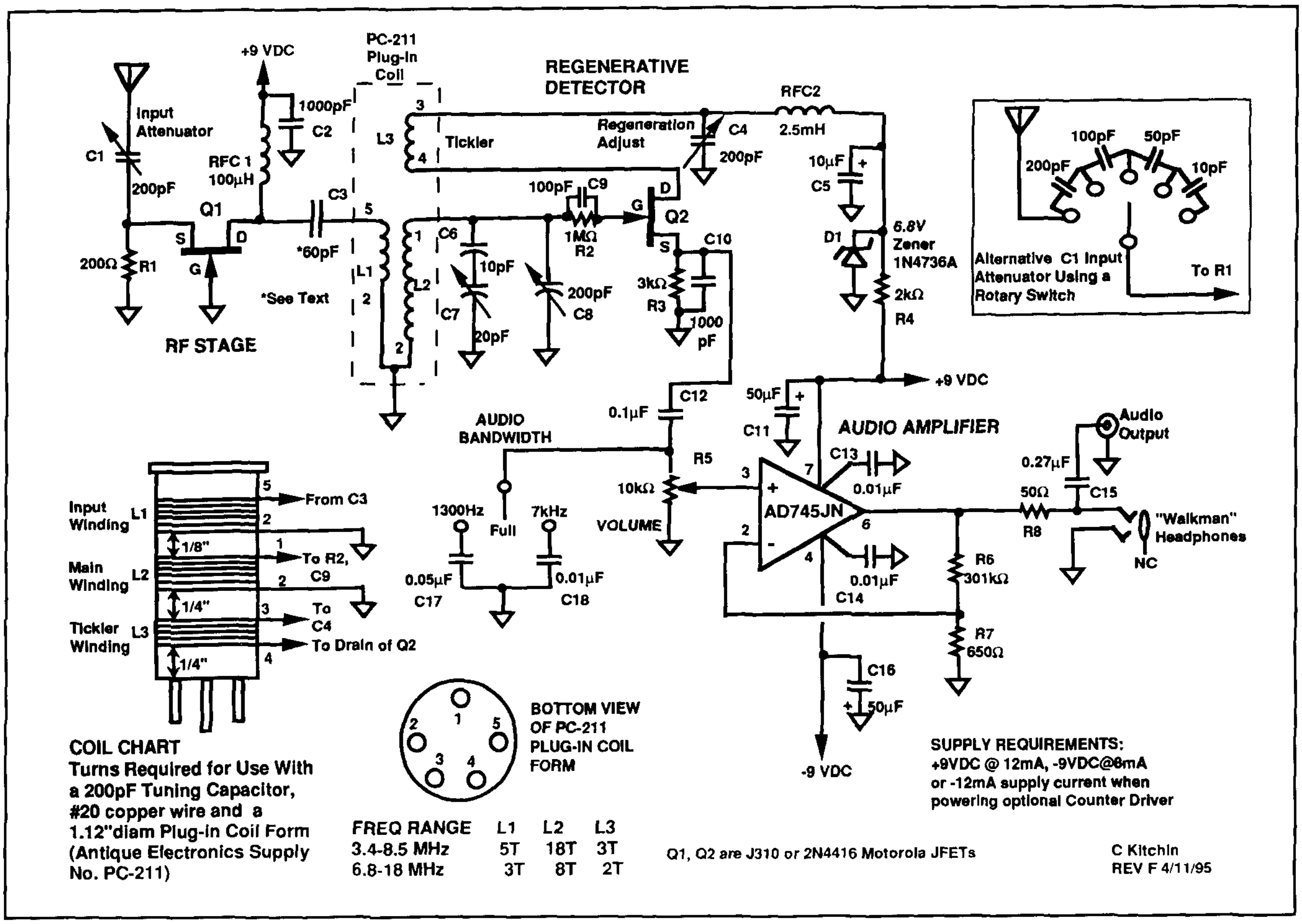


Figure 13. A high-performance JFET shortwave receiver.

it's in an oscillating condition, greatly improving CW and SSB performance. For most JFETs, a 1-meg resistor and 100-pF capacitor are about optimum for the network.

Pentode vacuum tubes have operating characteristics very similar to JFETS, except their output impedance's are much higher and their supply voltages are much greater. Input impedances' are similar, however. Old-timers claimed that tube circuits with very large values of grid leak resistor (2 to 10 megs) provided the best CW performance.

Bipolar transistors have high transconductance that provides very high gain within the regenerative loop. The signal input to the detector must necessarily be kept low to avoid overloading. The use of large amounts of negative bias is very helpful in making the regeneration level in these circuits easier to control. Bipolar detectors provide a high audio output level but, in the oscillating condition, the circuits I've built never seem to work as well as any of my JFET designs.

Detector loading issues. Detector loading is one of the most important issues and tradeoffs involved in the design of regenerative detectors. The ideal state is a condition where the maximum possible input signal can be applied to the detector while not loading it at all. **Figure 7** shows several methods of coupling the RF input signal to the detector.

Tickler turns. Always use the minimum number of turns on the ticker that will permit oscillation throughout the receiver's entire frequency range. This is best determined by trial and error. A good "cookbook" method is to use approximately 1/3 the turns on the tickler as are used on the main tuning coil, test the circuit, and then reduce the turns to the minimum needed.

Detector gain. In 1933, Robinson[4] wrote a classic article in which he detailed a series of experiments investigating how component changes within the feedback loop affect detector gain and selectivity. He recommended using pentode tubes instead of triodes and told readers to first optimize their detectors for best sensitivity in the nonregenerative state before connecting the regenerative loop. Although the use of a high-gain device within the regenerative loop does provide a higher audio output signal,

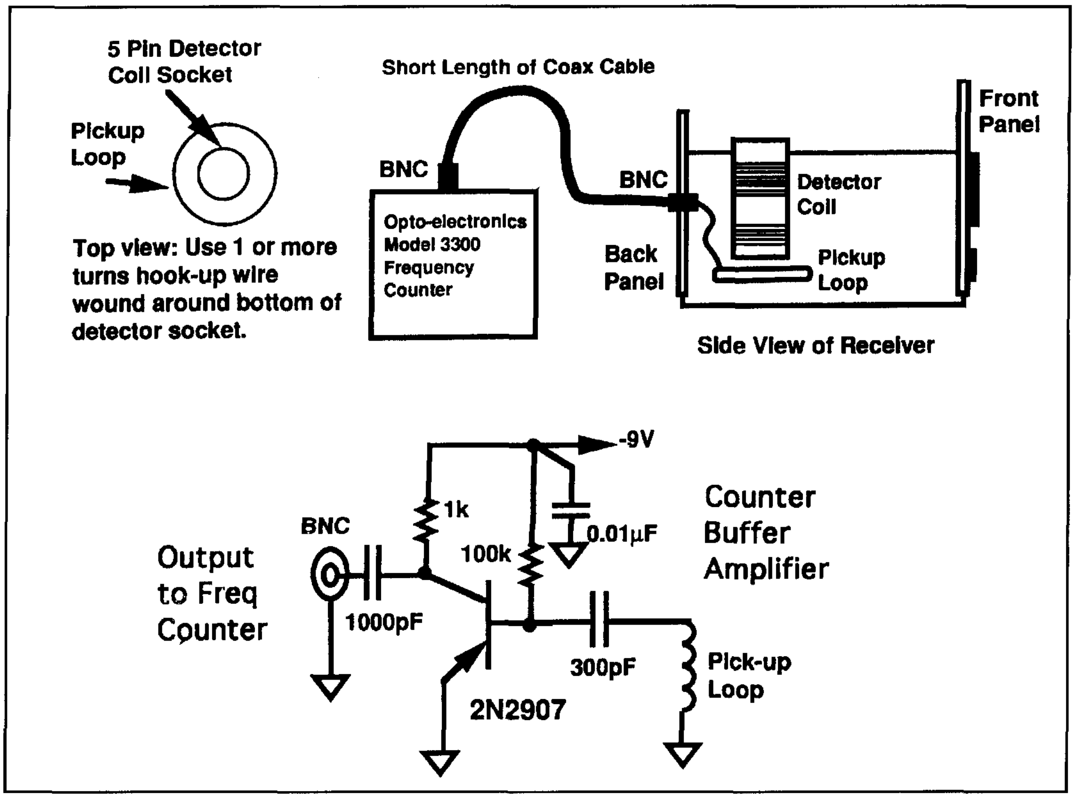


Figure 14. Optional connection to external frequency counter.

the benefits of high gain are misleading. As with any receiver, overall gain is far less important than the signal-to-noise ratio of the first stage. Higher gains within the regenerative loop also contribute to its instability.

Pulling. "Pulling" is an effect that causes two oscillators to become synchronized if they are operating close to the same frequency. In a regenerative receiver, this causes the detector to lock onto the centers of strong RF signals. This can make it difficult to receive strong SSB or CW signals, because the detector needs to oscillate at a frequency just above or below the carrier to produce the necessary heterodyne. Pulling can also cause problems when the detector is receiving AM signals and not oscillating. When many stations are close together, the receiver will tend to jump around, locking onto whichever station is stronger. Pulling can be prevented by simply adding an RF gain control or input attenuator to the receiver.

Receiver power supply. Hum modulation can arise from excessive power supply ripple or from having any power transformers or AC line voltage anywhere on the receiver. Your best bet is to power the receiver with batteries or use a well-filtered power supply that's physically isolated from the receiver chassis by several feet.

RF stages. Use an untuned RF stage preceding the detector. In a regenerative receiver, only a small amount of RF gain is needed; detector isolation is a much more important issue. The use of an untuned stage greatly eases construction, as a tuned RF stage will usually oscillate unless its coil and that of the detector are both shielded. Shielding should be avoided because it increases the dielectric losses that can ruin the receiver's selectivity. A tuned stage can also cause tracking problems. For a single ham band, an RF stage tuned to the center of the band will normally suffice; however for general coverage use, the LC circuit of the RF stage also needs to closely track the detector circuit over frequency.

Extracting the audio output from the detector. **Figure 8** shows several methods of extracting the audio output from a JFET detector; **Figure 8D** is the preferred connection.

Practical regenerative receiver circuits

Classic regenerative receivers of the 1920s. **Figure 9** shows an amateur "breadboard" shortwave receiver from the 1920s, including its parts list and 1927 prices! **Figure 10** shows some commercial plug-in coils available in the 1920s and a block diagram of another receiver from the same period. Note their use of variable input coupling, throttle capacitor regeneration controls, vernier dials, and detector shock mounting. Also note that they simply connected their high-impedance headphones in series with the plate of the audio tube, which was common practice in those days. Because they used their receivers in the oscillating condition (for receiving CW), and they had sensitive, very narrow bandwidth headphones, both the sensitivity and selectivity of these early receivers was probably pretty good. Those interested in building regenerative receivers using vacuum tubes should refer to David Newkirk's article in *QST*.[5]

An ultra low-cost shortwave receiver for the beginner. Figure 11 answers the challenge: "How many parts do you really need to build a useful shortwave receiver?" The entire circuit uses less than $10 worth of parts (not counting speaker or headphones) and operates from almost any battery. Battery current is only 6 mA for the entire receiver. This is an excellent project for a boy (or girl) scout's merit badge, or as a gift for a young child. Regenerative detector Q1 amplifies microvolt level input signals up to hundreds of millivolts to drive an LM386 audio op amp. Regeneration is simply (and a bit crudely) adjusted by potentiometer R3. R1 and C3 provide "grid leak" bias for the JFET while R4 adds source bias. Some circuit layouts may require a few pF between the drain of Q1 and ground to permit the circuit to oscillate. For best results, use a 50 foot long wire antenna with this receiver.

A bipolar regenerative shortwave receiver. Figure 12 shows the circuit for a very sensitive shortwave receiver using 10-cent bipolar transistors. With bipolar transistors and an RF stage, this receiver is sensitive enough to be used with just a short whip antenna. Transistor Q1 is used as an untuned RF amplifier with L1/C1 forming a high-pass filter to block interference from any nearby AM broadcast stations. Q2 is a high-gain regenerative RF amplifier connected in a Hartley oscillator configuration. Potentiometer R6 allows smooth regeneration by increasing the negative bias on Q2, and the amount of positive feedback simultaneously. Diodes D2-D4 provide simple voltage regulation. The output from the regenerative stage drives diode detector D1, which connects directly to a two-stage audio amplifier. An op amp or LM386 circuit could be substituted for the audio stage to reduce the parts count. To further limit the parts count, cost, (and performance) omit the RF stage and increase C3 to approximately 5 pF. Then connect an external antenna directly to C3.

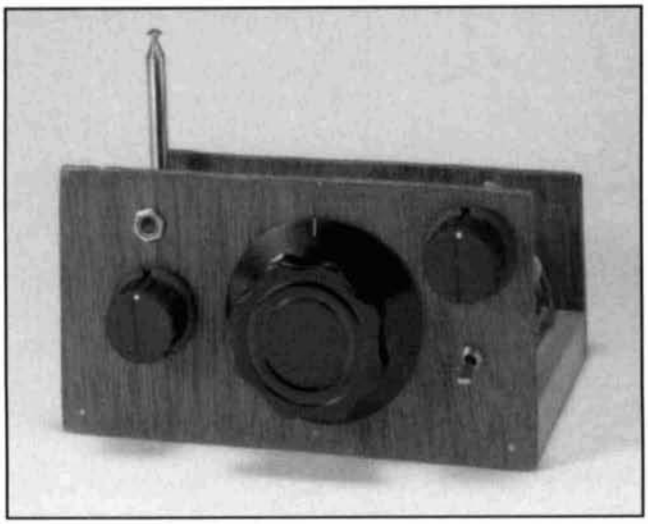


Photo A. Sensitive bipolar receiver, external view.

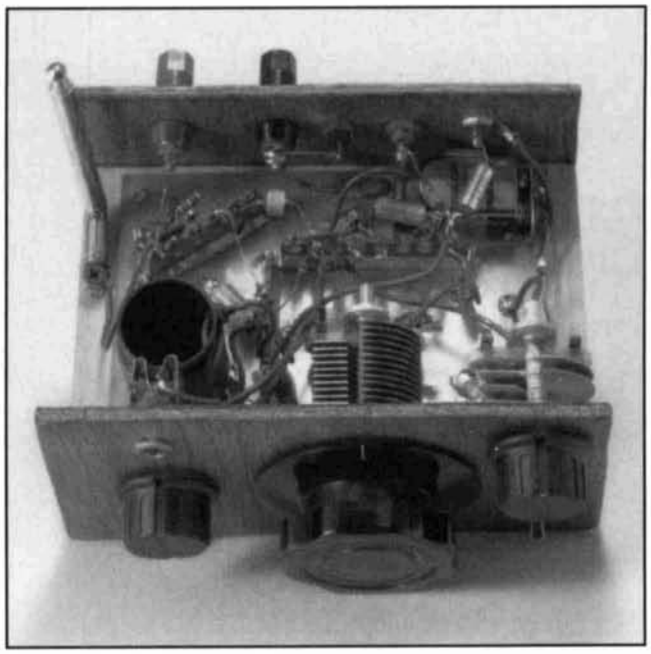


Photo B. Sensitive bipolar receiver, internal view.

A high-performance JFET shortwave receiver

Figure 13 shows a highly sensitive and selective shortwave receiver appropriate for either general coverage or ham band reception. It features an RF stage with a built-in input attenuator, a high-performance JFET regenerative detector, selectable audio filtering, and a high-gain low-noise audio output stage. The overall performance of this circuit equals or exceeds that of many superheterodyne designs, yet it has a very low parts count and draws less than 12 mA from its two 9-volt batteries.

JFET Q1 operates as a grounded gate RF stage to improve sensitivity and isolate the detector from the antenna. Capacitor C1 is a variable input attenuator that's very useful for eliminating "pulling" effects or for increasing the receiver's selectivity when receiving strong

Photo C. High-performance JFET receiver, external view.

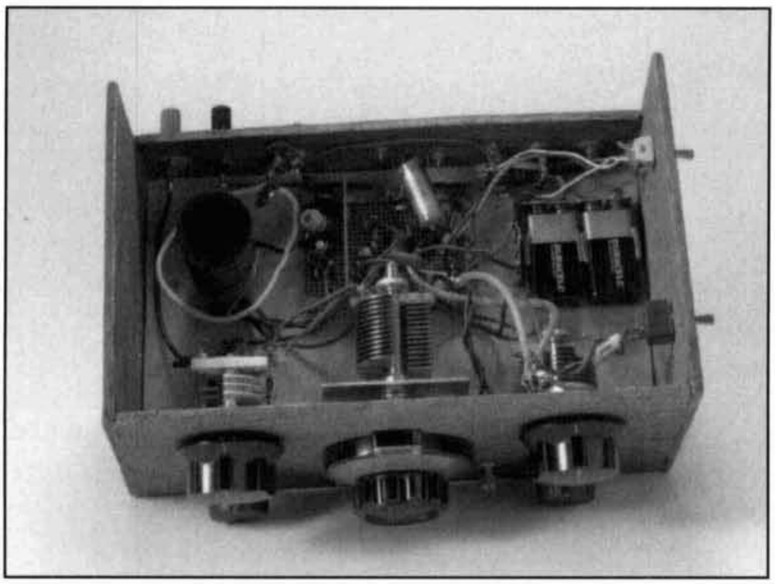


Photo D. High-performance JFET receiver, internal view.

signals. RFC 1 provides a DC path for Q1's drain. The RF output from Q1 is AC coupled to L1, which then inductively couples the signal to the detector. Note that RFC 1 can be almost any value above 50 μH as the inductance of L1 predominates. Capacitor C3 sets the amount of detector loading with 60 pF being about optimum. Secondary winding L2 and capacitors C6-C8 select the received signal, while tickler winding L3 provides regenerative feedback. R2/C9 are the usual "grid leak" arrangement that together with R3 set a very high level of operating bias for the JFET, making regeneration control much smoother.

Capacitor C4 is the "throttle capacitor" regeneration control, while RFC2 prevents the RF signal from being lost through the supply. Zener diode D1 regulates the drain voltage of the detector, so it's very stable in the oscillating mode. Commercial plug-in coil forms (see list of parts suppliers) are used to allow multiband operating (bandswitching would be difficult with 3 windings). When winding the coils, the same winding on each coil should occupy approximately the same space (expand or compress turns as necessary) and the spacing between windings should also be the same. This helps to reduce the distributed capacitance. The frequency range of this or any of the receivers in this article may be extended up or down by simply winding additional plug-in coils.

The audio output is extracted from the JFET source and low pass filtered by the switched audio filter. Audio feeds from the volume control R5 to an Analog Devices' AD745 op amp. This JFET op amp provides high gain and very low noise, high quality audio. The op amp output drives series connected (common not used) Sony "Walkman" type stereo headphones.

The optional buffer amplifier of **Figure 14** can be added to drive an external low-cost frequency counter. A pick-up loop located close to the detector coil extracts some of the signal from the detector when it's oscillating or close to oscillation. A bipolar transistor amplifies the loop signal and isolates the load presented by the BNC from the detector coil. The BNC cable connects directly to a low-cost commercial frequency counter.

To obtain the best frequency stability from this receiver on CW and SSB, adjust the regeneration control so the detector is operated well beyond the oscillation threshold. For single sideband, use the input attenuator to reduce the signal strength enough to avoid "pulling." For AM reception, use two hands—one for tuning, and the other for regeneration—and use the input attenuator to help separate stations on a crowded band.

Construction guidelines

Use a wooden chassis. The original concept of a breadboard was just that: a hard wooden cutting board. There are some very good reasons for using this type of chassis. First, the detector coil must be kept as far away as possible from any metal object. A metal chassis, shield cans, metal sides and back panels all absorb energy from the circuit and add to the dielectric losses, which directly affect the overall circuit "Q" and the receiver selectivity. The optimum setup for a homebrew receiver is to use a wooden base, sides, and back, and a grounded metal front panel. A wooden front panel can also be used if the metal bodies of the controls are grounded and a small, grounded, sheet metal plate is used between the fine tuning control and the back of the front panel (to reduce hand capacitance effects).

As shown, fine tuning can be incorporated into these receivers by adding a small variable capacitor in parallel with the main tuning capacitor. For general coverage receivers, use a vernier dial on the main tuning capacitor; for ham receivers, put the vernier on the fine tuning control for maximum band spread on the amateur bands.

Build the electronics on a small fiber glass board and screw that down to the wooden base board. I prefer poplar for the receiver base because it's commonly available and fairly hard. Oak looks better, but is more difficult to cut and drill. Softwoods like pine are okay, but dent easily.

Before winding a coil, I recommended that you test the coil form in a microwave oven for a minute or so. If it heats up, the dielectric absorption is too high. When winding coils, first drill two small holes in the coil form at the beginning of each winding. Next, feed the wire through the first hole, then out again through the second. If you're using insulated hook-up wire, simply tie a knot at the point in the wire where it enters the form—this will keep the wire from loosening up later on. Then wind the coil tightly onto the form. When the winding is finished, drill two more holes at the end of the form, to hold the winding in place, and feed the wire through. Be sure to sandpaper the ends of copper enameled wire before soldering connections. When the coils are finished (and working correctly) use Q dope to cement the windings firmly to the form.

New plug-in coil forms are available from Antique Electronics Supply (see list of parts suppliers). Old vacuum tubes can be "cannibalized" and their bases attached to thin-walled PVC drain pipe, plastic pill bottles, or plastic film cans. With plug-in coils, run the wires from each winding inside the coil form and solder them to one of the pins, making sure all the coils are wired exactly the same. The completed coil form then plugs into a companion tube socket attached to the receiver's base. Be sure to locate the coil at least 1 inch away from any metal object.

Component layout for these high-frequency RF circuit designs is very important. All ground leads should be kept as short as possible. Try to keep the audio wiring physically separated from the RF wiring. The volume controls should be connected using shielded wire. Also, connect a separate wire between the ground side of the control and chassis ground to avoid control loops through the shield. The ICs should have their power supply bypass capacitors located right at the chip using short leads to the ground. Most of the circuits for this article were built using Vector type 4112-4 plug board, a predrilled universal breadboard with a ground plane on one side. Standard low-cost fiber glass board can also be used if a copper-clad board is located below it on spacers and all grounds are made to the copper board.

Testing

Always wire receiver circuits **backwards**; i.e., first the audio stage, then the detector, then the RF stage. As each stage is completed, test it before going on to the next.

To test any of these receivers, first connect the batteries, plug in a set of headphones, and turn the volume control up half way. Test the audio stage with an audio oscillator or just place your finger on the top of the control and listen for a buzz in the headphones. Test the detector by first seeing if it oscillates. Starting from minimum, slowly adjust the regeneration control until the detector produces a "live sound" (a large increase in the background noise). If the detector refuses to oscillate, carefully check the wiring and, if it seems okay, try swapping the wires to the tickler winding. Once the detector is oscillating, temporarily connect an antenna in series with a 5 to 20 pF capacitor to the top of the main tuning coil. Tune in a strong station, adjusting the regeneration level and volume for best reception. When the receiver is completely wired, test the RF stage by connecting the antenna to the receiver's input and tune in the same station. Reception should be at least as good as without the RF stage.

Future developments

Since regenerative receiver circuit development basically ended many years ago—before homebrewers had low-cost semiconductors, op amps, or zener regulators at their disposal—it should now be possible to design some new features into this traditional circuit. These could include some type of regenerative AGC circuit for AM signal reception—using an op amp and rectifier, etc., to maintain a constant level of regeneration as the receiver is tuned. Crystal oscillator circuits could also be made regenerative to produce a very high gain, high "Q", low drift, single frequency amplifier (for IF stages, etc.). Applying regeneration in a limited controlled way into several RF stages simultaneously (to raise their gain and "Q") is also a possibility. I would very much like to correspond with others interested in exploring this very fascinating technology.

Parts suppliers

Parts suppliers for the receivers described in

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VILCCDVOVE INTO CONCEPTION

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ICE WILL BE FOUND IN MOST OF THE BAYS ON FRIDAY. MORE ICE WILL MOVE

TO SOUTH OF CAPE RACE AND SOME WILL BE FOUND IN THE VICINITY OF

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QJLTHE PERIOD.
 ICE WILL MOVE INTO THE TONGUE OF
WATER TO
EH OLN.


LABRADOR...ICE WILL PACK AGAINST THE COAST DURING THE

PERIOD.+MODERATE ICE PRESSURE EXPEMTD.


END/DD
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Figure 15. Sample RTTY and FAX printouts. (Courtesy John Hartman, NM1H.)

this article may be purchased from the following sources:

Tuning capacitors, tubes, tube sockets, transformers, knobs, potentiometers, wire, resistors, fixed capacitors, and other miscellaneous components: (new) Antique Electronics Supply, 6221 S. Maple Avenue, Tempe, Arizona 85283. Phone: 602-820-5411. (surplus) Fair Radio Sales Co., P.O. Box 1105, 1016 E. Eureka Street, Lima, Ohio 45802. Phone: 419-223-2196.

Plug-in coil forms: Antique Electronics Supply (above).

RF chokes and coils: Antique Electronics Supply (above); Digi-Key Corp., 701 Brooks Avenue, South, P.O. Box 677, Thief River Falls, Minnesota 56701-0677. Phone: 800-344-4539.

Vernier dials: Fair Radio Sales Co. (above); Ocean State Electronics, P.O. Box 1458, 6 Industrial Drive, Westerly, Rhode Island 02891. Phone: 401-596-3080.

Knobs, headphone jacks, 2N2222 transistors, National Semiconductor LM386 ICs, miscellaneous items: Radio Shack.

The Analog Devices AD745 op amps and the Motorola J310 and 2N4416 JFETs: Newark Electronics.

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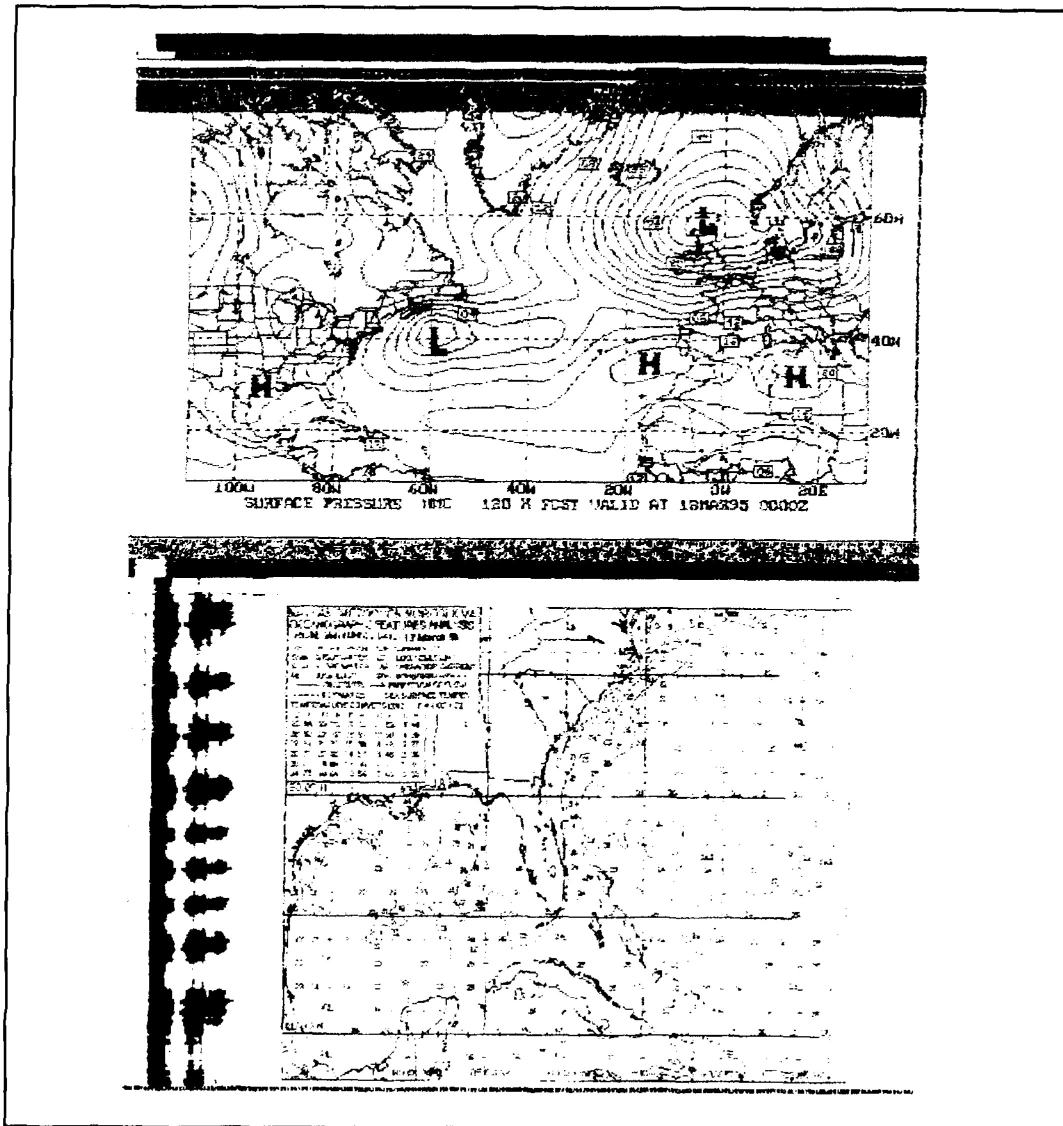
EDITOR'S NOTES

A. The operation of the coherer detector is more mechanical than electron-



ic, and is not a true rectifying detector. The presence of strong RF currents would cause the fine metal filings to "coher," or bond together, thus reducing the electrical resistance path to the recording instrument. In operation, the coherer more or less acted as a one-shot detector device. (Coherers usually drove sensitive relays which in turn operated crude mechanical recording devices.) A secondary function of the relay arm employed a mechanical "bell-type" clapper arm, which upon detection of a signal, tapped the coherer, resettling the filings in preparation for the next "signal."

The first coherer was demonstrated by Parisian Eduoard Branley in the 1880's, it used iron filings packed loosely in a glass tube capped with two electrodes. By 1895 Marconi had discovered that using a mixture of nickel and silver filings greatly improved the coherer's sensitivity.

B. Many substances were tried as detectors, the most popular was lead ore, or galena crystal, a natural substance. Despite the sensitivity advantages offered by galena and other point-contact mineral detectors, they all suffered from sensitivity to burn out from strong RF fields and induced currents from nearby lightning strikes. Elaborate protection schemes were used to isolate the detectors when the spark gap transmitters were fired up. They were also finicky to adjust, and were easily jarred out of alignment. Several fixed Carborundum detectors in my collection still function to this day.

To counteract these problems, a detector was developed using Carborundum—which is man-made and produced in high temperature electric furnaces. The Carborundum detector used a relatively high pressure contact system, and was immune to damage from strong RF fields and unaffected by vibration. Unlike any other detector, the Carborundum detector often used a small variable bias voltage source to improve its performance. The Carborundum detector was also better suited for use in high impedance circuits.

C. Early superhets were as complicated to tune as their TRF cousins. As an example, RCA's Radiola 26, an early "portable" superhet made in 1925, employed separate tuning dials for oscillator and RF stages.

D. To reduce cost, and eliminate expensive variable capacitors whenever possible, Variometers often were employed as main tuning elements. A variometer's winding was done in two sections, the first on a fixed cylindrical form, the second winding was wound on a rotatable coil form inside of the first. A shaft tuning arrangement allowed the inside coil to be rotated nearly 180 degrees, for series aiding or opposing tuning resulting a wide range of inductive tuning.

Although similar in design to the Variometer, the Variocoupler employed two discrete windings. Variocouplers were used to control antenna coupling, oscillator feedback, regeneration and interstage coupling.

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Appendix A: Armstrong's 1914 Circuit

In 1914, while conducting experiments in his home workshop, Edwin Howard Armstrong discovered regeneration, which is certainly one of the most significant discoveries in the history of radio. While others before him may have inadvertently caused circuits to oscillate,

Armstrong was the first person to clearly understand what was happening and to develop practical circuits using this principle. In his classic 1915 paper, Armstrong[1] introduced this new technology to the world.

Armstrong's original circuit for the regenerative detector is shown in **Figure 4**. In this 1915 circuit, a vacuum tube is the amplifying device. The tube is drawn horizontally with the filament on the left, the grid in the center, and the "wing" (today called the plate) at the right.

Small changes in grid voltage would cause large current variations in the plate or "wing" providing gain. A coupling unit supplied positive feedback via the interaction of two coils—L2, a coil in series with the main tuning coil and a feedback or "tickler" coil, L3. Armstrong varied the amount of feedback by manually adjusting the spacing of the two coils; the two windings were brought into close proximity with each other, the mutual coupling between the coils providing the feedback path. Two batteries operated the circuit: an "A" battery (usually a 6-volt car battery) to power the filaments and a 22.5 or 45 volt "B" battery to supply the voltage to the "wing" (plate) circuit.

Appendix B: Independent Evaluation of the High Performance JFET Receiver of Figure 14 by Amateur Extra Class Licensee John Hartman, NM1H.

My Evaluation of Charles Kitchin's Receiver.

Those who think that the days are gone when you could build your own simple communications receiver that rivals professional amateur gear in basic performance, should consider this rig. I had the opportunity to evaluate it one night and was impressed by its sensitivity, selectivity, excellent audio, very low noise, and surprisingly good stability.

I connected the receiver to my all-band Windom antenna and tuned up on the 80 and 40-meter amateur bands. The antenna loading control proved handy in preventing overload and blocking due to the better than 1 microvolt sensitivity of this receiver and its lack of AGC. CW was very easy to copy on the novice and extra bands. However, with the audio bandwidth switch in the "full" position, more than one station was heard, due to the 15 kHz or so bandpass of this receiver. Using the audio filters, I was able to attenuate the higher frequency CW tones. This, combined with no noticeable drift, allowed me to easily copy CW. Tuning up to the voice part of the amateur bands, I was surprised to be able to easily tune in SSB. The audio quality was better than any Japanese transceivers I've heard. I could detect no background hiss, and again, no drift.

Being a serious shortwave listener, I then tuned in HCJB, Equator. I was treated to what I can only say was AM high fidelity (be sure to use a high-quality headset for this rig). The audio was very good. I could even hear the announcer's breathing!

Now for the acid test. I connected my MFJ 1278 multi-mode TNC to the receiver's audio output jack. I then tuned to the 4 and 8 MHz marine bands. Not only was I able to copy almost perfect RTTY, I was also able to print several weather FAX maps. Again, no drift and no frequency tweaking was necessary on these RTTY and FAX printouts (see examples, **Figure 15**). Coast station CW decoding was a problem due to the 200-Hz input filter on the MFJ and the relatively coarse fine tuning control on the receiver.

So, what are the other negatives, besides the coarse fine tuning control? Well, this is a regenerative receiver, so you **will** have to learn to fiddle with regen, antenna coupling, and volume for best results. But, it's like driving a car with a standard shift. Once you get used to it, it can be fun! Like most hackers turned amateurs, Kitchin's workmanship left a little to be desired. He used loose, multi-stranded wire for point-to-point wiring in the RF stage, so there were some microphonics. However, Kitchin's use of a pick-up loop, buffer amplifier, and frequency counter for a direct frequency readout solves the classic regen "where am I on the band?" problem. I did find some frequency readout error, but this could have been my tuning.

All in all, this is a very good performing receiver that's definitely worth the effort to construct—and, for a fraction of the cost of anything you could buy (minus the $100 frequency meter). This would make a nice project for a son or daughter to build, as well as the serious SWLer. In the words of Winston Churchill, if he had been a hacker: "never in the course of human endeavor has so much performance come from so few components."

John Hartman

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BUILD A 2-CHIP, 80-dB RF POWER METER

Build this project as a stand-alone power meter or an accessory to your handheld DMM

Typical low-cost power meter circuits available to the radio amateur sample a portion of the signal being measured, rectify it, and provide a DC output voltage proportional to that signal. Laboratory-quality power meters are not only more accurate, but also orders of magnitude more expensive. These use calibrated thermal sensors or square-law detectors (i.e., diodes) with limited dynamic range.

Fortunately for the amateur radio fraternity, there is a third method of measuring power: the log amp. Low-cost log amps are widely available because they are used in cellular radio. Cellular standards require that the receiver measure the received signal strength and report the level back to the transmitter. One purpose is to minimize interference by using the minimum power necessary to maintain communications. The function is called Received Signal Strength Indicator (RSSI) and is required by the AMPS (Advanced Mobile Phone System) and GSM (Group Speciale Mobile or Global System for Mobile Communications) cellular radio standards, among others.

In contrast to the limited dynamic range (about 30 dB) of the thermistor and diode-based sensors used in laboratory power meters, a well-designed log amp can measure signals from less than –75 dBm to more than +5 dBm—an 80-dB dynamic range. Because of this sensitivity, simple resistive attenuators preceding the log amp can shift the dynamic range upwards by the amount of attenuation. A 30-dB attenuator placed in front of a –75 dBm to +5 dBm log amp can shift the dynamic range to –45 dBm to +35 dBm (+36 dBm ≈ 2W). Spectrum analyzers use log amps to measure the level of the IF to produce displays with 50 to 80 dB of dynamic range.

How does a log amp measure power? The output of a log amp is a DC voltage or current proportional to the log of the input voltage. The term "logarithmic converter" might be more accurate than log amp, but the name log amp has stuck. In the case of the AD606 log amp, which has a voltage output, the transfer function is:

$$V_{LOG} = V_Y \log_{10} \frac{V_{INPUT}}{V_X} \qquad (1)$$

where V_{LOG} is the output voltage, V_{INPUT} is the input voltage, and V_Y and V_X are fixed voltages that determine the log amp's slope and intercept, respectively. V_X is called the logarithmic intercept because, when $V_{INPUT} = V_X$, the argument of the log is 1, $V_{LOG} = 0$, and the transfer function of the log amp (when plotted on a logarithmic X linear Y scale) crosses the X axis at this point. V_Y is called the logarithmic slope because the slope of the transfer function is proportional to this voltage. In terms of the classic slope-intercept equation:

$$y = mx + b \quad \textbf{(2)}$$

the output, y, equals V_{LOG}, the slope, m, equals V_Y, and the intercept, b, equals V_X.

Notice that a log amp responds to the input voltage, V_{IN}. As long as the input signal is a sinewave, as it usually is in amateur applications:

$$P = \frac{V^2}{R} \quad \textbf{(3)}$$

If you take the log of both sides, and perform a bit of manipulation, you get:

$$\log_{10}P = 2\log_{10}V - \log_{10}R \quad \textbf{(4)}$$

which says that the log of the power is proportional to the log of the input voltage (assuming a sinewave). Assuming that the input signal is a sinewave, and the input impedance is 50 ohms, then the AD606 transfer function becomes:

$$V_{LOG} = V_Y (P_{IN} - P_X) \quad \textbf{(5)}$$

where P_{IN} is the input power in dBm and P_X is the logarithmic offset (derived from V_X) in dBm. (The transformation from **Equation 4** to **Equation 5** is sometimes described in the literature as: "Then a Great Miracle Happens!")

The values for the slope and intercept, respectively, for the AD606 are V_Y = 37.5 mV/dB and P_X = –88.33 dBm, both at 10.7 MHz and a +5 volt supply. The slope falls off gradually versus frequency; at 45 MHz it is 35 mV/dB.

In both cellular radio systems and the power meter circuit I'm going to describe, the log amp's output is digitized by an A/D converter. The RSSI output of the AD606 is proportional to the power supply voltage. Thus, the AD606's log output can be expressed not only as 37.5 mV/dB, but also as (0.0075 x VPOS)/dB. The power supply (reduced by a simple voltage divider) can serve as the voltage reference for the A/D converter. This allows the log amp's output and the A/D converter's reference to track over power supply variations because both use the power supply as the reference.

Circuit description

Now, on to the circuit. The power meter circuit (**Figures 1A** and **B**) uses just two active devices: the AD606 log amp and the Harris ICL7136 3-1/2 digit DMM IC. The ICL7136 is a low-cost, low-power combined A/D converter and 3-1/2 digit display driver—a complete DMM on a chip. It is available from several manufacturers (and distributors) and in several incarnations. The +5 volts configuration of the ICL7136 was copied exactly from the data sheet and was not amenable to circuit changes.

This meter circuit takes the nominal output of the AD606, which is 0.5 to 3.5 volts for a –75 dBm to +5 dBm input, scales it, and level shifts it. A simple three-resistor network (R1, R2, and R3) transforms the AD606's output of 37.5 mV/dB to 1 mV/dB and fits the result within the +1 to +4.5 volt common-mode range of the ICL7136; remember, the circuit is DC coupled!

Resistors R1, R2, and R3 do both the level shifting and the gain scaling.* The output of this network appears at the ICL7136 at INHI. The voltage at INLO, from the wiper of R9, provides a fixed offset to make the DMM IC's display read –75 to +5 mV for a –75 to +5 dBm input to the AD606.

Note that the AD606 has an on-board low-pass filter with a cutoff of 2 MHz. The input circuit of the ICL7136—C4 in parallel with R1-R3—forms an additional 1-pole low-pass filter with a 6-Hz corner frequency that removes any residual ripple on the AD606's RSSI output. This ripple is at twice the frequency of the AD606's input because the detector cells are full-wave rectifiers.

The resistor string of R4 through R7 sets the ICL7136's reference voltage. Adjust R6 for 100 mV between pin 35 (REF LO) and pin 32 (COMMON). This voltage is proportional to the power supply, as is the AD606's output. A simple test—grabbing the voltage-control knob on the bench power supply and twisting it—will show that the readout remains stable over ±5 percent power supply changes! The display is shown on the ICL7136 data sheet and isn't critical; the details are omitted to save space.

Calibration

Now that you know how the circuit works, it's time to calibrate the power meter. First, short the input and adjust R9 (the intercept adjustment) for a reading of 80 mV between INLO and INHI. This step sets the "logarithmic intercept," which is where the AD606's VLOG output "bottoms out" to its internal noise level of about –88 dBm. To set the slope, you can

---

$$* \quad \frac{R2 \| R3}{R1 + R2 \| R3} \times VLOG = \frac{1}{38} \times 37.5\text{mV/dB} = 1\text{mV/dB and } \frac{R3}{R1 \| (R2 + R3)} \times VPOS = 2.43\text{V}$$

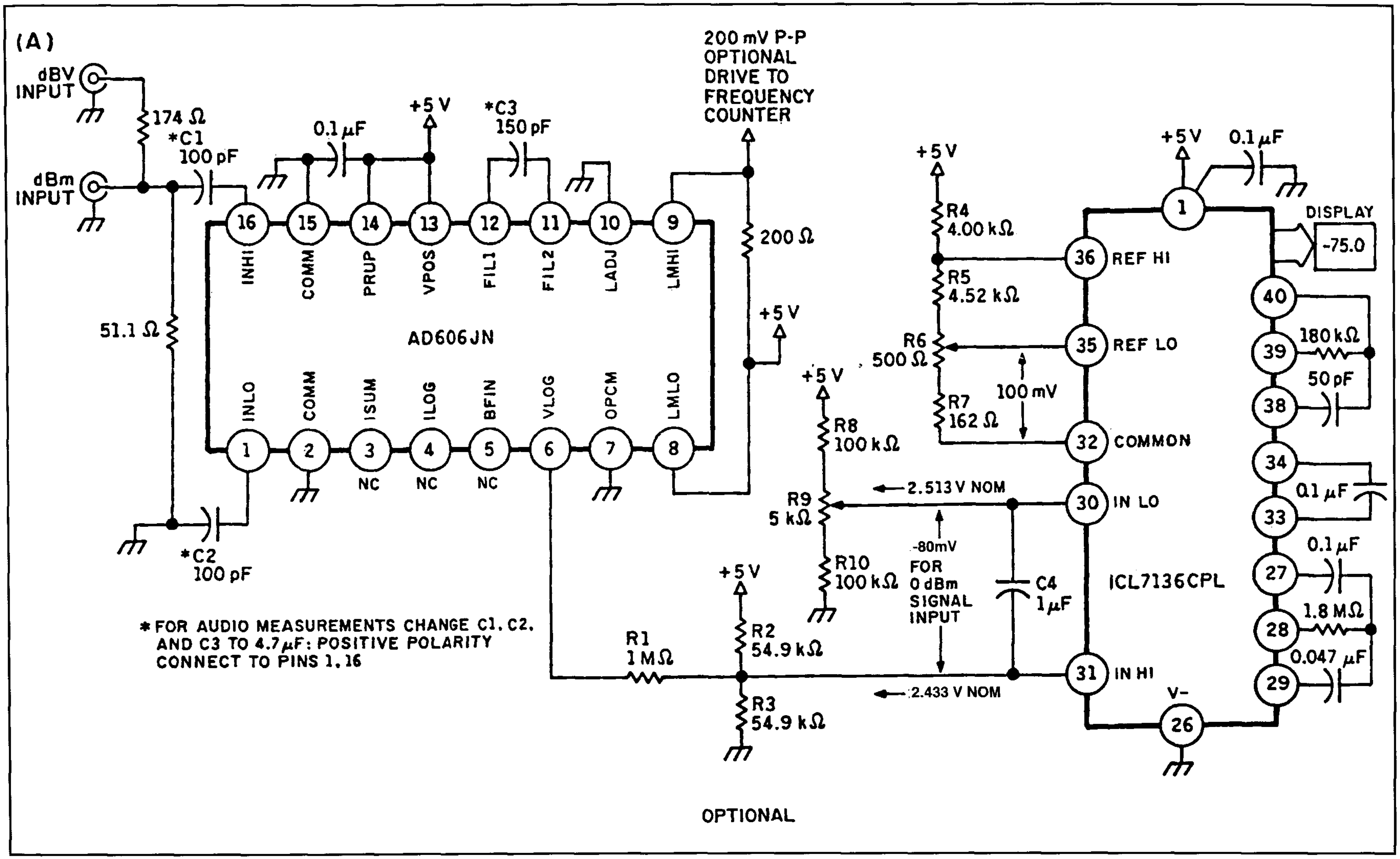
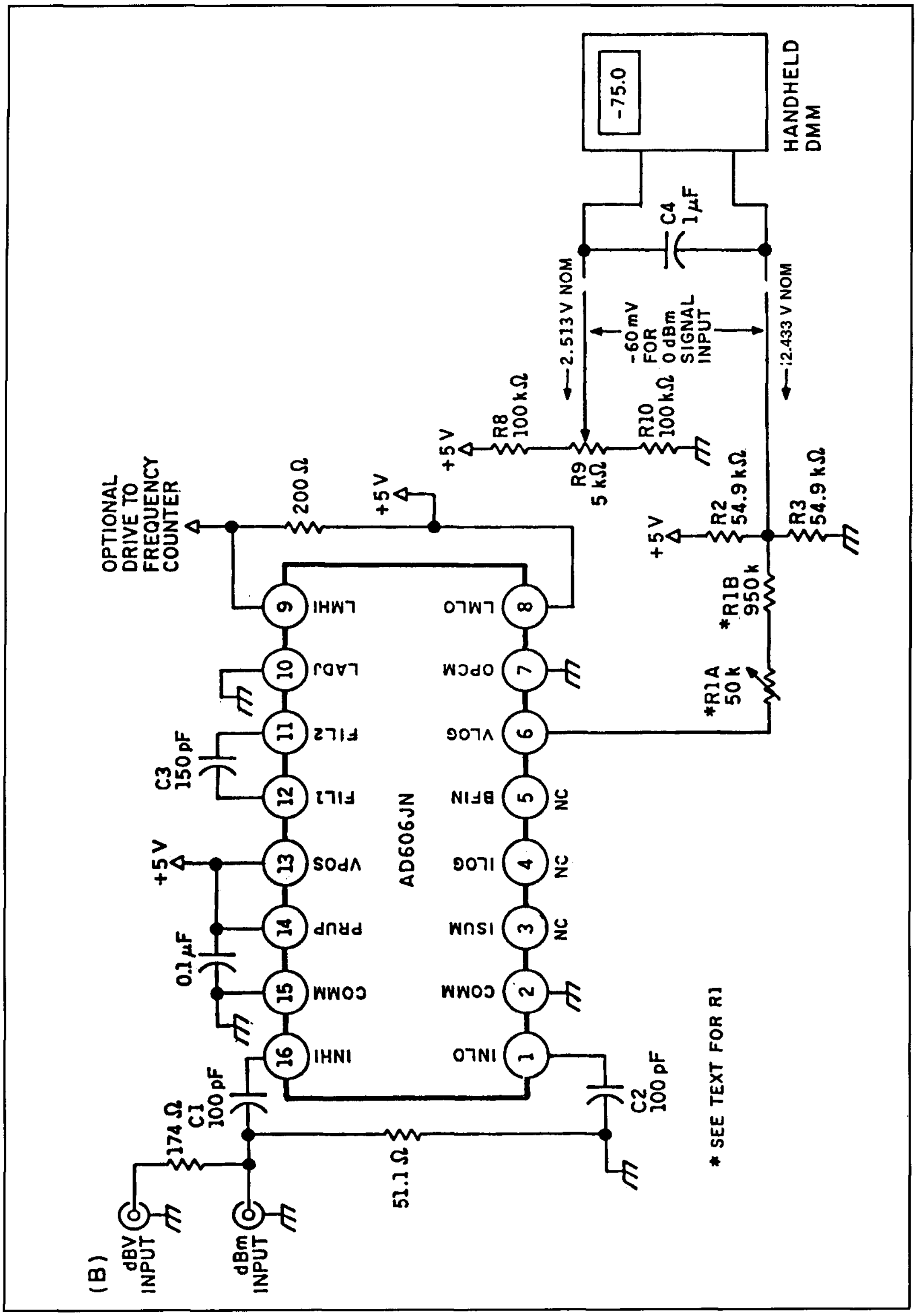


Figure 1. (A) A low-cost RF power meter. (B) Modifications for the handheld DMM.



use a calibrated signal generator and an attenuator or an uncalibrated signal generator and one or two accurate attenuators.

To adjust the power meter with a calibrated signal generator and an attenuator, attach the attenuator to the signal generator, set the output (including attenuation) to –80 dBm, and adjust R9 for a reading of –80 dBm. Next, remove the attenuator and set the generator for +5 dBm. Adjust R6 for a reading of +5 dBm. Now check the entire 80 dB range. You should be pleasantly surprised to find the power meter accurate to within 1 dB for the entire range.

If you don't have a calibrated signal genera-

Log Amp Circuit Topology

The AD606 log amp uses a technique called "successive compression" to generate its logarithmic transfer function. In a successive compression log amp, the log amp uses cascaded limiting amplifiers to approximate the logarithmic transfer function (**Figure A**). Each amplifier has the same fixed gain, which we'll call A. Each amplifier stage also has a detector that rectifies its output to produce a DC current. The output current of all the stages are summed to generate a single current.

For small input signals, each amplifier operates in its linear region and the overall gain of the log amp is A^N, where N is the number of amplifiers. As each stage limits, its output no longer increases and the gain drops from A^N to A^{N-1} to A^{N-2}, and finally to A. Consequently, the change in gain caused by successive compression provides a piecewise-linear approximation to the logarithmic function (**Figure B**). **Figure B** shows the transfer function of a successive-approximation log amp.

In practical implementations, the detector is a half-wave or full-wave rectifier. A full-wave rectifier has its ripple at twice the input frequency; as a result, higher ripple frequency is easier to filter out. The rise time of the log output depends on the corner frequency and number of poles in the ripple filter.

As a practical note, the cutoff frequency of the log amp's internal low-pass filter depends on the IF used for the cellular standard. In AMPS or ETACS, where the modulation mode is narrowband FM, the IF is typically 450 kHz, and the filter's voltage output response provides a "slow" RSSI. In GSM, DECT, and PHP applications, the IF is typically 10.7 MHz or higher, and the filter's voltage output response provides a "fast" RSSI. In the AD606, which was designed with IFs from 10.7 (21.4 MHz ripple) to 45 MHz in mind (90 MHz ripple), the low-pass filter has a 2 MHz cutoff and a "fast" RSSI response. The AD606 consists of seven cascaded amplifier stages, each with a gain of 10 dB (x3.162), and a side chain consisting of two cascaded amplifier stages, each with a gain of 10 dB (x3.162).

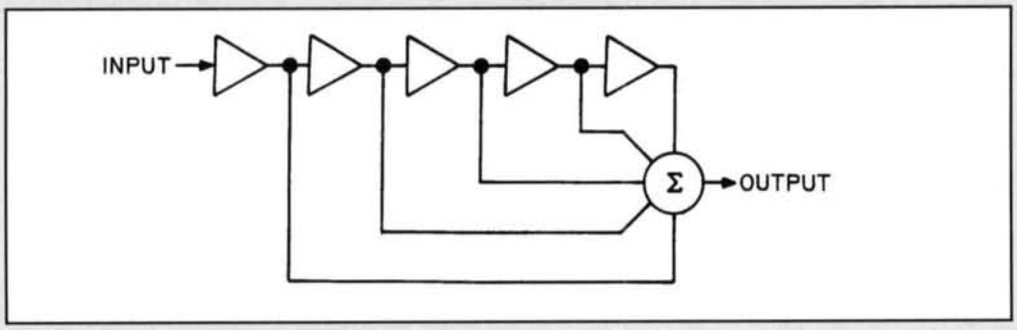


Figure A. Basic multi-stage log amp architecture.

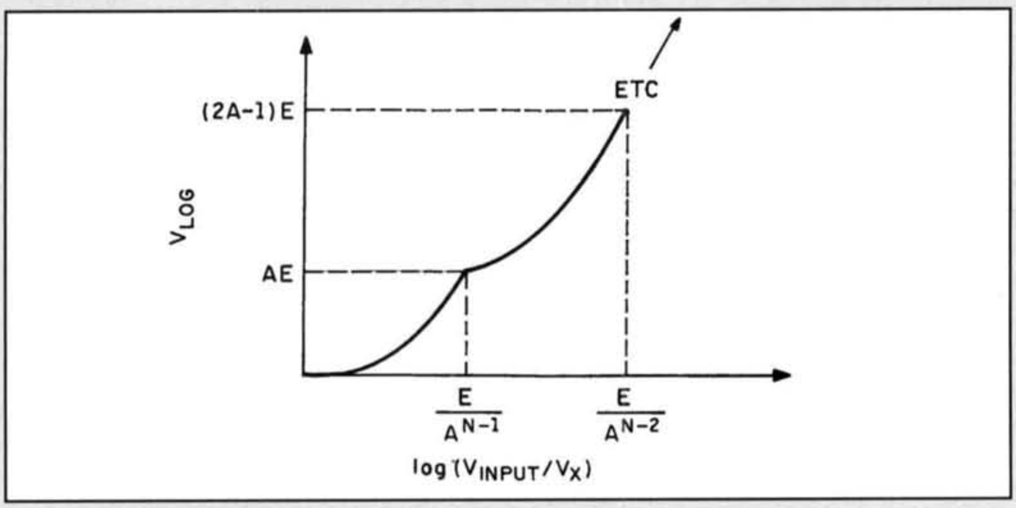


Figure B. The location of the "Lin-Log" and "Mid-Log" transition points.

Input waveform	Peak or rms	Intercept factor	Error (Relative to a DC input)
Square Wave	Either	1	0.00 dB
Sine Wave	Peak	2	–6.02 dB
Sine Wave	rms	1.414 ($\sqrt{2}$)	–3.01 dB
Triwave	Peak	2.718(e)	–8.68 dB
Triwave	rms	1.569($e/\sqrt{3}$)	–3.91 dB
Gaussian Noise	rms	1.887	–5.52 dB

Table 1. The effect of waveform on log intercept.

tor, the adjustment's a bit trickier. Connect a signal generator and take a reading. Next, insert one or two accurate attenuators.** Now make a second measurement and adjust R6 so the reading is the correct number of dB below the first reading. For example, if the attenuator is 20 dB, adjust R6 so the second reading is 20 dB below the first. Now remove the attenuator and adjust the slope once again to obtain the correct number of dB between the two readings. This adjustment is easier if you can use two calibration points as far apart as possible to minimize errors in the slope.

To extend the low-frequency limit of the power to audio frequencies, simply change C1, C2, and C3 to 4.7 µF. This also lets you calibrate the power meter using an audio source and a true RMS DMM as a reference.

You can also substitute a handheld DMM for ICL7136. Connect the positive lead to junction of R1, R2, and R3 and the negative lead to the wiper of R9. Resistor R9 still provides the offset adjustment, but you need to split R1 into the fixed resistance and a variable resistance, R1A = 950 k (750 and 200 k in series) and R1B = 100 k, to provide a slope adjustment. The DMM and AD606 now use separate voltage references, so the AD606 should be powered by a well-regulated 5-volt supply rather than a battery.

**If necessary, you can easily build an accurate 20-dB attenuator that works at 10 MHz using 1 percent metal film resistors. An R-2R ladder network, for example, provides 6.02 dB attenuation per tap and can be built using only one resistor value in series and parallel combinations!

Accuracy

This power meter actually measures the log of the input voltage because it uses a log amp as the detector. The log amp's output increases linearly with an exponential increase in input power. (Remember, $P = V^2/R$ and $\log V^2 = 2 \log V$.) As long as the input signal is a sinewave, the power meter's reading is accurate. This isn't a problem in most RF or audio applications because the signal being measured is usually a sinewave. **Table 1** shows the effect of waveform on logarithmic intercept for non-sinusoidal signals.

Measuring frequency

The limiter output of the AD606 may be used to drive the high-impedance input of a frequency counter. In the absence of an input signal, the 100 dB or so gain in the limiter tries to amplify the AD606's input noise. Thus, the output swings 200 mV p-p on noise until an input signal, even one as low as –80 dBm, is connected. ■

PRODUCT INFORMATION

Hy-Gain Introduces DX77 Advanced Vertical Windom with "No Ground Radials" Required

Telex Communications, Inc. has introduced its new Hy-Gain DX77 Advanced Vertical Windom. The design of this vertical antenna provides no-compromise performance without the need for ground radial wires. It puts the world at your fingertips on HF ham bands, 10 through 40 meters, including the WARC bands. The DX77 AVW can handle 1,000 watts of RF output. Automatic band switching and low angle of radiation allows for enhanced DX capabilities. The DX77 is designed with double wall tubing, steel mast clamps and all stainless steel hardware. The 29 foot vertical also features a new easy-tile mount that makes lowering it for tuning easier, and facilitates mounting on a pole, chimney, rooftop or deck. It can also be used for portable operation. The DX77 comes with Telex's 2 year limited antenna warranty.

For further information, write to Telex Communications, Inc., 8601 E. Cornhusker Highway, P.O. Box 5579, Lincoln, NE 68505; or phone: 402-467-5321; fax: 402-467-3279.

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QUARTERLY DEVICES

Radio Adventures Corp's CMOS CW keyer chips

This column usually features high-tech or consumer-electronic devices that have a secondary application for amateur radio projects. However, this time, I'll present a family of devices especially designed for the amateur experimenter and builder. The product is a family of CMOS CW-keyer chips, and the company is Radio Adventures Corp. of Seneca, Pennsylvania.

Radio Adventures Corp. was founded by Lee Richey, WA3FIY. After a successful career designing and manufacturing industrial control systems, Lee sold his control business and settled into an early retirement. However, it wasn't long before his amateur radio hobby grew into a new enterprise. RAC's first product was a simple direct-conversion W1AW code-practice receiver kit, but Richey claims the company's *real* focus is on developing special-purpose preprogrammed microcomputers such as the RAC keyer chip.

All in the family

According to application literature, RAC Iambic Keyer/Controllers are based upon the Microchip PIC-series microcontroller. This particular platform supports a customized keyer design that provides a wide range of features, while also offering extremely low power consumption and competitive pricing. There are currently three devices in the C1 family—plus a newly added economy version called the C2. The C2 has fewer features and consumes more power, but also costs about 1/3 less.

The C1 family provides all the essential keyer functions we've come to expect from the well-established Curtis chips over the years—along with some interesting added features and control functions. **Table 1** provides a function-by-function comparison between the C1, C1A, C1S, and C2 (note that C1S is identical to C1A, except the transmit interlock feature is omitted in favor of a dedicated pin for switching of the Autospace function). Here's a brief rundown of what the C1's primary features are—and what they mean to dedicated CW operators:

- Iambic A and B operation, selected by raising or lowering the state of the mode pin.
- Variable speed range (10–40 w.p.m.). Timing is crystal (or ceramic resonator) controlled for high accuracy over large changes in voltage or temperature.
- Adjustable spacing (to compensate for transmitter waveshape distortion), or standard one-dot spacing.
- Adjustable weighting (to compensate for transmitter characteristics or personal preference), or standard 3:1 weighting.
- Adjustable sidetone with sufficient internal drive for headphones or an efficient speaker. Pitch variable from approximately 500 to 1000 Hz.
- Built-in transmit sequencer with three outputs to provide full QSK and semi-break-in keying control (more details on this below).
- Bug Mode (single or double paddle) sends dots according to speed setting, while dashes are sent manually. Autospace mode can be set for automatic spacing after dots, or for automatic spacing after dots and dashes.
- Straight Key Mode allows the use of a straight key, mechanical bug, or tune switch. Autospace function may also be used in this mode. Mechanical key contacts are electronically de-bounced to provide clean high-speed QSK switching.
- CMOS technology allows operation over a wide voltage range (3 to 6 volts), wide temperature range (0 to 70 degrees C), and low

Features	C1	C1A	C1S	C2
Iambic A	X	X	X	X
Iambic B	X	X	X	X
Dot Memory	X	X	X	X
Dash Memory	X	X	X	X
Space Compensation	X	X	X	
Weight Compensation	X	X	X	
Bug Mode	X	X	X	
Speed Range	10-40	10-40	10-40	10-40
Sidetone (hz)	500-1000	500-1000	500-1000	500-1000
Straight Key / Bug Input	X	X	X	X
Debouncing	X	X	X	X
Autospace	X	X	X	X
Semi Breakin	X	X	X	
Transmit Sequence	X	X	X	X
Interlock	X	X		
Sleep (low power)	X	X	X	
Crystal (Mhz)	2.097	2.000	2.000	2.000
Xtal included with part		X	X	X
Voltage (vDC)	3-6	3-6	2.6-6	3-6
Operating Current	<2mA	<2mA	<2mA	<5mA
Idle Current	<1uA	<1uA	<1uA	<1mA
Package (Pin Dip)	18	18	18	18
Price	$14.95	$14.95	$14.95	$9.95

Table 1. Keyer matrix.

operating current (1.5 mA average, 5 μA in sleep mode).

- High output drive—more than 10 mA in output current available to drive switching transistors and sidetone monitor speaker.

Sequential T/R Control and Interlock

As outlined above, the C1 features a built-in transmit sequencer that allows you to custom-designed your radio's T/R switching system. Three different outputs are available for full and semi-QSK operation. For semi-automatic T/R switching, the chip's SEMI pin is timed so that output goes low when you key—and stays low until 3/4-second after the last element is sent. (The 630 mS "hold" time is typically long enough to prevent T/R switching between individual letters or between words strung together in a sentence.)

Also, whenever the key is closed, the ANT and TX logic pins go high in sequence to provide full-QSK switching outputs. The chip is internally timed so the ANT output changes state 5.5 mS *before* the TX pin. This sequencing ensures that the receiver front end is fully disconnected—and the load transferred to the transmitter—before the PA comes on. Sequencing may also be used to shift a VFO or BFO offset in advance, preventing chirp or a shortened character on the first CW element. When an element is completed, the TX pin goes low immediately. The controller then tests the INTERLOCK pin, which is driven externally. If the interlock is held low for any reason (parasitic RF output, or whatever parameter is being monitored), the antenna transfer will be prohibited. If the interlock input is high, antenna transfer will lag transmitter shutdown by 5.5 µS. Sequencing can be used in homebrew designs to provide for antenna transfer, VFO shifting, receiver muting, and transmitter keying in both semi and full-QSK modes. Note that the interlock input function is only available in the C1A version.

Using the C1S in a keyer circuit

C1-based keyers may be as simple or complex as you wish to make them. The schematic shown in **Figure 1** depicts a C1S "minimalist"

C1S Pinout Description

Pin #	Pin Name	Pin Type	Function
1	SPACE	ANALOG	Variable DC voltage to vary the space to weight ratio. Tie pin to ground to enter "bug" mode. Tie pin to VDD to provide standard 1:1 dit to space ratio.
2	WEIGHT	ANALOG	Variable DC voltage to vary the dit to weight ratio. Tie pin to VDD to provide standard 1:3 dit to dah ratio.
3	SEMI	OUT	Active low output during keying operation. Pin goes high approx 620 mS after last element is sent.
4	RESET	IN	Active low input to reset the C1.
5	VSS	POWER	Ground.
6	TONE	OUT	Sidetone output.
7	AUTOSPACE	IN	When high, AUTOSPACE inserts spaces between elements being sent by straight keys or bugs.
8	MODE	IN	When high and when using a paddle, C1 operates in Iambic A mode. Low for Iamic B mode. When high and using straight key or bug, spaces are inserted after key opening.
9	ANT	OUT	On key closure, ANT goes high 5.5 mS before TX is keyed. On key opening, ANT goes low 5.5 mS after TX goes low.
10	TX	OUT	TX goes high 5.5 mS after ANT. TX goes low after element is sent in keyer mode or 5.5 mS after key opens in straight key or bug mode.
11	SK	IN	Straight key input. Also used for semi-automatic (bug) key. Debounced. Active low.
12	DIT	IN	Dit input from paddle. Active low.
13	DAH	IN	Dah input from paddle. Active low.
14	VDD	POWER	Power, 2.6 to 6vDC. Less than 5 uA during sleep, less than 2.0 mA while operating.
15	XTAL2	OUT	Crystal connection.
16	XTAL1	IN	Crystal connection. May be used for external clock input.
17	SPEED	ANALOG	Variable voltage to set keyer speed. Speed varies inversly with input voltage.
18	PITCH	ANALOG	Variable voltage to set sidetone pitch. Pitch varies inversly with input voltage. Tie PITCH to VDD to provide default 750 Hz pitch.

Table 2. C1S pinout description.

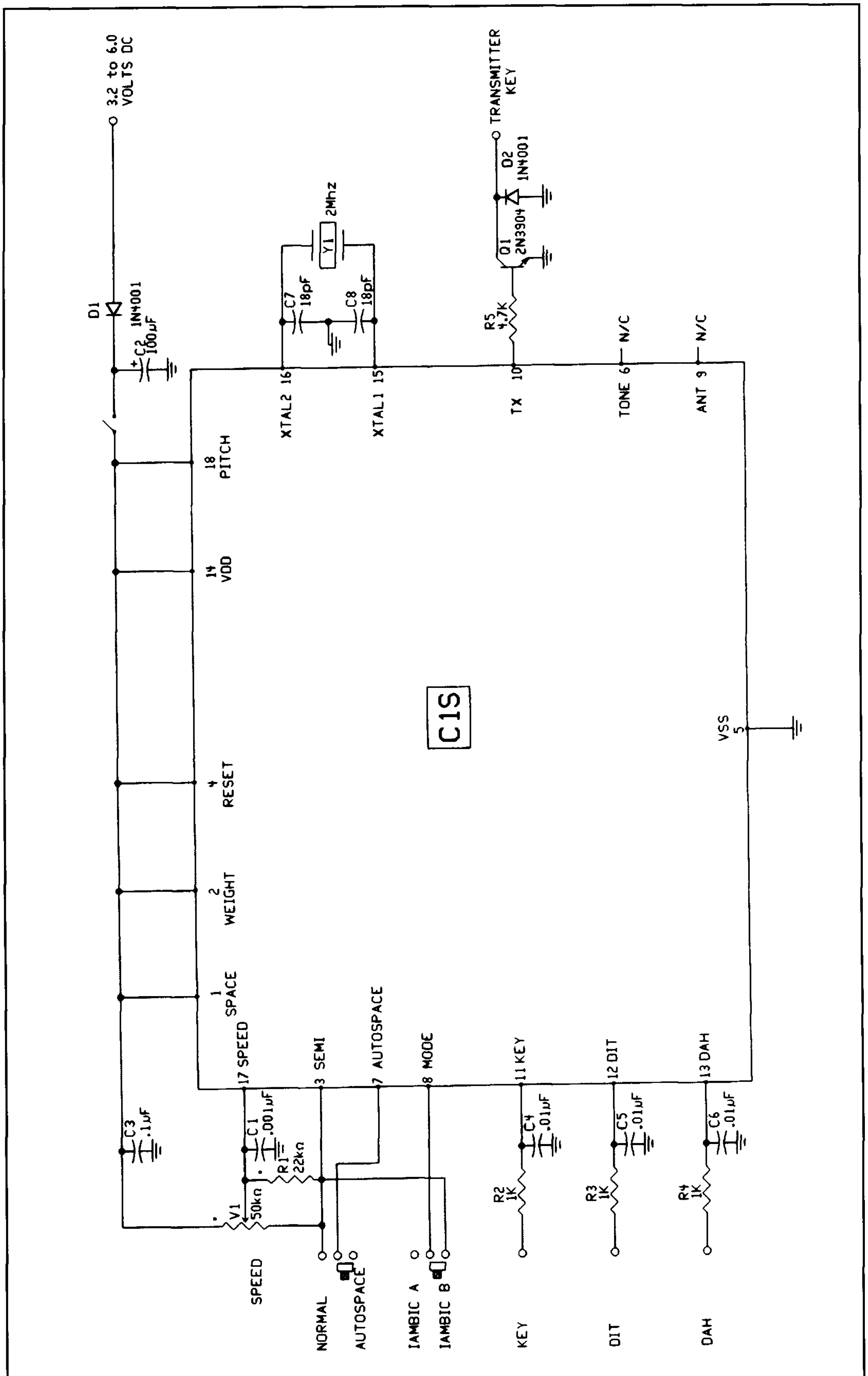

Figure 1. Low component-count keyer application.

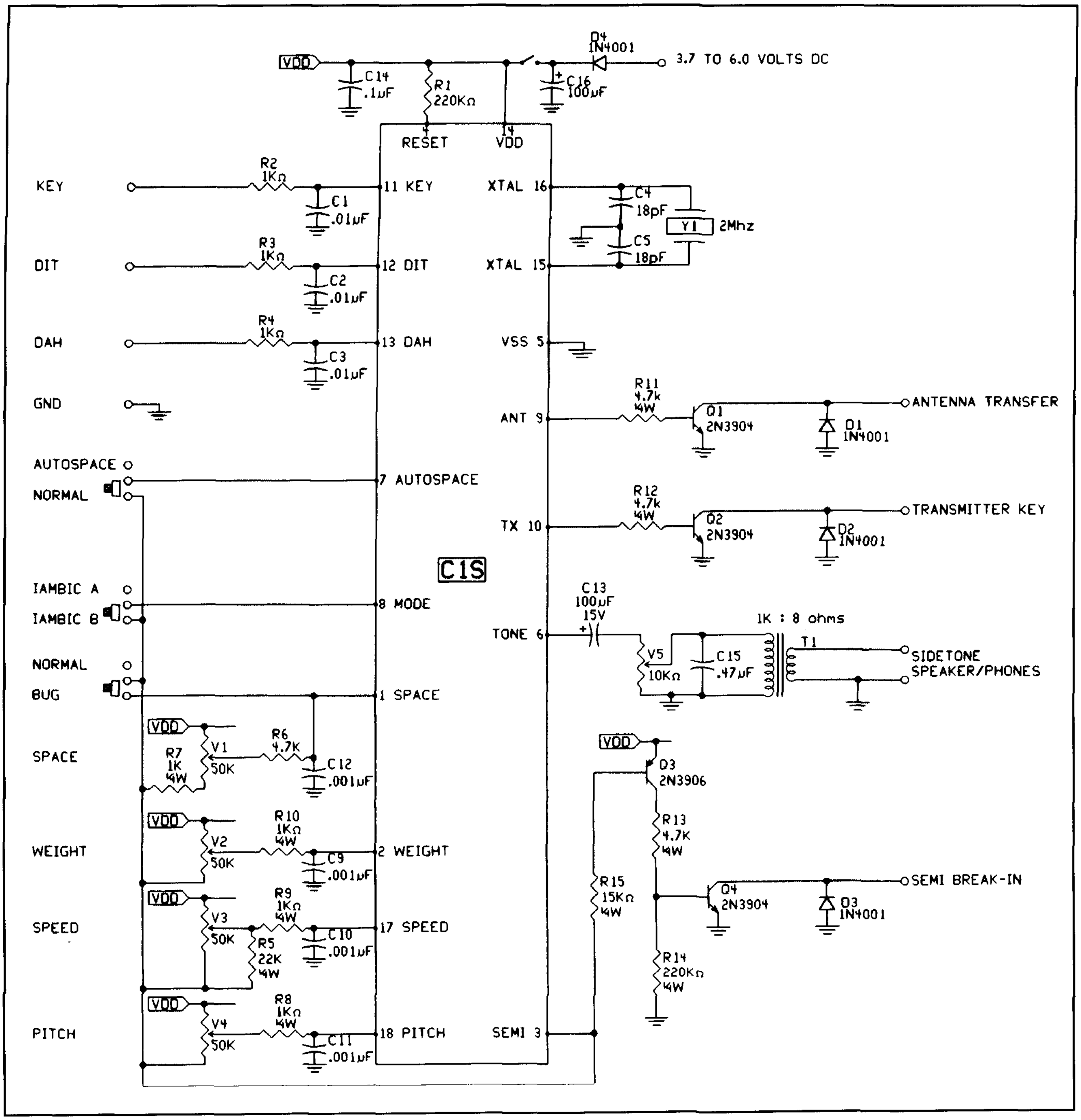

Figure 2. C1S application diagram.

circuit that uses your transceiver's built-in T/R switching and sidetone functions. Note that speed is adjustable from 10 to 40 w.p.m., but default settings are used for parameters such as weighting and spacing. RAC markets a C1S-based keyer similar to this called the "CodeBoy." The CodeBoy is available in kit or assembled form (see **Photo A**). The unit measures a scant 1-3/16 x 3-1/8 x 2-7/8 inches, weighs about 3-1/2 ounces, and draws power from an internal Lithium battery. Unlike the diagram shown in **Figure 1**, the CodeBoy uses FET output switching for transmitter keying.

Figure 2 shows a more complex circuit that provides access to the full range of C1S features. A minimal number of external parts are required to access each of these functions. Basically, you can customize your own design using only the specific functions you need, and either omitting or setting defaults for the rest. A pinout description for the C1S is shown in **Table 2**. The documentation package that

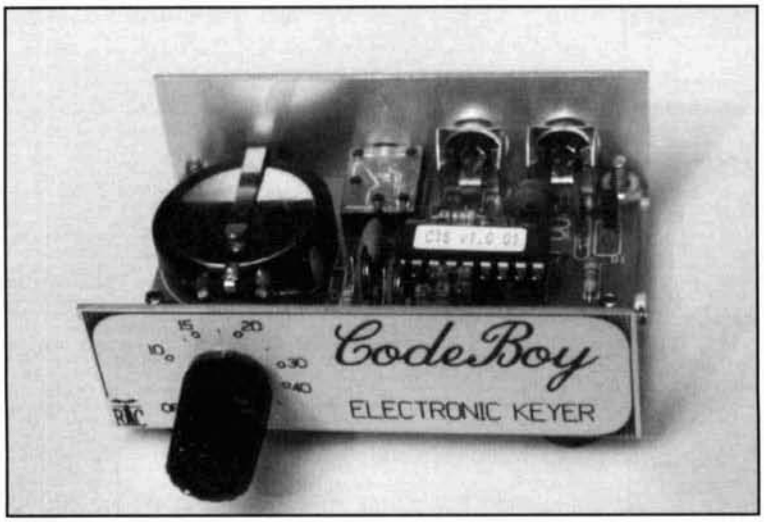

Photo A. Inside the CodeBoy keyer.

accompanies these devices is detailed, and you should have little difficulty understanding I/O interface requirements. When working with the C1 and C2 chips, remember that they are CMOS devices and subject to damage from static discharge.

How to order

Because RAC caters almost exclusively to the amateur experimenter market, they cannot offer design samples to home builders. To order RAC products, you may call them at (814) 677-7221, or write them at Radio Adventures Corporation, P.O. Box 339, Seneca, Pennsylvania 16346. RAC's FAX number is (814) 677-6456, and they may be reached on-line via the Internet at rac@usa.net.

The C1A, C1S, and C2 come complete with a 2-MHz ceramic resonator and a full documentation package. Single lot price is $14.95, $12.95 each for lots of 10 to 49, and $8.49 each for lots of 50 and up. The C2 sells for $9.95 in single lots, $8.95 for lots of 10 to 49, and $5.95 for lots of 50 and above. CodeBoy keyers are available for $37.95 in kit form and $54.95 assembled. RAC is also in the process of introducing the C5, a low-cost 50-MHz programmable frequency-counter chip suitable for transceiver frequency readout applications. Preliminary data sheets are currently available, and pricing will be identical to C1-series chips. A catalog of all RAC products is available upon request. RAC accepts VISA and MasterCard. Please add $3.75 S&H for single items, and add 5 percent for credit card orders under $25.

Conclusion

RAC microcontrollers appear to offer some very interesting possibilities for designers and builders of amateur equipment. The RAC philosophy is to provide special-function microcontroller chips for communication applications with programming already in place. This, in turn, enables designers to use the device off-the-shelf, without having to configure EPROMS and load data. Richey feels that advanced technology has brought us to a point where designing high-performance home-built equipment has never been easier—and I couldn't agree more! As usual, we invite readers to submit projects using "Quarterly Devices" for publication. If you have a unique application for the C1 or C2 CMOS keyer, please let us know! ■

PRODUCT INFORMATION

New Function Generators From B+K

B+K Precision now has new function generators that offer much higher levels of overall performance than previous models at their respective prices. A proprietary integrated circuit reduces the number of parts and doubles the bandwidth for improved reliability. The new line consists of four models, with bandwidths from 2 MHz to 20 MHz. Model 4010 has 0.2 Hz to 2 MHz bandwidth and all the basic functions, including sine, square, triangle waves, TTL and CMOS outputs, variable waveform symmetry, and variable DC offset. Model 4011 is a step up with 0.2 Hz to 5 MHz bandwidth and a built-in 4 digit LED counter. Model 4017 has 0.1 Hz to 10 MHz bandwidth, a 5 digit counter, and lin/log sweep. Model 4040 has 0.2 Hz to 20 MHz bandwidth, AM, FM, and burst operation in addition to all the features of the 4017.

Sine wave distortion is under 1%, with better than 0.35dB flatness across the specified bandwidth. Square wave rise time is under 50nS. CMOS output level is variable from 4 to 14.5 volts peak to peak. Duty cycle is variable (15:85:15). All models are compact, lightweight, and stackable, with tilt stands for convenient use. Bright red LED counters are used, with coarse and fine frequency adjustments for accurate setability. The function generators operate off 120 or 230VAC. All are backed by B+K's 1-year warranty.

For more information or the name of your nearest B+K distributor, contact B+K Precision, 6470 West Cortland St., Chicago, IL 60635; phone 312-889-1448.

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MAGNETIC UNITS AND FORMULAS

A Handy Guide for People Working in Magnetics

Back in 1980, while employed at SRI International, I worked on a project involving large air-wound coils and had to become familiar with a lot of magnetic units. For my own education I made up a list of these strange-looking terms, their definitions and the relations between them, and the derivation of some of the practical design equations. I have many times since found it handy to have all this stuff together, and I think other people working in magnetics will also find it useful.

The test is essentially as I wrote it in 1980, except for the addition of the definition of gamma—a small unit of magnetizing intensity convenient for magnetic prospecting and similar uses.

Basic quantities

μ =	permeability	
	webers per ampere-meter	(MKS system) (meter/kilogram/second)
	oersteds per gauss	(cgs system) (centimeter/gram/second)
μ_o =	permeability of vacuum	= 1 (cgs system)
		= 10^{-7} (MKS system)
μ_r =	μ/μ_o = relative permeability (dimensionless)	
	(Note: because μ_o = 1 in the cgs system, in that system μ = μ_r. But μ is not dimensionless.)	
B =	flux density, or magnetic field, or magnetic induction	
	gauss	
	maxwells per cm^2 or per $inch^2$	
	lines per cm^2 or per $inch^2$	
	tesla	
	webers per m^2 or per $inch^2$	
Φ =	flux	
	maxwell	
	line	
	weber	
H =	magnetizing intensity	
	oersted	
	gamma	
	ampere-turns per unit length	
	magnetomotive force (mmf) per unit length	

Definitions of units

		Symbol
Ampere	Unit of current	I
Ampere-turn	Unit of magnetizing force (cgs system)	NI
Gamma	Unit of magnetizing intensity [1]* = 0.00001 oersted or 10 micro-oersted	γ
Gauss	= 1 maxwell or line per cm^2 (cgs system)	B
Gilbert	Unit of magnetizing force or magnetomotive force = 0.4π ampere-turn = 1.257 ampere-turn	F
Henry	Unit of inductance, flux linkages per ampere; volts induced per time rate of change of current, amperes per second	L
Line	Unit of magnetic flux (cgs system) (same as maxwell)	Φ
Line/inch^2	Unit of flux density = 6.45 gauss	B
Maxwell	Unit of magnetic flux (cgs system) (same as line)	Φ
Oersted	Unit of magnetizing intensity = 1 gilbert per cm of path length = 0.4π ampere-turn per cm	H
Tesla	Unit of flux density (magnetic field) = 10^4 gauss = webers per m^2 (MKS system)	B
Weber	Unit of magnetic flux (MKS system) = 10^8 maxwells or lines	Φ

Note: The Earth's magnetic field is of the order of 0.6 oersted.

Derivation of the transformer equation

Faraday's Law for a single-turn circuit:

$$\epsilon = -\frac{d\Phi}{dt} \qquad (1)$$

ϵ in volts
Φ in webers

That is, the induced emf in a circuit is equal to the rate of flux change through it.
For N turns:

$$\epsilon = -N\frac{d\Phi}{dt} \qquad \textbf{(1a)}$$

In the cgs system:

$$\epsilon = -N\,(10^{-8})\,\frac{d\Phi}{dt} \qquad \textbf{(1b)}$$

Φ in maxwells or lines
(Lee 2)[2]

For sinusoidal flux variation:

$$\Phi = \Phi_{max} \sin \omega t \qquad \textbf{(2)}$$

$$\frac{d\Phi}{dt} = \Phi_{max}\,\omega \cos \omega t \qquad \textbf{(3)}$$

Substituting (3) in (1b):

$$E_{max} = -N\Phi_{max}\,\omega\cos wt(10^{-8}) \qquad \textbf{(4)}$$

* References are on the last page

However, the effective value of ϵ_{max} is $E = .707\ \epsilon_{max}$ and $\omega = 2\pi f$. Then:

$$E = (.707)2\pi fN\Phi_{max}(10^{-8}) \qquad \textbf{(5)}$$

$$E = 4.44\ fN\Phi_{max}(10^{-8}) \qquad \textbf{(5a)}\ \text{(Lee4)[2]}$$

Then:

$$10^8E = 4.44\ NBAf \qquad \textbf{(6)}$$

where E = volts RMS
N = turns
f = frequency, Hz

and B = gauss — or B = maxwells/inch^2 or lines/inch^2
A = cm^2 — A = inch^2

In the MKS system:

$$E = 4.44\ NBAf \qquad \textbf{(6a)}$$

where B = teslas (or webers/m^2)
A = meters^2

Solving (6) for N:

$$N = \frac{10^8E}{4.44BAf} \qquad \textbf{(7)}\text{(Lee 32)[2]}$$

This is the usual form of Faraday's equation used for the design of transformer windings.

In many cases, information on cores is published as flux density in gauss and core area in inches^2. The following form of equation (7) may then be used:

$$N = \frac{3.49(10^6)E}{BAf} \qquad \textbf{(7a)}\text{(Lee34)[2]}\ B \text{ in gauss, } A \text{ in inch}^2$$

Note that this formula has incompatible units.

Inductance

By definition*

$$\mu = \frac{B}{H} \qquad \textbf{(8)}$$

where μ = permeability
B = flux density
H = magnetizing intensity

---

* Equation 8 is analogous to Ohm's law, which is usually given as:

$$R = \frac{E}{I}$$

where R is in ohms, E is electromotive force (emf) and I is current in amperes.

Ohm's law can also be written as:

$$G = \frac{I}{E}$$

where G = 1/R (conductance in mhos or siemens)

Then G corresponds to μ in the magnetic circuit, E (emf) corresponds to H (magneto motive force or mmf), and I (current flowing as a result of emf acting on G) corresponds to B (flux density set up by mmf acting on the magnetic circuit).

In vacuum or equivalent medium $\mu = \mu_o$ and in the cgs system:

$$\mu_o = \frac{B}{H} \tag{8a}$$

where B is in gauss and H is in oersted
But in vacuum $\mu = \mu_o = 1$ and:

$$B(\text{gauss}) = H(\text{oersted}) \tag{8b}$$

Recall that the oersted is defined as 0.4π ampere-turn/cm.

Substituting this in (8b):

$$B = 0.4\pi NI/l \tag{9}$$

where B is in gauss, NI is in ampere-turns, and l is the magnetic path equivalent length* in cm.

Inductance may be defined as flux linkages per ampere. Lee[2] gives (his Equation 37) in cgs units:

$$L = \frac{\Phi N}{10^8 I} = \frac{BAN}{10^8 I} \tag{10}$$

Substituting Equation 9 in Equation 10:

$$L = \frac{1.25N^2A}{10^8 l} \tag{11}$$

where L is in henries, N is turns, A is in cm^2, and l is the equivalent path length in cm.

or,

$$L = \frac{3.19N^2A}{10^8 l} \tag{11a}$$

where A is in $inches^2$, l is the equivalent path length in inches, and other units as above.

Inductance carrying direct current

In an inductor having a core of ferromagnetic material and carrying appreciable direct current, the core permeability is not quite constant, but depends on the level of DC magnetization in the core. The relative permeability μ_r for the DC component B_{dc} of flux density in the core may be found from the published B-H curve for the core material. For the AC component B_{ac}, the relative permeability is the incremental permeability μ_Δ defined by the slope of the B-H curve over the interval of

---

* The equivalent path length l is the path length in vacuum or equivalent medium that has the same effect as the actual path having a given physical length and core permeability. In a core having length l_1 and permeability μ_r.

$$l = \frac{l_1}{\mu_1}$$

In a path which is partly of permeability μ_1 and partly of permeability μ_2.*

$$l = \frac{l_1}{\mu_1} + \frac{l_2}{\mu_2}$$

If the path is a core of path length l_c and relative permeability μ_c, in series with an airgap of length l_g and relative permeability $\mu_o = 1$ (in the cgs system), then:

$$l = l_g + \frac{l_c}{\mu_c}$$

Because l is defined as a path length in vacuum, Equation 8 holds and Equation 9 is valid.

AC magnetization centered on the operating point established by the DC magnetization. The incremental permeability may differ from the DC permeability, sometimes quite substantially so.

For a core with a substantial airgap, such that the equivalent path length l in the core is due principally to airgap, an average of 0.85B crosses the gap, because of fringing. Therefore, the DC component Equation 9 becomes:

$$B_{dc} = 0.47\pi NI_{dc}/l_g \quad \textbf{(12)}$$

With l_g in inches:

$$B_{dc} = 0.6NI_{dc}/l_g \quad \textbf{(12a)}\ \text{(Lee 35)[2]}$$

where B_{dc} is in gauss, NI_{dc} is in ampere-turns, and l_g is the airgap in inches. This equation, like Equation 7a, has incompatible units.

For the AC component of flux, transpose Equation 7a to obtain:

$$B_{ac} = \frac{3.49(10^{-8})E}{NA_c f} \quad \textbf{(13)}\ \text{(Lee 36)[2]}$$

where B_{ac} is in gauss, E is rms volts across the winding, N is turns, A_c is in inches^2, and f is in Hz. However, the sum of B_{dc} and B_{ac} is B_{max}. Typically, B_{max} should not exceed:

12,000 gauss for 4% silicon steel at 60 Hz
16,000 gauss for grain-oriented steel at 60 Hz
10,000 gauss for 50% nickel alloy
8,500 gauss for 4% silicon steel at 400 Hz
12,000 gauss for grain-oriented steel at 400 Hz

For other materials and at other frequencies, the recommendations of the core manufacturer should be followed.

The inductance of a coil having a core with a substantial airgap can be found by substituting the appropriate expression for equivalent path length in Equation 11a:

$$L = \frac{3.19(10^{-8})N^2A_c}{l_g - (l_c/\mu_\Delta)} \quad \textbf{(14)}\ \text{(Lee 38)[2]}$$

where L is inductance in henries, N is turns, A_c is core area in inches^2, l_g is airgap in inches, l_c in core path length in inches and $\mu_\Delta$is the incremental permeability.

Choice of toroid core for power inductor

The choice of core for a power-handling inductor is determined by the energy storage needed. The energy stored in an inductor is:

$$W = 1/2LI^2 \quad \textbf{(15)}$$

where W is in joules or watt-seconds, L is in henries, and I is in amperes.

The energy storage capability of cores is usually given as the LI^2 product, which, as Equation 15 shows, is proportional to joules.

For toroid cores having no airgap, i.e., uncut, the path in the toroid has uniform permeability, and the equivalent path length l for a path of length l_c and relative permeability μ_r is simply l_c/μ_r. Then substituting in Equation 9:

$$B_{max} = \frac{0.4\pi\mu_r NI_{max}}{l_c} \quad \textbf{(16)}$$

where l_c is in centimeters.

Because toroids are available in a wide range of permeabilities, the question arises as to what permeability should be chosen to obtain the largest LI^2 product from a core of a given physical size. Solving Equation 16 for I_{max}, we obtain:

$$I_{max} = \frac{B_{max} l_c}{0.4\pi N \mu_r} \tag{17}$$

The inductance obtainable with a given toroid is often given as A_L, the inductance of a 1000-turn coil on the toroid. If we let N = 1000 in Equation 17, the LI^2 product will be:

$$LI^2 = A_L \left(\frac{B_{max} l_c}{1257 \mu_r} \right)^2 \tag{18}$$

Where L is in henries, I is in amperes, A_L is in henries, Bmax is in gauss, l_c is in cm, and μ_r is relative permeability. Note, however, that the published values of A_L are usually in millihenries.

If we compare two cores with the same dimensions and B_{max} but different μ_r, by examining the squared term in Equation 18 we see that decreasing μ_r will increase LI^2_{max} as the second power of μr. However, decreasing μ_r has the opposite effect on A_L; decreasing μr will decrease A_L as the first power of μ_r. The net effect is that, for a given B_{max}, LI^2_{max} will vary inversely as μ_r. For LI^2_{max} limited by core saturation, therefore, it is advantageous to select as low a value of μ_r as possible. (Note that this differs from the case of an AC transformer, where for a given core geometry and number of turns one wishes to minimize the magnetizing current in the primary, hence tends to choose a material having as high a value of μ_r as possible.)

The allowable value of LI^2_{max} for a given size core may also be limited by temperature rise. A power inductor will have core loss (eddy current loss and hystersis loss) due to the alternating component of current, and copper loss (I^2R loss where R is the resistance of the winding) due to both AC and DC components of current. For an inductor carrying substantial DC (the usual case), the copper loss will usually predominate.

For a given weight of copper, a core with lower μ_r will need more turns and therefore a smaller wire size, thus more I^2R loss. From this standpoint, it is advantageous to choose the value of μ_r as high as possible, which conflicts with the criterion for choosing μ_r as determined by core saturation.

Reference 3 provides a method of finding the LI^2_{max} capability of a core as limited by temperature rise. It also gives tables of LI^2_{max} for various cores as limited both by core saturation and by temperature rise. It turns out that a toroid having μ_r of 60 will be limited by core saturation, while a core with μ_r of 26 may have a slightly larger LI^2_{max} product; but the core with μ_r of 60 may require less copper and result in a more economical design. Cores with μ_r greater than 60 or less than 26 will have lower LI^2_{max} capabilities. On balance, then, a core with μ_r of 60 will usually be the best choice, but in some cases a μ_r of 26 may work out better.

REFERENCES

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3. *Switching Regulator Design Guide*, Publication No. U-68A, 1974, Unitrode Corporation, Watertown, Massachusetts.

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THE SOLAR SPECTRUM

Monitoring radio bursts from solar flares

In addition to its more well-known phenomena, the Sun is the source of several kinds of radio emissions, reflecting different aspects of solar activity. Certain features—solar flare waves and material ejection for example—should be detectable in the characteristic radio frequencies emitted by the Sun's atmosphere (corona), and indeed they are. Experienced radio amateurs may find the recording of such information quite interesting. Accordingly, following a brief outline of this aspect of solar activity, our collaborator, Peter D. King. G8KJP (Electronic Engineering Department, Cambridge University) describes his technique for measuring these effects.

Solar radio emission bursts are recorded at fixed and sweep frequencies by a worldwide network of observing stations. In the United States, these data are compiled by several organizations within the National Oceanic and Atmospheric Administration (NOAA), and are made available to a number of interested scientists and researchers. The bursts (**Figure 1**) are ranked on an intensity scale of 1 to 3 with 3 the highest. They are classified as follows:

- **Type I** are short bursts—about one second long—which usually occur in large numbers (i.e., a Type I noise storm). They occur in conjunction with large solar active regions, flares, and eruptive prominences, and are recorded at frequencies of 80 to 200 MHz.
- **Type II** are slow drifting, large-intensity bursts 5 to 30 minutes in length that are associated with powerful flares, proton emission, and magnetic shock waves. They are monitored at 20 to 150 MHz, and are indicative of

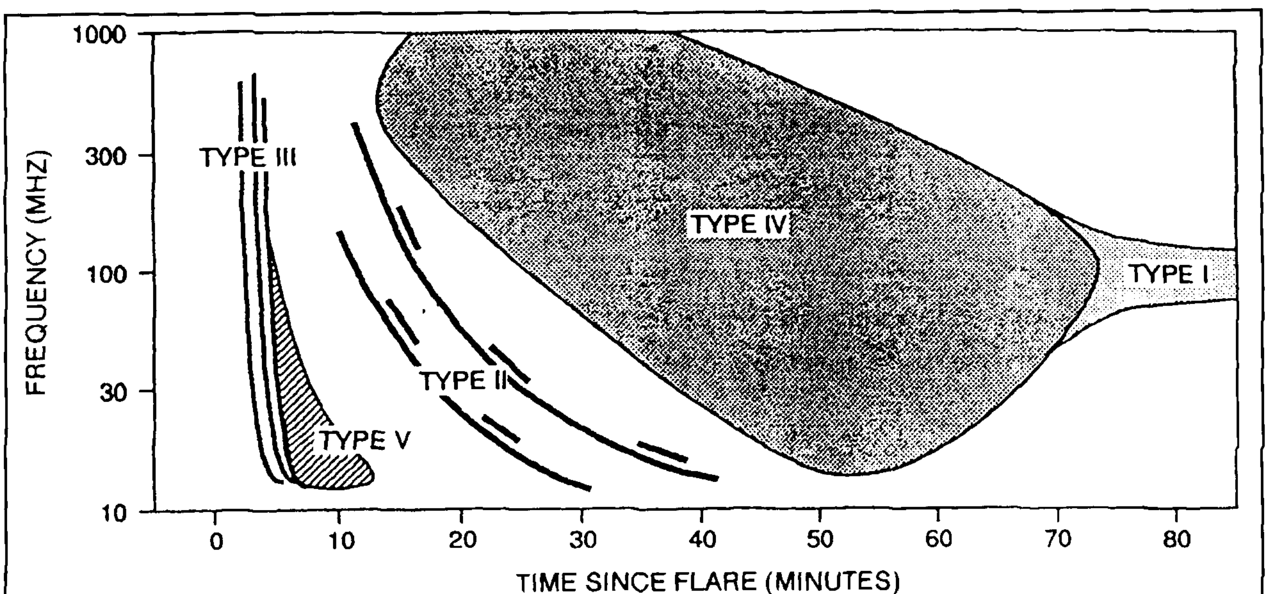

Figure 1. This diagram illustrates each major type of solar radio burst in a typical configuration following a large solar flare (see text). Note that not all of these features are observed after every flare. Source: *The New Culgoora Radiospectrograph Technical Report* IPS-TR-93-03.

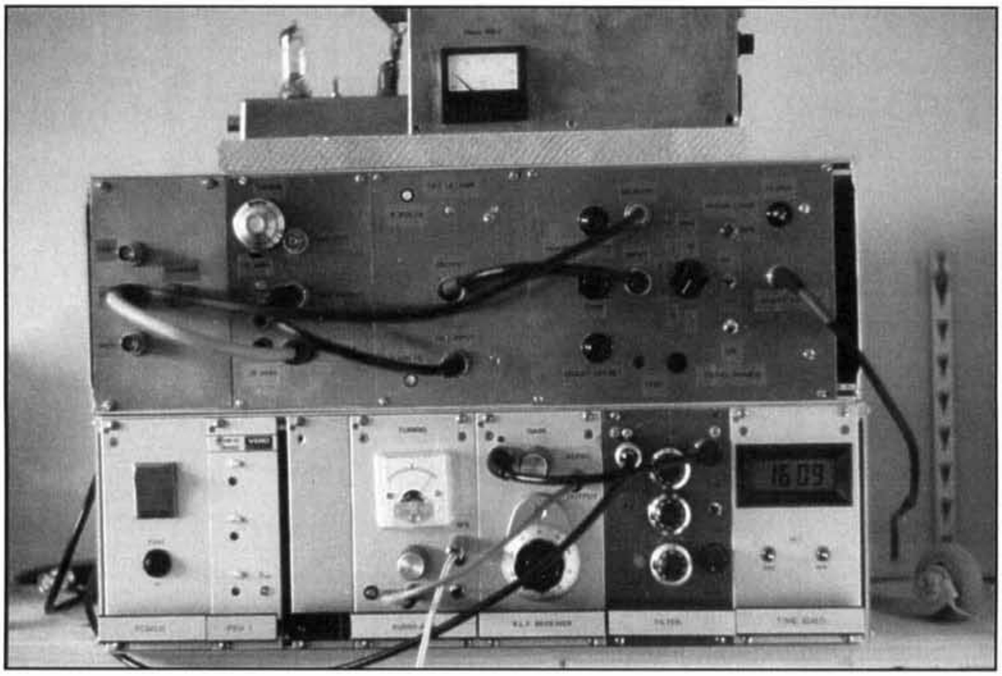

Photo A. The King VHF radio telescope. Note the vacuum tube noise generator to the top of the photo, which is used to calibrate the system.

a shock wave moving outward through the Sun's atmosphere at a speed of 1000 to 1500 kilometers per second.

- Type I and **Type III** bursts are the most common of these emissions. The latter are fast drifting bursts 1 to 3 seconds in length, which are recorded at between 10 kHz and 1 GHz. Type III bursts are rooted in flares and active regions, and often occur in groups. The frequency drift of such bursts corresponds to an electron stream moving at 100,000 km/sec.
- **Type IV** bursts are classified as 'stationary' (duration hours to days), or 'moving' (duration 30 minutes to 2 hours). The former are recorded at frequencies of 20 up to ~1000 MHz, and occur in conjunction with flares and proton emission. On the other hand, the rarer moving Type IV bursts (20 to 400 MHz) are associated with eruptive prominences or solar magnetic shock waves. Type IV bursts

Photo B. The two 110-MHz antennas that form the basis of the interferometer.

ELEMENTS	GAIN	BEAM WIDTH	ANTENNA CAPTURE
Five	6.5 dB	±28 degrees	Four square meters each

Table 1. 100-MHz antenna parameters (modified from broadcast 88 MHz).

begin about 10 to 20 minutes after certain major flares.

- **Type V** bursts follow some Type III bursts and last about 1 to 3 minutes. Their source is the same as Type III, although they are generally recorded at 10 to 200 MHz.

When monitoring these phenomena, Peter King's VHF radio telescope is configured as a 110-MHz switched-phase radio interferometer, offering the advantages of interference and increased sensitivity over other types of radio telescopes. A schematic diagram of Peter's instrument array appears in **Figure 2**, and **Photo A** shows the arrangement of the electronics components. (All photos courtesy of Peter King.)

The two antennas (**Photo B**, **Table 1**) are normally fixed and set at an angle of 25 degrees. However during the winter, when the Sun is low in the sky, the angle is decreased to 15 degrees. Consequently, the beamwidth of the antenna (±28 degrees) allows the Sun to be tracked with good results throughout the year. Both antennas point south, and are located on an imaginary line running from east to west; thus, the interferometer relies on the Earth's rotation for tracking the Sun.

The antennas are spaced three wavelengths

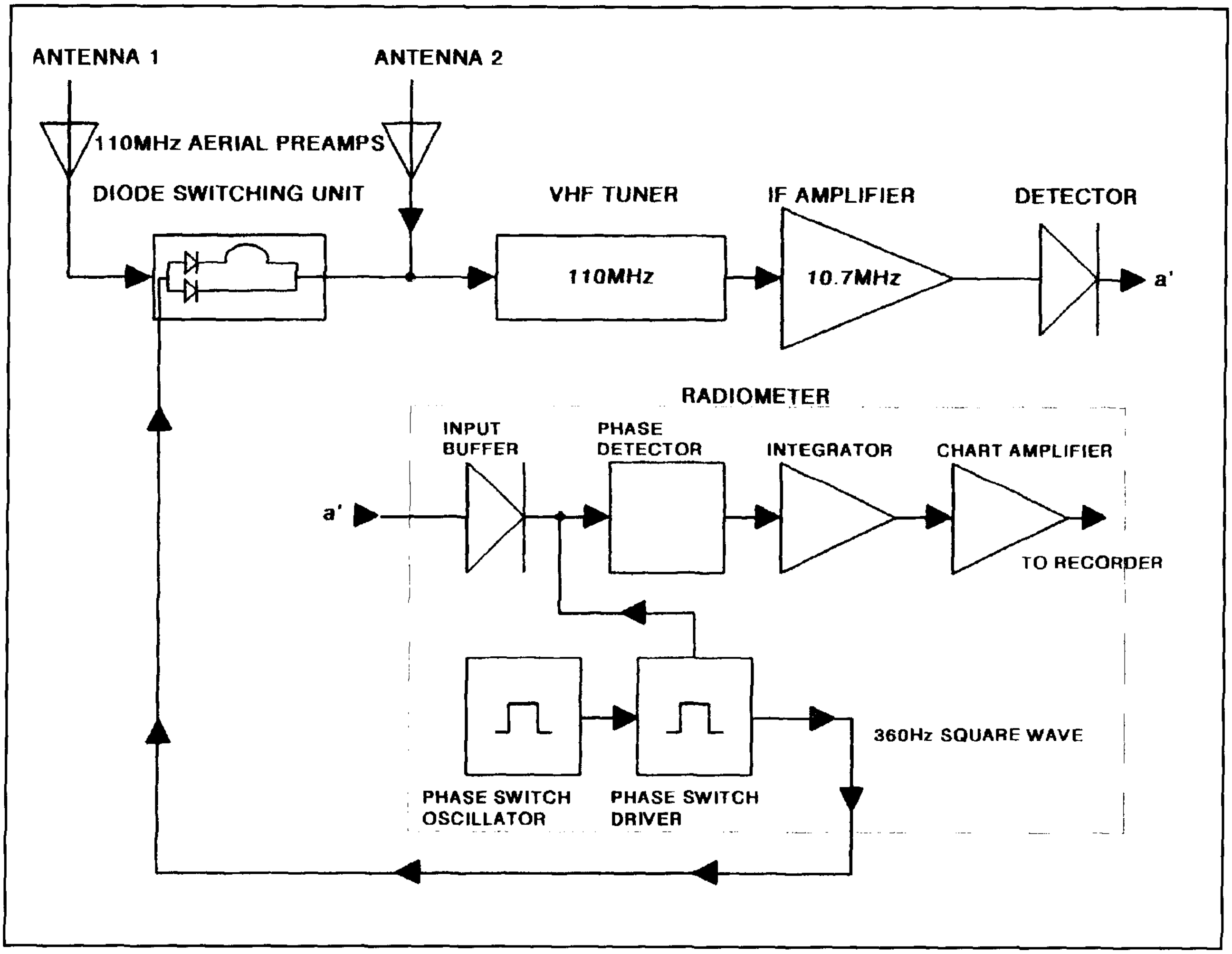

Figure 2. Block diagram of Peter King's radio interferometer, which he uses to monitor solar radio bursts at 110 MHz. Note that the output of the adjustable voltage regulator (Motorola LM317T) should be set by the 5-k resistor to give an output of 2 to 4 volts. The maximum voltage on the AGC line should not exceed 5 volts.

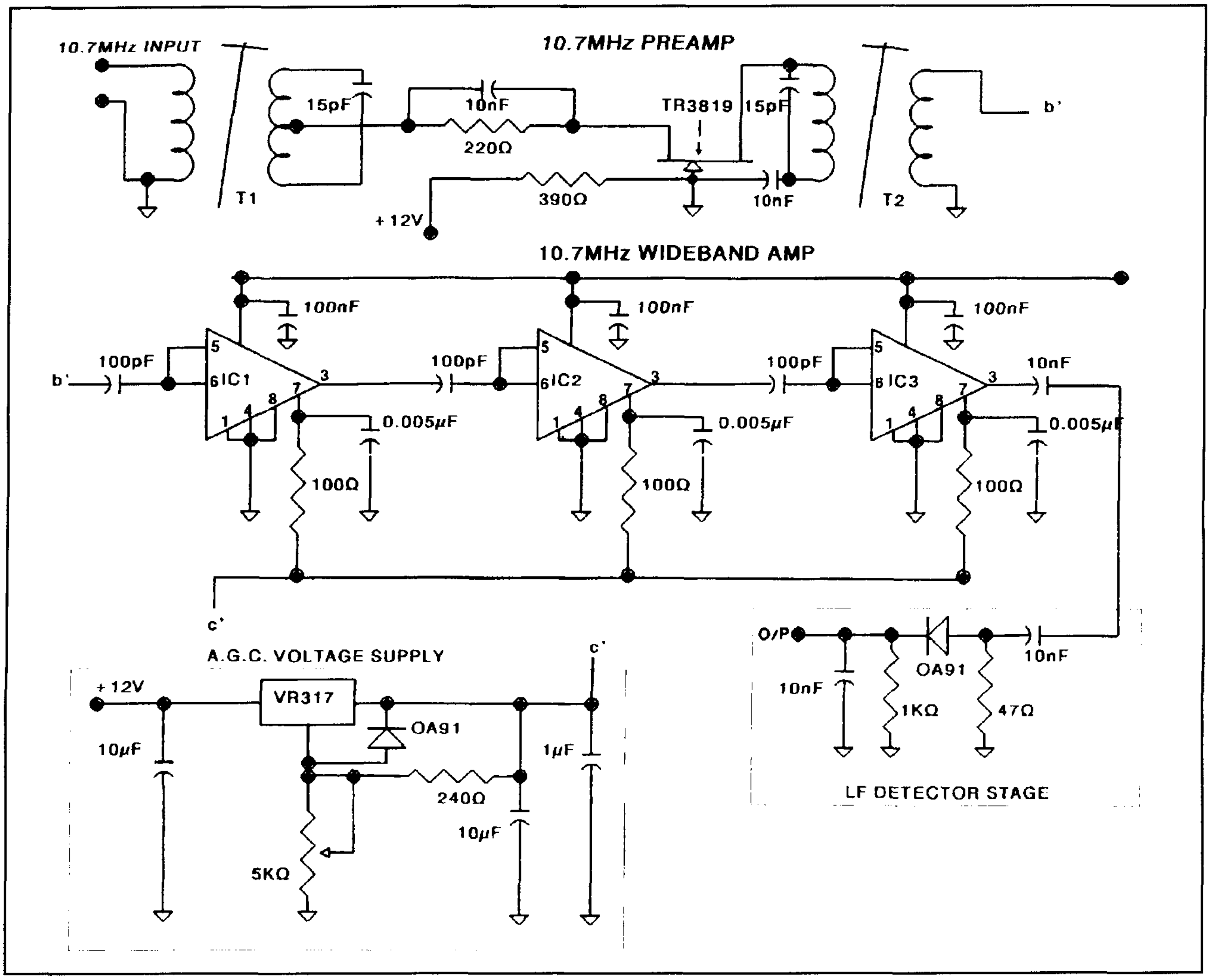

Figure 3. Schematic for the 10.7-MHz preamplifier, 10.7-MHz wideband amplifier, AGC voltage supply, and detector.

apart, giving an interference fringing of three lobes. According to Peter, a longer baseline might well result in losing flare effects in the lobes. (Twenty wavelengths produces twenty fringes. Therefore, a small flare could be lost in the fringes.) The time-constant in Peter's system is 2 to 3 seconds—ideal for recording Type II and III bursts.

In the schematic representation, the radiometer and antenna switching unit are adopted from circuits shown in the May through September 1978 issues of *Sky and Telescope* magazine. Peter tested numerous circuits throughout the years and believes that these are superior. The integration time is switched between 1 and 10 seconds. The interval could easily be increased to 30 seconds or longer, but the shorter time-constants appear to be better for solar observations.

The varicap tuner is a FM broadcast, front-end tuner-head consisting of an RF amplifier and mixer oscillation, resulting in a IF output of 10.7 MHz. This approach provides a fast way into mixer/oscillator designs—not always an easy task. An equivalent tuner should be easy to locate in the U.S.

The IF amplifier shown in **Figure 3** consists of a 10.7-MHz matching input transformer followed by a FET, then another matching transformer. These in turn, feed into a main IF amplifier consisting of three Plessey SL1612 IF amplifiers (available from Cirkit Distribution Ltd., Park Lane, Broxbourne, Herts, England EN10 7NQ).

The ICs are each regulated by a DC voltage via the 100-ohm resistor on Pin 7, which also acts as an AGC voltage. The bandwidth of the amplifier is 250 kHz, with a nominal gain of 90 decibels. The gain of the 10.7-MHz IF preamp is 23 dB, so in most cases the AGC voltage must be applied to the IF ICs; otherwise, the entire amplifier could become unstable. The

TUNER

Power Amplification	dB	typical
Noise Figure	dB	5.5 dB
IF Bandwidth	MHz	500 kHz
RF Bandwidth	MHz	1.7 MHz
Image Rejection	dB	80 dB
IF Rejection	dB	100 dB
Tracking 88-110 MHz	dB	1.5 dB

Source: Cirkit (EF 5402) 1979

COAX DOWN-LEADS

CABLE CHARACTERISTICS	ATTENUATION	75Ω IMPEDANCE
75Ω	1.3 dB for 10 meters	—
75Ω	0.7 dB for each lead	5 meters each lead

10.7 MHz IF PREAMPLIFIER **RF/IF AMPLIFIER IC R.S. 1612C**

PARAMETERS	CONDITIONS	MIN	TYPE	MAX	UNITS
Voltage Gain	Rs = 50Ω Rl = 600Ω	31	34	38	dB
Cut-off Frequency	(–3 dB)	—	15	—	MHz
Noise Figure	Rs = 600Ω	—	3.0	—	dB
Max Output Signal	Max AGC	—	1.0	—	V rms
Max Input Signal	Max AGC	—	250	—	mV rms
AGC Range	Pin 7, O.0 V to 5.1 V	60	70	—	dB

Source: Radio Spares Data Sheet, 1984 (England).

Table 2. Component characteristics.

detector can be any germanium diode. Component characteristics are outlined in **Table 2**, and **Photos C** through **E** show several of these devices.

Peter records these data using a Rickadenki four-pen chart recorder (**Photo F**) running at a speed of three centimeters per hour. He notes that this method may seem old fashioned in this day of digital record-keeping, but his preference remains the pen-type recorder and the clear, easily analyzed record it produces. Although the system is still under development, Peter says it has already proved to be an excellent performer.

Peter also records solar flare-induced sudden ionospheric disturbances for the *AAVSO Solar Division*, using Arthur Stokes' gyrator-tuned VLF receiver (described in the Spring 1994 issue of *Communications Quarterly*). This aspect of the King system uses an intriguing variable gain filter as a noise reducer along with several other innovative accessories. We'll go into more detail about that setup in a future

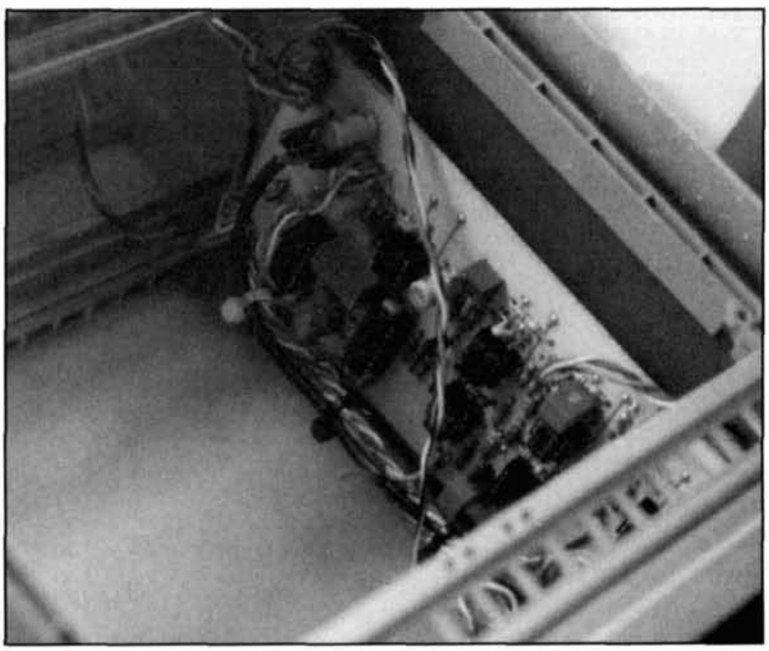

Photo C. The radiometer printed circuit board.

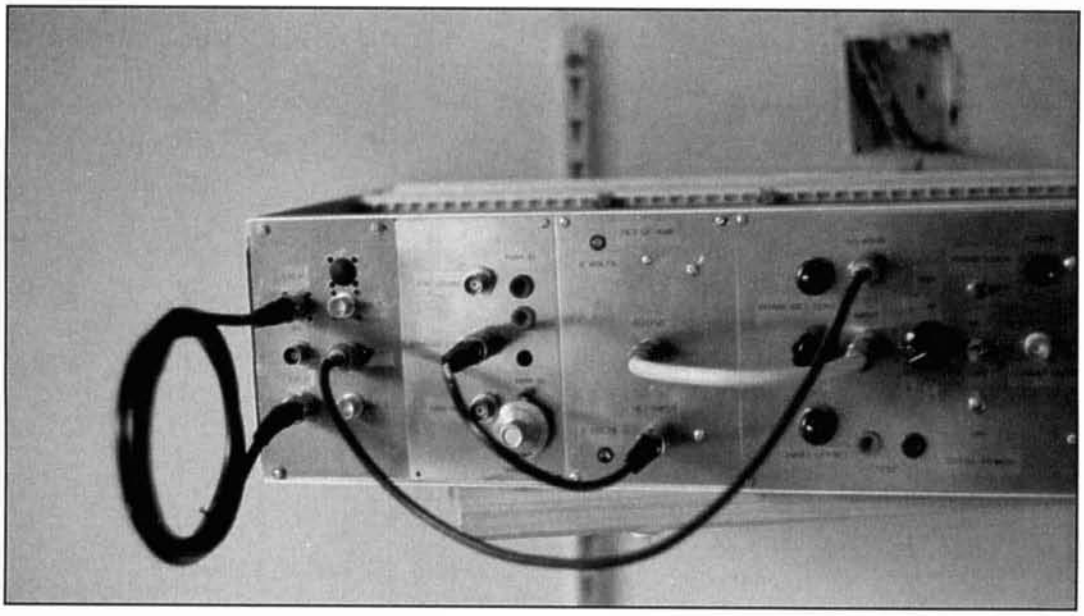

Photo D. The half-wavelength loop attached to the antenna switching unit.

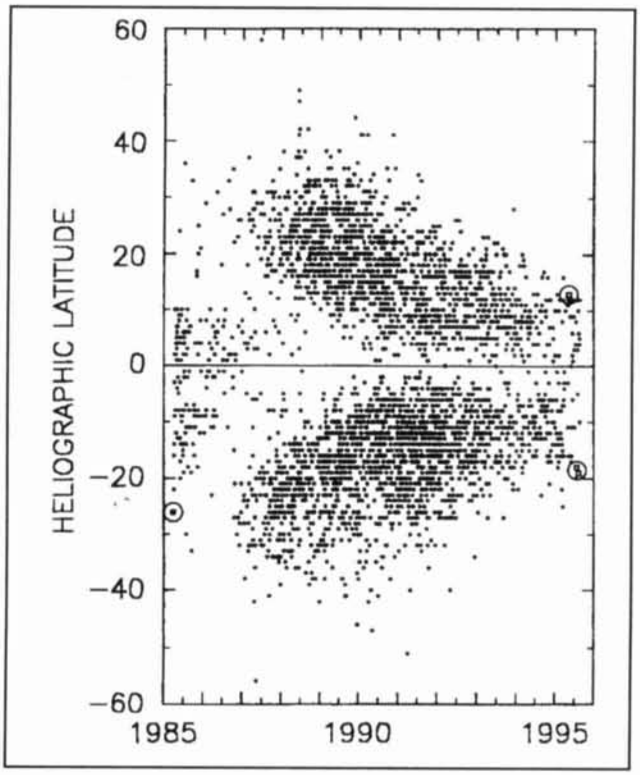

Figure 4. Butterfly diagram showing the heliographic locations of all emerging sunspot groups during solar cycle 22 (through August 1995), as determined by the *NOAA/USAF SOON* network. The first spot group of cycle 22 is represented by the filled-square symbol at the far lower left. The probable first groups of the new cycle (cycle 23) are depicted by the open-square symbols to the extreme right.

column; in the meantime, interested parties can contact Peter through the following addresses:

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Cycle 23 Sunspots Observed at Learmonth Observatory

During mid-August, John Kennewell, Chief Physicist at Learmonth Solar Observatory in Australia, provided the astronomical community with one of the earliest reports of activity from the coming solar cycle (cycle 23). His report and analysis of these events follows.

As of August 13, a total of three sunspot groups potentially belonging to new solar cycle 23 have been observed at Learmonth Solar Observatory, jointly run by the *USAF Space Command* and *IPS Radio and Space Services* (**Figure 4**). Using hourly full-disc magnetic diagrams via a network connected to the recently established NSO/GONG site on the North West Cape of Western Australia, observers on August 13 reported the third potential new cycle sunspot group. These groups were also observed by Kitt Peak magnetogram and at the Big Bear Solar Observatory in California. Details of the groups are:

Date	NOAA Region	Latitude	Carr. Long.	Max # Spots	Class
May 13	7872	N13	321	2	Bxo
Jul 26	7893	S18	125	3	Axx
Aug 13	7899	S20	215	2	Bxo

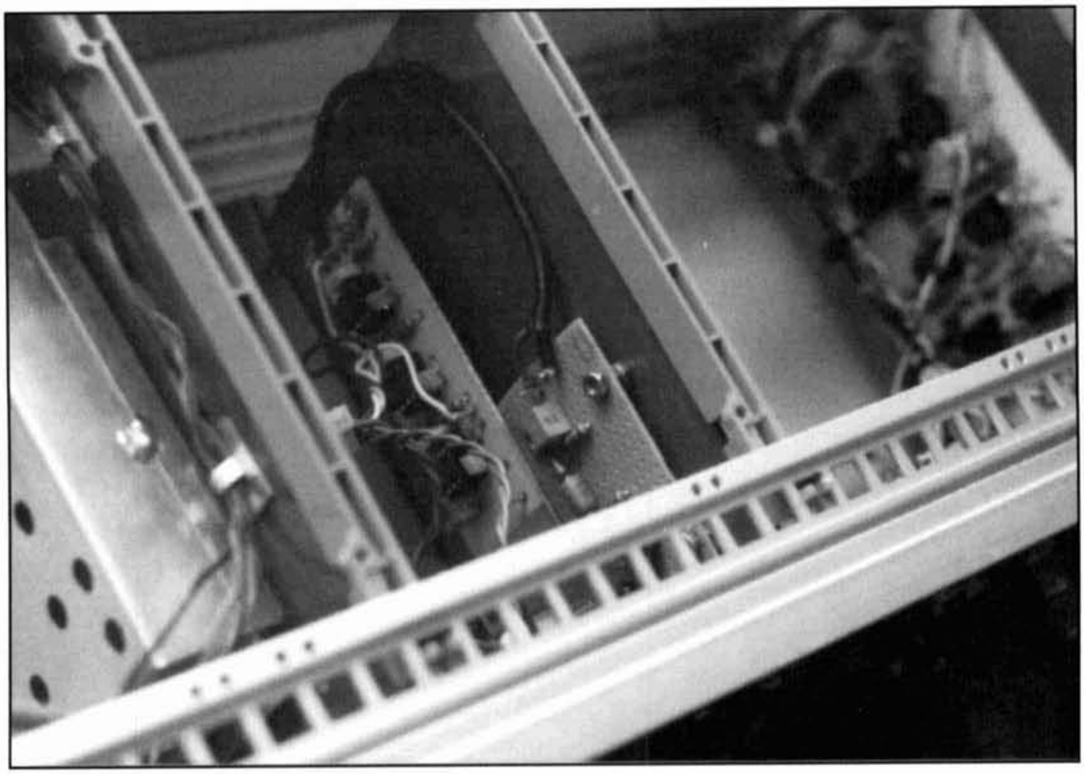

Photo E. The IF amplifier and detector circuits.

All of the above regions had the correct hemispheric magnetic polarity for cycle 23. (The magnetic polarity of regions in the Sun's Northern and Southern Hemispheres reverses each solar cycle.) Although it did not develop sunspots, a fourth region located at 38 degrees north latitude on August 24 also showed new cycle polarity. However, the low latitudes of these sunspot groups could call into question their cycle allegiance. In recent cycles, the latitudes of the first spotted region has typically lain between 25 and 40 degrees.

It is not uncommon for a region belonging magnetically to one hemisphere to become

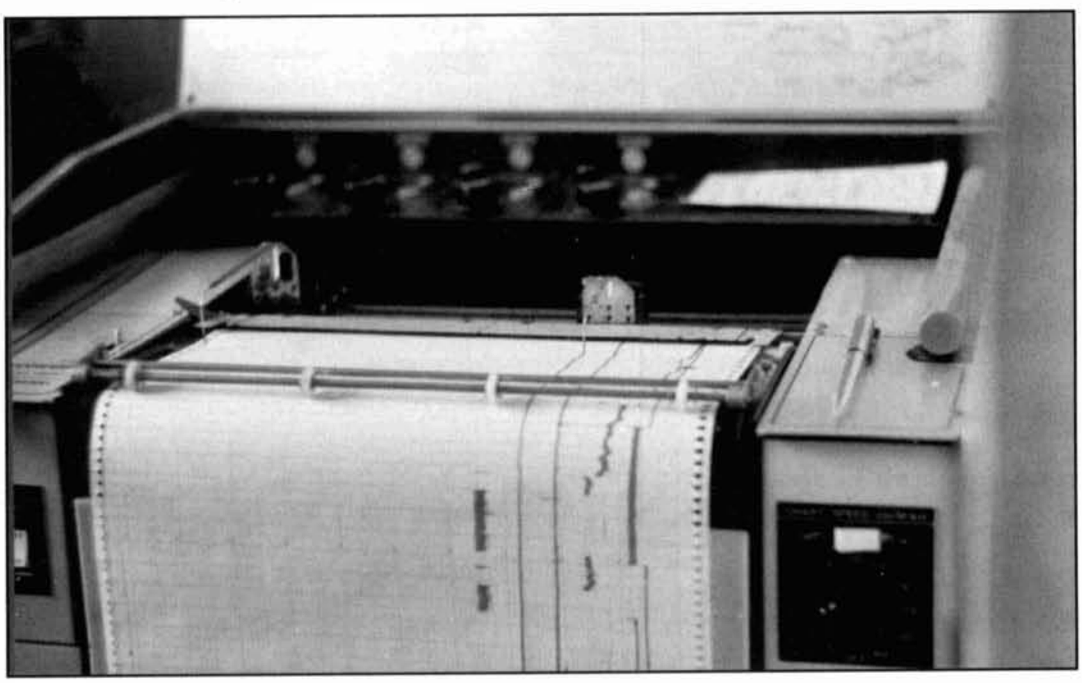

Photo F. Peter King uses this four-pen chart recorder to register the results of his radio emission measurements.

'lost' and 'stray' into the wrong hemisphere. Most such groups are located at latitudes less than 10 degrees, while from 10 to 20 degrees we have a grey area. However, in lost groups the leader polarity is typically poleward of the trailer. Such was not the case for Regions 7893 and 7899. Both of these regions had the leader inclined toward the equator. In fact, the axis of Region 7893 was inclined almost 45 degrees equatorward.

The major significance of new cycle spots lies in their predictive value for the minimum of the current cycle which coincides with the onset of the new cycle. Typically, minimum does not occur until at least 12 months following the appearance of the first spot group of the new cycle. Details for the last three cycles are:

Cycle	First Spot	Minimum	Latency
20	Sep 1963	Oct 1964	13 months
21	Nov 1974	Jun 1976	19 months
22	Mar 1985	Sep 1986	16 months

On the basis of past behavior we would thus expect the next solar minimum to occur between June and December 1996 and the duration of cycle 22 to lie between 9.7 and 10.3 years, noticeably shorter than the average cycle length of around 11 years. ■

PRODUCT INFORMATION

New DC/AC 600A Clamp Meter From B+K

B+K Precision's new Model 366 clamp meter measures AC and DC current and voltage, resistance, and line frequency. It has audible continuity and diode tests, plus peak hold and data hold.

Model 366 measures AC and DC current in two ranges at 2% accuracy with 100mA maximum resolution. The jaws accommodate conductors up to 1.5 inches in diameter. DC voltage is measured at 0.8% accuracy up to 1000V. AC voltage up to 750V is measured in two ranges at 1.2% accuracy and 100mV maximum resolution. Resistance measurements to 2000Ω are accurate to 1%. Frequency can be measured to 2000Hz with 1 Hz resolution and 1% accuracy. Data Hold instantly freezes the display reading, while Peak Hold captures peak voltage and current readings as short as 100ms.

The Model 366 measures 8.2 x 2.6 x 1.3 inches and weighs 11.6 ounces. It comes with 9V battery, carrying case, test leads, and instructions. For more information, contact B+K Precision, 6470 W. Cortland St., Chicago, IL 60635; or call 312-889-1448.

New EZNEC Antenna Software From W7EL

EZNEC overcomes the limitations of MININEC programs because it is based on NEC-2 code. EZNEC has all the advanced features of its predecessor ELNEC including 3-D antenna view display, with currents and antenna pattern superimposed; group editing features to help make complex antenna descriptions; the ability to rotate wires and change wire lengths or antenna height with a single entry frequency sweep; multiple plots on the same screen; and one-key analysis to display gain, front/back ratio, beamwidth, and other important information. In addition to ELNEC features, EZNEC also offers 500 segments for very complex antennas; transmission line models, including stubs; color printing with DeskJet printers; copying of plots to Windows documents; automatic segmentation; comprehensive automatic modeling guideline check; and full NEC-2 ground model.

EZNEC requires an 80386, 80486, or Pentium processor; coprocessor (built in to 486DX and Pentium); and at least 2 megs of available extended memory. Required disk space depends on the amount of available RAM. The program requires less than 2 megs. If only 2 megs of RAM are available, EZNEC will require 9 megs of disk space to analyze a 500-segment antenna, or no additional disk space for up to about 200 segments. Less disk space is needed if more RAM is available. An EGA or VGA monitor is required. Supported printers are HP LaserJet, color and monochrome DeskJet, and Epson dot matrix, or compatible types. EZNEC can send a setup string to the printer to put it into an emulation mode. The program can be run under Windows as a DOS application.

EZNEC is $89.00, postpaid to the United States and Canada. ELNEC is $49; MaxP, $25. Visa and MasterCard accepted. For more information, or to place an order, write to Roy Lewallen, W7EL, P.O. Box 6658, Beaverton, OR 97007; or call 503-646-2885; or e-mail w7el@teleport.com.

New Hewlett-Packard Catalog

Hewlett-Packard Company's 1995/96 Power Products Catalog (Literature 5963-3906 EUS) is now available. The 66-page catalog contains the latest technical information on alternate-current sources, direct-current power supplies and electronic loads, power-test systems and solar-array simulators for the lab or factory. Extensive listings of application information, dimension drawings, front-panel photos and product specifications are also included.

Copies of HP's 1995/96 Power Products Catalog can be obtained in the United States by calling the HP DIRECT support number: 1-800-452-4844.

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MODELING AND UNDERSTANDING SMALL BEAMS: PART 3

The EDZ family of antennas

The Extended Double Zepp (EDZ) has been a lonely antenna for most of its life. Even its vertical cousin, the Extended Single Zepp, seems to have changed its name to the 5/8-wavelength ground plane to avoid identification with the EDZ.[1]* By contrast, everyone knows about the family of antennas related to the half-wavelength dipole: the quarter-wavelength ground plane, the 2, 3, and more-element Yagis, the ZL Special. The EDZ also spawns a family of antennas that include parasitic and phase-fed beams. I'll try to fill in some of the family tree without necessarily recommending everything that the computer modeling says is

*This tongue-in-cheek introduction does have a serious point. The Jones handbook mentions the 5/8-wavelength vertical, especially as a broadcast antenna, as early as 1936—*if not before*. In my collection of old handbooks, the EDZ is not even mentioned, under that or any other name, into the 1950s.

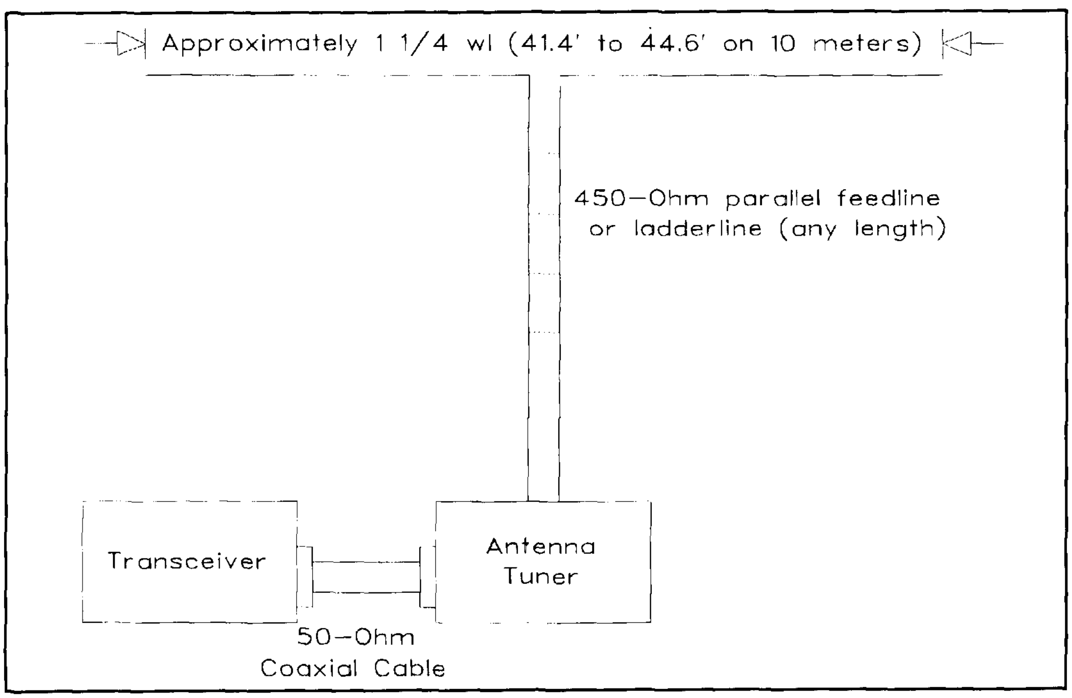

Figure 1. The basic structure of a horizontal single-wire EDZ antenna.

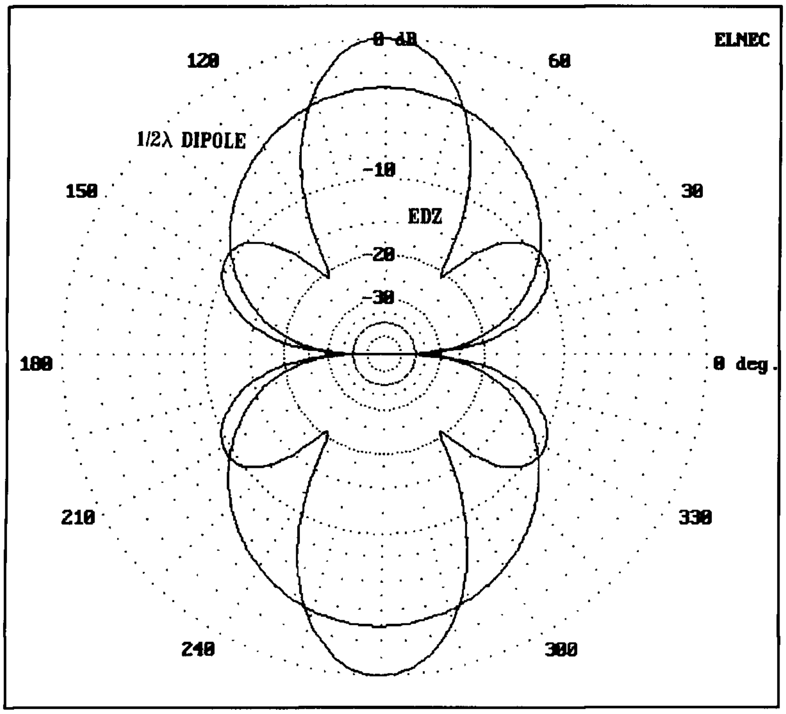

Figure 2. Free space azimuth patterns of a 1/2-wavelength dipole and a single-wire EDZ antenna.

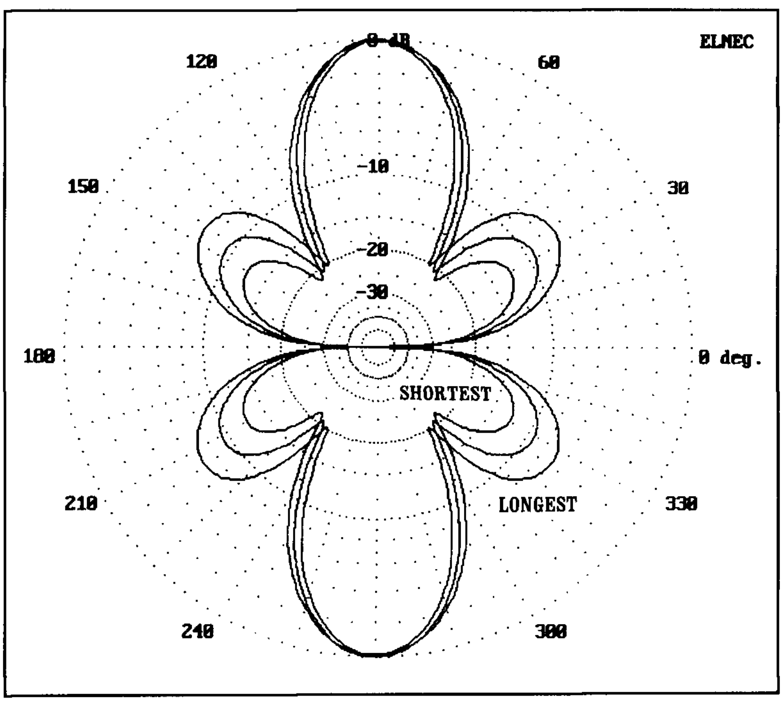

Figure 3. Three free space azimuth patterns for short (41.4 foot), medium (43.0 feet) and long (44.6 foot) EDZ antennas.

theoretically possible with the EDZ. What looks good on the computer may not work in the backyard. However, I hope the ideas we'll consider may spur someone else to realize some of the potential shown by EDZ beams.

Here's a sample: A good 2-element dipole-based Yagi shows about 3-1/2 dB gain over a similarly placed wire dipole. It would take 4 to 5 elements to achieve 6 dB gain over that same piece of wire. Suppose one could make an 2-element antenna with the same 6 dB gain over the original wire. That fact would qualify the EDZ beam as a small beam in boom length, although not in element length. Both parasitic and "135-degree" phase-fed versions of the EDZ promise to give the computer-indicated performance. However, achieving that performance will impose severe restrictions that mark the antennas for special purposes under narrowly defined circumstances.

For this study, all antennas are referenced to 10-meters, with any exceptions clearly noted. Within limits, the results can be scaled, at least within the upper HF region of the spectrum.

The basic EDZ

Before turning to the more distant cousins of the EDZ, let's begin with the basic horizontal EDZ. This is a wire antenna, about 1-1/4 wavelengths long, fed in the center. It presents a high, complex impedance at the feedpoint, ordinarily necessitating the use of parallel feeders and an antenna tuner. **Figure 1** suggests the basic setup.

The chief advantage of the EDZ over a resonant half-wavelength dipole is bidirectional gain. (See **Figure 2**, which provides patterns for free space half-wavelength and EDZ dipoles.) Relative to a wire dipole, whether in free space or over real ground, the EDZ provides about 2.9 dB gain over the resonant dipole, with a bandwidth in the main lobe some 45 degrees narrower. For the cost of wire, the EDZ provides some significant advantages.

Some authors have provided formulas for cutting the EDZ. The most common is based on the long wire length formula:

L (in feet) = 984(N - 0.025)/f (in MHz)

where:

N = number of wavelengths, or
L (in feet) = 1205/f (MHz).

Beers prefers a constant of 1218, while recent editions of *The ARRL Antenna Book* call for 0.64 λ per side, which yields a numerator of 1258.[2,3]

Most wire models using no. 12 to no. 18 show their maximum gain in 10-meter models with a length closer to 1225/f (MHz). However, such number-crunching devices are misleading. They obscure the fact that selecting the length of an EDZ is always a compromise between gain and another factor that dipole builders do not have to confront: side lobes. **Figure 3** overlays three

EDZ patterns for wires of different lengths. The pattern with the greatest gain also has large side lobes and a narrow beam width. The other patterns show slightly less gain (less than –0.1 dB), but the shortest model (L= 1180/f (MHz)) also has the least off-axis gain. The cost of smaller sidelobes is a higher capacitive reactance at the feedpoint. In the end, selecting a length for an EDZ may be determined less by absolute gain potential than by the amount of off-axis QRM to be tolerated. **Table 1** summarizes the modeled variations of gain and front-to-sidelobe ratio for 10-meter antenna lengths from 41.4 to 44.6 feet in free space.

Had **Figure 3** showed further shortening of the antenna wire, the side lobes would have eventually disappeared—just as the antenna length approached a single full wavelength. Moving in the other (longer) direction, the main lobe gain quickly falls off to yield the traditional 6-petal pattern of a 1-1/2 wavelength antenna. **Figure 4** illustrates these extremes.

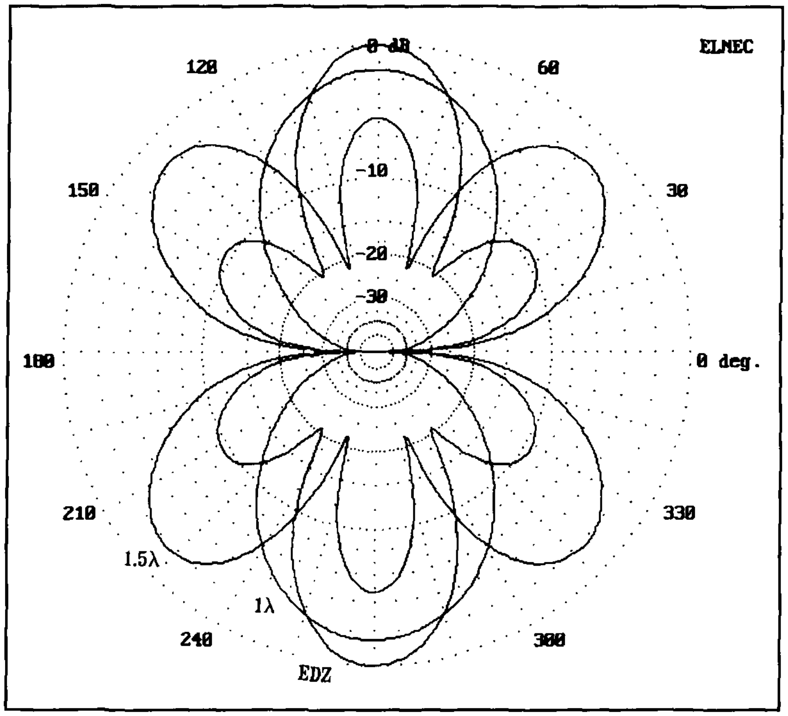

Figure 4. Comparative free space azimuth patterns for 1, 1.25, and 1.5-wavelength wire antennas.

With the EDZ, there are also subtleties occasioned by antenna height. Within the usual amateur backyard building limits (20 to 70 feet antenna height), antenna gain will vary, with peaks at the 5/8, 1-1/8, and 1-5/8 wavelength heights and minima at the 7/8, 1-3/8, and 1-7/8 wavelength heights. Unfortunately, the 7/8 wavelength height, about 30 feet on 10 meters and 35 feet on 12 meters, is often a tempting and convenient amateur construction height. However, moving up a quarter wavelength in height (to about 39 feet on 10 meters) can increase gain by 1.6 dB over the 7/8 wavelength point and decrease the main lobe takeoff angle by 4 degrees, with a resulting increase in radiation at the lower angles most favorable to DX. Above a height of 1-1/2 to 2 wavelengths, the gain fluctuations with height become insignificant, but the radiation at low radiation angles continues to increase. Half-

EDZ Antenna Length vs. Gain and Sidelobes

Length (Feet)	#12 Copper Wire Gain (dBi)	#12 Copper Wire Front-to-Side Lobe Ratio (dB)	#18 Copper Wire Gain (dBi)	#18 Copper Wire Front-to-Side Lobe Ratio (dB)
41.4	4.94	–15	4.90	–16
41.8	4.98	–13	4.94	–14
42.2	5.01	–11	4.96	–11
42.6	5.02	–10	4.98	–10
43.0	5.03	–9	4.98	–9
43.4	5.02	–8.5	4.97	–8.5
43.8	5.00	–8	4.95	–8
44.2	4.96	–7	4.91	–7
44.6	4.90	–6.5	4.85	–6.5
	$\Delta G = 0.13$ dB $\Delta F\text{-}S = 8.5$ dB		$\Delta G = 0.13$ dB $\Delta F\text{-}S = 9.5$ dB	

Notes:
1. Antenna model is for 28.5 MHz in free space.
2. Gain figures are recorded to 2 decimal places for comparison purposes only. Single digit differences in the first decimal column are unlikely to be significant to performance.
3. Front-to-side lobe ratios are estimated from antenna plots. Higher accuracy is not required to show the trend in sidelobe growth with antenna length.

Table 1. EDZ antenna length versus gain and front-to-side lobe ratio.

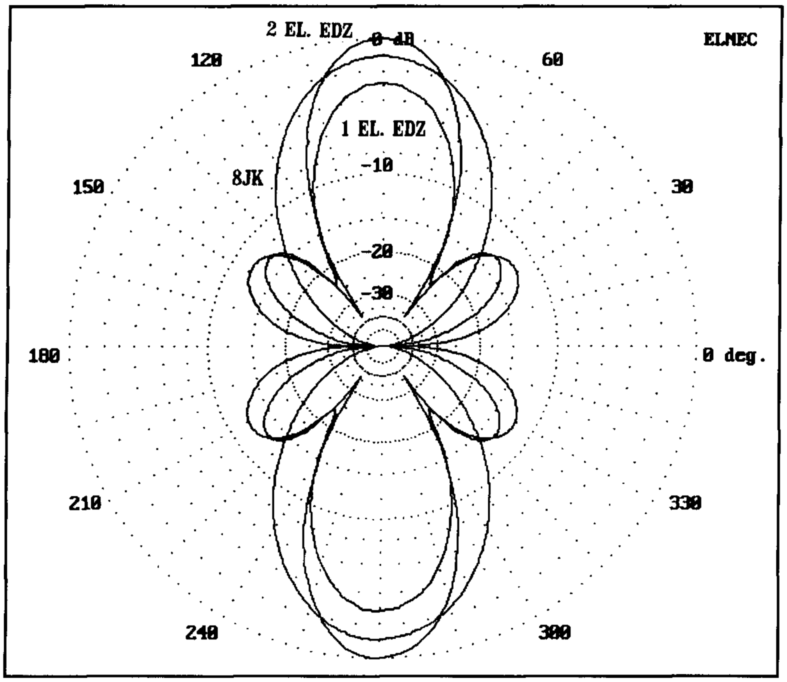

Figure 5. Comparative free space azimuth patterns for the 8JK, the single-wire EDZ, and the 2-element 180-degree phase-fed EDZ array.

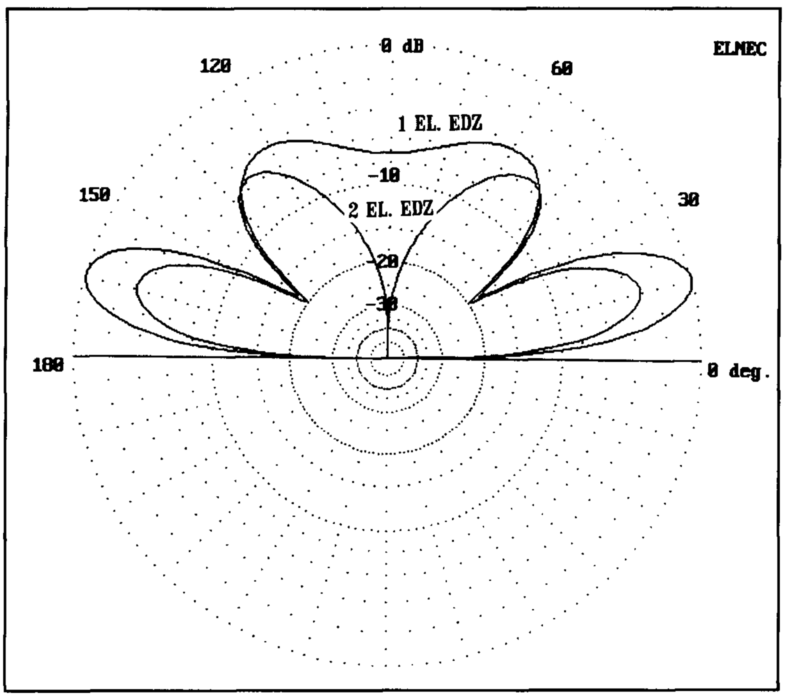

Figure 6. Elevation patterns over real ground for the single-wire and the 2-element 180-degree phase-fed EDZ antennas.

wavelength dipoles also show such fluctuations at about the same heights, but to a much lesser degree. Gain variations with height result from reflected antenna currents being reinduced into the wire at phase angles that vary according to antenna height. The same phenomenon also creates a varying feed point impedance.[4]

The EDZ is a nonresonant antenna, displaying great capacitive reactance. Within the range of reasonable lengths (about 41.5 to 44.5 feet on 10 meters), the antenna shows a feedpoint impedance ranging from 175-j930 Ω at the short end to 110-j640 Ω at the long end. The impedance—both the resistive and reactive components—falls off more rapidly as the length passes the midpoint (43 feet), where the impedance is about 150-j840 Ω.

Most commonly, hams feed the EDZ with open-wire or similar parallel transmission lines and an antenna tuner. This style of operation permits the operator to use the antenna on other bands in a way similar to the use of center-fed (double) Zepps in the 1930s. (I suspect that this fact contributed much to the name "extended double Zepp.") More recently, Yardley Beers, WØJF, reminded us that impedance matching does not have to be performed at a distance from a highly reactive antenna. He developed a system of transformer matching between the antenna and a coaxial feedline. The secondary of the transformer not only provided the step-up ratio for the resistive component of the impedance, but also provided the inductive reactance to compensate for the antenna's natural capacitive reactance.[3]

Stub matching to a 50-ohm coaxial feedline is also possible. One selects a line (for example, 450-ohm parallel line) and, by calculation or experiment, chooses a length that results in a 50-ohm resistive impedance when a suitable stub is connected in parallel across the junction of the matching section and the main feedline. A 44-foot long no. 14 wire 10-meter EDZ at about 35 feet above average ground would require a 450-ohm (0.95 VF) matching section just over 5 feet long and a parallel shorted stub of the same material just over 1.2 feet to provide a perfect match to 50-ohm coax. Setting more precise dimensions than these would require information on the antenna's feedpoint impedance over the actual terrain of the site. Stub matching should result in a very reasonable 2:1 SWR bandwidth of over 800 kHz on 10 meters. See the Appendix to review the characteristics of stub matching and a method of calculating the elements of such a system.*

The 180-degree phased 2-element EDZ

John Reh, K7KGP, was perhaps the first in

* K7KGP's 12-meter EDZ, shown in recent *ARRL Handbooks*, can mislead builders, because his stubless match applies only to the antenna feedpoint impedance figures he lists. (See *The ARRL Handbook*, page 33-11.) The matchline he specifies leaves an ohm or 2 of remnant reactance, much too little to be of concern. Except in very rare cases, other feedpoint impedance figures will require a stub to compensate for reactance at the junction of the matchline and the main feedline. The Appendix to this article provides a method of directly calculating both the matchline and stub elements of a stub-matching system.

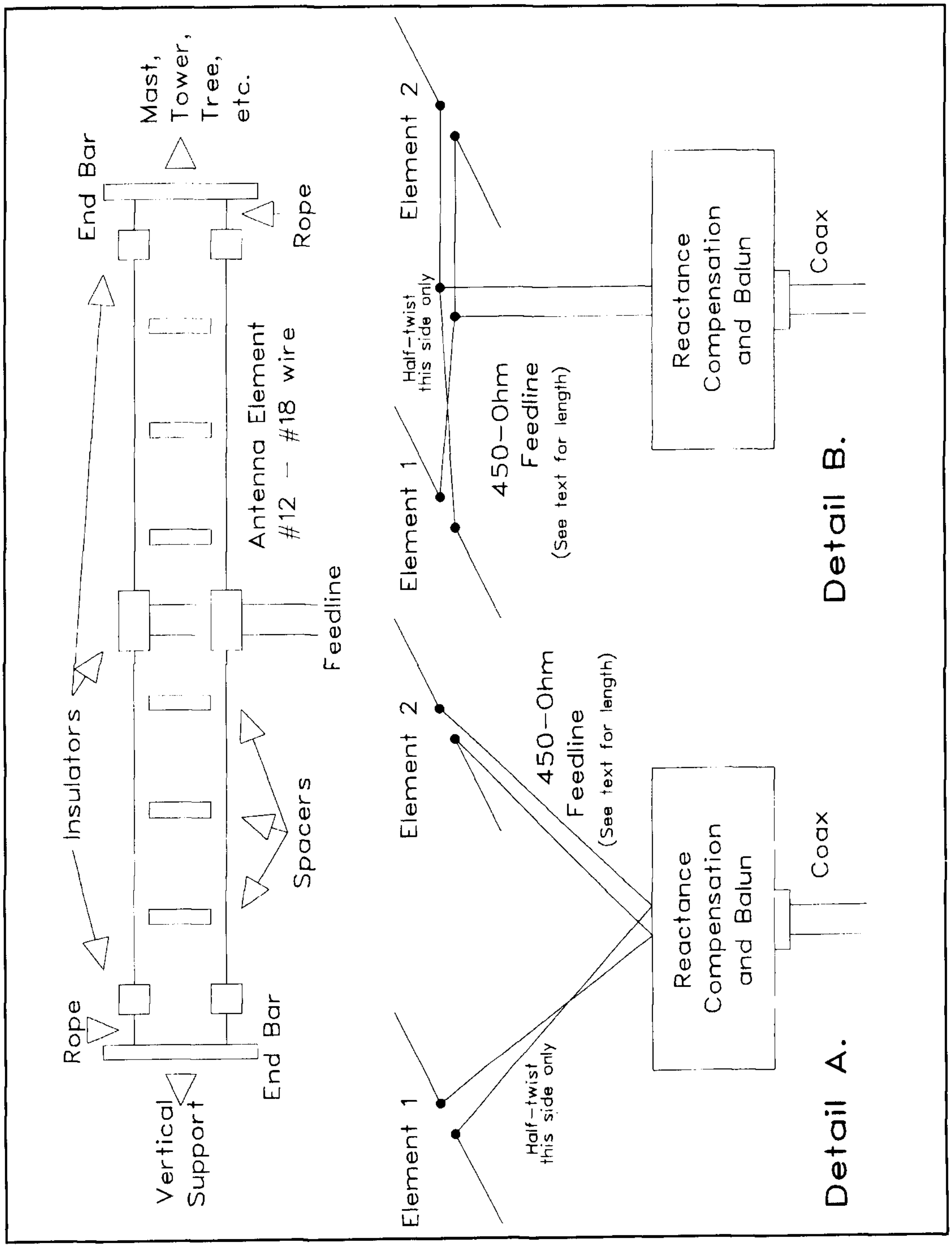

Figure 7. General construction outline of a 2-element 180-degree phase-fed EDZ array with two different methods of feed.

recent times to experiment with 2-element arrangements of the EDZ, developing a 180-degree phased array of identical elements spaced 1/8 wavelength apart. The antenna is an extension of one version of the "two-section W8JK," which used 1-wavelength elements. **Figure 5** provides free space patterns of the 8JK and the phased EDZ antennas, both bidirectional arrays, along with a single-element EDZ. The phased EDZ provides about 1.1 dB gain over the 8JK and about 2.9 dB gain over a single EDZ. This is equivalent to about 5.8 dB gain over a 1/2-wavelength wire dipole equally situated.

One advantage of the 180-degree phased EDZ array, like all other 180-degree phased arrays of any length, is the immunity of the antenna to variations in gain and impedance with changes in height. From 0.5 to about 2 wavelength, the

Relative Free-Space Gains of Various Antennas in the EDZ Family	
Antenna	**Gain (dBi)**
1/2L Dipole	2.07
1-el. EDZ	4.85
2-el. reference Yagi	6.30
2-el. EDZ, 180-degree phase fed	7.85–8.10
2-el. EDZ, parasitic	8.70–9.25
2-el. EDZ, 135-degree phase fed	8.95–9.30

Notes:
1. The "2-el. reference Yagi" refers to the modified W6SAI 10-meter beam used as a standard of broadband 2-element design throughout this series.
2. All values are derived from computer models and, except for the reference Yagi, average several designs using, as relevant, different element lengths, spacings, and wire sizes.

Table 2. The relative free-space gains of various santenna in the EDZ family, along with a reference dipole and Yagi.

gain of the array climbs quickly and then more slowly to essentially flat-top above 1.2 wavelength. The impedance remains quite constant, with the reactance varying by less than ± 1 ohm. The difference in these characteristics from their counterparts in a single-wire antenna is due to the cancellation of radiation vertically (both incident and reflected), thus reducing the complexity of radiation interactions with the elements with changes in height. **Figure 6** compares the elevation patterns of a single wire EDZ with its 180-degree phased counterpart over medium earth at a height of 35 feet, about 1 wavelength at 10 meters.

Figure 7 shows the general outline of the phased array, along with two feed systems. Dimensions of the elements aren't critical. Neither is spaced exactly. Construction can consist of two no. 18 copperweld or no. 12 to 14 copper wires with spreaders every 5 to 8 feet. For 10 meters, use 4.5-foot lengths of thin wall PVC conduit. Hacksaw slots into the ends at points 4.3 to 4.4 feet apart. Drill the ends of the cuts to pass the wire with friction. Leaving burrs on the holes increases friction and holds the wires in place. Press the wires into the slots until they reach the holes.

A 2-year test of this system showed no tendencies for the spreaders to slip from their initial positions, even without the use of adhesives or additional wire ties. Single-end supports (towers, trees, guyed masts, etc.) are adequate for the antenna if the element ends, extended by 3/16 to 1/4 inch diameter sun-resistant synthetic rope, are attached to a longer and sturdier PVC length. Schedule 40 material is strong enough to permit the installation of eye-bolts. The end ropes pass through the eye-bolts and down to a tie-off point for raising and lowering the antenna.

Feed systems are numerous, but only two are shown here. Section A in **Figure 7** shows individual feedlines brought together with a reactance-canceling stub, while Section B shows a taut section of feedline between the elements, with a section dropped vertically to the coaxial cable junction. Either system will work over a narrow frequency range. It is possible to model feedline sections and correlate the results with feedline calculations using standard formulas.[5]* Calculations used standard 450-ohm line with a velocity factor of 0.95, while MININEC models used no. 18 wire spaced 0.083 feet (1 inch) apart and NEC models specified transmission line lengths, impedances, and velocity factors. Of course, MININEC models don't treat feedline as feedline, but as part of the radiating structure where the fields tend to cancel each other. Nevertheless, the resulting figures came within construction variations of each other.

For an array consisting of two 44-foot no. 14 elements spaced between 4.3 and 4.4 feet at 28.5 MHz, the feedpoint impedances of the elements are each approximately 20-j650 Ω. For reasonable variations in these dimensions (up to a half foot shorter and wire as thin as no. 18 copperweld), the resistive component will vary by an ohm or two, while the reactive component may range between –600 and –800 Ω.

To achieve a Vee-shaped junction of two feedlines that in parallel produce a resistive impedance of 50 ohms when a compensating parallel stub is added, one needs a pair of lines,

*See Terman, among other sources (for example, Johnson's *Antenna Engineering Handbook*, 3rd edition), for the basic formulas for calculating the impedance, current, and voltage along a lossless transmission line for any length from the load. Fortunately, these formulas are amenable to simple basic programming that, in addition to figures for specific line lengths, will produce charts of results for any desired interval. Such charts permit estimation of desirable line lengths within trimming range. As previously noted, it is also possible to use these formulas to calculate required stub-matching systems; see the Appendix.

each nearly 5 feet long. Remember, one of the two lines has a half-twist to place the elements 180 degrees out-of-phase. A shorted stub about an inch to an inch and a half long provides the proper compensation where the elements join at the point of the Vee. Is the stub essential? Without the stub, the antenna's feed impedance at the junction of the Vee is about 3 ohms resistive and 12 ohms reactive.

Only general figures are given here, because exact numbers depend upon knowledge of all the antenna and feedline variables for a given installation. In the area of line length required for a 50-ohm match, the impedance shows a rapid change per unit length. Consequently, a very fraction of an inch of line length may separate the impedance values generated by slight variations on a given version of the antenna. Trimming must be done in small increments. Series or parallel capacitors of a capacitive stub will compensate for the remnant reactance.

Using a taut 4.4-foot line between elements, with a further feedline centered at the 2.2-foot mark yields a different situation. Each line—a bit under 25 degrees in length—shows an impedance of about 8.5-j300 Ω at the junction point of the two, for a parallel combination of 4.25-j150 Ω. Various models of this structure provided values of 3.9 to 4.3 ohms resistive, with a reactive component of 140 to 175 ohms. Connecting a single length of 450-ohm feedline vertically from this junction created a usable stub-main feed junction about 1.5 feet down the line. A very short, shorted stub (about 2 inches long) in parallel across the junction will provide the 50-ohm match to a coaxial feeder. Due to the very low resistive component of the junction, special care should be taken to ensure as lossless and weatherproof a set of connections as possible. Moreover, the very high reactance-to-resistance ratio indicates that the match will have a quite narrow bandwidth.

Models also suggest an untried scheme as a variant on **Figure 7B**. Without putting a twist in the line, divide the antenna structure down the middle of the taut feedline, making sure each half is fed in the center of its side of the line. Cross-connect the two resulting feedpoints in a parallel connection (essentially putting the twist at the connection), and the feedpoint impedance will be about 4.5-j150 Ω, ready for the same connecting line as above.

Of course, one may opt not to use coax and instead run the parallel feeders all the way to the station antenna tuner. Several factors recommend this method in preference to a stub match. First, system tuning is critical. Matching sections require lengths over which both resistance and reactance vary by great amounts per unit of line length. With either the Vee or the flat-top stub-feed systems shown here, the 2:1 SWR bandwidth is just over 100 kHz wide at 10 meters.

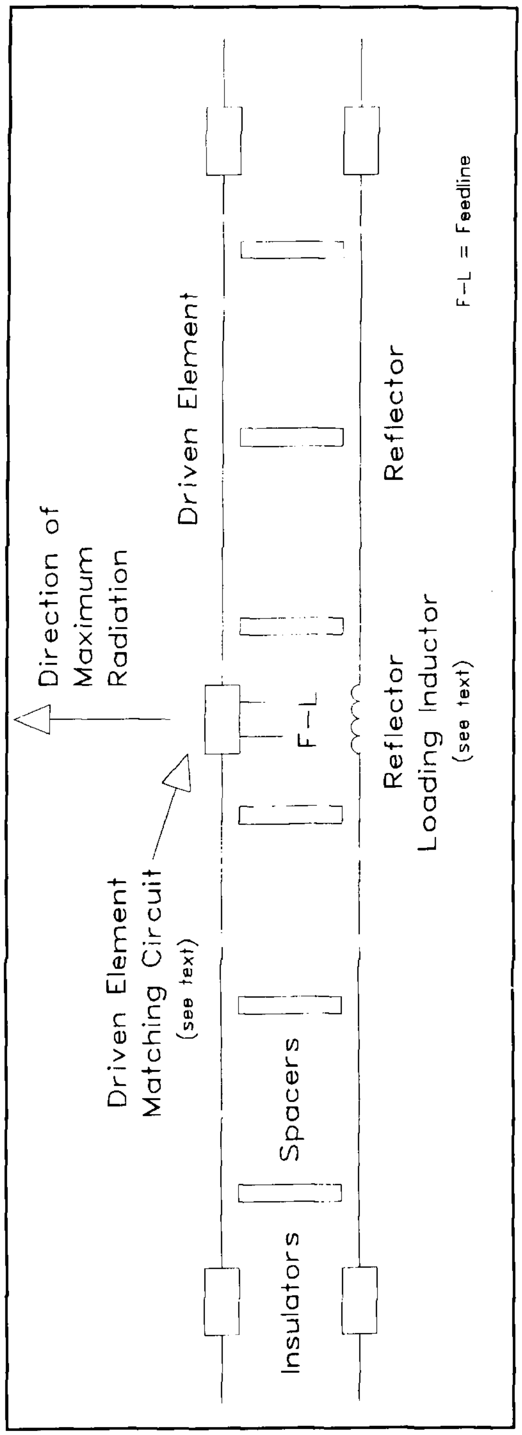

Figure 8. General construction outline of a ZL1LE 2-element EDZ parasitic beam.

Second, the dimensions just given apply to a single height for the experimental antenna; alterations of antenna height from the 35-foot model height will require total recalculation. Indeed, the more critical the dimensions of a

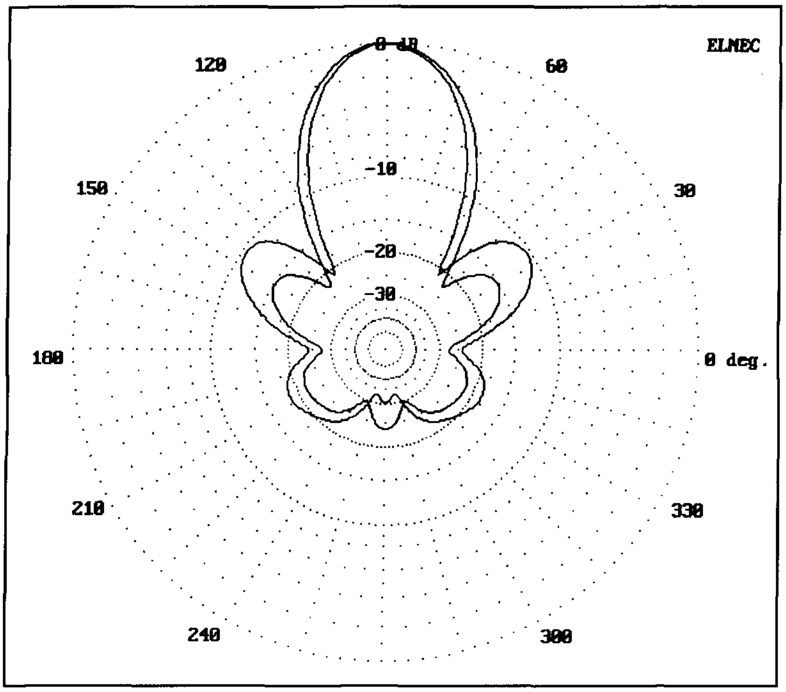

Figure 9. Azimuth patterns for two versions of the ZL1LE parasitic EDZ beam at 35 feet above real ground.

matching line and stub, the greater ease of adjustment a good antenna tuner will provide.

The resulting array produces a pair of opposing narrow main lobes (28 to 32 degrees) with sidelobes about 50 degrees off-axis and down about 12 to 14 dB—depending upon the exact choice of element lengths, spacing, and wire size. At a wavelength in height (35 feet at 10 meters), the take-off angle is about 13 degrees with a –3 dB point at 7 degrees above ground. Models also suggest that interaction between the antenna elements and the feeding-phasing structure may reduce gain by up to a dB from the theoretical optimum. The missing power reappears in the 90 degree off-axis directions, reducing the front-to-side ratio by a small but determinant amount. Nonetheless, for a fixed array where both forward and reverse directions may be useful but do not usually result in QRM, this antenna may be worth the work it takes to prune it to a particular frequency and to a match with the transceiver.

Parasitic EDZ beams

If one could make the 2-element EDZ unidirectional, one might achieve a little more gain, plus have the advantage of reduced QRM from the rear. Theoretically, there are two ways of achieving this goal: a) create a parasitical beam, and b) phase-feed the rear element in a manner similar to the ZL Special. **Table 2** shows a comparison of the free space gains of models of the full EDZ family, along with a standard half-wavelength wire dipole and the 2-element Yagi that has been used as a reference in Parts 1 and 2.

The idea for a two-element beam based on the double extended Zepp was first presented to me in 1991 by Brian Egan, ZL1LE. For his computer studies, Egan proposed a 15-meter model having element spacing of 100 inches, with one element fed and the other a loaded reflector. My concern for keeping the wires properly spaced led me to consider closer spacing, something in the neighborhood of 1/8 wavelength spacing, as used in the ZL Special. The result was the development of two different, but related antenna concepts. One is a double extended Zepp version of the ZL Special, with phased feed. The other, following ZL1LE's lead, uses two elements of the same length, with one fed and the other loaded as a reflector. The computer says both should work quite similarly. The ZL Special version (referred to as the 5/4ZLS hereafter) offers the potential for eliminating loading coils. The symmetrical antenna (referred to as the ZL1LE hereafter) offers the potential for reversibility, allowing me to orient it toward Europe and toward down under just by moving an accessible feedline.

The original ZL1LE antenna used wide spaced elements. The modified version shown in **Figure 8**, uses 1/8 wavelength spacing (4.39 feet) between two equal length elements. The forward element is fed, while the rear element is parasitic. However, to achieve any forward gain and significant front-to-back ratios, the rear element must be loaded inductively. With a carefully selected load, and a load inductor with minimal losses, the antenna is capable of potentially superior performance. The gain at most heights averages across the design bandpass better than 14.5 dBi, or about 6.5 dB greater than a dipole of equal height and orientation. The front-to-back ratio for various models runs from just under 20 dB to more than 30 dB.

Figure 9 shows the azimuth patterns for two model beams at a 35 foot height. The pattern with smaller sidelobes and a smaller front-to-back ratio (about 23 dB) uses no. 12 copper elements 41.67 feet long, spaced 4.39 feet apart, and requires a reflector load of 1035 ohms. The version with larger sidelobes and a higher front-to-back ratio uses 44 foot elements at the same spacing, with the same wire and a 685-ohm reflector load reactance. As with any member of the EDZ family, the balancing of various characteristics determines the final choice of design. In any event, one must model the parasitic EDZ beam over real ground, as in **Figure 9**, to gain a perspective on the actual characteristics. The free space test yields a pat-

Comparisons of Antenna Models Based on the Extended Double Zepp

Height (Feet) Z=10+jX_L	Gain (dBi)	Front-to-Back Ratio (dB)	Front-to-Side Ratio (dB)	S/L Gain (dBi)	Beam Width (°)	Source Resistance R (Ω)	Source Reactance $-X_c$ (Ω)	Load X_L (Ω)
Single Wire EDZ (42.8' Elements, #18 Copper)								
F.S.	4.9		10	–5	32	162	840	
20	11.3		10	1	36	128	860	
25	10.0		10	–1	34	150	810	
30	9.6		10	–1	34	193	830	
35	11.0		10	1	34	159	870	
40	11.0		10	1	32	137	830	
45	10.0		10	0	32	173	820	
2-Element EDZ, 180° Phase-Fed Array (42.8' Elements, #18 Copper)								
F.S.	7.8		15	–7	30	26.4 (x2)	820	
20	12.2		17	–5	32	26.3	822	
25	12.8		16	–3	32	25.2	820	
30	12.7		15	–3	32	26.8	819	
35	12.9		15	–3	32	27.0	820	
40	13.2		15	–3	32	25.9	820	
45	13.2		15	–3	32	26.1	819	
2-Element Parasitic EDZ Beam (42.8' Elements, #18 Copper)								
F.S.	8.7	16.4	11.8	–3.1	32	89.8	775	950
20	14.3	18.0	13.5	0.8	34	80.7	792	905
25	13.6	11.9	13.0	0.5	34	76.6	777	945
30	13.1	18.1	12.3	0.7	32	97.1	760	980
35	14.4	25.3	12.8	1.6	32	96.0	775	935
40	14.4	13.2	12.0	2.4	32	79.1	785	930
45	13.8	14.3	11.9	1.9	32	88.3	768	970
2-Element Parasitic EDZ Beam (41.67' Elements, #12 Copper)								
F.S.	8.8	15.4	14.0	–5.0	34	102.0	830	1055
20	14.5	16.9	15.8	–1.3	36	94.8	852	1000
25	13.7	11.1	14.0	–0.3	36	86.2	832	1050
30	13.5	16.9	14.7	–1.2	34	109.3	813	1095
35	14.5	23.4	15.1	–0.6	34	110.0	830	1035
40	14.6	12.5	14.0	0.6	34	89.2	843	1025
45	13.9	13.4	14.0	–0.1	34	99.1	822	1080

Notes:
1. F.S. = free space
2. Sidelobe figures estimated from graph for bidirectional antennas.
3. Loading coil assumed to have an approximate Q of 100.

Table 3. Comparisons of antennas based on the Extended Double Zepp as modeled at typical amateur construction heights and optimized (where necessary) for maximum front-to-back ratio.

tern whose rear lobes look like miniatures of the forward lobes—a picture that doesn't hold over real ground.

Table 3 compares the modeled performance of two versions of the ZL1LE antenna to the performance of a single-element EDZ and a 2-element 180-degree phase-fed array. Also noted are the necessary changes in loading inductor for the parasitic element to achieve maximum front-to-back ratio at each height. Below a height of about 2 wavelengths, the parasitic EDZ beam in almost any form is height sensitive with respect to gain, front-to-back ratio, and the required loading inductance to achieve maximum front-to-back ratio. A height of about 1 wavelength provides the best combi-

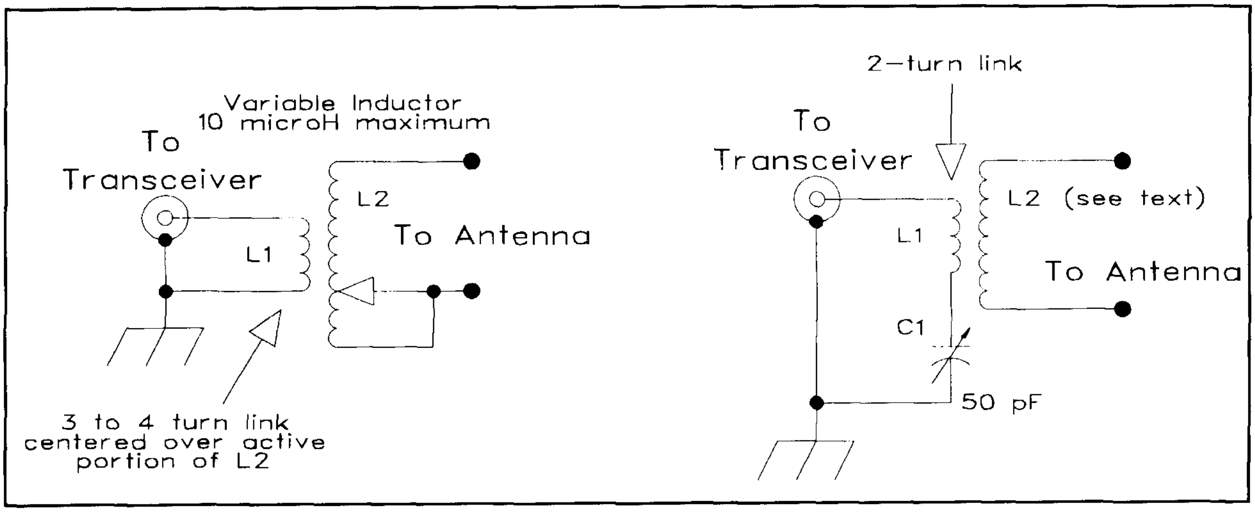

Figure 10. Two feed systems for the ZL1LE parasitic EDZ beam.

nation of gain and front-to-back ratio. In fact, experimental models of the antenna appear to lose much of their DX potential if not placed at least 1 wavelength above ground.*

The source impedance for the antenna shows a large capacitive reactance that requires compensation. Assuming the use of a suitable inductance to eliminate the reactance, the feedline impedance, now only resistive, is roughly twice that of the normally used 50-ohm coaxial cable. A 2:1 quarter-wave matching section of 75-ohm cable cut to design center frequency would likely yield an acceptable match. A linear choke, such as the W2DU ferrite choke balun, would be apt between the feedline and the matching section.

Because this antenna requires inductors in both elements, one to cancel the series capacitive component of the source impedance and the other to load the reflector, the system holds potential for being used as a fixed beam whose direction is reversible. However, if the components are mounted at the antenna, it's unlikely that anyone would climb a structure to readjust the inductor values. Fortunately, the inductors need not be mounted at the antenna.

Low-loss parallel feedline—450-ohm is recommended for its ability to withstand power and weather—permits both matching inductors to be mounted closer to the ground. A wavelength (assuming 0.95 velocity factor) at 28.5 MHz is about 32.8 feet, and a half wavelength is 16.4 feet. For 10-meter antennas at 20 or 35 feet, feedline runs to a platform near the ground are feasible with a length of line that permits the impedance conditions at the element centers to replicate. Using rotary or tapped inductors, one can adjust the loading and the compensating inductors quite easily. A chart of settings would ensure quick adjustment. Installing a coaxial connector near each coil would permit shifting the feedline from one element to the other, thus permitting the direction of the beam to be reversed.

The reason for employing this scheme is to achieve a reversible fixed beam. A beam with 44-foot elements was constructed for reversible parasitic operation. I mounted a rotatable 5-foot plank on the end of a 4-by-4 sunk in the ground. One end of the plank held the reflector inductor, the other held the driven element matching system. The antenna could be reversed by rotating the plank and reconnecting the feedlines. A modified version of the Beers matching system converted the driven element impedance to 50-ohm coax values. **Figure 10** shows two systems tried with equal success. One uses a rotary coil with a fixed 3-turn link of no. 18 solid hook-up wire. The other uses a fixed 2-turn link over a fixed coil of 8 turns of 1.5 inch diameter, 10 turns per inch stock, with a 50 pF variable capacitor in series with the link to control the degree of coupling. Either system amounts to installing an antenna tuner at the antenna, a multiple of a half wavelength below it.

Tuning the inductor to maximum front-to-back ratio requires a variable inductor and a signal source several wavelengths behind the antenna. Coil variability should range of 4.0 to 6.5 μH to cover a 10-meter reactance range of 770 to 1100 ohms. A 10 μH variable inductor picked up at a hamfest provided sufficiently

*A 2-element Yagi at 35 feet provided stronger signals on 10 meters on the eastern United States to VK/ZL path than an initial experimental model of the EDZ beam at 25 feet, despite the fact that models estimated roughly equal radiation in the 5 to 10-degree elevation region. Raising the wire antenna resolved the problem. However, the experience impressed upon me the importance of choosing antenna heights where the lowest required path angle clears fields of obstructions, such as nearby woods with 70-foot trees and the like. The effect can be dramatic.

sharp tuning for tests. Of course, it is essential to weatherproof the tuning components for this scheme to be successful.

Test results showed that the parasitic EDZ beam has excellent gain and reasonably good front-to-back ratio for the frequency to which everything is tuned. However, without retuning—especially the reflector inductor—everything goes to pot very quickly as one tunes off frequency, especially downward. SWR curves don't necessarily provide significant information in this connection. After adjustment of all the variables for 28.5 MHz, an SWR meter in the coaxial feedline showed under 2:1 between 28.1 and 28.9 MHz. However, below the design frequency, the beam had lost its unidirectional characteristic.

Table 4 models the effects of off-frequency use of the parasitic EDZ. If the value of X_L is low by 10 percent, the reflector becomes a director, and the beam reverses its direction. Using values of X_L optimal for a frequency 0.5 MHz higher in the 10-meter band results in performance similar to that of the bidirectional array. A higher X_L value, optimal for a lower frequency, results in a gradual drop in gain and a more rapid drop in front-to-back ratio. If one must choose a single inductor value for the reflector coil, the best choice is the optimum value for the lowest operating frequency in the band. However, the large variation in optimum inductance required across a wide band like 10 meters suggests that the parasitic EDZ beam is best used as a fixed-direction, fixed-frequency or narrow-band antenna. For that use, however, it is inexpensive compared to a Yagi with similar gain and front-to-back ratio.

The 2-meter potential for the ZL1LE, where materials and dimensions make the antenna self-supporting, remains untested. Using 0.75-inch diameter aluminum tubing, one can construct a beam with a driven element 8.2 feet long and a reflector 8.6 feet long, spaced just over 10 inches apart. Models indicate a reflector load between 250 and 305 ohms at 144.5 MHz. The source impedance is about 50-j290 Ω, which simplifies the process of matching the antenna to coaxial cable to the elimination of the reactance alone. Due to the increase in element diameter to element length ratio, bandwidth increases over HF wire models and may cover a full megahertz of the band without undue loss of gain or front-to-back ratio, if the antenna is optimized near the low end of the desired frequency range.

Such antennas almost exist. MFJ advertising includes a double 5/8-wavelength vertical; that is, two such antennas end to end. The same ads indicate that some directionality will result if the antenna is mounted on the side of a tower, which apparently forms an untuned reflector. Perhaps, some day, a manufacturer who can control the reflector loading reactance within tight specifications may produce a true 2-meter ZL1LE antenna.

The horizontal advantages of the 2-meter antenna for low-end CW and SSB operations, despite bandwidth restrictions, are the same as for HF models. However, **Figure 11** shows an azimuth pattern of the antenna mounted vertically, with the center 35 feet above ground. The bandwidth to half-power points is about 135 degrees and the gain is over 13 dBi. The utility of such an antenna at a ham station for both repeater and packet work seems obvious, and three of these antennas arrayed around a tower

Performance of a 2-Element Parasitic EDZ with Nonoptimum Reflector Loading

Measurement Frequency	Optimum Frequency for X_L Used	Optimum Load Reactance (X_L)	Reactance Used (X_L)	Gain (dBi)	Front-to Ratio (dB)
28.0 MHz	28.0 MHz	1030 Ω	1030 Ω	14.6	22.3
28.5		905	1030	13.7	8.0
29.0		795	1030	13.0	5.2
28.0 MHz	28.5 MHz	1030 Ω	905 Ω	–12.4	–0.9
28.5		905	905	14.6	22.3
29.0		795	905	13.6	8.0
28.0 MHz	29.0 MHz	1030 Ω	795 Ω	–12.9	–3.4
28.5		905	795	–12.5	–0.9
29.0		795	795	14.6	22.0

Note:
Modeled antenna used 2 42.8' elements, #18 copper wire space 4.39' apart at a height of 20' over medium earth. Similar results were obtained with other "equal-element" parasitic models at various heights above ground.

Table 4. Performance of a representative 2-element parasitic EDZ with nonoptimum reflector loading.

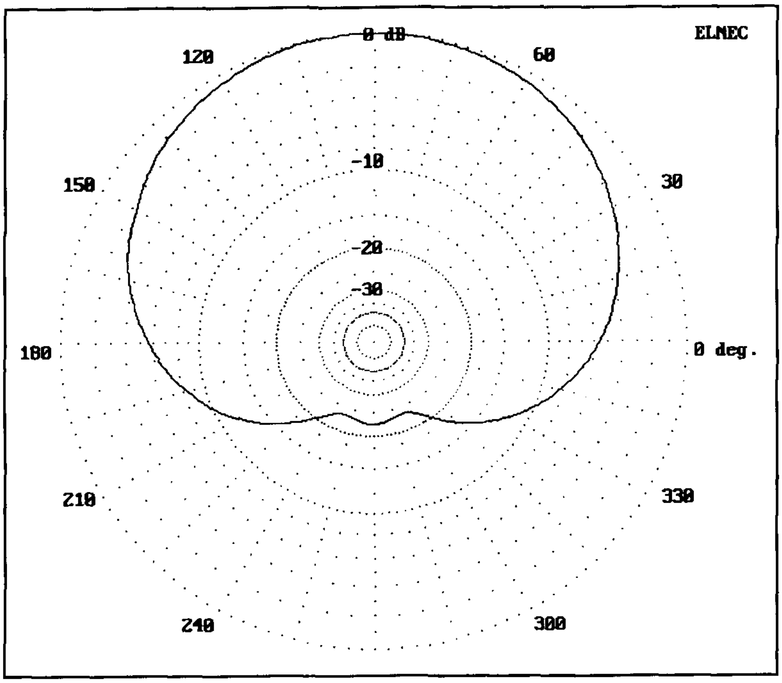

Figure 11. Azimuth pattern for a 2-meter version of the parasitic EDZ beam mounted vertically with the center 35 feet above real ground.

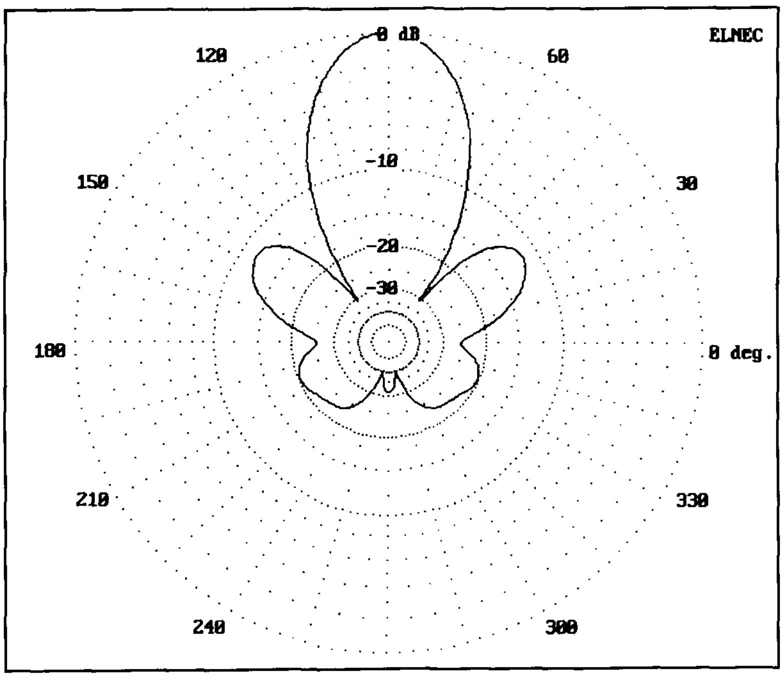

Figure 12. Azimuth pattern of a phase-fed EDZ beam (54ZLS) 35 feet above ground.

might well increase the range of any repeater. My lack of suitable VHF test equipment causes me to leave the development of a working model of the 2-meter ZL1LE to others.

Phase-fed EDZ beams

The phase-fed version of the two-element EDZ beam consists of two unequal length elements spaced 4.31 feet apart for 28.5 MHz. The directly fed forward element is 42.3 feet long, while the phase-fed rear element is 44.7 feet long. Configured as a wire beam, this assembly is unidirectional. Computer models show a peak gain of 14.7 dBi across the 1 MHz design bandwidth, with an average front-to-back ratio of about 20 to 23 dB, peaking at about 30 dB. The beamwidth is a narrow 34 degrees. All of these figures apply to a 35-foot height for the antenna. **Figure 12** shows the pattern of the antenna as optimized for 28.5 MHz.

Experience with ZL-Special would suggest that with 1/8th-wavelength spacing, the rear element should be fed 135 degrees out-of-phase with the forward element. Modeling suggests otherwise. The model whose pattern appears in **Figure 12** is current phase fed at 143 degrees to achieve maximum front-to-back ratio. Since the proposed method of feed is a twisted parallel feedline section, the modeling technique was altered. The rear element is modeled in the opposite direction from the forward element, and the phasing directly applied as –37 degrees to the rear element source point. This technique yields identical pattern figures, but provides correct information on current and voltage amplitudes and phases for use in calculating phasing lines.

The chief difficulty in implementing the 5/4ZLS version of the 2-element EDZ is feeding two elements with highly reactive components in such a way as to ensure that equal power flows to both elements and that the rear element is about 143 degrees out of phase with respect to the forward element. The source impedances for the model under discussion are 100-j830 Ω and 5-j660 Ω for the forward and rear elements, respectively. I have been unable to discover any cable of any length that will provide the proper phasing in the manner of the traditional ZL Special. Moreover, the demands of this antenna may exceed even the flexibility of a phasing network. The high reactances at the feedpoint place any network in a region that is experiencing very rapid changes in current phase and in impedance per unit length. Regretfully, the 54ZLS has had to be consigned to the realm of antennas with theoretical potential but no present feasibility.

Nevertheless, models do suggest that the dimensions of this antenna, when fed as a Yagi, will produce excellent unidirectional gain and a good front-to-back ratio with a reflector load of 675 ohms. In essence, the ability to adjust the reflector loading substitutes for a phasing line in bringing the reflector current to the correct magnitude and phase to achieve a deep rear null. However, no unidirectional EDZ array will do much better than about 20 dB front-to-rear ratio when including all parts of the rear lobes.

It's worth noting, however, that what we call "phasing" lines in antennas like the ZL Special are actually impedance transformers, with or without the half twist. Explorations of models using approximately 1/8th wavelength half-twist parallel transmission lines between two identical elements produced an alternative means of matching an EDZ to 50-ohm coaxial cable. A two-element version of the 10-meter EDZ with such a connecting section made up from 600-ohm (0.95 VF) parallel line shows an almost perfect impedance for coaxial-cable feed. The gain of the antenna is about the same as a single element EDZ (about 11 dBi at 35 feet over medium earth), with a slight (less than 0.9 dB) difference between the two main lobes. The SWR is less than 2.5:1 over the first MHz of 10 meters. For the cost of an additional element and some separators, the builder can produce a coax-fed adjustment-free EDZ.

Summary

Understanding the possibilities for EDZ arrays and beams depends on understanding the basic properties of the single element EDZ antenna. Two-element arrays, either bidirectional or unidirectional, are possible and feasible—if they fit the operating needs and circumstances of a particular station. As wire beams, they are fixed and thus fit for point-to-point communications. Their narrow bandwidths and beamwidths reinforce this type of use. However, they are not forgiving of casual building and tune-up practices.

If these factors aren't deterrents to building one of the EDZ family of beams, but part of the needs of a station, then the operator can expect considerable gain over many other types of wire antennas. Moreover, the cost of these antennas, including wire elements and feedlines, is well below the cost of Yagis with equal gain (and equal front-to-back ratio for the ZL1LE). EDZ beams may have a small, but not insignificant, niche in the spectrum of amateur antennas. I don't recommend them, since recommendation would require a detailed knowledge of too many variables. However, I do recommend further experimentation, modeling, calculation, and ingenuity in pursuit of getting the most out of this interesting family of antennas. EDZs and other nonresonant arrays may have gone unjustly neglected in our dipole-and-coax age.*

*All patterns shown in this discussion were plotted on ELNEC 3.02, but figures cited have been cross-checked on various programs and by various means of calculation.

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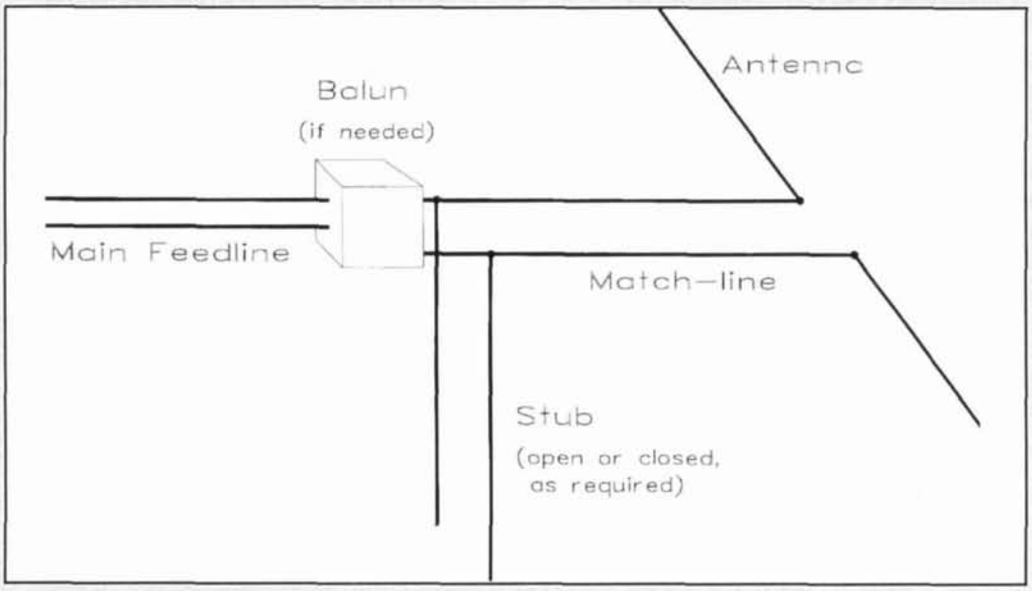
Appendix: Stub matching—a review

As long as hams wish to use or experiment with antennas like the Extended Double Zepp and others that present complex feedpoint impedances, stub matching will remain one alternative method of matching the antenna to a 50-ohm feedline. However, most discussions of stub matching appear almost wholly in qualitative terms. In this Appendix I will attempt to convert stub matching techniques into quantitative terms. We'll review the basic concept of stub matching, present the basic equations for calculating the elements of a stub-matching network, and use a simple BASIC implementation of those equations to solve a couple of exemplary problems.

The stub-matching system

Most antenna manuals provide simple equations for calculating the reactance of both shorted and open transmission line stubs. However, these treatments regularly omit similar equations for calculating the length of the line between the antenna and the stub-feedline junction. With this in mind, let's begin again.



1. Basic elements of a stub-matching network.

Figure 1 shows the basic structure of a typical stub-matching system. It consists of the antenna with its complex feedpoint impedance, a length of matching-feedline (the Line) leading to the critical junction, a reactive Stub, and the main feedline (the Feed) leading to the power source, which is ordinarily a transmitter or transceiver. The functions of the antenna and the main feedline are well-known, but the functions of the other two elements require brief comment.

The matching-feedline operates as an impedance transformer. When there is a complex antenna feedpoint impedance or a mismatch between the antenna impedance and the matching-feedline impedance, the overall impedance, as well as the resistive and reactive components of that impedance, will vary along the line. These values are normally given as series values. If the line type (that is, its characteristic impedance) is properly chosen, at some point along the line, the resistive component of the impedance will be of such a value that its corresponding parallel value will equal the characteristic impedance of the main feedline. This point defines the correct length of matching feedline to use.

Ordinarily, at the junction of the matching-feedline and the main feedline, there will also be a reactive component to the overall impedance. Although usually given as a series value, it too has a corresponding parallel value. A reactance of the opposite type but the same magnitude will compensate for the junction reactance. In this exercise, the compensating reactance will be composed of a feedline stub, even though lumped components (capacitors or inductors) are also usable with somewhat greater losses in some instances. Compensating for the parallel reactance will leave a parallel resistance equal to the main feedline. With the reactance compensated, the resulting series resistance value will be the same, thus effecting a match to the main feedline.

Calculating the matching-feedline and stub lengths

Although often left to graphical analysis along with some miscellaneous calculations, the calculation of match-line and stub systems can be direct. With the advent of home computers and BASIC, the reputed tediousness of the calculations is no longer a hindrance. Indeed, a simple computer program is faster than most graphical methods (some of which have been computerized).

The process begins by understanding that along a match-line, we are seeking the point at which the parallel-equivalent value of the series resistance is equal to the characteristic impedance of the main feedline. Associated with these values is a value of series reactance and its parallel equivalent. If we call the series resistance and reactance the target values, then we define R_T and X_T. Let Z_F be the characteristic impedance of the main feedline. Then, using the series-to-parallel resistance conversion equation:

$$Z_F = \frac{R_T^2 + X_T^2}{R_T} \qquad \textbf{(1)}$$

Solving for X_T^2, we get:

$$X_T^2 = Z_F R_T - R_T^2 \tag{2}$$

Before using **Equation 2**, we must calculate the reflection coefficient, ρ, (actually its square) of the antenna-to-match-line system. Let the match-line characteristic impedance be Z_M. Then, using the antenna feedpoint impedance, $R_L \pm jX_L$, we can calculate:

$$\rho^2 = \frac{(R_L - Z_M)^2 + X_L^2}{(R_L + Z_M)^2 + X_L^2} \tag{3}$$

Using this figure for ρ, we can then determine the value of series resistance at the point in the line defined by **Equation 2**, using that equation to remove reactance values from the calculation of R_T:

$$R_T = \frac{Z_M^2(1 - \rho^2)}{Z_F(\rho^2 - 1) + 2Z_M(\rho^2 + 1)} \tag{4}$$

The target value of reactance is, of course, the square root of **Equation 2**.

The equation, in various forms, for calculating the impedance, Z_{in}, anywhere along a transmission line back from a load, Z_L, is well known.[1,2] That equation can be rewritten as separate equations for R_{in} and X_{in}, which will be more useful for present purposes. We shall use equations for lossless lines for three reasons. First, the lengths of line involved—all well under a wavelength—have losses far less significant than other potential error factors that enter the use of matching stubs. Second, for most types of transmission line, the most imprecise figure is the velocity factor of the line to be used, and most hams don't have access to laboratory-grade measuring equipment to bring experimental determination of that figure under 5 percent. Third, physically replicating a calculated antenna, especially one with a significant reactive component at the feedpoint, usually results in departures from calculated values. Nevertheless, a calculation of the anticipated matching line and stub lengths will do much better than put one in the ballpark: it will allow one to make a close play at the plate.

Since we wish the matching line to yield a resistive impedance component that correlates with the characteristic impedance of the main feedline, we'll begin with the formula that has appeared in *The ARRL Handbook* in the '80s and early '90s.[3,4]

$$R_{in} = \frac{R_L(1 + \tan^2\ell_r)}{\left(1 - \frac{X_L}{Z_O}\tan\ell_r\right)^2 + \left(\frac{R_L}{Z_O}\tan\ell_r\right)^2} \tag{5}$$

where R_L is the resistive component of the antenna impedance, X_L is the reactive component of the antenna impedance, Z_O is the characteristic impedance of the matching section transmission line, and R_{in} is the resistive component of the impedance at a distance ℓ_r from the antenna along the line. In this exercise, R_{in} is precisely our target value of resistance, R_T. For our purposes, let's assume that ℓ_r is in radians; although, in general, it might also be in degrees relative to a wavelength at the frequency of interest for the antenna.

The matching line length calculation simply requires us to solve **Equation 1** for ℓ_r and to convert that length in radians into degrees and feet. A rewrite of **Equation 1** yields a quadratic:

$$\tan^2\ell_r\left(\frac{X_L^2}{Z_O^2} + \frac{R_L^2}{Z_O^2} + \frac{R_L}{R_{in}}\right) - \tan\ell_r\, 2\,\frac{X_L}{Z_O} + 1 - \frac{R_L}{R_{in}} = 0 \tag{6}$$

Solving for ℓ_r, we obtain:

$$\ell_r = \arctan \frac{\frac{X_L}{Z_O} \pm \sqrt{\frac{X_L^2}{Z_O^2} - \left(\frac{X_L^2}{Z_O} + \frac{R_L^2}{Z_O^2} - \frac{R_L}{R_{in}}\right)\left(1 - \frac{R_L}{R_{in}}\right)}}{\left(\frac{X_L^2}{Z_O^2} + \frac{R_L^2}{Z_O^2} - \frac{R_L}{R_{in}}\right)} \tag{7}$$

Note that there are two solutions, because for every 180 degrees of line length (under mismatch conditions), there will be two points at which the resistive component of the impedance has the same value.

The limiting case occurs where the value under the radical in **Equation 7** drops to less than zero. This condition indicates that, with the combination of line values chosen for the antenna impedance values measured or derived from a modeling program, the resistive component never reaches the chosen main feed line characteristic impedance. The solution to this problem is usually to select a different transmission line for the matching line section.

Equation 7 returns two lengths in terms of radians along a wavelength. We can convert these lengths to a more familiar measurement in degrees by the equation:

$$\ell_d = \frac{180\,\ell_r}{\pi} \tag{8}$$

where ℓ_r is the length in radians and ℓ_d is the

STUB

```
10 'file STUB.BAS
20 CLS:COLOR 11,1,3:CLS
30 PRINT"        General Solutions for Stub Matching,":PRINT"        given Antenna R & X
plus Line, Stub, & Feed Zo":PRINT"        L.B. Cebik, W4RNL":PRINT
40 PRINT"For any antenna load R and X, this program finds the Line and Stub length needed to
match any feedline Zo, if a match is possible with the proposed Line, Stub, and Feed Zo val-
ues.":PRINT
50 INPUT "Enter Antenna Load Resistance in Ohms ",RL
60 INPUT "Enter Antenna Load Reactance in Ohms ",XL
70 INPUT "Enter Frequency (in MHz)        ",FQ
80 INPUT "Enter Zo of line (from antenna to stub)    ",ZL
90 INPUT "Enter Velocity Factor of Line (as decimal)        ",VFL
100 INPUT "Enter Zo of Feed (from stub junction to rig)        ",ZF
110 INPUT "Enter Velocity Factor of Feed ",VFF
120 INPUT "Enter Zo of Stub (from line-feed junction)        ",ZS
130 INPUT "Enter Velocity Factor of Stub ",VFS
140 RLS=(RL*RL):XLS=(XL*XL):ZLS=(ZL*ZL):RIS=(RI*RI)
150 RHOS=(((RL-ZL)*(RL-ZL))+XLS)/(((RL+ZL)*(RL+ZL))+XLS)
160 RT=((ZL*ZL)*(1-RHOS))/((ZF*(RHOS-1))+((2*ZL)*(RHOS+1))):RI=RT
170 'IF (ZF*RT)-(RT*RT)<0 THEN 350 ELSE XT=SQR((ZF*RT)-(RT*RT))
180 A=(XLS/ZLS)+(RLS/ZLS)-(RL/RI):B=2*(XL/ZL):C=1-(RL/RI)
190 IF A=0 THEN A=1E-08
200 NUM=((B*B)-(4*(A*C))):IF NUM<0 THEN 350
210 TLP=(B+SQR((B*B)-(4*(A*C))))/(2*A)
220 TLM=(B-SQR((B*B)-(4*(A*C))))/(2*A)
230 LP=ATN(TLP):LM=ATN(TLM)
240 PI=3.141592654#
250 LPD=(LP*180)/PI:LMD=LM*180/PI
260 IF LPD<0 THEN LPD=180+LPD
270 IF LPD<0 THEN LMD=180+LMD
280 LPF=(LPD*VFL)/(.3660131*FQ):LMF=(LMD*VFL)/(.3660131*FQ)
290 PRINT"Possible line lengths are A.";LPF;"feet and B.";LMF;"feet."
300 LR=LP:GOTO 360
310 RIA=RI:XIA=XI:PRINT"For Line length A., RS= ";RI;"Ohms and Xs= ";XI;"Ohms."
320 LR=LM:GOTO 360
330 RIB=RI:XIB=XI:PRINT"For Line length B., Rs= ";RI;"Ohms and Xs= ";XI;"Ohms."
340 GOTO 440
350 IF NUM<0 THEN PRINT"There are no possible solutions with this combination of antenna
```

2. Program listing for STUB.BAS.

length in degrees. Transformation of these lengths into feet involves the equation:

$$L_f = \frac{\ell_d \, VF}{0.366 f_{MHz}} \qquad (9)$$

where L_f is the required length in feet, f_{MHz} is the frequency of interest in MHz for the antenna, and VF is the velocity factor of the matching section transmission line.

Using the value of ℓ_r, we may calculate the remnant reactance by using the *Handbook* formula for X_{in}:

$$X_{in} = \frac{X_L(1 - \tan^2 \ell_r) + \left(Z_O - \dfrac{R_L^2 + X_L^2}{Z_O}\right)\tan \ell_r}{\left(1 - \dfrac{X_L}{Z_O}\tan \ell_r\right)^2 + \left(\dfrac{R_L}{Z_O}\tan \ell_r\right)^2} \qquad (10)$$

where all of the variables have the same meaning as in **Equation 7**. Applying **Equation 10** to the two lengths resulting from **Equation 7** will yield opposing values of reactance. We may choose to match either with a stub. Alternatively, we may calculate the reactance

```
impedance and line impedance.":GOTO 710
360 IF RL=0 THEN RL=1E-08
370 RA=RL/ZL:XA=XL/ZL:T=TAN(LR):TS=T*T
380 DA=(1-(XA*T))*(1-XA*T)):DB=(RA*T)*(RA*T):DN=DA+DB
390 RS=RA*RA:XS=XA*XA
400 RN=RA*(1+TS):XK=XA*(1-TS)
410 XM=((1-RS)-XS)*T:XN-XK+XM:RZ=RN/DN:XZ=XN/DN:RI=ZL*RZ:XI=ZL*XZ
420 IF LR=LP THEN GOTO 310
430 IF LR=LM THEN GOTO 330
440 PRINT"For a record of these calculations, press <Print Screen>."
450 PRINT:PRINT"Press <C> to continue."
460 I$=INKEY$:IF I$="c" OR I$="C" THEN GOTO 470 ELSE 460
470 CLS:PRINT:PRINT"Stub Calculations:":PRINT
480 PRINT"Option A: Rs=";RIA;"and Xs=";XIA;" Ohms
490 XPA=((RIA*RIA)+(XIA*XIA))/XIA:XCOMPA=(-1*XPA)
500 PRINT"The required parallel stub reactance to compensate is";XCOMPA;"Ohms."
510 LRL=ATN(XCOMPA/ZS):LDL=(ABS(LRL)*180)/PI
520 IF XCOMPA<0 THEN LDL=180-LDL
530 LFL=(LDL*VFS)/(.3660131*FQ)
540 LRC=ATN(ZS/XCOMPA):LDC=(ABS(LRC)*180)/PI
550 IF XCOMPA>0 THEN LDC=180-LDC
560 LFC=(LDC*VFS)/(.3660131*FQ)
570 PRINT:PRINT"The required SHORTED STUB length is ";LDL;"degrees or";LFL;"feet.
580 PRINT"The required OPEN STUB length is";LDC;"degrees or";LFC;"feet.
590 PRINT:PRINT:PRINT"Option B: Rs=";RIB;" and Xs=";XIB;"Ohms
600 XPB=((RIB*RIB)+(XIB*XIB))/XIB:XCOMPB=(-1*XPB)
610 PRINT"The required parallel stub reactance to compensate is ";XCOMPB;"Ohms."
620 LRL=ATN(XCOMPB/ZS):LDL=(ABS(LRL)*180)/PI
630 IF XCOMPB<0 THEN LDL=180-LDL
640 LFL=(LDL*VFS)/(.3660131*FQ)
650 LRC=ATN(ZS/XCOMPB):LDC=(ABS(LRC)*180)/PI
660 IF XCOMPB>0 THEN LDC=180-LDC
670 LFC=(LDC*VFS)/(.3660131*FQ)
680 PRINT:PRINT"The required SHORTED STUB length is";LDL;"degrees or";LFL;"feet.
690 PRINT"The required OPEN STUB length is";LDC;"degrees or";LFC;"feet.
700 PRINT:PRINT"Press <Print Screen> to complete the record of calculations."
710 PRINT:PRINT"For another run, press <A>; to quit, press <Q>."
720 I$=INKEY$:IF I$="a" OR I$="A" THEN 10 ELSE IF I$="Q" OR I$="q" THEN 730 ELSE
720
730 RUN "C:\basic\menu.bas"
```

values directly from the square root of **Equation 2**, assigning the signs this way: the reactance associated with the shorter line length will have the sign of the reactance at the antenna feedpoint.

For the stub calculations, we shall first convert the reactance into a parallel value to facilitate mechanical connections for the stub.

$$X_P = \frac{R_S^2 + X_S^2}{X_S} \tag{11}$$

where R_s is the main feedline characteristic impedance, X_s is the calculated input remnant reactance, and X_p is the equivalent parallel reactance the stub is to compensate.

A reversal of the signs of the reactances provides the values that must be returned by appropriate compensating stubs. The length of a shorted stub, when the desired reactance is known, is given by:

$$\ell_s = \arctan \frac{X_{in}}{Z_O} \tag{12}$$

and the length of a corresponding open stub is given by:

$$\ell_O = \text{arccot} \frac{X_{in}}{Z_O} \quad \textbf{(13)}$$

where X_{in} is the desired reactance, Z_O is the characteristic impedance of the transmission line used for the stub, and ℓ_S and ℓ_O are the lengths of shorted and open stubs, respectively. Because the values of ℓ_S and ℓ_O are in radians, they can be converted into feet by the same means used to convert the length of the matching line.

The final step is to select the best combination of matching line and stub for the proposed antenna. Ordinarily—for least loss and mechanical simplicity—the combination with the shortest matching line and stub is most desirable.

A simple utility BASIC program for stub matching

The calculations for a stub-matching system lend themselves to a simple utility program in BASIC or almost any other language. **Figure 2** is the listing for my own program, replete with my personal programming quirks. Lines 10-130 set up the input values for the calculation. Lines 140 through 170 calculate the target resistance value along the match-line. Lines 180-440 calculate the length of the matching-line section and the series resistance and reactance values at that point. The equation is broken down into components to precalculate repetitive parts. Line 200 catches the case where the value under the radical is less than zero. Lines 360–410 calculate the reactance for each of the solutions to **Equation 7**, once more with the relevant equation broken down into segments or normalized. (These lines also recalculate the input resistance of the matching line; I put this in while setting up the program as a check and never took it out, since it involves only a few extra lines. The technique is useful for error catching during the program writing process. However, using R_T and X_T and bypassing these steps would shorten the program somewhat.)

```
                 General Solutions for Stub Matching
       given Antenna R & X plus Line, Stub, and Feed Zo
                        L.B. Cebik, W4RNL

For any antenna load R and X, this program finds the Line and Stub length needed to match
any feedline Zo, if a match is possible with the proposed Line, Stub, and Feed Zo values.

Enter Antenna Load Resistance in Ohms    141.36
Enter Antenna Load Reactance in Ohms     –693.56
Enter Frequency (in MHz) 28.5
Enter Zo of Line (from antenna to stub)   450
Enter Velocity Factor of Line (as decimal)  .95
Enter Zo of Feed (from stub junction to rig)  50
Enter Velocity Factor of Feed      .66
Enter Zo of Stub (from line-feed junction)   450
Enter Velocity Factor of Stub      .95
Possible lines lengths are A. 5.038553 feet and B. 5.485493 feet.
For line length A., Rs= 41.10245  Ohms and Xs= –19.12316 Ohms.
For line length B., RS= 41.10246  Ohms and Xs= 19.12327 Ohms.
For a record of these calculations, press <Print Screen>.

Press <c> to continue.

Stub Calculations

Option A: Rs= 41.10245  Ohms and Xs=  –19.12316 Ohms.
The required parallel stub reactance to compensate is  107.4669 Ohms.

The required SHORTED STUB length is 166.5685 degrees or 15.16963 feet.
The required OPEN STUB length is 76.56851 degrees or 6.973203 feet.

Press <Print Screen> to complete the record of calculations.

For another run, press <A>; to quit, press <Q>.
```

Figure 3. Typical output sheet from STUB.BAS.

The results of the calculations so far can be recorded on paper by a program pause and a <Print Screen> command. As two screens of material will fit on one piece of paper, do not <Form Feed> at this time. Lines 450–710 calculate the required parallel reactive components and the stub values that will compensate for them, two values for each line length. A second <Print Screen> will combine this information with the input values for a complete record. A sample double screen printout appears in **Figure 3**. The antenna is a 10-meter (28.5 MHz) Extended Double Zepp with a 450-ohm stub match system for a 50-ohm feedline where the EDZ models a feedpoint impedance of 141-j694 Ω. Option A is 5.0 feet with a shorted stub of 1.2 feet or an open stub of 9.4 feet, and option B is 5.5 feet with a shorted stub of 15.2 feet or an open stub of 7.0 feet. Option A and a shorted stub provide the mechanically simplest system to implement. You can truncate the decimals in the results almost anywhere, because in most cases, results to the nearest ohm and tenth of a foot will be close enough to allow for antenna system adjustment.

One caution, however, is necessary with the use of calculating programs. Unlike graphical solutions, calculating programs give no feel for the sharpness or broadness of the results; that is, how small physical variations from the calculations will affect the adjustments. In general, the higher the ratio of reactance to resistance at the antenna feedpoint, the sharper the curve. In these cases, small physical variations may require extensive adjustment of the calculated lengths.

Recent *ARRL Handbooks* have presented an interesting 12-meter EDZ cut to a length that provides a feedpoint impedance of 142-j555 Ω.[5] With 450-ohm transmission line (VF=0.95), both options yield 5 feet 5 inches of matching line with negligible reactance, obviating the need for a stub. The impedance presented by the matching line to the coax is 55 ohms. In fact, using the program with a feedline impedance of 50 ohms produces a "no possible solution message." The lesson is that before giving up on a combination, try raising or lowering the feedline impedance by 10 percent to see if a solution emerges. The resulting SWR on the coax will be well within limits. However, K7KGP's antenna is quite unusual, and exact reproduction or scaling for other bands may require extensive on-site adjustment.

These examples only scratch the surface of the usefulness of a utility BASIC program in making matching-section calculations. The limits of stub matching are far wider than these examples. Of course, modeling the results on NEC with transmission-line capabilities permits all calculations to be verified.

REFERENCES

1.Terman, *Radio Engineer's Handbook*, page 186.
2. Kuecken, *Exploring Antennas and Transmission Lines by Personal Computer*, pages 180–181
3. *The ARRL Handbook*, 1987 edition, page 16–2, (normalized version of equation).
4. *The ARRL Handbook*, 1992 edition, page 16–3, (non-normalized version).
5. *The ARRL Handbook*, 1992 edition page 33–11.

PRODUCT INFORMATION

Philips ECG Introduces New Semiconductor Master Replacement Guide

Philips ECG has introduced their 16th Edition ECG® Semiconductor Master Replacement Guide. The Master Guide features over 14,000 additional cross references and nearly 200 new devices, including new product families. With more than 276,000 crosses to Industrial and Entertainment part numbers, the 16th Edition Master Guide is a comprehensive source of replacement information for electronic equipment servicers. Expanded selector guides are also provided to simplify choosing the best ECG replacement type for numbers that are not crossed.

The Guide includes a total of approximately 4,100 components. Among the new devices added are 124 modules and ICs used in VCRs, TV, audio, PCs and industrial equipment applications. Functions include voltage regulators, motor drivers, signal processors, decoders, AFPOs, small signal subsystems and deflection circuits. Also added are a number of transistors, rectifiers, diodes, gas filled surge arrestors, optoelectronic devices and accessories.

All ECG products and literature are available through authorized Philips ECG distributors. To locate the nearest distributor, consult "Electronic Equipment & Supplies" in the telephone directory yellow pages, or call toll-free, 1-800-526-9354.

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QUARTERLY COMPUTING

Putting a PC's serial port to work

As microprocessors rapidly infiltrate every item imaginable, odds are excellent that the next amateur radio equipment you buy will include a means of communicating—probably an RS-232-C serial port—with a host personal computer. In this column, we'll examine what RS-232-C is all about, and review a pair of interesting products that attach to a PC's serial port.

Roots of RS-232-C

Promulgated in 1969 by a committee of the Electronic Industries Association, the RS-232-C standard (or RS-232-C for short) solved a thorny problem—how to specify the interconnection of remote terminals via telephone lines to centralized mainframe computers. A typical link consists of a modem or data set (both referred to as DCE or data-communications equipment) and a teleprinter, CRT terminal, or computer (all referred to as DTE or data-terminal equipment).

RS-232-C defines electrical signals, the mechanical interface, circuit functions, and certain application-specific subsets of the standard. Surprisingly, nowhere does the standard define a specific connector; however, an appendix informs us that ". . . commercial products. . . will perform satisfactorily, such as connectors meeting MIL-C-24308 (MS-18275) or equivalent standards."

Developed sometime after World War II for rack-and-panel mounted avionics equipment, an MIL-C-24308 connector consists of a squashed keystone- or D-shaped metal shell surrounding an insulator containing pins in two offset rows. Commonly referred to as D-subminiatures, these popular connectors are available in sizes from nine to 50 pins.

Weird variants exist—for example, Apple's notoriously hard-to-find 19-pin external floppy-disk drive connector as used on the first-generation Macintosh, and exotic military versions containing coaxial and high-voltage contacts. For our purposes, we'll examine only the nine-pin (DB-9) and 25-pin (DB-25) models.

Mapping the pinout

RS-232-C defines a female connector for the DCE side of the interface, and a male connector for the DTE side. **Table 1** lists functions of all 25 pins, the subset chosen by the IBM PC's designers, and a further subset of nine pins used in PC/AT computers. Depending on its age and configuration, your PC may include both 9- and 25-pin serial port connectors. A PC's parallel printer port uses a 25-pin female D connector; don't attempt to attach an RS-232-C device via an adapter!

In the PC's early days, manufacturers and hobbyists sometimes played fast and loose with RS-232-C connectors, occasionally wiring PCs as DCE devices or as DTE devices fitted with the wrong connectors. Nowadays, serial ports enjoy greater standardization, however, some manufacturers use uncommitted pins for applying power to specialized peripherals such as optical scanners.

Table 1 reveals that most of RS-232-C's signals go unused, at least in the IBM PC-compatible universe. A PC operates as a DTE device, while modems, mouses, and other serial periph-

Signal Function	RS-232-C Pin Pin Number	IBM PC-XT DB-25M Pin Number	IBM PC-AT DB-9M Pin Number
Protective Ground	1	1	n/a
Transmitted Data	2	2 (output)	3 (output)
Received Data	3	3 (input)	2 (input)
Request to Send	4	4 (output)	7 (output)
Clear to Send	5	5 (input)	8 (input)
Data Set Ready	6	6 (input)	6 (input)
Signal Ground	7	7	5
Received Line Signal Detector (Carrier Detect)	8	8 (input)	1 (input)
Reserved	9	n/a	n/a
Reserved	10	n/a	n/a
Unassigned	11	n/a	n/a
Secondary Received Line Signal Detector	12	n/a	n/a
Secondary Clear to Send	13	n/a	n/a
Secondary Transmitted Data	14	n/a	n/a
Transmission Element Signal Timing	15	n/a	n/a
Secondary Received Data	16	n/a	n/a
Receiver Element Signal Timing	17	n/a	n/a
Unassigned	18	n/a	n/a
Secondary Request to Send	19	n/a	n/a
Data Terminal Ready	20	20 (output)	4 (output)
Signal Quality Indicator	21	n/a	n/a
Ring Indicator	22	22 (input)	9 (input)
Data Signal Rate Selector	23	n/a	n/a
Transmit Signal Element Timing	24	n/a	n/a
Unassigned	25	n/a	n/a

Notes:
1. n/a refers to signals that are not available.
2. Input and output directions describe signal flow to and from the PC (DTE).

Table 1. This table contains the classic definitions of RS-232-C signals and their implementation on the original IBM PC's 25-pin DB-25M (male) connector and the more recent nine-pin DB-9M (male) connector as used on most PC/AT compatibles and clones.

erals are configured as DCE devices. Signal names are usually abbreviated (e.g., Data Terminal Ready to DTR).

A straightforward modem-to-PC connection uses a nine-wire cable terminated in a DB-25M (DCE connector) on one end and a DB-25F on the other (which plugs into the DTE PC). The cable is wired straight-through; that is, the DB-25M's pin 2 connects to the DB-25F's 2, pin 3 connects to 3, and so forth. Most RS-232-C cables are shielded, and the shield connects to pin 1 in both connectors. Note that the original

Function	DB-25M or DB-25F Pin	DB-25M or DB-25F Pin	Function
Protective Ground	1	1	Protective Ground
Transmitted Data	2 (output)	3 (input)	Received Data
Received Data	3 (input)	2 (output)	Transmitted Data
Request to Send	4 (output)	5 (input)	Clear To Send
Clear to Send	5 (input)	4 (output)	Request To Send
Data Terminal Ready	20 (output)	8 (input)	Carrier Detect
		6 (input)	Data Set Ready
Signal Ground	7	7	Signal Ground
Carrier Detect	8 (input)	20 (output)	Data Terminal Ready
Data Set Ready		6 (input)	

Table 2. Wiring for a null-modem RS-232-C cable that's used for connecting two DTE or two DCE devices together. Note that pins 6 and 8 are connected together at each connector and to pin 20 of the opposing connector.

RS-232-C standard defines separate signal and protective grounds, while most PCs tie both to chassis ground.

Although RS-232-C specifically defines only DTE to DCE connections, you can connect two DTE or two DCE devices together via a crossover or null-modem cable or adapter that's equipped with two male or female connectors and wired as shown in **Table 2**. You can easily design a similar null-modem cable for nine-pin ports.

A fully implemented PC-to-modem interface uses all of the signals listed in **Table 2**, but you can build a functional RS-232-C interface using only lines 2 (Transmit Data), 3 (Receive Data) and 7 (signal ground). Instead of using hardware handshake signals (e.g., Data Terminal Ready and Request To Send) to notify the host that it's okay to send additional data, you design your software to recognize status characters inserted in the data stream.

Referred to as Device Codes 1 and 3, or X-On and X-Off, these nonprintable status characters occupy low addresses in the ASCII character chart. Upon receipt of an X-Off character, the host suspends data transmission until the receiving peripheral sends an X-On character, signifying that it's ready to continue.

Physical issues

You can purchase DB-25 connectors with solderable, individually crimped or mass-terminated pins. No one technique works best. Of course, manufacturers can afford specialized crimping tools, but the rest of us must use what's readily available—solderable connectors that work best with high-temperature wire whose insulation won't turn to mush when heated. Slide small pieces of heat-shrinkable tubing over individual conductors to insulate the soldered connections.

Not all connectors withstand soldering heat well—translucent nylon melts easily, allowing pins to wander unless you use a mating connector as a fixture. Look for connectors made with blue, heat-resistant diallyl phthalate insulators or other hard, thermosetting plastics.

Individually crimped connectors don't damage wire insulation, but applying too-vigorous pressure with a hand crimping tool can bend the pins. Also, removal of an erroneously inserted pin from a connector body requires a special tool and a measure of skill. Extracted pins may break when reinserted.

Mass-terminated RS-232-C connectors work well with ribbon cable for wiring inside an enclosure. While shielded ribbon cable is available, it's not commonly found as surplus, and unshielded ribbon cable cheerfully radiates interference to nearby receivers.

A variety of commercially-assembled 9-to-25 pin adapters can help you match cables and devices to your PC's serial ports. Custom-wired adapters sometimes appear as surplus. Perform a continuity check on any unlabeled adapter before putting it to work.

Electrical issues

An RS-232-C signal assumes one of two states—a logic 1 (a mark, in teleprinter parlance) consisting of a voltage between –3 and –25 volts, or a logic 0 (space) consisting of a voltage between +3 and +25 volts. A 50-volt one-zero swing no doubt strikes today's circuit designers, accustomed to noise margins of a volt or two, as an extravagance. However, RS-

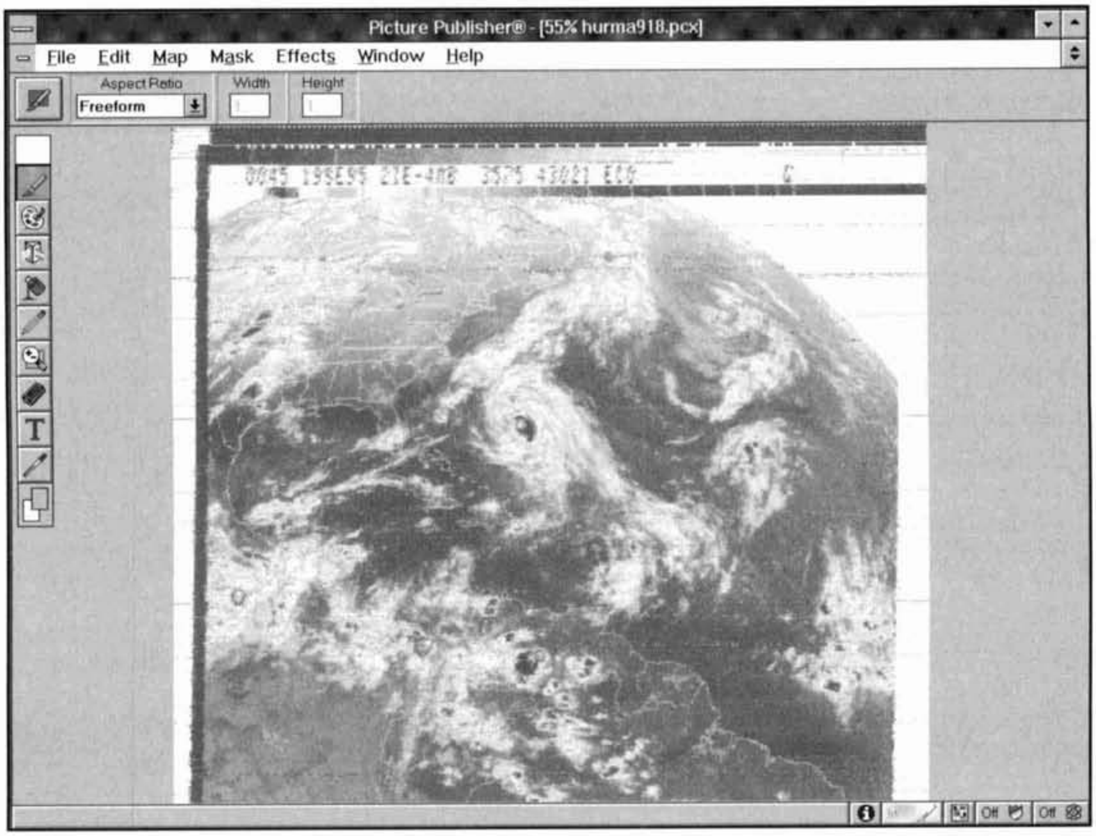

Figure 1. Received from station NAM in Cutler, Maine, this annotated infrared scan shows Hurricane Marilyn as of September 18, 1995.

232-C's ancestral teleprinters included noisy motors and spike-generating solenoids.

On the receiving side, an RS-232-C input circuit presents a resistance of from 3,000 to 7,000 ohms and an effective shunt capacitance of 2,500 pF or less. The latter value sets an upper limit on input plus shunt cable capacitances. While the standard recommends cables of 50 feet or 15 meters, you can extend the distance by slowing the data rate, using lower-capacitance cable, or using driver and receiver circuits that handle ±25-volt power supplies.

In addition, RS-232-C imposes rate-of-change limits—a 1 ms maximum transition time for a one-to-zero or zero-to-one transition, and a 30 volt-per-μs maximum rate of change. The standard applies for data-signalling rates from zero to a maximum 20,000 bits per second (or 20 kBPS—19.6 kBPS is the closest commonly used rate).

In practice, RS-232-C's 25-volt limit is rarely used—most serial ports obtain +12 and –12 volts from a PC's internal power supply. Using low-capacitance coaxial cable and data rates of 4800 BPS or under can extend the range to hundreds of feet.

Nowhere does the RS-232-C standard spell out data formats. You can use synchronous (clocked) or asynchronous (start-stop) modes, and the latter can assume any combination of start, data, parity, and stop bits.

In asynchronous mode, a character or burst of data can arrive at any time. Thus, each character begins with a low-going start pulse, followed by a number of data pulses that can assume either a low or high state. A parity bit (if used) follows, taking on a high or low state depending upon the value of the previous data. Finally, one or more logic-high stop bits terminate the character.

After World War II, radio amateurs adapted surplus Baudot-coded teleprinters that presented a current-loop electrical interface. Limited to five data bits or 32 binary words, the Baudot code covered the 26-letter alphabet, ten digits, and assorted punctuation characters by adding shift and unshift characters to reposition a teleprinter's type bars.

Baudot's 32-character base and time-consuming shift/unshift operations were inadequate for high-speed serial printers, and designers turned to other coding schemes, including EBCDIC (IBM's binary-coded decimal system) and USASCII (USA Standard Code for Information Interchange), or ASCII for short.

Most PCs and modern amateur radio systems use ASCII codes consisting of one start bit, eight data bits, no parity bit, and one stop bit.

This is often referred to as 8-N-1 format. Data rates run from 300 to 57,600 bits per second. A 7-N-2 format—one stop bit, seven data bits, no parity and two stop bits—is also used.

Putting the port to work

Beyond RS-232-C's obvious applications, a host of interesting and somewhat offbeat products take advantage of its strengths: low cost, uncomplicated data protocols, and good noise immunity.

Its electronics packaged in a slightly oversize RS-232-C connector shell, the $129 PC HF Fax Plus system from Software Systems Consultants of San Clemente, California provides radio amateurs and SWLs interested in copying weather-facsimile (WEFAX) images, radioteletype (RTTY) traffic, and Morse code with a compact and inexpensive hardware/software combination demodulator.

While you can receive satellite images directly via VHF and microwave frequencies, the U.S. Navy, Coast Guard, and agencies of other governments rebroadcast satellite video and infrared scans on selected frequencies in the HF bands. To receive these, you'll need an HF receiver that's equipped with either a beat frequency oscillator (BFO) or upper and lower sideband modes. Garden-variety AM-only shortwave receivers won't work.

For best results, use a balanced dipole antenna fed with coaxial cable or balanced line driving a balun to match your receiver's input impedance. Route the antenna feedline as far away from your PC and other noise sources as possible.

To begin, you plug the PC HF Fax demodulator into a spare RS-232-C serial port on your PC and install the software, which runs under MS-DOS 2.1 or higher versions and requires 640 KB of memory. You can also run the software under Windows 3.x in standard mode, or under Windows' enhanced mode if your PC includes a 486DX-2/66 MHz or higher processor.

The fax software can display gray-scale images, provided a VGA or higher graphics adapter is present in your PC. The Telex/RTTY/CW software runs only under DOS or standard-mode Windows.

If you're unfamiliar with the sound of HF fax and RTTY signals, you'll appreciate the tape cassette that's included as part of the package. One side provides an HF fax tutorial, while the other covers various RTTY, Telex, and CW modes which you can use to test your demodulator. An inch-thick manual provides detailed instructions for both the fax and Telex software. As a bonus, lists of frequencies and fax schedules provide a head start on finding interesting signals to copy.

Other accessories include a quick-reference card, a 9-to-25 pin serial-port adapter cable and three floppy disks containing the software and VESA video drivers for a generous assortment of VGA display hardware.

I used a Drake R8 communications receiver and a 200-foot ladder-line-fed dipole to tune the U.S. Navy's fax transmissions from Cutler, Maine. I received my first weather map within a half hour of opening the package—an interval prolonged by a search for an adapter to match the demodulator's 1/8-inch phone plug to the Drake's RCA-style audio line-output jack.

Figure 1 shows a view of Hurricane Marilyn as captured on September 18, 1995. A combination of local QRM and late-summer thunderstorm QRN added streaks and dashes to the image. The HF fax software includes several useful tools that alter image brightness and contrast, and translate a gray-scale image into false color, black/white, or blue/gray charts.

Other handy tools magnify a selected portion of the image, and display a histogram showing the distribution of image pixels versus brightness. You can examine the intensity of single pixels and straighten distorted images or align an image's display margin. A print mode produces gray-scale or line drawings on your PC's printer (a wide range of printers are directly supported or emulated).

There's more: you can control an RS-232-equipped receiver from a second serial port to schedule reception of faxes from several stations. Both fax and Telex software feature handy tuning-oscilloscope displays of signal thresholds and logic levels.

In Telex/CW mode, the software achieved near-perfect copy on W1AW's 10 dB-over-S9 35-wpm code transmissions, while it encountered some difficulties with weaker signals or those under moderate QRM. Teletype reception also required a solid signal for best results.

The demodulator draws DC power from the RS-232-C connector, an unorthodox use of a PC's serial port. My curiosity aroused by the absence of a schematic or hardware description, I probed the demodulator's pins, discovering that the software sets pin 4 (Request to Send) high and pin 20 (Data Terminal Ready) low to provide a few mA at ±12 volts.

Recovered FSK image or RTTY data emerges from the demodulator via pin 6 (Data Set Ready), and a 22-ms square-wave clock appears at pin 5 (Clear To Send). Clearly, wizard-class software does most of the work, as not much hardware can fit into an RS-232-C connector housing. My guess is that the demodulator contains a micropower phase-locked loop circuit.

Note that the fax and Telex software require separate setup procedures from within each

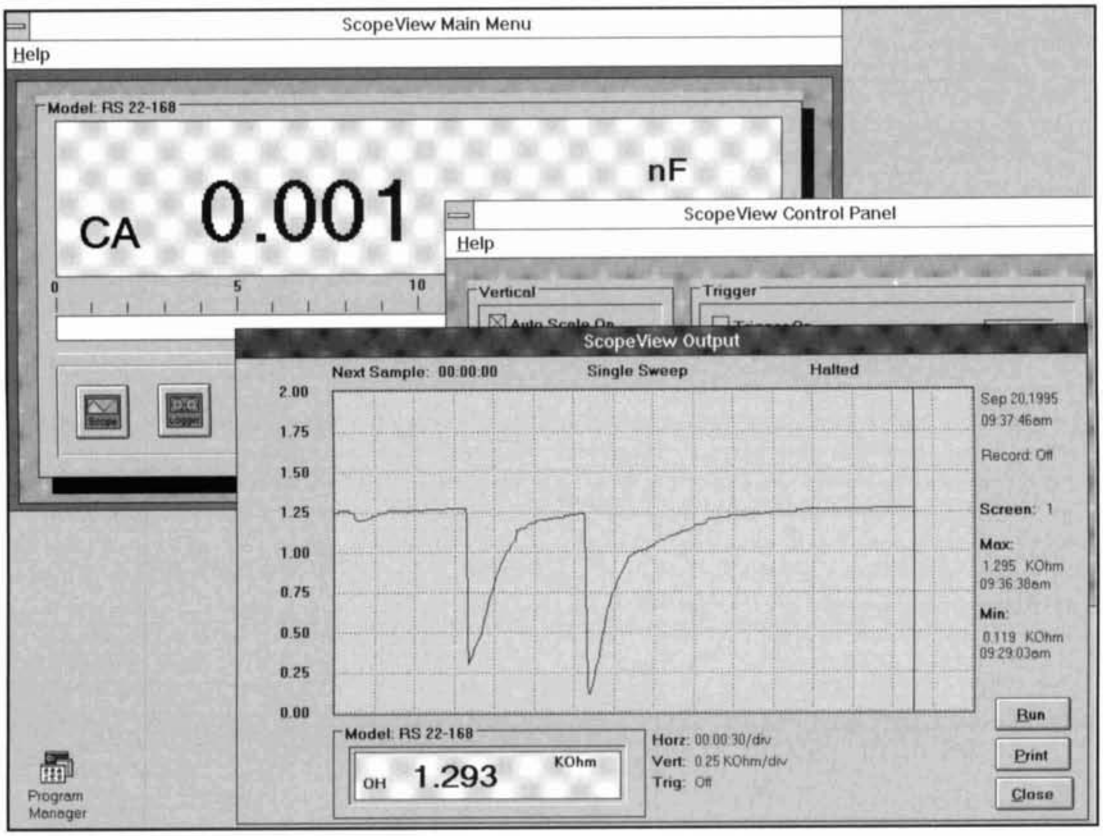

Figure 2. Radio Shack's Model 22-168 Windows software displays a 1-k NTC thermistor's resistance versus time—a match held nearby produced the sharp resistance decreases. An earlier capacitance reading appears in the main menu window.

program, a detail which I interpreted as a bug. I left a message for tech support on Software Systems Consulting's BBS and received a prompt reply to my question.

In general, you must read the manual carefully to appreciate all of the software's features—a task made difficult by the manual's too-tight plastic comb-style loose-leaf binding. While crowded and noisy HF bands may eventually whet your appetite for clearer images downloaded via VHF or UHF satellite services, the PC HF Fax Plus package will help you explore alternative communications modes at a modest cost.

Low-cost data acquisition

The advent of PCs has sparked an explosion of new applications and low-cost hardware/software products for data acquisition. One example of this is Radio Shack's Model 22-168 LCD Digital Multimeter priced at $129.99.

The handheld Model 22-168 contains a serial interface that works with an IBM-compatible PC's serial port. Software running under DOS or Windows displays and captures data to files on your PC's hard disk drive.

Running at a fixed 1200-BPS rate, the data link sends characters consisting of seven data bits, no parity, and two stop bits. A data frame consists of 14 alphanumeric characters terminated by a carriage return (CR) character.

The meter's liquid-crystal display measures a generous 2-5/8 by 1-1/2 inches and includes an analog bargraph and a slew of annunciators; digits measure 5/8 inches high by 5/16 inches wide and offer acceptable off-angle visibility. A built-in stand lets you position the meter at an optimal viewing angle.

A total of 38 measurement ranges include DC and AC voltage and current, resistance (to 2 gigaohms), capacitance (to 200 μF), frequency (to 20 MHz), diode checking, high/low logic levels, audible continuity and bipolar transistor hFE (current gain). A dual-mode capability simultaneously displays AC voltage and frequency, a useful feature for checking power supply ripple.

Controls consist of six push buttons for power and functions, and a rotary range selector switch. The function pushbutton provides a one-way toggle through ten menu choices. Connectors include four recessed banana jacks for test leads, two side-wiping sockets for capacitor leads, and an 8-pin socket for transistor leads.

The Model 22-168 sports interesting features:

auto-hold traps a four-second-old reading to an inset display, and data-hold traps an operator-selected reading to the display. You can view maximum and minimum readings, and enter a reference value to compare against a reading. You can store and recall up to ten measured values in memory.

A 44-page instruction booklet that lacks a schematic diagram covers the instrument's ranges, accuracy and functions. DC voltage ranges provide the best (±1.5 per cent) overall accuracy—a realistic value for a 3 1/2-digit instrument. Other ranges offer less accuracy. For example, the frequency range specifies only ±4.5 per cent overall accuracy—not adequate for precision RF work but still useful.

Safety features include two internal fuses, and test probes with textured grips and finger-stop rings. Measurement limits are 1000 volts DC, 750 volts AC, and 20 amperes (AC or DC). Of more importance, optocoupler integrated circuits isolate the RS-232-C data link from the measurement circuitry. The manual doesn't specify meter-to-computer breakdown voltage, but the instrument carries an "Underwriter's Laboratory Listed Tester 1T10" label—an assurance that it's safe to use within the measurement limits.

Internally, the Model 22-168 uses a mixture of through-hole and surface-mounted components. The range switch contacts consist of gold-plated PC-board traces, a common design choice in low-cost instruments. The heads of two PC-board retaining screws come perilously close to PC board traces and pads, but addition of a pair of plastic washers would preclude potential trace cut-through and short circuits.

In operation, the Model 22-168 draws approximately 70 mW from a standard 9-volt alkaline battery. A two-second delay between making contact and hearing the buzzer sound mars the audible continuity-test mode—that's too long for efficient testing.

The Model 22-168's software deserves a column in itself—versions for DOS and Windows are included. Both provide sampling rates best suited to slowly-changing phenomena such as power-supply drift or troubleshooting intermittents where outages last at least ten seconds. **Figure 2** shows a thermistor's changing resistance as a heat source is applied twice—the text at the right side lists maximum and minimum readings and their times of occurrence.

Both packages acquire data in ASCII format, simplifying export to other programs (although you may have to edit header and label text to meet their requirements). Of the two versions, I preferred the more-polished Windows software although both deserve documentation beyond their on-line help files. Also, to simplify component characterization, the meter could use a single-point data-acquisition mode in which pressing a PC key traps a measurement.

Radio Shack's Model 22-168 multimeter and software offer radio amateurs an inexpensive and self-contained data-acquisition system that opens paths to interesting applications (e.g., logging SWR while tweaking an antenna). Between the Model 22-168 and Software Systems Consulting's PC HF Fax Plus package, you just may find that your PC is running out of available serial ports—but that's a topic for another column!

Purchasing information

To purchase PC HF Fax Plus, contact Software System Consulting, 615 S.El Camino Real, San Clemente, California 92672, (714) 498-5784 (voice) or (714) 498-0568 (fax).

You can buy the Model 22-168 LCD Digital Multimeter from local Radio Shack stores, or contact Radio Shack, Tandy Corp., 700 One Tandy Plaza, Fort Worth, TX 76102. ■

PRODUCT INFORMATION

Kantronics' KPC-3 Now GPS Compatible

Kantronics' KPC-3 TNC now has GPS capabilities. To receive and re-transmit GPS (global positioning system) data, the KPC-3 connects to GPS receivers with NMEA-0183 interfaces. Standard features in the KPC-3 version 6.0 include: Multiple string parsing. Users select as many as four of the GPS unit's NMEA data strings. Storage of outgoing data in tracking buffers. GPS data can be stored for later retrieval and is accessible via the KPC-3's mailbox. Time-slotted location broadcasting based on the GPS clock. Users specify beacon start time and amount of time between beacons, so multiple stations report without collision. Remote access. System operator can reconfigure the GPS unit from a remote location; and the KPC-3 version 6.0 is also APRS compatible.

For versions prior to 6.0, Kantronics offers an EPROM upgrade. Contact your local authorized Kantronics dealer or Kantronics for more information. Kantronics, 1202 E. 23rd Street, Lawrence, KS 66046; phone 913-842-7745; fax 913-842-2021; BBS 913-842-4678

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INSTRUMENTS FOR ANTENNA DEVELOPMENT AND MAINTENANCE: Part 3

SWR and other precision measurements

For too many hams, the only antenna measurement they make is that of standing wave ratio (SWR). Many tales of poor antenna performance have been traced to a concentration on SWR and neglect of other measurements. Use the SWR measurement for what it should be: the measurement of conditions between a properly working antenna and a properly working transmitter or receiver. In this connection, don't forget that the SWR of an ideal dummy load is 1:1—but it doesn't radiate very well. Once again for emphasis, the job of an antenna is to radiate, not to have low SWR. The matching unit is the SWR control, whether at the antenna or at the shack.

Of course, SWR is of some importance. Coax attenuation does increase as SWR goes up; however, this is usually important only on VHF and above. More important is the possibility of puncture at high power, or of a short due to heating and softening of the dielectric at a high current point. Pay attention to line ratings when setting SWR goals. Problems like an unhappy solid state transmitter, RF on the mike cable, and narrow operating range disappear if an antenna tuner is used. It's easier to use this antenna tuner if a SWR meter is permanently in the line.

Early SWR indicators

Just after World War II, a simple SWR indicator appeared, as shown in **Figure 1A**. It works because there is simultaneous magnetic and electric coupling to the transmission line. These combine to separate the forward and reflected waves. When the two lamps are of equal brilliance, the SWR is infinite; when one is out, it is unity. **Figure 1B** shows an unbalanced line version. One of these makes a useful indicator of antenna change if it's left permanently in the line. Such an indicator can be useful in the emergency or field-day kit.

Modern meters

Modern SWR meters use variations of the principles used above. The meter in **Figure 2A** employs the same parallel line structure as in **Figure 1B**, but with separate lines for forward and reflected components. In **Figure 2B**, a RF

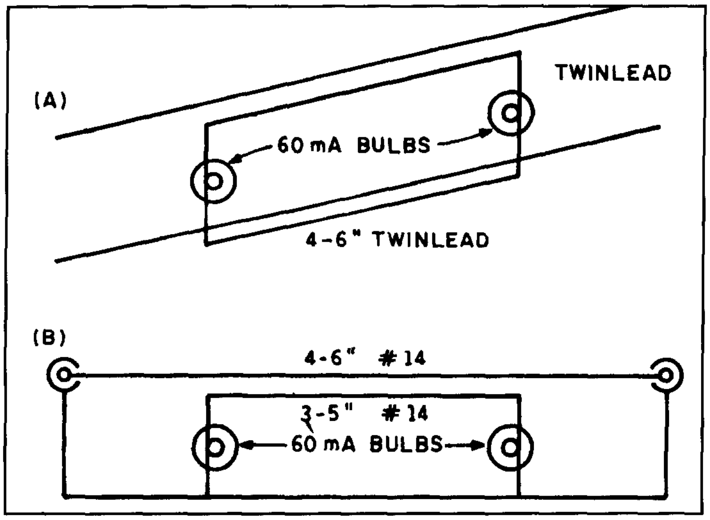

Figure 1. Balanced Line SWR indicator. (A) This post WWII open wire line indicator shows low SWR when the bulb towards the transmitter is bright as compared to the one towards the load. Dimension shown and bulb types suitable for a transmitter of 50–100 watts output. Sensitivity is adjusted by the spacing from the main line. (B) Coax version of (A).

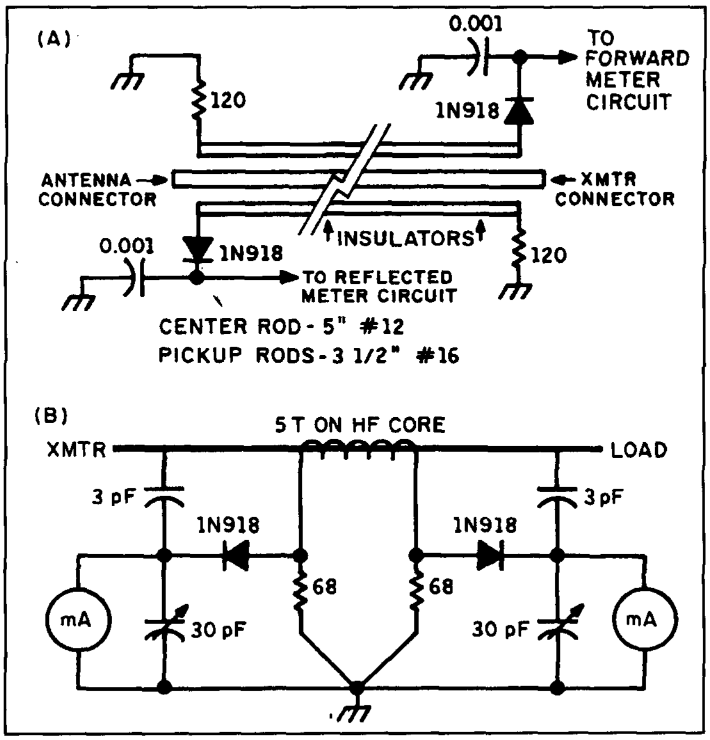

Figure 2. SWR measuring instruments. (A) Transmission line directional coupler, an extension of *Figure 1*. Sensitivity is much higher, due to use of the diode detector. A typical unit made for the CB band is usable for the low power range from about 0.1 to 10 watts. (B) Coupler/ indicator which separates the magnetic and electric field coupling to the line. Careful construction is needed to eliminate the effects of stray coupling, see the ARRL or other handbooks for design details.

transformer is used for the current component, and a capacitor for the voltage. Designs are available for the power range from about 1 watt to many kW, and for frequencies through UHF. Separate units or measuring heads for HF and VHF are best, but most types will correctly indicate 1:1 SWR over a wider frequency range than shown on specification sheets. Check using a matched dummy load. For example, a typical line type intended for CB use provides good results on 144 MHz, and is usable on 220.

Virtually all of the units on the market use a single meter, switched from forward to reflected. The two-meter type is much easier to use. However, they are more expensive, so you might want to add the second meter in a bolt-on box. You can replace the original calibration pot with a dual unit, or add the pot externally. The second meter doesn't need to be large. You can use the built-in meter when measuring the important quantity. Usually, this is reflected power during antenna adjustment, and forward power during station operation. One of these modified units between the match-box and the transmitter, plus calibrated dials on the match-box makes for fast tune-up/operation on multiple bands, over the entire extent of the band.

These SWR meters work on the basis of resistance comparison. If the resistor at the end of the pick-off line in **Figure 2A** is replaced by a capacitor, *the comparison is of reactance* components. Automatic antenna tuners use a form of this device to control the reactance cancellation components of the tuner, plus a normal SWR type to control the feed resistance-setting components. You could build a combination R-X indicator, but there are better techniques available.

Here's a trick to help tune up a Yagi antenna before raising it to the top of the tower. Point the antenna straight up, with the reflector a little above ground. Adjust the matching section so it's as close to 1:1 SWR as you can get. Now when the antenna is installed atop the tower, only touch-up adjustments will be necessary. If you're lucky, the SWR will be acceptable with no further tuning. This method is much easier than attempting to do all of the tune-up atop the tower.

Precision measurements to replace SWR

While it's possible to design an exact matching system from SWR information by using the "cut and try" method, the process is easier if you can measure the impedance. Also, the terminal impedance of the antenna tells a lot about what's going on. Any serious antenna work requires impedance measurement.

SWR alone does provide a measure of the magnitude of the impedance. However, a second measurement is necessary to find the angle of the impedance. This can be the position of voltage minimum (see later), but you may find it inconvenient to obtain. For an easy measurement, add series resistance to the line, **Figure 3A**, and measure the new SWR. The intersection of the two SWR circles on a rectangular or curved (Smith) impedance chart provides two resistance and reactance magnitudes; but, a third measurement is needed to select the correct value of the two possible solutions. This can be either the series capacitor or inductor as shown in **Figure 3B** or **3C**. The third circle on the chart now identifies the correct point. It's easy to calculate a computer solution of this resistance plot geometry. The equivalent shunt elements are also usable.

These series and shunt elements are useful in extending the calibration range of the measurement bridges discussed later. Use the series element for very low impedances, and the shunt for very high ones. Calculate the unknown from series and parallel impedance relations.

For greater accuracy and ease of use, you'll need an impedance bridge or similar device. The basic principle is shown in **Figure 4A**, and a typical series arm bridge in **Figure 4B**. Low-cost versions are made by combining a noise generator with a simple bridge, as "The Noise Bridge." They are available commercially, or you can build them using the designs in the ARRL, RSGB and Radio handbooks. Most bridges are only designed for 50-ohm lines and low reactances. Series elements as in **Figure 4B**, or equivalent shunts, can be used to extend the measuring range. Note that some designs indicate only resistance, depending on the depth of the null to indicate when reactance has been eliminated.

In the precision field, you may occasionally see a General Radio RF bridge for the 0.4 to 60-MHz range at a hamfest. The price tends to be high, because their usefulness is well known. More common, and far more inexpensive, is the HP VHF Bridge. This unit is excellent for scale model work in the 55 to 500-MHz range, and usable down to about 5 MHz (with some problems). The HP RX meter for 0.5 to 500 MHz turns up occasionally, as does the GR UHF Admittance Meter for 20 to 1000 MHz. You may see newer equipment such as network analyzers, vector impedance meters, but these items take more than loose change to buy.

All of these units have the same basic test set-up, as shown in **Figure 4A**. Source and generator may be as discussed above, or may be of one the many special types recommended by the manufacturer. One often finds old HP and GR catalogs at hamfests and used bookstores. These are the best guide to the identification of possible pieces of equipment, and offer many useful hints for use, too. You may also come across instruction and technique books. Fair Radio can supply copies for some

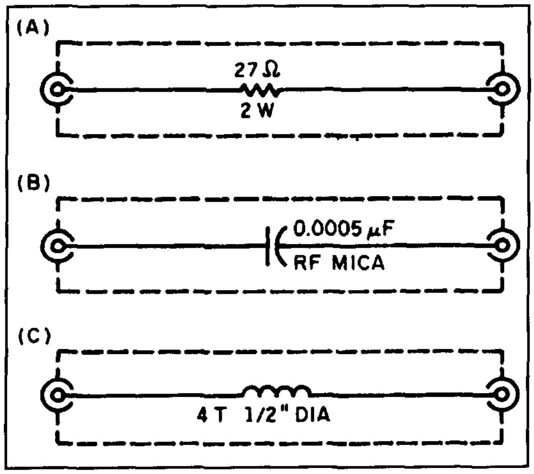

Figure 3. Series Elements for SWR to R,X measurement. Series or parallel resistance and/or reactance can be used to give R and X measurement from 2 or 3 SWR values. Use a rectangular or Smith chart, or computer program, for calculation. Typical values for 14 MHz are shown at (A) resistor, (B) capacitor, (C) inductor. For the parallel

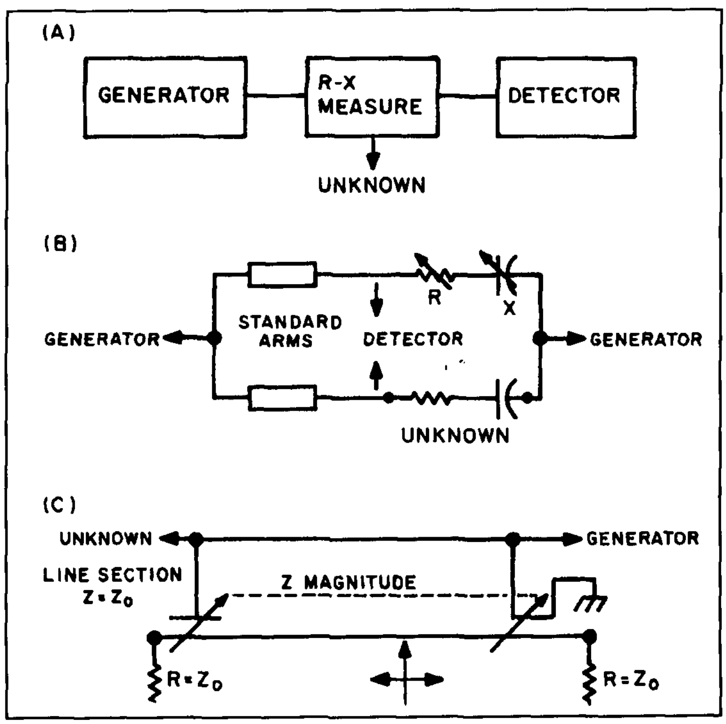

Figure 4. Elements of impedance measurements. (A) Basic measuring setup, of signal generator, bridge and detector. A wide band detector is often used with a single-frequency generator, but a narrow band detector (receiver) must be used if the signal source is wide band, such as a noise generator or a swept frequency oscillator. (B) Basic four arm bridge, usually designed for equality of the upper and lower arms. The standard arms may be a center tapped transformer, or two resistors, or may be more complex. The series adjustable arm type is good for dipoles below resonance, as shown for the series RC unknown element. Parallel arms may also be used, as in the GR RF bridge and some noise bridges. (C) Transmission line bridge. The resistive component of the unknown is determined by adjusting the relative capacitive and inductive coupling to the line ends, and the reactive component by the fractional wavelength departure from line center at balance. This is the principle of the HP RF bridge. Other designs are found in the literature.

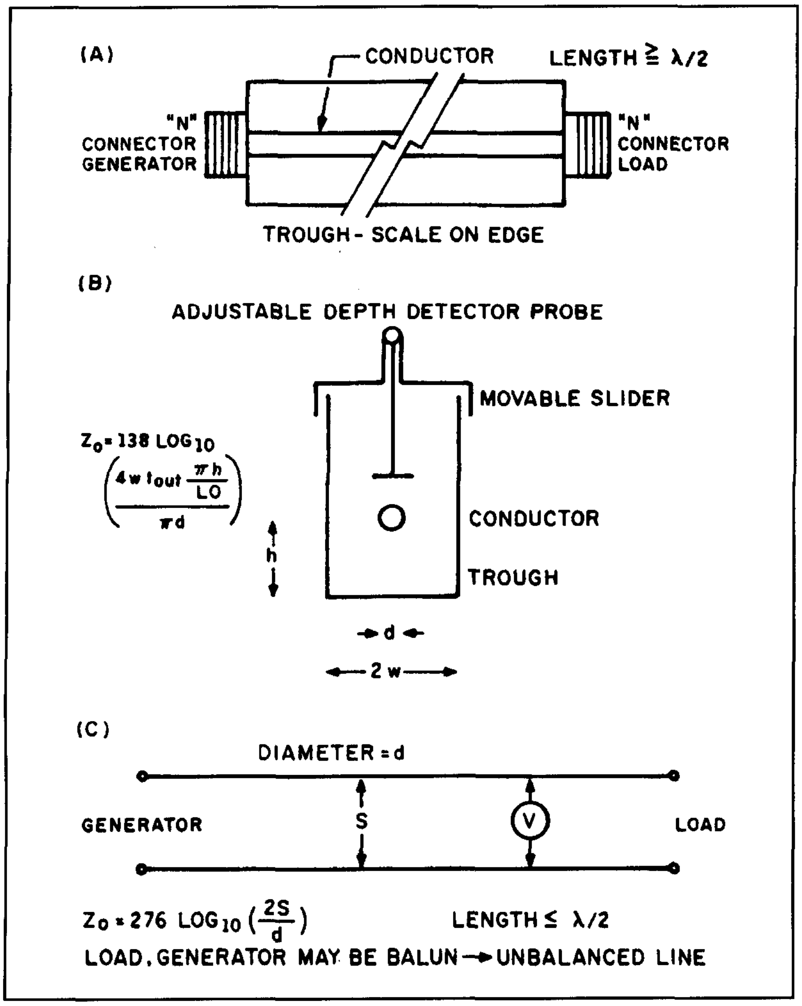

Figure 5. Transmission Lines for Measurement. (A) Top view of a slotted or trough transmission line, usually designed for a Zo of 50 ohms. A probe moves along the open top of the line to give the voltage variation along the line, which gives the SWR. The position of the voltage minimum gives the second measurement needed to calculate R and X. (B) Cross section of a trough line. The probe end is usually smaller than the conductor diameter. Such lines can be built with 1 x 3 inch extruded aluminum, or can be sheet metal folded around a 2 x 4 for forming. The equation gives the conductor diameter and position for the design impedance. A 6 foot length is good for 144 MHz and above. (C) Elements of a Lecher wire system, which may be designed for 270, 300, 450 and 600 ohm impedance. If room is available, useful at HF, for measurement, or for check of the SWR accuracy of another instrument.

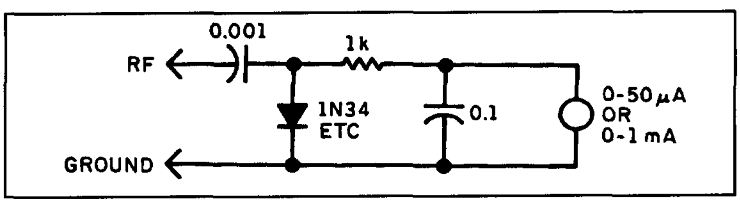

Figure 6. Germanium diode RF voltmeter for low voltage measurements. A IN914 silicon diode will withstand higher voltages, with a small loss in sensitivity.

equipment, and it may be possible to obtain microfiche copies from HP. Call your nearest office for ordering information.

The ARRL Handbook, *The RSGB Handbook*, and *The Radio Handbook* contain descriptions of RF bridges or equivalent measuring equipment for home construction. *Reference Data for Radio Engineers* and Terman's *Radio Engineers Handbook* include comprehensive discussions of theory plus material on practical use.

These devices aren't difficult to use. Setup can be a little tedious. Evaluation of the measuring results used to take some work, but there are now computer programs that produce evaluated results when you punch a few keys. The actual measurements are more fun than a chore.

The slotted line

For some reason, amateurs haven't paid much attention to the slotted line (**Figures 5A** and **B**) or to its open wire analog—Lecher wires (**Figure 5C**). A 3 or 6 foot length of either makes an excellent impedance measuring element on UHF and VHF, and a 20 or 40 foot temporary Lecher wire in the backyard is good for 6-10 and 6-20 meter work. The traveling detectors create the greatest problem, but they can be the simple RF voltmeter of **Figure 6**. The terminals should contact the wire for very low power; but, it's better to have a high enough level to allow capacitance probe coupling to the conductors. The necessary measurements include the ratio of the highest and lowest voltages, and the position of the lowest.

Commercial slotted lines, both coax and waveguide, generators and detector/indicator elements for the VHF through SHF range aren't uncommon at hamfests. These precision devices are really nice for the upper bands.

Older editions of *The ARRL VHF Manual* had a lot of data on this family of devices. *The RSGB VHF-UHF Manual* is very good, and provides complete construction data for a slotted line. *The Radio Handbook* also contains some information.

Calibrated lines

Impedance bridges of any form will give you the impedance at their terminals. But in antenna work, the item of interest is the impedance at the antenna. You must also account for the impedance transforming effect of the transmission line. The very old procedure involved plugging numbers into equations. The Smith chart was invented to get the same answer via fast graphical plotting. Today a computer program solves the equations, and provides both number values and Smith Chart plots.

Any of these solutions take some time, and aren't convenient when you're developing an antenna by cut-and-try. It's easier to have the measurements provide the antenna values directly, which is possible if you remember that the impedance values repeat each half wave of line. The trick is to get the right line length. This can be done by cutting the line to the value PK*LAM/2, where LAM is the wavelength at the test frequency and PK is the propagation constant stated by the manufacturer—usually 0.67 for coax.

Because the propagation constant does vary some from batch to batch, and because the end connectors change the effective length, it's better to simply measure the effective length. To do this, short one end of the line. A small disk with a center hole will work if there's no connector. One of these disks can also be used to short a cable connector receptacle. Connect a small coil to the other end, and measure the resonant frequency with a grid-dip meter (see part 4). For greater accuracy, measure with coils of 1, 2 and 3 turns, and plot the frequency against the number of turns. Project the curve to the zero-turn axis to obtain the true resonant frequency. The line length is 4 times the wavelength; i.e., 4*299.8/f.

The transformation is exact at only one frequency. Assuming that the antenna is near resonance, a small change in frequency has little effect on the resistance measurement. The reactance value change is approximately the impedance of the line multiplied by the fractional wavelength change. If the frequency change is appreciable, use the Smith Chart or computer.

Another way to use line sections of known effective length is to move a SWR minimum point into the range of the measuring device. This lets you use a short slotted line below its normal lower limit. For a 1-meter line, extensions of 1, 2, 3, etc. meters length are needed.

A line section can also be used as a shunt for the antenna terminals, to bring the terminal impedance to a value suited to the measuring instrument. If you use a T connector, remember to compensate for its length.

If you take time to make up any of these lines, label them carefully, and save them for future work. But don't store them in sunlight, in a damp area, or with too tight a coil. Treat them like the precision tools they are.

Next time

In the fourth and final part of this series, I will discuss field strength meters, grid dip oscillators, a few mechanical devices, and the antenna range. ■

PRODUCT INFORMATION

New Line of Standard Transformer Designs From Toroid Corp.

Toroid Corporation has introduced their new line of standard transformer designs with UL and CSA safety approvals for the North American and Overseas markets.

The new line of auto-transformers offers a low cost alternative to full isolation transformers for step-up or step-down applications. Auto-transformers can safely be used when the user's equipment already has a properly designed full isolation transformer installed for the line frequency at which the equipment will operate. Step-down or step-up auto-transformers can also be installed in equipment to drive components like certain types of motors that do not require an isolation transformer.

The size and weight of the toroidal transformer construction are approximately 50% smaller than a conventional, stacked E/I lamination transformer. Using an auto-transformer will save space and money. Toroid offers five standard power ratings: 250VA, 500VA, 750VA and 1,000VA. All five are available for step-down from 230V/50–60 Hz to 120V/60 Hz and vice versa. The 250VA step-up transformer, P/N AZ8 5615 measures 3.9" OD x 1.7" in height, while the dimensions of the 1000VA, P/N AZ8 5618 are 5.5" OD x 2.3" in height. Weights are 2.6 lbs. and 7.1 lbs. respectively. Prices range from $27.90 to $45.00 in lots of 100 pieces. NEMA steel enclosure with or without IEC style receptacles, cords and plugs are available at extra cost.

For more information, contact Christian Ennerfelt, Sales Manager, Toroid Corporation of Maryland, 2020 Northwood Drive, Salisbury, MD 21801-7805; phone 410-860-0300; fax 410-860-0302.

Custom Filters In A Flash

For quick solutions to your low-pass filtering problems, the Shape Shifter Model SL55 from Gatewave provides both design software and reconfigurable hardware. The stepped-impedance tubular filter covers cut-off frequencies from 500 to 5500 MHz with up to 11 sections. Typical insertion loss is 0.2dB at 1 GHz and 1dB at 5 GHz. The specific parameters of the kit's hardware are embedded in the synthesis routines of the software, allowing for design and construction of custom lowpass filters for laboratory use. The software may also be used to analyze user-defined configurations. Macintosh and PC platforms are supported. For more information, contact Gatewave, 565 Science Drive, Madison, WI 53711; phone 800-797-9283; or fax 608-238-5120.

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QUARTERLY REVIEW

The Ten-Tec 1253 9-band SW Receiver

Ten-Tec's recent excursion into the kit arena promises to help fill the void left when Heath company decided to concentrate on educational materials in lieu of their traditional line of electronic kits. Dubbed "T-Kits" by the manufacturer, Ten-Tec's new line is highlighted in their A-95 catalog. I felt a review of Ten-Tec's 1253 9-band receiver kit might be an appropriate compliment to Charles Kitchin's (N1TEV) presentation on regenerative receivers, which also appears in this issue.

Audience

This is a beginner's receiver, and is aimed at novice SWLs or hams on a limited budget. The kit is surprisingly complex, considering its modest price, and should provide novice builders many hours of enjoyment, exposure to a wide variety of different kinds of electronic parts, and a chance to enhance their soldering and assembly skills. The receiver is fun to use, and I suspect the dad or Elmer involved in this project may be caught playing with it when he thinks no one is looking!

For clubs, training classes, and family activities, Ten-Tec also offers a 4-band regenerative receiver for $24—the model 1054. This includes a drilled and silk-screened front panel, but lacks the cabinet, speaker, and knobs. If class size permits, it's possible to save some money by ordering the 1054C CLASS-Pak. It contains five 1054 kits, instructor's notes, and a spare parts kit. (The 1054C is $110 in the Summer 1995 A-95 Ten-Tec kit catalog.) Note that there are some differences between the circuitry used in the 1054 and the 1253 kits reviewed here.

HRO styled tuning dial

Ten-Tec doesn't supply a band-by-band calibrated tuning dial. This is in part due to the confines of limited front panel space for nine bands, and in part due to the use of fixed component values for band coils. Tolerance variations of the varactor diodes and coils would require extensive alignment points for proper dial accuracy and tracking. Instead, Ten-Tec provides a 0 to 22 logging scale over an approximate 270-degree tuning range on the front panel. A sample logging chart in the manual provides a method of logging frequency versus dial indication for each band. This is similar to the system used by National in the early HRO series of receivers during the 1930s and 1940s. The band coverages overlap enough so no frequency gaps should exist at band edges due to component tolerances.

An overview of the 1253

Ten-Tec's model 1253 receiver kit covers from 1.76 to 1.99 MHz (160-meter band), 3.3 to 4.150 MHz (80/75 meters), 5.5 to 6.9 MHz (49 meters), 6.8 to 8.5 MHz (40 meters) 8.5 to 11.0 MHz (31 meters), 10.1 to 13.2 MHz (25 meters), 12.5 to 16 MHz (19, 20, and 21-meter bands) 14.7 to 18.5 MHz (16 and 17 meters), and 18.5 to 21.5 MHz (13 to 15 meters) for a total of nine electronically bandswitched ranges. The detector is regenerative, permitting SSB, CW, and AM demodulation. An internal 3-inch speaker or earphones may be used. Power is supplied by 8 internal alkaline C cells, or optionally from an external DC source.

First impressions

I first contacted Ten-Tec as a consumer looking for advice and information on inexpensive receiver kits. The salesman went into great detail explaining the features and virtues of each kit. He was courteous, knowledgeable, and seemed sincerely committed to the quality of Ten Tec's kit line. He left me with the feel-ing that Ten-Tec was the kind of company I'd like to do business with.

Rugged metal enclosure

I had expected to find the receiver housed in a clamshell metal cabinet with plastic sides. I wasn't prepared for what I found when I unpacked the 1253 kit. Although the cabinet top, bottom, and sides are made of a clamshell steel design, the front and rear panels are made from a U section of aluminum. When put together, the receiver bottom is actually double-wall metal construction! Five control knobs protrude through the front panel. Rather than having unattractive nuts and washers show, Ten-Tec uses a rear aluminum panel to affix these controls; thus, only the knobs show on the front panel. While the speaker could have been mounted to the cabinet top (which is grilled to allow audio to pass through) with simple metal clips, Ten-Tec uses a second aluminum plate with a three-inch punched hole and mounting screws for this purpose. Another custom-made steel plate supports the battery holders inside the case. All exposed metal is painted black; lettering for control markings is white. The hefty enclosure measures 5-3/4 inches wide, by 3-1/2 inches high, by 5-1/2 inches deep. Rubber feet are also included. Ten-Tec also offers a wide line of enclosures for homebuilders and commercial applications.

Getting started: read the manual!

Assembly of the T-Kit was left to my 12-year-old son, who had no prior experience with kits or electronics. I figured this would be a fair test of the manual and the intended audience for the kit. I limited the assembly sessions to 2 hours a day—answering questions and offering advice as the work progressed. In the manual Ten-Tec advises that, while the kit is intended for beginners, the presence of an "Elmer" is suggested to help the builder if problems arise.

One thing I noted was a lack of drawings in the manual. Heath was a master at illustration, and their manuals were liberally sprinkled with drawings of mechanical and electrical details. Despite the lack of artwork, the Ten-Tec manual is otherwise good. It runs some 60 pages, moving beyond the assembly and providing other important information. The manual describes how the radio works, what to use for antennas, the best times of the day to listen, what to listen for on the different bands, and how to use the regeneration control for best reception of different modes—all in a style that expertly relates to a young reader.

Here's an abridged excerpt from the manual: "In theory, your receiver's regeneration control adjusts the level of feedback...In practice this control is like a 'joystick' for optimizing receiver performance. Your ability to handle this 'joystick' saves you many dollars over today's cost of receivers which perform similar functions automatically . . . Once you know how to use it, it's a fun control!"

Caveats

Ten-Tec warns against rushing through the assembly to avoid mistakes. For example, both 3.3-ohm resistors and 3.3-μH molded coils are used in the kit. Both have similar packages and orange-orange-gold color codes. Ten-Tec mentions this in the manual, and also separates the resistors and coils in different plastic zip bags as a precaution. Some care and patience is needed in some areas of mechanical assembly. One such area is where the front panel plastic push-button mates with its matching momentary contact switch on the digital band selector board. This needs to be accurately spaced to prevent binding the switch or excessive slop.

To their credit, Ten-Tec lets the builder stop and actually power up the receiver as certain milestones are reached during assembly. The first test takes place when the audio amplifier stages are completed. At this point, the builder is instructed how to apply power and how to check to see that the circuits he's just finished wiring actually work. This boosts the builder's confidence, and discourages him from rushing through the kit to see if it's really going to work when finished. I like this aspect because I feel the first time kit is an experience that the builder should savor, and not rush through.

The circuit

The RF section contains three J310 JFETS. The first serves as a buffer, isolating the detector from variations in antenna loading, and also helps reduce RF radiation from the regenerative detector. A second JFET is used for the regenerative detector. The third JFET is used as a buffer following the regenerative detector. Bandswitching is provided by pin diode switching of nine different band coils. Molded RF chokes are used here. A CMOS 74HC4017 decade counter is used for band selection. Band selection is accomplished by clocking the 4017 via a front panel push-button switch. A corresponding LED indicates the band selected by the appropriate 4017 output, which also forward biases the appropriate pin diode for the desired inductor for that band.

A 7805 provides voltage regulation for the detector, regen control, and digital logic circuits. The power for the detector and regen control are further isolated and filtered by an MPS6514 pass regulator. Another MPS6514 is used for an AF preamplifier. The final audio amplifier is a 9-pin TDA2611A SIP packaged audio IC for ample audio.

PC boards

The electronics are assembled on two pc boards. One board is used for the digital circuitry involved with the electronic bandswitching; the second pc board is used for the RF and audio sections of the receiver. Both boards are silk-screened with component legends. The digital board is double sided with plate-through holes. The RF/audio board is single sided and while it's not plated, a protective factory coating protects the board from corrosion for ease of soldering during assembly.

There are five variable controls on the front panel: main tuning, fine tuning, RF gain, regeneration, and volume. A power switch and power indicator LED are present, as are nine LEDs for band indication. There's also a front panel push-button for band selection. An internal trimpot is used to roughly set the regeneration point for the receiver, leaving the front panel control for fine adjustments over the regeneration point. This trimpot is the only alignment point in the receiver. Front panel markings are silk-screened in white lettering and markings over the black panel finish.

Final notes

It took about six days for my son, Tom, to finish the kit. As I expected, some time was needed to instruct him on proper soldering techniques, and to keep him from rushing through what he thought were the easier areas of the kit assembly. Some confusion arose over where some of the mechanical parts would be used. For instance, the antenna connector is made up of two cone-shaped insulators and screw hardware that provide for a long wire antenna termination. It took some figuring on his part to ascertain the purpose of some of the items. Pictures and diagrams would have been a plus in the manual to guide Tom over these areas. At one point, the instructions require the builder to prepare the ends of some RG-174 coaxial cable for soldering. The length of RG-174 supplied with the kit had no RG markings, so I had to step in and help. Otherwise, assembly went without a hitch.

The final reward

When the kit was finally deemed finished by dad the "Elmer," the batteries were at last installed and the internal regeneration trimpot was adjusted per the directions in the manual. My long wire antenna was connected, and within several seconds my son had the sounds of several shortwave broadcast stations filling the room. Beaming with pride, he scanned the bands—quickly learning what settings were optimum for the RF gain and regeneration controls. Tom soon mastered the techniques for tuning in CW and SSB signals, when I pointed out which bands were for amateur use. I don't think this is an experience he will ever forget—building his first radio and hearing it work the first time. Hopefully, this will be something he can share with his own son. I hope I'm there to see it! ■

PRODUCT INFORMATION

Philips ECG Expands Software Program

Philips ECG introduced an expanded version of ECG Semiconductors INSTANT CROSS™ Floppy Disk Program, now cross-referencing over 14,000 additional industry part numbers.

The entire database from the recently published 16th Edition ECG Semiconductor Master Replacement Guide (212R) is now available in Version 2.0 of their INSTANT CROSS Software for IBM PCs and compatibles. A helpful feature of the enlarged program is the complete ECG Product Index file selected from the Main menu. Entering an ECG part number will display that number and following numbers plus device description and case style. The Index file may also be scrolled or paged up and down to view other ECG types.

The ECG software operates on IBM-PCs and compatibles that have 640K of RAM, a hard drive and 3 1/2 or 5 1/4 inch floppy disk drive. The program versions that are available will support 360K/1.2M and 720K/1.44M floppy disk drive; also supported are monochrome, CGA, EGA and VGA monitors.

All ECG products and literature are available through authorized Philips ECG distributors. To locate the nearest distributor, consult "Electronic Equipment & Supplies" in the telephone directory yellow pages, or call toll-free, 1-800-526-9354.

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SL-11R	•	•	7	11	2⅝ × 7 × 9¾	12
SL-11S	•	•	7	11	2⅝ × 7⅝ × 9¾	12
SL-11R-RA		•	7	11	4¾ × 7 × 9¾	13

RS-L SERIES

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RS-4L	3	4	3½ × 6⅛ × 7¼	6
RS-5L	4	5	3½ × 6⅛ × 7¼	7

RM SERIES

MODEL RM-35M

• 19" RACK MOUNT POWER SUPPLIES

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
RM-60A	50	55	7 × 19 × 12½	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50
RM-60M	50	55	7 × 19 × 12½	60

RS-A SERIES

MODEL RS-7A

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A		•	2.5	3	3 × 4¾ × 5¾	4
RS-4A	•	•	3	4	3¾ × 6½ × 9	5
RS-5A		•	4	5	3½ × 6⅛ × 7¼	7
RS-7A	•	•	5	7	3¾ × 6½ × 9	9
RS-7B	•	•	5	7	4 × 7½ × 10¾	10
RS-10A	•	•	7.5	10	4 × 7½ × 10¾	11
RS-12A	•	•	9	12	4½ × 8 × 9	13
RS-12B		•	9	12	4 × 7½ × 10¾	13
RS-20A	•	•	16	20	5 × 9 × 10½	18
RS-35A	•	•	25	35	5 × 11 × 11	27
RS-50A	•		37	50	6 × 13¾ × 11	46
RS-70A	•		57	70	6 × 13¾ × 12⅛	48

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46
RS-70M	57	70	6 × 13¾ × 12⅛	48

VS-M AND VRM-M SERIES

MODEL VS-35M

• Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES

MODEL RS-12S

• Built in speaker

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	•	•	5	7	4 × 7½ × 10¾	10
RS-10S	•	•	7.5	10	4 × 7½ × 10¾	12
RS-12S	•	•	9	12	4½ × 8 × 9	13
RS-20S	•	•	16	20	5 × 9 × 10½	18
SL-11S	•	•	7	11	2¾ x 7⅝ x 9¾	12

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

TECHNICAL CONVERSATIONS

(from page 5)

point I nearly gave up. However, after making some make impedance measurements on the basic double-toroid model I was able to construct an appropriate matching network that overcame the matching problem. Once this was done, the antenna performed quite well for such a small device.

I continued work on this antenna (the Mk2 and 3) after the validation work was done and the results were published the following August; I have included a copy for your information. However, Reference 5 in the article published in *Communications Quarterly* gives the impression that the method of validation was unsound because it was performed in a short timescale. It also says that "the impedance of the antenna system was very much better than implied." I didn't imply anything; I measured it and stated it as you can see.

I also enclose a copy of a book that gives details of the test equipment used (page 3.9) to measure the antenna impedance. The measurement was made via a half-wave length of coax with a ferrite choke to reduce antenna currents on the feeder.

A copy of the James Curum European patent on the toroid antenna in the RSGB library shows that it was filed on the 9th of July, 1981.

Peter Dodd, G3LDO
Technical Editor, *Radcom*

CORRECTIONS

Go Figure . . .

Figures and photos presented a bit of a problem in our last issue. Please note the following changes:

In Chip Cohen's (N1IR) article "Fractal Antennas: Part 1" in our summer 1995 issue, readers should note that Figure 12 is Photo A, and Figure 13 is Photo B. Thus, for Figure 14 in the text, see Figure 12, and so on for all later figures. All captions are correct save for Figure 5, which should read: A von Koch fractal for iterations 0, 1, and 2. Readers should also note that aspects of this research are patent pending.

Photo A on page 99 of Rick Littlefield's (K1BQT) "Tech Notes" piece "A Featherweight 6-meter Beam" (Summer 1995) showed a single featherweight Yagi rather than a stacked array. Somehow the wrong photo got into the works. Sorry Rick!

Brad Thompson's (AA1IP) "Quarterly Computing: Building a PC Toolkit" was also a victim of the artwork imp. The captions for Figures 1 and 2 of his summer 1995 article were switched. To view everything in the correct perspective, simply read the caption for Figure 2 when looking at Figure 1, and vice versa.

Call Me!

Dave Hershberger, W9GR, "A Few Words About DSP," Summer 1995, send along his new e-mail address and phone/fax number. Dave writes:

I just got my summer CommQuart, and just as it hit the mailboxes, my internet e-mail address changed! What was published with my article was "dlh@gvgdsd.gvg.tek.com." People sending e-mail to that address will have their mail bounced back to them. Here is my NEW e-mail address:

w9gr@psyber.com

I also have a new phone line and fax machine. My new daytime phone/fax is 916-265-6438. ■

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EZNEC ("Easy-NEC") captures the power of the NEC-2 calculating engine while offering the same friendly, easy-to-use operation that made ELNEC famous. **EZNEC** lets you analyze nearly any kind of antenna - including quads, long Yagis, and antennas within inches of the ground - *in its actual operating environment*. Press a key and see its pattern. Another, its gain, beamwidth, and front/back ratio. See the SWR, feedpoint impedance, a 3-D view of the antenna, and much, much more. With 500 segment capability, you can model extremely complex antennas and their surroundings. Includes true current source and transmission line models. Requires 80386 or higher **with coprocessor**, 486DX, or Pentium; 2Mb available extended RAM; and EGA/VGA/SVGA graphics.

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Both programs support Epson-compatible dot-matrix, and HP-compatible laser and ink jet printers.

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PRODUCT INFORMATION

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For information about the ADG608 and ADG609, contact Analog Devices, Inc., 181 Ballardvale Street, Wilmington, MA 01887; or call 617-937-1428; or fax: 617-821-4273.

2nd Edition of HPs *The Electronic Instrument Handbook* In Second Printing

Hewlett-Packard Company announced that the second edition of *The Electronic Instru-ment Handbook*, which sold out after six months on bookstore shelves, now is in its second printing. This handbook was first published by McGraw Hill in 1973, and has been updated to include the latest technological advancements.

The 900-page book, edited by Clyde F. Coombs, Jr., a retired HP manufacturing manager, and written by HP engineers and scientists and by engineering professors is a reference used by engineers, students and high-tech writers. It covers everything from the basics of electronic instruments to measurements to the role of software in "virtual" instruments.

The handbook is endorsed by the Electronics Book Club and Electronic Engineers Book Club. It is available at college and technical bookstores and at some larger superstores for $79.50. *The Electronic Instrument Handbook* can also be ordered direct from McGraw Hill by calling them at 1-800-722-4726.

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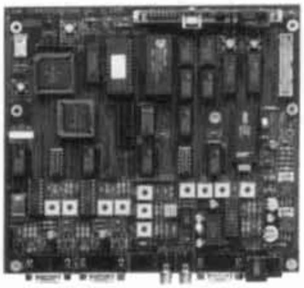

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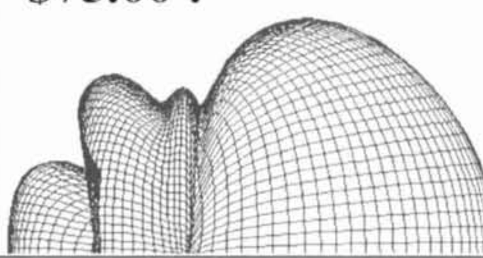

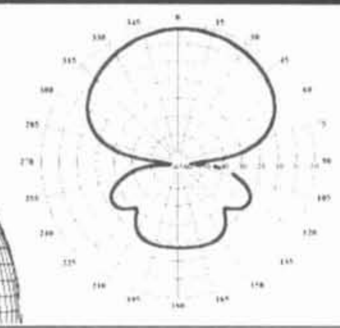

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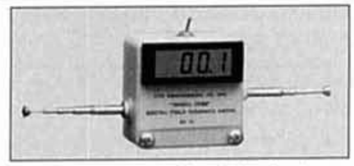

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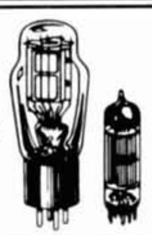

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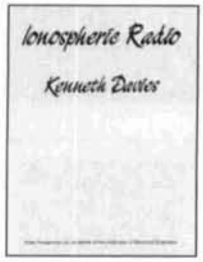

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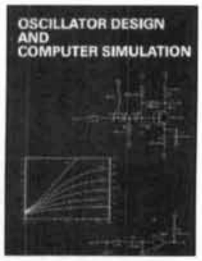

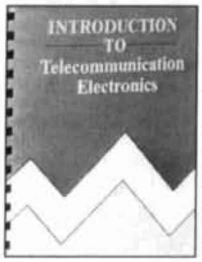

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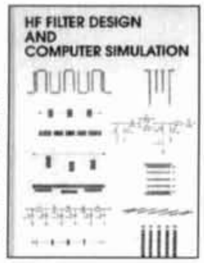

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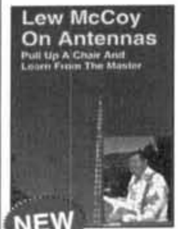

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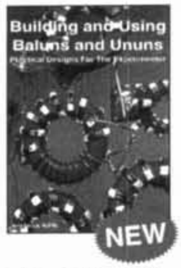

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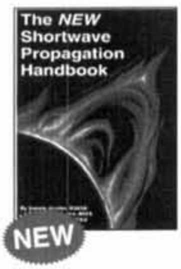

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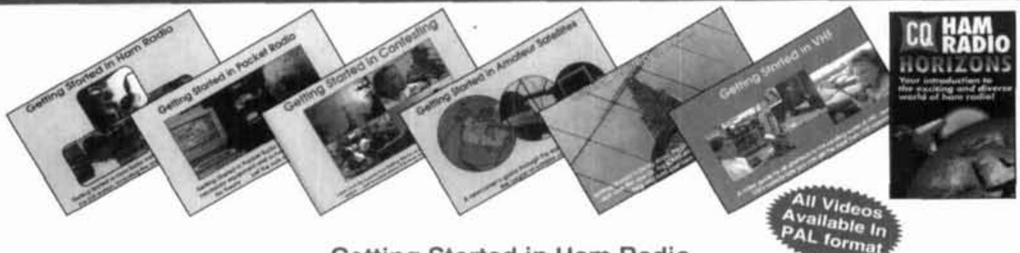

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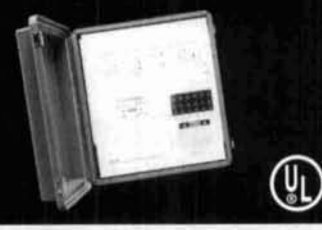

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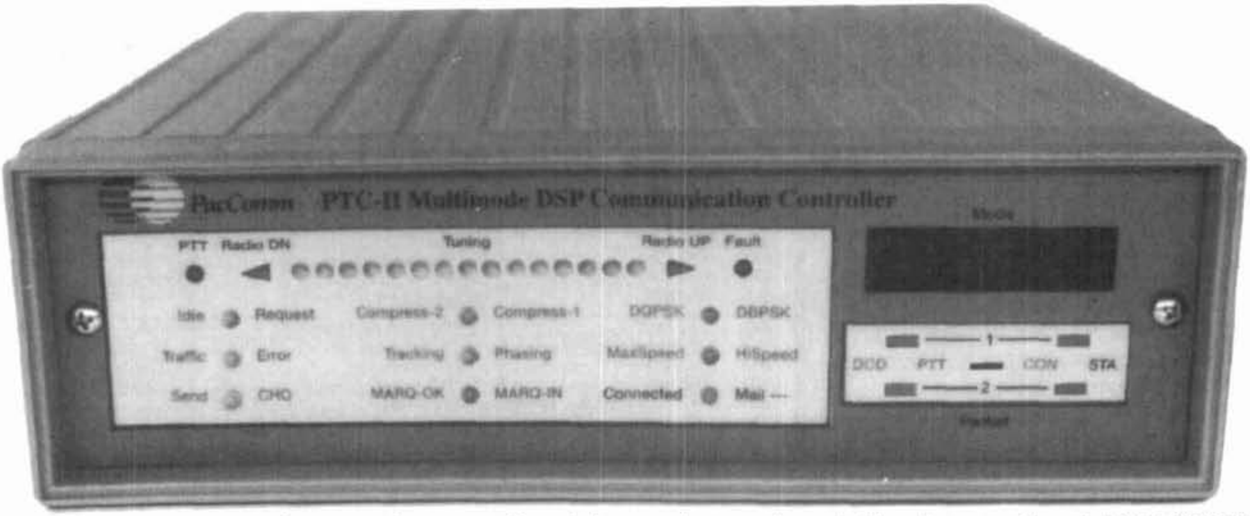

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The PTC-II is a new multi-mode controller and "communications platform" which contains very powerful and flexible hardware and firmware.

It is being built in the United States by PacComm under license from S.C.S., the group that developed both the original PACTOR and PACTOR-II.

The PTC-II offers the most robust HF digital protocol available to radio amateurs, but it should not be overlooked that the PTC-II is configurable as a triple-port multimode controller supporting packet data rates of 1200 and 9600 bps and numerous other modes.

What is PACTOR-II?

Like PACTOR-I, PACTOR-II is a communication system. There is a unique protocol, carefully designed and optimized for excellent HF performance, new and powerful modems realized via Digital Signal Processing, and a hardware platform to allow the firmware to realize its full operating potential.

PACTOR-II is the name of the new protocol including modulation specifications. PTC-II is the name of this specific hardware realization of the PACTOR-II architecture.

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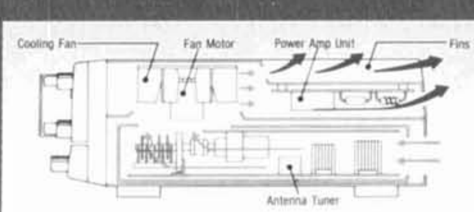

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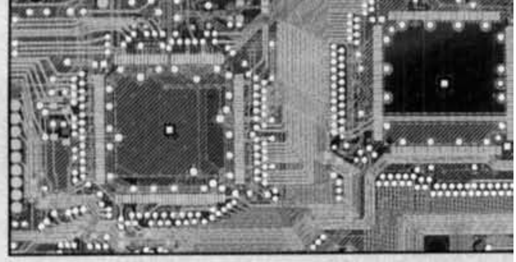

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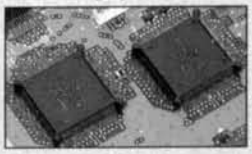

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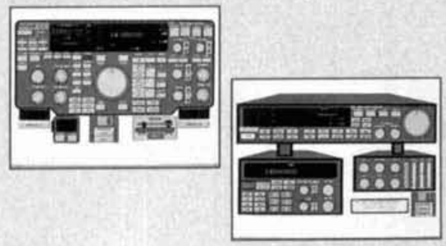

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