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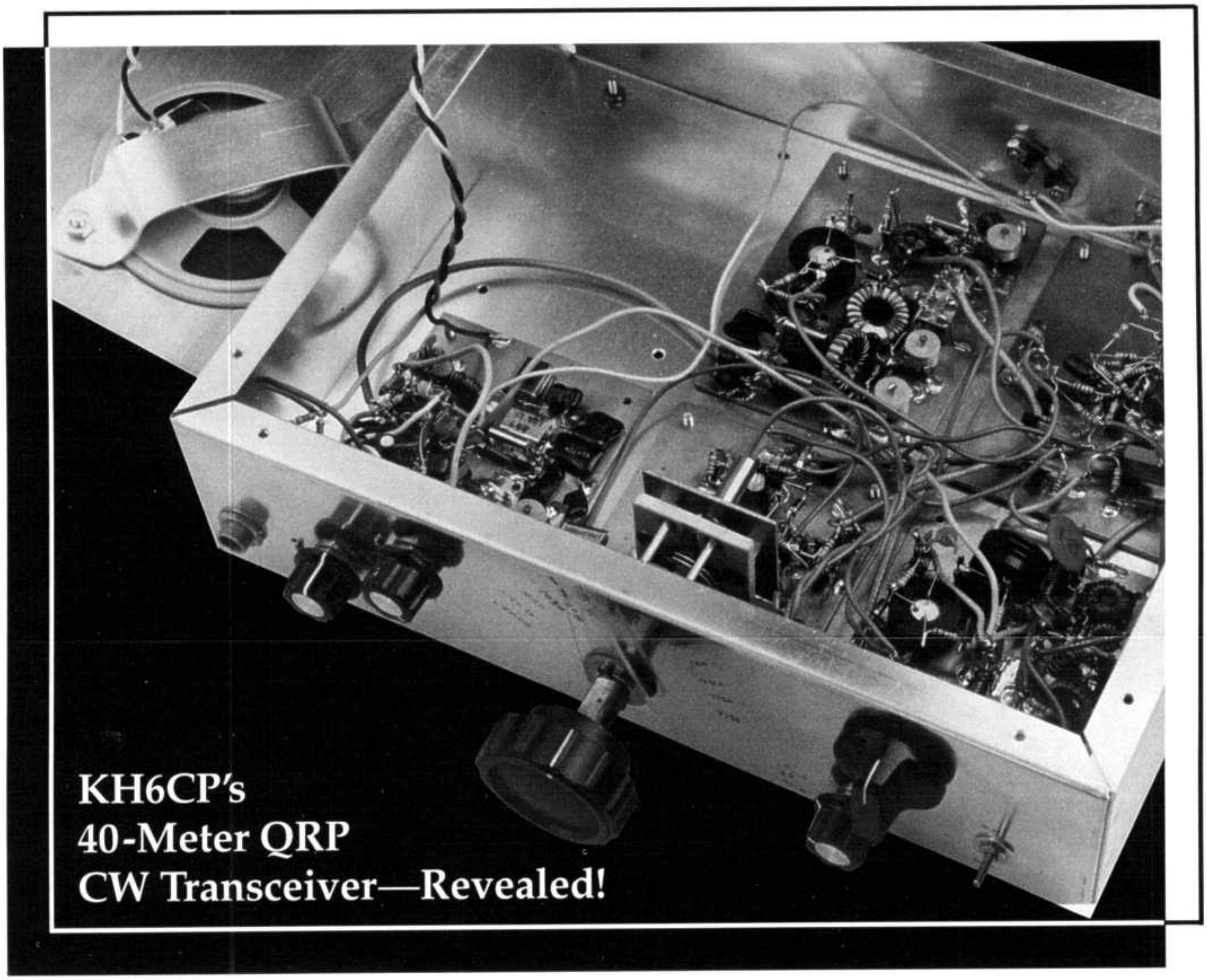
March 1993



KH6CP's
40-Meter QRP
CW Transceiver—Revealed!

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About the Cover:

No, it's not dead bugs, it's dead-bug construction—of a 40-meter QRP CW transceiver from KH6CP. *(Photo by Kirk Kleinschmidt, NTØZ)*

133

TABLE OF CONTENTS

COLUMNS

MARCH 1993 QEX ADVERTISING INDEX

Purpose of *QEX*:

1) provide a medium for the exchange of ideas and information between Amateur Radio experimenters

2) document advanced technical work in the Amateur Radio field

3) support efforts to advance the state of the Amateur Radio art

All correspondence concerning *QEX* should be addressed to the American Radio Relay League, 225 Main Street, Newington, CT 06111 USA. Envelopes containing manuscripts and correspondence for publication in *QEX* should be marked: Editor, *QEX*.

Both theoretical and practical technical articles are welcomed. Manuscripts should be typed and doubled spaced. Please use the standard ARRL abbreviations found in recent editions of *The ARRL Handbook*. Photos should be glossy, black and white positive prints of good definition and contrast, and should be the same size or larger than the size that is to appear in *QEX*.

Any opinions expressed in *QEX* are those of the authors, not necessarily those of the editor or the League. While we attempt to ensure that all articles are technically valid, authors are expected to defend their own material. Products mentioned in the text are included for your information; no endorsement is implied. The information is believed to be correct, but readers are cautioned to verify availability of the product before sending money to the vendor.

Empirically Speaking...

Why Write?

Elsewhere in this issue you'll find some "Bits" that call for papers for upcoming VHF/UHF/microwave and digital conferences. The papers submitted to these conferences find their way into the proceedings books from each conference. These books are normally available at the conference and later can be ordered from ARRL or, in some cases, the organization sponsoring the conference.

If you aren't familiar with these conferences, you should be. Some of the most fascinating efforts underway to advance amateur technology are first reported at these conferences. So you certainly should be reading these proceedings! But you may wonder why you would want to submit a paper yourself. Of course, the main motivation for writing a conference paper is to share what you have discovered with those other amateurs who have a common interest in a particular facet of technology. But conference proceedings serve another purpose: they serve as a snapshot of the current amateur state of the art with respect to the conference subject. As such, they provide a means for amateurs *and others* to keep up to date with amateur technology. We emphasize "and others" because we have an interest in ensuring that regulators, legislators and other services are aware that Amateur Radio continues to foster experimentation and development of new technology. (If you doubt that it does so, scan some of the recent volumes of the proceedings from these conferences!)

Naturally, periodicals such as *QST*, *QEX* and other amateur magazines serve this function, too, and we would never discourage you from submitting your work to these! But nonamateurs—and many amateurs, for that matter—may not be regular readers of the amateur press. The conference proceedings provide an all-in-one introduction to the latest amateur technology. Of course, while even the mere existence of these proceedings is evidence that amateurs are doing something, these publications best serve the purpose of enlightening others—amateurs and nonamateurs alike—when they are chock full of good material. Which is where you come in.

This Month in *QEX*

QEX Contributing Editor Zack Lau, KH6CP, takes you on a tour of a 40-meter QRP CW transceiver he designed. Rather than just present the circuit and describe how it works, Zack presents some insight into the design decisions he made along the way.

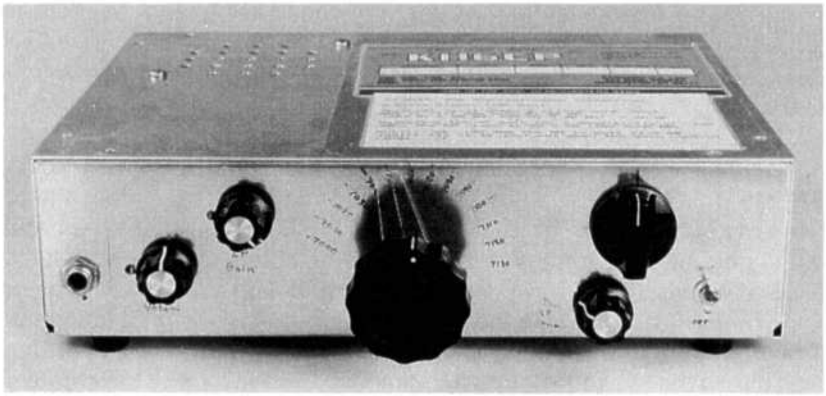
Think only parasitic antennas can be stacked? Not so! Al Christman, KB8I, has designed a unique wire antenna that acts like stacked inverted vees but can be easily constructed and mounted on a vertical antenna support, is easily fed at a single point, and gives respectable performance. His computer analysis is given in "The Double-Diamond Quad."

Zack Lau also contributes his "RF" column this month (we're shifting "RF" to the odd-numbered months), discussing a 432-MHz driver amplifier suitable for use between a transverter and a commercial power-amplifier module. (We're thinking of renaming this magazine *QEZ!*)

—KE3Z, email: jbloom@arrl.org (Internet)

Birth of a 7-MHz Transceiver

By Zack Lau, KH6CP/1
ARRL Laboratory Engineer
email: zlau@arrl.org (Internet)



Tired of canned printed-circuit-board projects that don't leave any room for innovation? This project shows one of many possible ways to build projects that can be easily modified to take advantage of the latest technology as it becomes available. The project is a 40-meter QRP CW transceiver. While you can just build the circuits shown if you like, our discussion of the project will focus on the *why* more than the *how*. It's at least as useful to understand why a designer made the choices he did as it is to understand how the circuit works.

The design goal of this project can be stated simply: a clean QRP CW transceiver with a minimum of glitches and decent performance. As we discuss the design, keep in mind that it is almost always possible to improve the performance of an individual circuit, but making the cost and complexity of the circuit fit the need, and integrating the circuit into the design as a whole are the essence of good design. These issues drive many of the design choices discussed here.

Fig A shows the block diagram of the transceiver. Since it is a transceiver, some of the circuits are used in both transmit and receive modes. While this choice complicates the initial design, it dramatically lowers the total parts count.

The Receiver

Choice of mixing frequencies is what often makes or breaks an HF receiver. Choose the wrong ones and you will have to struggle to get something that even sounds halfway decent. A particularly *bad* choice is to have the IF and local oscillator approximately equal, such as using a 3.59-MHz IF and a 3.41-MHz LO. Not only can it be difficult to keep the local oscillator out of the IF, but there are spurious responses that can't be filtered out easily. If necessary, one can substitute high-Q tuned circuits for the broadband transformers to filter out unwanted signals. Don't forget, though, that the purpose of having an IF in the first place is to make things easy. It makes little sense to choose one that makes it a real pain to get circuits to work!

For most amateurs, the choice of IF is determined by what crystal filter they can get their hands on. For this 40-meter unit, I found that using 12-MHz crystals resulted in a Cohn filter design that neatly matched 50 ohms. I also considered other frequencies, but ruled them out for the following reasons: With 10-MHz crystals, the image is at 13 MHz, which is a little close if you use a top-coupled filter with its characteristically poor high-side rejection, although picking off your receiver input signal after the transmitter's low-pass filter is somewhat of a help. On the other hand, with 16-MHz crystals, you need a 9-MHz VFO. With VFO stability degrading as you go up in frequency, 9 MHz may be a little too high, though there is a commercial rig, the ICOM IC-502, that used an incredibly high VFO frequency of 36 MHz. That's probably too high, as evidenced by '502s with significant drift problems.

A disadvantage of the 12-MHz VFO frequency is that the oscillator tunes "backwards"—the highest VFO frequency corresponds to the lowest operating frequency. This pretty much rules out using capacitors with built-in reduction drives, as they usually have stops that prevent you from having a capacitor that is at minimum capacitance when you have rotated the shaft fully clockwise. And the capacitors I've seen don't appear to be rugged enough to be modified easily.

Mixing, RF Switching and Filtering

As with any receiver design, one major decision was what to use as the first receive mixer. I opted for the moderately high-performance approach of using a Mini-Circuits SBL-1 doubly balanced mixer. I also decided to take advantage of its bilateral operability, using it on transmit as well. The Siliconix Si8901 quad JFET mixer can also be used bilaterally, but these devices are difficult to obtain. The Plessey SL6440 is a good, strong mixer, too, but it is still quite difficult to get unless you want to buy a kilobuck worth of parts! Like most active mixers, it is also unilateral.

Using the mixer for both transmit and receive requires that it be switched between the active signal paths. Rather than use sometimes hard to find or expensive PIN diodes for mixer switching, I chose to use 1N4007 rectifier diodes. Unlike the 1N4001, the 1N4007 has the required intrinsic layer (signified by the *I* in PIN). 1N4007s are also used to switch the band-pass filter, which is also used bilaterally so that only one filter has to be aligned.

One of the difficulties in getting maximum performance out of an SBL-1 is terminating it properly. The most straightforward way of doing this is with a broadband amplifier with a good input SWR (as indicated by low return loss) across a frequency span that includes as many of the mixing products as possible. It is possible to get better performance if you use a diplexer—but a diplexer, which is a network of two or more filters, has to be properly aligned to work. A swept return-loss measurement setup is often needed to adjust a diplexer properly, due to the LO and intermediate frequencies typically used in MF and HF systems. A significant advantage of the SBL-1 is its wide IF-output specification—unlike some active mixer designs that will not accommodate a VHF IF because they have too much stray capacitance. Conceivably, the SBL-1 could even be used with a UHF IF. Getting suitable oscillators for those frequencies can be a problem, though.

While perhaps not important to the casual builder, the problems with using a broadband-amplifier termination for the mixer are those of excessive gain and degraded intercept point. The first might seem pretty strange, as you would think you want as much gain ahead of the filter as possible to override the noise figure of the filter plus the following stages. But crystal filters often have very weird distortion properties—I've reduced the level of two-tone test signals by 10 dB and seen the distortion products drop by only 10 dB, rather than the 30-dB drop expected. So, for the ultimate in dynamic range, you often want to *limit* the gain ahead of the crystal filters. And in fact, the poor return loss of the crystal filter at the IF isn't as important as the termination it presents at the image frequency.

The problem of the degraded intercept point is more straightforward. The broadband amplifier has to handle all the signals at the output of the mixer. At minimum, the intercept point should be degraded by at least 3 dB, since the unwanted image is just as strong as the desired sig-

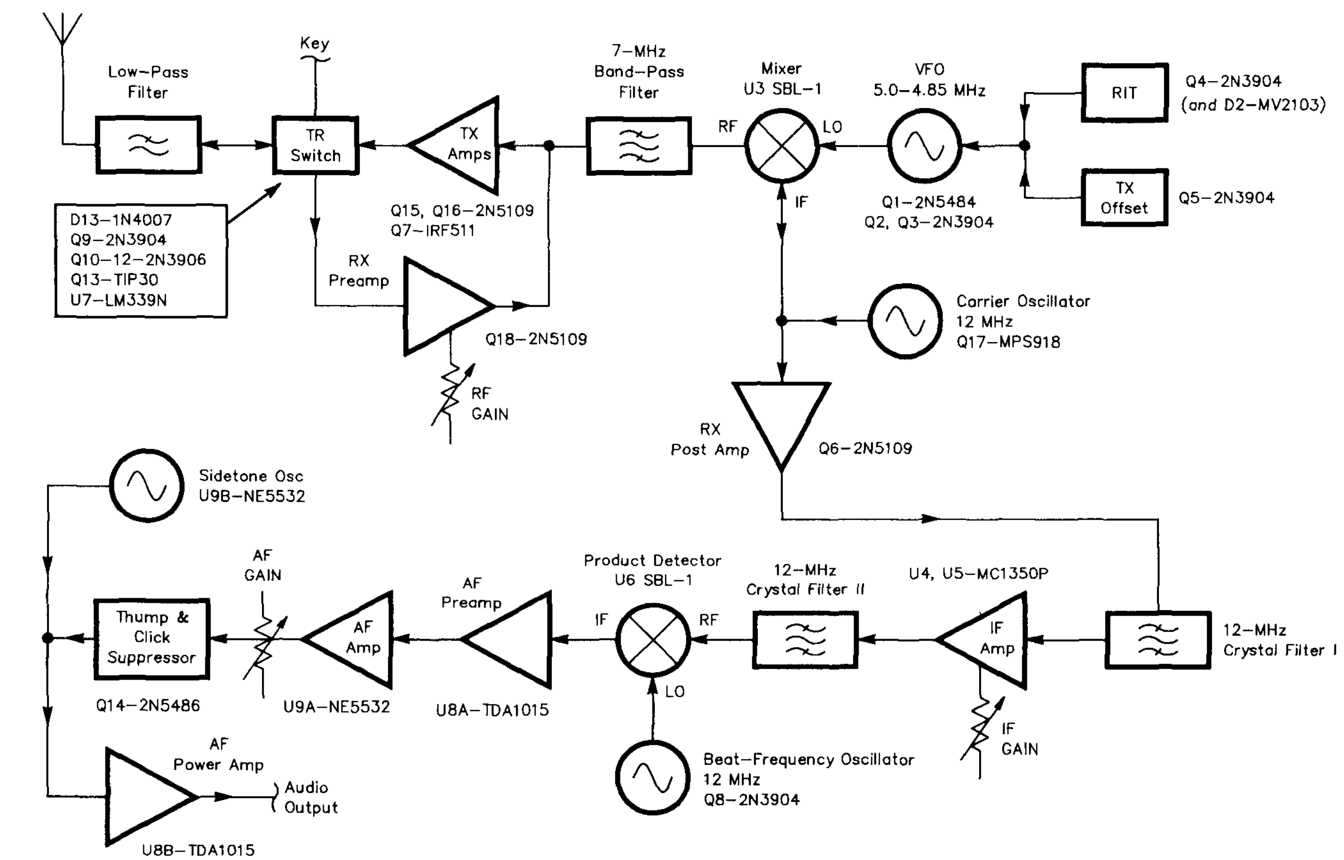


Fig A

nal. The other signals, such as the higher-order mixing products and the local oscillator feedthrough make things even worse.

Terminating the mixer with a diplexer can reduce the sorts of problems encountered with a broadband amplifier termination. A question that comes up when designing diplexers is how good do they have to be. A paper by Paul Drexler in the *1987 Mid-Atlantic States*

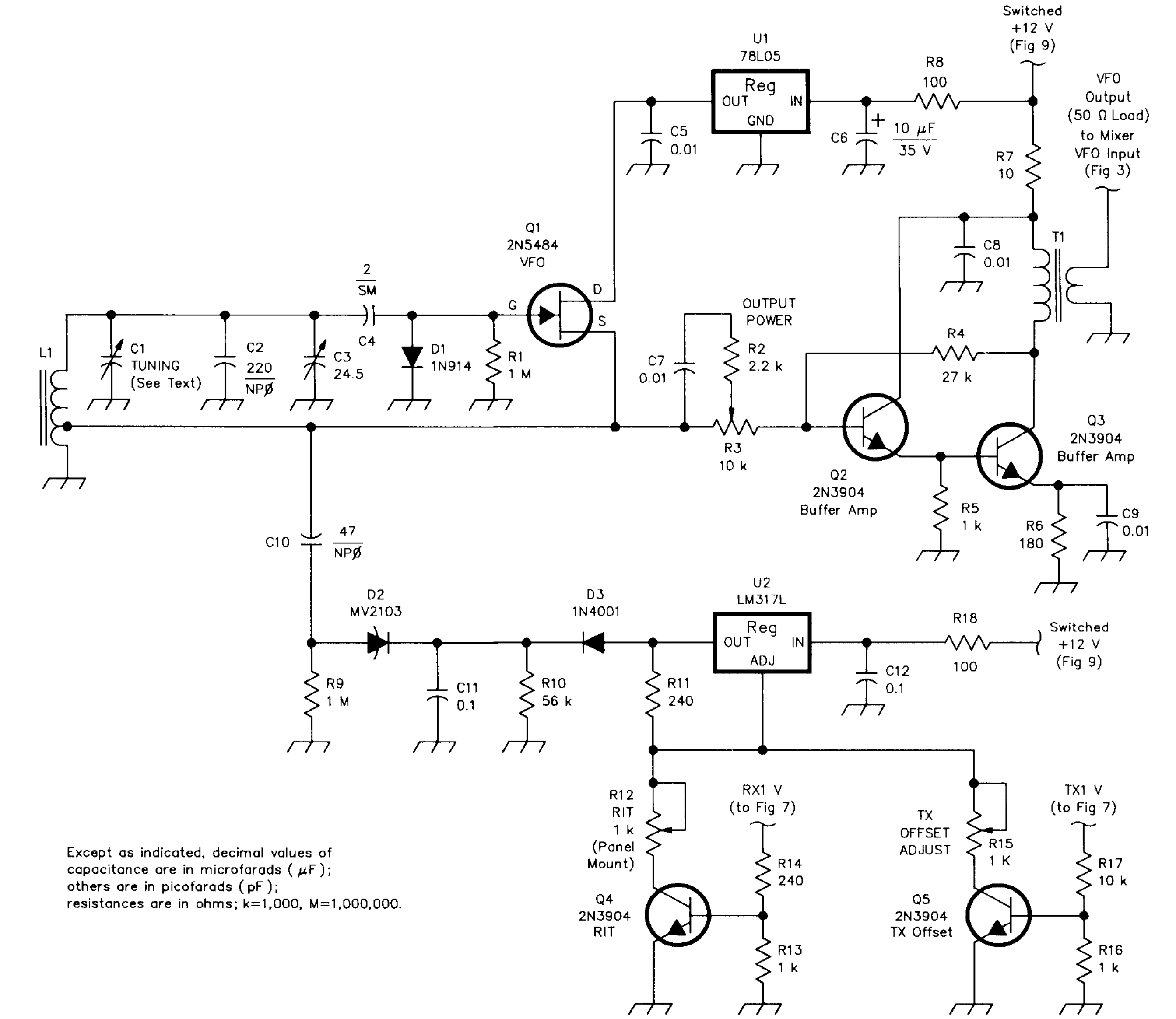


Fig 1—Schematic diagram of the VFO circuit.

C1—23.1 pF, air dielectric, variable. (Millen 21020) (Main tuning control that should have a reduction drive).
C3—2.4 to 24.5 pF air dielectric trimmer. Johnson 189-509 type.
D1, D5, D6, D7, D8, D9, D10—1N4148 or 1N914 switching diode. (Some of these are shown in other figures.)
D2—MV2103 varactor tuning diode.
D3—1N4001 rectifier diode used for temperature compensation. Do not substitute.
L1—29 turns #26 enameled wire on T-50-7 iron-powder toroid. Tap 9 turns up from ground.
Q1—2N5484 JFET. 2N5485 or 2N5486 will work with increased drift. The MPF 102 may also work.
Q2, Q3, Q4, Q5, Q8, Q9—2N3904, PN2222, 2N2222. BJT. (Some of these are shown in other figures.)
T1—Broadband transformer, 5:1 turns ratio. 10 turns of #30 enameled wire on an FB-2402-43 balun core (primary). Secondary has 2 turns of #30 enameled wire over primary. #26 or #28 enameled wire may be used if the larger FT-37-43 toroid core is used.
U1—78L05 5-volt regulator. A 78M05 or 7805 regulator will work but is physically much larger.
U2—LM317L adjustable voltage regulator. A LM317M or LM317T will work but is physically much larger.

VHF Conference seems to indicate that there is a difference of several decibels in output intercept between using a 20-dB resistive pad (40-dB return loss or better) and a diplexer with 20 dB of return loss or better, with the advantage to the pad. Clearly, return loss is critical in this scheme. Of course, getting 40 or 50 dB of return loss out of a wide-band filter network is awfully difficult in the HF and VHF spectrum, so perhaps there is room for better approaches. By the way, for those accustomed to thinking in terms of SWR, going from a 20-dB to 40-dB return loss is lowering the SWR from 1.222 to 1.020.

There is a totally different approach that one can take to terminating mixers. If you know what you are doing and have the test equipment available, excellent results are often possible with nonresistive terminations. Edward Meade Jr, apparently did just that about 20 years ago, getting a +15-dBm output intercept and 5 dB of conversion loss out of a mixer quite similar to today's SBL-1 at VHF. His material is still in today's *Handbook,* in case you want to know what can be done with a properly designed low-pass-filter termination, as well as how to do it. Since his approach requires a relatively low IF, it might be something to consider at HF if you use a single-conversion 455-kHz IF with sufficiently sharp front-end filters to get rid of the really close image response. His approach actually makes more sense than the textbooks' gospel of broadband termination—just try and design a diplexer that works into the UHF spectrum with parts available to amateurs in the 1970s! Even today, while you can use chip components to terminate images in the SHF spectrum, you will likely find that the output transformer of the mixer has nowhere near the required amount of bandwidth, and your carefully crafted SHF circuit does you no good. Theory is fine for ideal parts, but compromises are often needed when dealing with stuff you can actually get your hands on.

In the end, I chose to go with a broadband termination. The receiver post amp (Q6, Fig 2) is designed to present this termination to the mixer. A resistive pad is needed at the output of this stage to prevent the poor out-of-passband impedance of the 12-MHz crystal filter from affecting the mixer through the amplifier. (A difficulty with this simple feedback amplifier design is that the isolation between the input and output isn't very good.) I prefer to use metal-film resistors, as older carbon-composition resistors often have oxidized leads that require more work to solder properly, and there is little difference in RF performance. Using an exotic high-power JFET in this stage would produce better isolation, but such devices aren't easily available to most amateurs. (The Motorola MRF-136 TMOS device looks like an interesting transistor to try, featuring very high intercepts and a low noise figure.) An adventurous experimenter might want to try using feedback amplifiers with directional couplers, but I haven't seen a ready-to-go design.

While mixer performance is important, so is the performance of the circuit in front of the mixer. I decided to implement an RF gain stage (Q18, see Fig 8) with a true RF gain control. Varying the amount of Q18's emitter resistance that gets bypassed by capacitor C85 varies the gain of the stage. A disadvantage to this approach is that the input impedance of the amplifier increases as the gain drops. Also, you need a decent potentiometer: an inductive wire-wound pot won't work. But these disadvantages are offset by the advantage that the dc bias remains the same, so that signal handling isn't degraded when you need it most. Testing shows this isn't *entirely* true, as the two-tone dynamic range does degrade as the MDS increases from −130 to −120 dBm as the RF gain control is varied. But the dynamic range does increase as expected in going from −141 to −130 dBm. The RF gain stage also works to further isolate the input of the transmit amplifier chain from its output, improving transmitter stability. Removing the stage may cause difficulties in obtaining a stable transmitter. The selectivity ahead of the RF amplifier—a single tuned circuit and a low-pass filter—isn't great, but it is similar to that of many rigs. An additional band-pass filter could be placed between the single tuned circuit and the RF amplifier for even better performance.

Designing broadband amplifiers like that of Q18 requires solving a couple of circuits simultaneously. The most basic circuit is the dc biasing circuit—you usually want a certain collector or drain current that doesn't vary too much with temperature or component tolerances. Next, you want a flat gain characteristic while maintaining input and output impedances close to your design goal, typically 50 ohms. Increasing the emitter degeneration resistor decreases gain and increases the input impedance, while increasing the shunt resistor from the output to the input has the opposite effect—increasing gain and decreasing the input impedance. Finally, you want the amplifier to be stable regardless of load. Usually, the first two requirements are easily combined, while the third is often tacked onto the final result.

The VFO

Given the problems in obtaining a stable VFO, why not use a synthesizer? They usually offer improved stability, but only at the cost of increased complexity. It appears possible to combine direct digital synthesis (DDS) with a phase-locked loop (PLL) to eliminate the major problems of both—the phase-locked loop can remove the spurs generated by the DDS, and the DDS can help to reduce the lockup time/phase noise trade-off of PLLs. But this ends up with a circuit as complicated as the rest of the transceiver! Of course, one can reduce the amount of control circuitry by using a computer, as shown in the 1988 to 1990 ARRL *Handbooks.* In the end, I decided to keep it simple and use a VFO.

The VFO used in this radio, shown in Fig 1, is a JFET Hartley circuit, Q1, followed by a two-stage bipolar buffer circuit, Q2 and Q3, that drives the mixer. Since there is a bit of variation in the power output from one JFET transistor to another, I decided to allow the gain of the buffer to be adjusted with R3. (Of course, you could use a selected JFET and skip the adjustment circuit.) A JFET with a lower drain-to-source resistance or higher I_{DSS}

rating usually provides more power output, though usually at the expense of frequency stability. Improved performance, perhaps another 6 dB of IMD dynamic range, could be obtained by waveshaping the local oscillator to provide a good square wave. Since the IMD occurs primarily at the switching transitions, shortening the length of those transitions improves performance. While not the ultimate in stability, this circuit does provide good phase noise performance. Yes, you can still have a phase noise problem even without digital circuits! (See *QRP Quarterly* July-October 1992, "Experimenter's Corner: A High Performance DC Receiver," by Denton Bramwell, K7OWJ.)

One of the trade-offs that has to be made when using tuning diodes is to choose between phase noise performance and tuning range. By using lower voltages on the diode, you can get much more range out of the diode, but the Q decreases, and therefore the phase noise gets worse. For casual applications this isn't important, but you lose much of the benefit of improved mixer dynamic range if your receiver is unable to eliminate strong signals because the selectivity is smeared by poor phase noise performance.

Use of a 78L05 three-terminal, 5-volt regulator prevents the oscillator frequency from varying with changes in the supply voltage. Don't forget to bypass the input terminal of this regulator properly with a 0.33 µF or larger capacitor; I have seen this type of regulator oscillate if not properly bypassed. One I saw took off at a few MHz and amplitude modulated the oscillator it was connected to!

The VFO inductor uses a Type 7 Micrometals iron-powder coil. While not as easy to get as the common

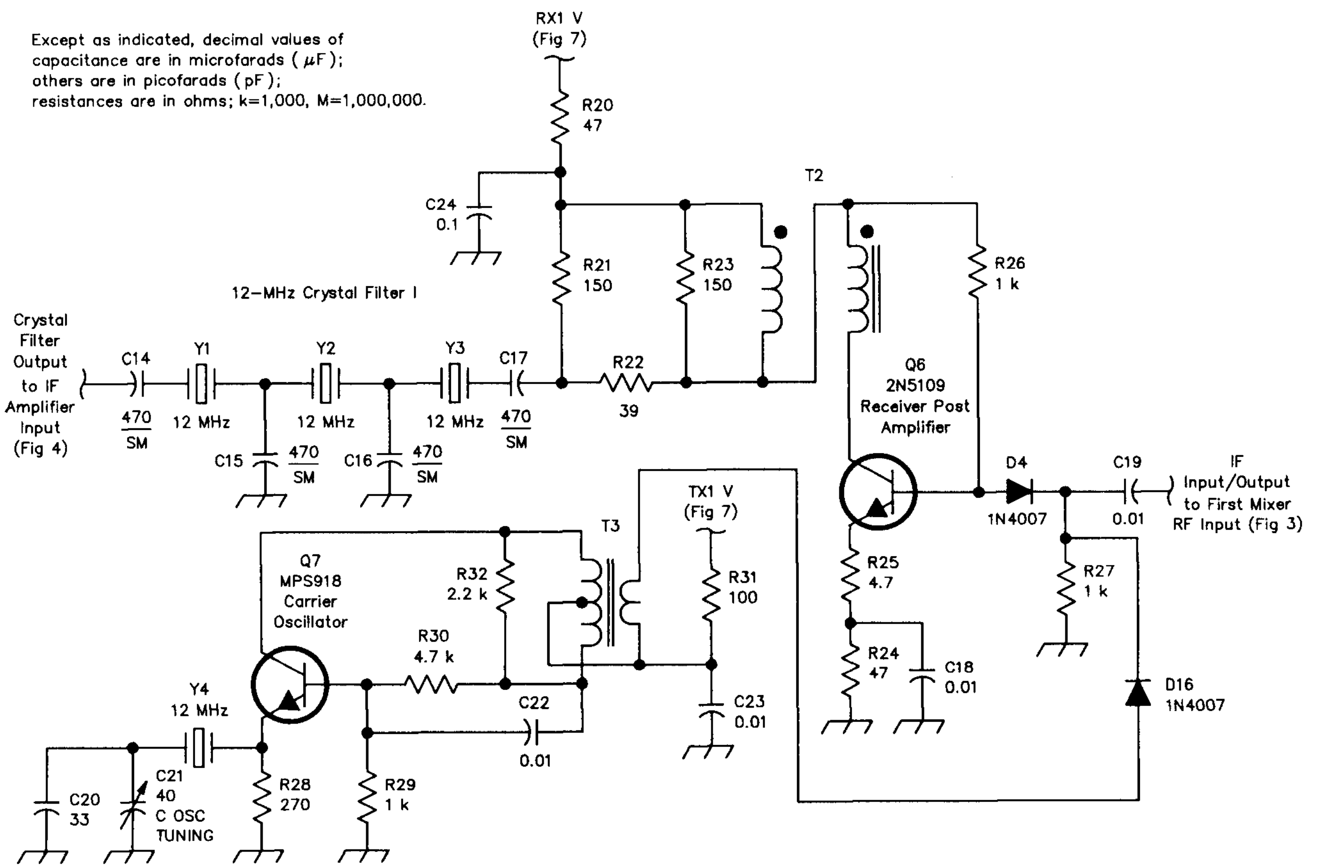


Fig 2—Receiver post amplifier and carrier oscillator circuits.

C20—Exact value depends on the crystal used. It might even be omitted if Y4 is appropriately low in frequency compared to Y1.

D4, D11, D12, D13, D16—1N4007 rectifier diode used as a PIN switching diode. More expensive PIN diodes such as the 1N5767 may be used. (Some of these are shown in other figures.)

Q6, Q15, Q16, Q18—2N5109 BJT. (Some of these are shown in other figures.)

Q7—MPS 918. A 2N5179 or 2N5109 will also work.

R32—May not be necessary. Added to prevent amplitude modulation present on one of the oscillators built.

T2, T6, T7—Broadband transformer. 10 bifilar turns #28 enamel wire on FT-37-43 toroid core. (Some of these are shown in other figures.)

T3—Broadband transformer. 10:1 turns ratio; 20 turns of #26 or #28 enameled wire on an FT-37-43 core (primary). Tap is 13 turns from the collector. Secondary has 2 turns of #26 or #28 enameled wire over the primary.

Type 6 core, it has a slightly lower temperature coefficient (30 versus 35 ppm/°C). It doesn't hurt to get the most stable parts possible, given that we want a VFO stable to within at least 10 ppm, if not better. You may wish to anneal the core by boiling it in water and letting the water and coil cool slowly, but I've not done measurements to see how effective this is.

One of the critical circuits I almost forgot to include is a receiver incremental tuning (RIT) control, which is the first thing I would have noticed if I used it on the air! RIT is really important if you are attempting to work someone using a DC receiver/transmitter combination that has no frequency offset capability. They can only work stations that can operate split frequency. RIT is implemented by U2, an adjustable three-terminal voltage regulator IC (see Fig 1). The voltage applied to D2, the varactor tuning diode, is controlled by R12 when receiving and by R15 when transmitting. I haven't seen this particular method of obtaining transmit and receive offsets used previously. It allows you to add transmit incremental tuning (XIT) fairly easily, though the biggest problem is likely to be finding the panel space! A compromise may be to use a single control and a DPDT switch to choose either XIT or RIT. Since a dedicated 8- to 10-volt regulator is needed in order to get enough range from a tuning diode, I decided to just switch the resistors that determine the output voltage. This allows the use of cheap NPN switching transistors (Q4 and Q5). There should be no interaction between the two variable resistors, and none has been noted. This method also has the advantage that it is usually possible to set up the RIT with just a voltmeter, although a frequency counter would eliminate any error caused by the transmitter pulling the VFO. This error, while sometimes severe if the transmitter operates on the same frequency as the VFO, is usually negligible in a properly designed heterodyne system. D3 and R10 are suggested by a Motorola application note as temperature compensation for the MV2101 series of varactor diodes, though I've not verified their results. An LM2931 low-dropout regulator was tried at U2 with mixed results: the high output capacitance required for stability hinders its transient response. Thus the circuit I tried doesn't change voltages quickly enough for QSK use. Wasting current with a small-valued resistor load would speed things up, but conserving current drain is usually a design criterion.

The BFO and CFO

You may be wondering why I don't just use the cheaper method of using the same oscillator for the BFO and carrier oscillator, swinging the VFO to obtain the proper offset. Yes, it's cheaper, and it works cheaper, too, as the offset you get depends on what frequency you are on. That's not too bad it you only cover 20 or 40 kHz of the band, but it certainly can be noticeable if you want the VFO to cover the entire band.

I added 47- and 100-Ω swamping resistors to the beat frequency oscillator (Q8 in Fig 5) after I discovered that the circuit had a tendency to take off on the third overtone of the crystal. This happened because of the poor load provided by the mixer, as demonstrated by the fact that such techniques as rolling off the gain of the oscillator with a capacitor from collector to ground had little effect. Not only isn't crystal overtone operation desired in this case, but I've found that using overtones the crystal manufacturer didn't expect you to use can lead to unreliable results, caused by spurious responses close to the desired frequency. As an example, I looked at the 5th overtone of an 18-MHz crystal and found two responses 300 kHz apart showing only about a dB of difference in series attenuation in a 50-Ω system. Careful crystal manufacturers make sure that the spurious responses are at least 3 or 4 dB down.

Crystal Filters

Since I know how much people hate to wind transformers, I designed these crystal filters to work well with 50-Ω input and output impedances. As a result, the only transformer needed to match the filters to the rest of the circuitry is the one that couples to the high-impedance MC1350 IF amplifier output (T4 of Fig 4). A second crystal filter is used just before the product detector. Since a

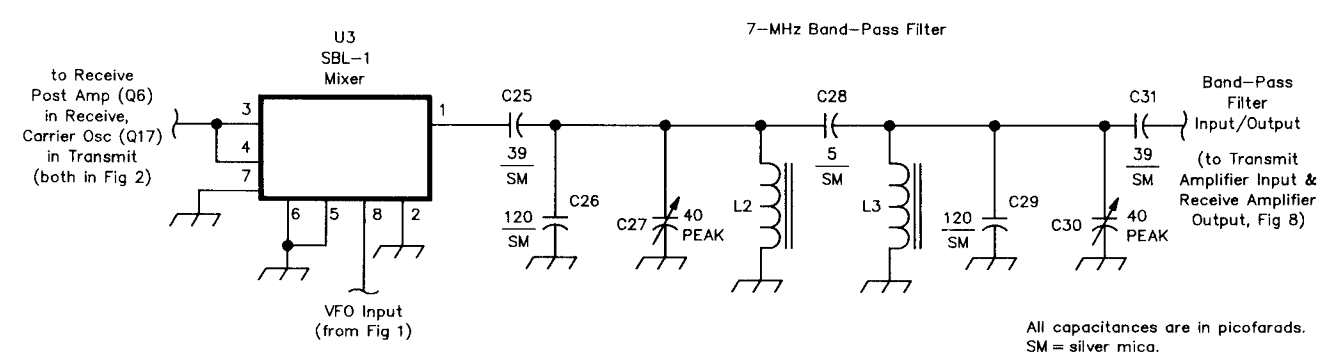


Fig 3—Mixer and band-pass filter.

L2, L3—22 turns of #20 enameled wire on a T-68-6 iron-powder toroid. (2.7 µH, Q=340 at 7.1 MHz)

U3, U6—SBL-1 doubly balanced mixer. (U6 is shown in Fig 6.)

product detector acts like a direct conversion receiver, it detects the audio image noise amplified by the IF amplifiers unless you filter out the noise.

AGC, Product Detector and Audio

You may wish to add your favorite AGC circuit, keeping in mind that the CW-bandwidth crystal filter adds a slight delay that may or not improve the AGC response. This delay *can* be helpful, as it can give time for your circuit to decide exactly what to do.

The product detector is another SBL-1 diode mixer, U6 in Fig 5. Use of the SBL-1 avoids local oscillator leakage resulting from poorly matched diodes, a common problem with home-brew diode mixers. If you do add AGC, the mixer's LO suppression of about 70 dB should be sufficient to prevent the carrier injection from upsetting the AGC, though be sure to avoid poor layout. Of course, balancing a mixer over a narrow range of frequencies isn't difficult, if you have the test equipment to make the adjustments. If you do have trouble with a feedthrough from a home brew mixer, you might consider using a 6-dB hybrid combiner/splitter to separately feed the AGC and product detector from the IF amplifier. This should give you another 40 dB of LO suppression, making even a mixer with lousy balance usable.

The TDA1015 audio chip used in this design (see Fig 6) puts out quite a bit of power; 4 watts is plenty for a quiet room. But this two-stage amplifier also has quite a bit of gain, and it will oscillate if you aren't careful with your circuit layout. You probably can substitute an LM380 and increase the gain of U9A to compensate. The LM386 is a possibility too, but, while I can't say it's bothered me, people have complained about the LM386 being "hissy" so I won't recommend it.

Like "The QRP Three Bander" I presented in the October 1989 *QST*, this rig uses an audio gate made with a JFET switch, Q14, to cut down on pops and clicks. This switch is controlled by the T-R sequencer, so we'll discuss it when we get to that part of the circuit.

A simple sidetone oscillator, shown in Fig 9, is made from an op amp section. It's not a particularly clean sounding tone, so you may wish to substitute your own favorite circuit.

The Transmitter

When trying to listen on one band while transmitting

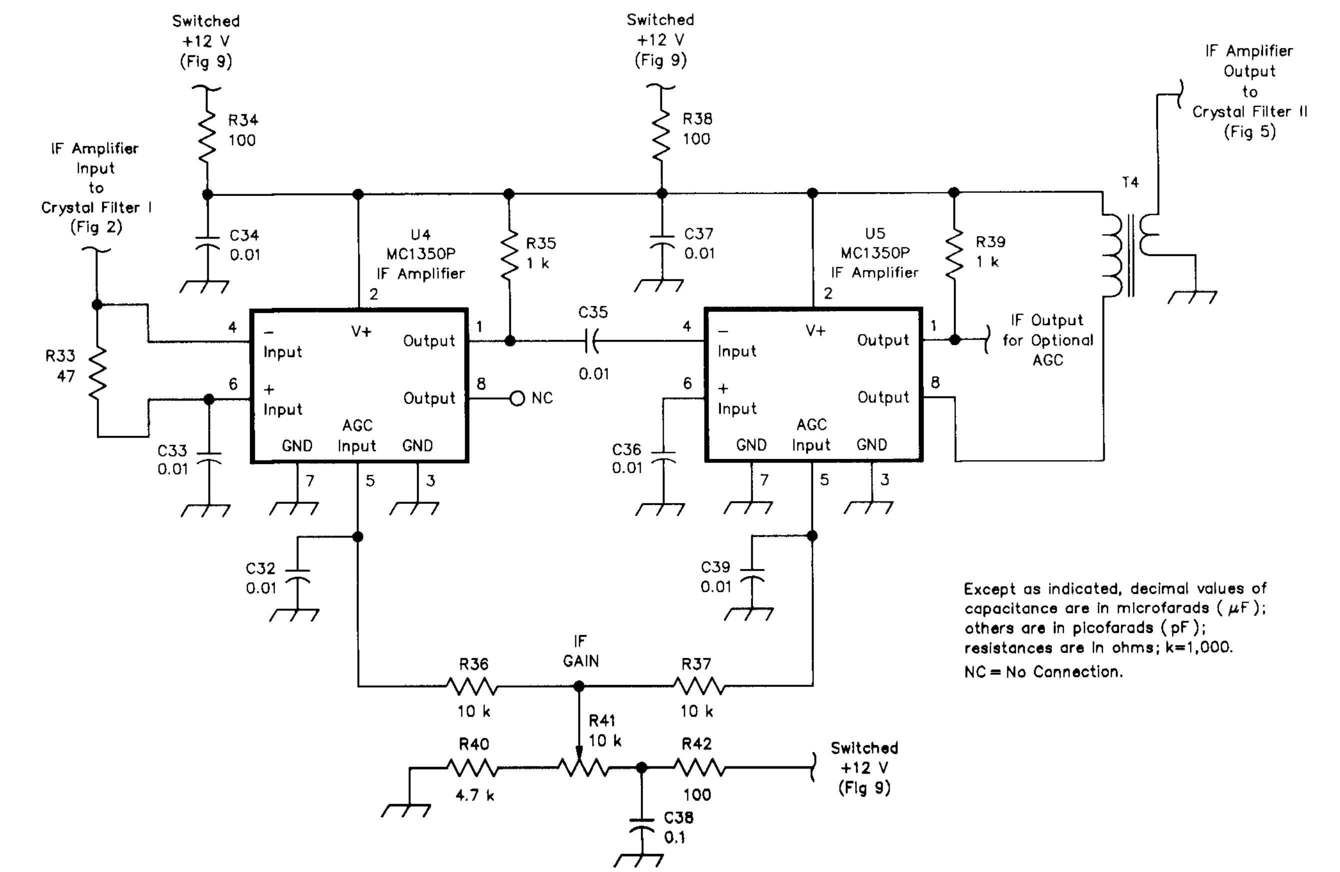


Fig 4—IF amplifiers.

T4—Broadband transformer. 19:4 turns ratio; 19 turns of #26 or #28 enameled wire on an FT-37-43 core (primary). Secondary has 4 turns of #26 or #28 enameled wire over the primary.

U4, U5—LM1350P IF amplifier.

on another, transmitter stability often makes a big difference in how successful you are. Even a low-level instability makes a huge signal in a nearby receiver. While solutions to stability problems are not trivial, it *is* possible to make a transmitter stable enough that you can listen to it with a receiver in the shack and have it sound clean. Of course, unwanted coupling paths and ground loops can have you chasing "problems" that don't really exist.

This transmitter uses a broadband amplifier driving a MOSFET transmitter output circuit developed by Mike Masterson, KA2HZA. Not only does the MOSFET run at high efficiency, but it's cheap and rugged as well! His design was reworked for use at the 5-watt-output QRP level. Those looking for a little more power might look at Mike's original article, "Three Fine Mice—MOuSeFET CW Transmitters," which appeared in the December 1986 *QST* and is reprinted in the ARRL book, *QRP Classics*. His circuit provides the component values for getting approximately 16 watts out. I spent a little time combining the transmit broadband amplifier, the receive preamplifier, and the switching circuit so that fewer parts are needed. In addition, a little less current is required, since some of the bias current for the transmit amplifier is also used to turn on the T-R switching diode.

Transceiver Switching

As many home brewers have discovered, merely having a transmitter and a receiver circuit doesn't make a transceiver. In fact, most will say that quite a bit of work is needed to get a radio that works well, with the difficulty coming in the circuitry that switches the unit between transmit and receive modes. To make things relatively straightforward and understandable, I used an LM339 quad voltage comparator to build a sequencer (Fig 7), a device that not only switches voltages sequentially, but switches them in reverse order when switching the other way. These voltages make it a lot easier to design a transmitter that doesn't oscillate when switching between transmit and receive and a receiver that has a minimum of unwanted noises. There are two diodes in the circuit, D6 and D7, that might not seem entirely necessary at first glance. D6 is used to raise the threshold voltage of the T-R switch, Q9. If the threshold is too low, you run into the situation where the transistor switches of some electronic keyers won't key the rig properly. Surprisingly, some commercial rigs have this problem. D7 is in series with a 68-kΩ timing resistor. This network equalizes the key-up and key-down timing—I thought it might improve the rig to have the times approximately equal.

A more flexible—and simpler—substitute for the LM339 sequencer is to use an LM3914 bar graph chip with an R-C generated ramp voltage on its input. In the bar graph mode, as opposed to the dot mode, you can ignore the first LED and use the other 9 sequenced outputs—enough for almost any transceiver or transverter application. The circuitry is also a bit simpler, the only real disadvantage being the higher cost of the chip. You'll probably also need at least one inverter, but this is just another PNP transistor and resistor. In the case of the LM339 circuit, you can do inversion by merely interchanging the comparator inputs.

A Darlington PNP pair, Q12 and Q13, switch the transmitter driver amplifier on and off with wave shaping. Not only does this reduce the current that has to be sinked by the comparator output, it also reduces the timing capacitances to more reasonable values. While you can buy 2.2-µF nonpolarized capacitors, they tend to be hard to find and more expensive than smaller metal-film capacitors. Ordinary electrolytics probably won't work because the voltage across them reverses.

To get rid of those annoying pops and clicks resulting from a receiver being energized before the transmitter is truly off, the sequencer disables the audio line via Q14 (see Fig 6) during the transition between modes. The timing capacitor connected from the gate of Q14 to ground is normally chosen to be the smallest that will cut off the annoying noises, though you certainly could increase it to suit personal preference. Not everyone appreciates hearing background noise between CW "dits." Should you wish to use this circuit, keep in mind that the circuit feeding the JFET source must always maintain a dc potential of several volts. If it drops below a certain threshold, annoying noises will be generated. Critical parameters of FETs—unlike those of bipolars—often vary considerably, so experimentation will be needed to find the threshold of your transistors, if you need to cut things that close. It's often necessary to use a pair of clipping diodes in the feedback loop of the op amp (U9A) to ensure that the JFET source voltage never drops below the threshold.

Finally, for a really polished result, you might consider designing your radio to be well behaved enough to be cycled on and off, no matter what it happens to be doing at the time. Or at a minimum, allow it to be turned on and off without making annoying noises. The circuit presented here needs more work in this regard, as the audio does make an annoying click as the supply voltage dies. It's a lot easier to solve this problem if the power supply is designed as a part of the rig—you want to disable the speaker before the voltage drops appreciably. Then again, you may just choose to live with the problem. We have seen some awfully complex projects make some equally awful noises when we turned them off!

An interesting design issue has to do with power supply bypassing. I elected to R-C bypass almost everything, even though it might not be necessary with a good power supply. Not only does this isolate noise, but it tends to simplify troubleshooting. A stage that has failed often draws abnormal current. This will easily be detected by the excessive voltage drop—or no drop at all—across the series decoupling resistor. R-C decoupling does run up the part count, of course. On the other hand, if you are doing a printed-circuit layout, it makes layout easier: the resistors bridge traces, sometimes eliminating the need for jumper wires.

Construction and Testing

At a minimum, you will need a voltmeter and an RF probe to build this project. For a project as complicated as this, I recommend you build and test each stage, one at a time. I'd probably start with the VFO, as it's the most difficult stage to align. Then again, someone looking for the easiest circuit to start with may wish to build the sequencer first. By tacking a 33- to 470-µF electrolytic capacitor across C69 and connecting some indicator lights to the outputs of the circuit, you can watch the behavior and see if it works properly.

If you have a calculator, you can do some playing around to see why breaking down a big project such as this into manageable chunks for testing is so important. Plug in a typical accuracy rate, say 0.95, and multiply this number by itself *n*–1 times, where *n* is the number of parts you used. For instance, a circuit with 10 parts means you do the multiplication nine times to multiply the 10 numbers together. Alternately, those with scientific calculators can do this with one step, using the exponent function. I get approximately 0.599 on my calculator. This isn't terribly good—this implies that 40% of the time the circuit won't work on the first try! It's even worse with 100 parts, as the expected chance of failure jumps to 99.99 %! (Obviously, those automated electronic assembly plants have accuracy rates much better than 95%.)

While I don't use it too often, it's often helpful to use one of the various "superglues" available. I've had pretty good luck with the thick viscosity stuff available from a local hobby shop. There are several different varieties available, and it seems important to choose the right one. While a recent discussion in *Model Airplane News* indicates these adhesives contain no cyanide, when heated they do give off a gas that seriously irritates the eyes.

Similarly, if you choose to use a thread-locking compound instead of lock washers, you should probably get the low-strength material designed to be removed, unless you have no intention of separating the parts again! Even if there is a solvent, it won't be too useful, as how can you get at the stuff between the threads? And keep in mind that the stuff is nonconductive—don't use it with fasteners you expect to provide good electrical contact!

While a "custom" chassis was used in the prototype, finding a suitable substitute shouldn't be difficult for

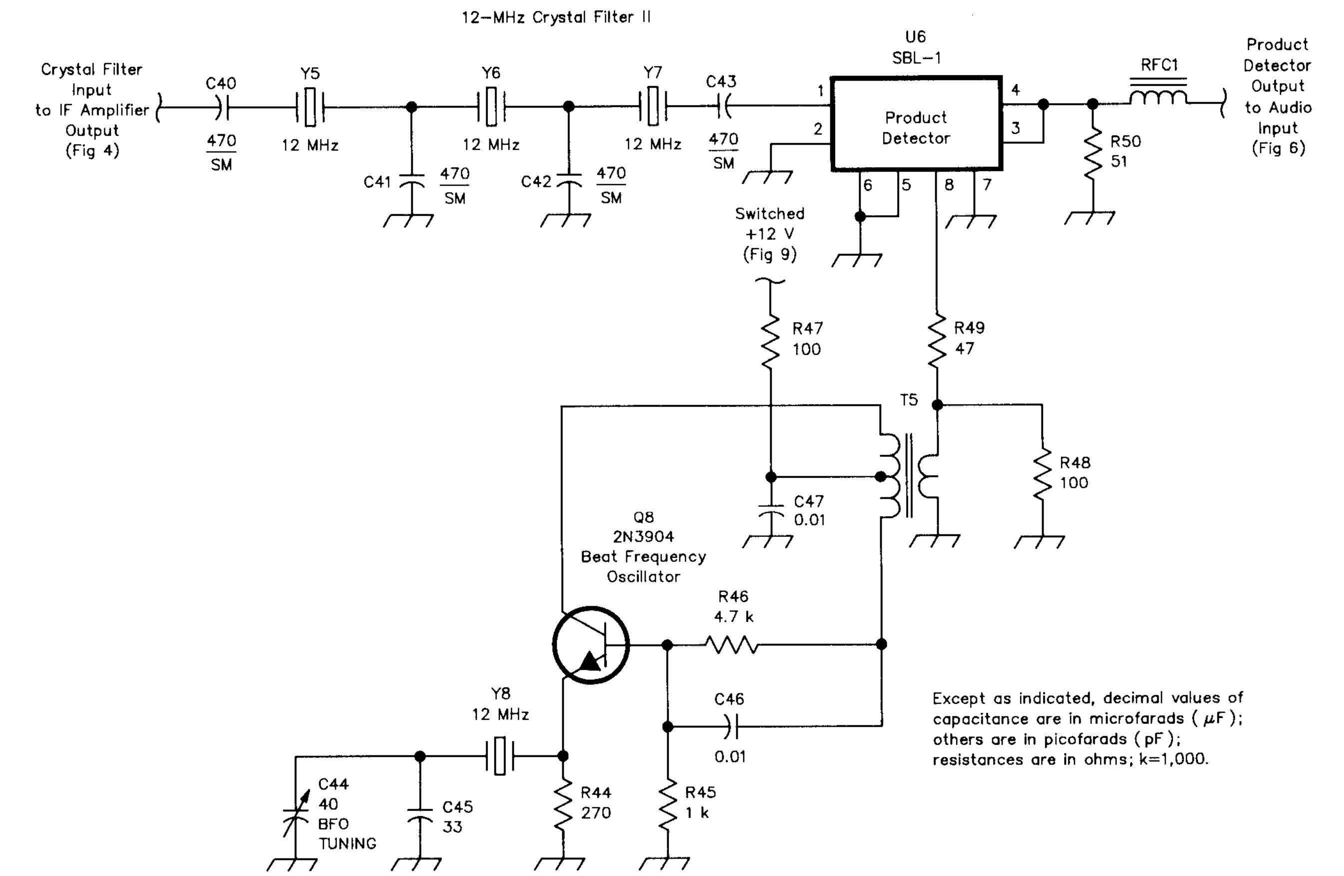


Fig 5—Second crystal filter, product detector and beat frequency oscillator.

RFC1—20 turns #28 enameled wire on an FT-23-75 ferrite core. An exact turns count isn't important—just fill the core with a single layer of wire.

T5—Broadband transformer. 5:1 turns ratio; 20 turns of #26 or #28 enameled wire on an FT-37-43 core (primary). Tap is 13 turns from the collector. Secondary has 4 turns of #26 or #28 enameled wire over the primary.

anyone who calls himself an experimenter. It's just an L-shaped piece of aluminum (bottom and back panels). Actually, you could probably even use steel, since the chassis doesn't have to carry RF currents. (It's not unusual for a steel chassis to sit around for a long time in the junkbox, simply because steel is harder than aluminum and much more difficult to work with, though cobalt drill bits do a fine job with steel. Some of the older rigs, such as the Drake TR-4, used copper-plated steel to provide good RF conductivity.) The circuits are built on pieces of 1/16-inch-thick circuit board roughly 2.7 by 3.7 inches in size. I tapped four #4-40 mounting holes in the corners of each piece to easily attach the boards to the chassis with 1/4-inch #4-40 screws. (I find holes threaded in glass epoxy circuit board to be adequate for attaching the boards. Don't attempt this with phenolic board—the corners will crack off.) Although I used double-sided board because it was available, single-sided board might be preferred. (Theoretically, having the copper and aluminum next to each other invites corrosion, but I haven't noted a problem. Nonetheless, this might be something to consider if you are going to use this building technique where corrosion is a problem, such as locations having high humidity or salt spray.)

Using 3/8-inch-long screws at the corners of the chassis instead of 1/4-inch screws allows rubber bumper pads to be affixed to the chassis with the screws used to hold down the circuit boards. I find the screw-attached bumper pads to be more reliable than the self-stick type.

Other mounting schemes I have tried include soldering nickel-plated brass nuts to the circuit board and using PEM nuts, but merely tapping the glass epoxy board has proven to be the most cost effective. (PEM nuts may be obtained from Small Parts Inc, a supplier that caters to the experimenter needing small quantities of metal stock and hardware. Unfortunately, they carry only steel and stainless steel PEM nuts rather than aluminum ones that might be preferred by amateurs.)

For those intent on making the ultimate chassis, one way to go is to make it out of aluminum bar and sheet stock. Having only flat panels certainly makes fabrication easier, and it allows you to use 6061-T6 stock, which machines quite well, for the panels. In other words, you can drill nice, clean holes without too much work, though the result isn't quite as clean as holes drilled in cast aluminum. (You may want to think twice about using aluminum alloys such as 3003. While they are easily formed, their softness sometimes makes them difficult to deburr, particularly if you get lazy and don't use sharp drill bits.) Don't try to bend 6061-T6, even if you have a sheet metal brake, as this alloy will crack, being too brittle for forming.

A difficulty with tempered stock such as 6061-T6 is that holes are more difficult to punch through the material. But there are ways of making the task easier. First, don't forget to use some sort of oil or cutting fluid to reduce friction. This makes a big difference. Also, remember that punching metal deforms it—less pressure is required if the stressed metal in the cutout has someplace to go. The easiest way to ensure that it does have a place to go is to drill a set of strain relief holes on the inside of the circle to be punched. I suppose you could also make progressively larger holes, but this seems to be a lot more work. To ensure that the final hole is truly in the right place, you can either make centering lines or use the big, final-size punch to make some indentations that help line things up. I find that the thin rings of material that result from using progressively larger punches are easier to remove from the punch than a solid blank. Finally, you can increase the torque by using a cheater bar or simply a piece of steel pipe that extends the length of your wrench. I don't want to get letters reporting what you have broken by overdoing it, though, so use caution!

As the cover photograph indicates, I used small-diameter Teflon-dielectric coax (RG-188 or similar) instead of cheaper RG-174, with its polyethylene dielectric that melts easily. Since the coax lengths involved are so short compared to a wavelength, it's unlikely that the exact impedance of the coax is important, so use what you have available. (If you do come across 25-ohm coax, though, save it for winding broadband matching transformers.) By the way, contrary to popular belief, you do not need lots of expensive test equipment to figure out the impedance of coax. In many cases, you can get pretty close by comparing it against known 50-ohm coax of the same dielectric type and guessing. After all, there really aren't that many impedances (25, 50, 52, 70, 75, 93 and 125 ohms, the 25 and 125 being awfully rare—don't ask me where to get it!) Of course, you *will* need test equipment if you don't know that the impedance goes up as ratio of the diameter of the outer conductor to that of the inner conductor goes up. What I *would* skip is coax with an aluminum shield. While some people insist that it can be soldered, why bother if you can get tinned or copper-shielded coax? Then again, if you find that your shiny coax braid won't take solder no matter how clean it is, perhaps you bought coax with an aluminum shield by mistake. Oh, well...

For audio cables, the preferred transmission line is twisted pair, rather than coax. Twisting the wires alternates the sense of the field pickup in the wire, causing the 60-Hz hum to cancel itself out. This is much more cost-effective than buying the special metals needed to get low-frequency magnetic shielding. (Even the ARRL Lab's expensive screen room does a poor job of low frequency shielding—I've picked up all sorts of noise on VLF inside it.)

Also, you might pay attention to where you place the RF amplifier. You want to put it close to the RF gain control; short leads for this potentiometer are essential for the amplifier to work properly.

Perhaps the biggest challenge in building a VFO-tuned rig is mounting the tuning capacitor. Often, you want it to be shielded, although I found it wasn't necessary for this radio. The challenge is to mount it securely without putting a torque on its shaft and creating backlash. (Backlash occurs when the energy you put into turning the capacitor shaft deforms the capacitor instead of moving its plates the way you want them to go. Backlash is the culprit when you start tuning in one direction and the radio tunes a bit in the opposite direction first and then hurries to catch up, or when you let go of the knob and the radio tunes back the other way a little!) One help to avoiding backlash is to *not* insist on nice right angles in boxes and mounting brackets. Making boxes from copper-clad circuit-board material facilitates this—you just solder their pieces together in whatever alignment is needed to prevent the capacitor and drive mechanism from binding.

Shielding is essential when you use an air-core VFO inductor—toroids, which are largely self-shielding, aren't quite so fussy. But air-core inductors don't have ferrite slugs to degrade stability, and very stable oscillators can be built using them.

Setting up a VFO is almost always a problem because toroids aren't precise enough components. Expect to have to add or remove a few turns in the process of get-

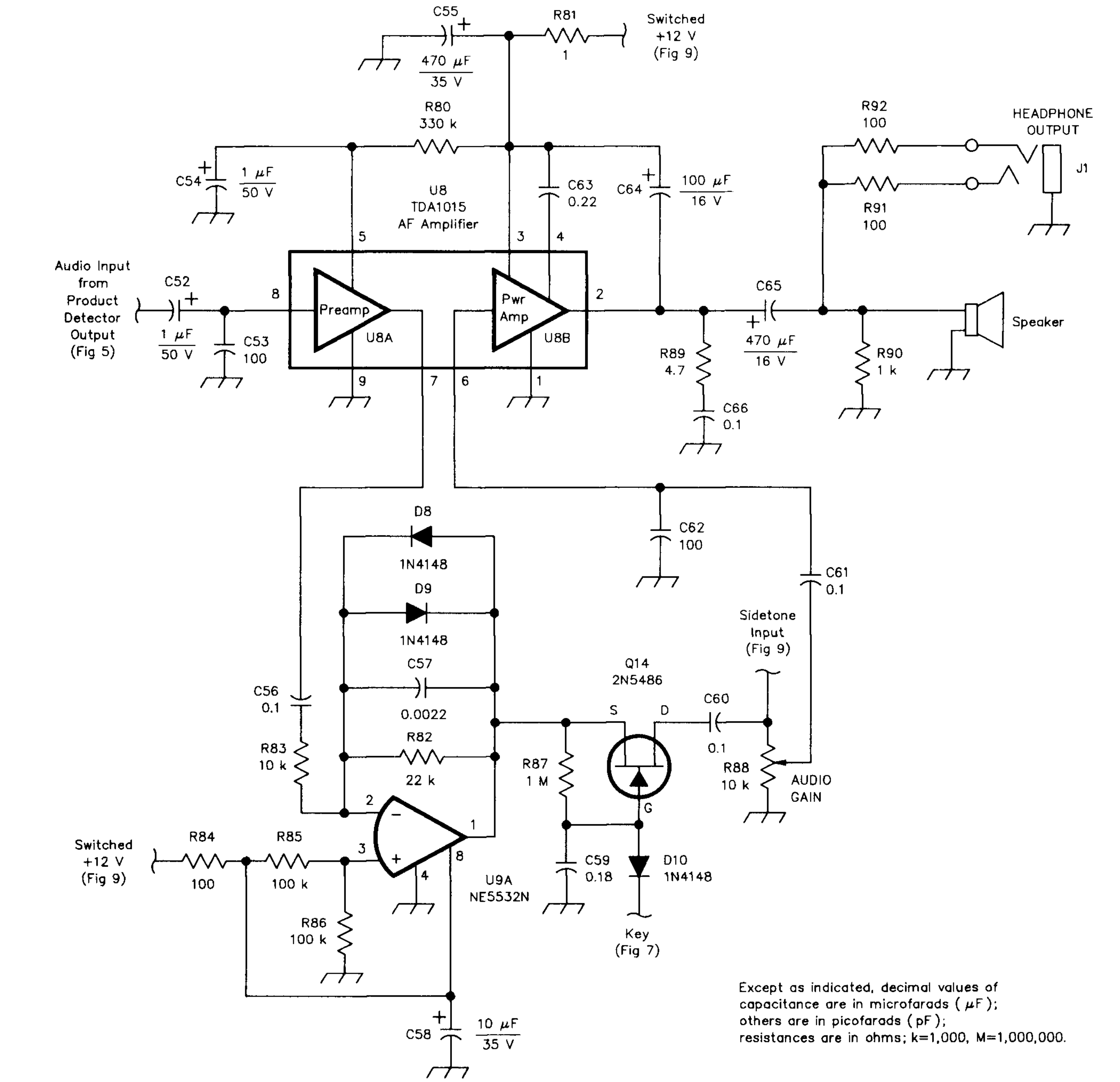


Fig 6— Audio circuits.

Q14—2N5486 JFET. 2N5484 or 2N5485 will work with slightly higher loss in the audio path.

U8—TDA1015 audio amplifier.

U9—NE5532N dual audio operational amplifier.

ting the VFO to cover the frequency range you want. A little fine tuning can be obtained by compressing the turns of the toroid coil to increase the inductance. A quick test with 12 turns of #28 wire on a T-25-6 core showed the apparent inductance going from 0.40 to 0.49 μH, with the Q dropping from 156 to 154 when the turns were compressed. The stray capacitance also changes, which is why I said the *apparent* rather than the actual inductance, but this distinction is rarely important for the cut-and-try experimenter. It's a bit more important for those trying to model circuits on computers. If you really want to figure out the stray capacitance, measure the inductance at several frequencies and calculate, keeping in mind that the permeability of some ferromagnetic materials is frequency sensitive.

Ground-plane construction is superior to printed circuits when "cutting and trying"—you can easily damage a printed circuit board by repeatedly unsoldering parts. Incidentally, the biggest cause of foil lifting from printed circuit boards is excessive temperature. If you have a temperature-controlled iron, set it to no more than 750 degrees. With ordinary irons, you will have to practice to see what the ideal warm-up time happens to be for your iron. What you should *never* do is let the iron sit around for half an hour and then use it to heat up a tiny circuit board pad.

One of the advantages to home brewing is that you get to optimize the circuit to meet your needs. Such personalization is easily done with the VFO. There are different ideas on how best to set up your tuning mechanism. Perhaps the simplest is just to cover a small portion of the band, say from 7.030 to 7.050. This range covers most of the QRP activity, for instance, and is the cheapest to implement. If you want to cover all of the CW band, the usual approach is to install a reduction drive to slow down the tuning rate to 50 kHz/turn or less. This invites mechanical complexity, so the "all thumbs" builder should probably avoid this rather than end up with a backlash-ridden drive. The compromise approach is to use two knobs, one for coarse tuning and the other for fine tuning. This actually makes a lot of sense for many amateurs. For instance, when you move frequency to dodge interference it's normally a few kHz from your current operating frequency, something easily done with a vernier control. On the other hand, someone trying to spot specific frequencies may get frustrated by a two-control system, unless a frequency counter is used.

While you could haul out the equations, here are some simplified rules for getting the tuning range you want. First, use an ordinary tuning capacitor, avoiding those with the unusually shaped plates. For small tuning ranges, such as a single amateur band, those special capacitors work worse, as they are designed to cover wide frequency ranges (such as 2 or 3 to 1). In general, the more capacitance you use to get the circuit tuned to resonance, the less range your tuning capacitor will have. Conversely, the more inductance you use the more range your tuning capacitor will have. So, if you try out your VFO and you want to cover a slightly larger chunk of the band, add a few turns to L1 and reduce the capacitance across the inductor.

The question always comes up as to what type of capacitors to use for building VFOs. The best capacitors to use are air variable ones, but they are often too expensive and bulky to be used where you just need a fixed capacitance. For fixed capacitors, NP0 capacitors seem to be a good choice, and can often be found in purchased capacitor assortments. I recommend you buy at least one of those assortments, especially if you can find one with lots of temperature compensating capacitors. These either have dots of paint on the top, or markings like *N220*. These tough-to-find parts are invaluable for getting your circuit temperature stable, though it's a time-consuming process that you may not feel is necessary. Most of these capacitors have negative coefficients, since they are made to compensate inductors, which commonly have positive temperature coefficients. You can also use polystyrene capacitors, but I avoid them, since they are so easily destroyed by errant soldering irons and solvents. (The errant soldering iron may be the biggest disadvantage to ground-plane construction. A drilled circuit board does work as a good shield to keep the parts and the hot iron separated.) Silver-mica capacitors can also be used, but they generally aren't as predictable as NP0 capacitors. Of course, if you get all your parts at bargain-basement prices, do keep in mind that some of the parts you buy may simply be of poor quality or mislabeled. (There are sometimes good reasons why something is unbelievably cheap.) I suspect the temperature-compensating capacitors are cheap because of the difficulty in selling them—imagine the space you need to carry all the different values and temperature coefficients, not to mention the time it takes for someone to need a particular value and coefficient.

I've also noticed that some of the beautiful-looking monolithic capacitors—the ones with the shiny epoxy coating—aren't particularly rugged when it comes to removing them from circuits. Their internal terminations seem to become unsoldered. While they are cheap enough to just throw away, you certainly don't want a broken or worse, intermittent unit in your circuit.

If you do choose to temperature compensate a VFO, keep in mind that the soldering iron affects the warm-up drift of the VFO. While I don't have any good numbers on how long you should wait for things to cool off, the way I do it is to make measurements before and after work, giving the circuit many hours to cool off. Also, keep in mind that the warm-up drift, the drift from turn-on to settling down, may differ considerably from the drift after things have "stabilized." It is possible for the circuit to be quite stable for the first few minutes, and then start to creep upwards in frequency, seemingly forever. This is not what you want! Also, the drift is sometimes temperature dependent—it will work fine at 70 degrees, for instance, and be pretty lousy when you show it off during Field Day when it's 90 degrees in the shade. The way to avoid such problems is to get the most stable parts you can, rather than trying to correct a highly temperature-sensitive coil with equally sensitive temperature-compensating capacitors.

I've avoided the use of a potentiometer for my main tuning knob. While they do work, I use that knob so much that wearing it out is quite likely. Worn out volume controls are quite common in "old style" non-digital radios and TV sets—and those really don't get constant use like a tuning control on an HF amateur transceiver. Trimmer potentiometers often have ratings of 100 cycles or less.

When laying out VFO parts, it's helpful to remember that high-Q tuned circuits often have high circulating currents. What you want to do is to minimize the path that these currents take, both to prevent radiation and to prevent unwanted signals from getting in and disturbing things. What this means is that you usually want to strap as much as your capacitance as possible directly across the inductor leads. This is best done by mounting the largest capacitor in the best spot and mounting the smaller ones further away. This also applies to other circuits, such as band-pass filters. If you do a decent job, you can

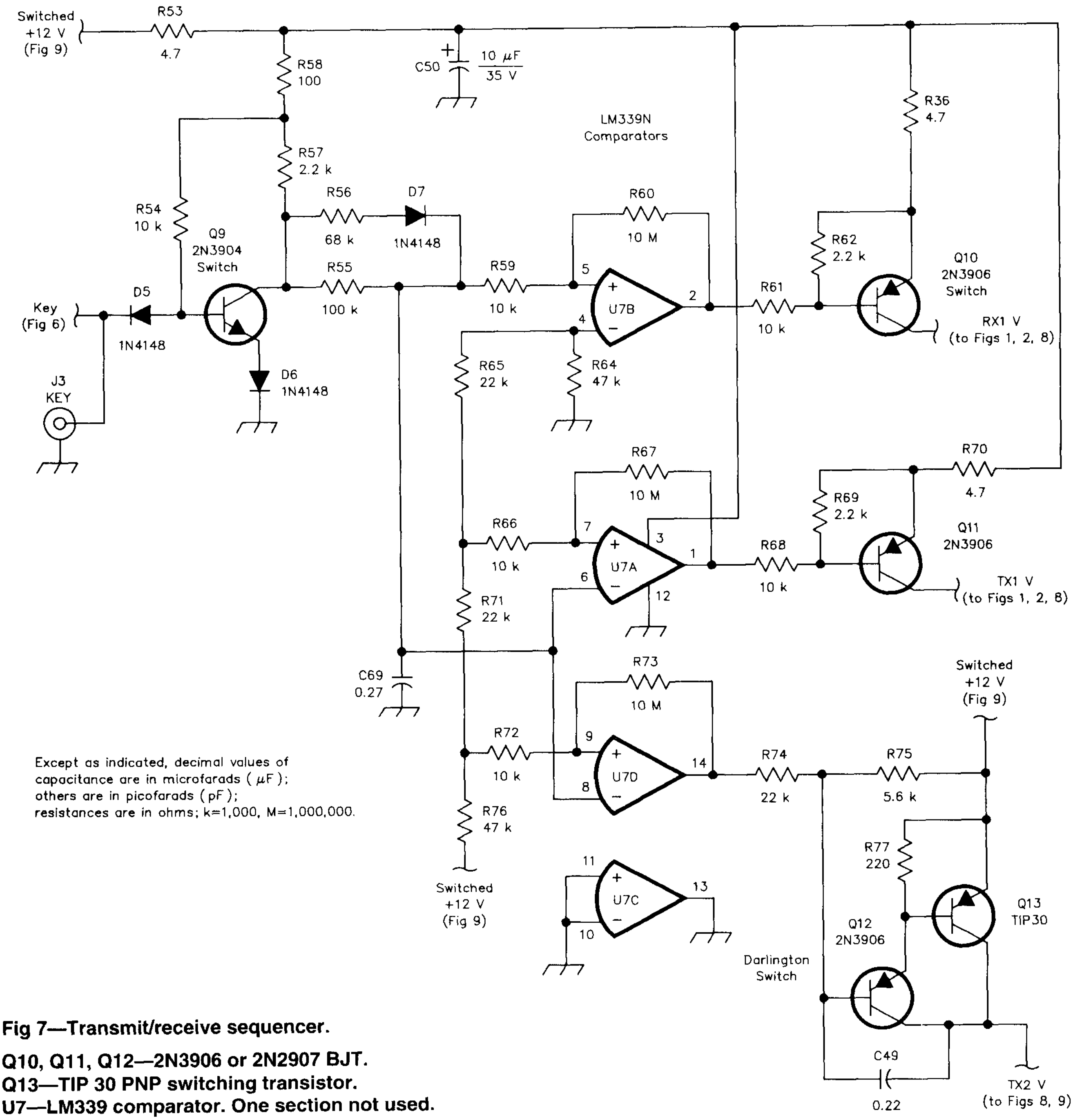


Fig 7—Transmit/receive sequencer.

Q10, Q11, Q12—2N3906 or 2N2907 BJT.
Q13—TIP 30 PNP switching transistor.
U7—LM339 comparator. One section not used.

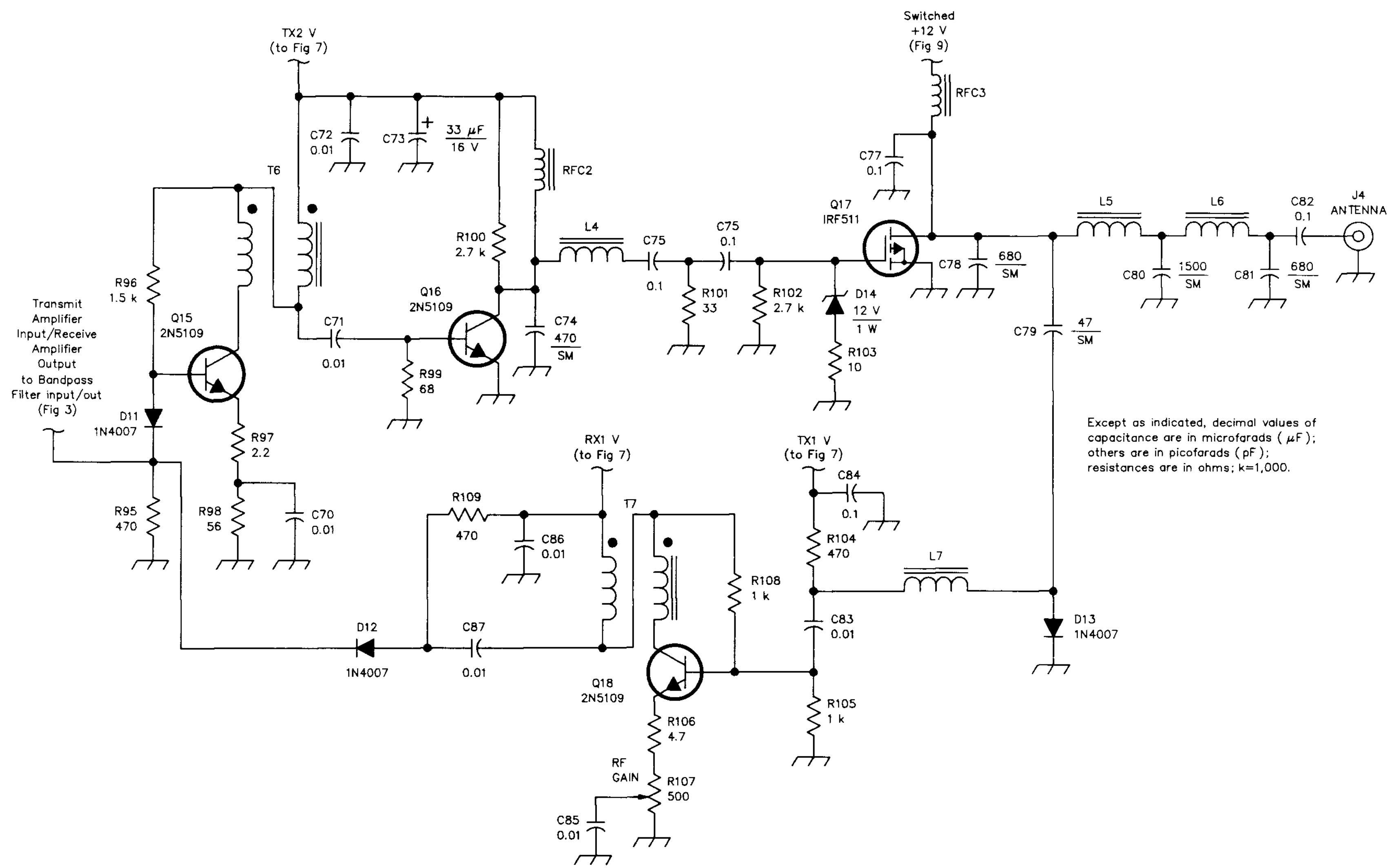


Fig 8—Transmitter amplifiers and receiver RF amplifier circuits.

D14—1N4742. 12-volt, 1-watt zener diode.
L4—12 turns #28 enameled wire on a T-44-6 toroid core.
L5—12 turns #26 enameled wire on a T-30-6 toroid core. (0.68 µH)
L6—15 turns #26 enameled wire on a T-37-6 toroid core. (0.88 µH)
L7—42 turns #28 enameled wire on a T-50-2 toroid core. (10.6 µH).
Q17—IRF510, IRF511. The IRF510, available from Radio Shack, is slightly more efficient.
RFC2—8 turns of #28 enameled wire on an FT-37-61 ferrite core.
RFC3—12 turns of #24 enameled wire on an FT-37-43 ferrite core.

often minimize the shielding needed to isolate the individual circuits from each other. Building this shielding is a time-consuming task, and shielding often makes circuit modification more difficult.

The 7-MHz band-pass filter shown in Fig 3 is pretty easy to align—if you build the transmitter you should be able to set it up by tuning for "maximum smoke" in the middle of the passband (tune your power meter for a maximum reading). It should then be tuned up for receive as well. In the process of developing a simple design I discovered that tapped inductors often have more loss than those with link coupling. But who wants to deal with all those leads, anyway? Hence the design with simple coils and capacitive coupling, even though capacitors are more expensive than wire.

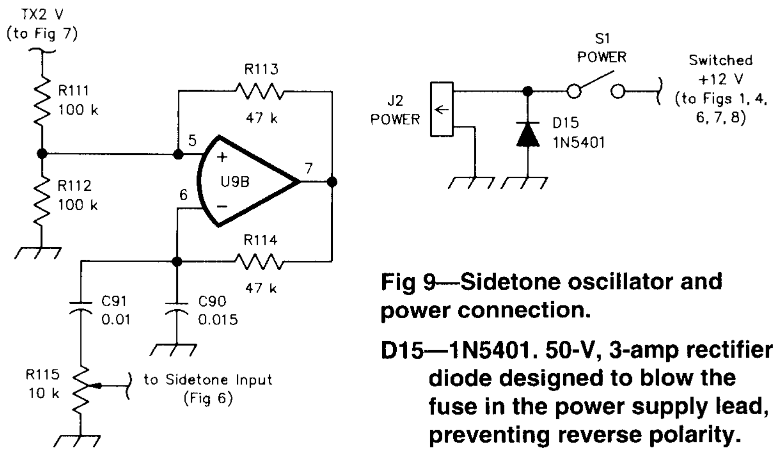


Fig 9—Sidetone oscillator and power connection.

D15—1N5401. 50-V, 3-amp rectifier diode designed to blow the fuse in the power supply lead, preventing reverse polarity.

The tricky circuits to align are the two oscillator trimmers. What you want to do is to set up the beat frequency oscillator so that when a signal is in the middle of the crystal filter passband, it comes out on your favorite CW tone's frequency. Then, the carrier insertion oscillator is set up so that you also hear that exact tone. If you have an attenuator or don't mind building one, one way to set things up is to feed the carrier oscillator into the receiver through the attenuator and then adjust it for maximum audio output. Then the beat frequency oscillator is adjusted for the proper pitch or frequency. You might have to go back and forth a few times to get everything really set right.

For those experimenters who are looking to test their crystal filters without a whole lot of expense, there is a way to do it if you have a way of measuring frequencies accurately. One of the oscillators can be used as a signal source. Add 20 dB or so attenuation to avoid damaging the crystal filter and add some inductance in series with the oscillator's crystal to generate lower frequencies. To measure the power output, make a logarithmic detector using one of the FM receiver chips that features a signal strength indicator. (The October 1988 issue of *QEX* shows an example using the Motorola MC3356P receiver chip.) These chips are temperature-sensitive, so your measurements may vary between winter and summer if you work in an unheated basement. A frequency counter easily measures the frequency, if you have one. With a bit more work, a modern general-coverage receiver could also be used.

To keep the VFO stable despite changes in supply voltage, a 78L05 regulator is used. Bypassing is recommended by the manufacturer on the input for stability, while the bypass capacitor on the output suppresses the noise generated by the regulator. Zener-diode regulators tend to be pretty noisy unless well bypassed, so one wasn't used here. (A 6.8-volt Zener is actually used as a noise generator in the *ARRL Handbook's* noise bridge.)

An interesting problem is what sort of dc power connector to use with QRP equipment. Phono, or RCA, connectors have serious problems, being easy to short out if you aren't careful. ARRL's Field Services Department has published a 12-volt-connector recommendation that appeared in the newsletter sent to affiliated clubs. A difficulty with this standard is that it is *almost* the same as that on the popular Heathkit HW-9, the difference being that the positive and negative leads are reversed! I've had a report of significant damage to an HW-9 from reverse polarity. I strap 3-A diodes across the power leads to prevent problems like this. In any case, if you are building portable equipment, miniature Molex connectors may make more sense than the recommended one.

You need to match the crystal frequencies, ideally within 50 Hz. Normally, a lot of 10 crystals should be sufficient to get the needed 6 matched crystals. One way to measure the crystals is to put them in an oscillator circuit, such as either the carrier insertion oscillator or beat frequency oscillator, then measure the output frequency. It should be possible to match the crystals using a partially built transceiver. You can use the receiver to listen to the relative frequencies of the carrier insertion oscillator. To avoid the "both sides of zero beat" problem, you probably want to make sure the crystals are matched at both a high and a low pitched tone. The problem is that a crystal 1 kHz above the BFO sounds just like one 1 kHz below the BFO—you could get a crystal that is off by twice the audio tone frequency used. Running a quick check with another audio tone eliminates this ambiguity. □□

The Double-Diamond Quad

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Introduction

After getting married in 1991, I moved into the small house which my wife had already been renting. In the center of the back yard I put up a 65-foot mast composed of several "recycled" sections of TV tower, with a length of steel pipe on top for good measure. A variety of single and multiband inverted Vs were hung from the top of this mast, with good success, but I had been searching for a high-performance wire antenna which could be used for working DX on 20 meters. This article describes the results of some computer-modeling studies which I have performed using *ELNEC*.[1]

Background

The inverted V is one of the most widely used antennas in Amateur Radio today. It is simple, easy to build, and easy to get working. It provides a fair amount of gain (compared to an isotropic source) and can be adjusted for a near-perfect impedance match to 50-ohm coaxial cable. Fig 1 is a plot of the broadside radiation pattern for a 20-meter inverted V, when mounted on my mast at an apex height of about 65 feet. The legs slope downward at a 45° angle, and their length has been adjusted to produce an input impedance which is almost purely resistive. Notice that there is a good low-angle bidirectional lobe at a take-off angle of 16°, with a maximum gain of about 6.5 dBi. Unfortunately, there is also a lot of

[1]ELNEC is available from Roy Lewallen, W7EL, PO Box 6658, Beaverton, OR 97007.

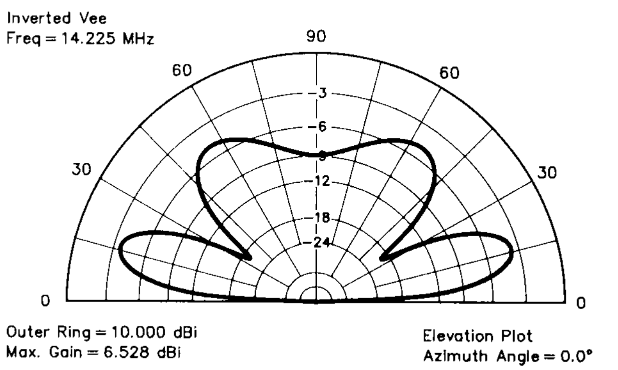

Fig 1—Broadside radiation pattern for a 20-meter inverted V with apex at 65 feet. (Beamwidth=16°, Take-off angle=16°.)

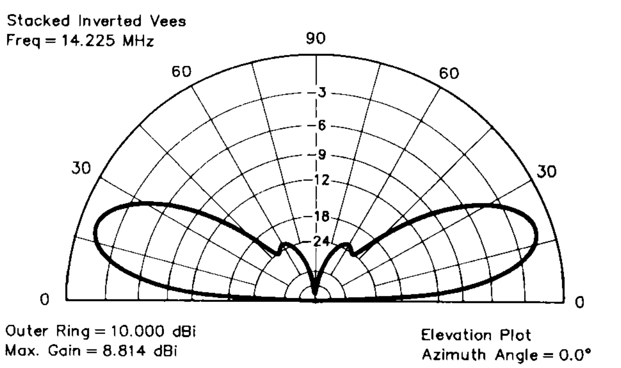

Fig 2—Broadside radiation pattern for stacked 20-meter inverted Vs in phase. (Beamwidth=21°, Take-off angle=19°.)

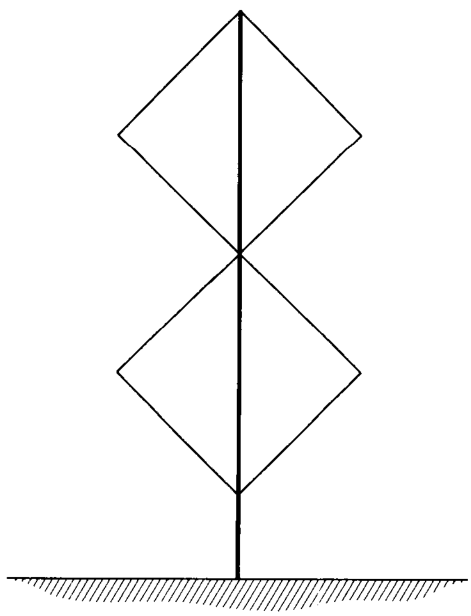

Fig 3—Physical configuration of the double-diamond quad antenna.

high-angle radiation from this antenna, which is not generally desirable for working DX on 20 meters.

Initially, I considered stacking two inverted Vs, one above the other, with half-wavelength spacing between them. Fig 2 shows the performance of such an array when the two radiators are fed in phase with equal-amplitude currents. As you can see, the broadside radiation pattern looks very good, and there is a null directly overhead. On the computer, I adjusted the length of each inverted V to achieve resonance (zero reactance), and found that the two driving-point impedances were different. I got a value of about 35 ohms for the upper antenna, and a value of around 44 ohms for the lower inverted V, which is slightly longer. This means that it would be necessary to design suitable passive networks in order to achieve ideal current drive to each of the two radiating elements. However, I wanted an antenna that was really simple, preferably a structure with only one feed point, and I didn't want to build any phasing or matching networks.

The Double-Diamond Quad

The antenna which is the subject of this article consists of two diamond-shaped quads mounted one above the other, as drawn in Fig 3. At the point where the two diamonds intersect, the diagonal wires simply cross over each other like a giant X. These two wires do not touch, so they must be insulated from one another, or spaced slightly apart.

This antenna system may be fed either at the top of the upper diamond or at the bottom of the lower one. In either case, the input impedance can be made purely resistive (no reactance) by pruning the overall wire length. According to *ELNEC,* the resulting impedance values will be about 315 ohms (top feed) or 319 ohms (bottom feed), so the installation of a 6:1 balun at either of these two possible feed points should provide a good match to 50-ohm coax. For the computer model, each side of each diamond was 18.46-feet long, and #12 AWG copper wire was used.

The double-diamond quad radiates a horizontally polarized signal when fed as described above. Fig 4 illustrates the elevation pattern perpendicular to the plane of the antenna (broadside) when bottom feed is used. The main lobe is very large and covers a much wider range of take-off angles than the inverted V of Fig 1. In addition, the maximum gain of the DDQ is nearly 2.5 dB higher than the single inverted V (8.9 versus 6.5 dBi), and the high-angle radiation produced by the inverted V is almost completely absent. Fig 5 shows that there is very little radiation in the plane of the antenna. This is confirmed in Fig 6, which is a plot of the azimuthal-plane radiation pattern at a take-off angle of 21°. Notice that the front-to-side ratio is better than 18 dB, while the half-power beamwidth is 84°.

If the DDQ is fed at the top, the radiation patterns which result are shown in Figs 7-9. When comparing top feed versus bottom feed, it can be seen that the performance is very similar in most respects. The only marked

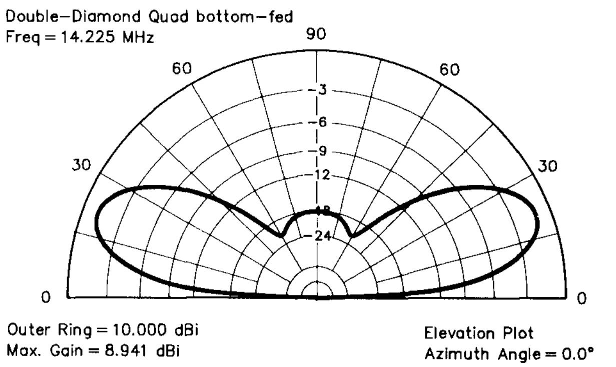

Fig 4—Broadside radiation pattern for the double-diamond quad antenna when fed at the bottom of the lower diamond. (Beamwidth=24°, Take-off angle=21°.)

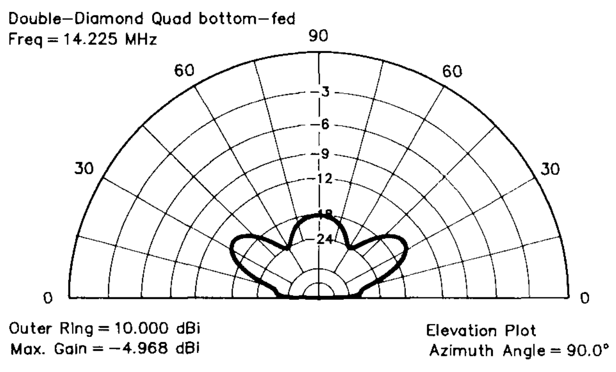

Fig 5—Radiation pattern for the double-diamond quad in the plane of the antenna when using bottom feed.

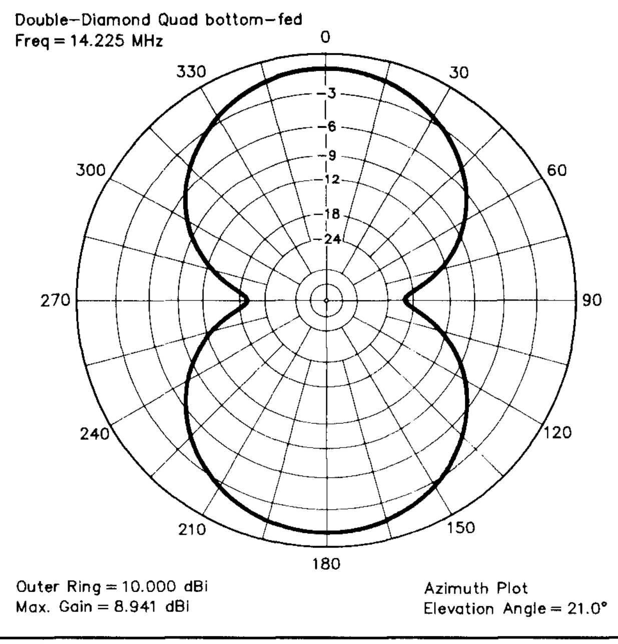

Fig 6—Azimuthal-plane radiation pattern for the double-diamond quad antenna at a take-off angle of 21° when fed at the bottom of the lower diamond.

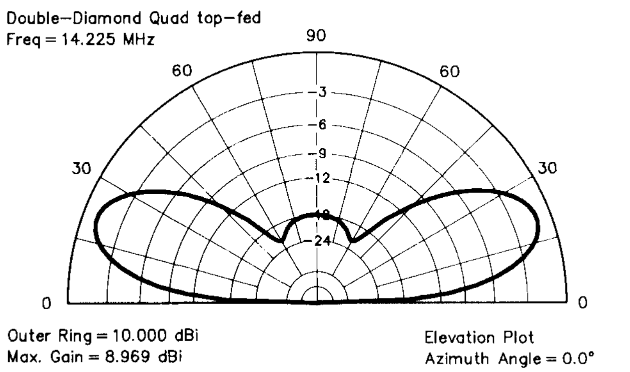

Fig 7—Broadside radiation pattern for the double-diamond quad antenna when fed at the top of the upper diamond. (Beamwidth=24°, Take-off angle=21°.)

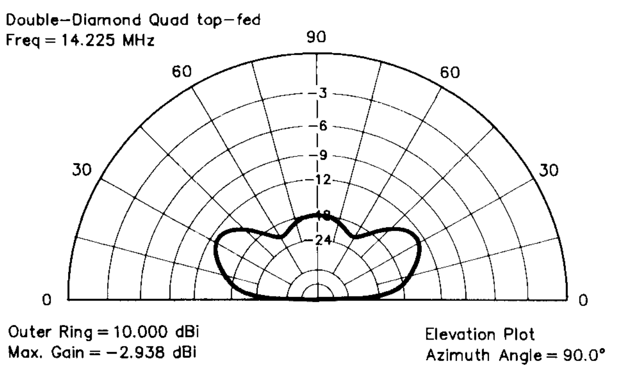

Fig 8—Radiation pattern for the double-diamond quad in the plane of the antenna when using top feed.

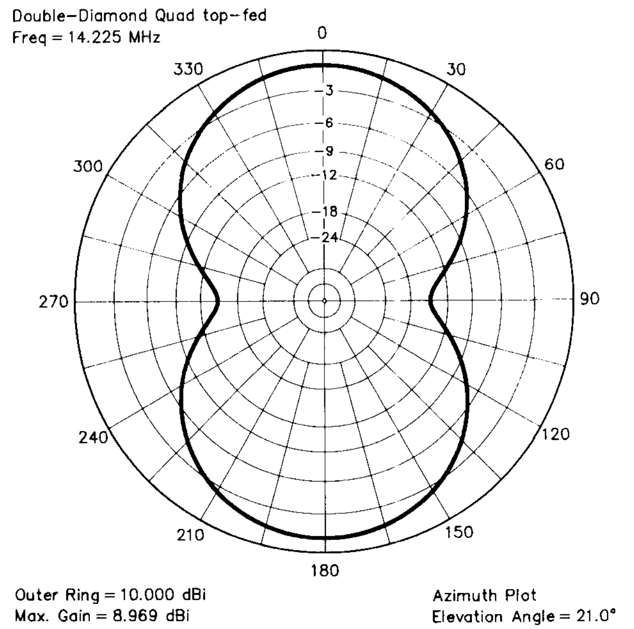

Fig 9—Azimuthal-plane radiation pattern for the double-diamond quad antenna at a take-off angle of 21° when fed at the top of the upper diamond.

difference is that the front-to-side ratio deteriorates by about one S-unit when the DDQ is fed at the top.

Modifications and Additions

If coverage of *all* compass directions is desired, it appears that a second DDQ could easily be mounted at right angles to the first, with switching provided by a simple relay box. Since the azimuthal beamwidth is nearly 90°, there would be only a small reduction in gain in those directions which are midway between the two main beams.

For unidirectional coverage, a second DDQ element could be placed about ⅛ wavelength behind the first. This second element could be actively driven, as in a phased end-fire cardioid array, or it could be adjusted for parasitic operation. The parasite could be made to function as either a reflector or director by switching an appropriate value of reactance into the circuit.

The double-diamond quad antenna may easily be modified to operate on 40 meters or one of the WARC bands, if desired. With a side length of 18.46 feet, the DDQ described here is resonant near 14.225 MHz, and has a total height of about 52.2 feet, so a DDQ for some other band can be scaled proportionally by using these dimensions as a starting point.

Conclusion

This article has described a simple high-performance wire antenna that provides good gain at low take-off angles while minimizing cloud-burner radiation. The double-diamond quad requires only one tall support, and presents an input impedance value which is easy to match to 50 ohms. It is the author's hope that this short presentation will stimulate other hams to experiment with wire antennas. if you build a DDQ, please write and let me know how you like it! □□

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A No-Tune Driver for 432 MHz

In looking at the September 1991 *QST* write-up of a 432-MHz transverter by Ed Krome, KA9LNV, I noticed he uses a Toshiba SAU-4 hybrid power amplifier module fed from a tuned MRF911 driver stage. While this works fine, it requires tuning of the driver. Since the power amplifier module is no-tune, it caused me to wonder if it wouldn't be useful to have a no-tune driver stage as well. Cheap plastic-case MMICs, such as the Mini-Circuits MAV-11, don't really provide enough power alone to drive this module properly, though they will work if you keep the system output power down to 10 watts PEP or less. The MRF559 driver described here will produce 15 to 17 watts out before high-order IMD performance starts to really degrade.

The heart of no-tune designs is in the use of microstrip tuned circuits. Use of microstrip makes the built-in tuning dependent only on component tolerances and the accuracy with which the circuits are etched. But the challenge of low-frequency microstrip lies in keeping the line sizes manageable. A full quarter wavelength at 70 cm on G-10 epoxy board is about 3.4 inches, a little larger than desirable for use in a low-power amplifier. (It's fine for a high-power transistor amplifier—the heat sink prevents miniaturizing the amplifier anyway.) With that in mind, I used relatively small matching sections for most of the design, although the MRF559 input network is nearly full size. An added complication was that I want-

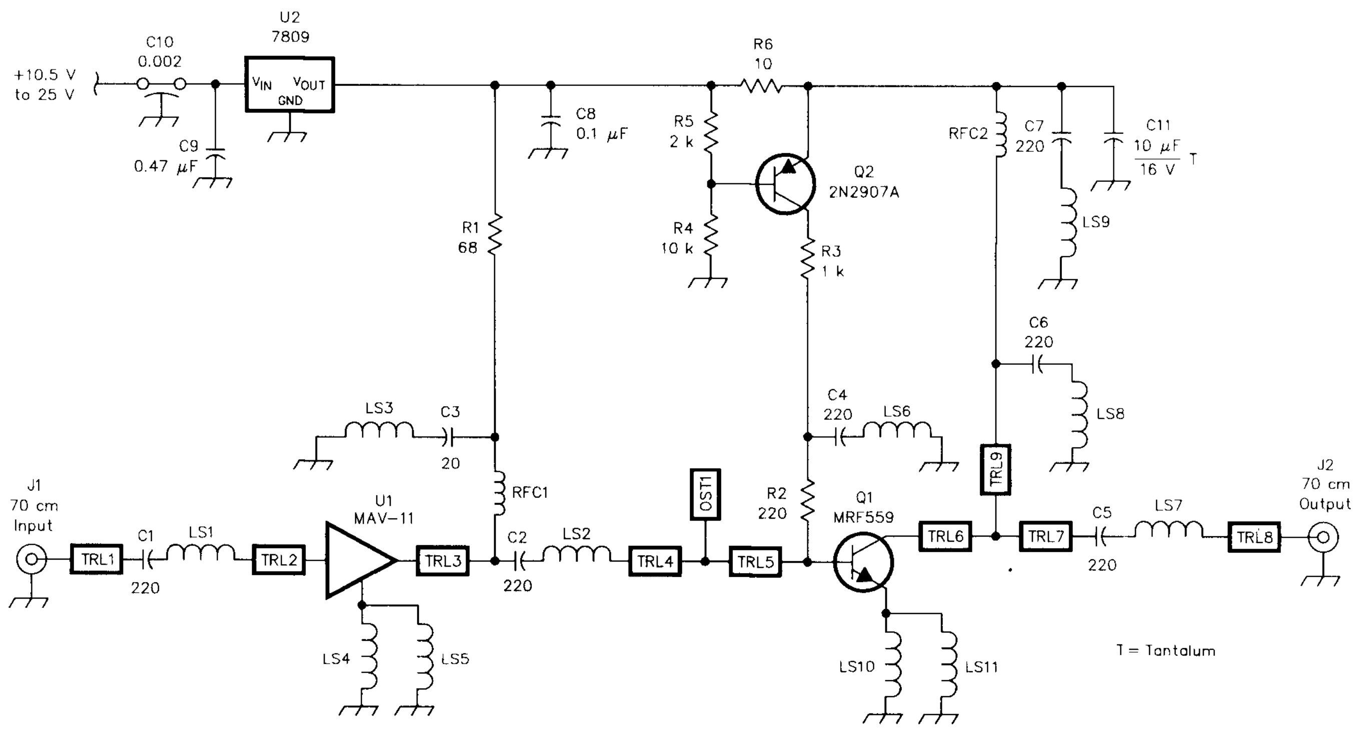

Fig 1—Schematic of the 432-MHz, no-tune amplifier.

J1, J2—Coaxial connectors. I use SMA connectors but it may be more practical to use direct coax connections.
LS1-11—Stray inductances useful for computer modeling.
Q1—Motorola MRF559.
Q2—2N3906, 2N2907, or 2N2907A.
R1—68-Ω, ½-watt resistor. A quarter-watt resistor can be used but will run a bit warm.
RFC1,2—10 turns no.28 enameled wire, 3/16 inch ID, closewound.
U1—Mini-Circuits MAV-11 or Avantek MSA-1104.
U2—7809 9-volt regulator, available from:
Ocean State Electronics
P.O. Box 1458
6 Industrial Drive
Westerly, RI 02891
1-401-596-3080
1-800-866-6626 (order line)

ed the bandwidth of the amplifier to be as large as possible in an attempt to minimize the effect of variations in circuit board fabrication and components.

To minimize the effect of the biasing circuitry on the RF performance, I chose to make my dc connections at low impedance points of the circuit—the base of the transistor and the short-circuit side of an RF-bypassed stub. This should also reduce the amount of radiation from the board, an important consideration if you are too lazy to mount everything in its own shielded box!

Since portable operation is a possibility, I wanted this circuit to run properly even if the battery voltage is a bit

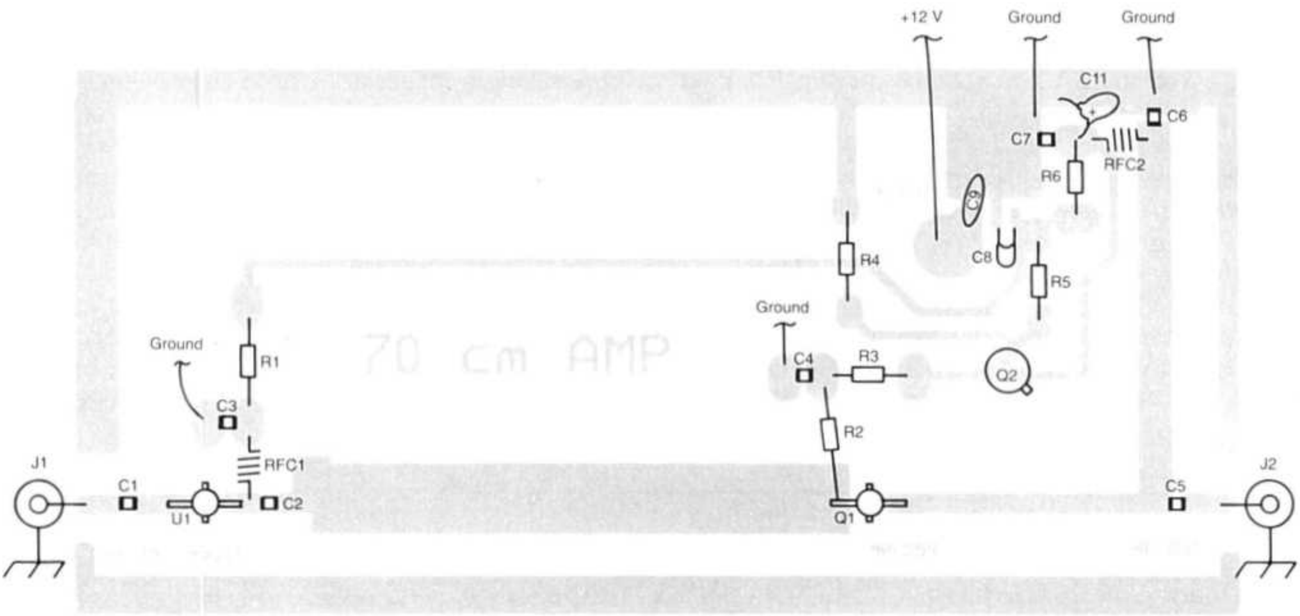

Fig 2—Parts-placement diagram of the amplifier.

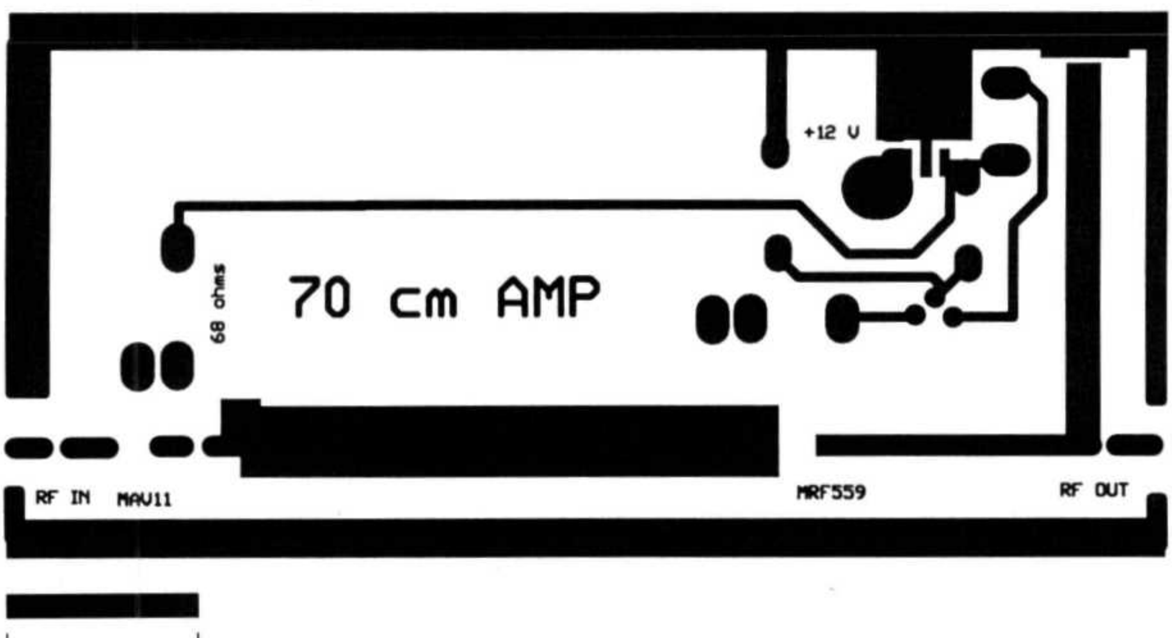

Fig 3—Circuit board layout of the amplifier.

low. This was accomplished using a 9-volt regulator from Ocean State Electronics (see Fig 1). (They also have those 10-volt regulators used to drive X-band Gunn diode transceivers.)

To ensure that the circuit works over a wide temperature range, I used an active bias circuit. The typical shunt diode circuit attempts to use a thermally coupled diode to compensate for the transistor's gain variation with temperature. The active bias circuit instead monitors the MRF559 collector current as it passes through a 10-Ω sensing resistor and adjusts the base current accordingly. The circuit should work fine from a supply voltage as low as 10.5 volts, though I haven't temperature tested it yet.

For those wishing to model the circuit on a computer, I have included a Microwave Harmonica linear analysis of this circuit (see Figs 4 and 5). Keep in mind that many of the parasitics included in the model are approximate, rather than rigorously measured quantities. Some

```
**430-450 MHz Broadband amp using the MRF-559

** 100 mA at 10 volts
blk

* input matching

w50:109mil
w19:370mil
win:?200mil?
pin:?250mil?
wout:?175mil?
pout:?2000mil?

  trl 1 2 w=w50 p=300mil sub1
  slc 2 3 l=0.5nh c=220pf
  trl 3 4 w=w50 p=200mil sub1
  tee 4 5 6 w1=w50 w2=w19 w3=win sub1
  trl 5 7 w=w19 p=?2800mil? sub1
  SRL 7 77 R=220 L=5NH
  SLC 77 0 L=0.5NH C=220PF
  ost 6  w=win p=pin sub1

  two 7 8 100 mrf559

  wrap 100 0 A=50mil Leads
  wrap 100 0 A=50mil Leads

  trl 8 9 w=w50 p=?1400MIL? SUB1
  tee 9 10 11 w1=w50 w2=w50 w3=wout sub1

  trl 11 22 w=wout p=pout sub1
  slc 22 0 l=0.5nh c=220pf

  trl 10 12 w=w50 p=100mil sub1
  slc 12 13 l=0.5nh c=220pf
  trl 13 14 w=w50 p=300mil sub1

  amp: 2por 1 14
end
freq
  step 250mhz 1500mhz 50mhz
  step 400mhz 500mhz 10mhz
end
data
 mrf559:motac file=\scompact\bank01\mot1.flp
 sub1: ms h=59mil er=4.8 tand=0.005
*guessed tand?
 Leads:ms h=59mil er=4.8 met1=SN 10mil
end
```

Fig 4—Microwave Harmonica source file used to analyze the MRF559 amplifier.

```
MICROWAVE HARMONICA PC V5.1B
File: \ws4\rf\432drv.ckt
15-JAN-93 02:49:11
```

Freq GHz	MS11 dB AMP	MS22 dB AMP	MS21 dB AMP	K AMP
0.250	-2.64	-1.76	14.47	1.1
0.300	-3.03	-3.03	15.83	1.1
0.350	-4.60	-5.52	16.98	1.1
0.400	-9.01	-11.81	17.52	1.1
0.410	-10.51	-14.25	17.48	1.1
0.420	-12.24	-17.59	17.39	1.1
0.430	-14.02	-22.31	17.24	1.1
0.440	-15.35	-25.51	17.04	1.1
0.450	-15.46	-21.37	16.78	1.1
0.460	-14.38	-17.45	16.48	1.1
0.470	-12.86	-14.78	16.14	1.1
0.480	-11.37	-12.89	15.76	1.1
0.490	-10.08	-11.49	15.37	1.1
0.500	-9.01	-10.42	14.95	1.1
0.550	-5.75	-7.74	12.81	1.1
0.600	-4.31	-7.09	10.86	1.1
0.650	-3.60	-7.38	9.26	1.2
0.700	-3.24	-8.34	8.03	1.2
0.750	-3.08	-10.05	7.15	1.2
0.800	-3.03	-12.94	6.58	1.2
0.850	-3.04	-18.29	6.28	1.2
0.900	-3.05	-25.40	6.17	1.2
0.950	-3.03	-16.57	6.22	1.1
1.000	-2.97	-11.70	6.40	1.1
1.050	-2.93	-9.17	6.75	1.0
1.100	-3.03	-8.05	7.33	1.0
1.150	-3.41	-8.04	8.15	1.0
1.200	-3.96	-7.10	8.64	1.0
1.250	-3.10	-3.13	6.93	1.0
1.300	-1.82	-1.05	2.52	1.0
1.350	-1.30	-0.42	-2.94	1.0
1.400	-1.13	-0.21	-9.26	1.2
1.450	-1.10	-0.13	-18.53	2.8
1.500	-1.15	-0.12	-30.85	26.7

Fig 5—Microwave Harmonic output file.

Table 1—Measured Performance

MRF559 gain stage at 432 MHz

P_{in} (dBm)	P_{out} (dBm)
–6.0	9.9
–4.0	12.0
–2.0	14.0
0.0	15.9
6.0	21.5
8.0	22.9*

*1-dB compression point

Note: With –8 dBm of input, the output of the MRF559 gain stage drops from +8.0 dBm at 420 MHz to +7.7 dBm at 450 MHz. The output intercept of the single stage was measured at +32 to +33 dBm.

Two-stage amplifier with a MAV-11 driver, –10 dBm input

Frequency	P_{out} (dBm)
350 MHz	19.0
400 MHz	18.0
420 MHz	17.6
450 MHz	16.7

Notes: The output intercept measures +33 to +36 dBm. It gets better as you reduce output from the 1-dB compression point. The two-stage amplifier actually measures better than the single stage amplifier. A second sample of the two-stage amplifier showed a +22-dBm 1-dB compression point and a +34-dBm output intercept. The 1-dB small signal bandwidth was 322 to 476 MHz. Changing to a different MRF559 resulted in a +32-dBm output intercept.

parasitics, such as the loss of the coupling and bypass capacitors, do seem to be negligible. For instance, I changed C5 and C6, which see the highest RF currents, from cheap NP0 chip caps to 100-mil ATC chip capacitors without seeing a significant change in performance. The MAV-11 input MMIC isn't included in the model, since its small-signal characteristics are well known.

Construction

The board is standard 1/16-inch-thick FR-4 or G-10 glass epoxy. All parts are soldered to the top of the board. The MRF559 and the MMIC are mounted by drilling small holes in the board, bending the grounded leads against the body of the device (you only get to bend them once) and soldering these leads to the ground-plane foil. The connections from the pads to the ground are made by "Z-wires." (Actually, I find it easier to make them "U" shaped.) These are just resistor leads soldered from the top foils to the ground foil. I used half-inch brass sheet stock to make an enclosure that could be shielded.

The amplification available from this circuit, shown in Table 1, is sufficient to drive the SAU-4 or perhaps your favorite 70-cm power amp circuit. □□

Correction

Author Frank Morrison, KB1FZ, "The Magic of Digital Filtering," February 1993 *QEX*, sends the following correction. Page 8, the second equation under *Band-rejection impulse response (weighting) sequence*, should read:

$$+2\cos\left[\frac{\pi}{N}\left(\frac{(N-1)}{2}+L+1\right)\left(n-\frac{(N-1)}{2}\right)\right]\frac{\sin\left[\frac{\pi}{N}\left(\frac{(N-1)}{2}-L\right)\left(n-\frac{(N-1)}{2}\right)\right]}{\sin\left[\frac{\pi}{N}\left(n-\frac{(N-1)}{2}\right)\right]}$$

Bits

Central States VHF Society—Call for Papers

The Central States VHF Society announces a call for papers for their annual conference. The conference is scheduled for July 29-31 and August 1, 1993, at the Lincoln Plaza Hotel in Oklahoma City. Emphasis this year is on the beginner to weak-signal VHF and above operating, with a special session devoted to this subject. If you are interested in presenting a paper or having a paper published in the *Proceedings,* contact Tommy Henderson, WD5AGO, 12476 E 13th, Tulsa, OK 74128, phone 918-438-0099. Authors who have not been previously published are especially encouraged to submit their papers. Papers should be sent to Tommy and must received by **June 1, 1993** for inclusion in the Proceedings.—*Joseph L. Lynch, N6CL, President, CSVHF Society* □□

1993 West Coast VHF/UHF Conference

This year's West Coast VHF/UHF Conference will be held May 21-23 at the Ventura Holiday Inn in Ventura, California. Registration begins at noon on Friday. Technical talks and expanded vendor exhibits will be held all day Saturday, along with noise-figure measurements, a banquet (with speaker) and awards ceremony. The conference concludes Sunday with a breakfast (with speaker), an indoor swap meet and antenna range measurements. Plans also call for a codeless Technician class during the weekend.

General admission will be $15. Reservations are required for the Saturday night banquet ($25) and the Sunday breakfast ($10). Proceedings will be offered to conference attendees at the special price of $10. Rooms are $58 (contact the Holiday Inn at 1-800-842-0800). The Technician class is $125 (contact Loraine McCarthy, N6CIO at 714-979-2633).

For registration forms and more information, contact Steve Noll, WA6EJO, 1288 Winford Ave, Ventura, CA 93004. □□

ARRL Conference on Digital Communications

A call for papers has been issued for the ARRL Conference on Digital Communications (formerly the Computer Networking Conference). Deadline for receipt of camera-ready papers is July 30, 1993. Hosted this year by the Tampa Local Area Network, the conference has been tentatively scheduled for September 11, 1993, at the University of South Florida in Tampa. Technical papers for the Conference on Digital Com-munications can be on any aspect of digital communications in Amateur Radio. For more information contact Maty Weinberg at ARRL Hq. □□