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OCTOBER 1990

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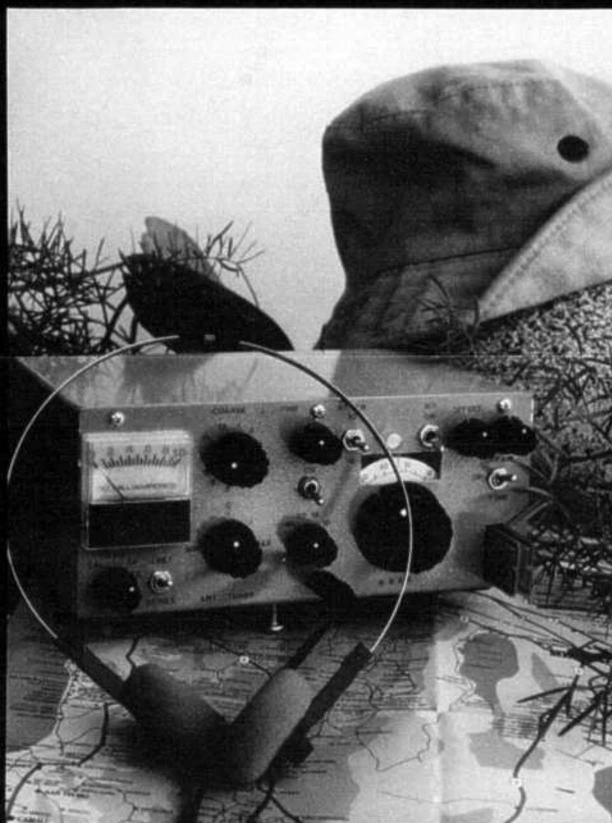

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Purposes of *QEX*:

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 double 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*.

Empirically Speaking. . .

London on Thames—Ontario

The London Regional Art Gallery and Museum, overlooking the Thames River in London, Ontario, might seem like a strange venue for the ARRL/CRRL 9th Amateur Radio Computer Networking Conference on September 22. It couldn't have been nicer. CRRL Vice President Harry MacLean, VE3GRO, and David Toth, VE3GYQ, were responsible for the hosting and moderating of the conference. If you attended, you might want to take the time to give them your personal thanks for a job well done.

A list of the papers presented at the conference is given later in the Gateway column, starting on page 25. As usual, the quality of the papers was high and should be a source of pride in the technical prowess of hams of the digital persuasion. If forced to single out one paper that stands out, it would have to be "CELP High-Quality Speech Processing for Packet Radio Transmission and Networking" by A. Langi, and W. Kinsner, VE4WK. But then again, the PACSAT papers by Harold Price, NK6K, are examples of some adroit protocol strategies to cope with the fact that Microsats are in a sky near you only for a few minutes at a time. Analysis showed that the usual procedure of having each station get a separate transmission of bulletin material from the satellite would not be possibile, owing to the limited time of each pass. Instead, a form of "broadcast" protocol was developed to give everyone a chance of receiving everything for them on one pass.

The ARRL Committee on Amateur Radio Digital Communication met on September 23 to discuss packet and other issues. Joining the group was new member Dale Sinner, W6IWO, a name that is very well known to RTTYers. At the top of the agenda was AMTOR. Digital Committee member Paul Newland, AD7I, has been working with manufacturers to review CCIR Recommendation 625 for possible amateur contributions to improvement of the protocol. Also, Paul and industry have been studying the introduction of adaptive ARQ into AMTOR. This technique was developed by the USSR maritime administration as a way of improving the efficiency of maritime SITOR communications. Soviet tests showed a 20% increase in throughput over non-adaptive ARQ. The basic idea is that when two stations make contact, a handshake occurs if both stations can operate the optional mode which permits transmission of 9 characters in a block instead of 3 if circuit conditions warrant. The normal 3-character mode is known as ARQ3, and the 9-character as ARQ9. If only one station is capable of ARQ9, both stations use ARQ3 as the default mode.

Committee member Eric Scace, K3NA, stayed away from the meeting in order to spend the weekend finishing AX.25 version 2.1. Eric reported that he had the new SDL diagrams nearly complete. The plan is to circulate the SDLs to a short list of reviewers prior to publication of a new AX.25 protocol document.

Glenn Elmore, N6GN, one of a number of observers at the meeting, talked about the future of Amateur Radio from several viewpoints: the role of Amateur Radio in society; services amateurs may want from future systems; and, possible system architecture. Some urgency for work on future systems was punctuated by the FCC's news release, just 3 days earlier, on their Second NOI on WARC-92, wherein they discussed an option of using the 2400-2450 MHz band for the broadcast-satellite (sound) service and a complementary terrestrial service—also known as digital audio broadcast (DAB).

Observers at the meeting asked about plans for the 10th Computer Networking Conference. There had been some discussion of going back to the Washington, DC area, where the first CNC was held in 1981, but there had been no follow-up action. Several other venues were talked about, and it was left that the Digital Committee would be willing to consider any *hard* proposals. The sooner the better, as planning the hosting end of the conferences takes time. Also, the Digital Committee's recommendation for a conference requires the approval of the ARRL Board, which meets next in January, 1991, or the ARRL Executive Committee, which meets more often. January, 1991 would be a good time to have such approval so that a call for papers can be issued giving prospective authors time to do their research and writing.—*W4RI*

Modifications and Improvements to the HW-9

By S. W. McLellan, ND3P
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Several articles have been written dealing with modifications and improvements to the HW-9.[1,2,3,4] After reading them and making my own changes/improvements/corrections to the rig, I thought I would share the modifications I made and the difficulties with the changes suggested by others.

Although the HW-9 "out of the box" played well, several problems were detected and many enhancements were desired. Most importantly, problems, such as instabilities, had to be overcome to make the HW-9 stable and reliable.

To avoid duplicating the entire HW-9 schematic, portions of the schematic are redrawn to illustrate the changes I made. It will still be necessary to refer to the complete HW-9 schematic to get the "whole picture." In addition, the original components are referred to by Heath's component designation numbers.

The most annoying (and destructive) problem was instability and overheating in the power output stage. My solution is shown in Figs 1 and 2. I (and others—see Ref 1) have noticed that the power output stage of the HW-9 suffers from instability problems. In particular, I noticed that a mismatch to the rig might "induce" oscillations in the power output stage, sometimes at VHF. The oscillation was evident from excessive current consumption, FM radio interference, or by a sudden increase in output power (as measured by a watt meter) as the drive level is increased. As will be discussed below, I found that I could not rely on the HW-9's output meter to measure the relative output power of the HW-9; AGC and IF amplifier problems interfered with the output meter's readings.

When I asked Heath about the power output amplifier instability problem, they said that Q402 should be a 2N5770 (417-293) and that would solve the problem. They happily sent me the replacement transistor and I installed it; the problem didn't go away.

I found that the main problem with the power output stage is the lack of adequate RF bypassing. Heath provided a single 0.1-μF ceramic bypass capacitor (C445) between the output transformer T403 and the two-

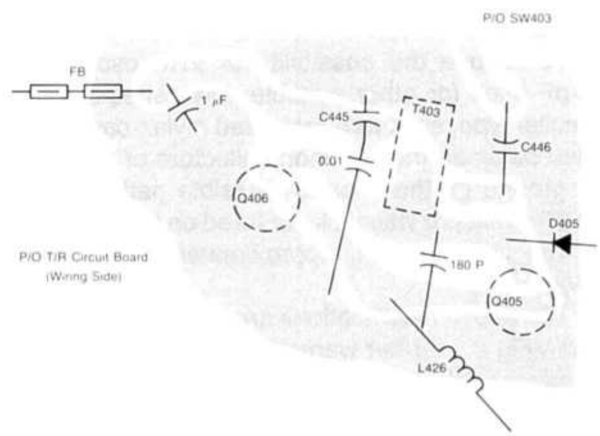


Fig 1—P/O T/R circuit board (wiring side view of power output stage)

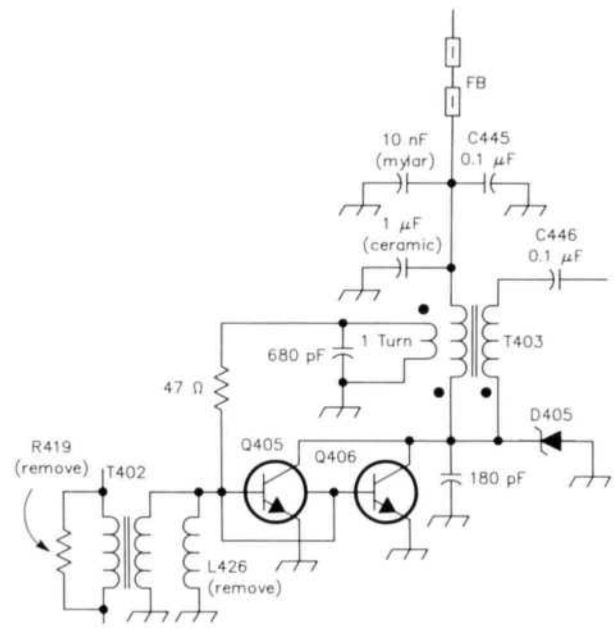


Fig 2—Power output stage

[1]Notes appear on page 9.

ferrite-bead choke. Further, the ground side of C445 is as far away from the emitters of Q405, Q406 (the power output transistors) as seemingly possible. Too much inductance in the ground path can lead to instability. I beefed up the bypassing by adding a low series inductance 0.01-μF mylar capacitor from the "hot side" of C445 to the ground trace nearest the emitters of Q405, Q406. This is shown in Fig 1, a magnified bottom view (wiring side) of the T/R circuit board. Note that the 0.01-μF capacitor provides the shortest possible path back to the ground and the emitters of Q405, Q406. In addition, I added a 1-μF ceramic bypass capacitor to improve the low-frequency bypassing. Both the 0.01 μF and 1 μF capacitors are tack-soldered on the underside of the T/R circuit board in approximately the positions shown.

To reduce the possibility of VHF oscillations, a 180-pF mica (or other suitable, low series-inductance capacitor type, such as a metalized mylar) capacitor was added between the common collectors of Q405, Q406 and ground by the shortest possible path. Again, the 180-pF capacitor was tack-soldered on the underside of the T/R circuit board, approximately in the position shown in Fig 1.

The above modifications are also shown schematically in Fig 2. I further wanted to "flatten" out the gain-vs-frequency characteristic of the power output amplifier so that the amplifier is more stable at low frequencies without reducing the gain thereof at high frequencies. For this I added frequency-dependent feedback. As a result, I have been unsuccessful in coaxing the amplifier into oscillation with the load mismatched. I removed R419 (330 ohms) and L426 (10 μH) and added a 47-ohm resistor to the bases of Q405, Q406. The 47-ohm resistor now acts, essentially, as the base load for the power output amplifier, removing the need for R419. Negative feedback is provided by a loose, one-turn, winding added to T403. In my HW-9, T403 is provided by Heath premade and the one-turn winding is wound such that the lead to the 47-ohm resistor is wound through the center of the toroidal transformer T403 and the other end of the winding is grounded. Check your T403 to make sure of the correct winding direction. The 680-pF ceramic capacitor reduces the negative feedback at high frequencies, such as on the 12 and 10 meter bands. I grounded the one turn winding to one lead of the 680-pF capacitor and soldered that lead to the ground trace in the circuit board by drilling a hole nearest the emitters of Q405, Q406. Remember, use the shortest leads possible to get the job done.

In Ref 1, emitter degeneration resistors are suggested to help with stabilizing the power output amplifier. I found that they were not necessary. Motorola provided emitter ballasting resistors inside of each transistor so external ones are redundant and significantly reduces the available output power of the amplifier.

I also added extra heat sinks to both transistors (I super-glued another finned heat sink onto each existing heat sink) which greatly reduced the case temperatures thereof. In addition, I heat sinked the driver transistor, Q404, to cool it down.

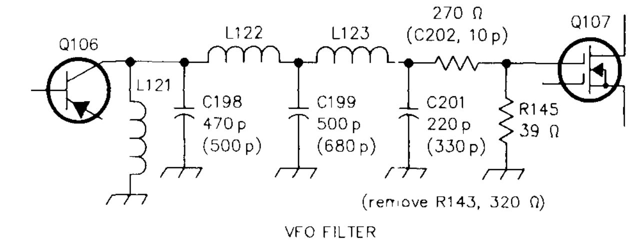


Fig 3—VFO filter

Stability of the VFO was greatly improved by soldering one end of a short piece of wire (I used a thin braided wire like Soder-wick®) to the frame of the variable capacitor C1 and the other end to the VFO shield. This was suggested in (Refs 1 and 3) and it went a long way in making the HW-9 easy to tune—especially when using the narrow filter.

One very annoying problem was the dropping off of output power from the HW-9 when tuning to the low-frequency end of each band. Similarly, the sensitivity of the receiver also decreased at the low end of each band. A major contributor to both problems was the VFO filter shown in Fig 3. A computer model of the Heath version of the VFO filter showed an approximately 4-dB variation in output signal voltage over the desired passband (about 5.75 to 6 MHz). Further, the harmonic suppression was not as great as it could be due to a high-pass coupling arrangement between the output of the filter and the first mixer. The new filter design utilizes the old inductors (L122 and L123) and changes the capacitors—only two new capacitors are needed, a 470 pF and a 220 pF. Preferably, the capacitors have a low temperature coefficient (NP0) or are of mica. The 500-pF capacitor (C198) is reused and moved to C199's position. The capacitor C202 is removed and a 270-ohm, ¼ watt, resistor is used instead. R145 is changed to 39 ohms, ¼ watt, and R143 is removed. The 270- and 39-ohm resistors reduce the signal level to the first mixer from the VFO filter to approximately that before the above changes were made. Note that to avoid static discharge problems, C202 should be changed first and then R145. The resulting filter has less than 0.5 dB of ripple over the desired passband and the second harmonic suppression is now about 45 dB below the fundamental. This change reduced the level of "birdies" on all bands and substantially flattened out the output power vs VFO frequency problem mentioned above.

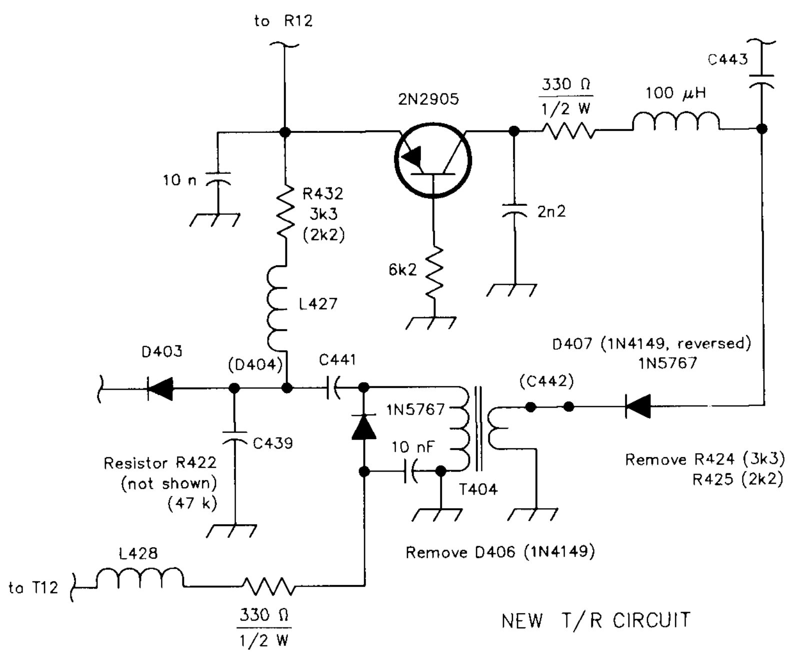
A perplexing problem showed up when I was monitoring the output of the HW-9 on another receiver and the output power varied. As the output power increased, the transmit frequency changed almost 1 kHz from minimum power to full power. Initially I thought it was due to power supply voltage droop. Instead I found that the RIT circuit (C179, D118, R126, etc) was being affected by the T/R circuit. When the output power is increased, the output signal through C443 forces diode D407 into conduction, placing a negative voltage on the R12 control bus. When the voltage on the R12 bus (+12 on receive) became sufficiently negative (I measured over 10 volts negative at full power), the emitter-base junction of Q103 breaks down and varies the voltage applied to the RIT control diode D118. Placing a diode in series with the emitter of Q103 would solve the problem but still applied negative voltage across electrolytic capacitors (which I removed anyway, discussed below) on the bus. The solution I chose was to redesign the T/R switch to solve the problem and improve the receiver performance. This is shown in Fig 4. Instead of D407 dragging the R12 bus negative, D407 is reversed, C442 is removed, and the additional transistor passes current to the D407 when in receive and isolates the diode during transmit. D407 is changed to a PIN diode so that it acts less like a rectifier and has a lower series resistance than the old diode when forward biased. Now, more of the input signal gets through to the transformer T404 during receive. Diode D406 is removed (it operated as a clamp during transmit) and another PIN diode is used as a shunt during transmit. During receive, the diode is unbiased, acting as an open circuit. During transmit, the lower forward resistance of the PIN diode further increases the isolation of the transmitted output signal from the rest of the receiver. With this design, only one diode (D403) instead of two (D403 and D404) is used for isolation. D404 is replaced with a strap and R422 is removed as it is no longer needed. The result is that on 10 meters I can now hear the background noise when I attach an antenna. Note that in Ref 1, the author suggests replacing the diodes with Schottky diodes; *don't do it*—it will worsen an already bad problem and the overload characteristics of the receiver will suffer having diode D406 with a lower forward voltage drop as a clamp across the high-impedance secondary of T404.

I also found an interaction between the transmit return adjustment (R131) and the RIT control due to the emitter-base junction of Q104 breaking down. To eliminate it, I placed a diode (here a germanium diode, such as a 1N3666) in series with the emitter of Q104 (anode to emitter).

Stability of the AGC amplifier was another major concern. I notice that the output of the AGC amplifier U302 was weakly oscillating. Heath evidently forgot to add a bypass capacitor across the power supply to U302; tacking a 0.01-μF ceramic bypass capacitor (circled in Fig 5) from pin 7 of U304 to ground solved the problem. The AGC set point had to be readjusted afterward. In addition, no compensation capacitor on U302 is provided. The lack of a compensation capacitor caused U302 to oscillate when it suppled current to C317 during increasing received signal strength. Adding a 100-pF ceramic capacitor from pin 1 to pin 8 on U302 (also circled in Fig 5) on the underside of the T/R circuit board eliminated the oscillation.

In Ref 1, a suggestion was made to increase the capacitance of C317 to slow up the AGC decay rate. I

Fig 4—New T/R circuit



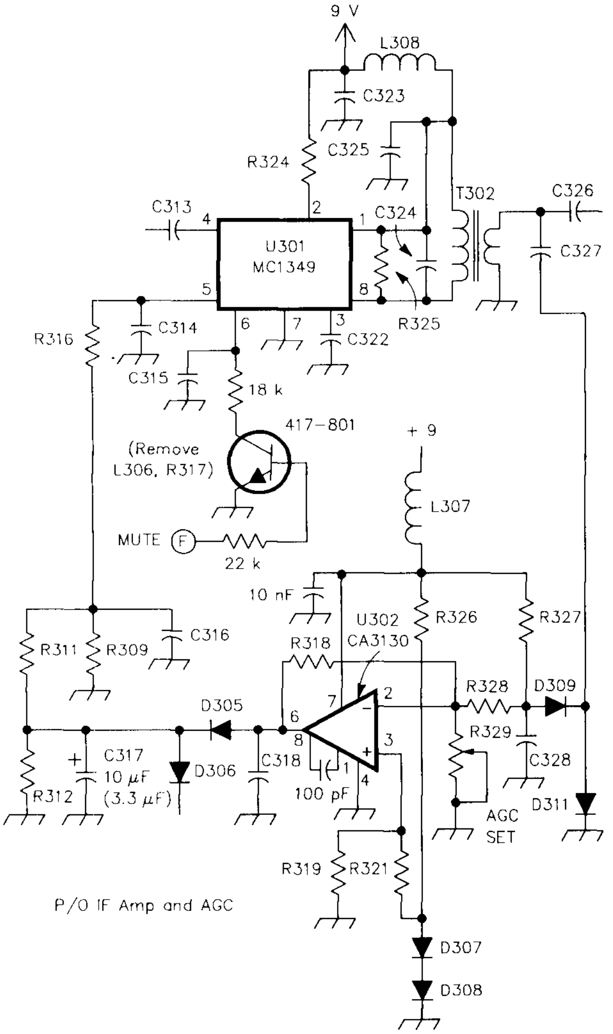


Fig 5—P/O IF amp and AGC

changed C317 from 3.3 µF to 10 µF which slowed the decay nicely. It was also suggested that the resistor R312 (47 kΩ) be changed; *don't do it!* The values chosen by Heath are critical for the proper bias current for U301 at maximum IF gain (minimum AGC). However, after I changed C317, when keying the rig the S-meter would "pin" at the top end even with low output power. But if I hold the key down, the meter reading would fall to the right level. I traced this to the timing of the voltages on the R12 and T12 (+12 on transmit) control busses. The voltages on the busses were not dropping sufficiently when switching from transmit to receive and vice-versa before the other bus went to full voltage. The overlap would leave the receiver IF on for a millisecond or so during transmit and remained transmitting when the receiver was turned back on. Changing C576 to a 0.01-µF bypass capacitor (it was 3.3 µF) solved the receiver turn-off delay problem. Changing R436 from 1 kΩ to 510 ohms solved the transmit turn-off delay problem. Now the meter "reads" correctly even at 35 WPM.

As mentioned above, I could not rely on the output meter to indicate the relative output power of the HW-9. The main difficulty with the output meter was a sudden jump of the meter to near full scale as the output power was increased. Correspondingly, the meter indication would not drop until the output power was nearly zero. The actual output power, as indicated on a watt meter, did not match the output meter's indication. This phenomena occurred mainly on the 80- and 40-meter bands. The problem was traced to RF energy being amplified by the IF amplifier U301 (Fig 5) and upsetting the AGC. I found that the gain of U301 was not being sufficiently reduced during transmit. Decreasing the resistance of R317 (originally 1.5 megohms) reduced the gain of U301 sufficiently to eliminate the problem but the AGC characteristics would suffer because the AGC bias current from pin 5 of U301 would decrease, decreasing the voltage across C317. The meter would then read well below zero when transmitting with low output power. However, by shorting pin 6 through a resistor to ground during receiver mute (further discussed below), the bias current from pin 5 of U301 would not change significantly and the gain of U301 was reduced sufficiently to eliminate the problem. I used the transistor that was Q303 (Heath part number 417-801, discussed below in connection with the audio thump suppression modification), removed R317, and replaced L306 with a 22-kΩ resistor. The base of the added transistor goes into the hole where R317 was to connect to the 22-kΩ resistor. An 18-kΩ resistor is soldered into the remaining hole for R317, the unsoldered end connecting to the collector of the added transistor, and the emitter of the transistor soldered to the ground lead of C315. Now the output meter correctly indicates the relative output power of the HW-9 on all bands even though the meter reads slightly less than zero during transmit with no output power. This modification should be made in conjunction with the audio thump suppression modification discussed below.

A small change that will improve the overload capability of the HW-9 receiver is to add an idler to the second mixer U401 during receive. As shown in Fig 6, I added a 51-ohm resistor in series with a tank circuit, which resonates at approximately the IF frequency (8.83 MHz), across the primary of T301. The resistor/tank circuit combination has low Q so that the resonate frequency need not be at exactly the IF frequency. The idler, during receive, provides a matched load to the IF port of the mixer U401 at all frequencies except at near the IF frequency. At the IF frequency, the tank circuit resonates and the 51-ohm resistor is decoupled. Note that the idler does not couple to the mixer U401 during transmit; steering diodes D301, D302 (not shown) decouple T301, C301, C302, etc, from the mixer during transmit.

A major change I made dealt with the IF filtering. As pointed out in Ref 4, the IF filter bandwidth is too wide for dense signal environments—such as in a contest or on Field Day. In addition, the crystal filter supplied (FL 301) is mismatched so that severe passband ripple

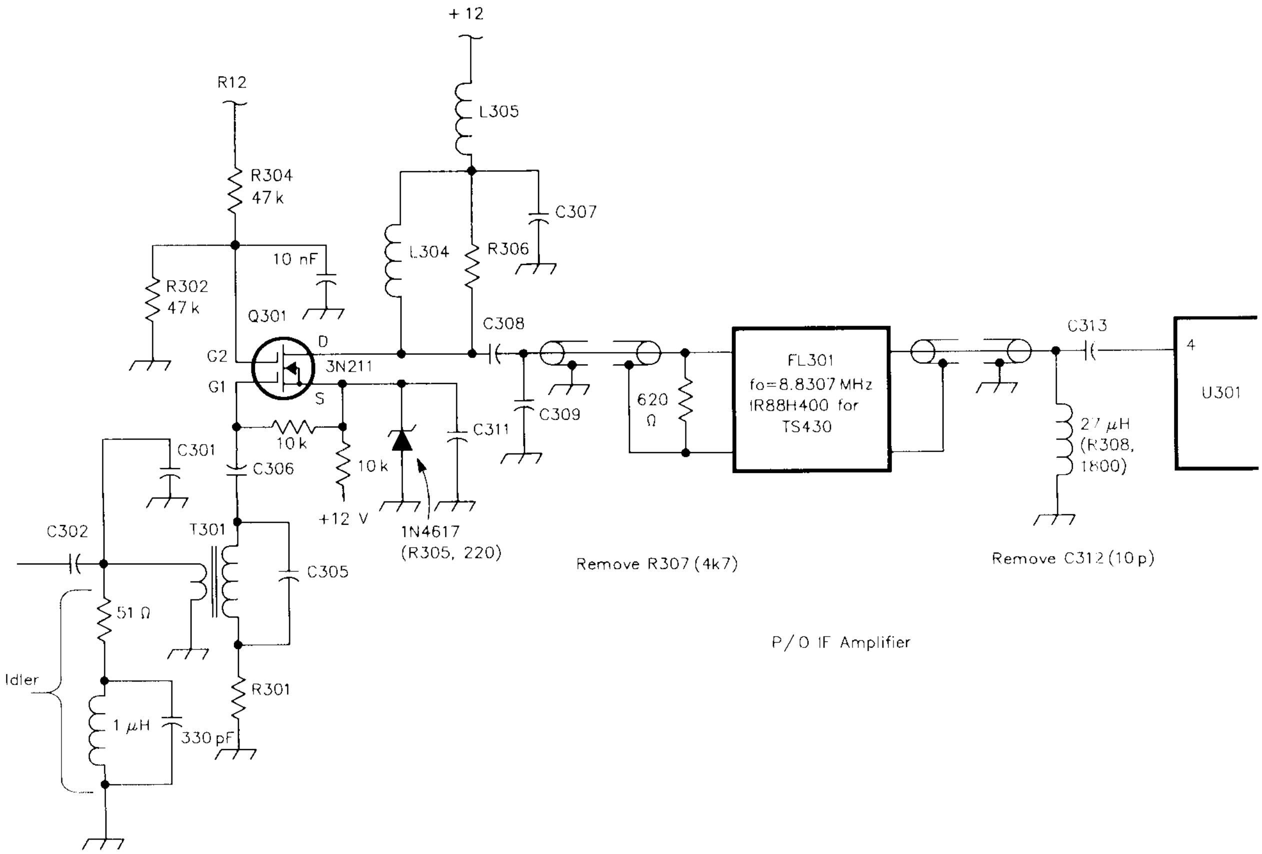


Fig 6—P/O IF amplifier

occurs. It was very annoying to have the S-meter vary more than 3 S units when tuning through a steady carrier in the pass band—the signal peaked when the audio note was over 2 kHz! Instead of trying to properly match the existing filter, I decided to replace it with a good (8-pole) crystal CW filter. Fortunately, the Kenwood TS-430 IF frequency is the same as the HW-9's—8.8307 MHz. I bought an International Radio IR88H400 crystal filter, an 8-pole, 400-Hz wide filter. The new filter is designed to be matched with a 600-ohm resistive load with a 10-pF shunt on both the input and output thereof. The new filter arrangement is shown in Fig 6. I removed the old filter and, because the new filter is much larger than the old one, attached a PC board on standoffs to the HW-9 T/R circuit board and mounted the filter on the new PC board. Short pieces of RG-174 couple the new filter to the T/R board. To properly match the output of the new filter, R308 was removed and a miniature inductor (27 μH) was put in its place. The inductor, combined with the input admittance of U301, provided the equivalent of 600 ohms/10 pF at 8.83 MHz. The input of the filter is matched with the 620-ohm resistor and C309 (10 pF). However, the insertion loss of the new filter is much greater than the old one; the first IF amplifier had to have its gain increased. Q301 was changed to a high transconductance dual gate MOSFET (3N211) and R307 removed. (Note that R306 could be changed to 600 ohms and the resistor across the input of the filter removed. However, I feel that it is easier to do the matching as shown.) L305 (missing from Heath's schematic) is routed to the +12 line instead of R12. To turn off the first IF amplifier when transmitting, the second gate of Q304 is controlled by R12. To assure that Q304 is completely cut off during transmit (R12 being approximately zero volts), the source of Q304 has a fixed voltage thereon suppled and stabilized by a 10-kΩ bias resistor and a 2.7-volt Zener diode, respectively. A visible LED and a silicon diode in series may be used instead of the Zener diode with a lower resistance bias resistor. R301 is no longer necessary but does not require removal.

As a consequence of using the narrow IF filter, the BFO transmit frequency must now be set correctly or you may not be transmitting on the same frequency you are receiving. Don't attempt to adjust the transmit return (R131) to compensate; you are compensating the VFO and the "compensation" will be good only for the one VFO frequency you adjust it for. Instead, adding a small trimmer capacitor to the BFO circuit will allow for proper

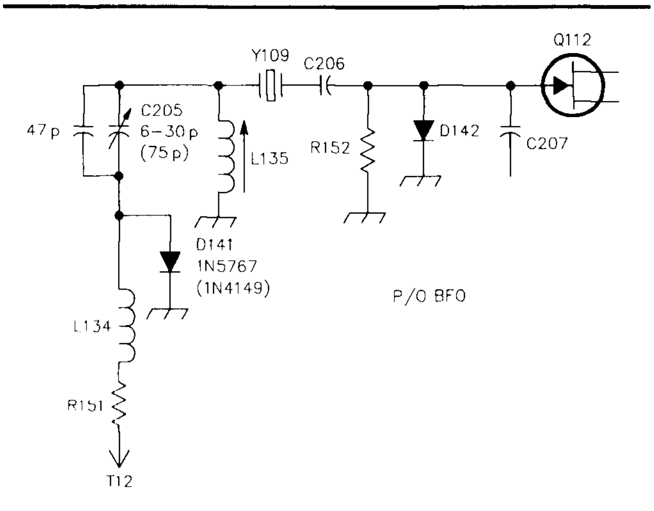


Fig 7—P/O BFO

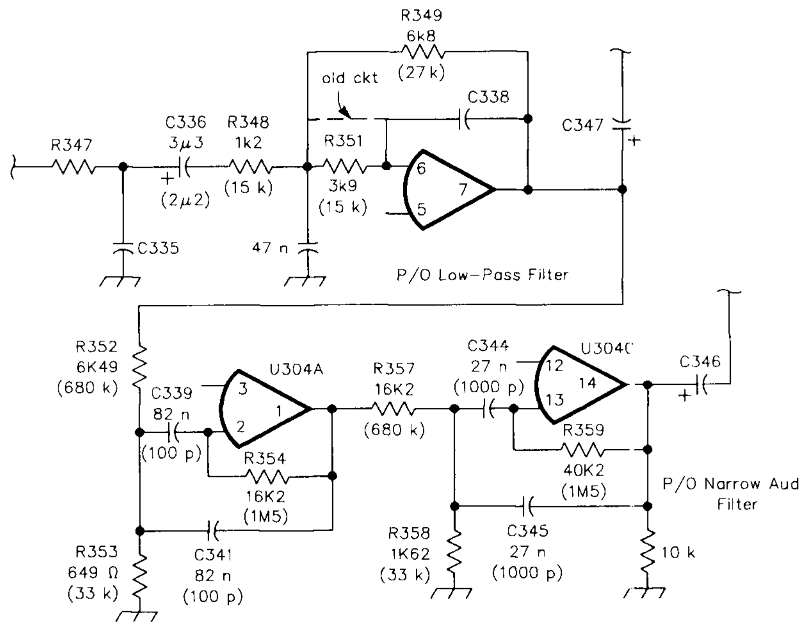
compensation, as shown in Fig 7. I removed C205 and placed a 6-30 pF trimmer in its place and bridged a 47-pF capacitor across the trimmer. I changed D141 from a IN4149 to a IN5767 (a pin diode) for better, more stable, control of the BFO frequency. Next, I drilled another hole in the BFO shield over the trimmer and reassembled the radio. This change means that the alignment of the BFO is now slightly different from the way Heath suggests. BFO alignment is now as follows: Using a frequency counter, first set the BFO frequency to 8.8314 MHz by adjusting L135. Next, key the HW-9 and adjust the new trimmer (C205) for a frequency of 8.8307 MHz. This gives you the correct 700-Hz shift from receive to transmit.

Another annoyance was the sidetone pitch being lower than the narrow filter center frequency; often I was off frequency and not "zero beat" since I was tuning the receive pitch to the sidetone pitch. Bridging a small resistor (I used a 270 kΩ) across R366 moved the pitch to near 700 Hz.

Ref 2 suggests changing the components in the narrow audio filter. The new narrow filter design works well and implementing the change is strongly recommended. The design is shown in Fig 3 using 1% resistors and metalized film capacitors for more consistent performance. However, I noticed some "fuzziness" when using the new narrow filter. The data sheets on the device (an LM 324) mentions that when the output of the op amp is AC coupled (here, the output of U304C), crossover distortion may result due to insufficient bias current in the output stage thereof. This was causing the fuzziness I was hearing. Adding a 10-kΩ or so resistor from the output of U304C (pin 14) to ground eliminated the distortion.

Another possible "mistake" is in the design of the low-pass filter of U304B. It is not clear to me what filter type it is; the design suggests a second-order multiple-feedback low-pass filter arrangement. I took the basic circuit from Heath and changed the component values and the circuit topology slightly to provide the second-order characteristic. The new circuit is shown in Fig 8. The resistance values of R348, R349, and R351 have been changed. The capacitance of C338 was not changed but instead of coupling to the junction of R348,

Fig 8—P/O narrow audio filter



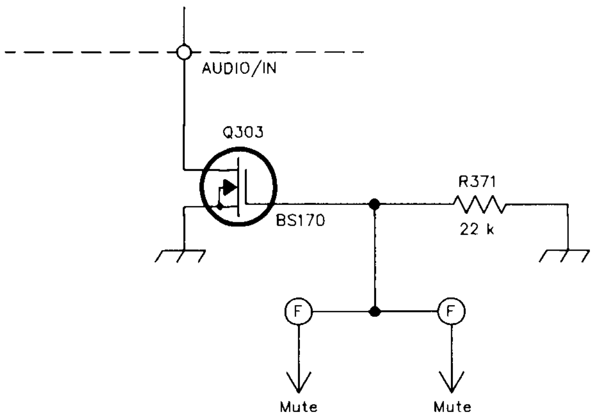


Fig 9—New audio mute circuit

R349, and R351, it couples to the inverting input (pin 6) of U304B. A 0.047-μF capacitor is added from the junction of R348, R349, and R351 to ground to achieve the second-order characteristic. The resulting low-pass filter has a cut off frequency of about 1.4 kHz. Note that capacitor C336 should be reversed per Ref 3 and the capacitance thereof increased to 3.3 μF (or more) for better low-frequency response.

In Ref 2 an elaborate design is presented to remove an annoying audio thump when the HW-9 goes from transmit to receive. A much simpler circuit that completely solves this problem is shown in Fig 9. Q303, originally a bipolar transistor, is replaced with an N-channel MOSFET, such as a BS-170 (Radio Shack 276-2074), with the gate thereof going directly to the mute line and R371, formerly going to the base of Q303, connecting to ground. The thump in the original Heath design results from collector-base junction of Q303 becoming forward biased during transmit. The FET does not have this problem and R371 now acts as a pull-down for the mute line. The thump is now gone. The old Q303 is reused in the IF amplifier disable modification discussed earlier.

Some operators may think that the receive bandwidth using the new, narrower, IF filter is too narrow for just browsing the bands. I feel, however, that the narrower bandwidth is required since the HW-9's narrow audio filter is not sharp enough for our crowded bands and the narrower IF filter avoids AGC pumping from adjacent strong signals. Still, it would be an interesting project to correct the matching to the original Heath crystal filter so that the pass-band ripple is reduced and still have the wider bandwidth when desired.

No reverse power voltage protection is supplied. I inserted an in-line fuse (2 amps) in the red wire coming from the power connector S1 to switch S2. A diode (such as a 1N4001) was bridged from the S2—fuse junction to ground so that the diode would be forward biased upon applying reversed power supply voltage to the HW-9, thus blowing the fuse. If you frequently get the voltages reversed, it may be more practical to mount the fuse in a holder on the rear panel or put it in line with the power cable so that the fuse can be replaced more easily.

I also suggest that you build in a keyer like that mentioned in Ref 2. It makes it much more convenient on backpacking trips if you don't want to bring along your keyer.

Be forewarned that some of these modifications are difficult and tricky; proceed at your own risk. It took over a year and a half to design and complete the modifications described above. It was both fun and frustrating to figure out the problem and solve it—usually the change interacted with other sections of the rig, causing more headaches and changes. I think my HW-9 is now more enjoyable to use both for contest operating and rag-chewing than the unmodified version. Some of the modifications can be made at any time. A few of the changes, such as the power amplifier changes and the AGC bypass capacitor and compensation capacitor additions, should be made at the earliest convenience.

References

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[2] "Improving the HW-9 Transceiver" by Hutchinson and Lau, *QST*, April 1988, pp 26-29.
[3] QRP Column by Bryce, *73 Magazine*, February 1989, p 86.
[4] Hints and Kinks column by Newkirk, *QST*, June 1990, pp 40-41.

The Safari-4: A High-Integration, 4-Band QRP Transceiver—*Part 1 of 3*

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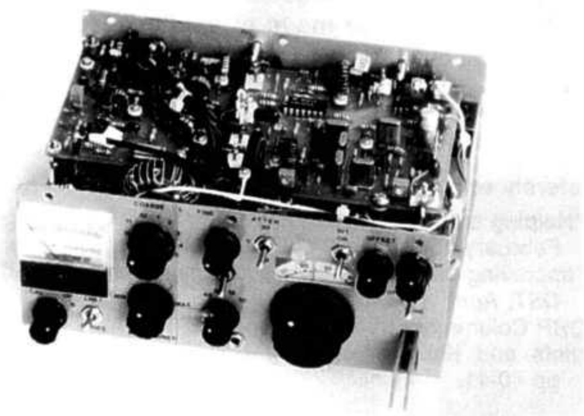
Everything but the antenna is packed into this sophisticated 1-watt CW transceiver—the perfect rig for the outdoor QRP fan. The dual-conversion design uses transverters to bring the art of multiband home-brew down to earth.



I had two goals in mind when I designed the Safari-4. First, I wanted a compact station for backpacking and camping trips—one that didn't require a full set of luggage to handle all of the usual accessories. Second, I wanted to design a multiband, dual-conversion CW transceiver that could be built initially for one band, but could be easily extended to other bands when I was ready. The Safari-4 satisfies these goals.

Anyone who has set up a portable QRP station will appreciate what's inside this rig: built-ins include a 12-volt battery and trickle charger, an antenna tuner, an absorptive SWR bridge and an electronic keyer with touch-sensitive paddles. This level of integration reduces setup to tossing a wire into a nearby tree and plugging in the headphones! And here's the best news: The entire rig can, with some patience, be shoehorned into a 3 × 7 × 5-inch (HWD) box—less than half the volume of a Heath HW-8.

If you've attempted multiband transceiver construction, or have been discouraged by it, you will be encouraged by the simple bandswitching scheme used here. Individual transverters are used for 40, 20, 15 and 10 meters; only +12 volts and low-impedance RF signals are switched. The transceiver can be built for any of the four bands, and expanded later by simply adding transverters.

Besides simple bandswitching, an advantage of using transverters is that the failure of the PA transistor (or other component) on one transverter won't affect operation on the other bands. This is an important consideration in field operation where on-site repair may be impossible.

Diode switching is one possible alternative to using individual transverters. But I rejected this method because the diode bias would have roughly tripled the transceiver's receive-mode current drain, which is just over 30 mA. This low current requirement significantly enhances battery life.

The transceiver has the following additional features:

General

- covers the first 70 kHz of 40, 20, 15 and 10 meters
- panel meter shows SWR, S-units, or battery voltage

Transmitter

- 1-watt output on 40 and 20 m, 0.5 watts on 15 and 10
- high-stability, low-frequency VFO (3.9 MHz)
- full break-in or selectable T-R delay
- low-noise, no-IF-amplifier design
- no audible birdies over entire frequency range
- Cohn 4-pole crystal filter (500 Hz @ −6 dB) using inexpensive 4.915-MHz crystals
- simple audio-derived AGC applied directly to NE602
- receiver incremental tuning (RIT)

Overall System Design

The Safari-4 is divided into several subsystems as shown in the block diagram, Fig 2. The antenna tuner is shown here as an L-network, one of two switchable configurations (L-net or series). The SWR bridge (Z1) is

[1]Notes appear on page 16.

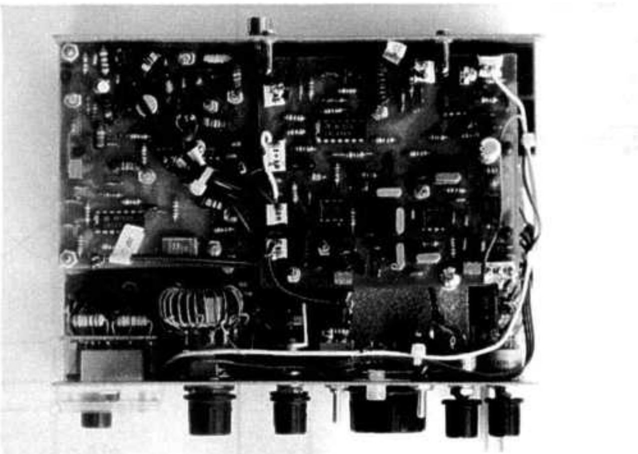


Fig 1—Top view of the Safari-4. The antenna tuner and metering circuits are just behind the front panel on the left-hand side of the box. Behind the meter, the stack of four transverter boards can be seen, with the 10-meter transverter on top (see Fig 12). To the right of the meter is the tapped transmatch inductor, attached directly to the coarse inductance switch. The variable inductor is in the middle front (see Fig 10). At the middle right, the 8.8-MHz IF board sits on top of the VFO (see Fig 13) and 12 V battery. On the extreme right are the keyer/control board, mounted vertically, and the resistive-touch paddle (see Fig 11).

of the absorptive type to protect the PA during tune-up. One of four band-switched transverters (Z6-Z9) follows the SWR bridge; here the signal splits into receive and transmit paths to the IF/AF module (Z5). At the bottom of the block diagram are the keyer/control module (Z4), and the separately-enclosed VFO/RIT module (Z2/Z3).

These modules are discussed in more detail below, and can also be seen in the top view, Fig 1. The interconnection of all of the modules is shown in Fig 3, the chassis wiring diagram.

The dual-conversion mixing scheme used has three important consequences: (1) low-cost crystals are employed in the receiver crystal filter; (2) the VFO is low in frequency, resulting in good stability; (3) there are no audible birdies or images on any band. I explored thousands of possible VFO, BFO, and conversion oscillator combinations using a computer program that I wrote specifically for this project.[1] (See sidebar, "Why Dual Conversion?")

In the configuration I chose, the transverter conversion oscillators operate above the RF frequency on all bands. For example, on 20 meters, the oscillator is at 22.870 MHz. In receive mode, subtracting the 14.0 to 14.07 MHz RF input gives a first IF of 8.870 to 8.800. In transmit mode, the 8.8-MHz IF signal is subtracted from the 22.870-MHz oscillator to yield the same 14-MHz RF output range.

The 8.8-MHz 1st IF signal is mixed with the VFO, at 3.885 to 3.955 MHz, to yield 4.915 MHz. The latter is a common microprocessor crystal frequency, resulting

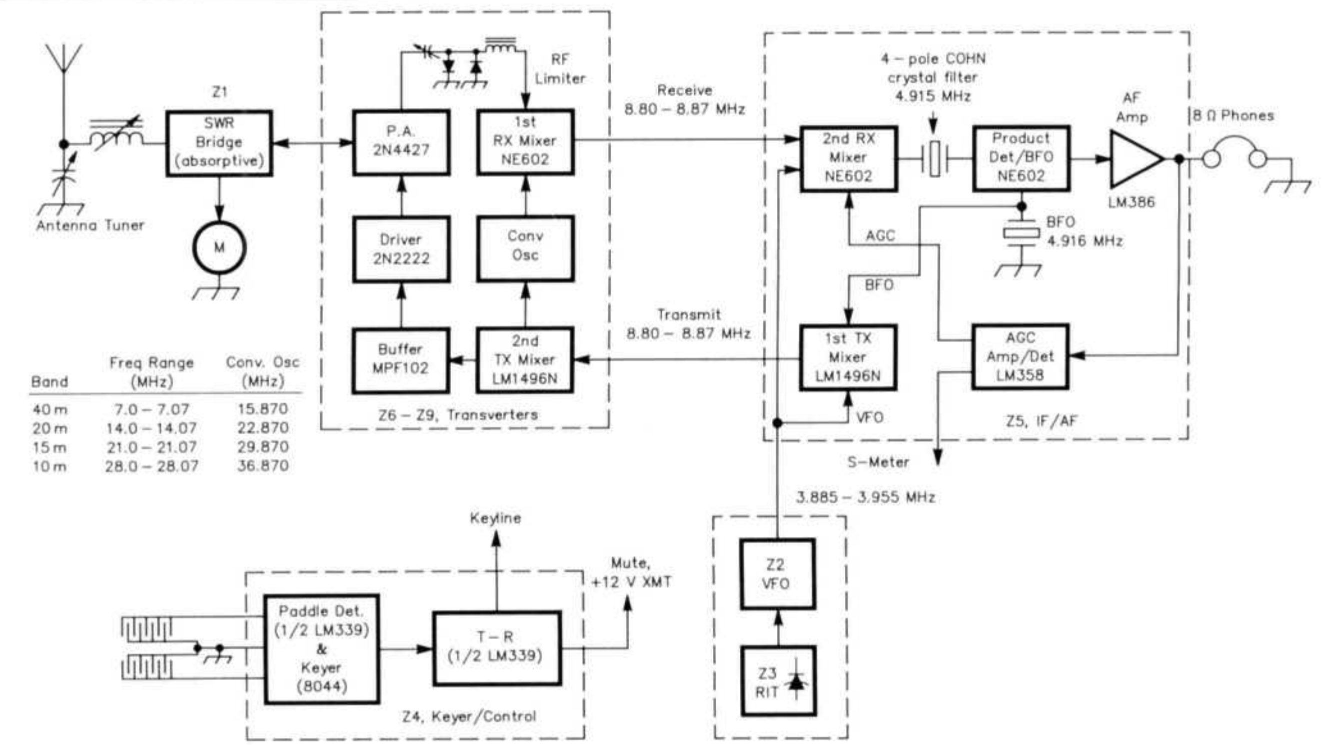
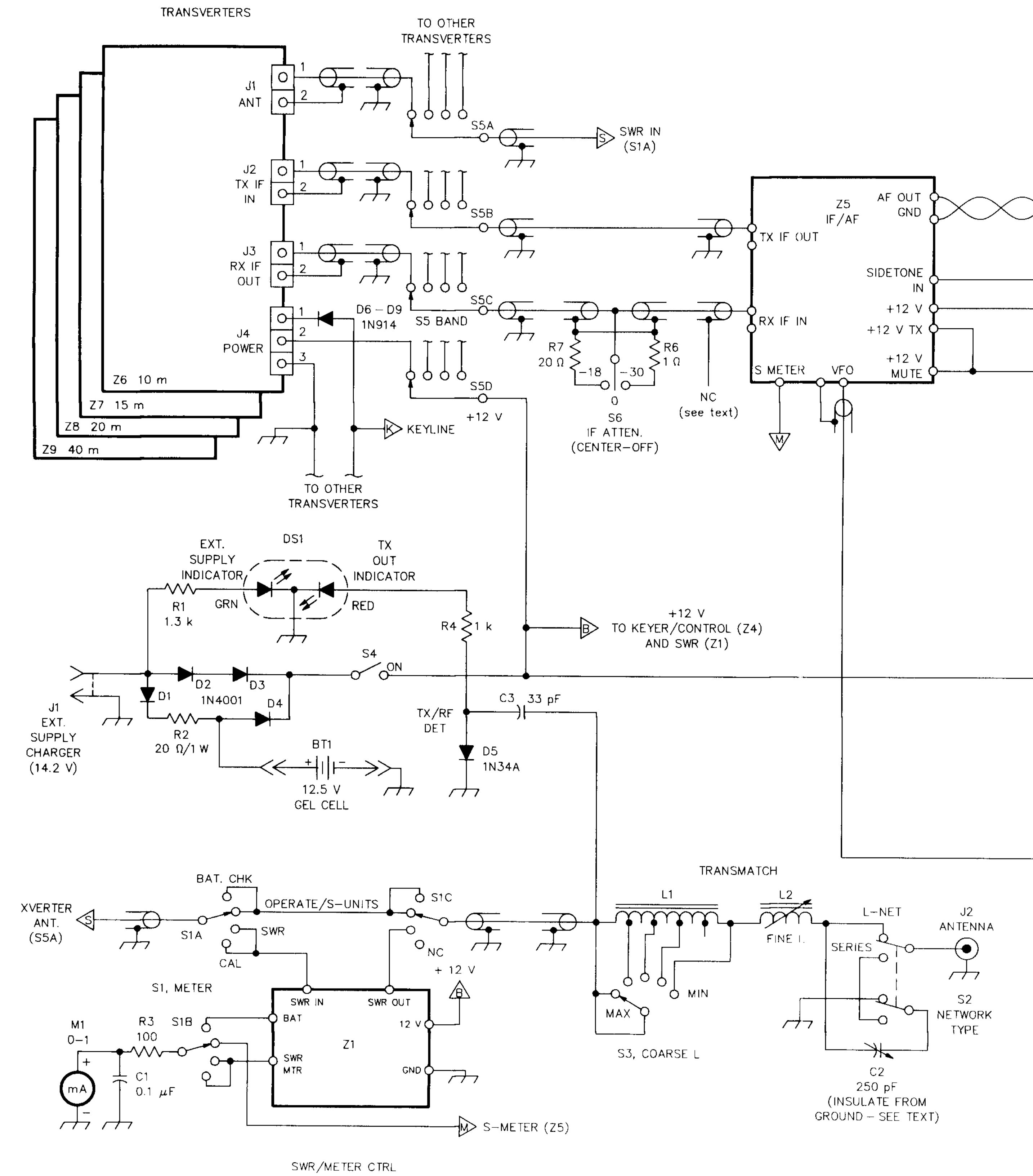


Fig 2—Safari-4 transceiver block diagram. One transverter is shown, at left, along with a table showing the conversion oscillator frequencies. (See text.)



TRANSVERTERS
TO OTHER TRANSVERTERS
J1 ANT
J2 TX IF IN
J3 RX IF OUT
J4 POWER
Z6 10 m
Z7 15 m
Z8 20 m
Z9 40 m
S5A
SWR IN (S1A)
S5B
S5C
S5D
S5 BAND
D6 – D9 1N914
KEYLINE
TO OTHER TRANSVERTERS
+12 V
Z5 IF/AF
AF OUT
GND
TX IF OUT
SIDETONE IN
+12 V
+12 V TX
RX IF IN
+12 V MUTE
S METER
VFO
R7 20 Ω
R6 1 Ω
−18
−30
0
S6 IF ATTEN. (CENTER−OFF)
NC (see text)
EXT. SUPPLY INDICATOR
DS1
TX OUT INDICATOR
GRN
RED
R1 1.3 k
R4 1 k
+12 V TO KEYER/CONTROL (Z4) AND SWR (Z1)
S4
ON
D1
D2
D3
1N4001
D4
J1 EXT. SUPPLY CHARGER (14.2 V)
R2 20 Ω/1 W
BT1
12.5 V GEL CELL
C3 33 pF
TX/RF DET
D5 1N34A
TRANSMATCH
L1
L2
FINE L
L-NET
J2 ANTENNA
SERIES
S2 NETWORK TYPE
MIN
MAX
S3, COARSE L
C2 250 pF (INSULATE FROM GROUND – SEE TEXT)
BAT. CHK
OPERATE/S-UNITS
S1C
XVERTER ANT. (S5A)
S1A
SWR
CAL
NC
+ 12 V
S1, METER
SWR IN
SWR OUT
BAT
12 V
Z1
SWR MTR
GND
M1 0-1
R3 100
S1B
C1 0.1 μF
mA
S-METER (Z5)
SWR/METER CTRL

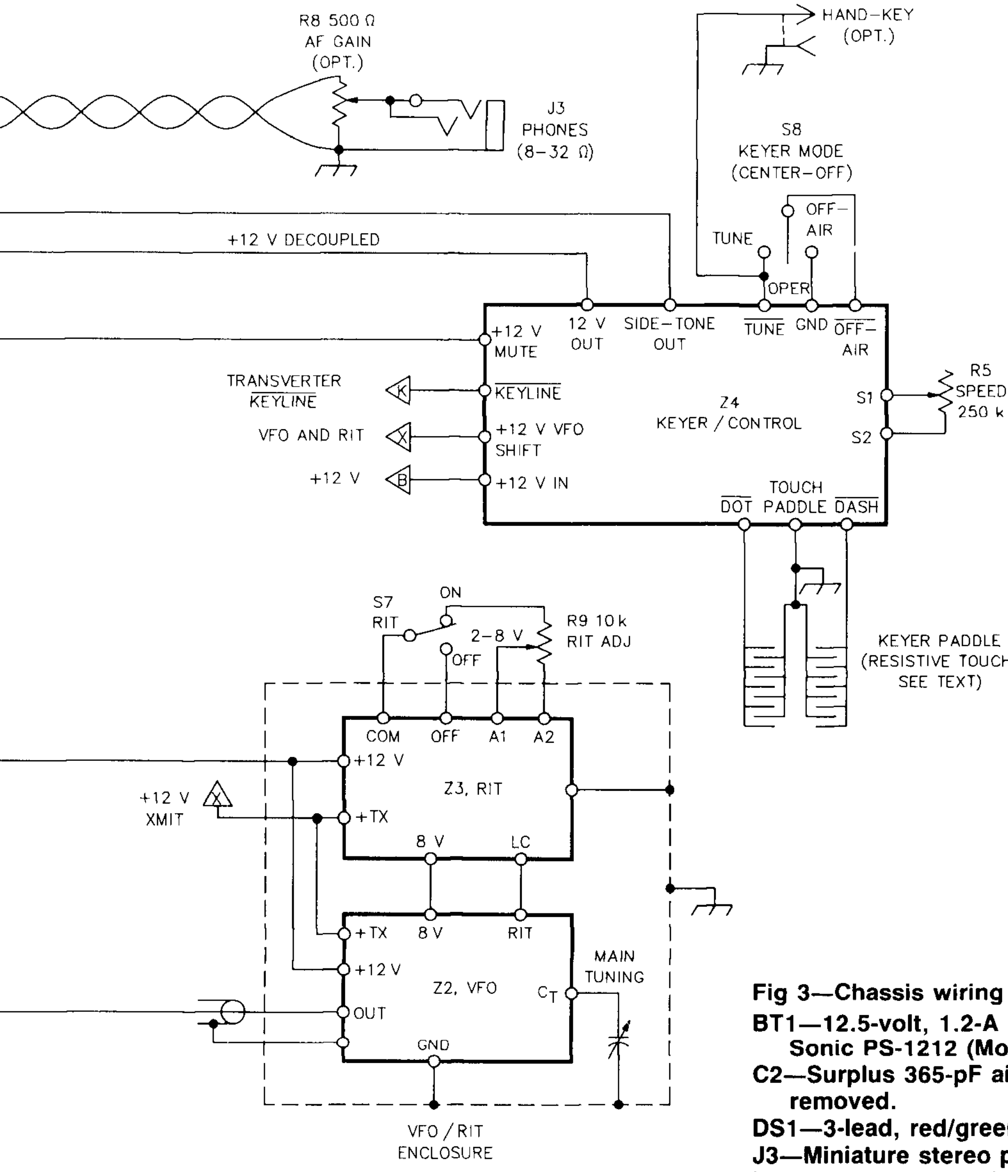


Fig 3—Chassis wiring diagram.

BT1—12.5-volt, 1.2-A gelled-electrolyte battery, Power Sonic PS-1212 (Mouser 547-PS-1212).
C2—Surplus 365-pF air variable with ⅓ of the plates removed.
DS1—3-lead, red/green LED, CAL-PAK CP26 (Gateway).
J3—Miniature stereo phone jack.
L1—26 turns no. 22 bus wire on Amidon T-106-2 core. Tap at 2, 4, 6, 8, 10, 12, 14, 17, 20, and 23 turns.
L2—Homemade variable inductor (see Fig 10).
S1—Subminiature 3 pole, 4 position, nonshorting rotary switch (surplus).
S3—1 pole, 12 position shorting rotary switch (Mouser 10WA164)
S5—Subminiature 4 pole, 4 position, nonshorting rotary switch (surplus).
S6, S8—Miniature toggle switch; SPDT, center-off.
Z1—See Fig 9.
Z2—See Fig 6.
Z3—See Fig 7.
Z4—See Fig 8.
Z5—See Fig 4.
Z6-Z9—10, 15, 20, and 40-meter transverters, respectively (see Fig 5).

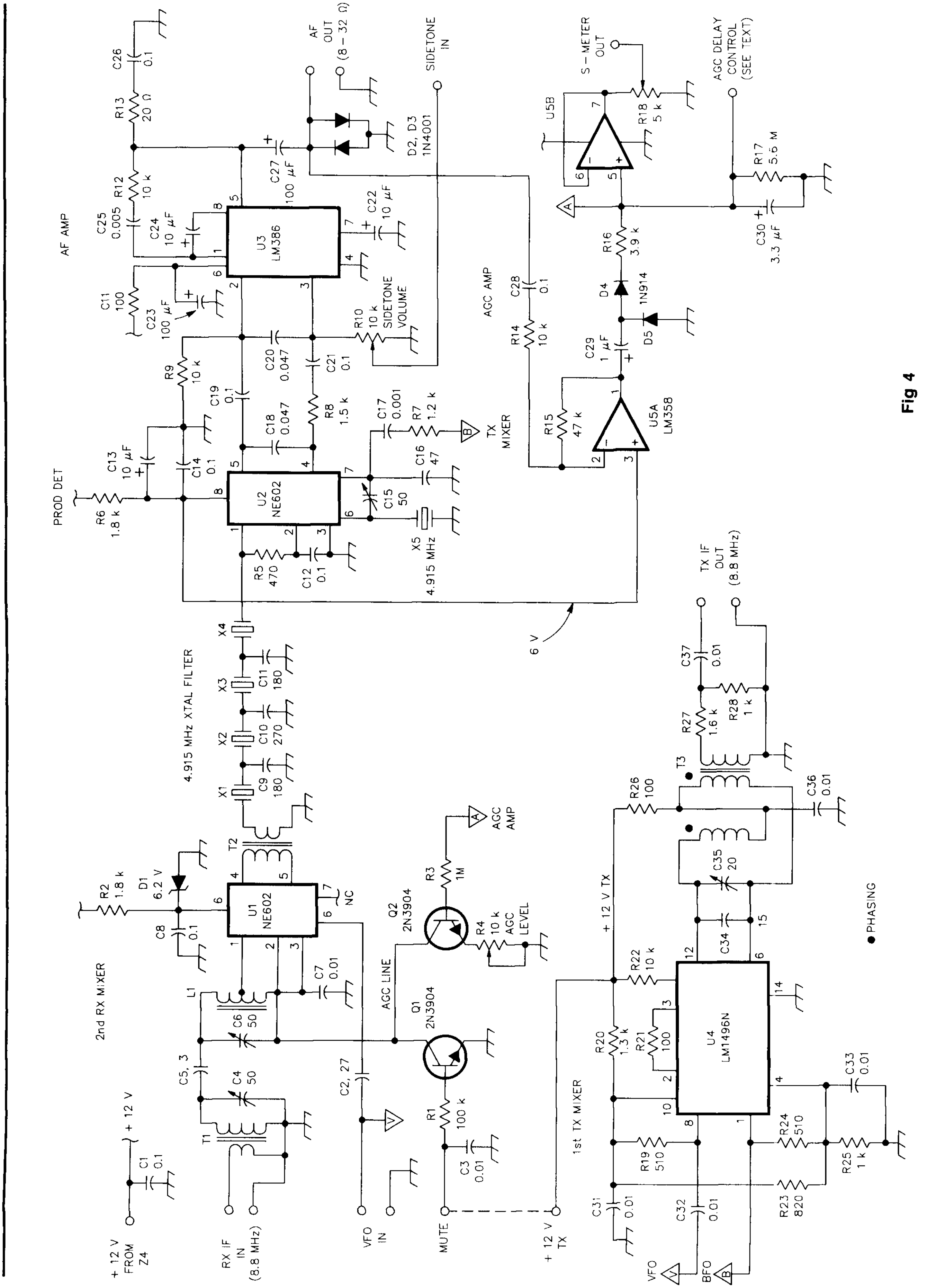

AF AMP
PROD DET
2nd RX MIXER
4.915 MHz XTAL FILTER
AGC AMP
1st TX MIXER
U1 NE602
U2 NE602
U3 LM386
U4 LM1496N
U5A LM358
U5B
Q1 2N3904
Q2 2N3904
D1 6.2 V
D2, D3 1N4001
D4 1N914
D5
X5 4.915 MHz
+ 12 V FROM Z4
RX IF IN (8.8 MHz)
VFO IN
MUTE
+ 12 V TX
AGC LINE
AGC LEVEL
AGC AMP
TX MIXER
SIDETONE IN
SIDETONE VOLUME
AF OUT (8 – 32 Ω)
S – METER OUT
AGC DELAY CONTROL (SEE TEXT)
TX IF OUT (8.8 MHz)
6 V
VFO
BFO
● PHASING
NC
Fig 4

Fig 4—(Opposite page) Schematic of the 8.8-MHz IF/AF board (Z5). The trimmer capacitors must have 0.2-inch lead spacing to be used on the author's PC boards.

C4, C6, C15—Ceramic trimmer, 9-50 pF (Mouser 24AA024).
C35—Ceramic trimmer, 3.5-20 pF (Mouser 24AA022).
L1—40t no. 26 enamel on Amidon T-50-2, center-tapped (8.6 μH; Qμ = 160).
T1—Secondary: 40t no. 26 on Amidon T-50-2; primary: 4t (8.6 μH; Qμ = 160).
T2—Secondary: 27t no. 26 on Amidon FT-37-43; primary: 9t.
T3—18 turns no. 28, trifilar, on Amidon T-50-2. Observe phase.
U1, U2—Signetics NE602N mixer/oscillator IC (Radiokit; also see Note 7).
U3—National LM386N AF amp IC (Digi-Key).
U4—National LM1496N balanced mixer IC. Author's PC board is laid out for plastic 14-pin DIP.
U5—National LM358N low-power dual op-amp IC.
X1-X5—4.91520 MHz, HC-18 case (Digi-Key X050); buy 10 and hand-pick (see text).

in a low-cost 2nd IF crystal filter with a good bandwidth for CW operation.

Receiver Design

The receiver is based on the Signetics NE602, which is a low-power, double-balanced mixer and oscillator in an 8-pin DIP (dual-inline package). The NE602 first appeared in *QST* in the "Neophyte";[2] Dillon's article is a good primer for those who haven't used the chip before. As a low-power receive mixer the chip excels, consuming only 2.5 mA and requiring few external components. (Unfortunately, it is not suitable for use as a transmit mixer.)

The NE602's on-chip oscillator is a nice feature, but can be used only in cases where low output amplitude and poor input-to-oscillator isolation can be tolerated. In particular, the on-chip oscillator should not be used as a transmitter VFO. The oscillator also fares poorly when used in some transceiver applications, since its frequency can be pulled by large RF signals appearing at the input of the mixer stage. In the Safari-4, an NE602 is used as the first receive mixer on each transverter board, and two more are used on the IF board—one for the second receive mixer and one for the product detector. Only the BFO uses the on-chip oscillator.

In something of a departure from typical superhet receiver design, I chose not to use an IF amplifier. This works for a couple of reasons.

First, the NE602 provides around 20 dB of gain per stage. With three '602s in the signal path, plus an LM386 AF amp with 40 dB of gain, total receiver gain is over 100 dB. The resulting design is quieter than other designs that I tried using the MC1350, CA3028, and other IF amplifiers. Second, I found a way to couple an AGC voltage directly to the second mixer, the details of which are discussed later. While this method suffers from a lack of dynamic range, it has the advantages of minimum parts-count and low current drain and is a welcome improvement over no AGC at all.

Transmitter Design

Both transmitter mixer stages are based on the LM1496N. This double-balanced-mixer IC is inexpensive, easy to use, and provides good conversion gain. A three-stage amplifier follows the second transmit mixer, and has an output of between 0.5 and 1.0 watt.

RF from the PA stage is routed to the receiver mixer input by a series tuned circuit with diode limiting, a technique described by DeMaw.[3] This, in conjunction with an effective receiver muting circuit, provides smooth full break-in operation (QSK).

Circuit Details

The following paragraphs provide circuit details for each of the subsystems.

8.8-MHz IF/AF Module (Z5)

The 8.8-MHz IF/AF module (Fig 4) is the heart of the transceiver, comprising the second receive mixer, product detector and BFO, AF amp, AGC, and first transmit mixer. The 8.8-MHz IF input and output are tied to one of the transverters through the bandswitch. On receive, the 8.8-MHz IF is mixed with the VFO (3.885 to 3.955 MHz) to provide a low second IF of 4.915 MHz. On transmit, the first transmit mixer adds the BFO (4.915 MHz) and VFO to provide a carrier output at 8.800 to 8.870 MHz.

The input to the second receive mixer, U1, is routed through a narrow band-pass filter (T1, C4, C5, C6 and L1). L1 is center-tapped to keep the Q of C6/L1 high.

Q1 and Q2 provide muting and AGC, respectively, by lowering U1's internal dc bias voltage and hence its gain. When the mute voltage is applied, Q1 saturates, cutting off U1 completely. On the other hand, the AGC circuit forward biases Q2 to vary U1's bias voltage from about 1.2 volts (maximum gain) to just over 0.6 volts (minimum gain). Zener diode D1 is required to regulate U1's supply voltage when the bias is lowered.

The output from the second receive mixer drives a Cohn 4-pole crystal filter; C9, C10, and C11 were chosen empirically. Hayward and DeMaw cover such filters in detail.[4] I achieved a filter bandwidth of about 500 Hz at −6 dB (1 kHz at −20 dB) without too much difficulty, keeping the crystals well matched. U2 forms the product detector and BFO. X5 resonates in its parallel mode, around 700 Hz higher than the series-resonant frequency of the filter, so the same type of crystal can be used. C17 and R7 isolate the BFO signal for use in driving the first transmit mixer, U4. Without them, the +12 TX transient will generate an audible pulse at U2's AF output.

U2's output filter (C18-C21 and R8-R10) was borrowed from the Neophyte, and improves the AF bandwidth looking into U3. In a slight variation from that design, R10 doubles as the sidetone level control.

The AF output stage is conventional except for the

Why Dual Conversion?

A simple receiver can be built using direct conversion, a technique in which an oscillator is mixed with the RF signal to directly produce an audio signal. For example, an RF signal at 7.100 MHz could be mixed with a 7.101-MHz oscillator to produce a 1-kHz audio output. (The mixer, in this case, is called a product detector.)

The next step up in receiver design is single conversion. Here, the RF input signal is mixed with a lower VFO signal to produce an intermediate frequency (IF). A BFO is then mixed with this IF, using a product detector, to produce the audio signal.

To illustrate, say you want to build a 21.0 to 21.2 MHz receiver. With single conversion you can choose a VFO frequency much lower than 21 MHz—say 5.0 to 5.2 MHz—to improve stability. A intermediate frequency of 16 MHz will result, since 21.0 − 5.0 = 16.0 MHz (and at the high end, 21.2 − 5.2 = 16.0 MHz). A BFO signal of 15.999 MHz can then be fed to the product detector. This will result in a 1-kHz audio output when the input RF signal is at 21.000 and the VFO is at 5.000 (since 21.000 − 5.000 − 15.999 = .001).

Notice that the RF signal, when mixed with the appropriate VFO signal, will always yield the same IF frequency of 16.000 MHz. This leads to another important benefit of the single-conversion design: We can provide a very sharp crystal filter at the IF frequency and get single-signal reception, meaning that you will only hear a station on one side of zero beat when tuning the VFO.

Let's clarify this idea. A direct-conversion receiver cannot provide single-signal reception, because there is no way to prevent signals on both sides of the VFO from producing AF outputs. For example, if the VFO in a direct-conversion receiver is tuned to 7.100 MHz, you'll hear a 1-kHz pitch from a station at 7.101 MHz or from a station at 7.099 MHz, or, in the worst case, both—twice the QRM!.

In the single-conversion receiver, though, we can eliminate one of the two responses with a sharp filter. Assume that the VFO is tuned to 5.100 MHz, and that we have a crystal filter in the IF signal path that will only pass 16.000 MHz. This means that a signal at 21.100 MHz will get through, because 21.100 − 5.100 = 16.000. A signal at 21.099 will not, however, because 21.099 − 5.100 = 15.999 MHz, which is out of the filter passband.

The single-conversion receiver design has a couple of major drawbacks. First, the IF frequency depends on the RF and VFO frequencies; as we saw above, the IF may be quite high (16 MHz). It is much harder to build a sharp crystal filter at this frequency than at 9 MHz or lower. Things get worse as the RF input frequency is increased or the VFO frequency is decreased. Second, the only way to build a multiband single conversion receiver is to alter the VFO frequency by premixing or by switching LC circuits.

A dual-conversion design, in which two IFs are used, can be used to overcome these problems. In such a receiver the RF input signal is mixed with a fixed-frequency conversion oscillator, resulting in the first IF. Next, we mix the first IF signal with the VFO to produce a second, lower IF. Finally, a BFO and product detector are used to produce the AF output as in the single-conversion design.

Consider a dual-conversion receiver that covers 14.0-14.2, 21.0-21.2, and 28.0-28.2 MHz. We'll use a first IF of 10.0 to 10.2 MHz, and a second IF of 4.4 MHz. The three RF input ranges are first mixed with conversion oscillators of 4.0, 11.0, and 18.0 MHz, respectively, depending on which band is selected. This gives us the first IF output of 10.0 to 10.2 MHz (eg, 14.0 - 4.0 = 10.0). Next, we mix the first IF with a VFO at 5.6 to 5.8 MHz, which gives us an output at 4.400 MHz—the second IF. A sharp crystal filter is fairly easy to build at this frequency. Also, bandswitching with this receiver design can be accomplished by selecting a conversion oscillator crystal and RF input filter for each band.

Dual-conversion receivers may suffer from unwanted mixer products and RF images if the conversion oscillator, VFO and BFO frequencies are not carefully chosen. A computer program was used to choose problem-free frequency combinations for the Safari-4 (see Ref 1).

addition of C25 and R12, which effectively remove high-frequency hiss. R13 and C26 provide a path to ground for RF parasitics, and should be mounted as close as possible to U3. A value of 10 μF at C22, rather than the 0.1 μF sometimes used, improves the large-signal distortion characteristics of the LM386.

The audio output is intended to drive low-impedance Walkman-style headphones. Since the response time of the AGC circuit is relatively slow, back-to-back diodes (D2 and D3) were added to limit the maximum output voltage when a large input pulse hits under quiet conditions. Inexpensive, lightweight phones are preferred, since their small transducers tend to saturate during noise impulses, further mitigating the slow AGC response time.

U5 is used as both the AGC amp and S-meter driver. An intermediate dc reference voltage for U5A is obtained from R6, which sets U2's supply at 6 volts. D4 and D5 form a full-wave AGC detector.

The AGC recovery time (decay) is a compromise. The values shown for C30 and R17 keep the gain constant during slow CW, but a smaller value resistor can be added in parallel to improve impulse recovery time. I use 2.7 MΩ.

The last major subsection of the 8.8-MHz IF board is the first transmit mixer. This circuit is identical to that used on the transverter boards, described later.

[This is the end of Part 1. Part 2 will be in the November issue.]

Notes

[1]The program, SUPERHET.PAS, runs on an IBM® PC, AT, or compatible, and is written in Turbo Pascal. The author will supply a copy of the program on 5.25-inch diskette, including source code and instructions, for $5.00 prepaid.

[2]Dillon, J., "The Neophyte Receiver," Feb 1988 *QST*, pp 14-18.

[3]DeMaw, D., "Some QRP Transmitter Design Tips," Feb 1988 *QST*, p 31.

[4]Hayward, W. and DeMaw, D., *Solid State Design for the Radio Amateur*, ARRL, Newington, CT, 1986, pp 85-87.

The Sporadic-E Season of 1989 in Review

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Sporadic E (E_S) continues to be among the most fascinating propagation modes commonly used in the amateur bands above 28 MHz. This is especially true in the 50-54 MHz and 144-148 MHz bands, where opportunities for long-distance communications are limited. In spite of more than half a century of experience with sporadic E, its appearance remains largely unpredictable and its causes are still imperfectly understood, even by professionals.[1] Long-term observations have provided general indications of when and where E_S is most likely to appear. US amateurs have come to expect sporadic E during the late morning and early evening hours of late spring and early summer, for example, but the dimensions of year-to-year variations, especially across the solar cycle, are still largely undocumented. Over the years, the unique observations of radio amateurs has contributed to filling this important gap, although a great deal of opportunity remains for furthering this work.[2]

This review of sporadic E during 1989 continues annual reports Patrick Dyer (WA5IYX) has made since the mid-1960s and so adds to a long run of amateur E_S observations. Analysis of this data also provides an additional opportunity to test the relationship between sporadic E and various measures of solar activity, an issue that is still unresolved among propagation researchers. This may be of special interest as we go through the peak of solar cycle 22. Even though a review of last year's sporadic E may be "old news" to readers more curious about the 1990 E_S season just ending, analysis of a full year's record adds to the growing understanding about the general behavior of this most unpredictable of propagation modes.

Data Collection

For almost 20 years, Dyer has meticulously recorded observations of E_S propagation in the FM broadcast band (88-108 MHz) from his homes in south Texas. The FM broadcast band is nearly ideal for this purpose, as it is filled with hundreds of stations scattered over a wide area of North America and broadcasting around the clock. It is also high enough in frequency that the observation record is not saturated with E_S events. Even so, the data indicate that sporadic E reaches 88 MHz an average of nearly 75 days per year. More significantly for radio amateurs, the FM broadcast band lies between the two most popular amateur VHF bands at 50 and 144 MHz. Conclusions drawn about E_S conditions in the 88-108 MHz band can be applied with certain precautions to the two amateur bands higher and lower in frequency.

All observations in this review refer to instances of commercial FM broadcast stations heard in San Antonio, Texas, in the 88-108 MHz band, unless otherwise stated. Dyer monitored the FM broadcast band everyday, roughly between 7:00 AM and 12:00 midnight local Central Standard Time, for the distinctive signs of E_S propagation. The early morning hours were not included in this demanding regime, but it is widely believed that few E_S events begin after midnight. Data used here does not include the distance or direction of individual stations heard. Single-hop E_S range is typically 1000 to 2200 km, and instances of double-hop propagation were suspected for stations beyond 2200 km. In addition, it was not unusual to hear signals from several different directions simultaneously, which meant that it was likely that two or more distinct E_S events were sometimes combined and recorded during a single listening period.

A Better-than-Average Year

The occurrence of sporadic E during 1989 was above the average for the previous ten-year period in terms of total minutes and number of days E_S was observed, but it was generally less than the totals for the previous two years. See Table 1. A tabulation of number of days with sporadic E by month reveals the expected peak during the May-August period and the much smaller period of E_S activity at the end of the year (included in Fig 1). The spring season got off to an above average start, and except for the nearly average June, remained at above average levels in terms of number of days per month E_S observed throughout the peak period. See Table 2 for a month-by-month accounting.

When the May-August peak period is broken down into minutes of E_S per week, as shown in Fig 2, the picture changes. The first half of the peak period now appears decidedly below average, except for the spectacular week of May 14. That week 675 minutes of 88-MHz E_S was logged, most of that occurring on May 14 and May 18. Although sporadic-E was not heard as high as 144 MHz in San Antonio during the week of May 14, other parts of the country reported 144-MHz E_S on May 14, 17, and 18. The second half of the peak period was well above average. During the extraordinary week

[1]Notes appear on page 19.

Table 1

Annual Number of Days and Total Minutes of 88-MHz Sporadic E Recorded, 1987-1989 and 1980-1989 Average

	1987	*1988*	*1989*	*1980-1989 average*
Days	89	80	85	74.2
Minutes	10175	8880	7750	7200

Table 2

Total Number of Days and Minutes per Month of Sporadic E Observed

Month	*Days*	*Minutes*
Jan	1	20
Feb	2	45
Mar	0	0
Apr	1	10
May	15	1190
Jun	19	1410
Jul	21	3720
Aug	14	915
Sep	0	0
Oct	1	40
Nov	5	130
Dec	6	190
Total	85	7750

Table 3

144-MHz Sporadic-E Events Reported During 1989

Date	*Time CST*	*Presumed Center of E_s Ionization*
May 14	2000-2140	Missouri
17	2045	Southern New Mexico
18	1930	Colorado
31	1845-2020	Missouri to Iowa
Jun 18	**0900-1000**	**Mississippi**
	1115	Arkansas
Jul 4	1500-1730	Virginia
5	1730-1745	North Carolina
6	1100	South Carolina
8	**1100-1150**	**Southern New Mexico**
	1455-1510	**Southern New Mexico**
	1735-1750	**Southern New Mexico**
9	0800-1020	Virginia to Ontario
21	2000-2110	Colorado

Note: Events in bold observed by WA5IYX. Sources: *Midwest VHF Report*, April and May/June 1989; *West Coast VHFer*, July 1989; *Northeast VHF News*, Jan-Feb 1989 (published Aug 1989); *QST*, Aug 1989, pp 71; WA5IYX and W3EP/1.

of July 2, 88-MHz sporadic-E was recorded every day for a weekly total just over 1700 minutes. July 8 was especially noteworthy. Sporadic-E MUF exceeded 88 MHz for 565 minutes and reached over 144 MHz for at least 75 minutes. Other parts of the country reported 144-MHz

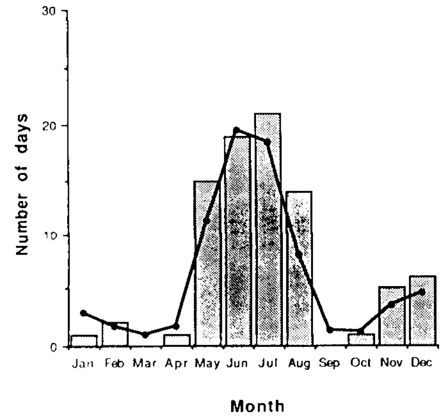

Fig 1—Number of days per month 88-MHz sporadic E recorded during 1989 (bars) compared with average number of days per month 1980-1989 (line).

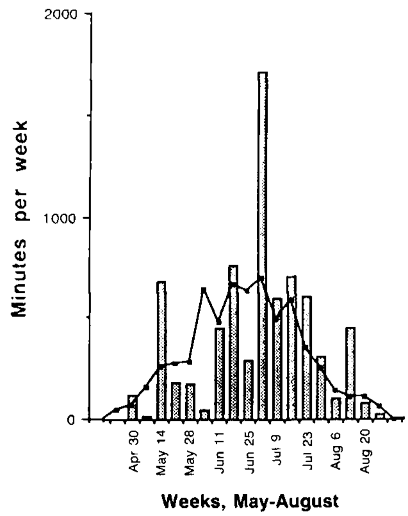

Fig 2—Minutes per week 88-MHz sporadic E recorded during May-August 1989 (bars) compared with average minutes per week 1980-1989 (line).

E_S on four additional days. Table 3 summarizes all known 144-MHz E_S events for the entire US during 1989.

Sporadic E and the Solar Indices

There is an on-going controversy over the relationship among the formation of sporadic E, the solar cycle, and the daily and hourly measures of geomagnetic field activity. Data from three previous cycles (including reports by Dyer) have suggested that sporadic-E activity declines just prior to the peak of the cycle.[3] This hypothesis is not completely accepted and ironically cannot be entirely supported by data in this study. As it now seems evident that Cycle 22 probably peaked during late 1989, most of the year fell into the period just prior to the peak. Contrary to previous trends, 1989 was not a below average year (see Table 1 again), although the downward trend from 1987 to 1989 is consistent with prior cycles. If previous patterns of E_S variation across the solar cycle hold up, sporadic E should reach a low point during 1990. In any event, the complete picture for the current cycle will not become evident for several more years.

Hourly and daily measures of geomagnetic activity, which are also a function of solar activity, provide another opportunity for hypothesis testing. Some researchers have suggested that E_S formation is retarded by high levels of geomagnetic activity, such as measured by the daily A-index. Although it is hazardous to draw any conclusions based on a year's worth of observations from one location, some interesting relationships appear in the data. A daily plot of E_S minutes and A-index for the May-August period (Fig 3) reveals an certain inverse relationship between the two phenomena.[4] When the A-index was above 30, sporadic-E formation appeared to be suppressed. Conversely, long-duration E_S events occurred primarily when the A-index was no higher than 15. A scatter graph of the same data (Fig 4) shows this more clearly. The one exception occurred on August 18, when 365 minutes of sporadic E was observed when the A-index was 29. We can offer no explanation for this anomaly.

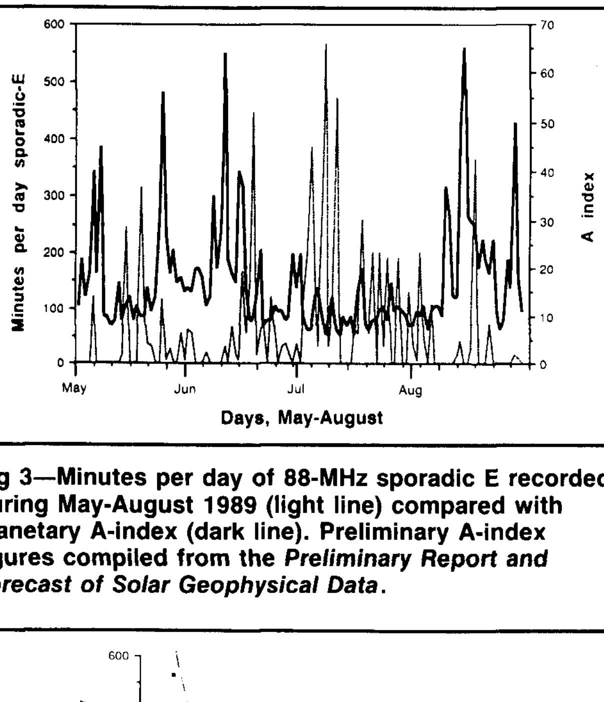

Fig 3—Minutes per day of 88-MHz sporadic E recorded during May-August 1989 (light line) compared with planetary A-index (dark line). Preliminary A-index figures compiled from the *Preliminary Report and Forecast of Solar Geophysical Data*.

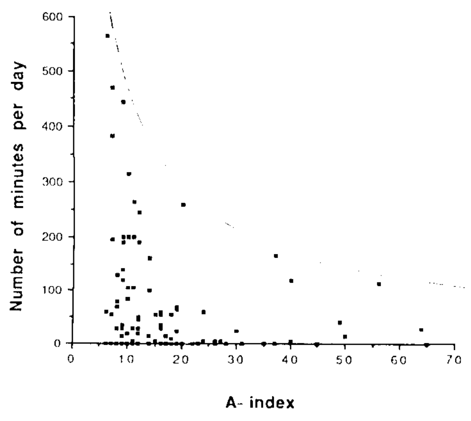

Fig 4—Minutes per day of 88-MHz sporadic E and planetary A-index, May-August 1989. The dashed line has no formal statistical significance, but it suggests a strong inverse relationship between the two variables.

Even if this suggested relationship can be substantiated from other observations, it has limited value as a predictor of sporadic E. A low A-index is no guarantee of E_S—although it may be one of several necessary conditions, as the data clearly shows that on many days when the A-index was no higher than 15, during the peak spring-early summer period, there was little or no sporadic E evident at 88 MHz. When this analysis is made for the entire year, the daily A-index bears no significant relationship to the appearance of E_S. Sporadic E was rare no matter what the A-index. Without a clear understanding of what other variables play a role in E_S formation, the function of the A-index as a predictor will remain uncertain.

Prospects and Acknowledgement

We hope to make a more complete report on the occurrence of 88-MHz sporadic E after the end of 1990, when 11 years of observations will have been accumulated using consistent criteria. Analysis of one solar-cycle's worth of E_S data will undoubtedly provide further insight into the behavior of this most perplexing propagation phenomenon. Thanks to Dr Michael Adams, Eastern Connecticut State University, for his help in creating the graphics.

About the Authors

Emil Pocock has been an ARRL Technical Advisor since 1984 and a frequent contributor to *QEX* and *QST*. He is an associate professor of history at Eastern Connecticut State University.

Pat Dyer became interested in VHF sporadic-E via TV DXing during the early 1960s and received his Technician class ticket in 1963 in order to study 50-MHz E more actively. Dyer earned a BS in physics from the University of Texas (Austin) in 1971 and subsequently worked on VHF sporadic E at the Office of Telecommunications in Boulder, Colorado. Dyer upgraded to Advanced in 1977. His articles on VHF propagation have appeared in *CQ* and *Ham Radio*.

Notes

[1]See Emil Pocock, "Sporadic-E Propagation at VHF: A Review of Progress and Prospects," *QST*, April 1988, pp 33-39.

[2]Some exemplary amateur studies of sporadic E include Mel Wilson, W2BOC, "Midlatitude Intense Sporadic-E Propagation," *QST*, Dec 1970, pp 52-58, and Mar 1971, pp 54-57; Patrick J. Dyer, WA5IYX, "Fifty-Megahertz E_S, 1964-1970," *Radio Science*, 7, Mar 1972, pp 351-53; and Michael Owen, W9IP, "The Great Sporadic-E Opening of June 14, 1987," *QST*, May 1988, pp 21-29.

[3]This is discussed in more detail in Pocock, "Sporadic Propagation at VHF," *QST*, April 1988, pp 34-35.

[4]E_S data was collected in days defined in terms of Central Standard Time, whereas the A-index is a given for the UTC day. Although a discrepancy of 6 hours exists between these two sets of data, the possible distortion is not great. The A-index peaks in Fig 3 could be shifted by a quarter of a day, for example, with no change in the conclusions.

VHF+ Technology

By Geoff Krauss, WA2GFP/3
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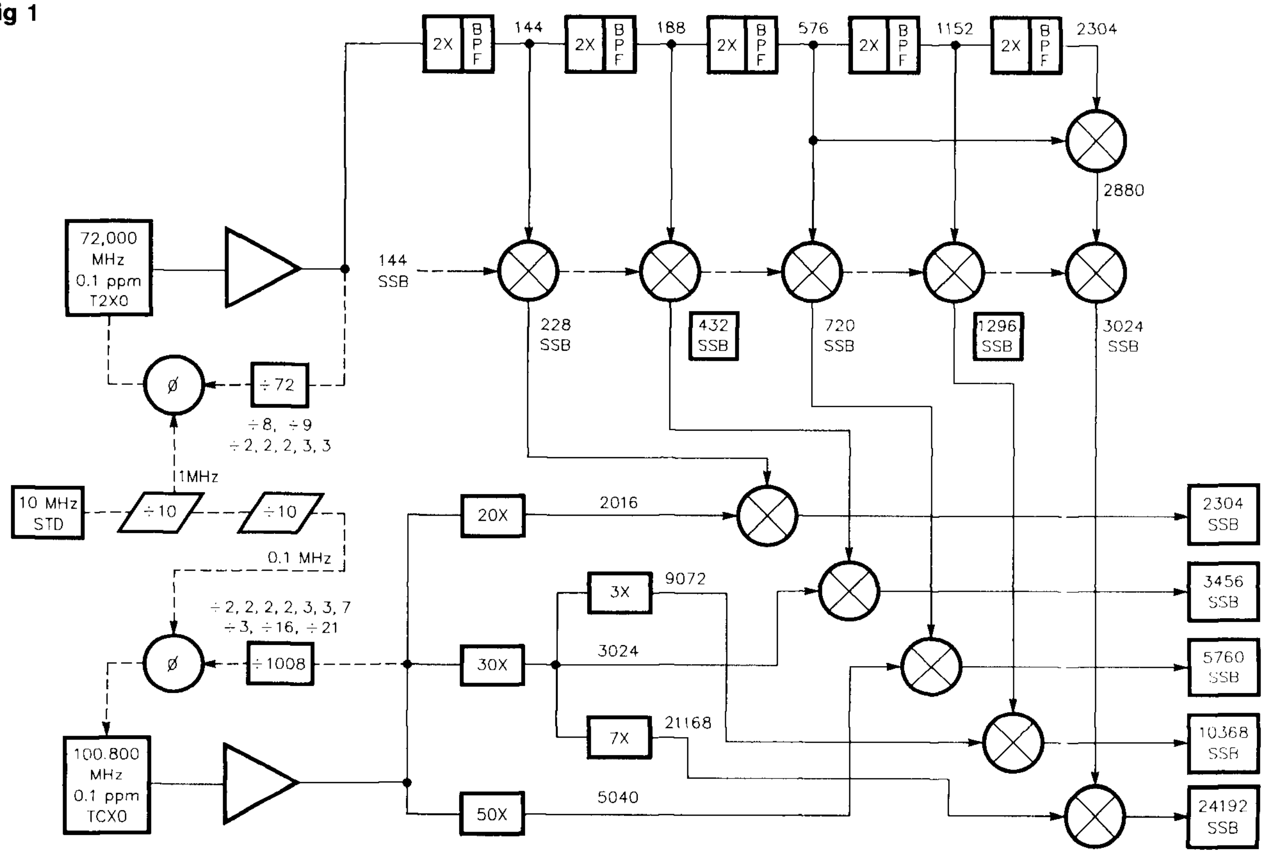
Dresher, PA 19025

Now is the time to better your performance—and the first thing you have to do is make a commitment! Plan to do *something* to better your station, contest rig, etc. It does not have to be a huge project; it does not have to involve large amounts of money, parts, time or other people (though those are often the projects that are, somehow, the most fun). But *now is the time!!* The doldrums of high winter get to most people, unless you are so busy you don't notice. Okay, you say, now that 'GFP has soapboxed, what is he planning for this winter? Having just moved, and with three (3) rooms of "stuff" to sort, reconnect and test (as well as setting up the shack in a "cold" nonradio room, planning tower and outdoor installations for the spring, etc), I got to thinking: This is the perfect time to get my home station updated and poised for the new millenium.

What is the *one* thing I would really like to do differently in this main station, compared to previous work? No sooner was the question asked, then the answer jumped at me: unified microwave-band setups, the smallest number of really stable oscillators used to derive all the needed signals to heterodyne a single common band to any ham band between 1 and 30 GHz. Simple task, right? Actually, I put a second constraint on the problem; I want to use at least some of the phase-locked loop (PLL) "bricks" that I've acquired over the years as part of the local oscillator chains. Is there a way, I pondered, to use just one external oscillator to drive a bunch of the brick sources to get the required signals? Hmmm. I can't seem to find one. How about if I use only two highly stable external oscillators? Well, what if I use a double-heterodyne plan? Can I substitute some extra

Fig 1



mixers and band-pass filters (both easy to build, if you don't need state-of-the-art performance), for extra LO chain components?

Flash of inspiration—see Fig 1; it even works out so that I can lock both external oscillators, if need be, to a very high precision HP standard I acquired someplace in my travels. Note that use of lots of MMICs is contemplated, both as overdriven frequency doublers and as signal amplifiers. Since the Avantek or MCL units are cheap and work well up to about 4 GHz, I should be able to use them to get a 5-10 mW drive to all but the three highest-frequency mixers. I may have to use a stage of GaAsFET amplifier at 5040, 9072 and/or 21,168 MHz, but the designs for the lower two are pretty straightforward these days. I don't want to think about designing an amp at 21 gigs, much less building one; it's the highest band in this scheme, and I won't worry about it now! Of course, being close to the problem has probably blinded me to at least one shortcoming of this scheme (other than the fact that the 72-MHz harmonics are at many lower band edges); if you care to comment, I'll certainly read what you have to say and publish any germane wisdom in future columns. If you don't want to work on such low-frequency stuff this winter, how about doing something at. . .

The Next-to-Final Frontier

For the VHF + er, optical communications are just slightly easier than communications at The Final Frontier—wavelengths between 10 and 1 millimeter (frequencies between 30 and 300 GHz). (By the way, how do you like that as a title for a recurring column, say in *QEX* in 2090!) Above 300 GHz is generally regarded as quasioptical territory, although the present work going on is, in fact, mostly in the truly optical, visible-light portion. In fact, almost all serious work at present is being done at one wavelength: 632.8 nanometers, which is a frequency of 471,089 GHz or about 471 THz (terahertz, or 10E + 12 Hz). This is the red emission line of a helium-neon (HeNe) laser, one of the least expensive visible laser sources currently available.

Safety: It makes no sense to me to discuss state-of-the-art amateur activity if my discussion prompts someone to do a potentially harmful act! Now, you could say that about almost any phase of ham radio (we all know at least one person who has been hurt, if only in the pocketbook, by indulging in VHF + work), and we all should know about beam-power restrictions and safety requirements if we work with any form of electromagnetic radiation, but this area *is* more dangerous because of potential eye damage. Why is it so necessary to use visible light? In order to avoid actual eye damage, you can either not work with high-power-density optical equipment, or plan ahead and avoid the hazards. In order to know where not to be/look, you have to be able to see the beam. This means use of a visible laser, either as the main beam, or as a collinear "guide" beam for a more powerful laser. Simply because the beam is so collimated (narrow and nondiverging) that even a relatively low beam power has a high-power density (ie, the amount of power in a unit area) and could be hazardous to the user's eye if improperly viewed. A simple calculation will show that a 1-mW laser output in a beam of about 3-mm diameter (and those are typical gas laser specifications) will have a power density of P_{out}/beam area, or about 15 mW/cm^2, which is well above the 1-mW/cm^2 normal limit (which actually is much too large when it comes to eye safety; this is due to the focusing properties of the eye components and the easily disrupted retina upon which the light is focussed). Some protection can be obtained by wearing safety glasses with a special lens that does not transmit laser light of a particular color (wavelength). The percent of transmission can be very low (eg, less than 10%), but that creates a problem in itself: You want to block as much laser light as possible (to minimize harm), but the less light seen, the harder to sense if the beam is on or where it is pointing! I have a pair of relatively cheap HeNe-protective goggles that cut out over 98% of the 632.8-nm beam, but I *cannot* see the beam when I wear the goggles. I use a key-activated lock on the laser power supply and 3-second delay (both of which are required by Federal laws) to make sure that *I* am the one controlling when the beam goes on, and use a positive beam shutter to then unblock the beam aperture only after I have double-checked the pointing of the beam and put on my goggles. Some VHF + ers are working with green-output helium-cadmium lasers which have power ratings on the order of 150 mW and power densities of better than 1 W/cm^2; extra-special care is obviously required.

Modulation—of the beam is required to convey information and has an oblique influence on the safety question. There are two basic types of coherent laser sources: The semiconductor laser and the gas laser. The semiconductor laser is a diode element that requires a few volts and, typically, less than an ampere of current for operation, so that the resulting beam is easily turned on and off. The binary pulsing of the diode current produces a beam which is similarly pulsed (bit rates in the megahertz range are typical). A typical form of information transmission might use a sampling analog-to-digital converter to convert speech into a digital data stream, which modulates the diode laser output. The down side is that most diode lasers produce nonvisible outputs, which are beams with wide dispersion angles (say 20-50 degrees); however, collimators (to get narrow beams which do not spread much over longer distances) are becoming cheaper and laser diodes operating well into the visible range are slowly becoming available, although not yet at amateur-affordable prices. The gas laser is affordable (one source is MWK Industries, 1269 W Pomona, Building 110, Corona, CA 91720) and well collimated, but not inherently modulatable; the laser turns on slowly, so that direct pulsing at any speed greater than one pulse per minute is not practical and any attempt to amplitude modulate does not give much depth of modulation. How is a gas laser modulated? With external mechanical beam occluders. A shutter can be mechanically moved into and out of the beam, cutting the beam off or letting it pass. The blocked beam is usually reflected back off the shutter and may pose a safety hazard if the shutter is not designed and tested to prevent harmful beam scatter/redirection. One particularly interesting method of beam blockage is by use of a mirror cemented to the cone of a speaker; the angle of beam reflection varies with the speaker drive (both amplitude and frequency). Another interesting shutter is a chopping wheel—a circular fan blade with periodic apertures, which can "chop" the beam into pulses. In fact, this method is the most used as the chopped beam can convey intelligence by small changes in chopping frequency and the chopped beam can be heavily filtered to remove outside influences (noise) before amplification. A number of articles and pamphlets have been recently published about these forms, see the Foltzer article in the August 1990 *QEX* and the Atkins articles in March-May 1990 *Ham Radio*.

Receivers come in two basic flavors: semiconductor (diodes or transistors which are photosensitive) operating either as a source or as a photo-variable resistance (with a voltage between 2 and 300 volts), and vacuum tubes (usually photo-multipliers) which have very high gains (as large as 10E + 7) but require fairly high operating voltage (700-2000 V) and relatively complex circuitry. If enough interest is shown, I might be persuaded to do a future column on these interesting "receiving" tubes.

Antennas can be almost anything having light-gathering or amplification properties; of particular interest on reception are lenses of large size, such as flat-plate Fresnel lens. Typical sizes can be 10 × 2 inches, which has a lot more light-gathering ability (10000 ×, or about 40 dB) than the 0.1 × 0.1-inch aperture of a semiconductor photo-diode. Similarly, beam expanders are often used for collimation of transmitted beams at long distances. Again, this esoteric topic deserves a separate column (or two) all to itself. If enough interest is shown, perhaps a real authority in the field can be persuaded to provide a tutorial for *QEX* readers.

Conducted by: Stan Horzepa, WA1LOU
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Wolcott, CT 06716-2702

NK6K TALKS ABOUT FUTURE PACSAT PBBS OPERATIONS

AMSAT-NA Area Coordinator Jim White, WDØE, has asked some questions of interest to OSCAR satellite users about the PACSATs and AMSAT. Software engineer, Harold Price, NK6K, has responded with the following answers.

WDØE: Which PACSAT will commence PBBS operations first?

NK6K: UO-14 is now about 80% functional with the PBBS software. AO-16 will be brought to that level, then LO-19.

WDØE: Will the TLM or Whole Orbit Data (WOD) format, frequency or any other parameters change?

NK6K: This is more of a spacecraft control than an application question. There is nothing in the new code that will require a change. WOD will probably end up as files and be downloaded or broadcastable. This means that it won't come down in frames sent to WOD, but the internal record format will be the same. I'd like to see some compression on the WOD, but we may be running out of code space. If it is compressed, the ground station software will know how to decompress it. We'll do an update to the Operating System runtime libraries next year that will give us back some program memory. There would be no frequency changes.

WDØE: Will people still be able to just digipeat through the PACSATs?

NK6K: This has not been decided yet. Digipeating is a good way to check out your station. I don't think there is much other demand for it in the long run. The Microsats are available now for digipeating, but you don't see much of this type of activity. I would not want to see digipeating supplant the intended use of store-and-forward data, but if there is a big demand, I suspect it would be addressed.

WDØE: How will the "user" ground software be distributed and tested?

NK6K: Jeff Ward, GØ/K8KA, and I have discussed two approaches. However, nothing has been finalized. The first approach is to make a minimum implementation of the user ground-based software available, including C language source code, as shareware. We want to get something out soon, and, since Jeff and I are IBM PC-based, availability of code will hopefully encourage others to write for Mac, Amiga, C-64, UNIX and others. The only downside is we don't want to leave the impression that all AMSAT offerings are shareware. The minimum implementation will not be automated and will not be pretty (colors, windows, pull-down menus, etc), but will be more portable. The second approach is for AMSAT-NA and AMSAT-UK to make available an "all-singing-all-dancing" PACSAT ground station program. This will be automated, easily interfaced to terrestrial PBBSs, and have a fancy menu-driven format. No source.

WDØE: How will WOD dumps and PBBS output be interleaved on the downlink?

NK6K: The broadcast protocol provides for demuxing broadcast files. This is probably how WOD will be done. These frames are sent to the address QST-1. Direct connect PBBS file transfer is done in frames sent to the user's address. The broadcaster controls the repetition rate for simultaneous broadcast files, based on the number of users requesting a file or the file's priority. The method for prioritizing broadcast vs direct connect (FTLO) is to be determined.

WDØE: How will people know the PBBS is running and available? Will they be able to tell by looking with an ASCII terminal program with TLMDC or similar software or will they have to have the ground software to tell?

NK6K: Much ado will be made. Also, you'll see frames sent to QST-1 with a PID of PBBS.

WDØE: Will the user ground software be available before the PBBS in the bird is turned on?

NK6K: Available, yes. Widely distributed? No. We have to have it running on the air to give it a good test, but don't want to subject it to simultaneous connects on the first day. Once a week's worth of testing by users has been done successfully, the shareware will be placed on CompuServe and elsewhere.

WDØE: Will you be testing PBBS and ground software together for a time prior to making ground software available? How will people know that is happening?

NK6K: Yes, we will. This is happening as we speak on UO-14. People will learn the same way as always, via official AMSAT nets and publications.

from AMSAT

WALKING MOBILE PACKET-RADIO STATION AROUND THE EARTH

(Editor's note: Fred Moore, KG6HQ, is one of those free spirits that actually practices what he preaches. He represents, I think, a little part of us all. And he is not some far-out weirdo. Fred founded the Homebrew Computer Club! See the book *Hackers* by Steven Levy. We hope you enjoy Fred's short article.)

I have been walking south from Vancouver since August 31, 1989, and crossed the Golden Gate into San Francisco on Earth Day. This walk, which I call the "Walking Rainbow," is a long-term noncommercial project to promote respect for the Earth by living simply, in harmony with nature. Less is best.

Combining high tech and low tech, I have designed and built two lightweight "tension-structure" carts to carry my camping equipment and supplies. (I don't need two carts, but bring them both so that other walkers who join will have use of a cart. Each cart can carry 300 lbs or more.) I am testing out a portable single-pot, wood-burning cook stove of my design, which is fuel efficient because the air is preheated before entering the combustion chamber. I also have plans for building a portable solar oven, portable yurt, power-assist motor for going up hills, corn grinding brake for slowing downhill, and other appropriate tools for nomadic living.

While in the San Francisco Bay Area for a couple of weeks, I plan to equip the carts with 12-volt photovoltaic modules to power a portable computer and packet-radio gear. The computer would be used to write and publish articles about my designs, prepare graphics and technical drawings, simulate and model the physics of these designs to optimize them, keep a diary for an eventual book about the walk, play with cellular automata programs for enjoyment and insights into how natural systems might work. The computer will also maintain and address list of contacts and be the terminal for packet-radio communications via the Microsats (while I am walking in foreign countries.) So, I am looking for a portable computer workstation that can be used in the field, such as an AGILIS System. I welcome help in locating sponsors for this project. Advice and technical assistance are also needed.

(Editor's note: I know that Fred is very interested in finding a way to get an inexpensive Macintosh Portable to take with him; any helpers out there?)

The walk route will take me down the coast to Los Angeles, east through Arizona, New Mexico, Texas, and south from there (Mexico, Central America, South America, etc). I'll be checking in to the packet-radio frequencies on 2 meters and 70 cm along the way and would be delighted to meet hams interested in this venture. Perhaps a few amateurs would be willing to act as base stations and gateways to various computer nets for the reports I'll be sending back when I'm south of the border. It is going to take some advance work to get permission to operate packet radio in other countries and I need to begin that process now. Any suggestions of contacts in Mexico or Central America who could assist in this process would be greatly appreciated. Thank you.

by Fred Moore, KG6HQ (784 Rosewood Dr, Palo Alto, CA 94303, telephone 415-327-1104) from *The NCPA Downlink* (NCPA = Northern California Packet Association)

KANTRONICS V3.0 FIRMWARE READY

Kantronics recently announced the release of firmware update version 3.0 for the following Kantronics KAM, KPC-2, KPC-4, KPC-2400 and KPC-1 TNCs.

The firmware upgrade includes several major features:

- Reverse-forwarding of personal messages,
- Software carrier detect,
- Host Mode, AMTOR-625 (KAM only) and NAVTEX/AMTEX (KAM only), and
- New commands that allow flexible use of command parameters.

The PBBS in all units now supports reverse-forwarding of private and traffic messages. Enter your messages into your mailbox, then arrange with your local full-service PBBS to reverse forward them in the middle of the night when the frequency is not so congested.

Kantronics has also simplified the use of the PBBS from the local terminal. No more pbl, pbr, pbk, pbs; just issue a connect and use it like you would use a remote PBBS. Commands are handled by the local unit and not transmitted.

REROUTE is a new command that lets you change the "to" field of a message already in the mailbox.

The KAM now supports the NAVTEX message format and remembers the last 256 messages received (50 from any one station). This also supports the new AMTEX format being used by ARRL to transmit their bulletins each day. You can copy only those messages you are interested in and avoid duplicate copies of the same bulletin automatically.

The recently revised Part 97 of the Amateur Radio rules allows the use of the CCIR 625 specification for AMTOR in the amateur bands. The KAM now fully supports this mode, allowing 7-character or 9-digit selcals. Enhancements include relinking and busy signalling to other stations.

For FEC operation, Kantronics has added the TXDFEC command, which allows you to specify extra phasing characters at the beginning of your Mode B transmission. This allows other stations the extra time required to tune in your signal and obtain a lock before sending data.

RESTORE is a new command that allows you to restore the PERMed parameter values. PBBS messages will not be destroyed if the PERMed value is large enough to hold all current messages. RESTORE D restores the factory defaults.

The Kantronics Host Mode is now included in the KAM and KPC products, allowing special terminal programs to access all features of the TNC while operating in packet-radio mode.

Software Carrier Detect has been added to the KAM and KPC-4 and has been improved in the other units. By turning this on, you can run open squelch and detect weaker signals. Carrier will only be detected when packet-radio signals are present.

BEACON intervals are now in 1-minute increments.

CWID is now available in all units and operates in the KISS Mode as well as in normal AX.25 mode.

MYAUTOST (KAM only) sets the call sign used for AUTOSTART in RTTY and ASCII. It accepts up to 7 characters, specifically for MARS operators.

PREKEY (KAM HF port only) allows you to set a delay between asserting the PTT of your radio and sending audio tones from the KAM. POSTKEY (also KAM HF port only) sets a delay between sending audio tones from the KAM and dropping the PTT of your radio.

Battery Backup keeps parameters and PBBS contents. SmartWatch does the same plus keeps date and time.

Units purchased after July 1, 1990 are entitled to a FREE update. For more information, call or write Kantronics, Inc, 1202 E 23rd St, Lawrence, KS 66046, telephone 913-842-7745, FAX 913-842-2021, BBS 913-842-4678.

TRANSPACK: NETWORKING THE SOVIET UNION

Andy Fyodorov, RW3AH, who lives in Moscow, and amateurs of the Soviet Amateur Radio Emergency Service (SARES) would like to establish an HF packet-radio network, based on the territory of the USSR, in such a way that packet-radio message forwarding may be streamlined from Alaska and the United States into Europe and Asia through the Soviet Union. They want to establish a system of PBBSs (to be called "TRANSPACK") in their country and get in touch with interested amateurs worldwide to work on this project. Of course, such a system will give a welcomed boost to packet radio in the USSR, in general, and allow more people to use this new exciting mode, not to mention benefits to hams worldwide by alleviating congestion of some more popular forwarding routes.

They would like to see it fit well into the existing system of HF PBBSs and are in need of practical advice and some support. If you can help, please write to Andy Fyodorov, Box 899, Moscow 127018, USSR

from Ed Kritsky, NT2X

FIRST LATVIAN HF PACKET-RADIO QSO

John Seney, WD1V, recently completed a 20-meter contact with the first active packet-radio station in Latvia, Oleg, YL1ZW. At the time of the contact, Oleg was using an ICOM IC-751A and power amplifier into a 6-element Yagi with an AEA PK-232 TNC. Oleg's location was in the town of Valka.

WD1V @ KB4N.NH.USA.NA

SAREX PACKET-RADIO LOGS AND QSLS

There have been several questions concerning the logs and QSL cards for the STS-35 SAREX packet-radio robot. Stations making successful two-way contacts with the robot will know they are in the log when their call sign and QSO serial number is sent in the QSL beacon ("WA4SIR>QSL:. . ."), which lists the last 22 contacts. The QSL information for the last 600 contacts is saved in the TNC's RAM and these logs will be dumped frequently to the laptop computer in the orbiter. If you work the robot several times (even with different SSIDs) your call sign will appear in the QSL beacon and logs only once. If your call sign appears in these logs, you will receive a QSL card commemorating the feat.

Stations heard by the orbiter (including those who make successful contacts) will be listed in the QRZ beacon ("WA4SIR>QRZ:. . ."), which lists the last 35 call signs heard. The contents of the QRZ beacons are not saved; the only record of them must be acquired by ground stations monitoring the QRZ beacons. So, to have logs to validate call signs for STS-35 SWL cards, we must rely on you. If you receive and save QRZ beacons and wish to submit them to help assemble the worldwide logs, you can submit them in several ways: as a packet-radio message sent to SAREX@W3IWI.MD.USA, as Internet SMTP mail addressed to SAREX@tomcat.gsfc.nasa.gov, on an MS-DOS disk mailed to SAREX, AMSAT, 850 Sligo #600, Silver Spring, MD 20910 or, if all else fails, as printer output to the Silver Spring address. Keep the messages brief. All that is needed is the ASCII beacon text with a time stamp. Any other material will be ignored. It won't be possible to return your disks if you choose to mail in the data.

from Tom Clark, W3IWI @ W3IWI

NETWORKING CONFERENCE PROCEEDINGS AVAILABLE

The ARRL/CRRL Computer Networking Conference was held in London, Ontario, Canada on September 22. The proceedings of this ninth edition of the conference are available now from ARRL HQ for $12.00 plus $2.50 ($3.50 UPS) for shipping and handling. The papers contained in this year's edition are:

"9600 Baud Operation" by Phil Anderson, WØXI, and Karl Medcalf, WK5M

"Digital Regenerator Modification for the TNC-2A" by William A. Beech, NJ7P, and Jack Taylor, N7OO

"Considering Next-Generation Amateur Voice Systems" by Jon Bloom, KE3Z

"AVC_R_ISA: A MAC Layer for NOS/net" by F. Davoli, A. Giordano, I1TD, A. Imovilli, IW1PVW, C. Nobile IW1QAP, G. Pederiva IW1QAN, and S. Zappatore IW1PIR

"A Built in TNC for the Toshiba Mod. T1000" by Frederic de Bros, KX1S

"A GPS Data Receiver" by Dan Doberstein, M.S.E.E.

"Physical Layer Considerations in Building a High Speed Amateur Radio Network" by Glenn Elmore, N6GN

"Hubmaster: Cluster-Based Access to High-Speed

Networks" by Glenn Elmore, N6GN, Kevin Rowett, N6RCE, and Ed Satterthwaite, N6PLO

"Adoption of the KA9Q TCP/IP Package for Standalone Packet Switch Operation" by Bdale Garbee, N3EUA, Don Lemley, N4PCR, and Milt Heath

"Network Routing Techniques and Their Relevance to Packet Radio Networks" by James Geier, Martin DeSimio, WB8MPF, and Byron Welsh, KD8WG

"Status Report on the KA9Q Internet Protocol Package for the Apple Macintosh" by Dewayne Hendricks, WA8DZP, and Doug Thom, N6ØYU

"Texas Packet Radio Society Projects: An Update" by Greg Jones, WD5IVD, and Tom McDermott, N5EG

"FlexNet - The European Solution" by Gunter Jost, DK7WJ, and Joachim Sonnabend

"MACA1 - A New Channel Access Method for Packet Radio" by Phil Karn, KA9Q

"Forward Error Correction for Imperfect Data in Packet Radio" by W. Kinsner, VE4WK

"The Network News Transfer Protocol and its Use in Packet Radio" by Anders Klemets, SMØRGV

"Packet Radio with Rudak II on the Russian Radio-M1 Mission" by Hanspeter Kuhlen, DK1YQ

"CELP High-Quality Speech Processing for Packet Radio Transmission and Networking" by A. Langi, and W. Kinsner, VE4WK

"The PackeTen System - The Next Generation Packet Switch" by Don Lemley, N4PCR, and Milt Heath

"The BPQ Node in an Expanding Network" by Karl Medcalf, WK5M, and Phil Anderson, WOXI, in cooperation with John Wiseman, G8BPQ

"Node Networking" by Donald R. Nelsch, K8EIW

"The 'Cloverleaf' Performance-Oriented HF Data Communication System" by Raymond C. Petit, W7GHM

"Frequency-Stable Narrowband Transceiver for 10100.5 kHz" by Raymond C. Petit, W7GHM

"PACSAT Protocol Suite - An Overview" by Harold E. Price, NK6K, and Jeff Ward, GØ/K8KA

"PACSAT Data Specification Standards" by Harold E. Price, NK6K, and Jeff Ward, GØ/K8KA

"PACSAT Protocol: File Transfer Level 0" by Jeff Ward, GØ/K8KA, and Harold E. Price, NK6K

"PACSAT Broadcast Protocol" by Harold E. Price, NK6K, and Jeff Ward, GØ/K8KA

"PACSAT File Header Definition" by Jeff Ward, GØ/K8KA, and Harold E. Price, NK6K

"A Fast Switching, Wide Bandwidth Transceiver for 70-cm Operation, The DVR 4-2" by Jerry Schmitt, WXØS, Phil Anderson WOXI, and Bruce Kerns

"Long Distance Packet Mail Via Satellite" by Mark Sproul, KB2ICI, and Keith Sproul, WU2Z

"Station Traffic System: A Traffic Handler's Utility Package with Integrated Packet Support" by Frank Warren, Jr., KB4CYC

"Comments on HF Digital Communications Part 1—Link Level Issues" by Tom Clark, W3IWI

"Comments on HF Digital Communications Part 2—Data Protocol Issues" by Tom Clark, W3IWI

"BULLPRO—A Simple Bulletin Distribution Protocol" by Tom Clark, W3IWI

"Some Comments on the 'H'ierarchical Continent Address Designator" by Tom Clark, W3IWI

PACKET RADIO COVERED IN SOUTHERN CALIFORNIA 6-METER BAND PLAN

At a meeting sponsored by the Southern California Repeater and Remote Base Association (SCRRBA), representatives of all Southern California 6-meter band users' groups vigorously discussed proposals, debated alternatives, and then unanimously adopted a new 50-54 MHz band utilization plan.

Highlight of the new plan is a "modular" approach to utilization of the spectrum authorized for repeater-based communications. (Based upon Rule Making petitions originating in Southern California, the FCC recently increased the repeater spectrum available on 6 meters from 52-54 MHz to 51-54 MHz.)

Recognizing that the exact pattern of utilization of the band for repeater communications in various parts of the country depends upon the presence or absence of Television Channel 2 broadcasting (54-60 MHz) in each area, the new Southern California plan divides the 51-54 MHz segment utilizing the new repeater plan. The coordination centered around two major proposals: 500-kHz and 1000-kHz input/output spacing (both involving repeater inputs lower than outputs, both utilizing 20-kHz channel spacing on an "even" channel raster, eg, 51.12, 51.14, 51.16, etc).

According to the plan, the following frequencies have been set aside for digital/packet-radio operation:

Sub-band	*Mode of Operation*
50.300 - 50.600	all modes:
50.620	digital (packet radio) calling
51.120 - 51.480	repeater inputs:
51.120 - 51.180	digital repeater inputs
51.620 - 51.980	repeater outputs:
51.620 - 51.680	digital repeater outputs

Activities above 51.100 MHz are set on 20 kHz-spaced "even channels"

from *Repeater Coordinator's Newsletter*

GATEWAY CONTRIBUTIONS

Submissions for publication in *Gateway* are welcome. You may submit material via the US mail or electronically, via CompuServe to user ID *70645,247* or via Internet to *70645.247*@compuserve.com. Via telephone, your editor can be reached on evenings and weekends at 203-879-1348 and he can switch a modem on line to receive text at 300, 1200 or 2400 bit/s. (Personal messages may be sent to your *Gateway* editor via packet radio to WA1LOU @ N1DCS or IP address 44.88.0.14.)

The deadline for each installment of *Gateway* is the tenth day of the month preceding the issue date of *QEX*.