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QEX 111 WITH Gateway

ARRL Experimenters' Exchange

MAY 1991

A Series Regulator Power Supply

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Experimenters' Exchange
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TABLE OF CONTENTS

COLUMNS

MAY 1991 QEX ADVERTISING INDEX

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. . .

Request for Comments

The Teleprinter Over Radio (TOR) protocol, known officially as CCIR Recommendation 476 and AMTOR within Amateur Radio, has been around for decades now. Actually, Rec. 476-4 (the latest version) is simply a historical document in that it is the specification for systems already built. Any new TOR systems marketed for the Maritime Mobile Service should be built according to Rec. 625.

The TOR protocol was developed at a time when controllers were not trivial to build. In fact, the hardware limited what was possible. Designers could only dream of adaptive systems—software-controlled equipment that could change according to circuit conditions. Rather than being able to cope with whatever transmission impairments might come along, a TOR system had to be built to satisfy bad, if not worst-case, conditions. It had to be robust enough to sell to ship owners, who were having trouble justifying the cost of an RTTY installation aboard ship. Also, ITA2 (or Baudot) was the code of choice for communications in those days, and TOR was designed to convert 5-level code into a unique 7-level code for transmission then back to a 5-level code upon reception. TOR has served the maritime industry and Amateur Radio well for years, and it has a well-earned reputation for robust performance on the HF bands.

One factor in any radio communications system these days is protocol stability, meaning: Don't change things simply because someone has a slightly better idea. Before users will tumble to a new protocol, there should be some perceived benefit and little or no trouble making the conversion. In the past that meant getting a new box, new software and possibly suffering through several versions before the bugs are exterminated. The state-of-the-art is better than that today in that present-day commercially manufactured communications controllers can accommodate considerable protocol change without becoming obsolete. Future boxes based on digital signal processing (DSP) hardware will be able to incorporate program modifications unanticipated by their designers. Whether or not DSP is involved, any computer-based controller should be flexible enough to accept upgrades in protocols.

Given that as background, the ARRL Digital Committee would be interested in your views concerning the design of a new protocol for use at HF that perhaps would be as robust as AMTOR, have greater throughput, be octet based and be capable of serving as a transparent link in a packet-radio network. Improved throughput could be achieved through use of more efficient modulation, coding and error-control techniques. An octet-based code is needed to handle ASCII, extended (8-bit) ASCII and even Asian languages, such as Kanji, which require 16 bits.

Digital Committee member Paul Newland, AD7I, has been asked to spearhead this effort. There is very little history to this project except for an exchange of views on some gross system parameters. Here are some questions to get started thinking about them:

- Should the transmission block length be fixed like AMTOR or should it be dynamically variable according to circuit conditions?
- Should there be a fixed symbol rate or should that be adaptive?
- Is binary FSK the thing to stick with, or should it use PSK or a multitone modulation scheme?
- What type of error-correction scheme is best?
- Can't we figure out a way to send whole call signs instead of shrunken ones now necessary in AMTOR?
- How would any new system be incorporated into today's or tomorrow's hardware without unnecessary obsolescence?
- What is the best way of using diversity reception to advantage?
- To what degree, if any, should it be backwards compatible with AMTOR?

If you have any ideas on features of such a future system, please write to the Digital Committee, ARRL Headquarters, 225 Main Street, Newington, CT 06111, USA. At minimum, this effort can be at least a way of drawing out some new thinking on HF data transmission. Potentially, however, we could eventually develop a better TOR for the Amateur Service and introduce it to the Maritime Mobile Service. After all, turnabout is fair play.—*W4RI*

The Series Regulator Power Supply: A Closer Look

By William E. Sabin, WØIYH
1400 Harold Dr, SE, Cedar Rapids, IA 52403

For many applications, the rapidly advancing technology of the switching regulator power supply has made it the preferable approach, especially where light weight, small size and high efficiency are very important. But for my basement laboratory requirements I finally decided to build a series regulator supply. During the lab-bench development of sensitive low-level circuitry it is necessary to be sure that the power supply is beyond reproach and not contributing, in confusing ways, to various problems. The "switchers" can be a later addition to the equipment design.

In the course of the initial design work it occurred to me that my understanding of the series regulator was inadequate. The excellent material in references 1 and 2 helped, but several other questions came up. I would like to share with you my additional investigations, and describe the design and construction of the supply.

Requirements

The requirements which I believed essential are listed below. Compromises in cost and complexity are also apparent in these specs.

1) Continuously variable output voltage from 4.5 to 25.0 V. The extra circuitry required to go to zero was not justified.

2) Load currents from 0.0 to 2.5 A, continuous duty.

3) Tight load regulation, better than 0.03%, no load to 2.0 A, 0.1% to 2.5 A.

4) Line regulation 0.01% at 2.0 A dc for 117 to 122 V ac.

5) Very low ac ripple, less than 2 *microvolts* RMS at 2.0 A load.

6) Very low random noise, less than 2 *microvolts* RMS in the 0.1 Hz to 500 kHz band.

7) Use off-the-shelf transformer and other easily obtainable parts.

8) Excellent response to load fluctuations and transients; low output impedance.

After reviewing the switching regulator literature, in particular references 2 and 3, I felt that I could meet these difficult specs much more easily with the series regulator approach, especially since size, efficiency and heat dissipation were not important constraints in this case.

Implementation

Fig 1 is a system diagram of the supply, showing the various elements involved in the up front design. The analyses, simulations, various tests performed and the wiring interconnect approach can all be discussed with respect to this diagram. Fig 1 also includes three test circuits.

A type 723 regulator chip was used because of its simplicity and because its reference voltage is brought out to a separate pin so that I could filter the reference noise, typical of Zener diodes, to a very low level with C5, as suggested in the data sheet for the 723 and later verified to be true. The current-limiting circuitry is also accessible at pins 2 and 3 and is activated by the voltage drop across R2 + R3.

The most important regulator considerations can be described as follows:

A) When the output is 25.0 V at 2.5 A at a line voltage of 117 V ac, the Vcb of Q1 and Q2, and also the difference between pins 11, 12 and pin 10 of the regulator chip, when the ripple waveform on C1 is at its minimum (trough) value, must be sufficient to avoid a dropout of regulation and an increase in ripple output. A large value of C1 is used to reduce ripple voltage. Also, the R1, C2 combination reduces the ac on the regulator chip by a factor of 25 and this helped to avoid the need for an extremely large value for C1. Recall also, that the current flow in Q1 and Q2 is not strongly influenced by collector voltage variations if the base-to-emitter voltage is constant. The alternative to these steps would have been a special higher voltage transformer, with which I did not want to become involved.

B) When the output voltage is 4.5 V at 2.5 A at a line voltage of 122 V ac, the power dissipation in Q1 and Q2 is about 65 W. The heat sink requirements are established at this condition. A room temperature of 20° C is assumed. To minimize heating, it is desirable to have a power transformer with as low a voltage as possible, and the steps taken in A) help to assure this. Minimizing other voltage drops that occur between the emitters of Q1 and Q2 and the output terminal helped to assure that a standard 25.2-V ac transformer would do the job. A bend-back circuit prevents overheating when the output is short circuited.

C) The series regulator is a good example of a feedback control system. The open-loop gain and bandwidth, the phase and gain margins and the transient response are important factors. The goal was to maximize the closed-loop performance of the regulator. The approach was to use a high value of open-loop gain and to establish the open-loop frequency response mainly by means of (a) the RC lowpass filter consisting of C6, the resistances which separate Q1 and Q2 and the output resistance of Q1 and Q2 and (b) a single small capacitor C4 at the regulator chip.

D) The mechanical construction should emphasize heat removal, but a cooling fan would not be used. The maximum load current would be scaled to a level which the components could tolerate. A bend-back circuit would be used to limit the maximum heat dissipation with a short circuited output.

Current Limiting

In Fig 1, when the voltage drop from pin 2 to pin 3 of the 723 regulator reaches about 0.62 V, the output current becomes limited to the value of 0.62 / (R2 + R3) A. R3 is a 0.11-Ω resistor which acts as a shunt for the digital meter which measures load current. Resistor R2 is switch selectable (shown in Fig 4) to produce three values of maximum current, approximately 0.1 A, 0.5 A and 2.5 A. This very simple approach will protect delicate circuits and PC boards from burnout destruction.

Test Circuits

Fig 1 shows three test circuits. One is an adjustable load test circuit which can be modulated linearly (almost) by a sine or triangle wave using a function generator which has a dc offset adjustment (so that the waveform always has positive polarity), or by a bidirectional square wave. This circuit is used to test the response to various kinds of load fluctuations and has proved to be very informative, as discussed later.

The second test circuit (loop gain tester) is inserted into the regulator loop so that a test signal can be inserted in series with the loop in order to measure the *open-loop* gain and frequency response. But you will notice that, at dc and very low ac frequencies, the loop is closed through Rb and Ra, and the dc output voltage is being pretty well regulated, something which is essential to the loop testing. By observing the magnitude (and *rate* of change) of the frequency response it is possible to deduce information about the phase shift[4]. With this information available, the gain and phase margins and therefore the regulation, stability, transient response and output impedance of the closed-loop regulator can be estimated.

The third test circuit is a two-stage opamp preamplifier and oscilloscope, to measure very small signals in the 0.1 Hz to 400 kHz range.

Open-Loop Testing

The test signal which is applied to points A, A′ is reduced 60 dB by Rd and Rc (for ease of adjustment). Capacitor C couples V_a, the voltage across Rc, to the 723 chip through Ra. Ra is roughly the resistance which the 723 sees in normal operation. The test signal is amplified by, at most, 74 dB on its way clockwise around the loop to the right-hand end of Rb. It is then attenuated by the factor 20 log (Rb / Rc) = 100 dB. This means that the "leak through" back to the 723 input is much smaller than the V_a that we started with, if the frequency is 2 Hz or greater. At dc and very low frequencies the regulator functions *somewhat* normally. Above 2 Hz, then, the magnitude of the open-loop gain at the test frequency is very nearly the ratio of $|V_o| / |V_a|$.

The first benefit of this tester was that it isolated an instability in the 723 chip. An oscillation at several hundred kHz was cured by C4 (33 pF) and C3 (100 μF / 50 V with very short leads). Normally, one would suspect the oscillation to involve the overall loop, but this was not the case. This kind of instability is common in feedback control systems, where everything appears to be functioning

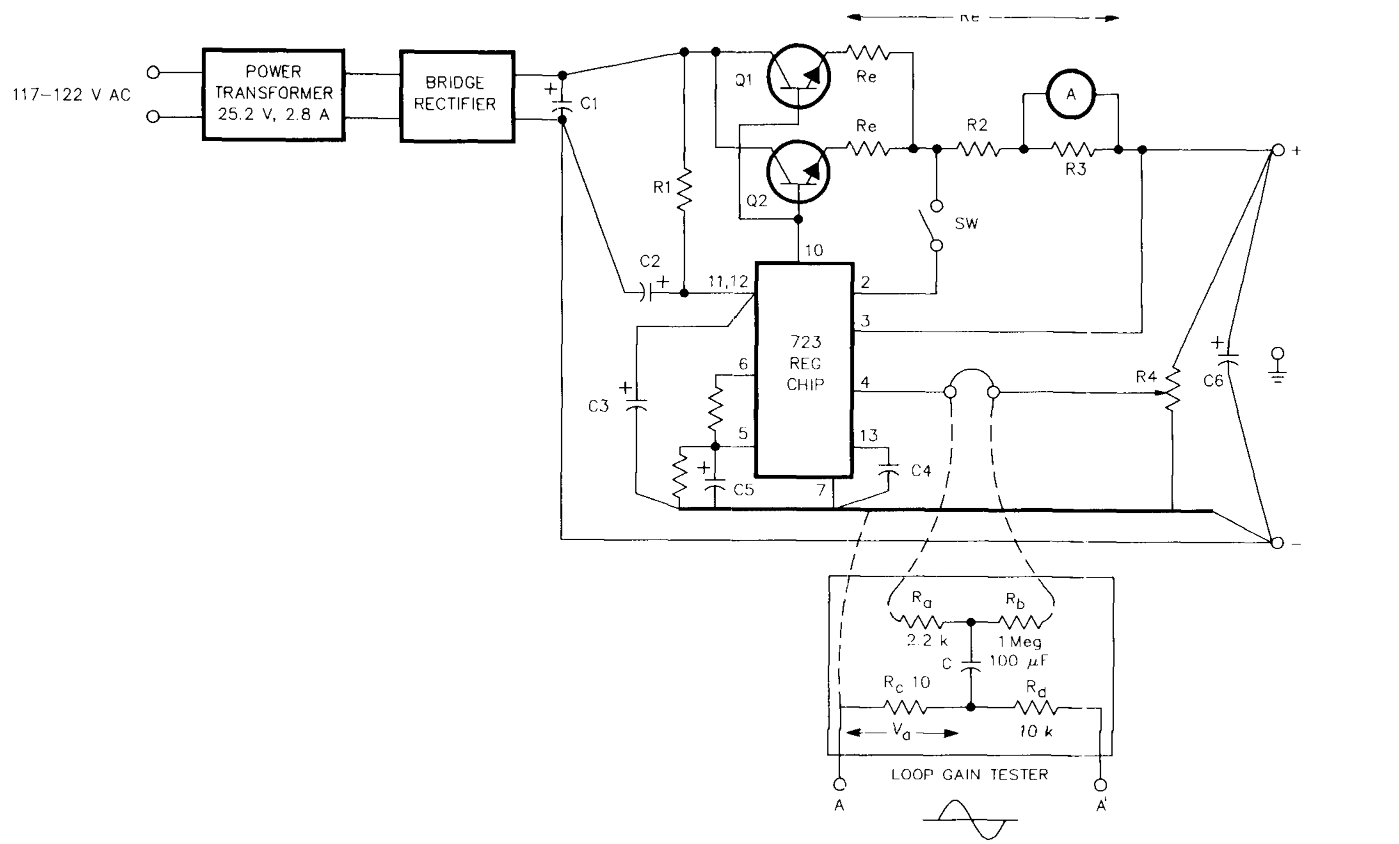


Fig 1—Simplified diagram used to discuss design principles.

(usually not to full specification) but an embedded element is not stable.

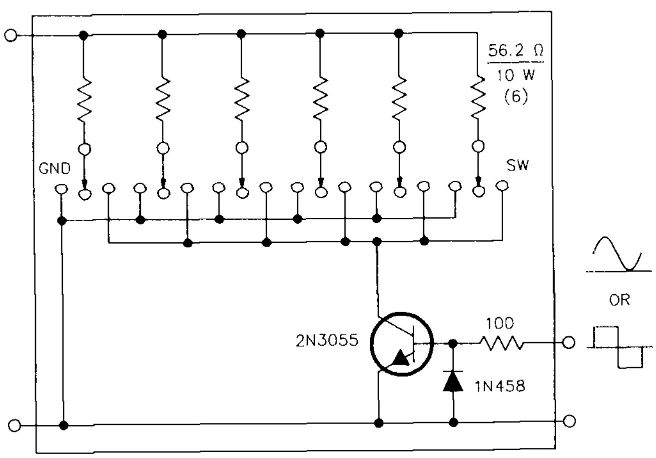
Open-Loop Frequency Response

Looking at Fig 1, the test signal is amplified by about 74 dB (on a voltage basis), on its way clockwise to the emitters of Q1 and Q2. It is then lowpass filtered by C6 and RE (the combination of Re + R2 + R3). It is then divided down by potentiometer R4 (when the output is 25 V, this division is greatest because the pot position is nearest to ground). The open loop gain is the product of these three factors and its greatest value is 59 dB at 25-V output and 74 dB at 4.5 V.

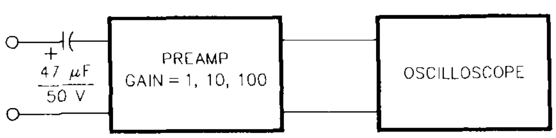
Fig 2 is the open-loop frequency response to the top of R4 when RE is set for the 2.5-A current range. At very low frequencies, the drop-off is due to the gradual closure of the feedback loop, as mentioned above. At higher frequencies the roll off is due to the combined effects of C4 and C6 and occurs at a 6 dB per octave rate (within the errors of my instrumentation). The corner frequency is about 280 Hz, which is $1 / (2 \times \pi \times RE \times C6)$ where $RE = 0.57\ \Omega$ and $C6 = 1000\ \mu F$. For comparison, a reference curve (6 dB per octave at the high and low ends) is superimposed. At about 1.2 kHz or so the reactance of C6 is roughly equal to the ESR (equivalent series resistance) of C6 (about 0.13 Ω for a small 1000 μF aluminum electrolytic, verified by direct measurement). Beyond this frequency, the impedance of C6 does not diminish and C4 takes over, thereby maintaining the 6 dB per octave roll off rate. Careful measurements and computer simulations of the regulator loop verified that C4 and C6 do in fact collaborate in this manner pretty well.

This was the effect I wanted. At 120 Hz (the major ripple frequency) the loop gain is maximum so that the regulator loop is working hard to suppress ripple output. At higher frequencies, the roll off rate of 6 dB per octave implies a loop phase shift in the neighborhood of 90°, which assures closed-loop stability and good transient response. Closed-loop transient response tests using a square-wave signal into the load test circuit verify the absence of ringing and large overshoots.

When RE is set to the 0.5 A or 0.1 A positions, the corner frequencies are 58 Hz or 12 Hz and the roll off rate remains 6 dB per octave as before. In these positions, the loop gain at 120 Hz is reduced, however the ripple voltage across C1 is also greatly reduced at these lighter load currents, and the final result is that the output ripple remains very low.



LOAD TEST CIRCUIT



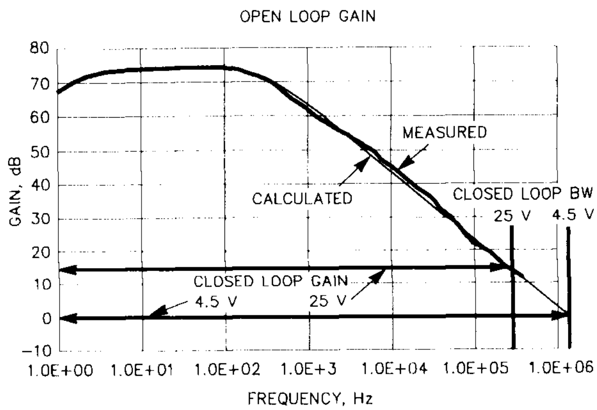


Fig 2—Regulator loop frequency response.

Closed-Loop Response

The closed-loop gain of the regulator is 20 times the log of the ratio of the output voltage, 4.5 V min to 25.0 V max, to the reference voltage, 4.5 V. Fig 2 shows the locations of the min and max gain values and also the corresponding closed-loop bandwidths. By locating the 280-Hz corner frequency fairly close to the 120-Hz ripple frequency, we have made the closed-loop bandwidth no wider than is necessary, which is commendable in a voltage regulator.

Another important parameter in a regulator is its closed-loop output impedance. Fig 3 shows a computer simulation of this. Mathematical analysis and actual measurements using the load test circuit with a sine-wave test signal corroborate the simulations quite well. Two results are shown. In (a) the C6 component (Fig 1) is removed and C4 is increased so that the 280-Hz corner frequency is maintained, as we discussed before. At low frequency, the output impedance should be RE (0.57 Ω) divided by the open-loop voltage gain (5000 max), which equals about 0.11 milliΩ). Above 280 Hz, though, the output impedance increases rapidly because the open-loop gain is decreasing. It will eventually reach the value of RE.

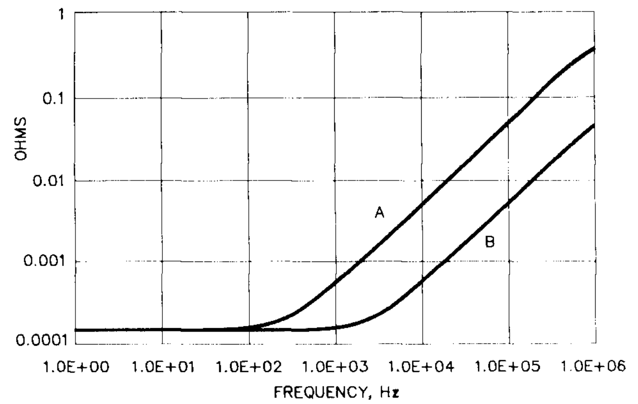


Fig 3—Output impedance magnitude.

In curve (b) the original values of C6 and C4 are used and the output impedance remains low out to about 1200 Hz and then increases but still remains much lower than in curve (a). This is because C4 is now *smaller* than in curve (a) and *it* mainly determines the frequency characteristic. In other words, the impedance of C6 (its reactance plus its ESR) is in parallel with the much lower output impedance of a high-gain feedback amplifier and therefore it is much less influential in determining the power supply output impedance. This situation gradually changes as frequency gets higher, as some study of the following equation will show.

$$\frac{1}{Z_{out}} = \frac{1}{RE}\left[\frac{K\beta}{1 + jf / f4} + 1\right] + \frac{1}{ESR}\left[\frac{jf / f6}{1 + jf / f6}\right]$$

K = amplifier gain
β = R4 divider ratio
f4 = corner frequency for C4
f6 = corner frequency for C6 and its ESR
RE see Fig 1

The result of this discussion is that the output impedance characteristic of the power supply is reduced at frequencies which may be significant in certain applications. Furthermore, it can be reduced to levels (by virtue of the feedback) which a practical capacitor may not be capable of. Of course, to take advantage of this lower impedance the regulator must be located extremely close to the application (remote sensing is also a possibility). Recall, also, that some small value of C4 was needed to stabilize the 723 chip.

Regulation and Wiring

When the line voltage was changed from 117 to 122-V ac, at about 25-V dc, 2.0 A, the output voltage varied less than 0.01%. When the dc load was changed from 0 to 2.0 A the output changed less than 0.03%. Heavy-duty binding posts reduced a small but significant voltage drop from the rear to the front of the front panel. When extremely tight regulation and low output impedance are important, the power leads to the load must be very short, heavy straps. Multiple loads should "fan out," both plus and minus, from the binding posts, not connected in tandem.

Fig 1 details the method of wiring the critical circuits and the following items are enumerated. Each was verified to be important to achieve the clean performance described above.

A) C1 is wired with short heavy leads to present minimum impedance for ac ripple. C2 returns to the negative side of C1.

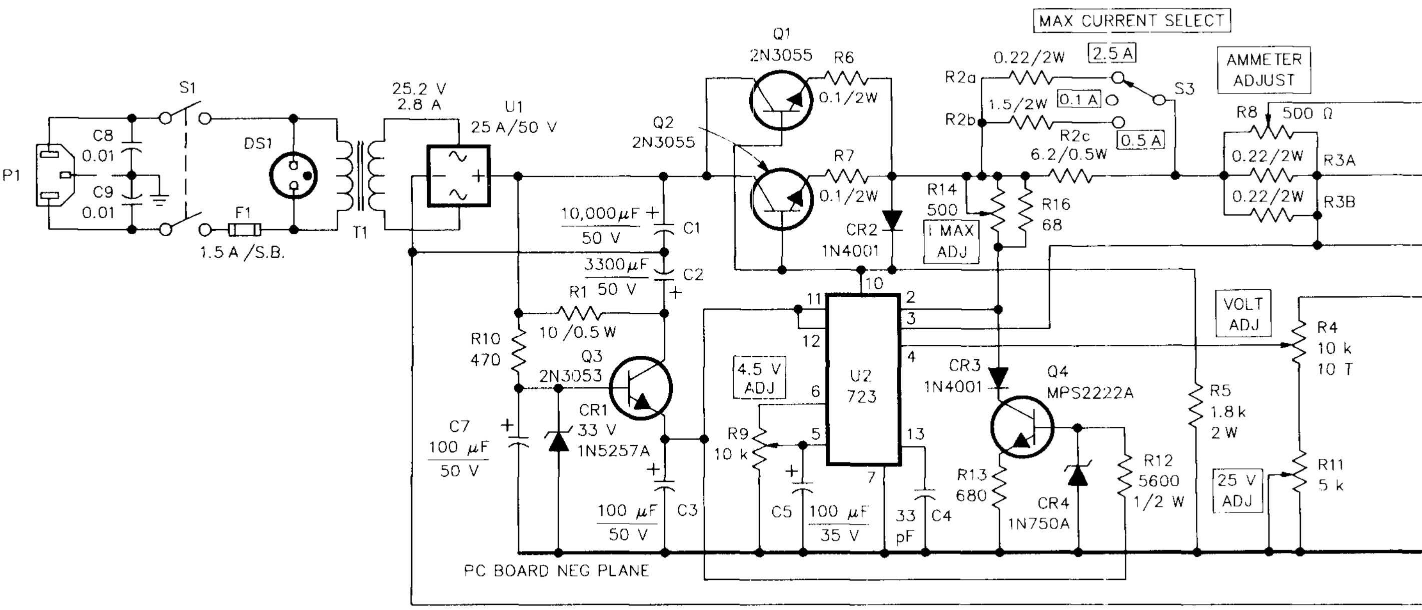


Fig 4—Complete schematic of regulated power supply.

B) C6 is connected directly to the binding posts. R4 returns to the regulator negative plane.

C) The regulator PC board negative plane is connected to the negative output terminal at a single point.

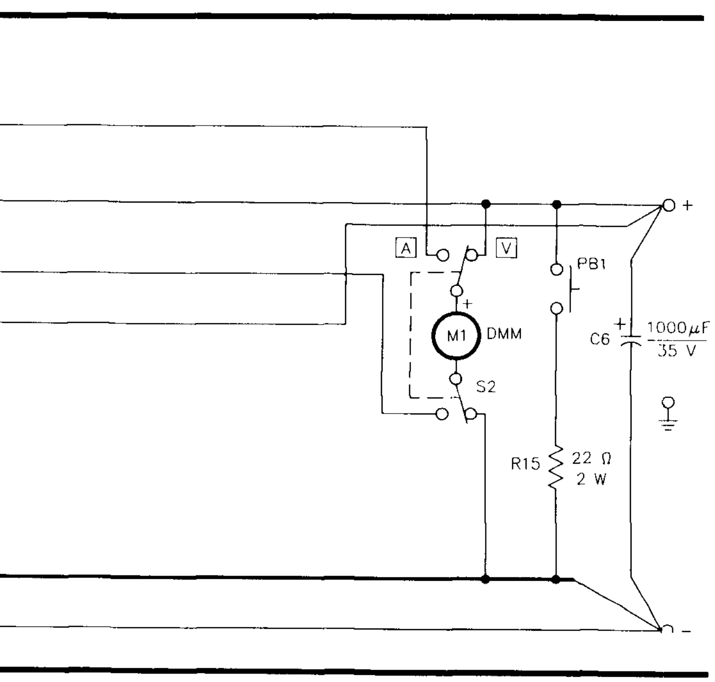
D) The bottom of C1 connects directly to the negative binding post with a heavy lead so that load current fluctuations prefer not to flow on the regulator PC board ground plane and thus influence the operation of the regulator in unpredictable ways.

Complete Schematic

The complete schematic in Fig 4 contains a few features not previously mentioned. The circuitry of Q3, R10, C7 and CR1 prevents the voltage on pins 11 and 12 of U2 from exceeding the 40-V max rating, especially at light loading and high line voltage. C7 eliminates a very small ac ripple at the dc output. As load current increases, the voltage at C1 decreases and Q3 then goes into saturation.

When R4 is turned in a direction to reduce output voltage, U2, Q1 and Q2 are turned off until C6 can discharge to the lower voltage. It was noticed that the emitter-to-base junctions of Q1 and Q2 were going into reverse breakdown (about 2.0 V) so that C6 could discharge through R5. The purpose of CR2 is to prevent this breakdown, which may or may not be dangerous (no actual problem was noticed). The purpose of R5 is to provide a minimum output loading on U2.

The bend-back circuit is interesting. Q4, CR4 and R13 provide a constant current through, and therefore a constant voltage drop across, R16. This is needed to make the current limiting, pins 2 and 3 of U2, work properly over the entire 4.5 to 25.0-V range. But as the current limiting action pulls the voltage at the top of R16 below about 4.0 V, diode CR3 quickly drops out of conduction, the drop across R16 goes toward zero, and the load current falls to and remains at about 1.9 A, thereby limiting the dissipation in Q1, Q2 and T1 and allowing short circuit protection for an indefinite time. This is a regenerative, positive feedback process. R14 is adjusted so that the bend-back starts at about 2.6 A. This circuit was modeled and perfected using a simulation program prior to breadboarding. See reference 1 for further discussion of bend-back circuitry.



The metering is done with a Heath Model SM-2300-A auto-ranging DMM which sells for $20. It is mounted on the front panel and is dedicated to the power supply. The voltage across R3 is 0.11 times the load current. R8 provides the required calibration of the ammeter (0.10 times load current). R3 is an ordinary wire-wound and not a high-grade ammeter shunt, but it is adequate since it does not heat up significantly at 2.5 A. The purpose of R9 is to set the reference voltage on pin 5 of U2 at 4.5 V so that the output can have that minimum value. R11 sets the 25.0-V upper limit. R4 is a ten-turn helipot, for easier adjustment. C8 and C9 are 2.0-kV ceramic, suitable for the ac line bypass function. ***A three-wire line cord is used so that the supply chassis is always tied to building ground, for safety reasons.*** The dc output floats with respect to chassis ground and performance is independent of the grounding connection. Push button PB1 gives a fast discharge of the capacitors (through R15) after turnoff if the load current is very small.

Construction

Figs 5(a) and (b) show the general construction method that I used. The cabinet and chassis surface are 0.062-inch aluminum plates connected by aluminum angle stock, drilled and tapped for 6-32 screws. Ventilation screens at the top and rear provide an excellent chimney effect. The chassis plate is tightly joined to the side plates for better heat transfer. The heat sink selection, one each for Q1 and Q2 (Wakefield 403A), was done according to the excellent discussion in reference 1 and need not be repeated here. The worst case 2N3055 junction temperature was calculated at 145° C, based on a measured (using a Radio Shack 271-110 thermistor that I calibrated myself) case temperature of 95° C and dissipation of 32 W each for Q1 and Q2. This temperature is a little higher than reference 1 recommends, but I consider it acceptable for intermittent lab usage.

The DMM is epoxied to a narrow aluminum strip which is screw mounted to the front panel. The battery compartment at the rear of the DMM is accessed by removing these screws. I really like the dedicated DMM arrangement because of its ability to resolve small changes in volts and amps. I also like the three-position maximum current-selector switch better than a continuous-adjustment potentiometer.

Fig 6 shows the underneath. The PC board is mounted on standoff insulators and the positive and negative output leads are close to the binding posts. All components and wiring are on one side of the board (with the help of a few jumper wires) and the other side is entirely ground plane, except for small circular areas where component through-pads are located. Silastic and pieces of Kraft paper are used to cover up the exposed 120-V ac.

My cabinet construction style is somewhat labor intensive, involving a lot of metal work, and the reader is encouraged to think of simpler approaches, for example,

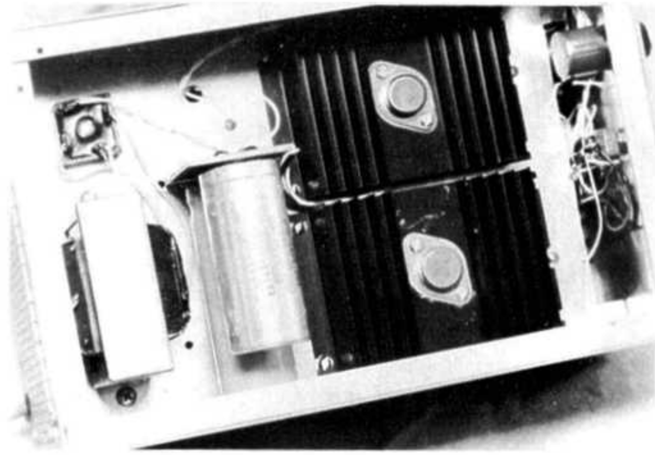


Fig 5—(a) Cabinet (b) top view showing two Wakefield 403A heat sinks.

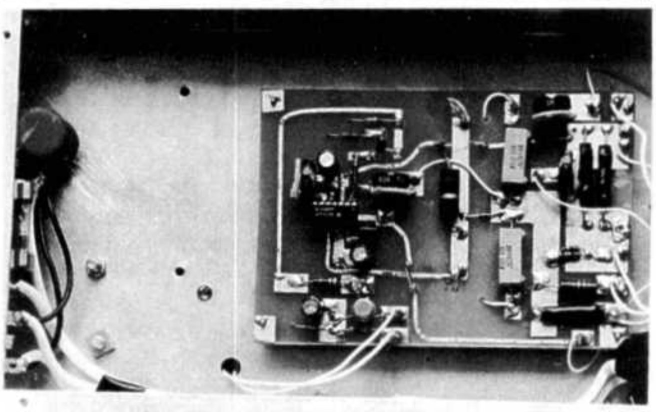


Fig 6—Underneath view.

a 7 × 11 × 2-inch chassis with bottom cover and rubber feet, a 7.5 × 6-inch front panel and some kind of perforated metal cover.

References

[1] *The ARRL Handbook*, 1991, chapters 6 and 27.

[2] DeMaw, Doug, "A 1.25 to 25 V 2.5 A Regulated Power Supply," *QST*, September 1989, and DeMaw, Doug, "Some Power Supply Design Hints," *QST*, November 1989.

[3] Brown, Marty, "Practical Switching Power Supply Design," Academic Press Inc, San Diego, 1990 (Motorola Series in Solid State Electronics).

[4] Van Valkenburg, *Modern Network Synthesis*, chapter 8, Wiley Book Co, 1967.

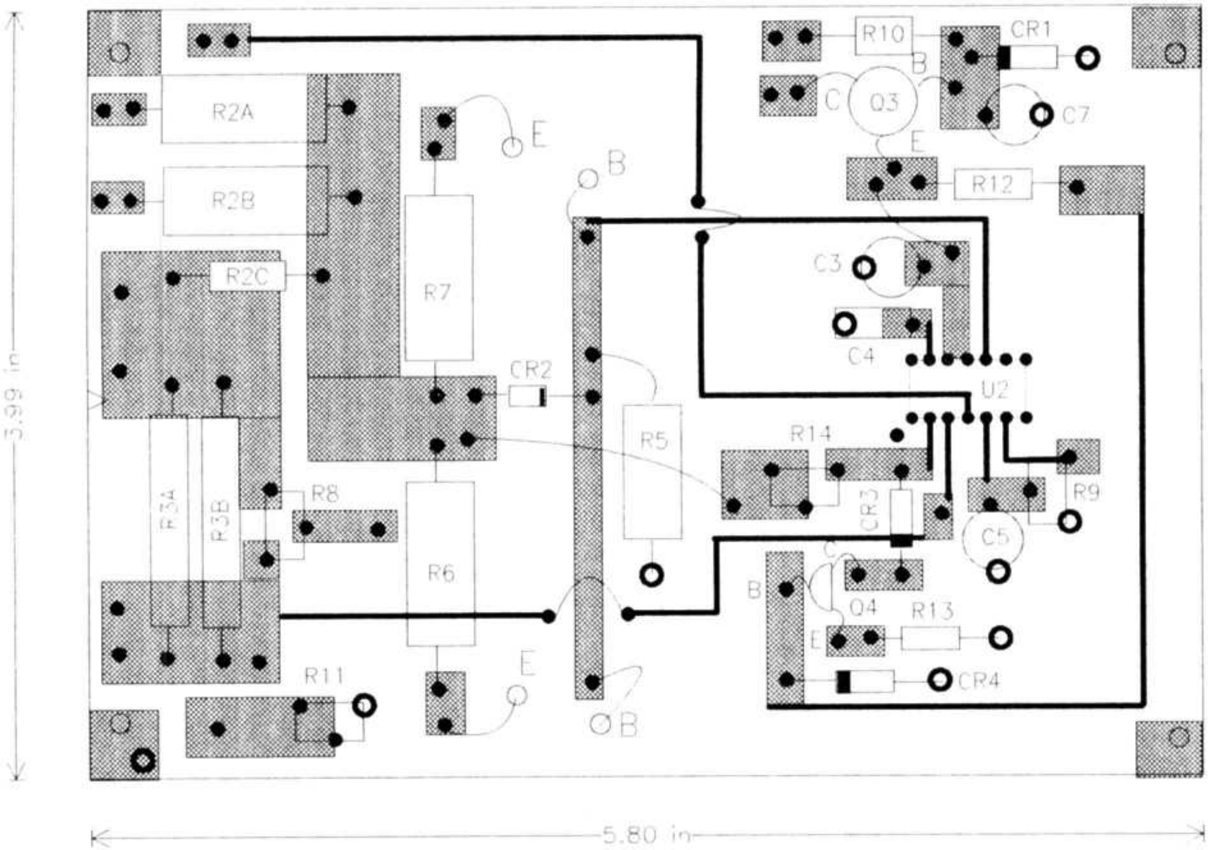


Fig 7—Parts placement.

Electrical Parts List for Power Supply

(RS signifies Radio Shack part number)

C1—10,000 μF, 50 V	CDE 10000-50-AC	$ 8.45
C2—3300 μF, 50 V	CDE 3300-50-AA	6.20
C3,C7—100 $n = muF, 50 V	RS-272-1044	0.89
C4—33 pF, 50 V		0.20
C5—100 μF, 50 V	RS-272-1028	0.79
C6—1000 μF, 35 V	RS-272-1032	1.59
C8,C9—0.01 μF, 2 kV	RS-272-160	0.99
CR1—1N5257A	Motorola	0.40
CR2,CR3—1N4001	RS-276-1101	0.50
CR4—1N750A	Motorola	0.20
DS1—Neon, 120 V ac	RS-272-704	0.90
F1—1.5 A SLO-BLO	RS-270-1284	0.65
M1—DMM	Heath SM-2300-A	20.00
PB1—Push button	RS-275-1547	0.70
Q1,Q2—2N3055	RS-276-2041	3.98
Q3—2N3053	RS-276-2030	0.79
Q4—MPS2222A	RS-276-2009	0.59
R1—10 Ω, 0.5 W		0.25
R2a—0.22 Ω, 2 W	Tresco PW3	0.50
R2b—1.5 Ω, 2 W		0.50
R2c—6.2 Ω, 5%, 0.5 W		0.65
R3A,R3B—0.15 Ω, 2W		0.81
R4—10-kΩ, 10-turn pot	Bourns 3540S	10.49
R5—1.8 kΩ, 2 W		0.25
R6,R7—0.1 Ω, 2 W		1.00
R8,R14—500-Ω pot	RS-271-226	1.40
R9—10-kΩ pot	RS-271-218	0.69
R10—470 Ω, 0.25 W		0.10
R11—5-kΩ pot	RS-271-217	0.69
R12—5.6 kΩ, 0.5 W		0.25
R13—680 Ω, 0.25 W		0.10
R15—22 Ω, 2 W		0.20
R16—68 Ω, 0.25 W		0.10
S1,S2—DPDT	RS-275-652	4.98
S3—SPDT (center off)	RS-275-654	2.39
T1—25.2 V ac, 2.8 A	Stancor P-8388	18.31
U1—Rectifier Bridge, 25 A, 50 V	RS-276-1185	2.69
U2—LM723 regulator	RS-27-1740	0.99
Heatsink (2)	Wakefield 403A	15.00
Heatsink grease	RS-276-1372	1.59
TO3 HDWE for Q1, Q2	RS-276-1371	1.98
Fuse holder	RS-270-739	0.50
Line cord, 6 foot	RS-278-1258	3.00
	Total Electrical Parts	117.23

Simple Crystal Filters

By Bill Parrot, W6VEH
5759 Tanner Ridge Avenue
Westlake Village, CA 91362

It is possible, with some frequency restrictions, to build a practical home-brew SSB or CW crystal filter at very little cost and with very little effort. These filters exploit the fact that "microprocessor crystals," with the required frequency spacings, are readily available. No special tricks are required except the need to know what crystals to use. It's one of those "Why didn't I think of this before?" situations.

This approach to "instant" filters probably was overlooked because, with the exception of a few articles on ladder filters, the building of home-brew crystal filters is becoming a lost art[1]. Many years ago, when WW2 surplus FT-243 crystals were plentiful, home-brew crystal filter design articles were commonplace. Today, much better crystals are available, but since they are sealed and are not easily modified for filter use, half-lattice filter articles are rare. The information provided below should be adequate to build one, but if you want the real details, you'll have to go back to some very old copies of *QST*[2,3,4,5].

Several rather loose rules and observations run through many of the older articles on half-lattice filters. It's handy to summarize some of the important ones here:

a) The filter bandwidth will be about 1.5 times the frequency spacing between the high and low groups of crystals. The precise width depends on many parameters, and is hard to predict accurately without making elaborate measurements and calculations.

b) Narrow filters are easier to construct and adjust than wide filters; very wide filters tend to have a dip in the center of the passband that is hard to eliminate.

c) These filters must be terminated with the proper impedance to keep the passband relatively flat. Wrong or highly reactive terminations will make an otherwise accurate filter practically useless.

d) Some adjustment of filter characteristics can be made by placing very small capacitors or inductors across a crystal. Fortunately, by a trick of circuit transformation, the same effect can be obtained by using a tuned circuit as the coupling element. Note that this tuned circuit is not necessarily peaked at the filter frequency; it is adjusted to provide the necessary reactance to optimize the filter's passband.

e) In a complex network, the crystal's reactances interact with each other. In a properly designed network this interaction is part of the design and is beneficial; in a simple home-brew design it is sometimes appropriate to provide some resistive isolation between sections to reduce undesired interaction. This can improve the filter's shape at the expense of some insertion loss.

f) The crystals in each group should be matched as closely as possible. A total frequency spread of about 100 Hz is appropriate for SSB filters, and 30-50 Hz spread is needed for CW filters.

g) Manufacturers frequently refer to a filter with, for example, eight crystals, as an "eight pole filter." I suppose this terminology is supposed to impress the buyer, but a review of the descriptive equations clearly shows that, within any crystal filter network, there are poles and zeros all over the place due to the interaction of the various elements. This crossed terminology causes no great problem, but is worth keeping in mind when trying to understand why the filter curves change in such complex ways when you are making adjustments.

h) Unlike simple LC filters, there is no mechanism that makes the upper and lower skirts of a crystal filter response curve symmetrical. If the skirts are closely matched, it is due to careful control of the crystal's specifications, and to careful attention to alignment.

Eureka!

Meanwhile, back at the bench, the design described below evolved accidentally out of an attempt to build a ladder filter, using 8-MHz crystals. I had been given a large sack of surplus microprocessor crystals, made up of two different brands. My first step was to build a test oscillator and sort a handful of them by frequency[6]. I was surprised to find that one brand measured about 3 kHz higher in frequency than the other brand. Checking with the original purchaser, I found out that one brand had been purchased as "series resonant" crystals, and the other as "parallel resonant" crystals.

Confusing Terminology

It is necessary to pause here to straighten out some terminology. These two terms, series resonant and parallel resonant, are unfortunate choices to describe crystals since the terms imply that there are major internal differences between the two types. Actually, as the *Handbook* explains, all crystals exhibit both series resonance and parallel resonance. The above terms don't describe the crystals themselves; rather, they are a purchasing shorthand way of saying, "The crystal has been manufactured to be on its specified frequency when used in a series (or parallel) resonant oscillator circuit."

[1]Notes appear on page 14.

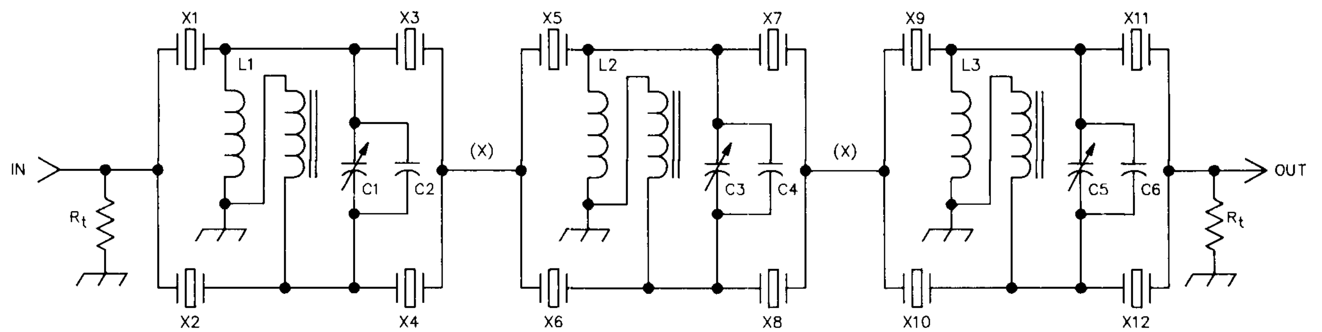


Fig 1—Typical 12-crystal half-lattice filter, configured for 6-MHz operation. Coils and capacitors are roughly resonant at the operating frequency, but may need to be padded for best alignment.

Odd-numbered crystals—6.0 MHz, series resonant
Even-numbered crystals—6.0 MHz, parallel resonant
L1,2,3—20 turns, no. 26 bifilar (40 turns total), Amidon T50-2
Rt—Termination resistors (approximately 510 ohms)
C1,3,5—5-70-pF trimmer capacitor
C2,4,6—47-pF NPO ceramic trimmer capacitor

For an 8-crystal configuration, the center section may be omitted and the points marked (X) joined. Optionally, low-value resistors may be inserted at (X) for some improvement in the filter curve.

If a crystal ordered is series resonant, no further purchasing description is required. However, it is ordered as parallel resonant, it is necessary for the buyer to specify the load capacitance, which typically ranges from about 18 to 32 pF. The result of all this, which is the key to this whole approach, is that there are two groups of microprocessor crystals that are being manufactured, both marked with the same frequency, but the groups actually differ just slightly in their real operating frequencies! The name of the game is to determine what nominal crystal frequency has the proper series versus parallel spacing to build a practical filter, then to find out if both types are manufactured for this chosen frequency, and finally to procure the crystals and build the filter.

The Search Goes On

Unfortunately, the catalogs don't tell you what the frequency would be if you tested a parallel crystal in a series oscillator. Theory doesn't help much since the standard equations assume you already know the crystal's parameters. Intuition, backed by a lot of experiments, tells me that the spacing between groups is very roughly (and I do mean "roughly") proportional to the nominal frequency.

As a preliminary experiment, just to see if I was on the right track, I built an eight-crystal filter using the 8-MHz groups of crystals. The first coil I tried was a bifilar-wound T37-43 with no tuning capacitor. The results were dismal. This configuration probably works if the crystals are precisely on frequency, but I needed some adjustment range. Changing to a tuned circuit, as shown in Fig 1, provided the necessary range.

After adjustment of the capacitors, and with some tweaking of the termination resistances, I was rewarded with a "textbook" filter curve except that, as predicted, the filter bandwidth was about 3.2 kHz, and as further predicted, it had some droop in the passband. This bandwidth was too wide for SSB use, but I proved to myself that it is possible to build a practical crystal filter using only off-the-shelf inexpensive crystals, without opening the cases or otherwise modifying them.

A Lower Frequency Approach

The search was on for the proper nominal frequency. I'll skip over a lot of intermediate experiments with various sets of crystals, and get to the bottom line; at least in my tests, crystals around 5 to 6 MHz have the proper series versus parallel spacing to make very good SSB filters! I can't provide a specific frequency that is optimum; there are too many variables here for exact specifications based on my limited tests. The parallel crystal load-capacitance

What is a "Microprocessor Crystal?"

"Microprocessor Crystal" is an industry term for a class of crystals used primarily in and around computers. They are manufactured as high-volume, low-cost parts, but are readily available for only certain frequencies. Since the computer industry is in a constant state of change, these crystals frequently show up in the surplus market at very attractive prices; $1.00 to $3.50 is the usual range.

It is important to understand that these crystals are not premium quality parts; the specifications are loose and they vary somewhat among manufacturers. There is no industry standard for size, tolerance or performance. Around 4-6 MHz the following is typical, but there are exceptions!

Frequencies Available (MHz)

3.2768, 3.3300, 3.5795, 3.6000, 3.6864, 4.0000, 4.1943, 4.1970, 4.4336, 4.4400, 4.5000, 4.5500, 4.7500, 4.9152, 5.0000, 5.0688, 5.1850, 5.5850, 5.7143, 5.7600, 5.9904, 6.0000, 6.1440, 6.4000

Frequency Tolerance: 0.005%

Case: HC18/U

Crystal Cut: AT

Maximum Drive Level: 2 mW

The catalog descriptions are usually somewhat cryptic. Typically, there is a column marked "Cap" alongside the frequency. If there is a value in this column, it is a parallel-resonant crystal, and the value is the necessary load capacitance. If the column is blank, the crystal is series resonant. This technique is used by most of the major suppliers; in the surplus and flea-market world you are on your own!

specification, the manufacturing processes, and probably several other factors, contribute to lot-to-lot differences. Several of the crystal groups I tried had questionable parentage, meaning I had no clue to the brand name or purchasing specifications. Fortunately, unless you require a precise filter bandwidth, almost any frequency within this range will make a reasonable filter.

Instant Skirts. . . Just Add Poles!

Commercial filters with 8 crystals (the so-called eight pole filters) are usually adequate for SSB use, but my home-brew filters were not quite as good as their commercial counterparts. This was expected; the crystals were not designed for filter use. On the other hand, adding four more crystals to a home-brew filter is a quick and simple way to make a major improvement in performance for only a few dollars more. In breadboarding, I started with four crystals, then added more just to see the improvement. I quit at 12 crystals, but it would be fun to find out how a home-brew rig with a 24-pole filter would sound. . .

Fig 2 shows the response curve of a 12-crystal, 6-MHz filter, adjusted in one case for minimum ripple, and in the other case for steepest right-hand response. The data cuts off about −36 dB because of the signal leakage limitations of my test equipment. However, the rate of descent is obvious. I could include many other response curves for other crystal frequencies, but the only differences would be the bandwidths. Table 1 summarizes the important tests. A careful study of the data tells you all you need to know.

About CW Filters

CW filters could be built from very low frequency groups following the plan above, but they can be approached another way if you can "borrow" a large sack of any one group of crystals (8 MHz or above is best) for sorting. Again, recalling that a filter is about 1.5 times wider than the spacing of the crystal groups, a 500-Hz CW filter would require two groups with a spacing of 333 Hz (obviously, the tolerances have to be much tighter than those needed in the SSB case). The typical manufacturing tolerance for microprocessor crystals is 0.005%, or 50 parts/million, which can also be expressed as 50 Hz/MHz. Using the 8-MHz example, you would expect a sack of crystals to have a frequency spread of from 400 Hz below nominal to 400 Hz above nominal. The usual "bell-shaped curve" applies here, with most of the crystals clustered around one frequency, but it should not be too hard to sort out two sets with 330-Hz differential spacing from a large group where the total spread is 800 Hz. I haven't built such a filter yet, but I do have the crystals sorted and set aside for it!

The Theoretical Considerations

The above approach to filter design tramples rough-shod on a lot of theory. According to the academics, one should measure the pole/zero spacing of each crystal, perform a long series of complex calculations and develop a specialized circuit before you ever pick up a soldering iron. The approach is appropriate if you needed a precise response curve, or if you were going into production, but it won't get you on the air over a weekend. These simple filters are adequate for their intended purposes, and all the calculations would prove is that you will need more precise crystals if you want a more precise filter.

Terminations and Adjustment

The terminating impedance needs a bit of discussion since it is an important adjustment in the filter. All crystal filters are fussy about their terminations; the name of the game is to get a fairly flat response across the passband.

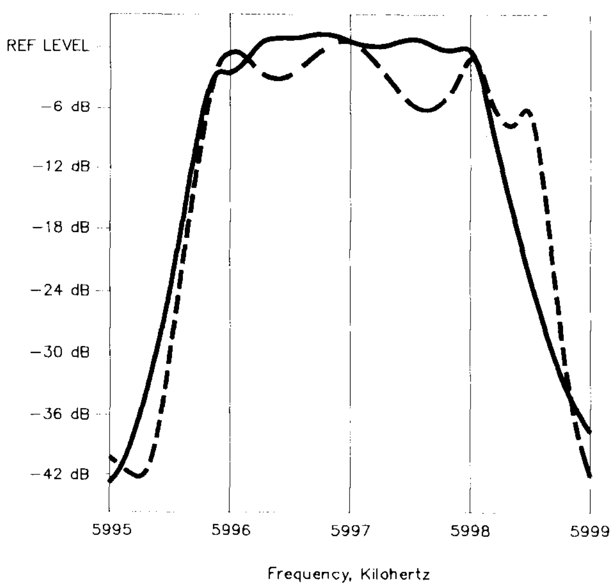


Fig 2—Typical response curve of a 6-MHz 12-crystal filter. One curve represents the flattest passband; the other the steepest right-hand skirt.

Table 1

Parallel Crystals:			
Lowest	3,684,666	5,996,880	7,995,638
Highest:	3,684,714	5,997,014	7,995,786
Average:	3,684,689	5,996,948	7,995,708
High-Low:	48	143	148
Series Crystals:			
Lowest:	3,685,739	5,999,359	7,998,719
Highest:	3,685,777	5,999,405	7,998,828
Average:	3,685,757	5,999,387	7,998,784
High-Low:	38	18	109
Group Spacing (Average):	1,068	2,439	3,076
Filter 6-dB Bandwidth:	1,250	2,500	3,230

Statistics, and resulting filter bandwidths, for three filters built using the parallel/serial crystal method. All frequencies are in hertz. Note that the filters are not quite as wide as those predicted by the *Handbook*. The variations in bandwidth versus group spacing are a function of filter tuning, and of the high-low spacing within a group.

Based on the data, the ideal center frequency for a 2.25-kHz filter would be around 5.4 MHz. The nearest standard values are 5.185 MHz and 5.585 MHz.

If the filter "looks into" the wrong impedance, you can get all kinds of weird response curves. The problem is doubly complicated if you prototype the filter since the driving and terminating circuits in your home-brew transceiver now have to duplicate the optimum values that you determined during bench testing.

Some authors seemingly ignore this impedance matching requirement and terminate filters in a very casual manner; perhaps the authors were lucky, but if you use such circuits you will probably be sorry. The best circuits are those where you have positive control over the filter terminations. As usual, DeMaw and Hayward to the rescue! Chapter 9 of *Solid State Design* shows several excellent circuits where the crystal filter terminations are well defined and controlled. Such circuits are highly recommended as a starting point for home-brew design.

There are two approaches to determining the proper termination resistance. The first is the formal, proper way to do it, where you use a very slow tuning oscillator, counter, and sensitive logarithmic voltmeter to plot the filter curve for various values of termination[7]. The second method is to use your finished rig itself as a test bed. If you go ahead and build your rig (or at least the receiver), and get it debugged, you have all the necessary elements to adjust the filter. All you need to do is to calibrate the S-meter and read it as you tune across a stable test signal.

In either case, the procedure is the same; adjust the variable capacitors for best (steepest) filter slope response, and for minimum passband ripple. Then try different values of termination resistance until you get a fairly flat passband. As a starting value, try 510 ohms. Don't expect perfection; one S unit (6 dB) of dip or ripple will hardly be noticed when you are on the air, and the filters in some high-priced commercial rigs are worse that this.

Sweep Generator to the Rescue

The best tool for filter alignment is a sweep frequency generator, but the usual wideband "TV Sweepers" won't do the job. I built a little sweep/alignment generator based on one of the VFO designs in the *Handbook*, with a varactor diode and ramp generator added to it. The short-term drift is less than 20 Hz, and it sweeps 3 kHz or less without drifting. With this setup, it takes me about five minutes to align a filter.

You can convert any stable VFO to a sweep generator by temporarily adding a varactor diode to the tuned circuit and driving the diode with a ramp generator[8,9]. This simple approach allows you to do swept-filter alignments in your receiver, as described above, by sweeping the local oscillator.

Don't Get Swept Up!

It would appear that a sweep frequency generator would be the perfect tool for filter alignment, but appearances can be deceptive. A rapidly moving swept signal tends to cause a very narrow filter to ring, and consequently the scope display is not a true picture of the filter's response curve. As an example of this (and a warning), I spent several hours fighting a 10-dB dip in one filter's passband as displayed on my sweep generator setup only to find that the filter had less than 2-dB total ripple when I measured it point-by-point with a simple test oscillator! If you have a sweep generator, by all means use it, but be aware of its limitations!

The Armstrong Approach

Not everyone has a scope, sweep generator, counter and log detector; all is not lost. The following assumes that you have a bare minimum of equipment, and are willing to settle for less-than-optimum filter alignment.

You will need some kind of a slow-tuning oscillator with a good dial that can be set (and reset) in 200-Hz increments. Also you need some kind of RF voltmeter, S-meter or equivalent readout device. If your readout does not have a dB scale, you will also need a conversion chart or "semi-log" graph paper.

Start with the oscillator set just below the filter's response starting point. Move upward in 200-Hz steps, graphing the results as you go. Continue until you pass the upper limit of the filter's response. The resulting curve will probably be a mess; find the center of the response curve, peak the capacitors to this frequency, and then rerun the response. The resulting curve will still not be proper, but it should be better. By inspection, you will find that the capacitor adjustments have very little effect on the lower frequency skirt, but have a major effect on the upper skirt and on the ripple in the passband.

Concentrate on getting the upper skirt of the filter smooth and sharp, and then recheck for excessive ripple in the passband. Try to find the combination that approximates the curves of Fig 2, but remember to quit when you get close!

Questions and Answers

The above leaves a lot of questions unanswered. My test equipment is somewhat limited, and I haven't had time to explore all of the possibilities of this technique. However, the following comments, which are in random order, may save you an SASE:

1) For optimum performance, each group of crystals should be from the same manufacturer. Small differences in the manufacturing processes cause parameter spread, which makes for a bumpy response curve.

2) Pay attention to shielding! If you want 50 dB of filtering, it stands to reason that you need at least 50 dB of shielding. Crystal manufacturers' don't put their products in expensive metal cans just to be neat. Try using a metal can, with interstage baffles, and pay careful attention to the signal leads going into and out of the filter if you want the filter to yield its ultimate performance capability. For a simple rig, elaborate shielding is not always necessary, but do pay attention to wire routing!

3) If nothing works, you probably destroyed the crystals—either while you were testing them, or when you installed them in the filter. Unlike most common components, crystals are fragile and require special handling. Never test a microprocessor crystal in a tube-based or other high-powered oscillator! Microprocessor crystal drive limits are low and strict; a high-powered oscillator will eat them for lunch.

There are three common mistakes that are easy to make when installing crystals. Never cut the leads of an unmounted crystal with diagonal cutters unless you cushion the lead with long-nose pliers. The shock generated in a lead wire by diagonal cutters can destroy a crystal! The second mistake is to use too much heat when soldering. Be sure your circuit pad is clean and then use a small iron very quickly! The third mistake is very simple; if you drop a crystal off the bench onto a hard floor, don't even bother to pick it up. . .

4) The circuit that I used is "old faithful," but many other half-lattice circuits and variations have been published. I ran some SPICE® simulations of various designs and found very little difference in performance among them. Adding 300-Ω isolation resistors between sections does seem to smooth the response curve. I considered putting FET isolation amplifiers between each half-lattice section, but never got around to it. If you have the time, try various approaches; if you find a better one, by all means share it with the readers.

5) The following is an attempt to answer the question, "Where can I buy microprocessor crystals?" There are four, and possibly more, sources available, but first some comments:

I have stressed that microprocessor crystals have very loose (or undefined) parameters. If you plan to buy just eight crystals and then build a filter, you will either be overjoyed or disappointed depending on where you got the crystals, and how well they happened to match each other. Building this kind of filter is an ideal project for a group or club purchase where you can re-sort a large bag of crystals into smaller, tighter subgroups.

Secondly, expect to do a bit of hunting for the particular combination you need. You may find parallel crystals from one source and serial crystals from another. Some may be easy to find; others will be harder.

Third, please, repeat, please do not write to me and ask where to buy crystals. Since I work in the electronics industry my sources are not appropriate for retail sales, so I can't help. Instead, try one of the following:

a) The surplus market is loaded with these crystals. Write for catalogs, not only from ham dealers, but also check some of the computer magazines. These sources offer the best prices, but also the most risk since you could wind up with crystals from several manufacturers mixed together. Also, check at hamfests and their related flea markets.

b) Some mail-order catalogs carry these crystals. For example, the Newark catalog has a few, both serial and parallel, but their prices are not competitive[10]. Digi-Key prices are better, and their selection is adequate[11].

c) Some crystal manufacturers and distributors do sell direct, but the minimums range from $25 to more than $100. If you can make a group purchase, this is the way to go. I've found that some manufacturers stock both serial and parallel versions, even though both are not listed in their catalogs. A bit of shopping is in order; prices may vary more than 3:1 among suppliers!

d) The stockrooms of many electronic manufacturing companies are a goldmine for this application. If you are fortunate enough to have the right "connections," you might be able to "borrow" their stock of some specific crystal to sort over a weekend, obviously offering to pay for those you keep (you might try offering a cash deposit to cover their value).

Conclusion

As you can tell from the information presented, I have only begun to explore the potential of this approach. However, rather than wait until I have all the details figured out, I'd rather see others, with better test equipment and fresh ideas, get involved. The main point presented here is that it is possible to build cheap and easy filters. I hope this idea, in turn, leads toward a renewed interest in simple home-brew transceiver designs.

Notes

[1]Hayward, "Designing and Building A Simple Crystal Filter," *QST*, July 1987.

[2]Weaver and Brown, "Crystal Lattice Filters for Transmitting and Receiving," *QST*, June 1951.

[3]Good, "A Crystal Filter for Phone Reception," *QST*, October 1951.

[4]Vester, "Surplus-Crystal High-Frequency Filters for SSB," *QST*, January 1959.

[5]Healey, "High-Frequency Filters for SSB," *QST*, October 1960.

[6]DeMaw, "A Tester for Crystal F, Q and R," *QST*, January 1990.

[7]Zavrel, "A Calibrated S-Meter," *Ham Radio*, January 1986.

[8]DeMaw, "A VFO With Bandspread and Bandset," *QST*, January 1989.

[9]DeMaw, "Tuning Diode Applications and A VVC Tuned 40-M VFO," *QST*, September, 1987.

[10]Newark has branches nationwide. Consult your local telephone directory.

[11]Digi-Key, PO Box 677, Thief River Falls, MN 56701-0677.

The ARRL Handbook and *Solid State Design* are both available from ARRL.

Correspondence

Congrats in Order

"The 'Cloverleaf' Performance-Oriented HF Data Communication System," by Ray Petit, W7GMH, in July 1990 *QEX* is fascinating. W7GMH appears to have more than solved the problem of robust HF digital communications! He is to be congratulated on his work. I would be very interested in experimenting with the Cloverleaf when the transceivers become available, and to get more details of the signals and algorithms when that information becomes available. A bank of twenty Cloverleaf signals is a likely candidate for the ARRL HF Packet Development Program.

The 1-kW solid-state amplifier in the same issue ("A Compact 1-kW 2-50 MHz Solid-State Linear Amplifier," by H.O. Granberg, K7ES/OH2ZE) uses two MRF154 power FETs, according to the cover photo. The article never gives the part number of the FETs. The price list from Richardson Electronics, dated 1 July 1989, lists MRF154s at $942.50 each, or $1885.00 for a matched pair. Once Motorola has cut these prices by 90%, the amplifier will become of more than academic interest. *—Peter Traneus Anderson, 990 Pine Street, Burlington, VT 05401*

LUSAT Frequency Changes

This information is of interest to those stations working towards fully automatic operation with the Microsat satellites.

The AMSAT automatic station in Ottawa (Canada) has detected a constant daily difference in the center downlink frequency between morning and evening passes of LUSAT-19.

The method of operation for the automatic station is to calculate the frequency on which the satellite will be acquired at the start of each pass. The receiver, ICOM 475, is then programmed onto the desired frequency within one second of the calculated time. If the receiver frequency is within about plus or minus 300 Hz, the TAPR PSK modem will achieve frequency lock and begin tracking the Doppler-shifted satellite frequency.

LUSAT-19 has proven to be slightly different in behavior than PACSAT-16 and WEBER-18. LUSAT-19 evening passes (00Z to 05:30Z) have been found to be 1.2 kHz in center frequency lower than the center frequency for morning passes.

After many months of observations, the LUSAT center frequency has been settled as 437.1270.

The TAPR PSK modem has been found to have slightly unequal capability for acquiring frequency lock with its PLL. This has been overcome by applying constant negative offset of 300 Hz with the real-time Microsat scheduler program. The Doppler frequency is taken as calculated in QuikTrak.

A small program modification to the Microsat batch integration programs for the raw QuikTrak tracking data now assigns a lower frequency (1.2 kHz) to evening LUSAT passes. The impact of this has been to permit close to 100% automatic acquisition of the evening LUSAT passes.

Further work in this area is anticipated in late 1991. By then the exact time and frequency of acquisition and loss of signal will be recorded on each pass. This effort is being undertaken to provide eventually for automatic management of the Keplerian element sets to a worst case tolerance of less than plus or minus 5 seconds. For fully automatic and highly autonomous operation of the AMSAT station the current Keplerian sets are frequently of inadequate quality for the needs of the work.

George Roach, VE3BNO, has been highly instrumental in confirming that the station hardware was performing normally. This was not an easy task. Isolation of the change to LUSAT required considerable evening trips to the site, thankfully George had a lot of budget preparation work to keep him busy between passes. . .*—Larry Kayser, VE3PAZ/WA3ZIA, 36 Glebe Ave, Ottawa, ON K1G 3N9 Canada*

Some Feedback

I just received your October '90 issue and feel compelled to write you with some feedback.

I like your magazine; it has a sense of spontaneity similar to the *National Contest Journal* (*NCJ*), to which I subscribe.

The HW-9 article was great. I don't own one, but I might buy one just to make the mods. It seems like a lot of fun.

The Safari-4 article is *very* frustrating. This is *not* N6KR's fault. As editors, you should question this 3-part article—what will it cost?, will someone make a parts kit available?, what skills are needed (I assume high, but how high?), what test gear is needed?, and how will it perform?

I suspect that most who get your magazine aren't outdoor QRP fans, but rather people who would (might) build it regardless. However, what are we getting into?

What I ask is hard to address, I suspect. I am writing software to interface rigs and TUs with the IBM PC—if I asked myself the same questions, the answers might not be too easy either. However, I wanted you to have this feedback, and *please* keep up the good work.*—Gordon Duff, KA2NLM, 6910 Shalimar Way, Fayetteville, NY 13066*

HW-9 Modifications Update

This is an update to my article which appeared in the October 1990 issue of *QEX*, entitled "Modifications and Improvements to the HW-9."

Since I wrote "Modifications and Improvements to the HW-9" last summer, I have found a few improvements to my original suggestions, as well as solving one of the last "problems" (as far as I'm concerned) with the operation of the HW-9.

I had suggested adding a 180-pF capacitor from the collectors of Q405, Q406 to ground to suppress VHF parasitic oscillations. Being a "belt-and-suspender" kind of person, I left the capacitor there even though I found that beefing up the power bypassing (paralleling C445 with another 1-μF ceramic capacitor and a 0.01-μF mylar capacitor) effectively suppressed the parasitic oscillations. The capacitance of Zener diode D405 (hundreds of picofarads) remains, contributing to the suppression of the oscillations.

But the "cost" of leaving the 180-pF capacitor there has reduced output on 10 m. I removed the capacitor and the output power increased while no trace of the parasitic oscillations has returned.

I also mentioned adding a diode in series with the emitter of Q104 to suppress the interaction between the RIT and the transmit return adjustment (R131) from pulling the VFO (see page 5 of the article). My explanation in the article for adding the diode was not quite correct. Nor was the solution of adding just one diode. Actually, I first found that the VFO frequency changed during transmit when the RIT control was turned to minimum resistance (counter-clockwise). This caused the breakdown in the emitter-base junction of Q103, not Q104, to occur, changing the voltage to the variable capacitance diode D118. I then found that the transmit return adjustment interacted with RIT control during receive (due to Q104 breaking down), although this is less of a problem. Therefore, I recommend that a diode also be placed in series with the emitter of Q103. A better solution is to simply replace Q103 *and* Q104 with n-channel MOSFETs. Remember to reset the transmit offset adjustment after any change.

The last of the "problems" with the HW-9 originally manifested itself as a change in pitch of a received signal after transmitting. I found as much as a 300-Hz shift in the pitch of a station I was in contact with immediately after signing over to that station. In addition, the pitch slowly changed to "normal" after a few tens of seconds. Similarly, the transmit frequency drifted during sending. The culprit was a shift/drift in the VFO frequency when going from transmit to receive, and vice-versa.

The shift/drift of the VFO was finally traced to D118, a 1N4002 rectifier, unconventionally used as a voltage variable capacitor for the RIT function. Evidently, Heath chose the 1N4002 for its relatively large change in capacitance over the voltage range of the RIT control (roughly 2 pF from 6.7 to 8.7 volts). However, for what-ever reason, this diode caused the shift/drift. I replaced the 1N4002 with a more conventional MV2101 voltage variable capacitance diode. This eliminated the problem. As a consequence, though, the RIT range has been slightly reduced. To compensate, the resistance of resistor R127 (33 kHz) may be reduced (I didn't bother) to 20 K or so to allow for a greater RIT range. The value of R127 should not be reduced too much; the minimum tuning (bias) voltage thereon should always be sufficient to keep the diode far away from conduction during the oscillation cycle of the VFO. This can be checked by measuring the oscillator output voltage (TP102) while the RIT control is changed. Should the output voltage change at the minimum setting of the RIT control (fully clockwise), the resistance of R127 is too low.

While the shield is off to replace D118, I took the opportunity to drip some wax onto the capacitors to add frigidity and reduce microphonics. Obviously, after any of the above changes, the VFO must be recalibrated per the instruction manual.—*Scott W. McLellan, ND3P, RD 1, Box 149H, Kempton, PA 19529*

QEX—still an experimenters' exchange?

Where have all the thought-provoking ideas gone? As one of the charter subscribers to *QEX*, I've always thought that the key item of distribution was ideas—not plans for replicating someone's nifty project. After all, someone else's project is never exactly what you want—it can always be done better, right? While it's unreasonable to expect everything in every issue, there certainly seems to be a trend towards project duplication. Why all the fuss about foreign parts? What happened to experimenting to find the best substitutions out of the junkbox?

Is there too much emphasis on polished projects? I suspect for every fully developed and completed project, there are many that still need a little work. It would be nice to see innovative circuits that people have gotten to work, even if there are still some problems to be ironed out of the rest of the project. Just be sure to point out what works and doesn't work—sometimes knowing which circuits are flaky from someone who apparently knows what he is doing is useful information. Similarly, I see no problem with publishing a design that you can't explain fully, as long as you don't make something up to confuse people trying to learn. Who knows, maybe some *QEX* reader has the missing piece you need and will help you out.

Finally, here's something to think about. . .

Does one way propagation exist on the HF bands?

One way propagation—signals apparently propagating in only one direction, is a very controversial topic. On CW, it can sometimes be argued that the phenomena can be explained by signals not lining up properly in receiver passbands. Dismissing the phenomena on SSB or RTTY is more difficult—nearly everyone sets their frequencies adequately enough to be noticed under strong signal conditions. Since is unlikely that QRM and poor operators account for all cases, I feel this topic may be worth pursuing—theories should be made to explain what we observe rather than the other way around.

First of all, I feel that ray-tracing arguments don't work. The paths for particles bouncing off objects are reciprocal. It doesn't matter how convoluted the path, there isn't anything that distinguishes direction of travel from the other. Therefore, I looked at devices that aren't reciprocal. What about circulators? An exotic microwave device that is passive, yet creates nonreciprocal paths. It certainly seems like a candidate for developing models about one way propagation.

Remembering the VHF/microwave proceedings, I looked up Tim Pettis' article on Faraday rotation. Why can't Faraday rotation explain one way propagation on HF? After all, it "is inversely proportional to the square of frequency" —its more pronounced as you go lower in frequency[1]. More importantly, it's nonreciprocal—if you have a rotation of 90 degrees one way you get −90 degrees going the other way.

Now that a suitable phenomena has been found, it's just a matter of integrating it with the existing observations and theory. Ideally, a good theory will even suggest experiments to help support itself. Conventional theory seems to suggest that the wave splits into two parts—and that the different paths allow for the waves to cancel upon reaching the receiving antenna. In this case, 90 degrees of Faraday rotation under the right conditions would allow wave cancellation on one end and wave strengthening on the

[1]Tim Pettis, KL7WE, "Spatial Polarization and Faraday Rotation," *Proceedings of the 22nd Conference of the Central States VHF Society,* [Lincoln, NE: ARRL] pp 95-105.

other—one way propagation. However, let's take a different approach. Suppose the Faraday rotation affects both of the split waves equally, or just affects a wave that doesn't really split. One way propagation could be explained if linear antennas were used. In this case, 45 degrees of Faraday rotation could result in a polarization mismatch in one direction but not the other.

That said, I offer the following experiments for those who would like to prove or disprove this explanation. First, assuming it is correct, one way propagation can be eliminated by using a circularly polarized antenna on one or both sides of the path. Secondly, someone using cross polarized antennas can always avoid being in a polarization null by switching to the other antenna. I've used the latter on 20 meters with good success. It may also explain why the best antenna for signal strength isn't always the best for working someone, although the other possible explanations are numerous.

The ideal experiment would be to measure the polarization and signal strength each way, every few minutes. This could be done with each station using a horizontally and a vertically polarized antenna. Transmitting could be done on one antenna, with the receiving station using both antennas and measuring the signals on each antenna. By plotting the horizontal and vertical signal strengths against time for both sites, it should be possible to determine whether any one way propagation occurs on this particular path. Perhaps some existing packet or AMTOR HF stations could be modified to take this data automatically. It wouldn't hurt to know what the path is really like before trying to optimize the hardware to improve data throughput.

Incidentally, if this explanation is correct, "one way" propagation really isn't one way. The signal still gets to the receive antenna on both paths, it's just that the polarity rotation prevents the receivers from picking up the signal equally. Of course, since there is experimental evidence supporting the split waves, the suggested experiment should show actual fading, rather than polarization rotation.—*Zack Lau KH6CP/1, ARRL Lab Engineer*

Bits

Call for Papers:
25th Central States VHF Society Conference

A call for papers has been issued for the 25th Central States VHF Society Conference. VHF Society president Rod Blacksome, KØDAS, and his staff are still looking for a few good people to speak at the conference and/or submit technical papers for the proceedings. Those interested should contact either Rod Blocksome, KØDAS, 690 East View Drive, Robins, IA 52328 (phone 319-393-8022) or Ron Neyens, NØICH, 8616 C Avenue Extension, Marion, IA 52302-9524 (phone 319-377-3207). Deadline for the receipt of papers is May 15, 1991. Please include an ASCII text file on a 5¼" IBM compatible disk along with a hard copy of your paper.

The 25th Central States VHF Society Conference will be held July 25-28, 1991, at the Sheraton Inn in Cedar Rapids, Iowa. An excellent and varied series of activities and technical presentations are planned. It should also be noted that this year marks the 25th anniversary of the CSVHF Society and will be well celebrated by all. With these points in mind, the 1991 CSVHFS Conference promises to be no exception to the high quality and superb technical presentations for which these events are traditionally famous. The conference is open to all members as well as nonmembers. It is a must-attend event for both the inexperienced and experienced VHF/UHF operator. For more information, contact Rod or Ron at the above addresses.

Microwave Update 1991

Microwave Update 1991, sponsored by the North Texas Microwave Society, will be held October 18-20, 1991, in Arlington, Texas. Technical presentations will be held Friday and Saturday, noise figure measurements on Friday night, and a Texas-Style BBQ will be served Saturday night. Special family activities are also planned.

If you're interested in presenting a paper, contact Al Ward, WB5LUA (2375 Forest Grove Estates Road, Allen, TX 75002). He'll give you information on topics and general guidelines for submitting papers (if you just want an Author Package, contact Maty Weinberg at ARRL HQ). September 1, 1991, is the deadline for receipt of papers.

For more information on Microwave Update 1991, please contact Al Ward, WB5LUA, at the above address

Components

By Mark Forbes, KC9C
PO Box, 445
Rocklin, CA 95677

Last month we talked about the basics of memory devices in the continuing discussion of digital electronics. I promised to begin discussion of microprocessors this issue, so here we go.

Microprocessors were invented about 1971. There is some discussion as to whether Intel or Texas Instruments actually "invented" the microprocessor, but the results of a recent lawsuit have made that discussion rather moot, since it concluded that an independent fellow had invented it and deserves royalties. No doubt that will be appealed.

Intel's development was the 4004. It was designed as a general purpose digital chip for implementing a calculator. While it was not very similar to the microprocessors in today's PCs, it was indeed the foundation from which those developed. What made it unique was that it could be programmed to do a number of functions rather than one specific job. Texas Instruments product was the TMS1000, but was only used internally for several years.

From Intel's 4004 evolved the 8008 and then the famous and popular 8080. The 8080 really started the "personal computer" idea when several experimenters built machines using this microprocessor. At approximately the same time, Motorola released their first processor, the 6800. A couple of years later, two spinoff companies improved on these two designs, and personal computers really got started. Zilog released an improvement of the 8080 (and later 8085) called the Z80, which became the backbone of the first popular PCs, the CP/M machines such as the Osborne, Xerox, Morrow and others. Meanwhile Synertec improved the 6800 and called it the 6502. Commodore and Apple jumped on the 6502 bandwagon.

Then in 1981, IBM used a newly developed Intel processor, the 8088, to power the IBM PC. Of course, they stole the market from everyone but Apple, and that 8088 family evolved into the 80286, 80386, and now the 80486. Apple now uses the Motorola 68000 family in their Macintoshes.

So much for a thumbnail history, but what do microprocessors do? If I were to define microprocessors in a single phrase, I'd say that microprocessors are user-definable digital circuit blocks. In other words, a microprocessor can be programmed by the user to perform just about any digital task. What used to require hundreds of digital logic chips can be done with a microprocessor and a half-dozen chips.

Microprocessors can be loosely classified into two areas: General purpose (for use in PCs and other computing devices) and single-chip microcomputer (for control functions, such as in automobiles, microwave ovens, etc). Most familiar is the general-purpose microprocessor. These devices contain most of the functions needed for a computer, but require external devices. Since they are general purpose, they can be used in a wide variety of applications.

A computer system has three main functional blocks: the central processing unit (CPU), memory, and input/output (I/O). The general-purpose microprocessor takes care of the CPU function; calculating, running the programs, managing memory and controlling I/O. While general purpose processors may have some memory, they do not have enough to operate alone. That's why your PC has memory chips (both RAM and ROM) and I/O devices.

The single-chip microcomputers, as the name implies, embody not only the CPU, but RAM and ROM memory, and I/O capability. Since everything is on one chip, very few external parts are needed. If you've ever looked at the computer controller board in your microwave oven or even a modern amateur transceiver, you'll find there aren't too many external parts. The TI TMS1000 was a single-chip microcontroller, and sold more units than all other microprocessors combined, yet most people have never heard of it because it is buried deep within its host product.

This month, I've just given you an overview. Next time we'll look closer at what a microprocessor is and what it does. If you have any specific questions, drop me a line and I'll answer them in the column.

New Battery Technology May Replace NiCds

A new battery technology has been developed with the advantages of NiCds that we all love, but with significantly higher capacity. The new cells are constructed from nickel-metal-hydride and are interchangeable with NiCd batteries, even in their chargers. The new batteries have 80-100% more capacity in the same weight compared to NiCds. They are also apparently more environmentally safe, since they don't contain any heavy metals like the Cadmium in NiCd batteries.

Sanyo developed the technology and has filed for 47 patents. Although production just started about the first of the year, already Motorola has signed up to use the batteries in some of their radio products, and Toshiba will be using the new battery in their laptop computers. Virtually every other battery manufacturer will be producing NiMH batteries within the next year or two.

The ever-popular AA size NiMH battery has a capacity of 1.2 Amp-Hr, compared to 500-600 mA-Hr for typical AA NiCd cells. The only drawback to the new technology is that they cannot provide high-current discharge rates. This limits the batteries to lower- powered applications, but they can be used in lower current applications such as most hand-held radios.

As with any newly developed product, they are rather expensive at the current time. Also, the initial demand is so great they will be difficult to obtain for quite awhile. But, keep your eyes on this one because they will find their way into the HTs of the near future.

Communications Controller

Siemens Components has introduced an enhanced serial communication controller chip, the SAB-82532. This chip can operate to 10 Mbits/s synchronously and to 2 Mbits/s asynchronously. It incorporates two separate serial channels. The chip implements X.25, HDLC, SDLC, and other protocols. It can interface directly to a microprocessor bus, making it very useful.

This chip could find its way into the high-performance packet networks of the near future. It has the capability to control the entire communications channel, with input directly from the microprocessor. The price is right too; in a 68-pin PLCC the chip costs about $20. For more information, contact Siemens Components, Inc, 2191 Laurelwood Rd, M/S M12P011, Santa Clara, CA 95054.

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

CALL FOR PAPERS: COMPUTER NETWORKING CONFERENCE

The deadline for receipt of camera-ready papers for the 10th ARRL Amateur Radio Computer Networking Conference is August 12. Those wishing to submit a paper(s) for this year's conference should contact Lori Weinberg at ARRL, 225 Main St, Newington, CT 06111, phone 203-666-1541, fax 203-665-7531, for paper guidelines and/or an author's package.

Topics will include, but are not limited to, HF packet-radio investigations, network development, digital signal processing, digital speech, hardware, software, protocols, packet-radio services, packet-radio satellites and future systems.

The 10th ARRL Amateur Radio Computer Networking Conference, hosted by the Northern California Packet Association, will be held September 27-29 in the San Francisco area.

HUBMASTER AND INTERMEDIATE-SPEED NETWORK DEVELOPMENT STATUS REPORT

This is an update of some of the recent moderate speed Amateur Radio network developments in Northern California that were previously described in the 9th ARRL Computer Networking Conference proceedings, QEX *and elsewhere.*

Our efforts are not only toward providing faster hardware and link-layer operation, but to more efficiently use Amateur Radio resources to provide amateurs with network layer access at moderate speeds. We anticipate that this will greatly improve current Amateur Radio applications and also provide a platform for completely new ones over wide areas. We hope that many "nondigital" amateurs will find the services such a network can provide to be both interesting and personally useful. We also hope that "nonradio" individuals with computer and networking interests will discover Amateur Radio as a wonderful environment in which to pursue their interests.

Doing all this requires new radio and digital hardware to handle the 0.25-Mbit/s to approximately 2-Mbit/s data streams. Economical operation at this performance level requires line-of-sight radio links. Hardware and new protocols are being devised that effectively utilize the hardware. We have developed an architecture that allows small groups of amateurs to share both the expense and performance in a manner similar to current narrow-band FM repeaters. One of the goals of this moderate speed solution is to solve the hidden transmitter problem without requiring users to suffer the additional complexity and expense of full-duplex radio hardware. We have found AX.25 not to be sufficient for these goals. Perhaps, most importantly, new kinds of coordination and cooperation among amateurs are becoming necessary to successfully implement all of this.

System Components

At present, the intermediate-speed radio hardware includes 900-MHz, 256-kbit/s 10-watt radios with 13-element Yagi antennas and a commercial collinear antenna for use at the hub (see the Hubmaster paper in the *9th ARRL Computer Networking Conference Proceedings* and "Gateway" in March 1990 *QEX* for details). Interface hardware to connect the radio(s) to the digital I/O is also necessary. Some of the 256-kbit/s radios with the previously described 2-Mbit/s microwave link hardware may be used on initial backbones.

The digital interface, MIO (Mundane or Multifunction I/O), can plug into a user's personal computer or be run stand-alone (remotely) as a hubserver and multiport switch/router to the backbone and other clusters. MIO has up to four SCC ports capable of driving a great variety of radio or wireline hardware. It is capable of emulating a conventional AX.25 TNC on one port while serving a 1-Mbit/s microwave backbone connection and a cluster of users running Hubmaster protocol at 256 kbit/s on other ports. A stripped down version of MIO can be run as the interface between the user's computer and 256-kbit/s radio.

In order for all this hardware to operate together, considerable software is necessary. This includes a packet-radio driver for KA9Q NOS, Hubmaster primary and secondary software with a software development tool kit including a debugger. Provision has been made so that software can be written in C++ and put into ROM or loaded into RAM either via over-the-air transfer for the remote case or through shared memory to the PC in the plug-in case. The tool kit will provide basic interface to the Hubmaster polling protocol allowing developers to port their own protocols into the environment.

Radio Hardware

For several months, two 10-watt radios, a prototype and a "pilot" unit, have been complete. A bit error rate (BER) tester was designed and built and BER and multipath distortion measurements have been made on-bench as well as across a variety of real paths from 50 feet to 7 miles using the first two radios. A quantity of commercially fabricated printed circuit boards have been made. One radio using the commercial boards is now virtually complete and working. Major parts for building 12 to 15 units are on hand. The intent is to build approximately six more radios for immediate deployment by N6RCE in Hubmaster protocol development and by N6GN for intercluster backbone communications.

A 13-element, 5-foot boom length Yagi antenna has been designed, built and measured. Performance is within a few tenths of a dB of expectations.

A commercial omnidirectional collinear antenna for hub use is being evaluated.

Interface circuits which provide logic level conversion between the radios and MIO are completed. Additional interfaces with provision for external clock/data encoding and decoding are in progress.

Digital Hardware

The basic MIO PC card has been completed. Ten boards have been commercially fabricated. Two of these have been loaded and the hardware completely debugged. Software development is proceeding. Additional boards are being assembled.

MIO is designed as a full-length IBM® PC plug-in adapter. The card sports a V40 microprocessor running at 8 MHz, 768-kbyte DRAM, up to 256-kbyte EPROM and a pair of Zilog ESCCs. It is designed for 8-MHz (ISA), 8-bit slot PC buses, but will function in any AT slot.

MIO can function in any of three ways; entirely stand-alone, only drawing power from the PC or it can interface with the host processor. Interface is via a shared memory window. The size can be either 8 or 64 kbytes. The shared memory window address and size are set by writing to a control I/O register. Possible base addresses for the window are C8000, CC000, D8000 and DC000.

This control register also allows the host PC to generate interrupts to the V40 and to program the IRQ line used by MIO to interrupt the host.

The V40 is an integrated microcomputer containing an interrupt priortizer, four DMA channels and three timer/counters. The DRAM refresh controller and external bus arbitor for sharing DRAM with the host PC are also utilized.

Software

A software tool kit supports development and debugging with Borland C++ V2.0. A complete operating shell is supplied including linking, relocating and loading of code written on a PC platform.

A run-time kernel provides TOD keeping, interrupt management, DMA management, buffering, protocol manipulation and easy access to the Zilog ESCC (give us your buffer, we'll get it there).

The programmer will be able to write software on a PC platform, compile and link on the PC with Borland TC++, load it onto the adapter and debug it with the remote debugger supplied with TC++. (TC++ has a facility, TDREMOTE, that allows you to use one PC to run the program to be debugged and to use a second PC to run the debugger. The two debuggers communicate via a serial link.) This feature allows you (from the comfort of your home station) to develop software running on hardware located at distant sites such as Hubmaster hubs or hilltop backbones.

System Status, Timetable, Targets and Priorities

All development is currently being performed and funded by Kevin Rowett, N6RCE, and Glenn Elmore, N6GN. Limited personal resources have set the speed of progress. As hardware and protocols come on-line in Northern California, we hope others will step in to develop applications. Such applications are becoming increasingly important to the furtherance of the project. We also hope that hardware will become commercially available for use by others outside of Northern California to start developing and enjoying the benefits of higher speed user network access.

Current emphasis is on getting a minimal hub and backbone operating in time for the next ARRL Computer Networking Conference in September. Deployment of one or two clusters with an interconnecting 256-kbit/s backbone is planned. To accomplish this, Hubmaster protocol refinement and backbone hardware/software development are running in parallel at N6RCE and N6GN, respectively.

Future Targets and Checkpoints

Future directions all rest on the completion of a functioning hub and backbone. Once this is accomplished, the number and variety of options is very large indeed.

Additional radios on other 900-MHz and 1.2-GHz channels will need to be deployed to allow multiple physically close clusters to coexist. N6GN anticipates first modifying 900-MHz radios for 1.2-GHz operation and then building all future radios in the 1.2-GHz band. A minimum-expense combination low-power user radio/interface ("layer-3 TNC") is anticipated with the goal of further reducing the cost of user hardware. N6GN hopes to have an opportunity to apply some ideas to substantially improve the cost/performance of microwave backbone hardware.

In Northern California, we hope to add additional clusters and to extend the backbone so that the San Francisco Bay Area, Santa Rosa, north coast and Sacramento users will all have higher performance connectivity.

Clearly these directions and opportunities require the cooperation and efforts of many people with many different talents. We hope that, as others see functioning hardware and protocols with available development tools, they will join in building and extending the network both in expanse and services.

Since the function of the network is to support new and improved Amateur Radio applications, we hope that many new ones will emerge. Existing PBBS traffic and possibly DXPSN traffic may be supported. We hope that wider area connectivity will evolve. The San Francisco Bay Area to Southern California path is one early possibility. NNTP, "News" protocol, client and servers are already in use at lower speeds and this network should be much more able to support Usenet style distribution of dialog and information to amateurs over a wide area.

Many other applications including amateur fax, voice mail, digital voice and an NBFM-to-IP gateway, to name a few, are being considered.

Where to Find Out More

This is an update on our recent intermediate-speed networking efforts. For more information, we refer you to December 1989 issue of *Ham Radio,* the 1991 edition of *The ARRL Handbook,* "Gateway" in March 1990 *QEX* and particularly to papers in the *8th and 9th ARRL Computer Networking Conference Proceedings* for details of our approach and what we are doing.

—from Glenn Elmore, N6GN, and Kevin Rowett, N6RCE

STS-37 SAREX MISSION OVERVIEW

The recently completed STS-37 mission sported an all-ham crew consisting of Mission Commander Steve Nagel, N5RAW, Pilot Ken Cameron, KB5AWP, Mission Specialists Jay Apt, N5QWL, Linda Godwin, N5RAX, and Jerry Ross, N5SCW.

SAREX equipment on the flight included a 2-meter, 2.3-watt Motorola radio, Robot 1200C SSTV converter, Heath HK-21 TNC, a 70-cm FSTV receiver, a video camera and a monitor/VCR. Planned operations included voice, packet-radio robot, downlinking orbiter video via SSTV and uplinking FSTV video to the orbiter. During sleep periods and when no other SAREX activities were scheduled, the equipment was left in packet-radio robot mode.

—from AMSAT

TAPR ANNUAL MEETING HIGHLIGHTS

The annual meeting of Tucson Amateur Packet Radio, Inc (TAPR) was held on March 2 in Tucson. The highlights of the meeting were reported by our roving Gateway reporter, Jon Bloom, KE3Z.

Al Danis, W6HGF, who works in the Department of Defense, made some brief remarks regarding the use of TNCs by the Desert Storm troops. This case of Amateur Radio technology being used in the military came as a surprise to many of the listeners. Danis reported that several thousand TNCs were in use, mostly for logistical purposes. During fast movements, frontline units use TNCs because they can't stop to set up their full-fledged communications terminals. The 18th Airborne, for example, uses TNCs extensively. AX.25 is a *de facto* standard for this use and TNC usage is now being extended into the tactical networks.

Lyle Johnson, WA7GXD, displayed a beta-test TAPR digital signal processing (DSP) board. The board plugs into an IBM PC and will provide for medium-speed (up to 9600 or 19,200 bit/s) packet-radio operation using any needed modem standard. It also will be usable for other modes, such as the various slow video modes (WEFAX, SSTV).

Paul Newland, AD7I, reported on his METCON project. This project, which will be kitted by TAPR, consists of a small, inexpensive microcomputer on a PC board which is used for telemetry and telecommand. The microcomputer board controls relays and has digital input and output signals. It can be attached to auxiliary boards that measure voltages or temperatures.

Phil Karn, KA9Q, showed a PC plug-in card that can be used for high-speed packet. The card, manufactured by Hasenfratz & Associates of Thousand Oaks, California, can transfer packet data from an IBM PC/AT or compatible computer at up to 4 Mbits/s.

Karn reported on an authentication technique he is using via packet radio to log onto his UNIX system. The system uses a one-way function in a manner that lets him give a unique password to the system each time he logs on. This identifies him as the user with a high degree of certainty.

A special award was presented by TAPR to Harold Price, NK6K, for his efforts in the development of packet radio. Price has been involved in several critical packet projects from the TAPR TNC I to the packet-radio MicroSats.

Dewayne Hendricks, WA8DZP, reported on investigations he has been making into Part 15 spread-spectrum devices. These devices implement wireless data communications in a local area. Operating in the same 900-MHz band where amateur operations are permitted, some of these devices may be suitable for conversion to amateur use. A report should be available at the 10th Computer Networking Conference in the San Francisco bay area.

Hendricks also reported on the preparations for that conference. The sponsors are soliciting ideas on how to better organize the presentation of papers. Many attendees of recent networking conferences felt that insufficient time was available for some of the more interesting papers.

There was much discussion of the recent FCC citations of PBBSs. The TAPR board of directors has appointed a committee to consider what TAPR's position should be.

The results of TAPR elections were announced. The new TAPR officers are President and new director Bob Nielsen, W6SWE, Vice President Harold Price, NK6K, and Secretary-Treasurer and new director Greg Jones, WD5IVD, and new director Jerry Crawford, K7UPJ.

APPLE-AMATEUR RADIO USERS GROUP FORMED

In August 1990 Phil Endres, N9AVF, ran a survey to locate amateurs on packet radio using Apple computers. The results of that survey (nearly 70 responses) convinced Drew Hilliard, N6SIJ, to organize interested Apple users/Amateur Radio operators into an Apple Corporation recognized "users group," to be known as the "Apple-Amateur Radio Users Group" (AARUG). The purpose of AARUG is to provide support for amateurs using Apple II and Macintosh computers in any area of Amateur Radio, including packet-radio communications, satellite orbital and propagation predictions, contest logging, QSL preparation, etc.

AARUG will operate under two basic principles: (1) AARUG will operate entirely via packet radio. Information will be compiled and monthly newsletters distributed on packet radio (hard copy newsletters will not be prepared or distributed). So as to gain maximum visibility to potential members, AARUG messages and bulletins will use "AppleNet" as the subject line. The only requirement for membership is possession of a valid amateur license. No dues or membership fees will be requestor. Software known as "shareware," "freeware," or "public domain" will have distribution costs (blank media and postage) borne by the requester. (2) Since AARUG/AppleNet will operate on amateur frequencies, no advertisements for commercial interests or enterprises will be permitted. If AARUG/AppleNet sounds like something you'd like to participate in, please send a response to N6SIJ@AA6QN. As part of your response, please indicate computer type, amount of memory, number of drives and communication software used.

—from Drew Hilliard, N6SIJ, via John Seney, WD1V

HAMS' "KUWAIT CONNECTION" REVEALED

Amateurs everywhere now can know the full story of a Kuwaiti amateur who was on the air during the occupation of his country, sending AMTOR messages to the outside world (Gateway in April 1991 *QEX*). On March 21, Abdul Jabbar Marafi, 9K2DZ, gave the ARRL his consent to reveal his identity and to publish his story; a major article about the 'Kuwait Connection' will appear in May *QST*. What follows is a greatly abbreviated version of that article.

Usually signing on his other call signs, A92ET, or no call sign at all, 9K2DZ almost daily reported from Kuwait. At first the traffic was Health and Welfare messages, and Marafi turned immediately to his Amateur Radio friends, on the APLink network.

"AMTOR sounds like a cricket," Frank Moore, WA1URA, told the U.S. Department of State when he began faxing them copies of Marafi's missives from Kuwait City. "The transmitters pulsate on and off and are much more difficult to locate with a direction finder. It also means that one receives almost perfect copy," Moore said.

Marafi's APLink colleagues were quick to recognize the need to protect 9K2DZ by keeping the traffic quiet (at one time Iraqi troops were living in the house next to Marafi's, Clark Constant, W9CD, said). The traffic was not a total secret; as hams happened upon the APLink frequencies, bits and pieces of messages with Arabic names would appear on their screens.

Meanwhile, aboard the *USS John F. Kennedy* in the Red Sea, Navy First Class Petty Officer Scott Ward, N5DST, had long been using AMTOR and APLink to handle messages for the more than 5,000 Navy personnel aboard. Following the invasion of Kuwait, Ward's official Navy channels were given over entirely to traffic more important than Health and Welfare messages, leaving Amateur Radio and APLink as his only medium for messages home.

It wasn't long before N5DST/MM ran across the Kuwait traffic. Ward was immediately invited to join the effort because his location seemed perfect for a short hop to Kuwait. "I always would forward the Kuwait messages to WA1URA, their primary destination," Ward said, "knowing they would get to State if necessary. Then they would go up the chain of command aboard ship, through officers in touch with Central Command in Riyadh."

Bob Foster, WB7QWG/9, had been handling *Kennedy* messages for Ward long before the ship's deployment to Operation Desert Shield. "At first we discouraged Abdul from transmitting," Foster says. "Although we all were very curious, we also feared for his safety."

For N5DST/MM, the spectre of shipboard radio silence always loomed, as when YI1BGD, at the Radio Club of Baghdad, briefly accessed Ward's BBS and also attempted to access those of APLink operators TG9VT and OD5NG.

By now Abdul Jabbar Marafi was beginning to feel the strain, as were those in the network. "Abdul, W9CD, and I became concerned," Troost says, "that the Iraqis might be listening in and that we were giving away information they could use to repress Kuwaiti nationals, not only in Kuwait, but also in the USA, so most of that [Health and Welfare traffic] was stopped."

Other Kuwaitis around the world were picking up the slack as they became active on AMTOR and joined the system to handle "mail" between each other. At least two 9K2 stations were active from Switzerland and additional links appeared in Muscat and the United Arab Emirates, to where many Kuwaitis had fled.

It also was feared that Marafi had been compromised by his "discovery" by YI1BGD in Baghdad. But "there is no reason to believe," Moore decided, "that the YI station compromised our friend. We have excluded him from our computers. We also ask 'special questions' to our friend so he can give us a sign if he were in trouble and that has not happened."

"It was about October," Troost says, "that the Iraqis started to pay more attention to the old Amateur Radio operators in Kuwait and interrogated all the hams. [Following the Iraqis' visit]. . . all Marafi's operating was done behind the false wall in the basement." Today, 9K2DZ is active almost daily on APLink; now doing what hams ordinarily do. . . talking with his friends.

Abdul Jabbar Marafi, 9K2DZ, is part of an extended merchant family known as the "Sultans," according to the Wall Street Journal. He is described by family members as a businessman. He has been licensed since 1973 and also holds the call signs A92ET and SU1DZ in Egypt, where he owns a home. Marafi, a graduate of Cairo University, worked in the Kuwaiti Cable and Wireless division from 1953 to 1956, then in the Ministry of Telecommunications until 1978.

Marafi provided assistance right to the end, Moore said. The CBS News crew that beat the first contingent of US Marines to Kuwait City was grateful when Marafi arranged to find an apartment for crew members.

Following the liberation of Kuwait, portions of this story were released to the press, including the ARRL, on March 4. News stories appeared heeding the embargo on identifying 9K2DZ or mentioning Amateur Radio. On March 21, Marafi himself lifted that embargo at the League's request.

—from *The ARRL Letter*

LATEST PACKET-RADIO SOFTWARE RELEASES

The following packet-radio software was released during the past month:

G8PBQ node software, version 4.03
N2GTE Packet Message Switch (GTEPMS), version 1.2
PE1CHL-NET, version 910123
PRMBS Roserver, version 1.52
TAPR TNC 2 firmware, version 1.1.7a
WØRLI MailBox, version 12.00

All of this software is available for downloading from Compuserve's HamNet. All or some may be available on various ham-radio-oriented telephone BBSs. Some are available on disk (or ROM, in the case of TNC 2 firmware) from Tucson Amateur Packet Radio, Inc (TAPR), PO Box 12925, Tucson, AZ 85732-2925, phone 602-749-9479 (write or call concerning availability).

THE 4X1RU HF-VHF GATEWAY/PBBS STATISTICS

The nerve center of the 4XNET Packet Radio system is at the station of 4X1RU. For years, Jim has been providing the link for packet-radio messages not only between our VHF network and the rest of the world, but between the areas of the world for which Israel is a geographical crossroads.

One day, one of our scribes will visit Jim's station to give you a report on how he does it. But, in the meantime, from the monthly statistics that 4X1RU makes available, we have been able to glean information that casts light on the scope of Jim's operation.

It appears that 4X1RU has two packet-radio stations going simultaneously, one on HF, 20 meters, and the other on VHF, 145.675 MHz, which are tied into the same computer that does all the marvelous work. A station connecting to 4X1RU on one of those ports may download mail, bulletins or files (either text or public-domain software) from the gateway's data base. Conversely, one may leave mail or bulletins for forwarding in Israel alone or anywhere in the world if he chooses and may also upload files on to the PBBS for public use.

Within Israel, most of us are aware of 4X1RU being the PBBS that serves the center of the country (4Z4SV serves Jerusalem and the south, while 4X4HF serves Haifa and the north). 4X1RU, being linked with the rest of the world, provides us with the steady stream of bulletins covering all kinds of topics including DX, technical topics, satellite news and the almost countless topics that interest hams.

Mail on one of the local VHF PBBSs that is addressed to points abroad will reach 4X1RU, where it will be forwarded via 4X1RU's HF port onto its destination. Mail on the international system addressed to Israel will reach 4X1RU from where it is transferred to our VHF net, reaching the local PBBS of the addressee. Most of this is achieved automatically, but in cases where the addressing is faulty or incomplete, Jim has to intervene manually and make the necessary corrections for the smooth passage of the mail.

Now with the explanation out of the way, let's go to the statistics. The computer running the system records every operation on the system, so that Jim has a complete record of every operation for each month. The following figures are for the month of February:

The VHF port had 87 stations connected to it for a total of 151 hours (that is over 5½ hours solid per day) and processed 5466 messages. That includes bulletins, NTS traffic, mail and texts.

Connect time on the HF port was just over 60 hours during February with 5492 messages processed and 2872 bulletins forwarded. Its connects were with 11 European countries and the US. Attesting to the good shape of the system, both the HF and VHF ports were on-line every minute of the month with no down time!

Summing up, those are just some of the dry statistics. But they point to a superb operation by 4X1RU and we salute Jim for his tireless efforts that benefit us all.

—from *HaGAL International*

GATEWAY CONTRIBUTIONS

Submissions for publication in Gateway are welcome. You may submit material via the US mail to 75 Kreger Dr, Wolcott, CT 06716, or electronically, via CompuServe to user ID 70645,247, or via Internet to horzepa@gdc.portal.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*.