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November 1997

Single-Board Performance on 5760 MHz

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QEX

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About the Cover

Paul Wade, N1BWT's roving rig based on his 5760 MHz transverter. For details, read our lead article on page 3.

Features

Columns

November 1997 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

Things are looking up! I have begun to hear from readers, and the flow of articles has picked up a bit. This month we have a couple of letters to the Editor. Keep writing! Remember, if you have a neat technical idea, a short note will be welcome. Several of you have commented on the monthly versus bimonthly versus quarterly publication of *QEX*. So far, the consensus is that bimonthly would be fine, but quarterly is too long between issues (*QE* who?), and the issues would have too much material to be absorbed all at once. All this goes into the hopper as we plan *QEX* for the new year. Keep your comments coming; we are listening.

Some of you have commented on the almost total absence of information on switching regulator techniques in the amateur literature. This is a case where the amateur community has been lagging rather than leading in a new field. There are of course some reasons for this. A switching regulator is more complex to design than a simple transformer-rectifier-filter power supply, and a misdesigned switching regulator is more likely to incinerate itself. It is, however, no more complex than designing a power amplifier, something that amateurs have been doing very well for a long time. The rewards for building a well designed switching converter are many, including greater efficiency, improved dynamic performance and regulation, smaller size and weight and possibly lower cost. In the commercial world, switching regulators are substantially cheaper than linear regulators of comparable power. There is the misconception that switching regulators are intrinsically unreliable. This is not true and has not been the case for a very long time. A properly designed switching regulator can be very reliable. Just look at the millions used in PCs. Failures are relatively rare even though these power supplies are built for the lowest possible cost.

For amateurs to adopt this technology we must provide two things: First, we need basic explanations of how switching regulators work and how to design them. Second, we need good, simple examples aimed at typical ham applications. To help with the first requirement, I will be writing a series of *QEX* articles that explain the basic operation and design of switching regulators. While this will be helpful it is only part of the solution. Designing, building, testing and then writing up good examples takes a lot of effort, and I cannot do all that is needed. I need help!!!! Along this line, I am in the process of contacting hams in a number of semiconductor manufacturers' applications departments. A large number of application notes describe simple, but very useful, examples of switching regulators. Hopefully, we can get some of this information into *QEX* and provide a list of useful application notes and how to get them. In addition, I ask those readers who work in power conversion to look over their work and see if there are some examples that we can publish in *QEX*.

If we are going to drag the amateur community out of the power-supply dark ages, I will need your help. If we can generate some good stuff in *QEX*, it will eventually get into the *Handbook* and *QST* to reach a much larger audience.

This Month in QEX

This month we have filled our 32-page dance card! For microwave enthusiasts, Paul Wade, N1BWT, describes a single-board transverter for 5760 MHz. It's not a "no-tune" design, but pipe-cap filters and newly available MMICs provide good performance in a reasonable project.

For those of you deeply into data transmission modes, Ken Wickwire, KB1JY, goes over a software package for analyzing channel and station performance data from HAL CLOVER modems.

In line with our desire to provide simple projects along with more advanced material, Bill Latta, N4LH, shows us how to build very simple receivers using an inexpensive IC developed for AM broadcast receivers. He shows how to adapt these ICs for 80/75 m operation.

Zack Lau, W1VT, is back this month with the RF column. This time he provides the dope on a VHF local oscillator for KK7B's 23 cm no-tune transverter. As usual, his column is informative and thought provoking.—*Rudy Severns, N6LF; e-mail* **n6lf@arrl.org**

A Single-Board Transverter for 5760 MHz and Phase 3D

Pipe-cap filters and advances in MMICs place 5760 easily with your reach.

By Paul Wade, N1BWT

Introduction

Single-board transverters account for most of the activity on the microwave bands below 5760 MHz, but there has been a dearth of high-performance designs for higher bands. This transverter provides the compact convenience of a single-board unit without compromising performance, making it easy to get on 5760 MHz or the coming Phase 3D satellite without requiring scarce surplus components.

The main reasons for the scarcity and complexity of previous transverter designs for this band are a lack of suitable inexpensive gain devices and the difficulty in making printed filters. For instance, the only single-board transverter article to date, by KK7B,[1] was simply a bilateral mixer and printed filter; a high IF, 1296 MHz, was required because of poor filter selectivity.

Recently, the availability of inexpensive GaAs and heterojunction bipolar silicon (ERA-series) MMIC devices has solved the gain problem, and pipe-cap filters[2] offer a good alternative to printed filters. In a previous 5760 MHz article,[3] I suggested that the next improvement would be to add a transmit and receive amplifier stage to that dual-mixer board. This article goes a step further, by including a multiplier chain for the local oscillator and enough amplifier stages to make a complete transverter station.

Description

The heart of this transverter is the printed-circuit dual mixer,[3] a design that has worked well for several years—a number are in use. The receive mixer is augmented with a GaAs MMIC preamplifier preceding the pipe-cap filter. This preamp provides good noise figure with adequate gain. The transmit side has two stages of gain with two pipe-cap filters. One follows the mixer and there's one between the MMIC devices, to adequately filter out the LO and image signals. The LO chain consists of three active stages separated by filters: an MMIC working as a times-10 frequency multiplier, followed by a pipe-cap filter, an MMIC amplifier, a second pipe-cap filter, and another

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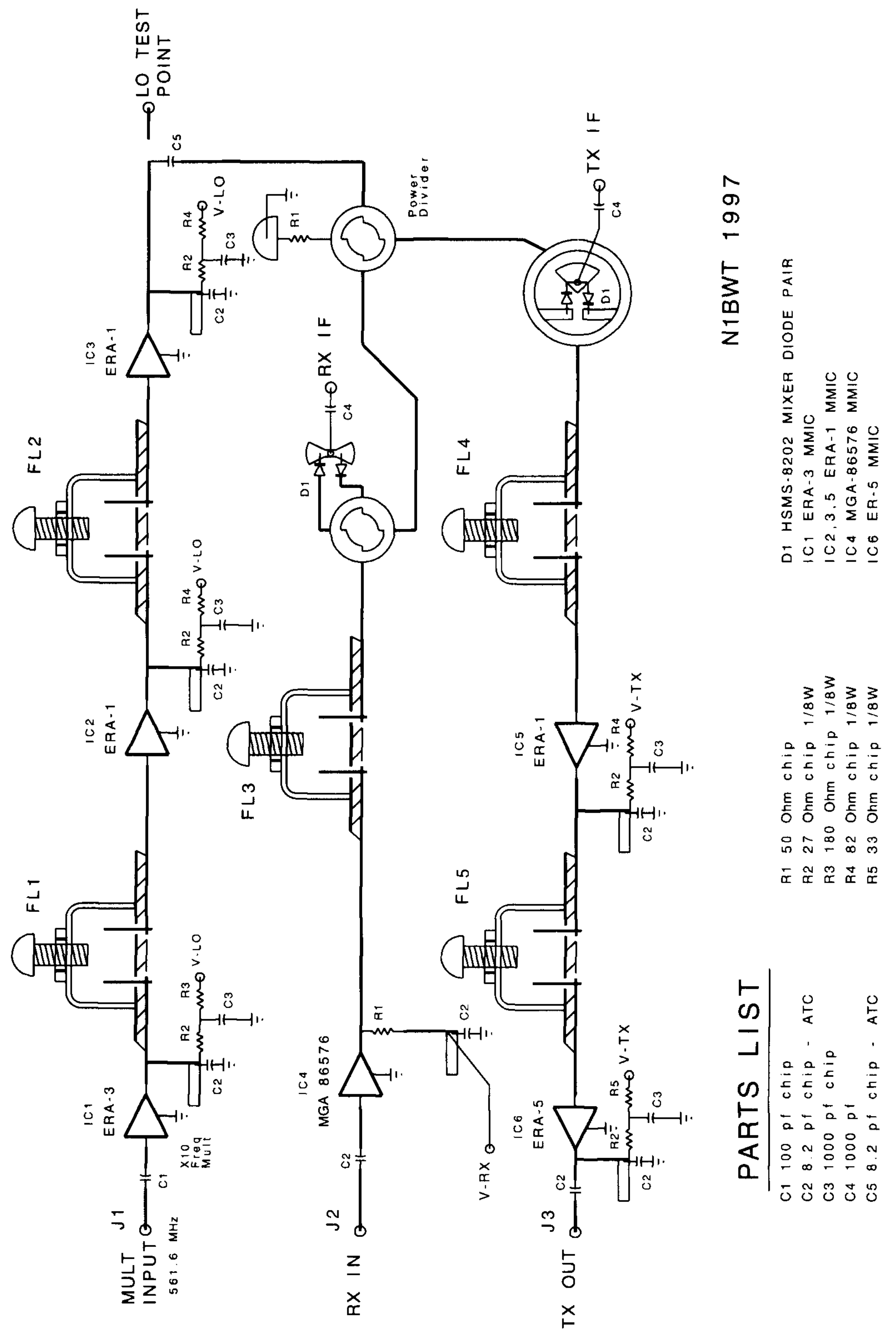
[1]Notes appear on page 14.

MMIC amplifier to raise the power to the level required for the mixers. The MMIC devices keep the schematic diagram in Figure 1 straightforward enough so that a separate block diagram is not necessary.

The local oscillator input, at 561.6 MHz (or 552.4 MHz for Phase 3D), comes from the same KK7B local oscillator board[4] used in lower-frequency single-board transverters. These boards are readily available and work so well that no alternative was seriously considered. When I built the LO board, I calculated the resistor values for operation at 9 V, so that it would operate at the same regulated voltage as the rest of the transverter. I also carved out a small block of Styrofoam to fit around the crystal for better thermal isolation, in the hope



5760 MHz SINGLE BOARD TRANSVERTER

FIGURE 1

of better frequency stability.

Multiplier

The new ERA series of MMIC devices from Minicircuits[5] offer usable gain up to 10 GHz at low cost, so they were obvious choices for the amplifier stages. However, an article by NØUGH,[6] which described using the ERA-3 as a frequency multiplier for 10 GHz, showed additional possibilities. Using a hobby knife, I made a breadboard of an ERA-3 followed by a pipe-cap filter on a scrap of Teflon PC board to test the multiplier performance. A ¾ inch pipe-cap filter can be tuned roughly from 4 to 7 GHz, so I tuned it to several different frequencies and varied the input frequency and power with a signal generator to try various multiplication factors, from ×4 to ×15. The results, plotted in Figure 2, show pretty good multiplier performance, but the curves have too many ups and downs for predictable performance.

While I was trying to understand the data in Figure 2, Steve, N2CEI, suggested that an article[7] on MMIC frequency multipliers by WA8NLC might offer some insight. The article referred me to Hewlett-Packard application note AN-983,[8] which describes how diode-frequency-multiplier performance is affected by the phase shift of the transmission line length between the diode and filter. Since WA8LNC showed that the same phenomenon applies to MMIC frequency multipliers, I calculated electrical line lengths for my breadboard and replotted the data as shown in Figure 3. Now we can see that some multiplication factors are more affected by line length than others, and we can design to optimize for the desired multiplication factor. Figure 3 also shows the ×10 multiplication used in this transverter to be relatively insensitive to line length; the output power change with output line length can be attributed to increasing multiplication factors.

Filters

Printed-circuit filters are not suitable for two reasons: Dimensions become very critical at bandwidths less than 10%, and radiation from them increases at higher frequencies (where the board thickness is a significant fraction of a wavelength). In a recent 3456 MHz design[9] KH6CP (now W1VT) used thinner Teflon PC material to reduce radiation from the filters, however, this requires that dimensional tolerance also be reduced proportionally—to the point where printed patterns would not be reproducible. If high-Q structures like printed filters are eliminated from the board, radiation is reduced, and then the thicker dielectric material is usable.

The pipe-cap filters[2] are made with readily available copper plumbing fixtures and offer good performance but require tuning. My previous experience[4] showed that a single pipe-cap filter did not provide adequate LO rejection at 5760 MHz, so multiple filters were required. I was uncertain whether it would be possible to tune up multiple filters without sophisticated test equipment, so I tested the tuning on the multiplier breadboard. I found that the tuning screw varied the frequency by 300 to 400 MHz per revolution, or about 1 MHz per degree of rotation. Also, the frequency could be set repeatably by measuring the height of the tuning screw. So, it is possible to preset the tuning screws close to the desired frequency, or to easily retune from 5760 MHz to 5668 MHz for Phase 3D. The difference in tuning should only be about one quarter of a turn, but the filters are sharp enough that retuning is required. Finally, since the filters are separated by amplifier stages, it is possible to tune them individually with minimum interaction.

Construction

Layout of the printed circuit board is shown in Figure 4. All components except the pipe-cap filters go on the top surface, as shown in the photograph, Figure 5. The pipe caps are soldered

Figure 2. ERA-3 Frequency Multiplier

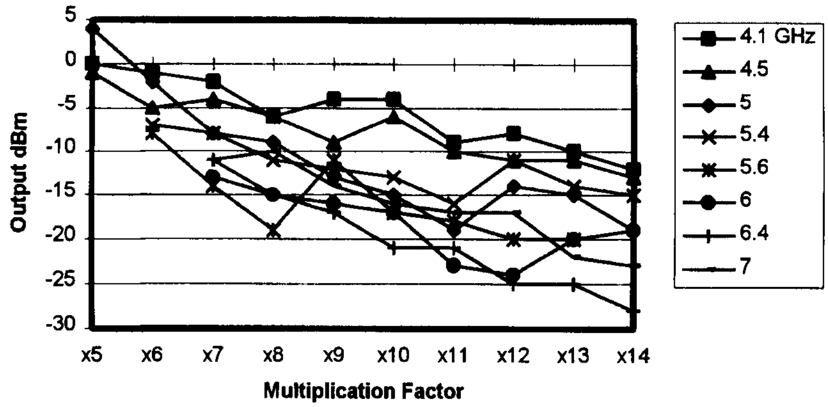
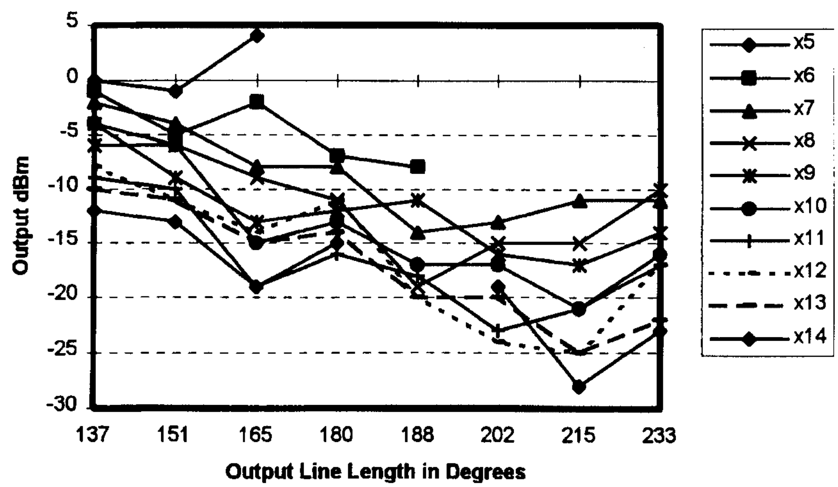


Figure 3. ERA-3 Frequency Multiplier



on the ground plane side. Since a torch is used to solder the pipe caps, it is a good idea to install them first, but not until all holes are drilled and a clearance area is cut in the ground plane around the probe pins for the filters.

A pipe-cap filter sketch is shown in Figure 6. This is the procedure I use to install the pipe-cap filters: in preparation, I drill tight-fitting holes for the probes and make clearance holes in the ground plane around the probe holes. For each pipe cap, I measure from the holes and scribe a square on the ground plane that the pipe cap just fits inside. Next I prepare each pipe cap by drilling and tapping (use lots of oil) the hole for an 8-32 tuning screw, then flattening the open end by sanding on a flat surface. Then I apply resin paste flux lightly to the open end, and to the area around the screw hole for a brass nut to extend the thread length as shown in Figure 6. The nut is not necessary, but it makes tuning smoother. The nut is held in place for soldering with a temporary stainless-steel screw. (Solder won't stick to it.) Next I center the open end of each cap in a scribed square on the PC board — the flux holds it in place. Finally, I fit a circle of thin wire solder around the base of each pipe cap and nut. The caps are soldered one at a time, starting with the center one. I hold each cap down with gentle pressure and heat it for a few seconds with a propane torch until the solder melts and flows into the joints, then let the solder harden before releasing pressure. Don't be shy with the torch—melt the solder quickly and remove the heat. Keep the flame on the top of the pipe cap to avoid damage to the PC board.

For the coupling probes, I use brass escutcheon pins that are 1/32" in diameter. The desired probe length inside the pipe-cap filter is 1/4", so I cut them to a length of 9/32", not counting the head, to compensate for the PC board thickness. The probes are not installed yet, but rather as part of the tune-up procedure. To install them, I put a small amount of flux under the head of each pin, then insert it into the tinned hole and solder.

The 5760 MHz end of the board, with J1, J2 and J3, needs a robust way to attach the connectors. The PC board was planned to fit into an extruded aluminum box (made by Rose Enclosure Systems) that is available from Down East Microwave.[10] The box is supplied with aluminum endplates, but I believe that the PC board ground plane must be soldered to the endplate around the connectors to provide a

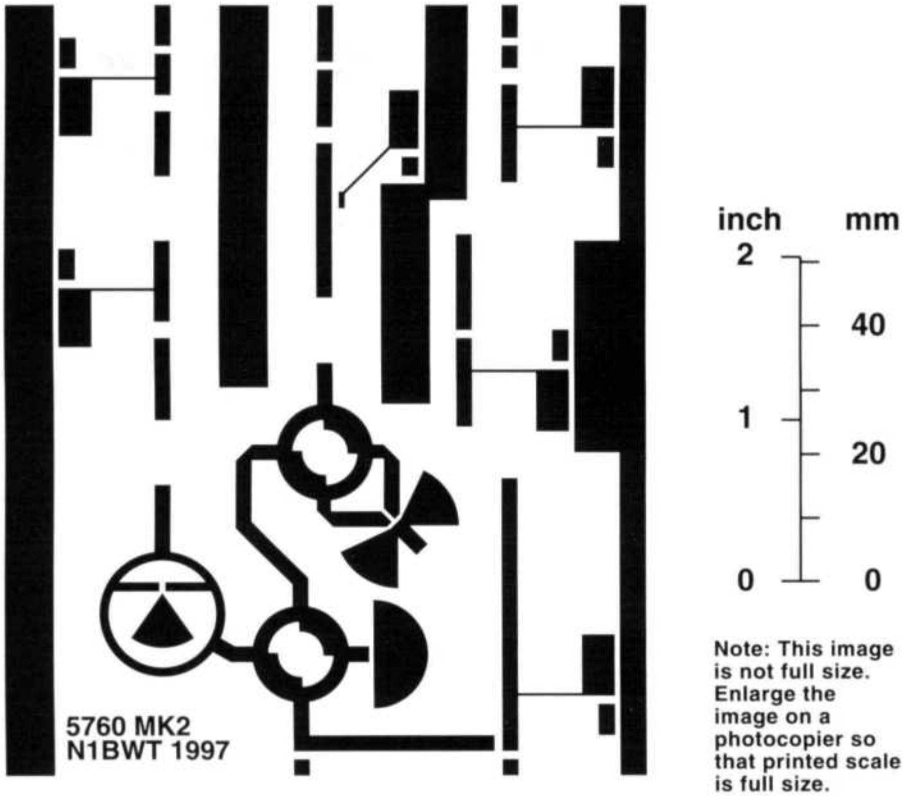


Figure 4. Printed circuit board layout for 5760 MHz single board transverter

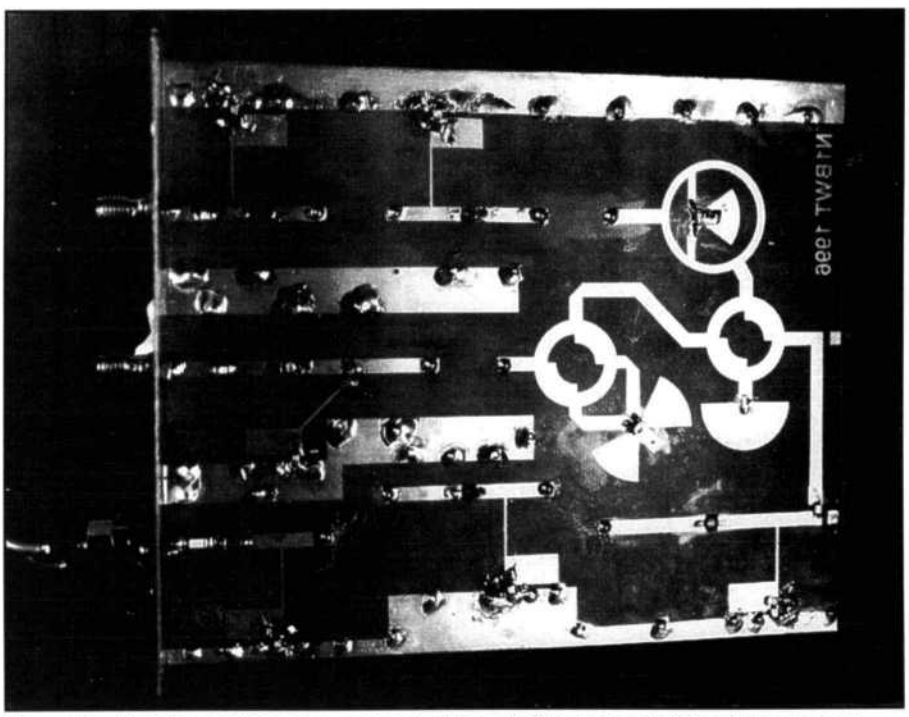


Figure 5—The top surface of a completed board. The parts on this side are very small compared to the board traces.

proper microwave ground. Aluminum does not solder well, so I cut out a brass endplate to match the aluminum ones supplied with the enclosure, drill holes for the connectors and solder the ground plane of the board (top and bottom) to the brass plate. The connector mounting screws only provide mechanical strength. Connector J1 need not be brought outside the box; I only did so for convenient testing.

Once the heavy soldering is complete, the other components may be installed, starting with the MMICs. On my handmade boards, there are no plated-through holes, so I ensure short connections for the ground leads by mounting the MMICs on the bottom (ground plane) side of the board and bending the input and output leads up through the board to reach the printed transmission lines. It takes a bit of trimming at each hole (with a hobby knife) for lead clearance, but the sides of the holes should fit tightly to keep the ground contact as close as possible. Alternatively, Down East Microwave sells boards with plated-through grounding holes, so all devices can be mounted on top of the board with short lead length.

Bias resistors and decoupling capacitors are also soldered on top of the board, but the dc connections are brought out through small clearance holes so that the dc wiring is on the far side of the ground plane and does not cause unwanted feedback. Resistor values shown in the schematic diagram, Figure 1, are for operation from a 9 V source, provided from a three-terminal voltage-regulator IC. The IC should maintain a stable voltage for the transverter and LO board—even with a partially discharged 12 V battery—to keep frequency and power output stable.

Next, solder the dc blocking capacitors (in series with the transmission lines) in place. The capacitor connecting the LO to the mixers, C5, is initially connected to the test point for tune-up, rather than to the mixers. Then the probe pins are inserted into filter FL1 only. FL2 is bypassed by soldering short pieces of enameled wire to each of the transmission-line ends connecting to FL2, with the two wires parallel and closely coupled to form a small capacitor. Now we are ready to begin tune-up; the other filters will be added and tuned one at a time.

Tune-up

The tune-up procedure will tune one filter at a time, adding additional sections of the circuit sequentially. Since the filters are separated by amplifier stages, interaction between them is minimal and repeaking previous adjustments is unnecessary.

The LO section is tuned first. An SMA connector is temporarily attached to the LO test point so that we can monitor output power here. I tune the transverters using only a power meter, and check the LO frequency with a surplus wavemeter. If you do not have a power meter, a diode detector is usable. Figure 7 shows one that can be quickly assembled "dead-bug" style on the flange of an SMA connector using a mixer diode pair and chip components—the values aren't critical. Figure 8 is a plot of the detector sensitivity I measured using a Simpson 260 analog VOM; analog meters make tuning much easier. The plot shows that the expected power levels should provide reasonable output voltage, but shouldn't be taken as a calibration curve, since sensitivity may change with frequency, temperature and different components.

Now install the tuning screw into filter FL1. I use 3/4 inch long flathead brass screws rather than the roundhead screws shown in Figure 6. When tuned to 5616 MHz, the head extends 0.29 inches above the top of the pipe cap. If you begin with the screw at approximately this setting, tuning should go smoothly. (*Note: A 3/4 inch flathead screw has an overall length of 3/4 inch, while a roundhead screw measures 3/4 inch to the bottom of the head. So a round head screw could be used for tuning by measuring the extension to the bottom of the head. However, in the enclosure I used, the extra height of the screw head prevented the cover from fitting properly. A 5/8-inch-long screw would also solve this problem, but was not available at the local hardware store where I found the flathead screws.)*

Next the LO source board at 561.6 MHz is connected to J1, and power is applied to the source and to IC1, IC2 and IC3. (If you can measure power, the LO power input to J1 should be +6 to +10 dBm.) Look for output power at the test point while adjusting FL1; there should be a peak within about a half-turn of the screw

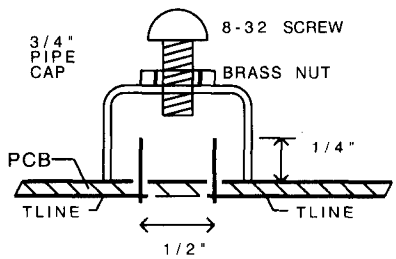
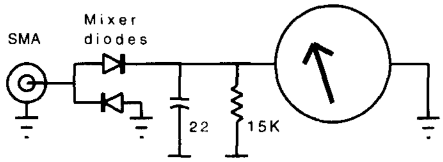
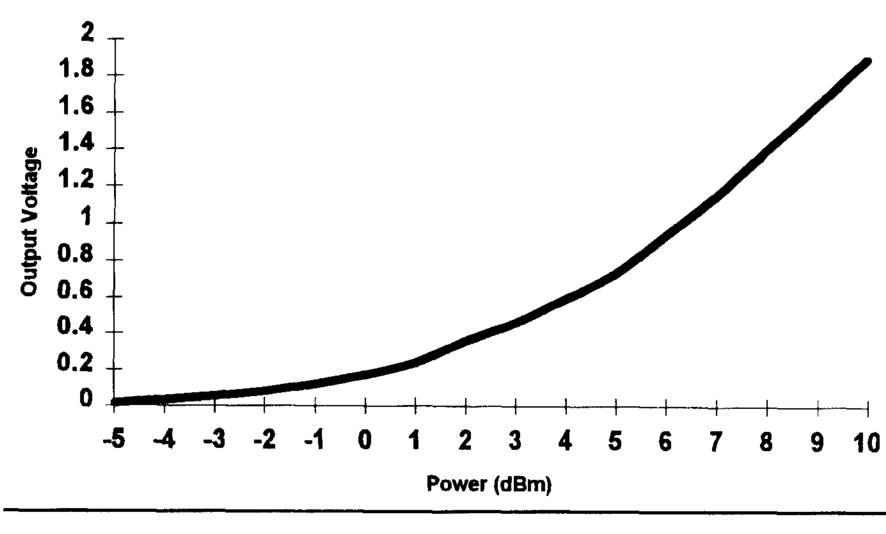


Figure 6. Pipecap Filter for 5760 MHz



Simple Diode Detector
Figure 7

Figure 8. Simple Diode Detector at 5760 MHz



from the initial setting. If much more tuning is required, it's possible that you have found the wrong harmonic. If you have any way of measuring frequency, check it now.

Once you are confident that filter FL1 is tuned to the correct harmonic for 5616 MHz output, peak the output and tighten the locknut on the tuning screw. Then proceed to filter FL2: Remove the bypass wires, insert and solder the probe pins and adjust the tuning screw to the same depth as FL1. Apply power again, peak the tuning screw of FL2, and tighten the locknut. If you are able to measure power, the output at the test point should be +5 to +10 dBm. If the power is slightly low, trimming excess metal from the ends of the transmission lines around the filter probe pins will help.

Now that the local oscillator is complete, the transmitter is next. Move blocking capacitor C5 to the transmission line connecting to the mixers and remove the temporary SMA connector from the LO test point. Insert probe pins into FL4, and bypass FL5 with a wire capacitor just as FL2 was earlier. Initially insert the tuning screw into FL4 the same depth as FL1. Move the output power indicator to J3, then apply power to IC5, IC6 and the whole LO chain. Turn the FL4 tuning screw counterclockwise, decreasing the screw depth to raise the frequency. You should find a peak within about one-half turn of the screw, since filter FL4 should be tuned 144 MHz higher than FL1. Peak the tuning and tighten the locknut.

Proceeding to FL5, remove the bypass wires, insert and solder the probe pins. Set the tuning screw to the same depth as FL4. Apply power again, peak the tuning screw and tighten the locknut. The transmitter is now complete, with a typical output power of +10 dBm.

The final step is to tune the receiver section. Insert and solder the probe pins in FL3, and apply power to the LO chain and to IC4 through the 3-terminal regulator, IC8 (Figure 11; IC8 and surrounding components are attached to the back of the board, dead-bug style). Peak FL3, while listening to another station or to a weak signal source,[11] and tighten the locknut. If you can measure noise figure, adjust the voltage to IC4 for best noise figure by varying the 910 Ω resistor attached to IC8. Otherwise, leave the voltage at approximately 6 V, since the voltage for best noise figure is between 5 and 7 V and not very critical.

Performance

As a barefoot transverter, this unit makes an excellent rover rig, with about 10 mW transmitter output and about 3.5 dB receiver noise figure. Both the LO and the transmitter outputs are pretty clean due to the double filtering. Figure 9 is a plot of the IF chain selectivity (when operated as a straight-through amplifier rather than a frequency multiplier). All LO

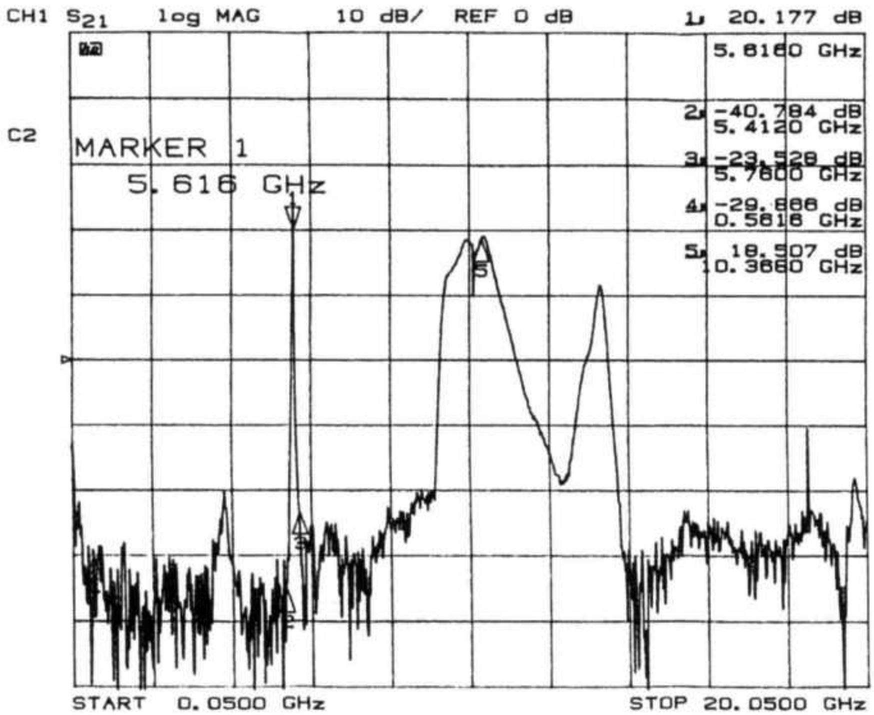


Figure 9—A plot of the IF chain selectivity (see text).

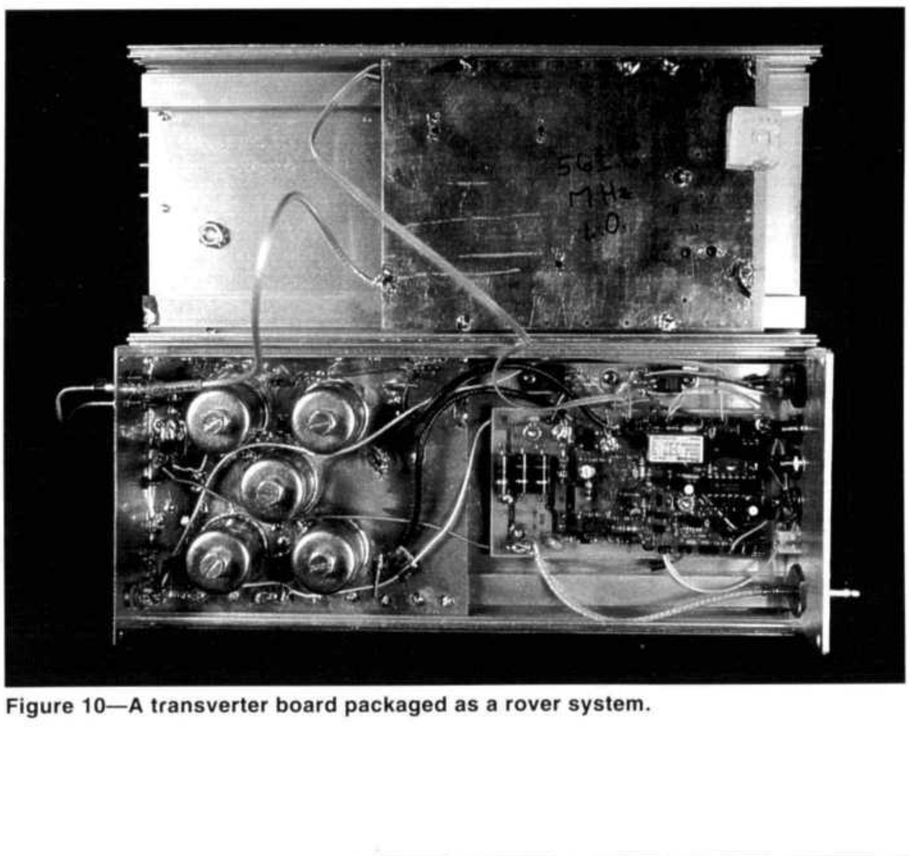


Figure 10—A transverter board packaged as a rover system.

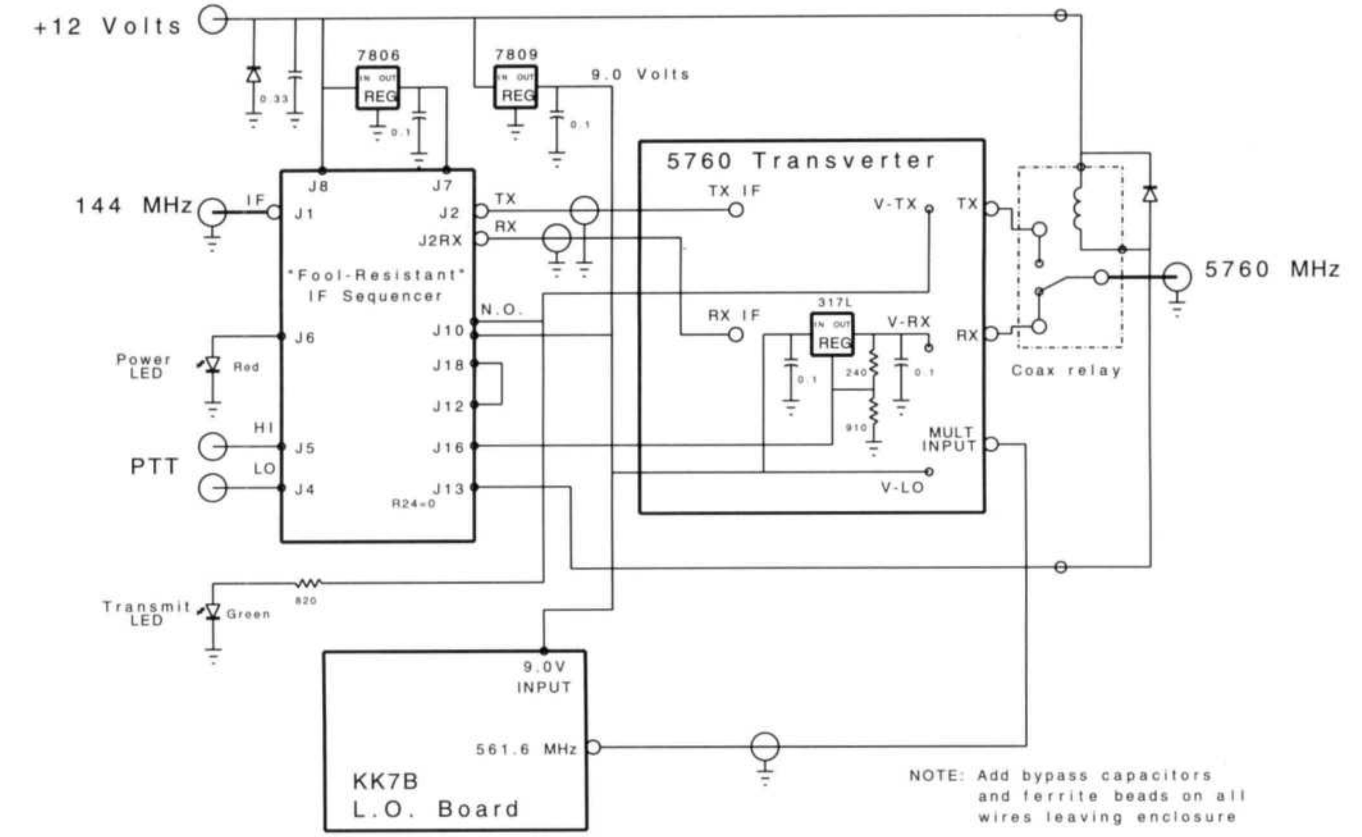


Figure 11. 5760MHz Transverter System Diagram

outputs at the test point are more than 60 dB down, except for the third harmonic of the LO sources at 1684.8 MHz, which is only 50 dB down. The transmitter output has some LO leakage, about 35 dB down from peak carrier output, and the image signal is about 45 dB down. Since the LO leakage is about 45 dB down when the transmit section is not powered, more shielding would be required for significant improvement. All other outputs are more than 60 dB down, except the second harmonic, which is only 30 dB down; as Figure 9 shows, the pipe-cap filters have little rejection above 10 GHz.

On the receive side, a single filter provides adequate image rejection to maintain a good noise figure.[12]

This performance is probably superior to lower-frequency "no-tune" transverters and should be an excellent foundation for a high-performance system, with the addition of power amplifiers and a low-noise preamp.

System

The transverter board does the RF work, but does not make a complete system without help. Figure 10 is a photograph of a transverter board packaged as a compact rover system in a Rose extruded box. One half of the extrusion contains the LO source board, mounted face down, while the other half contains the transverter board, also mounted face down, and an IF sequencer interface board.[13] The two RF boards are mounted face down so that the active circuitry is sandwiched between the PC board ground plane and the grounded wall of the case, thus maximizing isolation between different sections. The block diagram of this basic system is shown in Figure 11. The minimal rover system in Figure 12 uses the RF-sensing function of the IF board to operate with only a simple HT and a horn antenna.

Unless you are willing to move the

Continued on page 11.

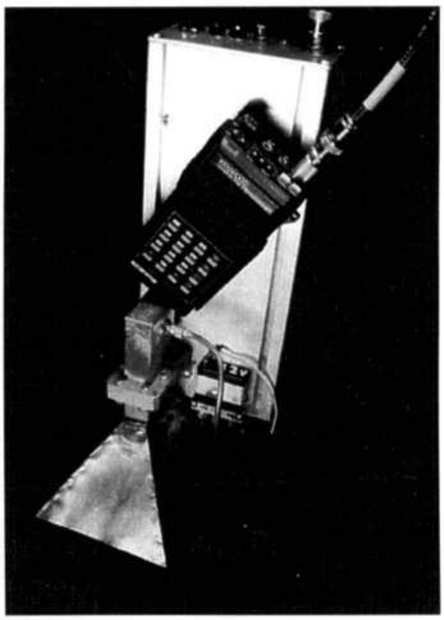
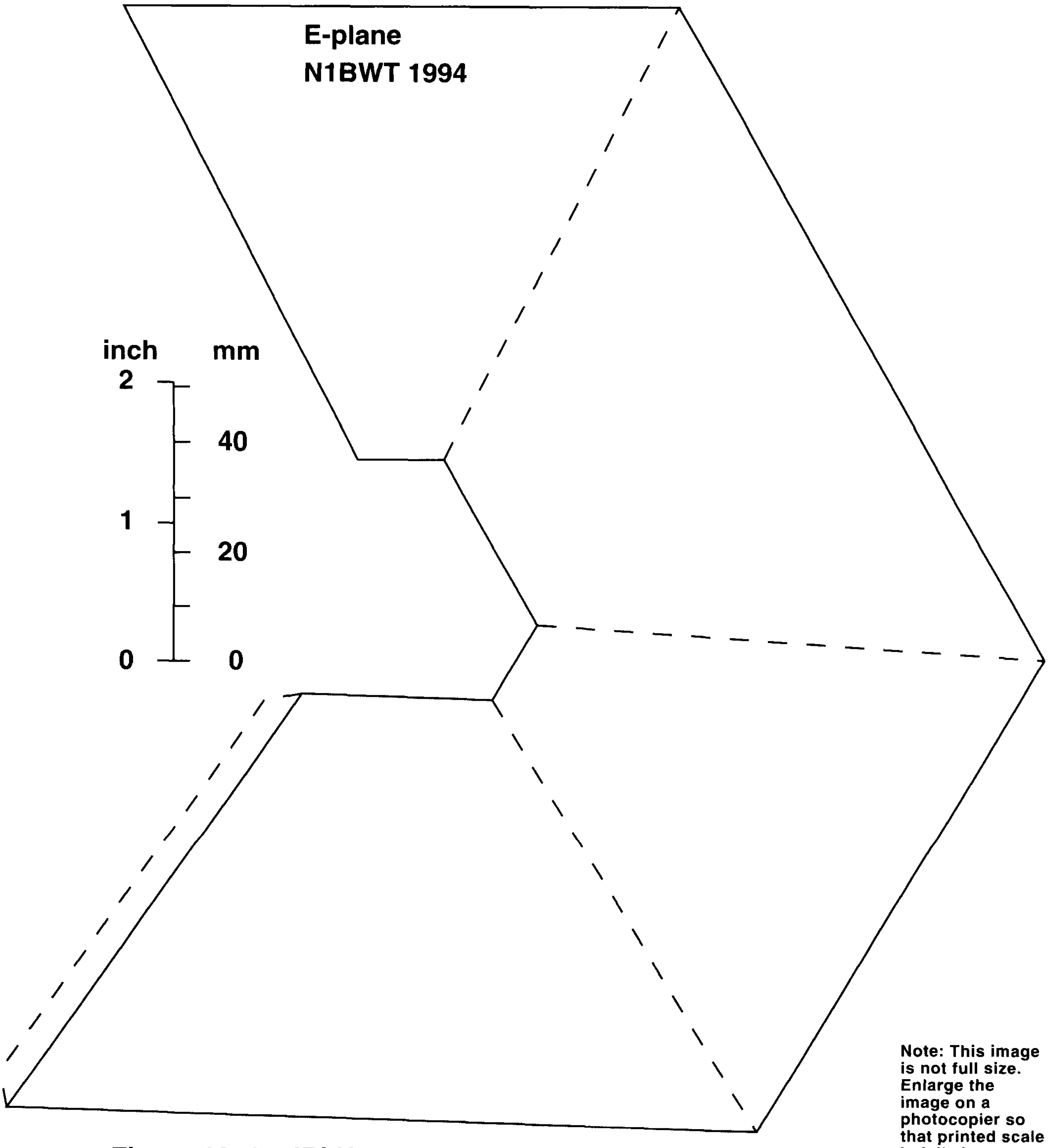


Figure 12—A minimal rover system. The RF-sensing function of the IF board allows operation with only a simple HT and a horn antenna.

Template for 14.88 dBi horn for 5760 MHz



Note: This image is not full size. Enlarge the image on a photocopier so that printed scale is full size. Alternatively, make a new full-size template with HDLANT (see Note 15).

Figure 13. 15 dBi Horn Template for 5760 MHz (WR-137 Waveguide)

antenna cable to switch from transmit to receive or to use separate antennas, an antenna switch is required. Down East Microwave stocks some affordable used SMA coax relays that require 24 V to operate. I disassembled one of these and rewound the coil (see the sidebar "Rewinding Coax Relays for 12 V Operation") to operate at 12 V,

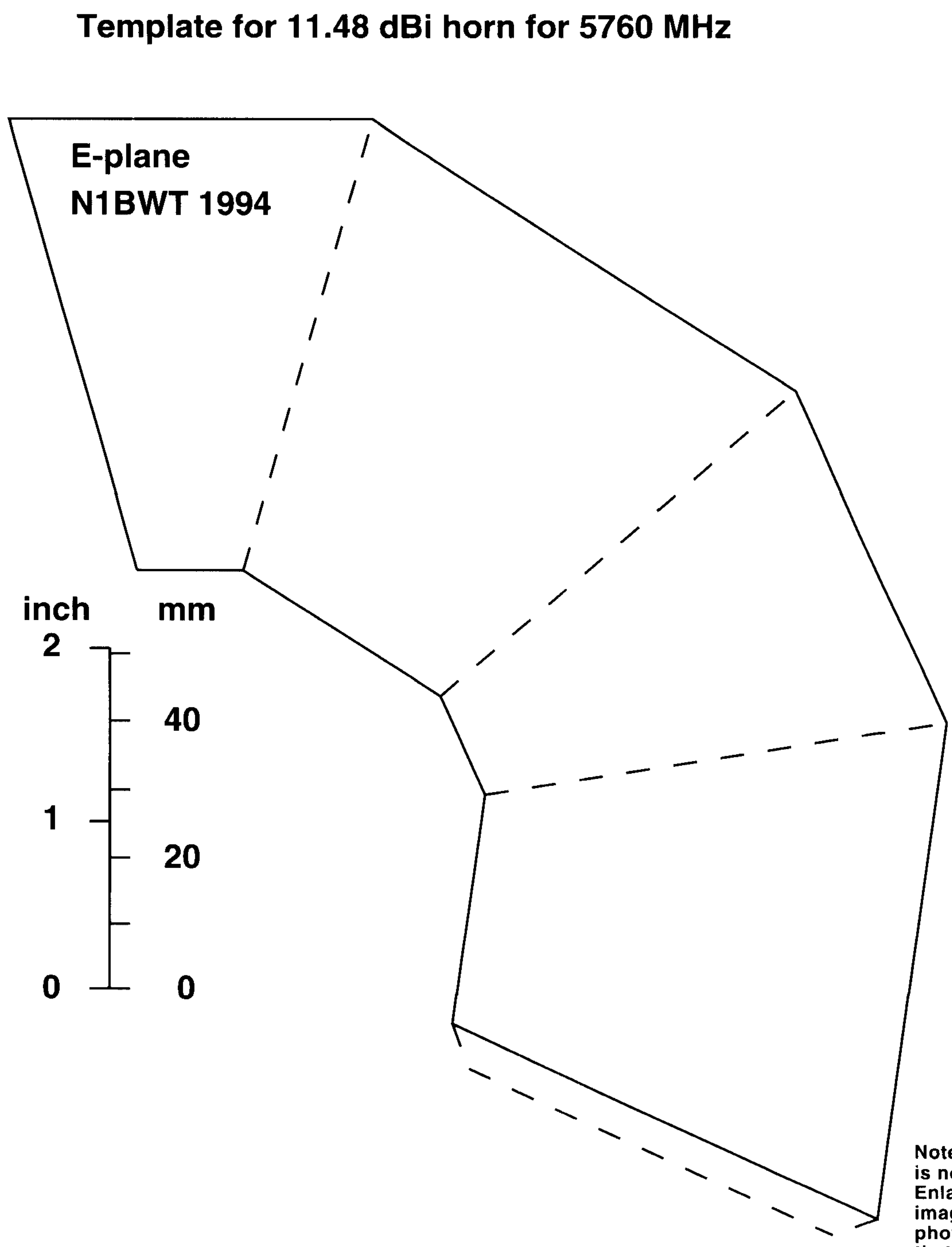


Figure 14. Feedhorn Template for RCA DSS Offset Dish 5760 MHz, WR-137 Waveguide

Phase center is about 7 mm. inside horn

Note: This image is not full size. Enlarge the image on a photocopier so that printed scale is full size. Alternatively, make a new full-size template with HDLANT (see Note 15).

Rewinding Coax Relays for 12 V Operation

Microwave operation from high places requires portable operation for most of us. The most convenient power source is usually the 12 V battery in the vehicle that gets us there, and modern solid-state devices work fine at 12 V, or less. Most surplus coax relays, however, are designed for operation at 28 V or more and don't switch reliably at 12 V. When available, 12 V coax relays are exorbitantly priced; so it would be nice if the higher-voltage relays could be converted. Since relays are ancient technology, digging through some ancient issues of *QST* yielded an article* that detailed the calculations necessary to rewind relays for different voltages. I will summarize them here because the back issue is probably no longer available.

Ham folklore says it is only necessary to remove turns from the coil until it works at 12 V. I'm told this often works for 28 V relays, but let's go through the numbers and see how well.

Calculations

The relative force generated by the coil to close a relay is conveniently measured in ampere-turns, simply the current through the coil times the number of turns. If we are rewinding a relay, rather than count thousands of turns we can simply fill the bobbin with wire and assume that the volume of wire is constant.

Using this assumption and standard US wire gauge (AWG) sizes simplifies the equations. The AWG wire diameters decrease geometrically with increasing AWG number, so that each size is approximately 1.12 times (or $10^{0.05}$) smaller than the preceding size. Using this relationship, we can calculate the number of turns per square inch, N, of bobbin cross section changing by a factor of $10^{0.1}$, or 1.26, per wire size, and the resistance, R, per cubic inch of winding changing by a factor of $10^{0.2}$, or 1.59, per wire size.

Since we are rewinding the same bobbin, area and volume are constant (k), so

$N^2/R = k$

Multiplying by I^2 / I^2, this becomes:

$(NI)^2 / I^2R = k$

we can recognize NI as ampere-turns and I^2R as watts, or power.

This means that—for any wire size that fills the bobbin—the same amount of power applied provides the same number of ampere-turns, so what we need to calculate is the wire size that will draw the same power at the desired voltage:

$V_1^2 / R = V_2^2 / R$

If V_1 is the original, higher, voltage and V_2 is 12 V, then we must increase the wire diameter by:

number of wire sizes = $10 \log(V_1 / V_2)$

Remember that a larger diameter wire has a smaller AWG number. The original article had a graph, but this is easily solved on a calculator, which they didn't have in 1956. I've summarized the most common voltages in the following table:

Common Relay Voltages

Original	*Desired*	*# of wire sizes*
28 V	12 V	4
48 V	12 V	6
115 V	12 V	10

Procedure

The rewinding procedure is straightforward: Peel the old wire off the bobbin, measure the wire size and rewind it with larger magnet wire as calculated above. Radio Shack carries several sizes of magnet wire, which may include the one you need. Mechanical details should be pretty much like the original relay; take notes during disassembly. The most difficult part is often prying the bobbin off the metal pole. Relays in sealed cans are a larger problem, and I welcome suggestions. (Letters to the Editor?—*Ed.*)

The fastest way to wind the new coil is to wrap masking tape around a dowel or pencil until the bobbin fits snugly over it, chuck the pencil in a variable speed drill or lathe, and run it *slowly* to wind the wire on. At *low* speeds, it's safe to guide the wire with your fingers.

Example

I found several excellent coax relays with N-connectors at a hamfest, quite cheap because they required 48 V. They were wound with #38 AWG wire. From the equation above, converting from 48 to 12 V requires wire roughly six gauges larger, so I rewound one with #32 AWG wire. The original winding required 48 V at 54 mA, pulled in at 35 V and released at 15 V. After rewinding, it draws 265 mA at 12 V, pulls in at 8.5 V, and releases around 2 V. The power is slightly higher now, because six wire sizes is an approximation, but I can be sure it will still operate on a low battery. My 903 MHz station now runs entirely on 12 V.

The SMA relay in Figure 11 (from Down East Microwave) was an easy one to rewind. The cover is held on with two screws, and the bobbin is easily removable. The 28 V coil is wound with pretty fine wire—I measured the diameter, guessed the enamel thickness, and estimated the wire size as #42 AWG. From the table above, the new wire should be four wire sizes larger, or #38 AWG. I rewound the coil with #38, reassembled it and tested it. The relay now switches solidly at 11 V. A better choice might be #36 AWG wire, which would give a little more voltage margin for low-battery operation, but I don't have any in the junk box. Incidentally, this relay has an unusual construction that will not operate with the coil voltage reversed, so try it both ways to find the right polarity.

Alternative

As mentioned previously, we could have just removed turns to increase the current until the relay draws the same power at the lower voltage. If we take off half the turns, the resistance drops in half. The original resistance of the 48 V relay is 48 V / 54 mA = 888 Ω. At 12 V, we need 216 mA for the same power, or a new resistance of 55 Ω, so we need one-sixteenth as many turns. We increased the current four times, so we end up with one-quarter as many ampere-turns as the original, or only one quarter as much force pulling in the relay. If we weaken the spring enough, it may work, but will it be reliable?

A few more trials convinced me that no matter how many turns of the original wire are removed from a 48 V relay, the force pulling it in at 12 V will only be 25% of that at 48 V. A 28 V relay isn't as bad—the force is only reduced by 12/28, to a bit less than half the original force. There is probably a combination of turns and spring bending that will work pretty well, but if you've done enough disassembly to remove some turns, why not take the rest off and rewind it for 12 V?

International

I haven't tried any relays from other countries, but I wouldn't be surprised if they use other wire size systems. In the UK, they may still use SWG sizes, which differ from AWG, but the relative sizes are close enough so that increasing the diameter by the number of sizes calculated above should work. So measure the wire, convert to the nearest AWG or SWG size, and go from there. I don't know what metric standard wire is available. (The Component Data chapter of the *ARRL Handbook* contains a "Copper Wire Specifications" table that lists AWG specifications with the nearest equivalent SWG.—*Ed.*)

Conclusion

Rewinding a surplus coax relay for 12 V operation requires only one simple calculation and perhaps an hour of work; why not try it rather than pay exorbitant prices or use inefficient dc-dc voltage converters?

* L. B. Stein, Jr, W1BIY, "Some Hints on Relay Operation," *QST*, Jun 1956, pp 21-25.

so the whole system in Figure 12 can be powered by an automobile battery.

The IF sequencer board also provides all the circuitry and sequencing needed to safely control external preamplifiers and power amplifiers for a high-performance system.

Antennas

Only a simple horn antenna[14] is needed for a basic rover system; I used my **HDLANT**[15] computer program (download from **http://www.arrl.org/qexfiles/**) to create the template in Figure 13. Use the template to build a horn that has 15 dBi gain—like the one in Figure 12—from a bit of flashing copper. Just tape a full-size copy of the template to a sheet of copper or brass, cut it out, fold on the dotted lines, and solder the metal horn together on the end of a piece of waveguide.

For a rover system with better performance, feed a small DSS dish[16] with an offset feed horn[17] (see template in Figure 14). This and a multimode transceiver—for CW and SSB capability—make up the excellent rover system shown in Figure 15. The system is mounted on top of a 10 GHz transverter. With a quick-change feed mounting arrangement I have a two-band rover station sharing the same dish. Of course, larger dishes can provide even better performance.[18]

Both of the antenna templates are for horns that mate with WR-137 waveguide (1.37" × 0.62" internal dimensions), but the larger WR-162 and WR-187 sizes also work fine at 5760 MHz. Surplus waveguide is available, so you should be able to find one of the three sizes or make a reasonable imitation from copper or brass sheet. Waveguide-to-coax transitions are more difficult to find, but easy to build: Figure 16 shows dimensions for a WR-137 transition. The coax probe is a section of 5/32" diameter brass rod with a hole drilled in it to fit over an SMA connector pin, which is soldered in position. Trim the SMA's Teflon insulation flush with the inside of the waveguide. This unit has a low SWR from 5.2 to 7.5 GHz and outperforms several commercial units I have tested. If yours needs adjustment, vary the length of exposed SMA pin (nominal 0.110 inches) by moving the 5/32" diameter section and resoldering it at the new location; repeat until good SWR is achieved.

Potential

Even the basic rover system in Figure 12 is capable of communications over *any* line-of-sight terrestrial path. I can make this claim because we have frequently demonstrated that a simple 10 GHz WBFM Gunnplexer system is capable of distances more than 100 km. This simple 5760 MHz system using NBFM has comparable power output and antenna aperture, but has a 13 dB advantage in receive bandwidth (approximately 10 kHz vs 200 kHz) and a noise figure at least 5 dB better. The improvement of 18 dB translates—using the inverse-square law—to eight times better range capability (more than 800 km) enough for all line-of-sight paths.

The improved rover system shown in Figure 15 adds another 10 dB of antenna gain and 5 dB better receiver bandwidth (even more for CW) to

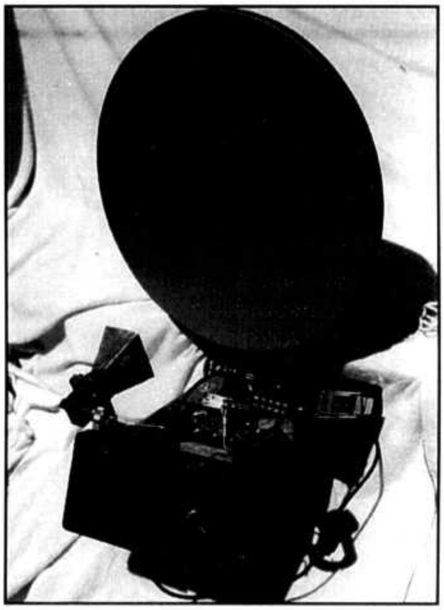
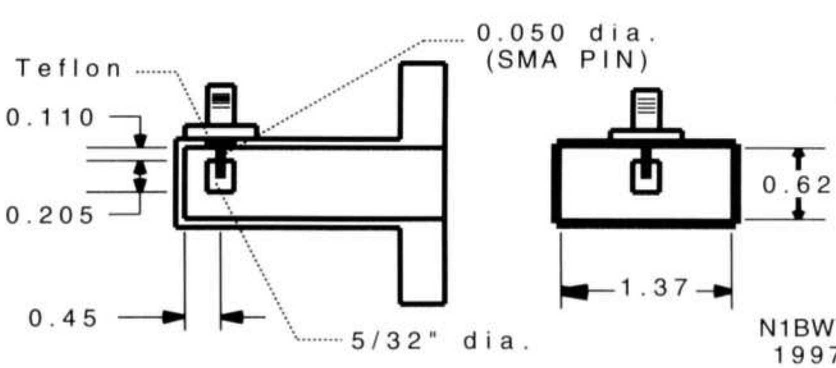


Figure 15—A two-band roving station that uses the 5760 MHz transverter.



WR-137 Waveguide Transition
Figure 16

provide the potential for over-the-horizon or partially obstructed paths.

From there, we can move up to larger antennas, a receive preamplifier and transmit power amplifiers—all the way up to EME capability.

Phase 3D

The Phase 3D satellite requires only a transmit capability at 5668 MHz. If there is another interested station not too far away, however, receive capability makes it possible to check out both systems before tackling the complexities of satellite tracking.

Published estimates[19] of Phase 3D path loss at this frequency are 181.5 dB at perigee and 201.4 dB at apogee. With the estimated 22 dBic satellite antenna gain, the system in Figure 14 should be able to provide an uplink signal that is weak at perigee, and lost in the noise at apogee. One solution would be a larger dish, but that would have a beamwidth narrower than the manageable 8° beamwidth of the DSS dish. A better solution would be a modest power amplifier; perhaps I can put one together before the satellite is launched.

Conclusion

This transverter provides the compact convenience of a single-board unit without compromising performance. While it is not a "no-tune" design, tune-up is straightforward and systematic. Now it is possible to get on 5760 MHz without any hard-to-find surplus components.

Notes

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[3]P. C. Wade, N1BWT, "A Dual Mixer for 5760 MHz with Filter and Amplifier," *QEX*, Aug 1995, pp 9-13.

[4]R. Campbell, KK7B, "A Clean, Low-Cost Microwave Oscillator, *QST*, Jul 1989, p 15. Also appears in *ARRL UHF/Microwave Project Book*, pp 5-1 to 5-9.

[5]Mini-Circuits Labs, PO Box 350166, Brooklyn, NY 11235-0003; tel 718-934-4500; **http://www.minicircuits.com/**.

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[14]P. C. Wade, N1BWT, "Practical Microwave Antennas, Part 1," *QEX*, Sep 1994, pp 3-11.

[15]P. C. Wade, N1BWT, "Practical Microwave Antennas, Part 2," *QEX*, Oct 1994, pp 13-22.

[16]MCM Electronics, 650 Congress Park Dr, Centerville, OH 45459; tel 800-543-4330.

[17]P. C. Wade, N1BWT, "More on Parabolic Dish Antennas," *QEX*, Dec 1995, pp14-22.

[18]P. C. Wade, N1BWT, "High-Performance Antennas for 5760 MHz," *QEX*, Jan 1995, pp 18-21.

[19]W. A. Tynan, W3XO, "Phase 3D, A New Era for Amateur Satellites," *Proceedings of the 30th Conference of the Central States VHF Society* (Newington: ARRL, 1996) pp 47-61.

Feedback

All I can say about my August *QEX* article is "Oops!" (see "Synthesizing Vacuum Tubes," *QEX*, Aug 1997, pp 17-21). The callouts in Fig 3 for a 2N5462 (Q1) and in Fig 5 for a 2N3822 are in error, Q1 should be an MPF3822 in both places. In Figure 5, the drain and source leads are interchanged in the layout so the figure is in error. Nonetheless, the source and drain are functionally interchangeable in JFETs, so the layout in Fig 5 is misleading but operable. My apologies to eagle-eyed readers.—*Parker R. Cope, W2GOM/7, 8040 E Tranquil Blvd, Prescott Valley, AZ 86314*

A Software Package for Analyzing CLOVER Channel Statistics

CLOVER's channel-quality data can help you assess propagation, troubleshoot channels and evaluate station changes.

By Ken Wickwire, KB1JY

1. Introduction

The HAL CLOVER modem and data-transmission protocols have now been in use for about five years, and the system has established itself as one of the most reliable and efficient means of HF data transmission available to hams. Several versions of CLOVER are available (the PCI4000, DSP4100, P38 and others). The versions are distinguished mainly by their maximum speed of data transmission, which is a function of the processing power of the associated hardware (computer or modem or both).

One of the capabilities of CLOVER that puts it in the forefront of HF digital developments, in and beyond the amateur world, is the system's real-time measurement and exchange of channel-quality data. These data are used by pairs of connected stations to adapt themselves automatically to changing channel conditions (including the effects of QRM). With the HAL software control interface, the data can also be recorded to an ASCII text file at any connected station. Every version of CLOVER can be used to record channel quality data, although there are minor differences (depending on the version's horsepower) in how many data fields each version fills.

We believe that these channel quality data have not been sufficiently appreciated, perhaps because of the lack (or ignorance) of software for analyzing and displaying them. This paper describes a relatively simple software package that performs statistical analysis of recorded CLOVER channel quality data and displays the data in graphical form. The paper's sections describe CLOVER channel quality data, the statistical analysis software, example plots, channel quality data in the FEC mode and some uses of CLOVER statistics. The last section gives our conclusions.

We hope the paper will encourage others to explore CLOVER statistics further and to write more useful packages. The statistical analysis software can be downloaded from the ARRL's ftp site or BBS.[1] The freeware graphing software we use to display the output of the statistics program can be downloaded from the Internet.[2]

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[1]Notes appear on page 21.

2. Description of the CLOVER Channel Quality Data for the ARQ Mode

As a pair of CLOVER modems pass information in the CLOVER Automatic Repeat Request (ARQ) mode, they also exchange information on channel quality, along with occasional commands to change modulation mode and other parameters. Some of these channel quality data appear on the screen of the HAL user interfaces to the various CLOVER modems (with the labels "MY" and "HIS," where MY is the recording station). These and several other data items may be recorded to disk by issuing the *Record Channel Statistics* command at the HAL user interface. With the latest CLOVER firmware and the HAL interface, channel statistics can be recorded during keyboard-to-keyboard chats, uncompressed ASCII text file transfers and compressed file transfers. (Compression using the PKLIB suite can be performed via the interface.) The channel data a station records and labels with its call sign ("MY" data) have been sent to it by the other station, which is the one that measured the data.

The recorded data, which are written in comma-delimited format to a channel statistics file after each ARQ link turn around, consist of:

- current time (HHMMSS),
- call sign of the described station (corresponding to the screen's "MY" or "HIS"),
- modulation mode (BPSM, QPSM, 8PSM, 8P2A etc),
- "bias" (1 = FAST, 2 = NORMAL or 3 = ROBUST; see below),
- uncompressed throughput for the most recent frame (in characters per second [cps]),
- signal-to-noise ratio (SNR, in dB, usually between 10 and 50),
- frequency tuning error (in positive or negative Hertz),
- phase dispersion (This is an empirically derived number between 0 and 256 that assesses multipath delay; low numbers are good. We call these numbers "degrees."),
- error-correcting capacity (ecc) used to combat bit errors (in percent; low percentages are good), and
- automatically adjusted percentage of maximum output power in use (a capability not provided by the P38; low levels are good).

The bias is a parameter that describes the error-correcting overhead (and thus the number of data bytes per ARQ block) chosen by the transmitting station for a communications session. Robust bias is usually chosen when the channel appears (sounds) bad and fast bias when it appears to be good.

The error-correcting capacity is the percentage of the maximum amount of forward error correction (FEC) coding overhead actually used to combat detected bit errors. An ecc of "XX" in the file means that the capacity of the Reed-Solomon decoder used by CLOVER was exceeded by the number of detected bit errors in the last received frame and that a request to send the frame again was issued by CLOVER's ARQ protocol.

For further details of how some of these data are calculated, see the HAL documentation supplied with the various modems and the descriptions published by Ray Petit and Bill Henry in the *RTTY Journal*, *QST*, *QEX* and elsewhere between 1990 and 1994.

Sample A is a snippet of channel quality data recorded during a recent file transfer from KB1JY to W1IMM using the P38 CLOVER modem. The HAL interface firmware labels channel quality data files with the format MMDDHHMM.TST; hence, for example, 05032038.TST. The first line of the snippet says that KB1JY was using 8PSM modulation and normal bias (2). His per-frame throughput was 19 cps, his average SNR 35 dB, his frequency error 4 Hz *below* the nominal carrier, his phase dispersion 21 and his used error-correcting capacity 000% (no error-correcting overhead was used for this frame). Since the P38 does not adjust output power, the last data item is zero.

3. The Statistical Analysis Program

Our software package, called clvrstat.c and written in ANSI C, first prompts the user for the channel-quality data-file name. It then reads the first and second lines of data to record their call signs, which it will later label as "MY" and "HIS." (If the user doesn't want the first call sign in a file to be called "MY" he can edit the file so that the first two lines carry the call signs he wants to be called MY and HIS in the output.) After reading the

Sample A

```
203935,KB1JY    ,8PSM  ,2,19,035,-4  ,021,000,000
203935,W1IMM    ,BPSM  ,2,02,029, 003,036,050,000
203954,KB1JY    ,8PSM  ,2,29,033,-4  ,023,000,000
203954,W1IMM    ,BPSM  ,2,02,030, 003,035,050,000
```

Sample B

```
Statistical Summary of CLOVER "status" data.
Input file is [ 05032038.TST] .
MYcall ("MY") = KB1JY; HIScall ("HS") = W1IMM.
Summary date: 13.04.97 20:27:09
Number of reports on KB1JY = 76
Number of reports on W1IMM = 76
E(MY_bias) = 2.0, sd(MY_bias) = 0.00
E(HS_bias) = 2.0, sd(HS_bias) = 0.00
E(MY_tput) = 9.80 cps, sd(MY_tput) = 11.60 cps
E(HS_tput) = 2.00 cps, sd(HS_tput) = 0.00 cps
E(MY_SNR) = 31.36 dB, sd(MY_SNR) = 5.22 dB
E(HS_SNR) = 27.93 dB, sd(HS_SNR) = 3.56 dB
E(MY_dfrq) = -3.53 Hz, sd(MY_dfrq) = 0.72 Hz
E(HS_dfrq) = 2.88 Hz, sd(HS_dfrq) = 0.82 Hz
E(MY_phs) = 23.37 deg., sd(MY_phs) = 7.38 deg.
E(HS_phs) = 34.57 deg., sd(HS_phs) = 5.84 deg.
E(MY_ecc) = 2.48%, sd(MY_ecc) = 6.36%
E(HS_ecc) = 29.86%, sd(HS_ecc) = 27.40%
Error correcting capability exceeded 5 times on MY link
Error correcting capability exceeded 4 times on HS link
maximum "MY" throughput = 29 cps, maximum "MY" SNR = 39 dB
maximum "HS" throughput = 2 cps, maximum "HS" SNR = 35 dB
Modulation counts:
MY: 52 BPSM 2 QPSM 22 8PSM
HS: 76 BPSM 0 QPSM 0 8PSM
```

call signs, the program runs to completion, producing nine output files (three files of statistics and data and six plotting scripts).

The Statistical Summary

The first output file is called MMDDHHMM.sum, where MMDDHHMM is the same as the prefix of the input file. This file contains a statistical summary of the channel quality data. Sample B is a summary file for data recorded during a file transfer from KB1JY to W1IMM over a near-vertical-incidence skywave (NVIS) link about 40 miles long:

Let's walk through the summary. The *MYcall* and *HIScall* were read from the first two lines of the input file. (Since the data were recorded at KB1JY, these would normally be KB1JY and W1IMM. If W1IMM also recorded data, they'd be W1IMM and KB1JY in his summary.) *Number of reports* is the number of lines of data for the call sign listed after it, which should be the number of frames sent from that call sign.

The next twelve lines give the mean E() and standard deviation of the bias, throughput, SNR, frequency tuning error (dfrq), phase dispersion (phs) and used ecc for the MY and HIS stations. (The standard deviation of the bias will be zero if the bias hasn't been manually changed during the recording session.) The program calculates ecc statistics only for actual ecc percentages; the "XX" symbols written to the *.TST data file to indicate exceeded error-correcting capacity are ignored in ecc calculations. (This means that ecc statistics for data collected when the error-correcting capacity was frequently exceeded may not mean much because of the small sample size.) Note that the biases here (2 = Normal bias) are constant since the bias wasn't changed during the session.

KB1JY had per-frame (uncompressed) throughput of about 10 cps and W1IMM, receiving the file, had the two-cps throughput characteristic of control blocks. SNRs in both directions were around 30 dB (fairly good) and both signals were about 3 Hz off frequency (typical). Phase dispersions were, on average, about 30 (tolerable for this NVIS link) and the file sender and file receiver had to use about 2 and 30% of their error-correcting capacity. (The receiver had to correct more errors than the sender, which is not surprising in view of the different amounts of information being sent in each frame, and the fact that the receiver was sending only Clover Control Blocks, which are more immune to channel noise and interference than data blocks.)

Then come the numbers of times out of the "Numbers of reports" that the error-correcting capacity at the receiver was exceeded at each station. When this happens, the ARQ protocol takes over for the forward error correction and asks for the offending frame to be resent. Since this was a fairly good link, the ecc was exceeded only rarely.

Next come the maximum uncompressed throughput in cps and maximum SNR in dB measured during the recorded session in each direction. These may be viewed as gross measures of how good the channel was during the transfer. Since the HIS station (W1IMM) was only ACKing data frames (with Clover Control Blocks, which always use BPSM modulation) and not sending data frames, its maximum throughput was the same as its average throughput (2 cps). Listed finally are the numbers of times each recorded modulation type was used in each direction. A large number of 8PSM modulations recorded for a file-sending station using a P38 indicate a good channel.

The Parsed Data Files

The next two files produced by the software comprise space-delimited lines of data for the two individual stations. These are the same data as those in the input file, but here they're sorted by call sign and given a slightly different time format. The two files are labeled as MMDDHHMM.MY and MMDDHHMM.HIS, where "MY" again corresponds to the station that collected the data in the MMDDHHMM.TST file. Sample C is a snippet of parsed data from a *.TST file collected at KB1PZ (the # lines are comment and label lines inserted by the program):

These parsed data files may be used for plotting the data on individual stations or for further analysis. Their column labels show that the statistics program keeps the individual station data in the same order as in the original *.TST file.

The Plot Setup Files

The last six files produced by the software are ASCII plot-formatting scripts that can be opened by the popular and powerful freeware program called *gnuplot*, which offers plotting capabilities that rival many of those in commercial plotting packages. Versions of gnuplot (properly written with a lower case "g") for the Macintosh, PC, Unix and several other operating systems are available. (The most recent version is 3.6, which is in final beta testing.)

The scripts tell gnuplot the names of the data files (in this case, MMDDHHMM.MY and MMDDHHMM .HIS) and how to format and label the plots. Our version of the statistics program, which runs on a Macintosh, produces gnuplot script files called mysnrtput 3.6, hissnrtput3.6, myphsecc 3.6, hisphsecc3.6, mydfreq3.6 and hisdfreq 3.6. (Users of other systems who use gnuplot would modify the source code if necessary to produce appropriate file

Sample C

```
#Status data for KB1PZ.
#Input file is [ 05170814.TST] .
#Summary date: 20.03.97 16:04:11
#(ecc = "XX" written as "110")
#H/M/S    mod  b tp SNR df  phs ecc pwr
07/49/05 BPSM 1 02 035 000 026 000 000
07/49/07 QPSM 1 08 036 000 026 000 000
07/49/27 8PSM 1 15 035 000 029 005 000
07/49/46 8PSM 1 23 035 000 028 016 000
07/50/06 8PSM 1 23 036 000 030 036 000
07/50/25 8PSM 1 23 037 000 028 015 000
07/50/45 8PSM 1 23 032 000 033 050 000
07/51/04 8PSM 1 23 032 000 033 017 000
07/51/24 8PSM 1 23 032 000 033 017 000
07/51/43 8PSM 1 23 028 000 031 002 000
07/52/03 8PSM 1 23 030 000 029 004 000
07/52/23 8PSM 1 23 030 000 029 004 000
07/52/25 8PSM 1 08 030 -1  036 110 000
07/52/45 8PSM 1 15 032 -1  035 110 000
07/53/04 8PSM 1 23 035 -1  031 007 000
```

names. Those who don't wish to use gnuplot can ignore the scripts or edit the clvrstat.c source code to stop creating them.)

The mysnrtput3.6 and hissnrtput3.6 script files cause gnuplot to produce plots of *SNR* and *throughput* versus time for MY and HIS stations. The myphsecc3.6 and hisphsecc3.6 scripts do the same for MY and HIS *phase dispersion* and *error-correcting capacity*. Finally, mydfreq3.6 and hisdfreq3.6 lead to time plots of MY and HIS *frequency tuning errors*.

4. Examples of gnuplot Graphs of CLOVER ARQ Channel-Quality Data

Figures 1, 2 and 3 illustrate the three kinds of channel-data plots we produce from CLOVER channel data with gnuplot. The data the plots are based on were recorded at KB1JY during a file transfer on 80 meters over a link that sometimes experiences a lot of multipath and hence a lot of phase dispersion. The plots cover about 14 minutes of data collection. The parts of each plot on the right and left correspond to periods just before and after the transfer when both stations were "idling"; that is, sending each other only CLOVER Control Blocks with BPSM modulation. (The "00" entries in the seconds field of the time-axis labels are the results of gnuplot's automatic axis-scaling and tic-mark placement, which can be overridden. The date and time in the lower left-hand corner give the time the plot was made.) Figure 1 shows SNR and throughput versus time.

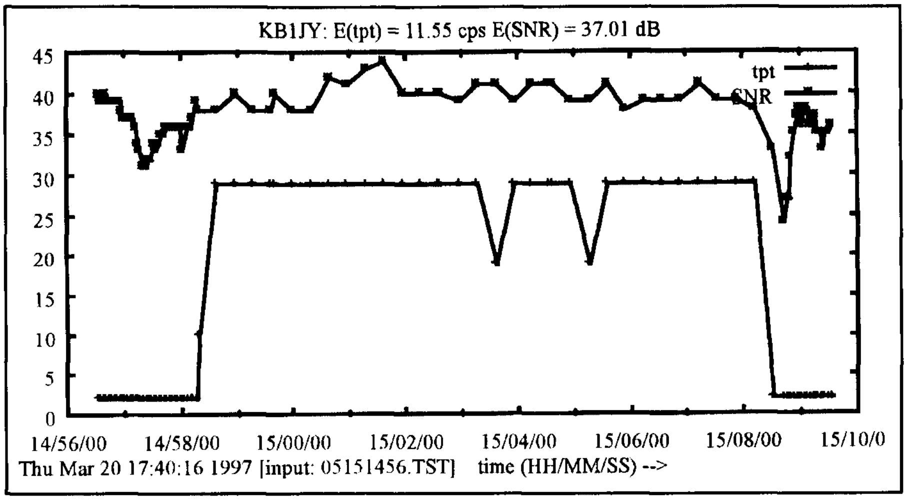

Fig 1—Throughput and SNR for KB1JY during a file transfer.

This graph shows, for example, that KB1JY—the file sender—had a throughput (tpt) of about 30 cps during the transfer with an SNR that varied about an average of 37 dB. (The average throughput and SNR here, calculated by the clvrstst.c program, were affected by the data collected before and after the transfer; these data could have been removed—raising throughput—by editing the *.TST file before running the program.)

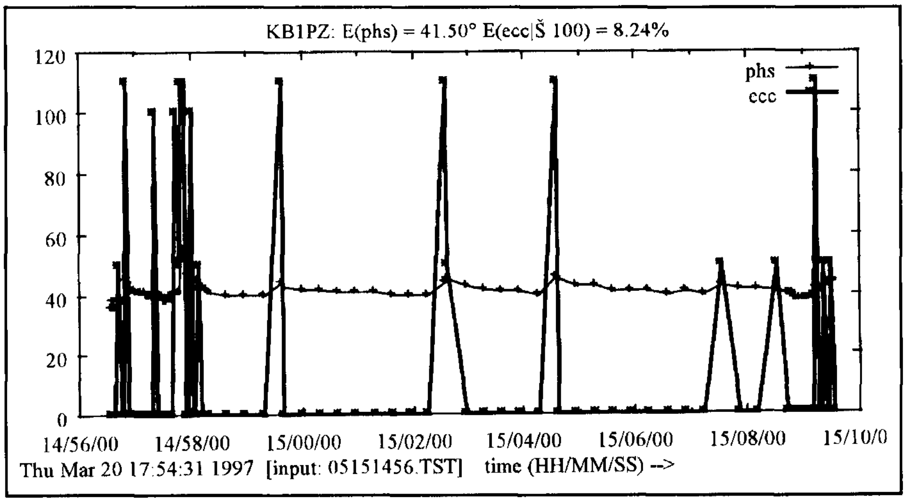

Fig 2—Phase dispersion (see text) and error-correcting capacity for KB1PZ during a file transfer.

Figure 2 depicts the phase dispersion (phs) and error-correcting capacity (ecc) at the file receiver during the transfer. To allow plotting of ecc values of "XX" (ecc exceeded), our software arbitrarily assigns a fictitious percentage of 110 to the "XX" value. You can see that the ecc was exceeded three times during the transfer. The graph's title confirms that the statistics program calculated the average ecc only for percentages less than or

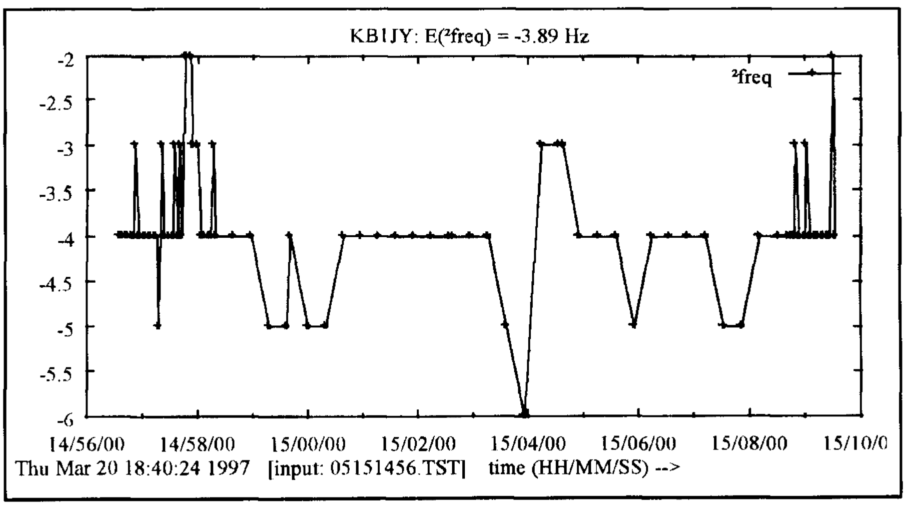

Fig 3—Frequency tuning error for KB1JY during a file transfer.

equal to 100. The figure suggests a possible reason for the sender's two drops in throughput during the transfer: an increase in phase dispersion that caused so many bit errors that CLOVER's Reed-Solomon error-correcting code could not handle them and forced the protocol to ask that a frame be repeated.

Figure 3 shows that KB1JY's radio was tuned about 4 Hz below the carrier frequency during the transfer. The excursion at about 15/04/00 to a tuning error of 6 Hz below frequency was probably related to the multipath that may have caused the throughput drops seen in Figure 1.

5. CLOVER Channel Quality Data for the FEC Mode

CLOVER stations can also communicate in the FEC mode. In the FEC mode, a station broadcasts CLOVER frames that are Reed-Solomon encoded for error correction. Any CLOVER station on frequency can attempt to decode and if necessary correct FEC transmissions. However, if an FEC frame has more bit errors than the R-S decoding can correct, the receiving station does not ask for a repeat, and the receiving operator lives with the erroneous transmission. FEC is sometimes used for sending CQs and to meet other broadcasting requirements.

Channel statistics may also be gathered by an FEC-receiving station. Sample D shows about forty-five seconds' worth of FEC channel data gathered by KB1JY while KB1PZ was sending short ASCII files in the FEC mode. The data fields have the same meaning as in the ARQ case but there is, with no ARQ, no alternation of call signs. That is, channel data are not exchanged by stations.

Notice that the bias is recorded as "0" in the FEC mode and that the 2DPSM (a form of differential BPSK) modulation mode was employed during PZ's manual modulation change at 22:57:37.

Our statistics program can also analyze and plot FEC statistics. Sample E is the output from analysis of the complete fifteen-minute FEC file (relabeled with the extension ".FEC") excerpted above.

Note that the program simply presents all the statistics twice, crediting each line of data to MY and HIS call signs that are in fact the same (KB1PZ). Recall that the channel data these statistics are based on are not sent back to the station sending in the FEC mode.

The gnuplot data and formatting files are also duplicated in the FEC case. Figure 4 shows the throughput and SNR of FEC transmissions from KB1PZ monitored at KB1JY. During the recording, KB1PZ twice changed

Sample D

```
225736,KB1PZ    ,BPSM  ,0,08,032,-2  ,032,000,000
225736,KB1PZ    ,BPSM  ,0,08,032,-1  ,032,000,000
225737,KB1PZ    ,2DPSM ,0,08,032,-1  ,026,000,000
225815,KB1PZ    ,2DPSM ,0,15,007,-1  ,025,000,000
225818,KB1PZ    ,QPSM  ,0,15,013,-2  ,032,006,000
225821,KB1PZ    ,QPSM  ,0,15,018,-2  ,031,006,000
225824,KB1PZ    ,QPSM  ,0,15,022,-1  ,030,000,000
225827,KB1PZ    ,QPSM  ,0,15,025,-1  ,030,000,000
```

Sample E

```
Statistical Summary of CLOVER "status" data.
Input file is [ 04102249.FEC] .
MYcall ("MY") = KB1PZ; HIScall ("HS") = KB1PZ.
Summary date: 20.04.97 00:00:41
Number of reports on KB1PZ = 556
Number of reports on KB1PZ = 556
E(MY_bias) = 0.0, sd(MY_bias) = 0.00
E(HS_bias) = 0.0, sd(HS_bias) = 0.00
E(MY_tput) = 12.43 cps, sd(MY_tput) = 5.92 cps
E(HS_tput) = 12.43 cps, sd(HS_tput) = 5.92 cps
E(MY_SNR) = 31.10 dB, sd(MY_SNR) = 4.39 dB
E(HS_SNR) = 31.10 dB, sd(HS_SNR) = 4.39 dB
E(MY_dfrq) = -3.23 Hz, sd(MY_dfrq) = 2.43 Hz
E(HS_dfrq) = -3.23 Hz, sd(HS_dfrq) = 2.43 Hz
E(MY_phs) = 32.02 deg., sd(MY_phs) = 4.34 deg.
E(HS_phs) = 32.02 deg., sd(HS_phs) = 4.34 deg.
E(MY_ecc) = 3.48%, sd(MY_ecc) = 7.31%
E(HS_ecc) = 3.48%, sd(HS_ecc) = 7.31%
Error correcting capability exceeded 21 times on MY link
Error correcting capability exceeded 21 times on HS link
maximum "MY" throughput = 24 cps, maximum "MY" SNR = 37 dB
maximum "HS" throughput = 24 cps, maximum "HS" SNR = 37 dB
Modulation counts:
MY: 309 BPSM 138 QPSM 86 8PSM
HS: 309 BPSM 138 QPSM 86 8PSM
```

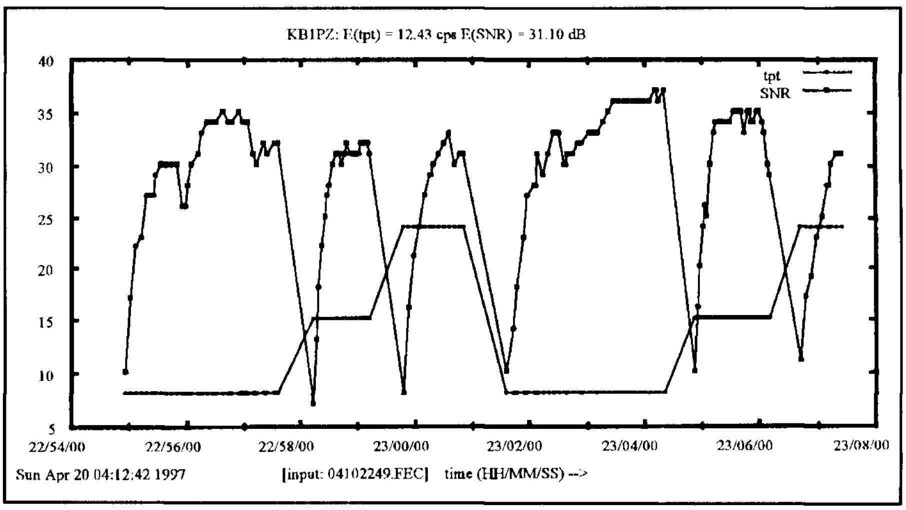

Fig 4—Throughput and SNR of transmissions in the FEC Mode.

his modulation mode from BPSM to QPSM to 8PSM (with corresponding increases in throughput). The short gaps between periods of fixed throughput correspond to pauses between transmissions.

6. What are CLOVER Statistics and Plots good for?

CLOVER channel statistics offer a powerful means for assessing more than throughput and protocol performance. They can also augment conventional means of gauging the effects of antenna changes (for the good or bad) on station performance. Instead of taking a few S-meter and SWR readings and hoping for the best, one can record CLOVER statistics such as SNR over extended periods while connected to one or more CLOVER stations and get a statistically reliable picture of how the changes really affect performance.

Likewise, recordings and plots of the frequency tuning error can expose problems like a dangerously drifting VFO and analysis of phase dispersion and SNR can help pinpoint times (around sunset) when the MUF on a link drops below an operating frequency.

Another fascinating possibility offered by the CLOVER statistics was pointed out by Ray Petit in several early descriptions of the protocol. Clover stations can be used as inexpensive *channel sounders*: turn on the statistics monitoring, connect to a station on a link and frequency you're interested in and record away! When you think you've measured enough, run the statistics through a statistical program like ours, make plots if you want and start to understand what's going on in the ionosphere when you use it.

The "sounder" mode of CLOVER channel monitoring allows one to gather long records of SNR data and thus to assess the variability of SNRs over time. This allows the statistically minded to augment and update the means and variances of the SNRs that are calculated by propagation prediction programs like IONCAP. (IONCAP's SNR statistics are based on variability measurements taken several years ago at a relatively small number of sites around the world.) Updated SNR statistics can lead to more accurate predictions of link behavior and hence to more accurate prediction and modeling of point-to-point and network performance.

Frequency drift in some rigs is detrimental to CLOVER performance (CLOVER can tolerate an offset of about 10 Hz before performance starts to suffer). Monitoring CLOVER's Frequency Tuning Error can often show when drift is a problem. Figure 5 below gives the tuning error at KB1PZ during about forty minutes of CLOVER ARQ operation, starting with a cold transceiver. The rig is an IC-751 with a standard reference oscillator and the noted drift is within the manufacturer's specified tolerances. (The IC-751 can probably be considered typical of solid-state rigs without temperature-compensated references.) The data from the first 15 minutes (up to about 21:55) show PZ's carrier drifting steadily below the nominal carrier. At about 21:55, PZ adjusted his carrier frequency upward. As his rig warmed up further, it continued to drift below frequency. Around 22:15, the rig appears to have achieved stability. These statistics suggest that PZ's VFO needs about 20 minutes of warm up for stable operation.

The channel statistics sometimes lead to discoveries that escape the notice of operators who look only at the CLOVER user interface and its throughput display. A few months ago we noticed in channel data collected during file transfers a curiously periodic rise in phase distortion and fall in SNR that occurred about every ten minutes for every pair of stations. Ignoring the periodicity at first, we barged into investigations of drifting VFOs, cycling air conditioners and a number of more foolish possibilities before realizing that the periodicity was not only real but was caused by the CW ID we had set to be sent every ten

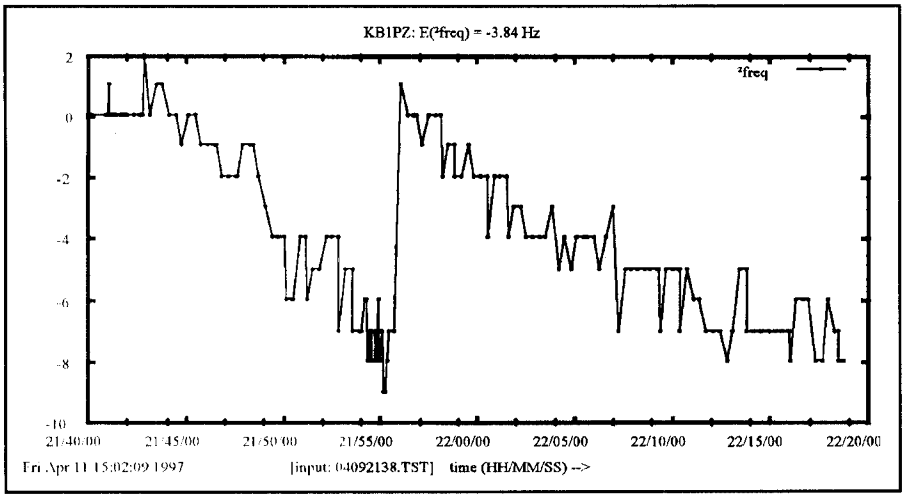

Fig 5—Frequency drift during a file transfer session.

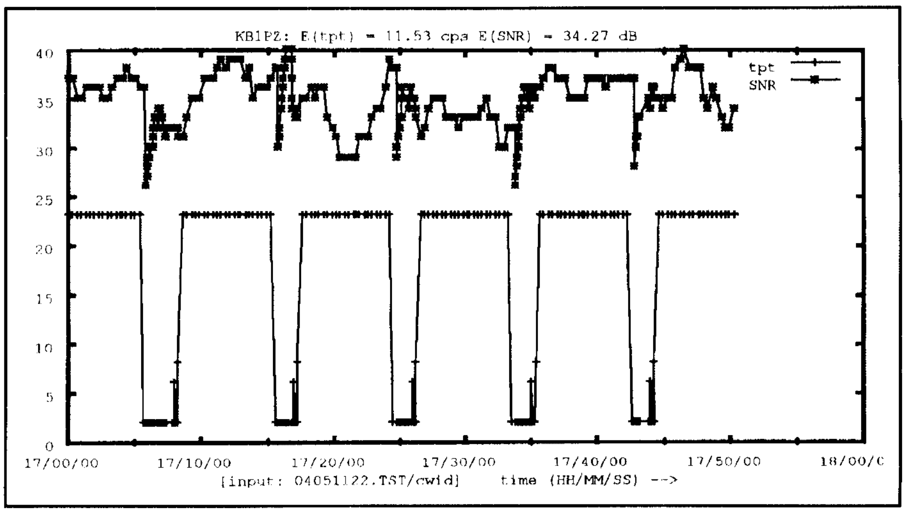

Fig 6—Effect of CW IDs on throughput and SNR during file transfers.

minutes! Since this ID is not sent with the CLOVER protocol, its occurrence temporarily perturbs ARQ operations, including synchronization and efficient demodulation and decoding. This leads to a slight drop in transfer throughput. The sudden drop in SNR on the right side of Fig. 1 was probably caused by the occurrence of a CW ID.

Figure 6 shows the phenomenon clearly. It depicts uncompressed throughput (tpt) and SNR at KB1PZ (reported by KB1JY) as functions of time and is based on an hour's worth of channel data collected during NVIS file transfers from KB1PZ to KB1JY.

7. Concluding Remarks

We hope this discussion of CLOVER channel-quality statistics will encourage other CLOVER users to see the system as much more than just a means of communication. It is also the first inexpensive way for amateurs to combine modern digital communications in the HF band with the study of how well those communications work *and* how the channels they use are changing. Although studying both of these aspects is not for everybody, their combination rides the wave of the future for HF, and the more hams who ride with it the better for the buoyancy of our hobby.

Acknowledgment

I am grateful to Mike Bernock, KB1PZ, Bill Henry, K9GWT, Bob Levreault, W1IMM, and Drew White, K9CW, for useful discussions of CLOVER channel statistics and for comments on the manuscript.

Notes

[1]You can download the clvrstat.c source code from the ARRL "Hiram" BBS (tel 860-594-0306), or the ARRL Internet ftp site: **oak.oakland.edu** (in the **pub/hamradio/arrl/qex** directory). In either case, look for the file CLVRSTAT.ZIP. It's in ANSI C, so you may want to modify the output file names (and perhaps add a command-line interface, clrscr(), etc.) for use with DOS.

[2] Versions of gnuplot for various operating systems can currently be downloaded from **http://www.cs.dartmouth.edu/gnuplot_info.html**. A useful newsgroup that will answer your questions about how to use gnuplot and where to get it is **comp.graphics.apps.gnuplot**.

The Belittler

Come take a look at N4LH's foray into low-voltage receiver design.

By Bill Latta, N4LH

The ZN414 chip is a favorite for building fun receivers for the AM broadcast band, operating on only 1.5 V at 0.5 mA, capable of good DX with only a loopstick antenna. I tuned one such receiver up to 160 meters and was able to copy strong SSB and CW signals by beating them against my bench signal generator.

I was challenged to see what sort of low-voltage ham receiver, if any, I could build using the ZN416,[1] which is a ZN414 with an additional 16 dB audio amplifier built in. Though rated to only 3 MHz, these ICs work to 4 MHz and higher. I soon built an 80/75 meter receiver on a small piece of scrap board, using a trimmer capacitor and toroid coil, with external antenna in place of the typical loopstick. The tiny receiver brought in numerous strong signals in short order.

[1]Notes appear on page 24.

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Louisville, KY 40222

Where, I wondered, could I find schematics for a very-low-voltage oscillator? Data available showed 12, 9, or 6.2 V as typical choices. You never know until you try. I decided to build a regular FET Hartley oscillator on another board and see how well it worked—if at all—at lower voltages. Reducing the dc input from an adjustable power supply showed the circuit oscillating well down to below *one* volt. I tried several 2N4416s and MPF102s in the circuit; all started promptly and kept oscillating vigorously. Ah, I love the thrill of discovery.

The oscillator was built with a good double-bearing capacitor to cover 3.4 to 4.0 MHz, and its output is terminated with a 1000 Ω resistor. When powered from the 1.5 V battery and placed close to the receiver board, SSB/CW signals were readable at all frequencies in the oscillator's range. (The receiver board trimmer, C1, was left peaked at midband.)

Of course, during these tests strong out-of-band stations were leaking through. An RF GAIN pot at the antenna helped a bit. Then I added a high-pass filter, but signals above the band were still very bothersome.

Next came a tunable 3.5-4.0 MHz band-pass filter that was built on another small board.[2] A 365 pF polyethylene-variable capacitor serves as a peaking control, which I labeled Preselect on the schematic (Figure 1). This filter ar-

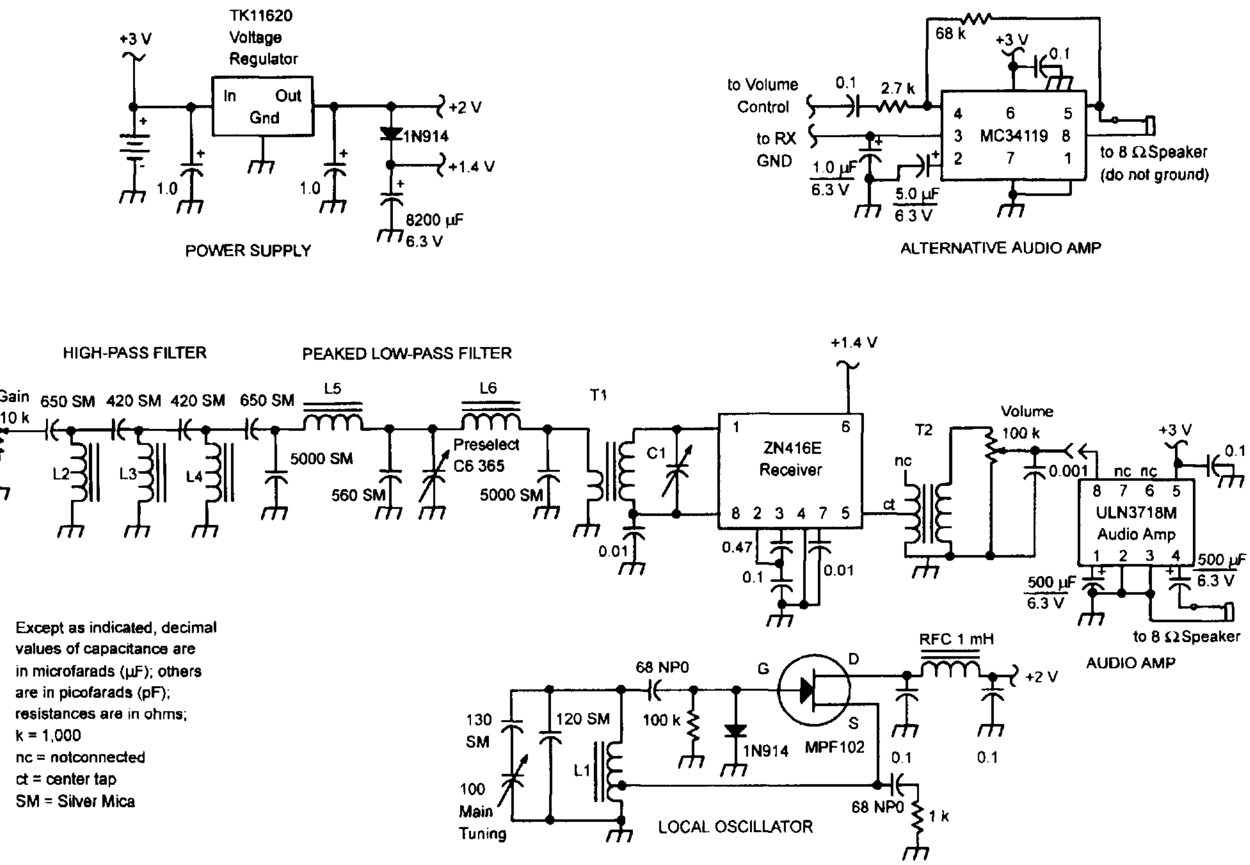


Figure 1—A schematic of the complete receiver and an alternative audio amplifier stage. The local oscillator board is proximity coupled (no physical connection) to the receiver board by spacing the two about 1 inch apart. Use 1/4 W, 5% tolerance carbon composition or metal film resistors. Use 10% tolerance disc ceramic capacitors unless otherwise noted.

C1—25-280 pF trimmer (ARCO464)
L1—43 t #26 enam. wire on T-68-7, or 44 t on a T-68-6 core, tap at 11 turns.
L2, L4—21 t #22 on T-50-2 core
L3—20 t #22 on T-50-2 core
L5, L6—30 t #22 on T-68-2 core
T1—70 t #28 on T-68-2 core with a 2 t link over the cold end
T2—Audio transformer to match low impedance to high. (Argonne AR-153, use one-half of input winding)

rangement does a good job of keeping signals above and below the band from interfering with reception.

Still, I was disappointed because the receiver could not "hear" weaker ham signals. I couldn't find a suitable audio amplifier chip for 1.5 V operation, so I decided to build another module that uses a 3 V audio amplifier IC. (Actually, I built two different audio boards. One uses a Sprague ULN3718M. The second one uses a Motorola MC34119P.[3] Both circuits perform well, as long as there is a matching transformer between the 64 Ω receiver output and the added audio module.) This extra amplification made it possible to hear many more signals.

The final receiver has several stages (see Figure 2). Miniature coax carries

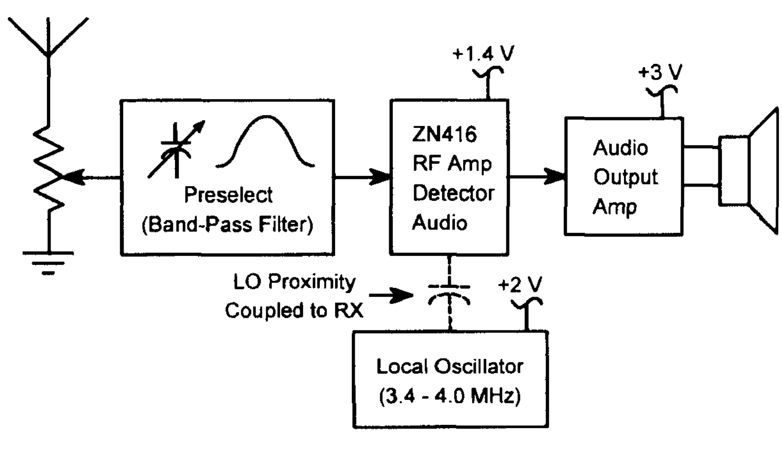


Figure 2—A block diagram of the receiver. Coaxial cable interconnects all the stages except for the LO. No wiring connects the LO and the other stages.

the signals between the various boards that hold the stages. I mounted all those stages in an aluminum enclosure (see Figure 3). A vernier dial slows the tuning rate. Now, an 8 inch speaker in a wooden cabinet (not shown) with a closed back gives comfortable volume, so I no longer need headphones.

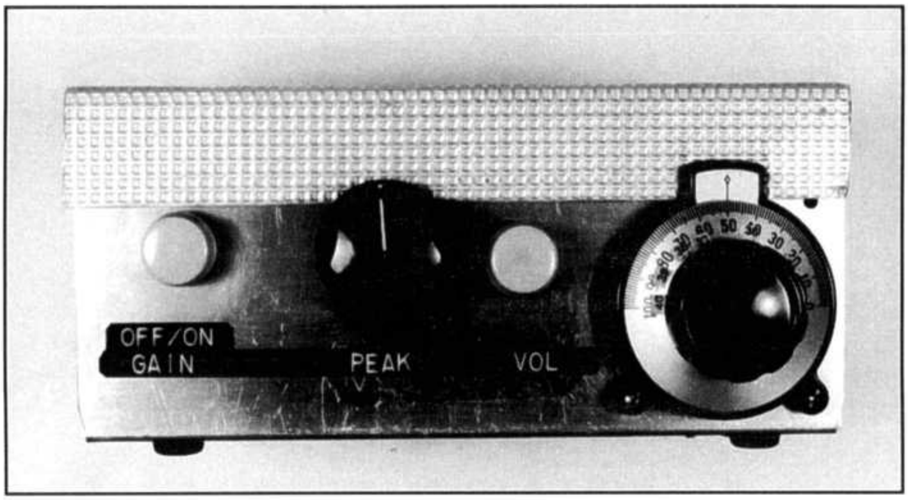
I can hear numerous signals by just connecting the antenna terminal to ground, or with a 20-foot indoor wire and separate ground.

As suspected, in-band selectivity is not always adequate, since this is a TRF receiver. Donald Duck sounds from strong up-band SSB signals may sometimes interfere when I'm listening to down-band CW signals. An audio CW filter could help reduce this problem.[5] Nonetheless, this is a fun receiver. For my purposes and location, I don't need it to be "competition grade." The sensitivity is good enough to hear strong European CW signals; the frequency stability and sound quality are very good.

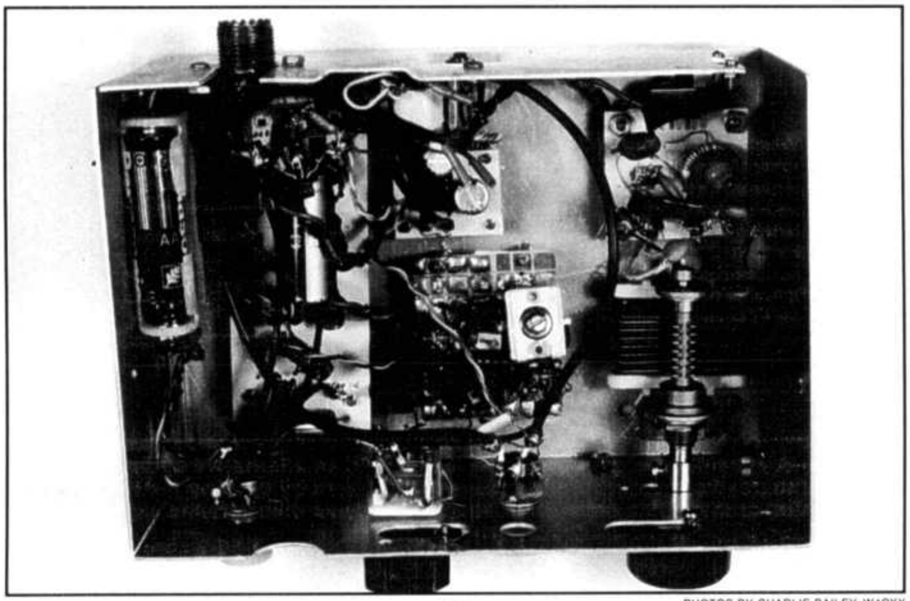
For the power supply, I ultimately used two AA cells. They provide 3 V for the audio stage. A 2 V, low-dropout regulator (TOKO TK11620),[4] powers the oscillator without sacrificing too much overhead voltage. A 1N914 diode connected to the 2 V output drops 2.0 V to 1.4 V for the ZN416 board. This circuit requires a decoupling capacitor to prevent motorboating.

Current drain is only 30 mA at rest, so batteries last me a long time. I named this receiver "The Belittler" for two reasons: It belittles the use of power, and others have belittled me for even trying such a thing. The design is, of course, not altogether proper. Perhaps someone could build an improved model using some of the ideas presented here—maybe a superhet running totally from 1.5 V.

There are many new ICs now being produced that will operate with low voltage and current requirements. Why not develop more efficient receivers? Light or other alternative energy sources could more easily power such equipment. At age 77, I may not be able to do much more of this work. So to others out there, the torch I throw! Bear it high, but keep the power low!



(A)



PHOTOS BY CHARLIE BAILEY, W4OXX

(B)

Figure 3—(A) shows the Belittler's front panel. (B) is a view inside the case.

Notes

[1, 3]ICs and data sheets from DC Electronics, PO Box 3203, Scottsdale, AZ 85271-3203.

[2]From *ARRL Handbook for the Radio Amateur* (Newington: ARRL, 1985), 62nd edition, page 30-11, Fig 15: 80-meter Preselector Filter.

[4]From Digi-Key Corp, 701 Brooks Ave South, Thief River Falls, MN 56701.

[5]After submission of this article, the author installed an 88 mH toroid with a 0.47 μF capacitor across it, and a DPST switch on the back of the housing. This "tuned audio circuit" connects across the outer terminals of the VOLUME control. When I switch the circuit in and tune CW signals to peak at about 750 Hz, there is a substantial improvement in the selectivity. The circuit also greatly reduces static noise.

RF

By Zack Lau, W1VT

VHF LO for the KK7B 23-cm No-Tune Transverter

With the impending launch of Phase 3D, there is renewed interest in 1269-MHz transmitting gear for accessing the satellite. Current predictions are that about 10 W to a RHCP helical antenna ought to work fine.

One popular transverter for getting on the band is a no-tune design by Rick Campbell, KK7B, that appears in the *ARRL UHF/Microwave Projects Manual*. Copies of the circuit-board artwork for noncommercial use are available from the ARRL Technical Department Secretary for an SASE. Unfortunately, current technology doesn't permit us to receive envelopes via telephone or e-mail. This transverter puts out about 10 mW. The easiest way to get to 10-W output is to use a cascaded pair of Mitsubishi hybrid amplifiers—an M67715 driving an M57762. A bit expensive, but there is enough power out of the driver to remotely mount the power amplifier and overcome as much as 4 dB of cable loss. Alternately, Greg McIntire, AA5C, has published "+28 dBm 1296-MHz Power Amp for No-Tune Transverters" in the *1995 Microwave Update* using a Motorola MRF 8372 driving a Siemens CLY5 GaAs FET.[1] Another option is to look for cheap surplus power GaAs FETs and tune them down to 1269 MHz. The impedance-matched versions typically have low-pass networks that essentially disappear at low frequencies.[2] Keep in mind that power GaAs FETs are supposed to have properly sequenced negative bias supplies—those hybrids are really the easiest way to go.

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[1]Notes appear on page 29.

A Simple 230-MHz LO

Most people use the 540- to 580-MHz local oscillator designed by Rick that appears in Chapter 5 of the *Projects Manual*. It conveniently uses 90 to 96.6-MHz crystals, allowing alignment with nothing more than an FM broadcast receiver. However, if you are willing to use a slightly higher frequency, the simpler circuit of Fig 1 will work. The second harmonic of a 112.5-MHz or 115-MHz crystal oscillator is used to drive the transverter. The simpler circuit is possible because the UHF LO was originally designed for a 13 or 9-cm transverter that needed the higher frequency. Rather than design a VHF LO from scratch, a known good design was pressed into service.

I used these frequencies because I already had the crystals. You will probably want to choose different frequencies to more closely meet your needs. Frequency choice is an important consideration for full-duplex satellite operation; you don't really want your transmit and receive IFs to overlap. You might want to move them as far apart as practical, to minimize interference.

An important consideration is the frequencies used by nearby stations. 145.01 MHz might be a very bad choice if it's used by a busy packet station. The same is true with busy 2-m repeater frequencies.

Can seventh overtone crystals be used instead? I have been able to get the circuit to oscillate just fine at 190 MHz with a 47-Ω resistor instead of the crystal and matching parallel resonant inductor after scaling L1, C3 and C4, but it didn't want to work with

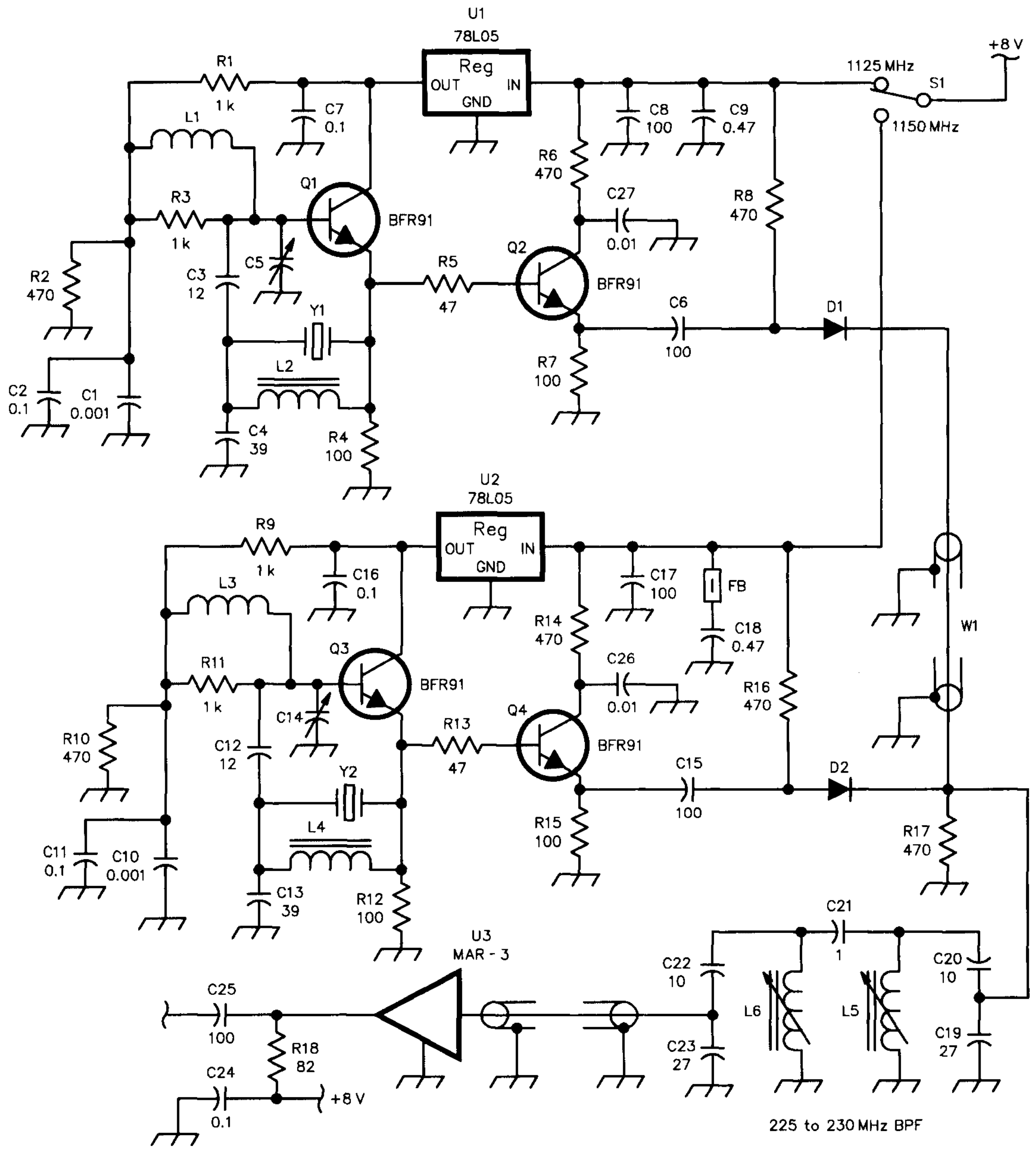

Fig 1—Schematic diagram of the VHF LO for KK7B's 23-cm transverter.

C5, C14—0.8 to 10 pF trimmer capacitor. Precise range not critical.
FB—FB-64-101 ferrite bead. See text.
D1, D2—HP 5082-3188 PIN diodes. Any inexpensive PIN diode ought to work here.
L1, L3—7 turns #28 enameled wire, 0.11 inch inside diameter (ID).
L2, L4—Parallel resonates with crystal's parallel capacitance at desired overtone frequency. 15 turns # 28 enameled wire on T-30-10 iron powder toroid.
L5, L6—Digi-Key TK2802-ND 52 nH tunable inductor. Toko E528SNAS-100073.
W1, W2—UT-085 semi-rigid 50-Ω coax—see text.

the seventh overtone crystal. It turns out that the crystal had a high series resistance, about 137 Ω. It was necessary to increase the feedback to compensate for this much higher loss. You have to be careful about increasing the feedback though—too much and you can damage the crystal. The output of Q2 should to be around +1 dBm. I measured the series loss by measuring the insertion loss in a 50-Ω system, then compared these measurements against an Amateur Radio Designer computer simulation.

The ferrite bead next to C18 is shown to illustrate a technique for preventing parallel resonances. If you use a pair of high quality capacitors next to each other, it is quite possible that the high series inductance of the larger capacitance may resonate with the smaller capacitor, forming a parallel-resonant circuit. This is precisely what you don't want—you want a low impedance at all frequencies. The ferrite bead effectively looks like a resistor at RF, reducing this parallel resonance effect.

Deciding how to switch frequencies is a trade-off. Probably the cheapest way to switch frequencies is to just switch crystals with a mechanical switch. I don't recommend it because there are two serious problems. The most serious is the difficulty in tuning the oscillator—you need to find an

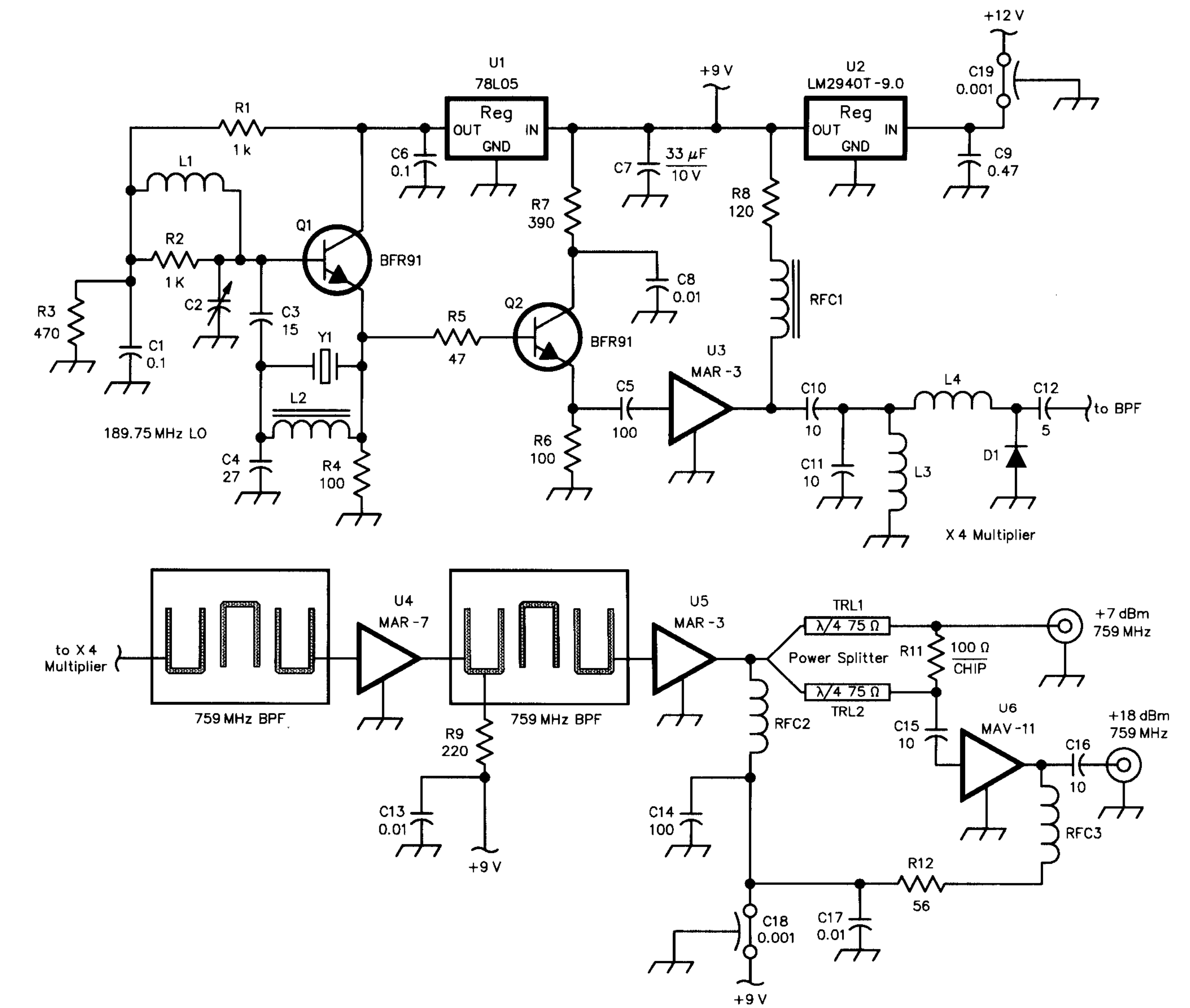

Fig 2—Schematic Diagram of a 759-MHz LO.

C2-0.8—10-pF trimmer capacitor, exact range not critical.
D1—HP 5082-2835 Schottky diode.
L1—4 turns # 28 enamel wire, 0.113 inch inside diameter (ID). Length is 0.2 inches.
L2—Parallel resonant with the stray capacitance of Y1 at 190 MHz.
L3—8 turns # 24 enamel wire, 0.188 inch ID.
L4—5 turns # 24 enamel wire, 0.102 inch ID.
RFC1—8 turns # 28 enamel wire on FT-23-63 ferrite core.
RFC2, RFC3—5 turns #28 enamel wire 0.102 inch ID.
TRL1, TRL2—2.72 inch lengths of 75-Ω, 0.70-VF conformable Teflon coax.
Y1—International Crystal Manufacturing Co, 477590, 189.75-MHz seventh overtone high-accuracy crystal for operation at 26°C.

acceptable compromise when tuning both frequencies. The second is that it puts extra stray reactances where it is least desirable, right at the frequency determining components.

I chose to use PIN diode switches at the output of the buffer amplifiers. While it uses more parts, it allows you to optimize the circuit for each crystal. It is advantageous to use "cold dc" switching—the circuitry is controlled by dc voltages, therefore minimizing spectral purity and EMI problems. This lends itself to remote-control applications, which is useful if you decide to mast mount the transverter.

The LO puts out about +7 dBm, though it dropped off to +5 dBm when I cooled it down to −24° C (−11° F). More output can be obtained by adding an MAR-4 or MAV-11 MMIC. Whether you need to add a low-pass filter depends on the circuit you are driving. It probably doesn't make much difference for single diode multipliers, however, it can make a difference with a number of other circuits that rely on a symmetrical sinewaves.

Interestingly enough, Rick Campbell's harmonic generator, using a 74ACO4 digital gate to develop a square wave, requires a symmetrical signal. If the output isn't a 50% duty cycle, you will get even harmonic energy, making the filtering of the desired output signal more difficult. Actually, what I said isn't quite right. You need whatever input signal is needed to generate the required symmetrical output signal. Obviously, a simple dc offset may be required to produce the most symmetrical waveform.

Construction:

I built the prototype on the groundplane side of the 1296-transverter board using point-to-point construction. This is an excellent technique for trying out prototypes—it is easy to add more filtering if necessary. The 115- and 345-MHz spurs were down 49 and 45 dB, respectively. If you need a cleaner signal, just add another 230-MHz band-pass filter. Trying to get 60-dB spurious suppression out of a no-tune transverter requires either external filters or much better filter shielding. Rick Campbell described how to put shields over hairpin filters in the *Proceedings of the 1993 Microwave Update.*[3]

The sharp reader will note two coax jumpers, W1 and W2. You might consider these to be microwave hook-up wire, used for joining two 50-Ω circuits together without creating an excessive impedance bump. Small semirigid coax works a lot better than RG-174 because there is little danger of melting it with the soldering iron having the shield short to the center conductor. You can even polish the shield with steel wool so it looks nice and takes solder easily.

While I found that it wasn't really necessary to tune the VHF filter inductors, you ought to position them to allow tweaking. I did this by keeping a nice clear area for the tuning stick. Mine were mounted on their sides. Alternately, some people prefer to put access holes in the side of the box.

A High-Quality 759-MHz Local Oscillator

This local oscillator design features a very high degree of spurious suppression—spurs within 700 MHz of the carrier are at least 76 dB down. This is quite a bit better than most published designs, though it may not seem all that surprising once you see the work required to achieve it.

I decided to start off with a 189.75-MHz oscillator to minimize the number of spurious signals that would need to be filtered out. Also, a 4× multiplier typically works a lot better than an 8× multiplier. I did this by scaling the typical Rick Campbell LO design and verifying it worked with a 47-Ω resistor in place of a crystal. When I plugged in the crystal, it didn't work. Measuring the crystal's series loss in a 50-Ω

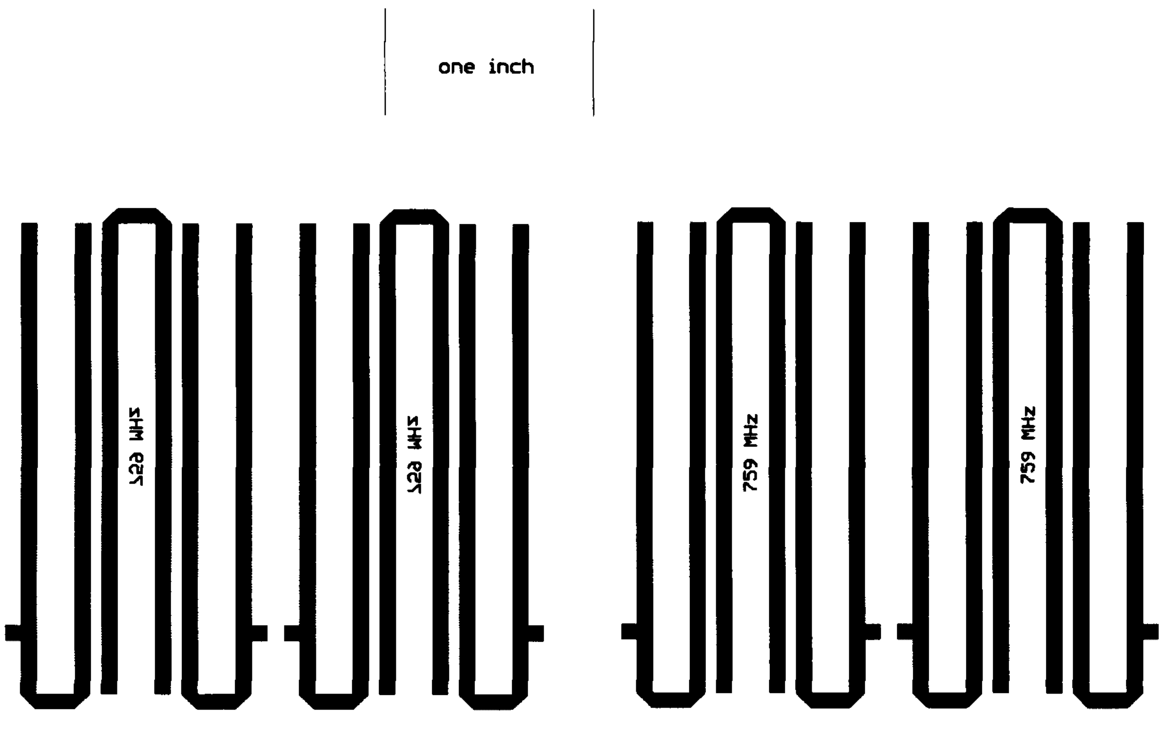

Fig 3—Circuit board etching pattern. The board material is 1/32 inch teflon with a dielectric constant of 2.55.

circuit, I found it had 7.5 dB of loss, which meant it had about 137 Ω of resistance, considerably higher than the 47 Ω the circuit was designed to have. I compensated for this by changing L1, C3 and C4, to increase the feedback till it worked just fine with a 150-Ω resistor. It oscillated just fine when I replaced the resistor with the crystal and a parallel resonating inductor.

The diode multiplier is a typical multiplier circuit that uses an inexpensive Schottky diode. The output is filtered with pair of hairpin filters. Instead of using the usual glass epoxy board, I chose to use woven Teflon, which provides lower loss and more repeatable performance. This allows a narrower bandwidth filter to be constructed. A narrower filter needs fewer filter elements to get the same attenuation of unwanted multiplier products, at least for the first 30 or 40 dB of attenuation. Models predicting more attenuation often fail because they don't correctly account for coupling around microstrip filters. Very narrow bandwidth filters often have problems with either excessive loss or fabrication tolerances. A really sharp filter isn't terribly useful if it eliminates the desired signal.

I found it necessary to install a shield between the filters. Because of the size of the filters, you have to worry about waveguide propagation, even if the slot is really thin. I found it necessary to install screws to break up the thin slot; four screws spaced an inch apart seemed to do a good job.

Like the oscillator and multiplier, I decided to construct the LO splitter and amplifier using point-to-point construction on the ground plane side of the filters. This made a compact assembly. The 70-Ω lines for the splitter are made of conformable coax. This is coax with a somewhat flexible tinned-copper shield that can be easily formed into shapes and soldered in place. RG-179 B/U can be used as a cheap substitute.

There is a good reason for putting low-pass filters at the inputs of mixers and multipliers, as opposed to including them at the outputs of the preceding stages. Diode mixers and multipliers will often reflect signals back at the source—often after clipping them and generating harmonics. Of course, a low-pass filter reflects these harmonics back to the diodes that generated them. If you vary the length of the connecting cable, you also vary the phase, often resulting in a substantial effect on circuit performance. The reflections can be minimized by adjusting the length of the transmission line between the filter and nonlinear device.

Construction:

The circuit board was etched using the PnP-Blue iron-on resist process sold by Techniks Inc of Ringoes, NJ.[4] While it works fairly well, I don't think any of the laser printer or photocopier techniques works well with large areas of solid copper. I find it easier just to mask off the ground area around the board with plastic mailing tape. I do this more for mechanical than electrical reasons—soldering both the top and bottom foils to the wall of the box results in a stronger assembly. I drilled three countersunk holes, two for the input and output and a third for the MAR-7 bias resistor. The countersinking cleared the ground-plane foil from the hole. The MAR-7 was mounted by bending the leads against the body of the part and bringing them through the 0.094-inch hole and soldering them against the ground plane.

I made a box around the Teflon circuit board using unetched double-sided circuit board. In addition to the shield between the filters, I also found it necessary to put a shield to separate the 189.75-MHz local oscillator from the 759-MHz amplifier and splitter circuitry constructed on the ground-plane side of the Teflon board. The feedthrough capacitor supplies +9 V to the 759-MHz amplifiers. Both of the shields attach to the cover using four 2-56 screws spaced one inch apart. The screws were needed to obtain the clean output signal. The oscillator also put out a clean signal without any covers, so it's unnecessary to do the initial testing and debugging with the covers installed. Installing just the covers reduced the spurs to roughly –40 to –49 dBc.

Notes

[1]McIntire, Greg, AA5C, "+28 dBm 1296 MHz Power Amp for No-Tune Transverters," *Proceedings of Microwave Update '95*.

[2]Malowanchuk, Barry, VE4MA, "Using IMFETS at other than rated frequencies," *Proceedings of Microwave Update '91*, pp 30-33.

[3]Campbell, Rick, "A Single Board No Tune Transceiver for 1296?," *Proceedings of the Microwave Update '93*, pp 17-38.

[4]Techniks Inc, PO Box 463, Ringoes, NJ 08551 orders 908-788-8249, fax 908-788-8837.

Letters to the Editor

New Harris Parts for DDC-Based Receiver

◊ Harris now makes a digital upconverter (DUC), the HSP50215VC, and a transmitting DAC, the HI5741BIB. These chips, when combined with the DDC-based receiver, permit the construction of a digital HF transceiver for the 160 through 15 meter bands.

For the transmit signal chain, the microphone audio is converted to bandlimited low-sample-rate I and Q signals either in a DSP or by using the ADC and DDC in the receiver. The I and Q signals are then applied to the DUC, which increases the sample rate to 50 MSPS and upconverts the signal to the desired RF passband. The DAC converts the DUC output to an analog signal that drives the transmitter linear amplifier stages.

From the spec sheets, the analog signal should have an SFDR of 70 dB or more, so the final signal purity will be limited by the linear amplifier chain.—*Peter Traneus Anderson, KC1HR, 990 Pine St, Burlington, VT 05401; e-mail* **traneus@emba.uvm.edu**

Interest in Cell Phone Conversions?

◊ I just read "Empirically Speaking" in the September *QEX*, and I like the idea of publishing correspondence. I, for one, would like to see what others are doing to experiment, in addition to Zack Lau and Peter Anderson. Not that they are not major contributors to the hobby, they are; but what is everybody else doing? Are most too busy(?), too shy, or too lazy to write about their projects?

I am trying to find interest in 900 MHz and conversion of cellular phones for use by amateurs in that band. Locally, there has been little interest when I presented the idea at ham club meetings. The response can be summarized as "If it won't get me more DX on 20, I don't want it!" I have converted two units, both now on 906.5 FM simplex. One is monitors the frequency at all times; it's connected to a voice actuated tape recorder in case of activity while I'm at work. So far, there's nothing on the tape (over the last three months).

Are *QEX* readers interested in this topic? If so would they like a *QEX* article about the conversion process? Thanks for an interesting journal!—*Matt Kastigar, NØXEU, 2118 Parkridge Ave, Brentwood, MO 63144-1638*

QEX invites you to share your ideas and comments with fellow hams. Send them to "*QEX* Letters to the Editor" c/o ARRL, 225 Main St, Newington, CT 06111-1494; e-mail **rseverns@arrl.org**. Please include your name, call sign complete mailing address, daytime telephone number and e-mail address on all correspondence. Whether praising or criticizing an item, please send the author(s) a copy of your comments.