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on the cover: John Webster, K1FWE (bottom left), and Doug Grant, K1DG (bottom right), operating during the 1986 K1AR Field Day effort. Both are SFRCC members. Top photo: Marty Durham, NB1H, "fixing things" on N1AU's tower.

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Novice enhancement and the future of Amateur Radio

Since going into effect last spring, Novice enhancement hasn’t caused any great upsurge in the Amateur ranks. Whether it should be considered a modest success or a complete failure seems to depend on who’s talking.

If the criterion is merely the decrease in Novice licensees compared with a year ago, Novice enhancement has failed. To me that’s a very shallow, superficial interpretation resulting from a cursory reading of the numbers and a misplaced belief that enhancement addressed a basic problem of Amateur Radio rather than an ancillary one.

Discounting the big Novice jump in April and May of 1987, when a lot of newcomers rushed in to take the Novice exam before it was expanded to cover new privileges, the Novice population hasn’t shown a significant change in the past year. There are, perhaps, many reasons for that. How many of last year’s new licensees didn’t even pause at Novice but moved up immediately? How many of them had been putting off becoming Amateurs and were stimulated to take the Novice exam before it got tougher? These and many other questions should be answered before the results of Novice enhancement can be properly assessed.

Some critics now say the problem with Novice enhancement is that it didn’t go far enough, and what’s really needed is to do away with the CW “boogie man”. Though I agree that the CW requirement has long intimidated — and will continue to intimidate — a vocal minority of prospective Amateurs, I also firmly believe that any attack on the CW issue, no matter what its outcome, will have no more effect on the long or short term Amateur growth problem than Novice enhancement did!

Whatever your feelings, neither Novice enhancement nor a no-code license addresses the basic problem. The problem isn’t our product, but its marketing. Amateur Radio is a great product, but if our potential customers don’t know the product exists, where to find it, or appreciate its many benefits, they aren’t going to buy!

Intelligent marketing is based on market analysis. Manufacturers who don’t understand this are doomed to slow growth and/or stagnation at best, and the bankruptcy court at worst. Analyses of recent licensees by the FCC and the VEs who are actually bringing the newcomers on board agree that the average new Amateur is an older, well-established adult. Our marketing effort has been aimed at youngsters, so it seems likely we’ve been targeting the wrong market. The ARRL seems to feel this way, and is now experimenting with a pilot program that encourages older residents of the Tampa/St. Petersburg, Florida area to become hams.

Before investing any great amount of money and effort in new sales pitches or product revisions, I suggest we put some of that money into a professional market study. This study should be directed primarily at those Amateurs who’ve joined us in the past 10 to 12 years and (when possible) those who’ve dropped out. It should include questions on how and why respondents became Amateurs, what they felt had helped, or what had hindered their developing Amateur Radio interest. When the study’s results are analyzed, the most cost-effective marketing strategy may become clear.

The ARRL and the Amateur Radio Industry Group have the capabilities for such a study. The two worked together well on the Archie’s Ham Radio Adventure comic book project, and might be willing to work together on this one. In the meantime, however, I feel that any further tinkering with the product isn’t going to solve the basic problems, only complicate them.

Joe Schroeder, W9JUV

This editorial is one person’s opinion about Novice enhancement and does not necessarily represent the views of ham radio. Ed.
Affordable DX-ing!

TS-140S
HF transceiver with general coverage receiver.
Compact, easy-to-use, full of operating enhancements, and feature packed. These words describe the new TS-140S HF transceiver. Setting the pace once again, Kenwood introduces new innovations in the world of “look-alike” transceivers!

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- All modes built-in. LSB, USB, CW, FM and AM.
- Superior receiver dynamic range Kenwood DynaMix™ high sensitivity direct mixing system ensures true 102 dB receiver dynamic range.

- New Feature! Programmable band marker. Useful for staying within the limits of your ham license. For contesters, program in the suggested frequencies to prevent QRM to non-participants.
- Famous Kenwood interference reducing circuits. IF shift, dual noise blankers, RIT, RF attenuator, selectable AGC, and FM squelch.

- M. CH/VFO CH sub-dial. 10 kHz step tuning for quick OSY at VFO mode, and UP/DOWN memory channel for easy operation.
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- RF power output control.
- AMTOR/PACKET compatible!
- Built-in VOX circuit.
- MC-43S UP/DOWN mic. included.

Optional Accessories:
- AT-130 compact antenna tuner • AT-250 automatic antenna tuner • HS-5/HS-6/HS-7 head-phones • IF-232C/IF-100 computer interface • MA-5/V-1 HF mobile antenna (5 bands) • MB-430 mobile bracket • MC-43S extra UP/DOWN hand mic • MC-56 (6-pin) goose neck mobile mic • MC-60A/MC-80/MC-85 desk mics. • PG-2S extra DC cable • PS-430 power supply • SP-40/SP-50B mobile speakers • SP-430 external speakers • SW-100A/SW-200A/SW-2000 SWR power meters • TL-922A 2 KW PEP linear amplifier (not for CW QSK) • TU-8 CTCSS tone unit • YG-455C-1 500 Hz deluxe CW filter. YK-455C-1 New 500 Hz CW filter.

Complete service manuals are available for all Kenwood transceivers and most accessories. Specifications, features, and prices are subject to change without notice or obligation.
tips for construction projects

Dear HR:

Boy, am I glad that I started building projects before reading Paul A. Johnson’s article in your March 1988 issue. I’m sure Mr. Johnson’s piece would have scared me away. Here are some suggestions for any of your readers who might be interested in project construction:

1. Don’t start with anything rf. Receivers, transmitters, tuners, lines are all difficult and require a lot of adjustment once they’re assembled. As I recently discovered, even a simple dummy load isn’t simple. Don’t start with a high-voltage or high-current power supply either. Anything over about 50 volts or 5 amps requires extra care and construction technique. Start with something like a 12 volt 3 amp power supply to run your HT in the house. How about a digital clock? Use a microcontroller and make a keyer. These suggested projects may not sound very exciting technically, but you’ll find that project construction is often more mechanical than electrical.

2. You don’t have to build most projects in metal boxes. For non-rf projects, plastic is fine. It’s inexpensive, easy to work, and doesn’t have to be painted. Jameco, Digi-Key, and your local Radio Shack all offer a selection of plastic enclosures. Stick with plastic and you won’t need a drill press; an electric hand drill is fine. You won’t need expensive and dangerous hack-saws, sabre saws, circular saws, or fly cutters either. Holes larger than your drill can make or odd-shaped openings can be cut quickly and easily with a reamer or some cheap files. Filing out openings in metal is an arduous task, but in plastic even cheap files cut quickly. Here’s a tip: to drill a nice hole in plastic, start with your smallest bit and work up to the final size using every bit between them in your index. Hot-melt glue guns work on plastic. A cheap pop riveter is another handy tool. If you do need metal, look for a prefab cabinet that will fit your project. Prefabs may seem expensive, but they’re a lot easier and you won’t need a lot of tools and equipment. Bud, LMB, and others offer an excellent assortment of cabinets ready to house most projects.

3. Plan! Document! Much of the work for my projects is done on paper. Start with a good schematic. If you’re using any integrated circuits, mark the pin numbers on your schematic. Draw pin diagrams of other parts like transistors next to the part on the schematic. Assign part numbers. Sketch how the project will be assembled, the layout of parts on circuit boards, and the chassis wiring. Then make a from-to wiring list. With all of this planning, your project will be a snap to build and will work the first time. If it doesn’t, all the documentation will help you find the problem fast. By the way, keep all of this paper so that if your project ever breaks, you or a friend ever want to build another, it’ll be easy. As you correct bugs or add modifications, document the changes.

4. Take your time. Measure twice, cut once. Make test fits as you move along. Check each electrical connection with an ohmmeter. Try to make every solder joint perfect. Use cable ties or lacing tape to form cable bundles. If you have extensive chassis wiring, use wire marker labels. Use heat-shrink tubing and cable clamps as necessary. In short, try to make each project a show piece inside and out.

While Mr. Johnson’s work certainly looks very nice, project construction does not have to be as difficult or as complicated as he makes it sound. You don’t have to be a machinist, and you really don’t need a lot of expensive tools either. By avoiding complicated projects (especially rf ones) at first, using plastic boxes whenever possible, planning carefully, and working slowly, anyone can enjoy building perfect construction projects. I know I sure do!

Chuck Gollnick, KA7QEN
Ames, Iowa 50010-1363

no contest

Dear HR:

In his article in the November 1987 issue, Bill Orr, W6SAI, says he can attest to the fact that allocation of even a small segment of the 30-meter band to SSB operation would be of great benefit to Amateur Radio. Instead of a bland statement, would he be a bit more explicit?

Surely he must be aware that another well-known author sparked a similar controversy in the columns of the RSGB’s RadComm magazine, and the consensus of opinion was against any change in the IARU’s recommendations. Could there be a conspiracy of authors on this subject?

I have used the 30-meter band almost since it’s inception, and the greatest problem is finding a space to work without causing QRM to priority users. Like many others, I have worked over 100 countries; the DX is there and occupancy will surely improve as we advance into the new cycle.

Would the SSB fraternity be as mindful of our non-priority status as the CWers have been? I doubt it, and it would not be long before we lost the band altogether.

I wholeheartedly agree with his 18 MHz sentiments and it would be a great shot in the arm to have the W/Ks on the band, but please no contests. As my friend SM3CIQ/U1F says; rather RTTY QRM than contests.

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design program for the grounded-grid 3-500Z

This program gives "no-compromise" answers

There are probably as many amplifiers in existence that use the 3-500Z as there are with any other power amplifier tube. The 3-500Z is an excellent tube with a well-deserved reputation for high power-handling capabilities and reasonable cost. There are no doubt hams with one or two spare 3-500Zs in their shacks who are thinking of building their own amplifiers. But of course, it's one thing to copy the design of another amplifier and a different matter to design one's own from scratch.

Circuit variations for amplifier designs are available from other sources and I will not discuss them here. This article covers only one mode of operation for the 3-500Z — grounded-grid class AB2 operation — probably the most prevalent use of the tube. I will discuss virtually all possible combinations of drive power, load resistance, drive impedance, plate voltage, and bias requirements for grounded-grid, linear operation. I have included a program which allows you to accommodate any set of normal operating conditions that can be realized on the constant-current curves. Figure 1 is the program listing.

This program's answers are probably a bit more precise than ones obtained with the "Tube Performance Computer". You might assume that linear divisions exist between constant-current curves on the tube charts; they don't. This is not a serious problem, as variations from one tube to another will usually be greater than those differences. It is important to remember that using the hand-calculated methods consumes much paper, time, and nervous energy. My program lets you change any of the input parameters and see the differences for each proposed operating condition in a few seconds, as compared with fifteen minutes or a half-hour required for hand-calculated answers — and it doesn't make mistakes!

The program has only 170 lines to enter; it runs completely within a few seconds after you enter the last input. (The answers appear in about 3 seconds on the Tandy-2000, and in 8 to 10 seconds on an early IBM PC.) When this program is compiled, answers appear in under a second. The program is very densely packed with numbers and equations — I know of no other way to define every current and voltage (including fractional values) that can be found on the tube chart, and still use fewer than 7 kilobytes of computer RAM. The labor of typing the program pays for itself many times over as it saves hours of effort during a design routine.

how to use the program

Figure 2 is a reproduction of the "EIMAC 3-500Z Typical Constant Current Characteristics" curves for grounded-grid, class AB2 operation. The operational area of the program (crosshatching) is superimposed on the curves. Stay inside the "box"; it includes every permissible or useful operating point. You decide on the placement of the operating (or load) line. It is drawn between two points, labeled "ip", and "Q". The first defines the maximum peak instantaneous plate current as well as the minimum plate voltage. The second defines the quiescent (no drive) value of the plate current, and occurs at exactly the plate-supply voltage. The chart also tells you the quiescent plate dissipation. This is not printed in the program output, but can be calculated by multiplying the rest-plate current by the plate supply voltage. The program requests: "Enter Plate Supply Voltage, Eb".

By W.J. Byron, W7DHD, 240 Canyon Drive, P.O. Box 2789, Sedona, Arizona 86336
fig. 1. The 3-500Z Design Program Listing.

10 PRINT "3-500Z Grounded-Grid Characteristics"
20 PRINT "Subroutines Copyrighted 1967, W.J. Byron"
30 PRINT "All rights reserved"
40 INPUT "Enter Plate Supply Voltage, Ebb":E3
50 PRINT "EXCESSIVE PLATE VOLTAGE!":GOTO 190
60 INPUT "Enter Peak Plate Current, Ip":I1
70 IF I1<1.6 THEN PRINT "EXCESSIVE PEAK CURRENT!":GOTO 290
80 INPUT "Enter Minimum Plate Voltage, Emin":E4
90 IF E4<150 THEN PRINT "HIGH GRID CURRENT AREA!":GOTO 210
100 IF I1<1.4 AND E4<250 OR E4>4500 THEN PRINT "OUT OF BOUNDS":GOTO 210
110 IF I1>4.4 AND E4<1450 OR E4<3000 OR I1=I1.4 AND E4<1500 THEN PRINT "OUT OF BOUNDS":GOTO 210
120 INPUT "Enter Cathode Bias Voltage (Zener)":E2
130 IF E2<0 THEN PRINT "Negative Cathode Bias Not Permitted!":GOTO 230
140 PRINT "CALCULATING..."
150 PRINT "RADIO FREQUENCY LINEAR AMPLIFIER"
160 PRINT "Cathode Drive, Class AB2"
170 PRINT "--------------------------------------------------------------------"
180 PRINT "--------------------------------------------------------------------"
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(3) Adapter. Connects 1/4" plug to 7/8" jack. #274-325 . 1.49
(4) HT Special Adapter. #274-381 . 1.79

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For RG-58/RG-62 Cables. #278-104 . 2.59
(2) 1084 Female. #278-105 . 1.39
(3) UG-914 Adapter. #278-115 . 2.19
(4) UG-306 Right-Angle Adapter. Male to Female. #278-116 . 4.29

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Voice Actuated, 49-MHz FM Operation
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(3) Regulated 2.5A, 13.8VDC Power Supply. UL listed. #22-120 . 39.95

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Radio Shack Has A Full Line of Ham Accessories For Every Budget
(1) Outdoor RF Seal. #278-1645 . 2.49
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(3) RG-59/8M Reducer. #278-204 . 2.99
(4) RG-58 Reducer. #278-206 . 2.99
(5) PL-258 Coupler. #278-1669 . 1.49
(6) M-358 T-Adapter. #278-1669 . 2.99

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(1) Premium-quality coax cable.
(2) Outdoor RF Seal. #278-1645 . 2.49
(3) PL-259 Plug. #278-205 . 2.99
(4) RG-59/8M Reducer. #278-204 . 2.99
(5) RG-58 Reducer. #278-206 . 2.99
(6) PL-258 Coupler. #278-1669 . 1.49
(7) M-358 T-Adapter. #278-1669 . 2.99

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Now you can get into this exciting mode with our all in one box TC70-1 70cm ATV Transceiver at the 1988 Spring Sale reduced price from $299 to only $269 delivered.

TC70-1 FEATURES:
- Sensitive UHF GaAsfet tuneable downconverter for receiving
- Two frequency 1 watt p.e.p. transmitter, 1 crystal included
- Crystal locked 4.5 MHz broadcast standard sound subcarrier
- 10 pin VHS color camera and RCA phono jack video inputs
- PTL (push to look) T/R switching
- Transmit video monitor outputs to camera and phono jack
- Small attractive shielded cabinet - 7 x 7 x 2.5"
- Requires 13.8vdc @ 500 ma. + color camera current

Just plug in your camera or VCR composite video and audio, 70cm antenna, 12 to 14 vdc, and you are ready to transmit live action color or black and white pictures and sound to other amateurs. Sensitive downconverter tunes whole 420-450 MHz band down to channel 3. Specify 439.25, 434.0, or 426.25 MHz transmit frequency. Extra transmit crystal add $15.

*Transmitting equipment sold only to licensed radio amateurs verified in the Callbook for legal purposes. If newly licensed or upgraded, send copy of license. Receiving downconverters available to all starting at $39 (TVC-2).

WHAT ELSE DOES IT TAKE TO GET ON ATV?
Any Tech class or higher amateur can get on ATV. If you have a camera you used with a VCR or SSTV & a TV set, your cost will just be the TC70 and antenna system. If you are working the AMSAT satellites you can use the same 70cm antennas on ATV.

DX with TC70-1s and KLM 440-27 antennas line of sight and snow free is about 22 miles, 7 miles with the 440-6 normally used for portable uses like parades, races, search & rescue, damage assessment, etc. For greater DX or punching thru obstacles: 15 watt p.e.p. Mirage D15N or 50 watt p.e.p. D24N or D1010N-ATV.

The TC70-1 has full bandwidth for color, sound, like broadcast. You can show the shack, home video tapes, computer programs, repeat SSTV, weather radar, or even Space Shuttle video if you have a home satellite receiver. See the ARRL Handbook chapt. 20 & 7 for more info & Repeater Directory for local ATV repeaters.

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12 June 1988
fig. 2. The 3-500Z Constant Current Characteristics Curves with the program boundaries marked. Stay inside the cross-hatched area. The load line is for the solution in fig. 4.

fig. 3. Diagram of the possible movement of the Operating Line. For use with table 1.

"Enter Peak Plate Current, \( i_p \)", "Enter Minimum Plate Voltage, \( E_{\text{min}} \)", and "Enter Cathode Bias Voltage (Zener)".

These four inputs define both ends of the operating line (which you may already have drawn on the curves), and are sufficient to determine all operating parameters. It isn’t actually necessary to draw the line, but it may help to visualize it; one appears in the figure to demonstrate the method.

The main program starts immediately after the last input and calculates a total of fifteen lines of data. Two are input repetitions (Plate Supply Voltage and the Zener Bias); the rest are results of internal calculation by the program. These are the numbers you want. The inputs are repeated in the line just below the heading for the program output as a record of what has been entered. Use them as starting points for any changes you want to make. After the listing there is a question: "Do you wish to change an input—Y or N?" If you enter "Y", the program in turn will ask you, "Which one?" in a menu, and you can change any one of the four inputs until the outputs are to your liking. Any other entry, including "N", will abort the program and you will have to "RUN" again. You can then use the "immediate mode" of BASIC to calculate, for instance, the quiescent plate dissipation (which as a rule of thumb should be somewhere between 30 and 40 percent of the maximum plate dissipation — 500 watts for this tube).

There are some constraints imposed on the initial inputs (see lines 190 through 240). These conditions do not exist at the "Do you wish to change an input?" prompt. Because you are typing the program yourself, you decide whether or not to include them (use your own good judgement). To do so, just duplicate the conditions stated in the lines identified above. The constraints result partly from the maximum values permitted by the manufacturer and partly from my work to limit both maximum grid dissipation and amplifier distortion. The program isn’t valid outside these limits; the manufacturer’s allowable values are the principal reasons for the lower limit of 250 volts for the minimum plate voltage.

Certain changes will occur when the positions of

<table>
<thead>
<tr>
<th>Movement in direction</th>
<th>Results of movement</th>
</tr>
</thead>
<tbody>
<tr>
<td>A:</td>
<td>Higher grid current</td>
</tr>
<tr>
<td></td>
<td>Higher plate current</td>
</tr>
<tr>
<td></td>
<td>Higher input and output</td>
</tr>
<tr>
<td></td>
<td>Higher plate dissipation</td>
</tr>
<tr>
<td></td>
<td>Lower drive impedance</td>
</tr>
<tr>
<td>B:</td>
<td>Reduced efficiency</td>
</tr>
<tr>
<td></td>
<td>Lower input and output</td>
</tr>
<tr>
<td></td>
<td>Reduced grid current</td>
</tr>
<tr>
<td></td>
<td>Increased plate current</td>
</tr>
<tr>
<td>C:</td>
<td>Reduced grid current</td>
</tr>
<tr>
<td></td>
<td>Reduced Plate current</td>
</tr>
<tr>
<td></td>
<td>Reduced input and output</td>
</tr>
<tr>
<td>D:</td>
<td>Higher distortion</td>
</tr>
<tr>
<td></td>
<td>(peak flattening)</td>
</tr>
<tr>
<td></td>
<td>Increased grid current</td>
</tr>
<tr>
<td>E:</td>
<td>Lower distortion</td>
</tr>
<tr>
<td>F:</td>
<td>Increased input and output</td>
</tr>
<tr>
<td></td>
<td>Increased efficiency</td>
</tr>
<tr>
<td></td>
<td>Increased plate dissipation</td>
</tr>
<tr>
<td></td>
<td>(Do not exceed mfg’s max Plate Voltage)</td>
</tr>
<tr>
<td>G:</td>
<td>Lower quiescent plate current</td>
</tr>
<tr>
<td></td>
<td>Lower quiescent plate dissipation</td>
</tr>
<tr>
<td></td>
<td>Increased distortion</td>
</tr>
<tr>
<td></td>
<td>(non-linear “crossover”)</td>
</tr>
</tbody>
</table>

<table>
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<tr>
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</thead>
<tbody>
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<td></td>
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</tr>
<tr>
<td></td>
<td>Higher input and output</td>
</tr>
<tr>
<td></td>
<td>Higher plate dissipation</td>
</tr>
<tr>
<td></td>
<td>Lower drive impedance</td>
</tr>
<tr>
<td>B:</td>
<td>Reduced efficiency</td>
</tr>
<tr>
<td></td>
<td>Lower input and output</td>
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<tr>
<td></td>
<td>Reduced grid current</td>
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<tr>
<td></td>
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<tr>
<td>C:</td>
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<tr>
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<tr>
<td></td>
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</tr>
<tr>
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<td>Increased input and output</td>
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</tr>
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<td></td>
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</tr>
<tr>
<td></td>
<td>(Do not exceed mfg’s max Plate Voltage)</td>
</tr>
<tr>
<td>G:</td>
<td>Lower quiescent plate current</td>
</tr>
<tr>
<td></td>
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</tr>
<tr>
<td></td>
<td>Increased distortion</td>
</tr>
<tr>
<td></td>
<td>(non-linear “crossover”)</td>
</tr>
</tbody>
</table>
EIMAC's new DX champion! The 3CX800A7.

Varian EIMAC continues to commit its development of reliable tubes for HAM radio.

The new, rugged 3CX800A7 power triode provides 2 kW PEP input for voice service or 1 kW cw rating up to 30 MHz. Two tubes will meet the new, higher power ratings authorized by the FCC.

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A data sheet and more information is available from Varian EIMAC. Or the nearest Electron Device Group sales office. Call or write today.

Varian EIMAC
301 Industrial Way
San Carlos, California 94270
Telephone: 415-592-1227
Ebb = 3150  Ip = .87  Emin = 250  Bias (Zener) = 5.2

Plate Supply Voltage = 3150 Volts
Cathode Bias (Zener) = 5.2 Volts
Zero Signal Plate Current = 67 mA dc
Single-Tone Plate Current = 278 mA dc
Single-Tone Grid Current = 90 mA dc
Grid Power Dissipation = 6 Watts
Peak RF Cathode Voltage = 87.8 Volts
Feed-through Power = 19 Watts
Grid Drive Power = 25 Watts
Total Cathode Drive Power = 50 Watts
Cathode Drive Impedance = 152.2 Ohms
Power Input = 878 Watts
PEP Power Output = 617 Watts
Plate Dissipation = 279 Watts
Plate Load Impedance = 6806 Ohms

Do you want to change an input - Y or N?

fig. 4. Program output for the design for two parallel 3-500Zs with 3150 plate supply volts, and 100 watts total drive.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Program Mfgr.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plate Supply Voltage</td>
<td>1500 1500</td>
</tr>
<tr>
<td>Cathode Bias</td>
<td>0 0</td>
</tr>
<tr>
<td>Quiescent Plate Current</td>
<td>51 65</td>
</tr>
<tr>
<td>Single-Tone Plate Current</td>
<td>403 400</td>
</tr>
<tr>
<td>Single-Tone Grid Current</td>
<td>142 130</td>
</tr>
<tr>
<td>Grid Drive Power</td>
<td>51 49</td>
</tr>
<tr>
<td>Cathode Impedance</td>
<td>134 94</td>
</tr>
<tr>
<td>Power Input</td>
<td>606 600</td>
</tr>
<tr>
<td>PEP Output</td>
<td>330 330</td>
</tr>
<tr>
<td>Plate Load Impedance</td>
<td>1602 1600</td>
</tr>
</tbody>
</table>

Table 2. Comparison of calculated and manufacturer’s data.

points "I_p" and "Q" are moved. Figure 3 is a schematic showing what to expect when shifting the points. Use it in conjunction with table 1. It is essential that you be familiar with these principles — that’s the only way you will accomplish your final design.

how well does the program work?

The proof is in the performance. A sample of the program output appears as fig. 4, which is also the demonstration of a design using two parallel 3-500Zs. It looks just like the manufacturer’s list of typical operating data.

Table 2 compares the program-calculated data with those published by ELMAC under the heading “Typical Operating Data”. The results of trying to match two sets (at 1500 volts and 3500 volts) are compared in the table. The manufacturer’s data came from the latest technical data sheets for the 3-500Z (the revision effective April 1, 1986). I have culled all except the directly comparable data from the table; they are remarkably close. Figures 5 and 6 are the program outputs for the 1500- and 3500-volt cases.

parallel operation

All the tables reflect data for one tube. If you choose a two-tube parallel operation, all currents and power levels must be doubled. All impedances (such as the cathode drive and plate load impedances) must be halved. Voltages remain unchanged.

A hypothetical design demonstration follows:

Suppose you have a power transformer that will deliver 3500 volts dc at no load. A typical power supply voltage will sag about 10 percent under load, so enter 3150 for E_bb. Now suppose that you have 100 watts of drive (PEP) from the exciter, and you also want to have the most power available from the amplifier.
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Plus you get high performance HF/VHF/CW modems, software selectable dual radio ports, precision tuning indicator, 32K RAM, AC power supply and more.

You'll find it the most user friendly of all multi-modes. It's a menu driven for ease of use and command driven for speed.

A high resolution 20 LED tuning indicator lets you tune in signals fast in any mode. All you have to do is to center a single LED and you're precisely tuned in to within 1 Hz and it shows you which way to tune!

All you need to join the fun is an MFJ-1278, your rig and any computer with a serial port and terminal program.

You can use the MFJ Starter Pack to get on the air instantly. It includes computer interfacing cable, terminal software and friendly instructions... everything you need to get on the air fast. Order the MFJ Starter Pack (MFJ-1283) for the C-64/128 and VIC-20 or MFJ-1284 for the IBM or compatible. $19.95 each.

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With MFJ's superclone of the industry standard -- the TAPR TNC-2 -- you get genuine TAPR software/hardware plus more... not a "work-a-like" imitation.

Extensive tests published in Packet Radio Magazine ("HF Modem Performance Comparisons") prove the TAPR designed modem used in the MFJ-1278 gives better copy with proper DCD operation under all tested conditions than the other modems tested.

Hardware DCD gives you more QSOs because you get reliable carrier detection under busy, noisy or weak conditions.

A hardware HDLC gives you full duplex operation for satellite work or for use as a full duplex digipeater. And, it makes possible speeds in excess of 56K baud with a suitable external modem.

Good news for SYSOPs! New software lets the MFJ-1278 perform flawlessly as a WORLI/WA7MLB bulletin board TNC.

Baudot RTTY
You can copy both shifts and all standard speeds including 170, 425 and 800 Hz shifts and speeds from 45 to 300 baud. You can copy not only amateur RTTY but also press, weather and other exciting traffic.

A high performance modem lets you copy both mark and space for greatly improved copy under adverse conditions. It even tracks slightly drifting signals.

You can transmit both narrow and wide shifts. The wide shift is a standard 850 Hz shift with mark/space tones of 2125/2975 Hz. This lets you operate MARS and standard VHF FM RTTY.

You get the American Western Union and the International CCITT character sets. Start your own unattended reception and selectable "Diddie".

A receive Normal/Reverse software switch eliminates retuning and Unshift-On-Space requires fewer errors under poor receiving conditions.

ASCII
You can transmit and receive 7 bit ASCII using the same shifts and speeds as in the RTTY mode and using the same high performance modem. You also get Autostart and selectable "Diddie".

CW
You get a Super Morse Keyboard mode, that lets you send perfect CW effortlessly from 5 to 99 WPM, including all prosigns -- it's tailor-made for traffic handlers.

A huge type-ahead buffer lets you send smooth CW even if you "hunt and peck".

You can store entire QSOs in the message memories... you want!

You can link and repeat any messages for automatic CFQs and beaconing. Memories also work in RTTY and ASCII modes.

A tone Modulated CW mode turns your VHF FM rig into a CW transceiver for a new fun mode. It's perfect for transmitting code practice over VHF FM.

An ASCII CW mode lets you QTY in CW. The CW receive mode lets you copy from 1 to 99 WPM. Even with sloppy fists you'll be surprised at the copy you'll get with its powerful built-in software.

You also get a random code generator that'll help you copy CW faster.

Weather FAX
You'll be fascinated as you watch WEFAX signals blossom into full fledged weather maps on your printer.

Other interesting FAX pictures can also be printed -- such as some news photographs from wire services.

Any Epson compatible printer will print a wealth of interesting pictures and maps.

Automatic sync and stop lets you set it and leave it for no hassle printing.

You can save FAX pictures and WEFAX maps to disk if your terminal program lets you save ASCII files to disk.

Pictures and maps can be printed to screen in real time or from disk on IBM and compatibles with the MFJ-1284 Starter Pack.

You can transmit FAX pictures right off disk and have fun exchanging and collecting them.

Slow Scan TV
The MFJ-1278 lets you exchange pictures with thousands of SSTVers all-over-the-world.

You'll not only see what your ham buddies look like but you can send your own pictures to them, too.

You can print slow scan TV pictures on an Epson compatible printer. If you have an IBM PC or compatible you can print to screen in near real time or from disk with the MFJ-1284 Starter Pack.

You can transmit slow scan pictures right off disk -- there's no need to set up lights and a camera for a casual contact.

You can save slow scan pictures on disk from over-the-air QSOs, audio tapes and other sources if your terminal program lets you save ASCII files.

The MFJ-1278 transmits and receives 8, 5, 12, 24, and 36 second black and white format SSTV pictures using two levels.

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You get iambic operation with dot-dash memories, self-completing dots and dashes and jamproof spacing.

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A weight control lets you penetrate QRM with a distinctive signal or let's your transmitter send perfect sounding CW.

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MFJ... making quality affordable
The program was run with initial inputs of 3150 volts, 1.0 A (chosen as a starting point), 250 volts, and 5.2 volts. The Zener was chosen as 5.2 volts because it is about the same bias voltage as that used in the Heath SB-220. It proved to be a good choice. By reducing \( i_p \) incrementally via the menu, the calculated drive power was reduced to 50 watts exactly (for one tube). It occurred when the max plate current \( (i_p) \) reached 0.87 A (see fig. 4). By "doubling and halving," the resulting numbers for two tubes in parallel are:

- **Plate Supply Voltage** = 3150 volts
- **Cathode Bias** = 5.2 volts
- **Zero Signal Plate Current** = 134 mA dc
- **Single-Tone Plate Current** = 556 mA dc
- **Single-Tone Grid Current** = 180 mA dc
- **Feed-Through Power** = 38 watts
- **Total Cathode Drive** = 100 watts

---

### Table: Program Output for 3150 Volts

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ebb</td>
<td>1500</td>
</tr>
<tr>
<td>Ip</td>
<td>1.31</td>
</tr>
<tr>
<td>Emin</td>
<td>$\approx 70$</td>
</tr>
<tr>
<td>Bias (Zener)</td>
<td>0</td>
</tr>
<tr>
<td>Plate Supply Voltage</td>
<td>1500 Volts</td>
</tr>
<tr>
<td>Cathode Bias (Zener)</td>
<td>0 Volts</td>
</tr>
<tr>
<td>Zero Signal Plate Current</td>
<td>51 mA dc</td>
</tr>
<tr>
<td>Single-Tone Plate Current</td>
<td>403 mA dc</td>
</tr>
<tr>
<td>Single-Tone Grid Current</td>
<td>142 mA dc</td>
</tr>
<tr>
<td>Grid Power Dissipation</td>
<td>14 Watts</td>
</tr>
<tr>
<td>Peak RF Cathode Voltage</td>
<td>117.9 Volts</td>
</tr>
<tr>
<td>Feed-through Power</td>
<td>38 Watts</td>
</tr>
<tr>
<td>Grid Drive Power</td>
<td>51 Watts</td>
</tr>
<tr>
<td>Total Cathode Drive Power</td>
<td>103 Watts</td>
</tr>
<tr>
<td>Cathode Drive Impedance</td>
<td>134.4 Ohms</td>
</tr>
<tr>
<td>Power Input</td>
<td>605 Watts</td>
</tr>
<tr>
<td>PEP Power Output</td>
<td>330 Watts</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>312 Watts</td>
</tr>
<tr>
<td>Plate Load Impedance</td>
<td>1602 Ohms</td>
</tr>
</tbody>
</table>

---

**Do you want to change an input - Y or N?**

---

### Table: Program Output for 3500 Volts

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ebb</td>
<td>3500</td>
</tr>
<tr>
<td>Ip</td>
<td>1.25</td>
</tr>
<tr>
<td>Emin</td>
<td>350</td>
</tr>
<tr>
<td>Bias (Zener)</td>
<td>15</td>
</tr>
<tr>
<td>Plate Supply Voltage</td>
<td>3500 Volts</td>
</tr>
<tr>
<td>Cathode Bias (Zener)</td>
<td>15 Volts</td>
</tr>
<tr>
<td>Zero Signal Plate Current</td>
<td>25 mA dc</td>
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<tr>
<td>Single-Tone Plate Current</td>
<td>396 mA dc</td>
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<tr>
<td>Single-Tone Grid Current</td>
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<tr>
<td>Grid Power Dissipation</td>
<td>11 Watts</td>
</tr>
<tr>
<td>Peak RF Cathode Voltage</td>
<td>116.6 Volts</td>
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<tr>
<td>Feed-through Power</td>
<td>37 Watts</td>
</tr>
<tr>
<td>Grid Drive Power</td>
<td>49 Watts</td>
</tr>
<tr>
<td>Total Cathode Drive Power</td>
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<tr>
<td>Cathode Drive Impedance</td>
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<td>Power Input</td>
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</tr>
<tr>
<td>PEP Power Output</td>
<td>991 Watts</td>
</tr>
<tr>
<td>Plate Dissipation</td>
<td>430 Watts</td>
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<tr>
<td>Plate Load Impedance</td>
<td>5001 Ohms</td>
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</tbody>
</table>

---

**Do you want to change an input - Y or N?**

---

*fig. 5. Program output for duplicating the manufacturer's data for 1500 plate supply volts.*

*fig. 6. Program output for duplicating the manufacturer's data for 3500 plate supply volts.*
Cathode Drive Impedance = 76 ohms
Power Input = 1756 watts
PEP Power Output = 1234 watts
Plate Load Impedance = 3403 ohms

The actual power output would be (1234 + 38) or 1272 watts because of the feed-through power. Total plate dissipation would be 559 watts. The design can proceed from here.

Always keep the manufacturer's maximum ratings in mind. Two appropriate values to monitor are the plate dissipation and the maximum plate voltage. Another more important one is grid dissipation, also calculated by the program. Normally you will never exceed all the maximum ratings at once — but stay alert to assure that it doesn't happen. This program should be accompanied by the manufacturer's tube data sheets.

**comments**

In table 2 there are two lines which show some discrepancies; neither of these is very important. They involve plate quiescent current and cathode drive impedance. These numbers agree with those calculated by hand from the EIMAC Bulletin No. 5 "Tube Performance Computer". Even so, the table 2 “worst-case” discrepancy (cathode drive impedance for the 1500-volt case) would result in a VSWR of only 1.4. Everything else came out much better than I expected.

My program is complicated and takes lots of time. But remember that when the work is done, the resulting program will simulate data for an operational circuit. It should be emphasized that "real" tubes may produce numbers which differ by as much as ± 10 percent in the main part of the characteristics curves. Though I have made many design calculations with the aid of this program, I certainly have not challenged every possible entry. I would be interested in your results. Let know if there are any “glitches” in the program, and send me your suggestions for improvements.

I plan to put several design programs on disk. I have done this same routine for the 3CX1200A7; the routines for the 8877 are half-finished. I will tackle each tube type in succession until most of the common tubes are covered. In the meantime, I hope you have good results with this 3-500Z design program.

**acknowledgment**

Thanks to Frank Chess, K3BN, who helped with the programming. It was his idea to echo the inputs at the heading of the output routine.

**references**

3. Available from Varian/EIMAC, 1678 Pioneer Road, Salt Lake City, Utah 84104

Ham Radio
designing a station
for the microwave bands: part 2

A complete 10-GHz Amateur SSB/CW station

Part 1 discussed why the microwave Amateur bands may be better than lower frequencies for many applications, though in the past Amateurs have viewed them as line-of-sight realms. It described some of the inherent advantages microwaves have for point-to-point communication, even over modern higher power hf, VHF, or UHF stations. These advantages make them very attractive for high volume, high data rate communications like those required for Amateur networking.

A local oscillator frequency scheme using common pc boards was presented. It can be used to get a station on all of the Amateur microwave bands with a minimum of redundant construction. This scheme uses conventional lower frequency components, readily available microwave oscillators, and only a small amount of additional microwave construction to produce a high quality narrowband station. Part 1 and the rest of this series demonstrate this approach by describing construction of a complete 10-GHz Amateur SSB/CW station — the station that holds one end of the current North American 10-GHz DX record of 414 miles.

spectral purity

The cornerstone of this station is a spectrally pure and stable 1010-MHz oscillator. Spectral purity, sometimes overlooked more than it should be even on the hf bands, is of particular importance when operating on microwave frequencies. This is because the “contamination” produced by angular (phase or frequency) modulation of a low-frequency reference signal is multiplied right along with the signal itself when a harmonic is used in a microwave system. The fact that drift and frequency errors are multiplied is well known to anyone who tries to “net” a pair of fm transceivers on 1200 or even 440 MHz. However, these frequency domain “imperfections” are members of a whole class of impurities given the name “phase noise”. Even a quartz oscillator in a modern hf transceiver exhibits this to some degree. In a well-designed oscillator the “cleanliness” of a signal is related to its operating frequency. On the Amateur hf bands these noise characteristics may be so small relative to normal signal-to-noise ratios that they are unobservable, except perhaps as an increase in background noise level down the band from a local “big gun”. Some of the early synthesized ham band transceivers exhibited this as noise “humps” a few kHz either side of the carrier frequency on both transmit and receive. Commercial Amateur equipment has improved to the point where fundamental overload or other factors usually come into play before the phase noise of the local oscillators is observed. However, as higher frequencies are required and higher harmonic multiples of reference oscillators are used, these unwanted components are multiplied. The relative amplitude of these unwanted signals follows a $20 \log N$ rule, where $N$ is the harmonic number. This means that on the tenth harmonic of a signal, the phase noise sidebands can be expected to increase by $20 \log 10$ or $20 \text{ dB}$. The 100th harmonic will be $40 \text{ dB}$ worse than the fundamental. Consequently, a “clean” signal at 10 MHz, one with say $-90 \text{ dB} \text{C}$ (dB relative to the carrier) noise sidebands or fm spurious signals, might be $60 \text{ dB}$ worse at 10 GHz, or $-30 \text{ dB}C$. On an S9 signal such noise might be barely audible; however, if the fundamental oscillator was only $-60 \text{ dB}$ the resulting microwave signal might be unusable for communications. Because the 1010-MHz oscillator and its harmonics provide a local oscillator signal for a narrowband station, spectral purity must be maintained. Although a PLL can serve to “clean up” a poorer oscillator at frequencies close in to the carrier, no improvement is made.

By Glenn Elmore, N6GN, 3528 Deerpark Drive, Santa Rosa, California 95404

June 1988
Beyond the PLL bandwidth. For this reason the best available oscillator should be used.

1-GHz reference oscillator

This oscillator can be used as the LO for a 1296-MHz signal mixer directly, as well as for the microwave harmonic downconverter reference at microwave frequencies. The active device is an inexpensive bipolar transistor. A coaxial resonator is made from pc board and brass tubing. Three separate buffered outputs are provided for phase locking, downconverter reference, and 1296-MHz signal mixer LO. The oscillator is tuned with the same UHF TV tuner diode used in the 100-MHz reference along with a short length of wire coupled to the resonator at the low-impedance end. This provides approximately ±3-MHz tuning range around a 1010-MHz center frequency.

Oscillator tuning is somewhat novel; it works in much the same way as “loop modulation” of early radio days. Free running high-power oscillators were used with a carbon microphone connected across a single-turn loop located in the vicinity of a frequency determining inductor. As the operator spoke into the microphone the resistive load across the loop varied, which in turn modulated the loop current. Because this was an induced current, it tended to produce an opposing flux which effectively varied the net tank inductance and frequency modulated the oscillator. The technique works, but be careful not to couple too closely or extract so much energy from the tank that you burn up the microphone — not to mention the operator!

The method used here doesn’t extract much power from the tank, as the load the varicap presents to the loop is mostly reactive. Any such dissipation is undesirable as it acts to lower the operating Q of the resonator. The varicap value and coupling wire inductance are chosen to be below self-resonance for any tuning voltage. This is done to limit the maximum current and control energy loss in the tuning circuit resistances. If the tuning circuit tunes too close to resonance, oscillation may stop. With nominal loop dimensions and the indicated varicap, the 1010-MHz oscillator tunes with a nearly straight frequency/voltage tuning curve. The 5-MHz tuning range is ample to maintain lock once the other loops and coarse tuning are adjusted to center the output frequency.

Two versions of this oscillator have been built. The first uses a quarter-wave line allowing physically smaller construction, but requiring a dielectric support for the high-impedance end of the line to obtain the lowest “microphonics”. The second approach uses a
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#8 CALLER CONTROLLED REVERSE AUTO PATCH WITH USER PROGRAMMABLE ACCESS CODES
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#10 SMART AUTO PATCH
#11 DIAL 9 FIRST, PBX FEATURE
#12 SEPERATE REVERSE AUTO PATCH ON/OFF COMMAND
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#22 ADJUSTABLE AUTO PATCH TIME OUT TIMER
#23 DTMF TONE MUTE AND COVER TONE MASKING
#24 REPEATER TIME OUT TIMER WARNINGS
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The 1010-MHz oscillator is built from pc board and copper flashing. The assembly is divided into three compartments — one for the resonator and one each for the oscillator and signal amplifier circuits. Minimum lead length is used when components are soldered directly to the circuit board material.

The half-wave line and, although longer, is simpler to construct. The quarter-wave version allows tuning versatility by "telescoping" the inner conductor with a length of the next smaller size brass tubing sliding through the center of the fixed tubing. It tuned it continuously from 800 to more than 1200 MHz.

The half-wave version has another advantage. When you place the oscillator transistor with its isolation amplifier on one end of the resonator and signal amplifiers on the other end, the resonator serves to isolate spurious signals which might be present in the downconverter/phase lock circuitry. This "autofiltering" makes it easier to achieve —80 dBc spectral purity at 1 GHz. Similar performance can be obtained with the quarter-wave version, but more stages of isolation and careful shielding are required.
fig. 3. 1-GHz oscillator.

fig. 4. A simple diode detector and voltmeter can measure signal power up through the VHF range. The circuit is useful for relative power measurement well past 1 GHz.

The BFR91 oscillator transistor is optimized to have maximum negative resistance at 1 GHz with the insertion of approximately 7 nH of inductance in its base lead. This inductor is just the 3/8 inch of lead length between the transistor package and the feedthrough capacitor ground on the end wall. The emitter is coupled into the resonator with a loop, also bypassed in a feedthrough capacitor on the same wall. Base and emitter biasing resistors are connected on the outside. The 1-GHz oscillator schematic is shown in fig. 1. Figure 2 shows the mechanical dimensions and positioning for the resonator, feedthrough capacitors, and coupling loops. A photo of the completed oscillator is shown in fig. 3.

collection

The oscillator could be built entirely of pc board, but I chose to make the end walls from copper flashing. This makes it easier to solder the brass tubing after the resonator box has been assembled. The sides, center, ends, and partition should all be punched or drilled before soldering. Holes for the oscillator emitter loop and buffer amplifier input loop are in the center wall. Amplifier transistor emitters and all bypass capacitors can be soldered directly to the board material with virtually no excess lead length. The oscillator emitter lead can protrude right through the center wall hole and be soldered to its coupling loop. An 8-32 brass nut should be soldered to the inside wall of the resonator so that a tuning screw can be inserted later. If possible use 1/8th or 1/16th watt resistors. The physically smaller packages should have less associated inductance. Choose feedthrough capacitors small enough to fit snugly against the brass tubing on the oscillator end. These must be soldered in place since their nuts would otherwise interfere with the brass tubing protruding from the end wall.

adjustment

Begin check-out without tuning screws and apply 12 volts. The oscillator emitter (measured at the outside of its feedthrough capacitor) should sit at about 3.5 volts, and the amplifier transistors should have 6 to 10 volts on their collectors. Collector currents of about 15 mA for the BFR91 and 40 mA for the BFR96 amplifiers are fine. All three outputs should have a load connected; a 50-ohm resistor may be tacked across an unused output as a temporary load. If a power meter or other calibrated detector is not available, an inexpensive power detector may be made (fig. 4). An approximate calibration curve useful through the VHF range is shown in fig. 5. At 1 GHz the curve may not accurately predict the detected power because of differing construction techniques and component characteristics, but the detector should still be useful for determining relative output powers and adjusting the 1010-MHz circuits. I built the detector right on the cable end of the same type of SMB coax connector I used throughout. You can use it to verify ECL outputs as well as oscillator performance.

A 1-GHz frequency counter or a spectrum analyzer

fig. 5. A plot of the detector output voltage as a function of input power shows a useful range from about 0 to +16 dBm (1 to 40 milliwatts).
is extremely useful for tune-up. If such test equipment is not available, build the 1-GHz harmonic downconverter described in the next section. Use it to convert the 1-GHz signal down to the hf range of a general coverage receiver or low frequency counter. If you use a receiver, couple the downconverter lightly or use an attenuator to avoid overload. Overloading can cause confusion because of images and other spurious responses.

With an applied fixed tuning voltage of 6 volts, insert the tuning screw and set the frequency to approximately 1010 MHz. Adjust the emitter loop slightly to assure oscillation while varying the tuning voltage over the 2-to-10 volt varicap tuning range. Reduce coupling by decreasing the area of the loop and positioning it further from the brass tubing. Use the minimum coupling to maintain output so you can avoid unnecessarily loading the resonator and degrading phase noise. This coupling is somewhat dependent on resonator loading by both the tuning circuit and buffer amplifier input loops. Adjust the buffer amplifier loop (made from the coupling capacitor lead) for minimum coupling consistent with maximum power out of the power splitter. Adjust the emitter loop to maintain output over the whole tuning range. Some iteration between these two adjustments may be necessary to arrive at the best settings. If you find that the oscillator dies at the high end of the tuning range, or just above 10 volts, you may need to lower the tuning circuit resonant frequency. Do this by lengthening the tuning inductor slightly. The values shown in the drawing should provide a good starting point and should work without modification. Extreme emitter loop over-coupling can cause "squegging", the output switching rapidly between two frequencies. This is not a problem if the above adjustment procedure is followed. Reduce coupling if you observe spurious sidebands on the unlocked oscillator or find low-frequency oscillations on the bias feedthrough capacitors.

The output amplifier on the signal side is followed by a power splitter made from two 2-inch lengths of semi-rigid coax. This is a simple way to provide two outputs. If only one 1010-MHz source is required, it may be omitted and the single BFR91 buffer amplifier used to provide +10 dBm for a signal mixer. The two-stage amplifier with a BFR96 in the output and the power divider can easily provide two +13 dBm (20 milliwatt) sources.

Once the loops are positioned for proper power output, all that remains is to readjust the tuning screw so the oscillator "free runs" right at the desired frequency. If you adhere to the dimensions for the half-wave version, the oscillator should run at about 1025 MHz with no tuning screw and only the 4.5 volts from the resistive divider on the tuning input. It should tune down mechanically to 1000 MHz without a significant change in output power. When you obtain the proper frequency, secure the tuning screw locking nut. Verify that approximately + and −2 MHz tuning is possible, respectively.

PLL harmonic downconverters

The downconverters themselves are similar, although implementation at 10 GHz is somewhat different from that at 1 GHz. Anti-parallel diodes are used with a diplexer arrangement to couple signals in and out. The downconverter block diagram is shown in fig. 6.

The anti-parallel diode pair is effectively an even harmonic mixer. Its simplicity and built-in protection from overload and static damage make it attractive for this application. Depending on harmonic number and phase-locked oscillator frequency, −30 to −40 dB conversion efficiencies are obtainable even with "ham shack" construction — i.e., discrete components or microstrip circuits cut out of Teflon™ epoxy pc board material with a small hobby knife. The high-pass filter couples the reference fundamental into the diodes; the low-pass filter couples the i-f out. The oscillator can be connected directly to the diode pair through a small capacitor.

At 1 GHz, packaged diodes and discrete capacitors and inductors can be used. Lead length should be kept to a minimum, but otherwise the circuit is extremely simple to build. The diodes generate considerable energy at odd harmonics of 100 MHz. However, the isolation of the 1010-MHz oscillator resonator, not to mention the buffer amplifiers, keeps this energy from showing up in the signal output. These sidebands are for the most part amplitude, not frequency modulated, and don't get "amplified" when higher harmonics of the 1010-MHz signal are used as a reference signal in the 10-GHz downconverter. The PLL i-f signal
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- Separate Volt and Amp Meters
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  - RM-35M: 25, 35
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## RS-M SERIES

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## VS-M AND VRM-M SERIES

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from the downconverter is approximately 30 dB below the reference or locked oscillator levels. This conversion loss is made up for in the bipolar amplifier and the two ECL line receivers on the phase-lock circuit. With 10 to 13 dBm reference drive, i-f output doesn’t change dramatically for 0 to 10 dBm oscillator input. Around -30 dBm PLL i-f power is typical for both converters — plenty to drive the last ECL line receiver before the phase comparator well into saturation. The i-f output may actually drop if the oscillator input level is increased too far. The 1-GHz harmonic downconverter schematic diagram is shown in fig. 7. Figure 8 is a photo of the completed 1-GHz downconverter and common PLL board.

The 100-MHz reference signal is bandpass filtered and amplified from the 0-dBm ECL levels. The filtering makes sure that any low level, low-frequency digital signals which might be present on the 100-MHz ECL output don’t “ride” straight through to the PLL i-f amplifiers. Diode drive of 10 to 20 milliwatts is adequate.

The 10-GHz downconverter is functionally the same as the 1-GHz version. Here, however, a pair (or half a quad) of diodes in a small package is used to avoid parasitic inductance and capacitance associated with the larger discrete diodes. Many of the filter elements are made using microstrip techniques instead of lumped components. Chip capacitors are used to minimize parasitic inductance.

Because most of the 10-GHz oscillator power is needed for converting the VHF signal to and from 10 GHz, a hybrid coupler is used to extract only enough to make the PLL downconverter operate. This hybrid has one of its input ports terminated with a discrete resistor. This termination needn’t be very good at 10 GHz, as the object of the coupler is simply to extract a sample of the energy (10 dB or so down) and its directivity isn’t particularly important. Use as physically small a resistor as possible with 0 lead length. All of the high-impedance lines may be made from some small diameter wire and soldered across the wider traces. Number 38 wire should be fine for this.

The signal mixer is shown with the 10-GHz downconverter and can be built on the same board at the same time. This makes it possible to get on the band as soon as the 10-GHz oscillator is locked and a VHF i-f is available. The 10-GHz harmonic downconverter is shown in figs. 9A and 9B. Figure 10 shows a 2:1
The 10-GHz and 1010-MHz downconverters are functionally identical. At 10 GHz, however, microstrip components replace discrete components. A hole is provided in the 1/32 inch Teflon board material under diode ring, D1, to allow shorting the diode leads to the backside ground. Radial transmission lines on these same leads help assure a low-impedance ground connection.

fig. 9. The 10-GHz and 1010-MHz downconverters are functionally identical. At 10 GHz, however, microstrip components replace discrete components. A hole is provided in the 1/32 inch Teflon board material under diode ring, D1, to allow shorting the diode leads to the backside ground. Radial transmission lines on these same leads help assure a low-impedance ground connection.

locking to 1010 MHz

The 1010-MHz common PLL circuit is nearly identical to that of the 100-MHz reference oscillator — only the loop filter values are different. For this loop, the phase comparator VCO input comes from the filtered and amplified output of the 1-GHz harmonic downconverter. A 35-MHz low-pass filter follows the downconverter; the PLL i-f is first amplified by a two-stage controlled-gain amplifier. I used this configuration instead of another ECL line receiver for two reasons: it allowed variation of the stage gain by changing a single resistor value, and the bipolar amplifier has lower bandwidth than the ECL line receiver. The rest of the PLL circuitry is identical to the 100-MHz phase lock except for the loop filter component values. The bandwidth of this loop is set to approximately 50 kHz.

Once the 1010-MHz oscillator is built and adjusted, you are ready to lock it up. Use one of the common PLL boards with the loop filter component values in part 1, table 2. Set the jumper wires on the phase comparator input for the “+” configuration. If the PLL board is working properly (remember that you can test it ahead of time by using it to lock up the 100-MHz oscillator), the loop should close and “pull in” the 1010-MHz oscillator exactly on frequency. This lock can occur if the 100-MHz loop is locked or free running, and the output frequency should be exactly 10.1 times the 100-MHz crystal oscillator frequency. Make sure you use the 10-MHz reference to lock at 1010 MHz and the 20-MHz reference if you are trying to lock to 1020 MHz.

Troubleshoot any problems by checking the PLL board and the 1010-MHz oscillator independently of each other. As long as the oscillator tunes over the correct range and the PLL board is working, there

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should be no difficulty in achieving and maintaining lock. Once this is done, you have an LO for use in a 1296/2304 station or as a reference oscillator for locking your 10-GHz oscillator.

10-GHz oscillator selection and locking

The 10-GHz oscillator is locked in the same manner as the 1010-MHz reference. The tuning circuit may depend on the type of oscillator available. Generally, only enough tuning range to overcome drift and instability is used. If too much tuning range is provided, the microwave oscillator might get on the “wrong side” of the downconverter reference frequency harmonic, giving an i-f with the wrong tuning sense. If this happens, the PLL amplifier tries to tune the oscillator in the wrong direction to acquire phase lock and the loop will remain saturated and unlocked. For a 20-MHz PLL i-f, 30 MHz of total electronic tuning range should be adequate, and this combined with about a 10-volt swing out of the loop amplifier suggests a 3-MHz/volt tuning sensitivity. If the microwave oscillator is unstable or drifts (necessitating a greater tuning range), an ECL divide-by-2 or divide-by-4 could be inserted right at the phase comparator input. Of course, this would produce a different locked output frequency, and all other i-f and oscillator frequencies in the system might have to be reselected. The loop filter component values would also have to be recomputed.

Selection of the 10-GHz oscillator depends upon what is available and within your budget. The M/A-Comm Gunnplexers™ work extremely well and require very little additional circuitry. If you have one of these as part of a wideband station, you may want to use the 10,220-MHz locking scheme. If there is already some broadband 10-GHz activity in your area and you
don’t want to give it up entirely, this approach will allow switching between modes. The Gunnplexer can be operated with its internal diode mixer for operation on 10220/10250 wideband duplex, or phase locked to 10220 and used with a 148-MHz SSB transceiver for 10368-MHz narrowband weak signal work. The wideband station can also be run phase locked with modulation of the 20-MHz reference signal in the 1020-MHz loop phase. (This should end any local discussions about who is or is not on the right frequency!)

The M/A-Comm Gunnplexers have electronic tuning and need only level shifting and scaling of the tuning voltage. A typical tuning curve for a GunnPlexer is shown in fig. 12. Driving the tuning input directly from the loop amplifier provides too much tuning range and could allow “latch-up” on the wrong side of the i-f, as mentioned before. It is a simple matter to scale the tuning input to reduce the approximately 7-MHz/volt sensitivity down to about 3. A circuit providing this scaling, as well as a regulated 10-volt bias supply, is shown in fig. 13. This circuit will maintain proper output and tuning even when the power supply voltage drops slightly below 12 volts. A low dropout regulator may be substituted for the LM317K for particularly low inputs. This is of concern primarily when mountain topping with discharged batteries as the only power source! The phase-locked Gunnplexer produces an excellent 10080-MHz signal (fig. 14).

Some means of tuning must be provided if an oscillator without an electronic tuning input is used. The Gunn oscillators in automatic door openers can be made to work by using bias voltage “frequency pushing”. These are very similar to Gunnplexers except for their lack of electronic tuning and a mixer diode. The tuning deficiency can be overcome by using the bias/tuning circuit in fig. 15. Here a three-terminal regulator sets the bias and tunes the oscillator for phase locking. To pick the nominal bias point, plot a frequency versus bias voltage curve for your particular oscillator — this will vary from oscillator to oscillator. Usually a range of bias can be found (often just on one side of maximum power output) that provides a fairly straight tuning curve or nearly constant tuning sensitivity. A plot of a typical bias-tuned oscillator is shown in fig. 16. The tuning resistor values are selected to tune over a 24-MHz range with 2 to 10 volts on the tune input. The nominal operating
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need to be in coax in order to use the downconverter and mixer. Although waveguide is well behaved and very low in loss, coax is versatile and convenient. I have used coax throughout the 10-GHz station, both at 1 and 10 GHz. Miniature SMB “snap on” connectors work well at 1 GHz and below, even when used with poor quality lossier coax cable. In the microwave region, 0.086-inch semi-rigid cable is a pleasure to work with; the cable and corresponding SMA connectors are fairly easy to find. To cut the cable to length, first score the outer conductor with a sharp knife; then grab each side of the score mark with a pair of needle-nose pliers and break. The Teflon dielectric can be trimmed away and the cable end slid into the connector or soldered directly to the circuit, depending upon the application.

If your oscillator is similar to the door-opener type, it probably has a waveguide output and will require a waveguide-to-coax adapter. These are often available as surplus but if you don’t have or can’t get one, it is easy to build an acceptable substitute. The version shown in fig. 17 made from a short length of commercial waveguide works very well, although you’ll need metal-working equipment. If your shack doesn’t include much more than a soldering iron, hacksaw, and file, the second version made from pc board in fig. 18 is for you.

After selecting your 10-GHz oscillator, build the appropriate tuning circuit. If a means of measuring 10-GHz frequency (a 10-GHz counter or spectrum analyzer with 1-MHz frequency resolution) is not avail-

![fig. 16. This is a tuning curve of a surplus Solfan™ oscillator of the type used in burglar alarm motion detectors and automatic door openers. Both output power and frequency are dependent upon bias voltage. By plotting a similar curve and selecting a useful portion of the tuning curve, you can find component values for biasing and tuning almost any similar oscillator. In this case, a bias of 8.25 volts + 1.25 volts will tune the output over approximately a + 12-MHz tuning range.](image)

point, with 6 volts applied to the tuning input, is set at the center of this range. If 24-MHz tuning is not possible, use the maximum available and recalculate the PLL component values for the different tuning sensitivities.

The three terminal regulators work in this application because they have several hundred kHz of bandwidth and can follow a 50-kHz bandwidth error signal without adding much additional phase shift. This is necessary for the loop to remain stable. The regulators do add some noise to the oscillator output when used in this configuration; reduction of this is the reason for splitting up and bypassing part of the voltage setting and tuning resistances. This technique is not the ultimate in low phase noise performance, but the ~90 dBc noise sidebands obtainable (1 Hz bandwidth) are more than adequate for Amateur use and will probably never be observed unless signal strengths are 30 or more dB above S9. The Gunnplexers, with their built-in tuning, will probably be at least 8 to 10 dB cleaner than this. Although I have not tried them, many of the oscillators in automotive radar detectors should work well. Another source of suitable oscillators is the type used for police radar guns. The NEC ND751AAM for 10 GHz (ND610AAM for 24 GHz) has similar characteristics. Any of these 10-GHz sources should have adequate drive power for the signal mixer described next.

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for most oscillators I have tried. Select the nominal bias voltage as the center of a reasonably straight 24-MHz range near the maximum power bias point, or in a mode-free region. Select the tuning scaling resistors from fig. 15 based on the change in bias necessary to produce 24-MHz frequency change; this gives 3-MHz/volt sensitivity at the tuning input. The sense of this tuning may be either positive or negative, depending on your particular oscillator. For the motion detector oscillator plotted in fig. 16 I chose a nominal bias point of 8.25 volts. Tuning resistor values were selected for the required volt change (approximately 2.5). These values cause the 2 to 10 volts from the loop amplifier to tune the oscillator over a 24-MHz total range. Resistor and capacitor values for some different oscillator sensitivities are shown in the table.

When you are confident that the oscillator is tuning correctly, preset it to 10080 MHz with 6 volts on the bias circuit tuning input. Do this by coarse tuning for a 20-MHz PLL i-f on the correct side of 10100 MHz. If the PLL board is functioning, locking should now be no more difficult than locking the 1010-MHz oscillator. Remember to select the proper wire jumpers based on "high side" LO and the tuning direction of your particular oscillator.

As for the 1010 MHz oscillator, troubleshoot any problems by separating the PLL components and testing them individually. Make sure that the PLL board works on a lower frequency loop. Verify that there are suitable 1-volt peak-to-peak ECL levels on both phase 10080 MHz LO INPUT FROM DOWNCONVERTER BOARD MATERIAL 0.025" DUROD, DOUBLE CLAD DIODE QUAD 2N4380 HEAT SINK OR METALLICS MS550-142-045 T1 T2 9 TURNS FLAT WOUND ON 0.050" NYLON CORE OR 1/4" NYLON BALLOON RTL 0.050" RADIUS, 60° TAN J5-FLANDED SMA CONNECTORS J6-SMC CONNECTORS NARROW TRACES ARE 0.056" WIDER TRACES ARE 0.083" W306 MHz 10368 MHz 258 MHz I-F

fig. 19. Closeup of the 10-GHz signal mixer.
comparator inputs. Also make sure that the 10-GHz oscillator is tuning properly. Be sure that there is no large (bigger than 1000 pF) bypass capacitor across its bias input; this could limit the fm bandwidth.

signal mixers

Commercial mixers that give good performance up through 2304 MHz are available at reasonable prices. Simple "rat race" mixers can be made on Teflon pc board for all bands up to and including 10 GHz; they don’t work as well at 24 GHz and above because of packaged diode size and parasitics. A diode mixer with less than 7-dB conversion loss at 10368 MHz (with 10080-MHz local oscillator injection) can be cut out of a piece of circuit board. This by itself (no amplifier, preamplifier, or transmit/receive switch) can give S9 signals between similar stations with 4-foot dishes separated by 10 miles!

The 10-GHz signal mixer uses the same diode ring and board material as the 10-GHz harmonic downconverter. Building it on the same piece of board material eliminates two connectors and some coax along with their associated losses. A balun is used to match the mixer diode’s input impedance to 50 ohms. You can make this balun from two toroidal cores, or use a VHF TV 300-to-75 ohm balun. Conversion loss of under 10 dB should be possible over a range of local oscillator powers. Low barrier diodes are indicated in the parts list, but medium and high barrier may be substituted if sufficient 10080-MHz oscillator power is available. Higher drive levels make higher i-f levels possible on transmit, and therefore higher 10368-MHz transmit power. To avoid serious distortion, i-f power should generally be kept at least 10 dB below the available local oscillator power. A close-up of the 10-GHz signal mixer is shown in fig. 19.

Build the signal mixer as part of the downconverter assembly, and you can be on the air as soon as the 10-GHz oscillator is locked and you have a suitable SSB i-f transceiver. Just hook it through a bandpass filter to your antenna!

Part 3 will discuss the following: a 260-MHz locked oscillator along with amplifiers and switching for the 280-290 MHz i-f transverter; and a two-stage, 16-dB gain, 2.5-dB noise figure 10-GHz amplifier that can be used on transmit and receive. Two such stations connected to modest size antennas should improve your DX possibilities and could help you break the current world 10-GHz DX record!

---

**ham radio**
I've discussed VHF/UHF/microwave and millimeter-wave radio propagation many times in this column, yet there is always more new material available. This month we will try to pick up where reference 3 left off and update the present state of the art (SOA) of radio propagation above 50 MHz.

DX records

For years I have felt that the greatest incentives for experimentation on the frequencies above 50 MHz are discovering new propagation modes and setting new DX records. However, published DX records were either scattered or incomplete and often without any mention of the propagation mode used. Most of the published records were worldwide, tending to favor regions where special geography or phenomena are present.

Several years ago I started publishing consolidated VHF/UHF/microwave and millimeter-wave DX records in "VHF/UHF World." At first only the more available worldwide records were included. Later EME (Earth-Moon-Earth) records were added.

I next published a list including only those DX records where at least one of the stations was located in North America. As a new twist, the suspected propagation mode was added. This made for many new DX opportunities above 50 MHz.

The "North America Only" list caught on like wildfire. Many new DX record claims were documented and other propagation modes on different frequency bands were added. These records have been published at least once a year in this column; we now publish new record claims at the end of each "VHF/UHF World."

This month is no exception. All three record tables have been updated. Table 1 shows the North America Only terrestrial records, table 2 lists the worldwide terrestrial records, and table 3 the worldwide EME records.

Each claim has been documented by at least one of the record holders. To facilitate new claims, I designed the VHF/UHF/SHF Record Verification Form in table 4. The form verifies when the claimed contact took place and shows the equipment required to make the record. The latter is particularly important since it sets the minimum equipment specifications required.

frequency bands

The list of frequencies available to Amateurs under FCC jurisdiction was published in reference 2; the microwave and millimeter frequencies were later updated and appeared in reference 3. There haven't been any changes of late.

However, there are some further frequency restrictions. The band most affected is the 70 cm (420-450 MHz). Any United States station operating within 100 miles of any PAVE PAWS radar installation and running more than 50 watts is required to obtain FCC permission. This currently affects Amateurs in New England, Georgia, Texas, Alaska, and California.

Amateurs operating in the 70-cm band near the missile test ranges in California, Florida, and New Mexico are also affected by the new rules. There have been additional restrictions placed on Amateurs operating in the 420-430 MHz region near the Canadian border and some Canadians are now affected in the 430-450 MHz region near airports using experimental wind-shear radar. These rules seem to be in a state of flux.

At the present time, Amateur restrictions on 33 cm (902-928 MHz) are
### Table 1. North American VHF and Above Claimed DX Records. (Notes 1, 2 & 3)

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Record Holders</th>
<th>Date</th>
<th>Mode</th>
<th>DX Miles (km)</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>50 MHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>EME</td>
<td>WA4NJP (EM84DG) - KH6HI (BL01XH)</td>
<td>88-02-15</td>
<td>CW</td>
<td>4530 (7289)</td>
</tr>
<tr>
<td><strong>144 MHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Aurora</td>
<td>KA1ZE (FN31TU) - WB0DRL/ WA0TKJ (EM18CT)</td>
<td>96-02-08</td>
<td>CW</td>
<td>1347 (2167)</td>
</tr>
<tr>
<td>Ducting</td>
<td>KH6GRU (BL01XH) - WA6JRA (DM13BT)</td>
<td>73-07-29</td>
<td>CW</td>
<td>2686 (4161)</td>
</tr>
<tr>
<td>EME</td>
<td>VE1UT (FN63XV) - VK5MC (QFO2EJ)</td>
<td>84-04-07</td>
<td>CW</td>
<td>10,985 (17676)</td>
</tr>
<tr>
<td>Spor. E</td>
<td>KD4WF (EM92LA) - NW70/7 (DM25GV)</td>
<td>87-06-14</td>
<td>SSB</td>
<td>1980 (3186)</td>
</tr>
<tr>
<td>FAl</td>
<td>W5HUQ/4 (EM90GC) - W5JUN (DM82WA)</td>
<td>83-07-25</td>
<td>CW</td>
<td>1228 (1976)</td>
</tr>
<tr>
<td>MS</td>
<td>K5UR (EM25WA) - K4EKG (FK68VG)</td>
<td>85-12-13</td>
<td>SSB</td>
<td>1960 (3153)</td>
</tr>
<tr>
<td>TE</td>
<td>KP4EOR (FK78AJ) - LU5DZJ (GF11LU)</td>
<td>78-02-12</td>
<td>SSB</td>
<td>2833 (4528)</td>
</tr>
<tr>
<td>Tropo</td>
<td>K1RJH (FN31XH) - K5WXZ (EM120W)</td>
<td>68-10-08</td>
<td>CW</td>
<td>1468 (2362)</td>
</tr>
<tr>
<td><strong>220 MHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Aurora</td>
<td>W3IY/4 (FM19HA) - WB5LUA (EM13QC)</td>
<td>82-07-14</td>
<td>CW</td>
<td>1145 (1842)</td>
</tr>
<tr>
<td>Ducting</td>
<td>KH6JK (BL11AQ) - W6NLZ (DM03TS)</td>
<td>59-06-22</td>
<td>CW</td>
<td>2639 (4086)</td>
</tr>
<tr>
<td>Spor. E</td>
<td>K6UGM (EM12MS) - W6HUQ/4 (EM90GC)</td>
<td>87-06-14</td>
<td>CW/SSB</td>
<td>932 (1499)</td>
</tr>
<tr>
<td>EME</td>
<td>K1WHG (FM43MK) - KH6HFZ (BL11CN)</td>
<td>83-11-17</td>
<td>CW</td>
<td>5058 (8128)</td>
</tr>
<tr>
<td>MS</td>
<td>K1WHG (FM43MK) - KOALL (EN16NNW)</td>
<td>86-08-12</td>
<td>SSB</td>
<td>1279 (2067)</td>
</tr>
<tr>
<td>TE</td>
<td>KP4EOR (FK78AJ) - LU7DZJ (GF05RJ)</td>
<td>83-03-09</td>
<td>CW/SSB</td>
<td>3670 (5906)</td>
</tr>
<tr>
<td>Tropo</td>
<td>VE3EMS (EN86QJ) - WB5LUA (EM13QC)</td>
<td>82-09-28</td>
<td>SSB</td>
<td>1181 (1901)</td>
</tr>
<tr>
<td><strong>432 MHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Aurora</td>
<td>W3IP (FM19PD) - WB5LUA (EM13QC)</td>
<td>86-02-08</td>
<td>CW</td>
<td>1182 (1901)</td>
</tr>
<tr>
<td>Ducting</td>
<td>KD6R (DM13NI) - KH6JAA/P (BK29GO)</td>
<td>80-07-28</td>
<td>CW</td>
<td>2550 (4103)</td>
</tr>
<tr>
<td>EME</td>
<td>K2UYH (FN20QG) - VK6FL (QF78VB)</td>
<td>83-01-29</td>
<td>CW</td>
<td>11,567 (18612)</td>
</tr>
<tr>
<td>MS</td>
<td>W2AZL (EN20VI) - W5LER (EN35IA)</td>
<td>72-08-12</td>
<td>CW</td>
<td>1019 (1640)</td>
</tr>
<tr>
<td>Tropo</td>
<td>WB5CZG (FN21AX) - WASJQB (EM12LQ)</td>
<td>86-11-29</td>
<td>SSB</td>
<td>1318 (2121)</td>
</tr>
<tr>
<td><strong>903 MHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>EME</td>
<td>K5JL (EM15DQ) - WB5LUA (EM13QC)</td>
<td>88-02-07</td>
<td>CW</td>
<td>187 (301)</td>
</tr>
<tr>
<td>Tropo</td>
<td>W2PGC (FN02OR) - K5SIW9 (EN52WA)</td>
<td>86-12-24</td>
<td>SSB</td>
<td>478 (769)</td>
</tr>
<tr>
<td><strong>1296 MHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Ducting</td>
<td>KH6HME (BK29GO) - WB6NMT (DM12KU)</td>
<td>86-08-13</td>
<td>CW</td>
<td>2528 (4068)</td>
</tr>
<tr>
<td>EME</td>
<td>K2UYH (FN20QG) - VK5MC (QF05EJ)</td>
<td>81-12-06</td>
<td>CW</td>
<td>10,562 (16995)</td>
</tr>
<tr>
<td>MS</td>
<td>W2AZL (EN20VI) - W5LER (EN35IA)</td>
<td>72-08-12</td>
<td>CW</td>
<td>1019 (1640)</td>
</tr>
<tr>
<td>Tropo</td>
<td>WB3CZG (FN21AX) - KD5RO (EN13PA)</td>
<td>86-11-29</td>
<td>CW</td>
<td>1287 (2070)</td>
</tr>
<tr>
<td><strong>2304 MHz</strong></td>
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<td></td>
</tr>
<tr>
<td>EME</td>
<td>W3WI/8 (FM08CK) - ZL2AQE (RE78JS)</td>
<td>87-10-18</td>
<td>CW</td>
<td>8658 (13931)</td>
</tr>
<tr>
<td>Tropo</td>
<td>KD5RO (EN13PA) - WBY10 (EN82BE)</td>
<td>86-11-29</td>
<td>CW</td>
<td>940 (1513)</td>
</tr>
<tr>
<td><strong>3456 MHz</strong></td>
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<td></td>
<td></td>
</tr>
<tr>
<td>Tropo</td>
<td>WA5TNY/5 (EM11AU) - WB5LUA/5 (EM24UQ)</td>
<td>86-10-19</td>
<td>CW</td>
<td>288 (464)</td>
</tr>
<tr>
<td>EME</td>
<td>W7CNK/5 (EM15FI) - KOE0/0 (DM79NO)</td>
<td>87-04-12</td>
<td>CW</td>
<td>498 (802)</td>
</tr>
<tr>
<td><strong>5760 MHz</strong></td>
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<tr>
<td>Tropo</td>
<td>K5PJR (EM2650P) - W5UGO/0 (EN00PH)</td>
<td>87-07-04</td>
<td>CW</td>
<td>332 (535)</td>
</tr>
<tr>
<td>EME</td>
<td>WA5STNY (EM12KV) - W7CNK/5 (EM15FI)</td>
<td>87-04-24</td>
<td>CW</td>
<td>174 (279)</td>
</tr>
<tr>
<td><strong>10.368 GHz</strong></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Tropo</td>
<td>N6GN/6 (CM89PX) - W6SFH/6 (DM94MS)</td>
<td>87-07-19</td>
<td>CW</td>
<td>414 (666)</td>
</tr>
<tr>
<td><strong>24.192 GHz</strong></td>
<td></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>LOS</td>
<td>WA3RMX/7 (CN93IQ) - WB7TNU/7 (CN95DH)</td>
<td>86-08-23</td>
<td>SSB</td>
<td>116 (186)</td>
</tr>
<tr>
<td><strong>47.040 GHz</strong></td>
<td></td>
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</tr>
<tr>
<td>LOS</td>
<td>WA3RMX (CN85PL) - WB7TNU/7W7TYR (CN85NH)</td>
<td>87-03-08</td>
<td>SSB</td>
<td>13.9 (22.4)</td>
</tr>
<tr>
<td><strong>76-149 GHz</strong></td>
<td></td>
<td></td>
<td></td>
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</tr>
<tr>
<td>LOS</td>
<td>None reported</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>474 THz</strong></td>
<td></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>LOS</td>
<td>K6MEP (DM040I) - WA6EJO (DM04KT)</td>
<td>79-06-09</td>
<td>LASER</td>
<td>15 (24)</td>
</tr>
</tbody>
</table>

Note 1. The records are listed alphabetically by mode. Ducting is suspected where the path is mostly over water. No efforts are made to separate out ducting on overland paths so they’re grouped under tropo.

Note 2. The information within the parentheses ( ) following the callsign is the grid square locator.

Note 3. Distances have been calculated assuming a spherical earth model using the actual latitude and longitude rather than grid square centers which are less accurate.

Note 4. Six-meters records, excepting EME, were left off since the primary propagation mode is often hard to distinguish. Also long-path QSOs have been reported during solar cycles 19 and 21 which exceed approximately 12,430 miles.
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Table 2. Worldwide Claimed VHF/UHF/SHF Terrestrial DX Records (notes 1 & 2)

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Record Holders</th>
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<th>Mode</th>
<th>DX Miles (km)</th>
</tr>
</thead>
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<td>88-02-15</td>
<td>CW</td>
<td>4530 (7289)</td>
</tr>
<tr>
<td>144 MHz</td>
<td>K6MYC/K76 (BK29AO)-ZS64LE (KG43RC)</td>
<td>83-02-18</td>
<td>CW</td>
<td>12,091 (19455)</td>
</tr>
<tr>
<td>220 MHz</td>
<td>K1WHS (FN43MK)-K7686 (BL11CJ)</td>
<td>83-11-17</td>
<td>CW</td>
<td>5058 (8139)</td>
</tr>
<tr>
<td>432 MHz</td>
<td>F9FT (JO29AG)-Z3LAAAD (RE66GR)</td>
<td>80-04-18</td>
<td>CW</td>
<td>11,679 (18793)</td>
</tr>
<tr>
<td>900 MHz</td>
<td>K5JL (EM15DO)-WB5LUA (EM13OC)</td>
<td>88-02-07</td>
<td>CW</td>
<td>187 (301)</td>
</tr>
<tr>
<td>1296 MHz</td>
<td>PA0S6 (JO11IW)-Z3LAAAD (RE66GR)</td>
<td>83-06-13</td>
<td>CW</td>
<td>11,595 (18657)</td>
</tr>
<tr>
<td>2304 MHz</td>
<td>W3IWI-8 (FM080C)-ZL2AOE (RE78JS)</td>
<td>87-10-18</td>
<td>CW</td>
<td>8685 (13931)</td>
</tr>
<tr>
<td>3456 MHz</td>
<td>W7CNIK/5 (EM15JF)-K06 (DM75NO)</td>
<td>87-04-06</td>
<td>CW</td>
<td>498 (802)</td>
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<td>5670 MHz</td>
<td>WA5TNY (EM12KJ)-W7CNIK/5 (EM15JF)</td>
<td>87-04-24</td>
<td>CW</td>
<td>174 (279)</td>
</tr>
<tr>
<td>10,000 MHz</td>
<td>None reported</td>
<td>None reported</td>
<td>None reported</td>
<td>None reported</td>
</tr>
</tbody>
</table>

Notes:
1. The information within the parentheses ( ) after the callsign is the grid square locator.
2. The distances shown have been calculated assuming a spherical earth model. The actual latitudes and longitude are used rather than grid square centers which are less accurate.

Table 3. Worldwide Claimed VHF/UHF/SHF EME DX Records (notes 1 & 2)

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Record Holders</th>
<th>Date</th>
<th>Mode</th>
<th>DX Miles (km)</th>
</tr>
</thead>
<tbody>
<tr>
<td>50 MHz</td>
<td>WA4JNP (EM84DG)-K16H1 (BL01XH)</td>
<td>88-02-15</td>
<td>CW</td>
<td>4530 (7289)</td>
</tr>
<tr>
<td>144 MHz</td>
<td>K6MYC/K76 (BK29AO)-ZS64LE (KG43RC)</td>
<td>83-02-18</td>
<td>CW</td>
<td>12,091 (19455)</td>
</tr>
<tr>
<td>220 MHz</td>
<td>K1WHS (FN43MK)-K7686 (BL11CJ)</td>
<td>83-11-17</td>
<td>CW</td>
<td>5058 (8139)</td>
</tr>
<tr>
<td>432 MHz</td>
<td>F9FT (JO29AG)-Z3LAAAD (RE66GR)</td>
<td>80-04-18</td>
<td>CW</td>
<td>11,679 (18793)</td>
</tr>
<tr>
<td>900 MHz</td>
<td>K5JL (EM15DO)-WB5LUA (EM13OC)</td>
<td>88-02-07</td>
<td>CW</td>
<td>187 (301)</td>
</tr>
<tr>
<td>1296 MHz</td>
<td>PA0S6 (JO11IW)-Z3LAAAD (RE66GR)</td>
<td>83-06-13</td>
<td>CW</td>
<td>11,595 (18657)</td>
</tr>
<tr>
<td>2304 MHz</td>
<td>W3IWI-8 (FM080C)-ZL2AOE (RE78JS)</td>
<td>87-10-18</td>
<td>CW</td>
<td>8685 (13931)</td>
</tr>
<tr>
<td>3456 MHz</td>
<td>W7CNIK/5 (EM15JF)-K06 (DM75NO)</td>
<td>87-04-06</td>
<td>CW</td>
<td>498 (802)</td>
</tr>
<tr>
<td>5670 MHz</td>
<td>WA5TNY (EM12KJ)-W7CNIK/5 (EM15JF)</td>
<td>87-04-24</td>
<td>CW</td>
<td>174 (279)</td>
</tr>
</tbody>
</table>

Notes:
1. The information within the parentheses ( ) after the callsign is the grid square locator.
2. The distances shown have been calculated assuming a spherical earth model. The actual latitudes and longitude are used rather than grid square centers which are less accurate.

still in effect in Colorado, Wyoming, White Sands Missile Range, and Region 3 areas. Operators in these restricted areas who have tried to obtain permission from the FCC have been unable to do so. Canadian Amateurs need special permission from DOC to use CW or SSB on this band. (It is presently designated as fm only!)

solar cycle update

Probably one of the hottest discussions on the hf and 6-meter bands these days is "when will the next solar cycle peak?" Near the sunspot peak there is a chance that F2 propagation will be possible on 6 meters. News of that peak is starting to come in. The new cycle, 22, has definitely begun and ended when it bottomed out at a smoothed sunspot count of Cycle 21 level of 118.6 sunspots in mid-1991. Because this cycle started statistically like the Sargent/Ohl were very close in predicting the peak of cycle 21. Based on available data and using this prediction method, it now appears that cycle 22 will peak at a smoothed level of 118.6 sunspots in mid-1991. Figure 1 shows this early data along with the final data on cycle 21.

The predicted peak of cycle 22 shows that it will be very flat and should stay above 100 sunspots from about July 1989 through June 1992.
fig. 1. This graph shows the smoothed sunspot numbers for solar cycle 21 as well as the forecasted numbers for cycle 22, per reference 8.

earlier than other cycles and rose abruptly, we will have to wait at least another year or so to see what if any modifications will occur.

This information is not very promising for 6-meter Amateurs as it usually takes a sunspot count above 150 to yield good F2 openings. However, minor sunspot peaks often occur during a cycle, albeit of short duration. No 6-meter operator active during the last solar cycle will ever forget the solar peaks in late 1979 that rivaled those of all previously recorded solar cycles.

The equivalent short-term sunspot number can be predicted using the solar flux measured at Ottawa on 10.7 cm. The value is updated daily and broadcast at 18 minutes after each hour on radio station WWV. Using the equation shown in reference 2, I have prepared fig. 2 which can be used to determine the equivalent sunspot number on any day. Remember also that the ionosphere usually has to be "pumped up" for four or five consecutive days to yield good long-haul F2 propagation.

The SOA in equipment, antennas, and propagation forecasting has greatly improved in recent years. Predictions of the MUF are now possible with improved accuracy using personal computer programs like MINIMUF® in conjunction with the sunspot number per fig. 2.

This information, along with increased 6-meter interest and improved operating methods (more on this shortly), as well as recent relaxations in licensing restrictions in western Europe and North Africa, means that there will be many more regions and DXCC countries represented during cycle 22. Let's hope it's a great cycle for 6-meter operators. Stay tuned!

fig. 2. This figure shows the correlation between solar flux and sunspot numbers.

backscatter

The backscatter form of propagation described in reference 1 is basically a form of reflection and indicates that a highly ionized region is present. Operators detecting this phenomenon can often work DX by aiming their antennas in the direction of the ionized region and "backscattering" their signals.

Backscatter also indicates that the MUF is very high; it was well used during solar cycle 21 to indicate the presence of an opening. Often western United States stations could work Hawaii while eastern stations could work Europe either by backscatter or by knowing that there was a high degree of probability of an opening in progress.

Ionospheric scatter

Ionospheric scatter was also described in reference 1. It is a form of "forward scatter" linked to the time
Table 4. VHF/UHF/SHF Propagation Record Verification Form.
(Please return to Joe Reisert, W1JR, 17 Mansfield Drive, Chelmsford, MA 01824)

<table>
<thead>
<tr>
<th>Band</th>
<th>Propagation Mode</th>
<th>Date of record (UTC):</th>
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<tbody>
<tr>
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<table>
<thead>
<tr>
<th>Time of record (UTC):</th>
<th>DX (miles)</th>
<th>(km):</th>
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<tbody>
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</table>

**Station 1**

<table>
<thead>
<tr>
<th>Call</th>
<th>Name</th>
<th>QTH for this QSO:</th>
<th>Lat*: Long*:</th>
</tr>
</thead>
<tbody>
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<tr>
<th>Grid Locator (6 digit):</th>
<th>Elevation ASL (feet):</th>
<th>(meters):</th>
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<tr>
<th>Location description:</th>
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<table>
<thead>
<tr>
<th>Antenna type:</th>
<th>Estimated gain (dBi):</th>
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<thead>
<tr>
<th>TX freq:</th>
<th>TX power:</th>
<th>Feedline loss:</th>
<th>Modulation type:</th>
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<tr>
<th>RX freq:</th>
<th>RX type:</th>
<th>RX desc:</th>
<th>Feedline loss:</th>
<th>Noise figure:</th>
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<tr>
<th>RX bandwidth:</th>
<th>Rcvd signal to noise ratio:</th>
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<tr>
<th>Other equipment description:</th>
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**Station 2**

<table>
<thead>
<tr>
<th>Call</th>
<th>Name</th>
<th>QTH for this QSO:</th>
<th>Lat*: Long*:</th>
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<th>(meters):</th>
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<th>Other equipment description:</th>
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**Other comments, weather conditions etc:**

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The information submitted above is correct to the best of my knowledge.

Submitted by: _____________________________ Call: __________ Home QTH:

_______________________________ Phone number (IAC _______ )

Record received and verified by: _____________________________ Date: __________

*Please list latitudes and longitudes in degrees, minutes, and seconds.*
of day (typically peaking broadly around noon local time) and to high sunspot activity. Ionospheric scatter can be used on 6 and 2 meters.

This form of propagation is used extensively in commercial service but seems to have been almost totally ignored by Amateurs. It does require reasonable antenna gain and high power, but is within the reach of many well-equipped 2-meter Amateur stations, especially those with EME or marginal EME capability.

As the sunspots increase, so will the possibilities of forward scatter. This represents an interesting challenge for Amateurs and is a good way to increase their grid square count in the 800–1300 mile region.

**TE (transequatorial) scatter**

Like forward and back scatter, TE propagation is a good mode for long DX, especially on 6 meters. It is best observed across the equator on more typical distances of 3 to 6,000 miles.

TE propagation is most often observed in the late afternoon and early evening for several weeks around the equinoxes. During the peak of the solar cycle around this time, highly ionized “patches” are often present approximately 10-20 degrees north and south of the “geomagnetic” equator.

Unfortunately, the geomagnetic equator is very far south in the North and South American sector. This limits North American TE propagation mainly to stations in the Caribbean and the extreme southern portions of the United States. Don’t let this discourage you; there are always new propagation modes and isolated openings to explore.

**equatorial FAI (Field Aligned Irregularities)**

Equatorial FAI was first discovered in 1977. It is still not fully understood and often referred to as TE propagation (see references 11 and 12). Like TE scatter, equatorial FAI depends on highly ionized patches that are typically located 10-15 degrees north and south of the geomagnetic equator at the same dates and times discussed under TE propagation above. However, the DX is slightly less. The most favored locations are paths from southern Europe to South Africa, Japan to Australia, and the Caribbean to southern South America.

While TE scatter is generally limited to below 100 MHz, equatorial FAI has been known to extend higher in frequency. Contacts as high as 220 MHz have been confirmed as shown in tables 1 and 2. Although some one-way 432 reports have been reported, I have been unable to document any two-way contacts above 220 MHz. Maybe during the peak years of solar cycle 22 the frequency barrier will be broken and two-way 432 MHz contacts will be completed. Any takers?

**midlatitude FAI**

When reference 10 was written, it was speculated that FAI propagation would be possible in mid-northern latitudes. It didn’t take long before this became a reality on 144 MHz.13

Midlatitude FAI propagation has many similarities to auroral propagation; both stations must be south of the ionized region and aim their antennas several degrees north of the great circle path. This type of propagation most often occurs in the evenings during the summer, especially on days when there has been sporadic E propagation on 6 meters. As shown in table 1, it has been successfully used out to a distance of just over 1200 miles on 2 meters. Until recently FAI has been slow to take hold, despite the fact that it should be usable up through 220 MHz.13

This is all changing now — contacts were reported during the summer of 1987 in the southern United States and a wide region of Europe (reference 14 through 16). In fact, well over 500 European contacts were reported using midlatitude FAI propagation during the summer period of 1986 alone! (See reference 15.)

As observed in Europe, the scatter region tends to be at the same height as sporadic E, typically 70 miles. These regions resemble aurora propagation; unlike the relatively small (1-2 mile thick) sporadic E clouds, they have large volume areas.

Midlatitude FAI signals tend to have rapid fading. They have been observed over several European locations but mostly along the 45-55 degree north latitude lines following the contours shown in reference 13. Those who can elevate their antennas have a greater possibility of success.

Midlatitude FAI propagation offers a great challenge to VHF Amateurs, especially in North America. This mode of propagation should be usable up through 220 MHz throughout the contiguous 48 United States. All it takes is some patience and a surge in activity. Who will be the first to report a 220-MHz midlatitude FAI QSO? It’s there for the asking!

**summary**

This month I’ve given you a status report on the latest DX records on the VHF/UHF/microwave bands. We’ve also discussed the latest prognosis for propagation using the solar cycle 22 peak and some scatter modes. Next month’s column will update other propagation modes. Until then, you can read the references cited.

**new DX records**

This has been a good month for new VHF/UHF DX records. First off, the 6-meter EME record has been extended. On February 15, 1988 between 1800-1845 UTC, Ray Rector, WA4NJP, Gillsville, Georgia (EM84DG) completed a two-way EME contact with Bert Ingalls, KH6HI, Ewa Beach, Hawaii (BL01XH), on 50.008 MHz using 1-minute sequencing. The distance was approximately 4530 miles (7289 km). Ray was using 1500 watts and Bert was running 1000. Both stations were using quads of four eight-element Yagis on 35-38 foot booms. Congratulations to Ray and Bert — 6 meters is now buzzing with EME activity.

Last month we reported the first ever 33-cm (902 MHz) EME QSO. That record didn’t last very long! On February 7, 1988 at 0500 UTC, Jay Lieb-
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OTHER CONFERENCES

Mid-Atlantic VHF Conference. This conference was sponsored by the Mt. Airy VHF Radio Club, Oct. 10-11, 1987. 11 papers cover everything from mountain topping to transceivers for the 3400 and 5600 MHz bands. 120 pages. $10.

MICROWAVE UPDATE 1987 held in Estes Park, Colorado, September 10-13, 1987. 17 papers on equipment, antennas and techniques for 902 MHz through 10 GHz. Much information on construction of 2.3, 3.4 and 5.7 GHz gear. 136 pages. $10.

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Tip: To increase the range of your hand-held scanner or transceiver, connect a Grove ANT-8 extendable whip antenna, equipped with standard BNC base.
mann, K5JL, Piedmont, Oklahoma (EM15DQ) completed a 33-cm EME QSO with AI Ward, WB5LUA, McKinney, Texas (EM13QC) over a distance of approximately 187 miles (301 km). Both stations were running approximately 150 watts and 24-28 foot dishes. Congratulations, Jay and Al. It looks as if 33-cm EME activity is just about to take off. Both records just discussed are included in tables 1 and 3.

Finally (although not yet a DX record), during February 1988 Rick Fogle, Grapevine, Texas (EM12KV) has been heard by Lucky Whitaker, W7CNK, Oklahoma City, Oklahoma (EM15FI) via 3-cm (10,368.1 MHz) EME. Likewise, Lucky has heard Rick via the same path. Rick uses a 10-foot dish and Lucky a 16 footer. Both have their preamplifiers and power amplifiers mounted right at the feed. Unfortunately, they have only one high-power (10-15 watt) TVT amplifier between them which they mail back and forth. Because of this, they can’t complete what is considered a conventional two-way QSO (use of two complete sets of gear all used during one operating session). Efforts are underway to get a second power amplifier. Good luck to Rick and Lucky as well as the other 3-cm operators who are also trying to conquer this elusive band. It seems that one of the last EME frontiers is about to be conquered.

**Important VHF/UHF Events**

### June
- **4** EME perigee
- **7** Predicted peak of the daytime Arietids meteor shower at 0150 UTC
- **9** Predicted peak of the Zeta Perseids meteor shower at 1020 UTC
- **11-13** ARL June VHF QSO Party
- **14** New moon
- **16-18** SMIRK (Six Meter International Radio Klub) Party Contest (contact KA9NNO)
- **21** ±1 month. Peak of midlatitude Sporadic E propagation
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- **28** Predicted peak of the Delta Aquarids meteor shower at 2100 UTC
- **30** EME perigee

### References
the Quad antenna: part 2, circular and octagonal loops

Analysis shows good performance with similar data.

Although there is a contradiction in terms, it is convenient to consider the circular loop, and arrays built of loops, as the first members of the Quad family. One reason is that all other versions can be regarded as departures from the 'ideal' circular figure. To the extent this is true, the performance of circular-loop antennas is thus representative of the performance of the entire family.

Another reason is that the theoretical analysis of the circular loop is far more advanced than for the other shapes. Extensive tables of calculated characteristics have been published, some with comparisons of measured performance. In contrast, while there are theories of square Quad loop and array performance, their complexity makes them impractical for calculation, even on mainframe computers.

theoretical basis of circular-loop analysis

As shown in fig. 1, only two quantities are necessary to specify the circular-loop antenna: the conductor size, usually given as its radius; and the loop size, also described by radius. It is often more convenient to use two derived descriptive quantities in theoretical discussion. The first is the normalized circumference of the loop at the specific frequency of interest given by the quantity kb, where k is defined as

\[ k = 2\pi/\lambda, \]

By R.P. Haviland, W4MB, 1035 Green Acres Circle North, Daytona Beach, Florida 32019

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**fig. 1.** A circular loop can be described by two quantities, the radius of the conductor (a) and the radius of the loop (b). For work, it is convenient to use two derived quantities, a conductor thickness factor (omega) and the circumference in wavelengths at the operating frequency (kb).
\( \lambda \) being the wavelength. The quantity \( kb \) is therefore the circumference of the loop in wavelengths. Conductor size is usually given by the relationship

\[
\Omega = 2\ln \left( \frac{2\pi b}{a} \right)
\]

where \( \ln \) is the natural logarithm, equal to 2.3 times the more common \( \log_{10} \) value. The value of \( \Omega \) is given in fig. 2 as a function of the ratio \( 2\pi b/a \), or loop circumference to conductor radius. Values less than 10 represent very large conductor diameters, and those over 20 very small conductors. High-frequency antennas will usually have values in the range 20-25, and self-supporting ultrahigh frequency antennas values of 10-15.

The loop is assumed to be fed at one point, usually taken as the angle reference. This induces a current, \( i \), in the loop at angle zero which, in turn, creates a field at the point designated by \( R,01 \), for example. The total field at this point is the sum of the fields produced by all points on the loop.

The field components also induce current flow in the loop. When equilibrium is reached (after a few rf cycles) the field close to the conductor must lie only at right angles to it. (If there had been a tangential component, a change in the current would have been induced, so equilibrium would not yet have been attained.) This observation plus standard field equations give the conditions for calculating current distribution, and therefore the drive impedance and radiation pattern.

While the concept is relatively simple, the mathematical operations are difficult. See the references, especially Storer,\(^1\) for details. For our purposes it is sufficient to note that the current distribution is given by:

\[
I(\theta) = \frac{V}{i \cdot \pi \cdot 377} \left[ \frac{1}{A_n} + 2 \cdot \sum \cos\left(n \cdot \theta\right) \frac{A_n}{A_n} \right]
\]

where the sum is for all values of \( N \) from 1 to infinity. This result was derived by Hallen.\(^2\)

This equation is simple, but its evaluation is complex. The quantities \( A \) involve series for which exact solutions are unknown. Even approximate solutions require further assumptions, two being that the conductor diameter is small compared to loop diameter and to operating wavelength. This restriction is satisfied by practical antennas.

Additionally, the infinite series in the equation tends to zero as the ratio of loop circumference to conductor radius increases. Practical self-supporting loops need an \( \Omega \) around 10-12 to have sufficient strength. Wire cage elements may be used to secure low \( \Omega \) factors at low frequencies.

**Fig. 2.** Values of the thickness factor (\( \Omega \)) as a function of the ratio of loop to conductor radius (or diameter). Practical self-supporting loops need an \( \Omega \) around 10-12 to have sufficient strength. Wire cage elements may be used to secure low \( \Omega \) factors at low frequencies.

**Fig. 3.** Current magnitude and phase on a large conductor circular loop one wavelength in circumference as derived by Storer.\(^3\)\(^,\)\(^4\) The loop is below resonance. The current on a conductor of essentially 0 radius is shown for comparison. Derived from transmission line theory, this is the first approximation to the current on any loop of the same circumference.
to become divergent when the number of terms is reduced for reasonable scale of calculation. Storer\textsuperscript{3,1} developed a method of calculation by keeping the first four terms of the series and replacing the remainder by an integral. He also published a set of ten curves giving the real and imaginary components of the series. With these, evaluation of the current distribution and drive impedance is reduced to some simple (but tedious) curve measuring and complex-number algebra. Unfortunately, the curves cannot be reduced to simple equation form, so this can't be avoided.

Rather than presenting these curves and usage details here, I will give only the results of examination of some specific loop designs. Subsequent analyses have given a table of drive impedances, which is more accurate for most work. These values are covered later.

For values outside the range given here, or to obtain the current distribution, you will need the Storer curves. (The reference 3 version is best, and is available at reasonable cost from Cruft Laboratory Library, Harvard University, Cambridge, Massachusetts.)

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simplification of the general equation above, with only three terms giving adequate accuracy. Approximate equations for each are given, and are easily handled by small computers. Note that the reactance for these small antennas is determined almost entirely by the loop circumference, with almost no effect on conductor size. In contrast, the input resistance varies with both.

current distribution on a circular loop

Figure 3 shows the magnitude of the current on a one-wavelength circular loop as calculated by Storer, together with the current magnitude on an ideal shorted transmission line made from one wavelength of conductor. Figure 3B shows the phase with respect to the driving voltage for both cases.

A number of important factors concerning the entire Quad family show on these curves. The first is that a one-wavelength loop is not resonant, as indicated by the fact that the angle of the drive impedance is not zero. Since the loop appears as a capacitance, it is below resonance. Unlike dipoles, loops must be longer than a wavelength to be resonant.

A second factor is that there is no point on the loop where the current goes to zero. Associated with this is the fact that the current at 180 degrees from the feedpoint is appreciably less than at the driving point. Similarly, these two currents are not 180 degrees out of phase, but somewhat more. The point of phase reversal is not at 90 degrees to the line of symmetry through the feedpoint.

One reason for the differences between the loop and the ideal shorted line is the greater separation of the sides. Currents are not constrained to be equal because of tight coupling, as in the ideal shorted line. Further, power is being radiated, causing a reduction in current when moving away from the feedpoint. (In a practical antenna, the current differences would be somewhat greater, as the analysis assumes zero ohms loss.)

These current curves show that the usual evaluation of a Quad — two separated dipoles with ends bent to touch — can’t fully describe the performance. This simple concept is useful in verbal descriptions, and can be a valuable tool in approximate analysis. But it must be remembered that numerical results are probably in error by a factor at least as large as the current error, or at least 20 percent. The effect of the error should be smallest for pattern calculations and drive resistance, but is likely to be sizeable for drive reactance and resonant frequency.

These detail current calculations are for loops with relatively large conductors, \( \Omega = 10 \). Other studies, plus the tables presented later, show that the current magnitude and angle move progressively toward the curve for the ideal transmission line as the conductor radius becomes smaller. This means that the two-dipole approximation is likely to be better at high frequency than at ultrahigh frequency because omega is large for practical conductor sizes. The current distribution for some other loops is also given in references 1, 3, 5, and 6.
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As a first approximation, the current distribution can always be assumed to be that of an ideal transmission line of the same conductor length. A second approximation can be sketched by "rounding all sharp corners," and decreasing the current away from the feedpoint. Greater accuracy will require tedious evaluation using Storer's curves or the MININEC technique (to be discussed).

**drive-point impedance**

Get the drive-point impedance by dividing the drive-point voltage by the drive current calculated above. Storer's gives tables of this impedance for loops from 0.05 to 2.5 wavelengths in circumference, and for omega values from 8 to 12 (large diameter conductors).

In considering the general problem of loop antennas, Wu developed another method for solving Hallen's equation. This gives the same values for resistance components as Storer, but the reactance values are quite different and agree more closely with measurements on real antennas. King, Harrison, and Tingley have used Wu's theory to calculate loop drive admittances for sizes from 0.05 to 2.5 wavelengths and for omega from 10 to 20. For those who have forgotten, or never had occasion to work with admittances,

\[
\begin{align*}
z &= R + jX \\
y &= 1/Z \\
y &= g + jb
\end{align*}
\]

(See also reference 10.)

Figures 4A and 4B show feed resistance (R) and reactance (X), respectively, for loops from 0.05 to 2.5 wavelengths circumference. The two values of omega shown, 10 and 20, are representative of very high frequency and high-frequency loops.

Over this range of loop diameters, three high-impedance or parallel resonance points are noted, corresponding to 0.5, 1.5, and 2.5 wavelengths circumference. For \( \Omega = 20 \), the low-impedance, zero-reactance points correspond to serial resonances at about 1.0 and 2.0 wavelengths circumference. For \( \Omega = 10 \), there is also a serial resonance at 1.0 wavelength circumference. There is no true serial resonance for a circumference of 2.0 wavelengths. Instead, the reactance becomes low and remains low and capacitive. This is the case for all values of omega less than 11, and for all but the first serial resonance.

The resistance at the serial resonance point changes markedly with the value of omega. For omega that is large (20 or so), the resistance also varies markedly with loop diameter. But for \( \Omega = 10 \), the resistance change is much smaller. (For \( \Omega = 8 \), the change is less than 2.1 for any circumference greater than 0.6 wavelength.)

These characteristics mean that thick loops are inherently broadband antennas and relatively easy to match. However, the pattern changes described in part 1 affect the desirability of this broadband operation, as discussed later.

Figures 5A and 5B show the feed resistance and reactance for frequencies of greatest interest (those close to the first series resonance) for loops from 0.8 to 1.2 wavelengths circumference. Figure 6 is derived from fig. 5B, and shows the resonant frequency as a function of conductor size. These three curves give the information needed to design practical loop antennas and arrays and to calculate their performance. Their use is covered further on.

**gain of loop antennas**

In considering loop gain, it was noted that gain should increase as loop size increases and that there are pattern changes as size increases. Specifically, gain on the axis of symmetry becomes zero for all loops with 2, 3, or more wavelengths in conductor length. These two effects are shown in the calculated gain curve of fig. 7 (see reference 11). For a circumference of 1 wavelength, the on-axis gain is 3.4 dB above isotropic, or about 1.4 dB above a dipole. (Based on measurements, Lindsay quotes approximately 4.0 dB, and Appel-Hansen quotes 3.4 dB above isotropic.)
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Figure 7 shows that larger loops are superior from the viewpoint of gain. A calculated gain value of 4.5 dB occurs for a loop circumference of 1.5 wavelengths. Above this circumference, the gain decreases, as the on-axis lobe splits. Remember, there is no radiation on axis for a circumference of 2 wavelengths.

In smaller sizes, the on-axis gain decreases as the circumference becomes less than a wavelength, and as the pattern changes toward the doughnut pattern of a small loop. The gain is about equal to that of a dipole at a circumference of about 0.65 wavelength, and about 1.5 dB poorer than that of a dipole for a circumference of 0.5 wavelength.

This brings up two important points. First, from the viewpoint of gain, the circular loop should be designed to operate away from resonance. Considering the factors of gain, lobe shape, and feed impedance, Ito et al. recommend a design circumference of 1.2 wavelengths, for a gain of 4.2 dB.

Second, a loop has good gain performance over a wide range of frequencies. For example, a 14-MHz loop would be slightly better than a dipole on 10 MHz, and about 1.5 dB poorer on 7 MHz. The gain would be near maximum on 21 MHz, better than a dipole on 18 MHz, and about as good on 24 MHz. The on-axis gain would be poor on 28 MHz, but there would be usable radiation in two lobes.

The acceptability of operation away from resonance is affected by the difficulty encountered in feeding the loop. From fig. 4, feeding a large loop on high frequency does not appear to be an unusual problem. The feed resistance would increase, but a matching section or transformer is needed. There would be an inductive component, easily compensated by a stub. Depending on conductor size, a very high frequency loop might require only a matching transformer.

Wideband operation would be a greater problem.

Figure 8. Relative geometry of a circle and an octagon. The difference in area for the same conductor length is very small.
You'd need a complex matching section at the antenna, or your transmission line would have appreciable VSWR. The combination of open-wire line and a wide-range "Match Box" would give excellent performance. This is especially interesting as the basis of a 3.5/7/10-MHz fixed-loop design.

I have made use of this wideband capability on many occasions. A 14-MHz loop is regularly used on bands from 7 to 28 MHz. Much of my 10, 18, and 24-MHz experimental work (KK2XJM) used this antenna. Other than Teflon™ insulated 75-ohm feedline and a pi-section tuner, no special precautions were taken.

**circular-loop radiation patterns**

Published data on radiation patterns of loop antennas varies from sparse to nonexistent. The equations needed to calculate the currents and patterns are solvable on a small computer only with a lot of programming and computing time. It has been necessary to approximate further to develop the pattern data which follows.

**the MININEC antenna program**

The chosen calculation technique (sometimes called "the method of moments") is the public domain program "MININEC," from Logan and Rockway, currently in its third version. It uses essentially the same initial assumptions as the Hallen approach above. But instead of applying the geometry exactly and then making simplifying assumptions, assume that the radiator is composed of a series of straight-line sections carrying constant current.

A solution's accuracy increases as the number of sections for a given geometry is increased. While the complexity of the solution is not greatly affected, the time required for the solution increases as the square of the number of segments. An IBM PC may take several hours to run the program; a very small computer like the Commodore 64 may need 12 to 24. Even so, this approach is the best generally available, and is quite practical.

The original MININEC was written for the Apple computer. I have translated it for the C-64 and the Amiga; KA4WDK has done the same for the PC. The third version was written for the PC, and I translated it for the Amiga; it also runs on the Macintosh.

MININEC originators used various conditions to examine calculation accuracy, including analysis of loops. One series used a ten-sided polygon approximation of a circle, with two current segments per side plus three for the feed side. Agreement with theory was to within 10 percent.

A second series approximated the circle using the circumscribed polygon. This shows good agreement (± 6 percent) for susceptance, down to four sides. Equally good agreement for conductance required 16 sides. With 22 sides, agreement was within 6 percent for sizes from 0.1 to 2.0 wavelengths. MININEC solu-
tions are found to be unreliable for circumferences less than about 0.01 wavelength.

The inscribed polygon approximation is not an ideal check of solution accuracy as the number of segments decreases. It introduces two added factors which change the results. One is the loss of area. (For eight sides this amounts to about 3 percent.) This reduces the gain by about the same amount. More important, the total conductor length decreases as the number of segments decreases, by about the same percentage. This change in wire length introduces a change in reactance near the series-resonant points, and a change in resistance near parallel-resonant points. Both may be relatively large.

A better approximation occurs when the conductor length is kept constant. It is not easy to evaluate the precise error this causes, but it appears that it is no greater than the sum of the inherent error of MININEC plus the area error in the approximation.

It also appears that the inherent error will be around 5 to 10 percent if two simple rules are followed in setting up a MININEC analysis:

1. use a minimum of two segments per section of conductor,
2. use a minimum of four segments per halfwave of conductor.

Practically, MININEC gives extremely good accuracy. An antenna can easily depart from its ideal value by 20 percent or more because of neglected factors like supports, tower location, and feed-antenna interaction.

**octagonal loop**

Calculated values for an octagonal loop should be a good approximation of circular-loop values. The octagon is also a useful antenna by itself. Probably the best-known type is the “Army Loop”—really a loop operating well below resonance, incorporating both a matching system and low-loss construction. Figure 8 shows the basic factors involved, and allows visualization of the small area difference from the circular loop.

The properties of the octagon for conductor lengths close to 1 wavelength are summarized by figs. 9A and 9B for drive resistance and reactance. These should be compared with figs. 4A and 4B to see the validity of the octagon approximation to a circular loop. As expected from previous discussions, the agreement in reactance near the series-resonant points, and a change in resistance near parallel-resonant points. Both may be relatively large.

- use a minimum of two segments per section of conductor,
- use a minimum of four segments per halfwave of conductor.

Practically, MININEC gives extremely good accuracy. An antenna can easily depart from its ideal value by 20 percent or more because of neglected factors like supports, tower location, and feed-antenna interaction.

**fig. 10. Feed or self-reactance of an octagon for circumferences around one wavelength. The point of equal values occurs at the same reactance as the circular loop but at a greater circumference. Curve slopes are almost identical.**

**fig. 11. Resonant wavelength of an octagonal loop as a function of conductor thickness. The curve is similar to that of a circular loop, but differs by as much as ±6 percent.**

**fig. 12. Horizontal radiation pattern for an octagon lying in a vertical plane and fed at the bottom. Only the horizontally polarized component is shown. The pattern is very similar to that of a dipole, but the gain is higher by about 1.25 dB.**
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For resistance is good — no more than a few percent difference. The agreement for reactance is not quite as good; this is the usual finding. However, neither error is beyond the expected range.

If you build a practical octagon antenna, use figs. 10 and 11 if the antenna is to be resonant or nearly so. If it is away from resonance, the values from table 1 give a good approximation. For best accuracy, the table values should be interpolated for the actual conductor diameter and equivalent radius. Alternatively, MININEC can be used to obtain values within 5 to 10 percent error.

Patterns of circular/octagonal loops

Because the octagon has been found to be a good approximation to the circle, we can use its pattern as a close approximation to those of circular loops.

Figure 12 shows the MININEC calculated horizontal plane pattern for a 1-wavelength octagon fed at
the center of the bottom segment — that is, with horizontal polarization. The lobes are very nearly the same as those of a dipole, but are slightly narrower to produce gain. The gain is 3.4 dB, essentially that of a circular Quad.

Figure 13 shows the vertical plane pattern in the plane of the loop, for the total radiation. Figure 14 shows the horizontal plane pattern for vertical polarization. There is a major difference from dipole patterns in these two figures. The loop has an appreciable vertically polarized component, zero on axis and maximum at right angles to this. This component is not present with dipoles and its importance is not clear. Vertically polarized component, zero on axis and maximum at right angles to this. This component is not present with dipoles and its importance is not clear. It may have an adverse effect because of interference received on the vertically polarized side lobes (and back lobe in beams), or it may tend to reduce fading due to variation of incoming signals and path splitting. The component may also be a factor in the reputation of received on the vertically polarized side lobes (and back lobe in beams), or it may tend to reduce fading due to variation of incoming signals and path splitting. The component may also be a factor in the reputation of Quad loops as good performers at low height. More aspects of these questions will be discussed in other parts of this series.

Figure 15, provided for comparison, shows the calculated current on an octagon. Because the example is for a relatively thin conductor, there is close resemblance to the distribution for a transmission line.

Figure 16 compares the calculated gain of an octagon and a circular loop. Some of the variation is due to the difference in areas. The rest appears to be related to differences in the calculation method. Practically, the difference is negligible.

General conclusions from this comparison of circular and octagonal loops are that both have good performance, and that the data from one can be used for the other with little error.

We will use this last finding in part 3, which is devoted to arrays composed of circular and octagonal loops. Design data for arrays of two to twelve elements will be given.

**references**


**ham radio**

**PART 3 — COMING UP IN OUR JULY ISSUE.**
a nifty bi-square beam for 10 or 12 meters

The miserable DX conditions at the bottom of the sunspot cycle are but a bad memory. True, the higher frequency bands tend to fizzle during the summer, but they'll be back again with a bang as soon as the cooler fall months roll around.

If you're interested in DX operation on either 10 or 12 meters, you'll eventually need a beam antenna. You can work a lot of "easy" DX with a dipole, but sooner or later you'll wish you had a beam for the more exotic DX stations. An easy solution is to buy a Yagi beam kit, but it's less expensive to build your own wire beam from scratch. Here's an inexpensive beam for your consideration.

The Bi-square beam (fig. 1A) is a derivation of the so-called "Lazy-H" array, a favorite of point-to-point stations in the maritime and fixed services. The Lazy-H consists of two half-wave dipoles in phase over a similar pair of dipoles. Spacing between the top and bottom dipole pairs is a half wavelength. Proper phasing of the pairs is achieved with a transposed open-wire transmission line fed at the center of the lower pair of dipoles with a quarter-wave, open-wire stub. The feedpoint impedance at the bottom of the stub is about 220 ohms.

A more practical version of the Lazy-H antenna is the Bi-square beam, shown in fig. 1B. This arrangement requires only a single center pole support. The Lazy-H dipole pairs are connected together at the outer tips, resulting in a diamond-shaped wire arrangement. You can eliminate the transposed line connecting the center of the pairs. The quarter-wave stub is retained.

The feedpoint impedance at the bottom of the stub is close to 150 ohms. There is a reduction in feedpoint impedance because the top and bottom radiating elements of the Bi-square configuration are closer to each other than they are in the Lazy-H antenna.

The Bi-square radiation pattern is a figure eight (bidirectional) at right angles to the plane of the array. The power gain over a dipole located at the center height of the array is about 5 dB.

building the bi-square beam

The Bi-square is an easy, inexpensive beam to build. You'll need about 100 feet of No. 16 enamel or Formvar™ coated wire and four insulators. The quarter-wave stub needs five spreaders cut from 1/2-inch diameter phenolic (or plastic) rod. One of the spreaders serves as the bottom insulator for the antenna wires. The diamond-shaped
antenna is open at the top (two insulators required). Overall height is a little less than 30 feet. I hung mine from a yard arm at the 45-foot level of my crank-up tower. The proximity of the metal tower to the plane of the loop didn’t seem to cause any harm.

Dimensions for the 10- and 12-meter versions of the antenna are given in fig. 2. The sides are pulled out by ropes and tied off to convenient points on nearby trees. The bottom of the quarter-wave stub is about 7 feet above the ground.

The yard arm holds the loop about 3 feet away from the tower. The loop isn’t quite in the vertical plane because I pulled the bottom of it 6 feet away from the tower in order to reach the bottom of the stub easily from the garage roof.

The Bi-square antenna’s bandwidth is very broad; the antenna may be cut to the dimensions given without further ado. Purists may wish to trim the antenna to a specific frequency in the 10-meter band. Design frequencies for the antenna shown are 28.5 and 24.95 MHz. The 10-meter antenna covers the whole band with an SWR of less than 1.5:1 — quite an achievement!

adjusting the antenna to frequency

It’s easy to set the resonant frequency of the antenna “on the nose.” The bottom of the stub (F-F) is shorted by a jumper that has a one-turn loop in the center. The loop is just big enough to fit over the coil of a dip oscillator. My shorting bar is made of two interconnected copper alligator clips so I can move it up and down the stub for adjustment. The dip oscillator is monitored in a nearby receiver. Move the shorting bar up and down the stub, an inch or so at a time, until the resonant frequency falls where you want it. Finally, cut the stub to the determined length.

matching antenna to feedline

As I stated earlier, the feedpoint impedance of the antenna is about 150 ohms. The antenna is symmetrical with respect to ground, and the feedpoint is balanced to ground. Two transformations are required to match the antenna to a 50-ohm unbalanced (coaxial) line. The 50-ohm point is first transformed from unbalanced to balanced by a 1:1 balun. The 50-ohm balanced condition is then transformed to 150 ohms. The first transformation is easy; I use a “Bencher ZA-1A” air-core balun which provides an excellent balance in the 10-meter region.

The transformation from 50 ohms to 150 ohms can be done in a number of ways. One is to use a ferrite toroid transformer (fig. 3). Take a core 2.4 inches in outer diameter and 0.5 inch high (Amidon FT-240-67, or equivalent) with a permeability of 40. Sand it to remove rough edges, and then wrap it with a layer of electrical vinyl tape. Wind 18 turns of No. 14 enamel wire around the core, tapped four turns from each end. Space the winding around the entire core. Fasten the completed transformer to a phenolic mounting plate with epoxy cement, and mount the assembly in a waterproof box for protection from the weather.

a linear matching transformer

The second matching scheme uses a linear transformer, (fig. 4). The design is based on a balanced L-network. The circuit (fig. 4A) was built using a receiving-type variable capacitor for initial tests. The dimensions shown allow adjustment of the capacitor which quickly drops the SWR on the transmission line to unity at the design frequency of the antenna. The last step is to replace the variable capacitor with a fixed one and substitute a section of transmission line for the network inductors (fig. 4B). This works like a charm. A 50-µF, 5-kV ceramic capacitor (Centralab 8505-50Z, or equivalent) is substituted for the variable unit. Place it in a plastic refrigerator jar to keep moisture away. The short line section is made up in the same manner as the quarter-wave stub.

results

For a few days the dipole was left in position as a comparison with the Bi-square. In all tests, the Bi-square outperformed the dipole (usually between one and two S-units on transmit). On receive, signals that were almost in the noise were perfectly readable on the Bi-square antenna. No doubt about it, the Bi-square delivers the goods!

a 15-meter version?

The Bi-square should work well on 15 meters if you have the space. Multiply all 10-meter linear dimensions by 1.34 to get antenna size for this band.

W5LDA 160-meter antenna

Jim, W5LDA, has an interesting 160-meter antenna that incorporates a simplified feed system (fig. 5). He uses
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**185**
his 54-foot tower (with a tri band Yagi atop it) as a vertical, top-loaded radiator. Rather than fooling around with a gamma match on the tower (which can prove to be very tricky), Jim made his tower into a voltage-fed unipole antenna. He fastened a wire to the top, brought it off at an angle, and voltage fed the bottom end. The natural resonance of the top-loaded tower is such that only a simple matching network is required.

The base of the tower, as well as the shield of the coax running to the beam, are grounded at the tower base. Each lead of the rotor cable (not shown) is bypassed to ground at the tower base with a 0.01-μF, 1.6-kV disc capacitor. The leads are also bypassed to the tower at the rotor. (Jim learned the hard way that bypassing is important, after he burned out the rotor potential-meter atop the tower running 1500 watts on 160 meters!)

The coax and rotor cables are buried in a hose and run to the shack. Twenty radials, each 65 feet long, are fanned out on the surface of the ground beneath the tower.

The end of the wire is at a high voltage point and is brought into the station via a ceramic feedthrough insulator. A simple L-network matches the antenna to 50-ohm coax running to the operating position.

The antenna is very high-Q (narrow bandwidth); the network must be readjusted for a frequency change. It is possible to achieve 80-meter operation of the antenna by retuning the network.

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"ferriting" out the problem

Ferrite refers to materials that behave similarly to powdered iron compounds. They are used in radio equipment as inductors and transformers. Although they were made originally from powdered iron (and indeed the name "ferrite" still implies iron), many modern materials are made of other compounds. According to the Amidon Associates, ferrites with a permeability of 800 to 5000 have manganese-zinc composition, while cores with permeabilities of 20 to 800 are of nickel-zinc.1 The latter are useful in the 0.5- to 100-MHz frequency range of interest to most Amateurs.

toroidal cores

This month’s column will answer your questions about ferrite cores, winding toroidal cores, and using ferrite inductor and transformer cores.

A toroid is a “doughnut” shaped object, so a toroidal core is an inductor or transformer form made of a ferrite material in the shape of a doughnut. Core nomenclature provides useful information about shape, size, and type of material. For example the number FT—xx—nn means a ferrite toroid (FT) with an “xx” size, and an “nn” material type. The “F” in “FT” is sometimes deleted on parts lists, and the core defined as a “T-xx-nn.”

Amidon has a chart that provides dimensions, a description of the properties of the different types of material, and other physical data. Some of these data are also found in The 1988 ARRL Handbook for the Radio Amateur, beginning on page 2-32 (the same material appeared in earlier editions as well).

Tables 1 and 2 are derived in part from both Amidon and ARRL sources; table 1 shows the sizes and table 2 the properties of various popular toroids. These tables do not contain an exhaustive list of the variety of toroids available or all the properties of the toroids mentioned. Using the nomenclature mentioned above and the tables, you can see that a T-50-2 core (which might be called for in the parts list of a ham radio article) refers to a core that is useful from 1 to 30 MHz. It has a permeability of 10, is painted red, and has the following dimensions: OD = 0.500 inches, ID = 0.281 inches, and height (i.e. thickness) = 0.188 inches.

toroidal transformers

One reader asked me about the winding protocol for toroidal transformers seen in Amateur books and magazine articles. My correspondent included a partial circuit (fig. 1A) as an example of his dilemma. He wanted to know how to wind it and proposed a couple of methods. At first I thought the answer was obvious, then I realized that I was wrong — to many people it is not.

All windings are wound together in a “multifilar” manner. If there are three, we are talking about “trifilar” windings. Figure 1B shows the trifilar winding method. For clarity’s sake, I have shown all three wires different-ly. Because most of my projects use No. 26, 28, or 30 enameled wire to wind coils, I keep three colors of each size on hand and use a different color for each winding.
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The dots in the schematic and on the picture identify one end of the coil windings. The “dot” and “no-dot” ends are different from each other, and it usually makes a difference to circuit operation (signal phasing) which way the ends are connected into the circuit.

Figure 2 shows two accepted methods for winding a multifilar coil on a toroidal core. Figure 2A is the same method as in fig. 1B, but shows an actual toroid rather than a pictorial representation. As previously shown, the wires are laid down parallel to each other. The method in fig. 2B uses twisted wires. The three wires are chucked up in a drill and twisted together before being wound on the core. With one end of the wires secured in the drill chuck, anchor the other end to something that will hold it taut. (I use a bench vise.) Turn the drill on slow speed and let the wires twist together until you achieve the desired pitch.

Be very careful when performing this operation. If you don’t have a variable speed electric drill that runs at very slow speed, use an old-fashioned manual drill. Remember to wear eye protection if you use an electric drill. If the wire breaks, or gets loose from its mooring at the end opposite the drill, it will whip around wildly until the drill stops. That whipping wire will cause painful welts on the skin, and can certainly cause eye damage.

Of the two methods for winding toroids, the method shown in figs. 1B and 2A is preferred. When winding toroids, at least those of relatively few windings, pass the wire through the “doughnut” hole until the toroid is close to the midpoint of the wire. Loop the wire over the outside surface of the toroid, and pass it through the hole again. Repeat this process until the correct number of turns is wound onto the core. Be sure to press the wire against the toroid form and keep it taut as you wind the coils.

Enameled wire is usually used for toroid transformers and inductors, and this can lead to problems. The enamel can chip causing the copper conductor to contact the core. On larger cores, like those used for matching transformers and baluns at kilowatt power levels, the practical solution is to wrap the bare toroid core with a layer of fiber glass packing tape. Wrap the tape exactly as if it were wire, but overlap the turns slightly to ensure covering the entire surface of the core.

On some projects, particularly those in which the coils and transformers use very fine wire (like No. 30), I have found that the wire windings tend to unravel after the process is completed. To prevent this, place a tiny dab of rubber cement or silicone sealer at the ends of the windings (see fig. 2C).

Mounting toroid cores

Now that you have a properly wound toroidal inductor or transformer, it is time to mount it in the circuit. There are three easy ways to do this. If the wire is strong enough, use the wire connections to the circuit board or terminal strip to support the component. If this is not satisfactory (and in mobile equipment or wherever else vibration is a factor it won’t be), try laying the toroid flat on the board and cementing it in place with silicone
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Seal or rubber cement. For the third method, drill a hole in the wiring board and use a screw and nut to secure the toroid. Do not use metal hardware for mounting the toroid! Metallic fasteners will alter the inductance of the component and possibly render it unusable. Use nylon hardware for mounting the inductor or transformer.

How many turns to use?

Three factors must be considered when making toroid transformers or inductors: toroid size, core material, and number of turns of wire. The toroid size is selected as a function of power handling capability or convenience. The core material is selected according to the frequency range of the circuit. The only thing left to vary is the number of turns. The size and core material yield a figure called the $A_L$ factor. These values are given for several popular toroidal cores in Table 3. The required value of inductance and the $A_L$ factor are related by the following equation:

$$ N = \frac{100}{\sqrt{L/A_L}} \quad (1) $$

Where:

- $N$ is the number of turns
- $L$ is the inductance in microhenries
- $A_L$ is the core factor in microhenries per 100 turns

**EXAMPLE**

Calculate the number of turns required to make a 5-$\mu$H inductor on a T-50-6 core. From Table 3 we see that the $A_L$ factor is 40.

Solution:

$$ N = \frac{100}{\sqrt{L/A_L}} = \frac{100}{\sqrt{5/40}} = 35 $$

Don't take the equation value too seriously; a wide tolerance exists on Amateur-grade ferrite cores. While this isn't too much of a problem when building transformers, it can be critical when making inductors for a tuned circuit. If you find that the tuned circuit takes considerably more or less capacitance when called for in the standard equation, and all of the stray capacitance is properly taken into consideration, then it may be that the actual $A_L$ value of your particular core is different from the Table 3 value.

Ferrite rods are also used in receiving antennas. Although few Amateurs have them, there are places where a ferrite rod antenna (or "loopstick") is used. For example, ferrite loops are common in radio direction-finding antennas. Some Amateurs report that they use a loopstick receiving antenna when operating on crowded bands like 40 or 75 meters. The small loopstick is extremely directional and is capable of nulling out interfering signals. Of course, one would not want to use the loopstick for transmitting, so there must be some means for switching between transmit and receive functions.

Mounting ferrite rods

Ferrite rods can be mounted several ways; two of them are analogous to the methods used on toroids. You can mount the rod using either its own wires for support or a dab of cement or silicone sealer to fasten it to the board. Although you can't use simple nylon screws the way you can on toroids, you can use insulating cable clamps to secure the ends of the rod to the board.

Questions, suggestions, and criticisms are welcomed. Send them to: Joe Carr, K41PV, POB 1099, Falls Church, Virginia 22041.

References

1. Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607
Optimized forward gain comparison

In part 1, we developed a means of analyzing quad antennas based on a mutual impedance versus spacing relationship and a fixed single quad element pattern. In part 2, we will examine different quad element configurations on a variety of boom lengths and attempt to maximize forward gain through fine tuning of the element lengths. Boom length will be limited to 1 wavelength or less because this is about the longest practical boom for 15, 20, or 40 meters. The intent here is to answer the questions: What is the best possible forward gain we can squeeze out of a quad of a given boom length, and how much has the optimized gain improved from where we started? Note that we are not taking front/back discrimination or bandwidth into consideration. Both parameters are important, but both detract from maximized forward gain. Finally, we can compare the computed maximized gain for a quad against the maximized forward gain for a Yagi.

Computational methodology

Antennas are modeled in free space using the assumptions outlined in part 1*. The mutual impedance between elements is assumed to be independent of element length. The self-impedance is approximated from interpolation of measured values. Gain is calculated from integration of field pattern over a sphere, and comparison of the forward field against the average power. Field points are updated every 10 degrees (both phi and theta, spherical coordinates).

Initial antenna design is based on general ARRL Antenna Handbook principles. Reflector length is \(1030/f = 1.05\) wavelengths. Driven elements are \(1005/f = 1.021\) wavelength. Directors are all \(975/f = 0.991\) wavelengths, and all elements are spaced equally along the boom. Starting with the reflector, the element length will be increased by 0.0025 wavelength and the gain again calculated. If the forward gain improves by at least 0.005 dBi, that element is incremented again in the same direction; if not, the element will be shortened until the gain starts to fall off. Once the reflector is optimized for forward gain, the same procedure is applied to the directors in order, and the process is repeated until no significant gain increase occurs with element change. The elements, except for the ones at the end of the boom, will then be moved along the boom in 0.0025 wavelength increments to attempt a further gain increase.

Two-element quad results

The optimized gain was about 8 dBi on a 0.115 or 0.172 wavelength boom (8 and 12 feet on 20 meters). Adding a third element was worthwhile, as gain increased to about 10 dBi for boom lengths of 0.258, 0.30, 0.343, 0.430, and 0.516. Initial performance of larger antennas was better. Quads on a 0.688 wavelength boom (48 feet on 20 meters) showed an initial gain of about 11 dBi which increased to just under 12 dBi with tuning. Quads on a 0.86 wavelength boom (60 feet on 20 meters) could be tuned for over 12 dBi, and on a 1.031 wavelength boom (72 feet on 20 meters) peaked at 12.7 dBi.

The relationship between maximum gain and boom length is compared against similar data developed for Yagi antennas1 (fig. 1). The two-element quad is slightly better than a two-element Yagi (by 0.5 dBi or so) but worse than a three-element full-size Yagi (by 1.5 dBi), and a three-element quad is slightly better than a three- or four-element Yagi (again by 0.5 dBi). The quad antennas did not show the same staircase gain phenomenon observed with Yagi antennas1, and forward gain rose smoothly with increases in boom length out to 1 wavelength. With Yagi antennas the gain versus boom length relationship is not smooth, but increases rapidly over a small extension in boom length and then plateaus. Thus, Yagi and quad antennas performed similarly out to a 24-foot boom on 20 meters, but past this point quad forward gain continued to increase and Yagi forward gain plateaued (fig. 2). Between 24- and 48-foot booms the quad showed up to 2.5 dBi gain over the same size Yagi. Above 48

By David Donnelly, K2SS, 8 Alder Street, Lincoln Park, New Jersey 07035

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*Yagi vs Quad, Part 1, Ham Radio, May 1988, pg. 68.
feet the Yagi closed the gap, reducing the quad advantage to about 1.5 dB.

Several conclusions are suggested by these results:

First, quad antenna gain, like Yagi gain, is basically a function of boom length and not the number of elements, as long as "enough" elements are used. Second, the quad may have 2 to 2.5 dB greater gain than a well-tuned Yagi, but this occurs only at certain boom lengths. The quad should work about 0.5 to 1.5 dB better for the same boom length. Third, the average gain increase expected with fine tuning is about 0.45 dB (range 0.0-1.55 dBi).

Lindsay's 440-MHz experimental results initially suggested that quads were somewhat better than Yagi antennas on the same-sized boom. However, experimental uncertainties led to questions about his conclusions. For instance, it was difficult to measure gain accurately on the basis of input power because of inefficiencies in the coupling system. This may have influenced Lindsay's measurements from the start, since he measured the gain of a quad loop as being 2 dB better than a dipole. The actual value, according to several sources, should be 1 dB. Another variable in Lindsay's study involved the lengths of Yagi and quad elements; they were cut by formula and not optimized for any particular factor like forward gain. Because fine tuning of a Yagi or quad may add another 1 dB or so to the forward gain, any gain difference between Yagi and quad antennas is reduced to the experimental noise.

Not all people have found quad antennas to be as good as Yagis. Driving around California with a 70-foot portable reference antenna, Wayne Overbeck com-
YAGI ANTENNA DESIGN by Dr. James Lawson, N4PV
Based upon the popular Ham Radio Magazine series, this book includes notes, charts, graphs as well as other additional information not found in the original text. W4PV was known worldwide as one of the most knowledgeable experts on antenna design and optimization. This book is full of his contest winning “trade secrets.” Eight chapters cover analysis, performance calculations, Simple Yagi antennas, Yagi antenna performance optimization, Loop antennas, The effects of ground, Stacking, Practical design, and Practical Amateur Yagi antennas. A wealth of information at a modest price—Lawson’s book should sell for much more—every Ham should get a copy for their bookshelf.

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Table 1. Element length, spacing, forward gain and front to back (at the horizon) for two, three, and four elements on 0.115, 0.258, 0.343, 0.430 and 0.516 wavelength booms (8, 12, 18, 20, 24, 30, and 36 feet on 20 meters). Row $S$ is the starting antenna and row $O$ is after forward gain optimization. The element length is specified first and the element position is in parentheses. $R$ is the reflector; $D$ is the driven element; $D_1$-$D_n$ are directors. Gain is specified over isotropic.

<table>
<thead>
<tr>
<th>Element Length (boom position)</th>
<th>Gain(dBi)</th>
<th>F/B(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R)</td>
<td>(De)</td>
<td>(D1)</td>
</tr>
<tr>
<td>S 1.050(0)</td>
<td>7.64</td>
<td>11.12</td>
</tr>
<tr>
<td>O 1.035(0)</td>
<td>8.00</td>
<td>11.12</td>
</tr>
<tr>
<td>S 1.050(0)</td>
<td>7.65</td>
<td>11.12</td>
</tr>
<tr>
<td>O 1.037(0)</td>
<td>7.78</td>
<td>11.12</td>
</tr>
<tr>
<td>S 1.050(0)</td>
<td>8.76</td>
<td>11.12</td>
</tr>
<tr>
<td>O 1.030(0)</td>
<td>9.84</td>
<td>11.12</td>
</tr>
<tr>
<td>S 1.050(0)</td>
<td>8.99</td>
<td>11.12</td>
</tr>
<tr>
<td>O 1.035(0)</td>
<td>9.72</td>
<td>11.12</td>
</tr>
</tbody>
</table>

pared low-angle signal strengths of Yagi and quad antennas at various QTHs against a portable reference antenna at the same height as the test antenna. A review of his results showed that quads were the same as or inferior to Yagi antennas, and certainly not 2 dB better than similar Yagi antennas. Why didn’t the quads perform better?

Theoretical results presented here suggest that quad antennas with correct coupling should be at least as good as Yagis, and certainly not worse. But two complicating factors were not included in the computer model: the efficiency of the feed system and the effects of nonresonant elements in the vicinity of the antenna. Most of the antennas showing the best performance in Overbeck’s study were monoband Yagis with a double driven element. This feed system, popularized by KLM, is known for its wide bandwidth and low SWR, and is also said to give excellent results on quad antennas. Perhaps part of the answer lies in a lower feed efficiency for direct feed quads. Attaching the coax to the quad loop is the most common method of feeding quads. Most quad antennas are implemented as tribands with wires for other bands in close proximity to the desired antenna. It may be that these other wires significantly degrade performance. Both of these factors deserve a closer look.

I didn’t intend to provide optimum dimensions for quad loops, and a caveat is in order if you wish to use the dimensions provided in Table 1. Although I believe in the results of computer modeling, the dimensions should be trusted only after confirmation on an antenna test range. I have not done this, and the actual performance peaks may be different. The constructed quad loops should have the same reactance as the computer antenna. The best way to do this is to direct-
Quad antennas offer a theoretical 0.5 to 2.5 dB advantage over a Yagi of the same boom length. Like the Yagi, a gain increase of about 1 dB may be obtained through fine tuning of the elements. The increase in forward gain as one goes to longer boom lengths is smoother for quad antennas than for Yagis, and quads do not show the gain plateau seen with Yagi antennas. Finally, the quad should be significantly better than the Yagi, especially between boom lengths of 25 and 45 feet on 20 meters. However, unknown variables such as feed system efficiency or the effect of other wires in close proximity to the quad may detract from its theoretical performance.

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Macket software for Macintosh

Macket provides power and flexibility for the packet operator with a Mac®. There are windows for entering text, displaying the receive buffer, and logging transmitted text. The windows support all the features Mac users expect. The input window allows control-based editing. Other features include text uploading and downloading, printing, and macro keys.

Macket works with all Pac-Comm TNCs, the TNC-200, TNC-220, Tiny-2 TNC, and the Microport-2 TNC as well as any TNC with an RS-232 port. When used with a TNC-2 clone that has the RXBLOCK command, Macket can display the user's conversations in a special window so that the conversation will not be mixed with monitored text.

Macket's suggested retail price is $39.95. The program, developed by 'S Fine Software' is available from Pac-Comm Packet Radio Systems, Inc., 3652 W. Cypress St., Tampa, Florida 33607. Circle #302 on Reader Service Card.

dual meter wattmeters

Encomm, Inc. announces the addition of several wattmeters to their Santec line. They are actually “dual” meter wattmeters in several different models. Model W-710 covers 1.6-60 MHz and has three power levels of 2k/200/20w. Model W-720 covers 1.8-200 MHz with power levels of 200/60/15w. The W-740 has the same power levels as the W-720 but with frequency coverage of 140-525 MHz. Housed in a sturdy metal case the meters are basically unaffected by stray rf fields.

Contact Encomm Inc., 1506 Capital Ave., Plano, Texas 75074 for more information. Circle #303 on Reader Service Card.

tri-band base and mobile antennas

NCG Company now has new Tri-Band SLC system antennas for operation on 145, 446 MHz and 1.2 GHz. Both are SLC (Super Linear Converter) system antennas. They are waterproof with lightning protection. The base antenna is model CX-901; the mobile is CX-801.

Features of the CX-901 include one-piece construction of heavy-duty fiber glass. The mast diameter is 1.25 to 2.5 inches, the length is 3 feet, 4 inches, and it weighs 1 lb., 14 oz. This base antenna handles 150 watts, with frequency and gain of 144-148 MHz 3.0 dB, 440-449 MHz 6.0 dB, and 1260-1300 MHz 8.4 dB.

The CX-801 mobile unit is a one-touch, fold-over stainless steel, with an N connector for low loss, high gain. The maximum power handled is 100 watts. The antenna is 3 feet, 3 inches long and weighs 12 ounces. Frequency and gain for the CX-801 is 144-148 MHz 3.0 dB, 440-449 MHz 6.8 dB, and 1260-1300 MHz 9.6 dB.

Both antennas are designed for use with the new Tri-Plexer CFX-4310 that allows receiving and transmitting on all three bands at the same time. With one CFX-4310 it is possible to use only one coax for all three transceivers. Using the new Tri-Band transceiver you can operate three antennas from one transceiver.
THE MOST AFFORDABLE REPEATER
ALSO HAS THE MOST IMPRESSIVE PERFORMANCE FEATURES
AND GIVES THEM TO YOU AS STANDARD EQUIPMENT!

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WIRED $975
VHF OR UHF

FEATURES:
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- CLEAN, STABLE TRANSMITTER, up to 18W output standard: 50W end to combat desense.
- *CLEAN. STABLE TRANSMITTER, up to 18W output standard: 50W, end to combat desense.
- R76 ECONOMY VHF FM RCVR for 16M, 6M, 2M, 22M. Without hel reson & etc. Kits only $129.
- Weather satellite & AM Aircraft receivers also available.

NEW

GaAs FET PREAMPS
at a fraction of the cost of comparable units!

LNG -(*)
GaAs FET PREAMP
ONLY $59!
Wired/Testsed

FEATURES:
- Very Low Noise: 0.7 dB VHF, 0.8 dB UHF.
- High Gain: 13-20dB, depending on frequency.
- Wide Dynamic Range: to resist overload.
- Stable: new type dual-gate GaAs FET.


LAN -(*)
MINIATURE GaAs FET PREAMP
ONLY $24/kit.
$39 Wired/tested

GaAs FET Preamp similar to LNG, except designed for low cost & small size. Only 5/8"W x 1 5/8"H x 1/4"H. Easily mounted in many radios.


LNS-(*)
IN-LINE PREAMP
ONLY $79/kit.
$99 Wired/tested

GaAs FET Preamp with features similar to LNG series, except automatically switches out of line during transmit. Use with base or mobile transceivers up to 25W.

* Specify tuning range desired: 120-175, 200-240, or 400-500 MHz.

HELICAL RESONATOR PREAMPS
Low-noise preamps with helical resonators reduce intermod & cross-band interference in critical applications.

MODEL IRR-(*) $49 vhf, $84 uhf.


ACCESSORIES

COR-3 Kit. Control dials and audio mixers needed to make a repeater. Tail & time-out timers, local spkr amp, courtesy beep.

* CWID Kit. Fixed-programmable, timers, the works.

* TD-2 DMF DECODER/ CONTROLLER Kit. Full 15 digits, switches 5 functions, toll call restrictor, programmable, much more. Great for selective calling too!

* AP-3 AUTOPATCH Kit. Use with above for repeater autopatch. Reverse patch and phone line remote control std.

* M-202 FSK DATA MODULATOR Kit. Run up to 1200 baud digital signals through any fm transmitter with full handshakes. Radio link computers, telemetry gear, etc.

* DE-202 FSK DATA DEMODULATOR Kit for rcvr end of link.

RECEIVING CONVERTERS

Antenna Input Range Receiver...Output
136-138 MHz...28.10 MHz
143-144 MHz...28.4 MHz
148-150 MHz...28.7 MHz
150-152 MHz...29.0 MHz
152-154 MHz...29.3 MHz
154-157 MHz...29.6 MHz
157-160 MHz...29.9 MHz
160-163 MHz...30.2 MHz
163-165 MHz...30.5 MHz
165-168 MHz...30.8 MHz
168-171 MHz...31.1 MHz
171-174 MHz...31.4 MHz
171-174 MHz...31.7 MHz
174-177 MHz...32.0 MHz
177-180 MHz...32.3 MHz
180-183 MHz...32.6 MHz
183-186 MHz...32.9 MHz
186-189 MHz...33.2 MHz
189-192 MHz...33.5 MHz
192-195 MHz...33.8 MHz
195-198 MHz...34.1 MHz
198-201 MHz...34.4 MHz
201-204 MHz...34.7 MHz
204-207 MHz...35.0 MHz
207-210 MHz...35.3 MHz
210-213 MHz...35.6 MHz
213-216 MHz...35.9 MHz
216-219 MHz...36.2 MHz
219-222 MHz...36.5 MHz
222-225 MHz...36.8 MHz
225-228 MHz...37.1 MHz
228-231 MHz...37.4 MHz
231-234 MHz...37.7 MHz
234-237 MHz...38.0 MHz
237-240 MHz...38.3 MHz
240-243 MHz...38.6 MHz
243-246 MHz...38.9 MHz
246-249 MHz...39.2 MHz
249-252 MHz...39.5 MHz
252-255 MHz...39.8 MHz
255-258 MHz...40.1 MHz
258-261 MHz...40.4 MHz
261-264 MHz...40.7 MHz
264-267 MHz...41.0 MHz
267-270 MHz...41.3 MHz
270-273 MHz...41.6 MHz
273-276 MHz...41.9 MHz
276-279 MHz...42.2 MHz
279-282 MHz...42.5 MHz
282-285 MHz...42.8 MHz
285-288 MHz...43.1 MHz
288-291 MHz...43.4 MHz
291-294 MHz...43.7 MHz
294-297 MHz...44.0 MHz
297-300 MHz...44.3 MHz
300-303 MHz...44.6 MHz
303-306 MHz...44.9 MHz
306-309 MHz...45.2 MHz
309-312 MHz...45.5 MHz
312-315 MHz...45.8 MHz
315-318 MHz...46.1 MHz
318-321 MHz...46.4 MHz
321-324 MHz...46.7 MHz
324-327 MHz...47.0 MHz
327-330 MHz...47.3 MHz
330-333 MHz...47.6 MHz
333-336 MHz...47.9 MHz
336-339 MHz...48.2 MHz
339-342 MHz...48.5 MHz
342-345 MHz...48.8 MHz
345-348 MHz...49.1 MHz
348-351 MHz...49.4 MHz
351-354 MHz...49.7 MHz
354-357 MHz...50.0 MHz
357-360 MHz...50.3 MHz
360-363 MHz...50.6 MHz
363-366 MHz...50.9 MHz
366-369 MHz...51.2 MHz
369-372 MHz...51.5 MHz
372-375 MHz...51.8 MHz
375-378 MHz...52.1 MHz
378-381 MHz...52.4 MHz
381-384 MHz...52.7 MHz
384-387 MHz...53.0 MHz
387-390 MHz...53.3 MHz
390-393 MHz...53.6 MHz
393-396 MHz...53.9 MHz
396-399 MHz...54.2 MHz
399-402 MHz...54.5 MHz
402-405 MHz...54.8 MHz
405-408 MHz...55.1 MHz
408-411 MHz...55.4 MHz
411-414 MHz...55.7 MHz
414-417 MHz...56.0 MHz
417-420 MHz...56.3 MHz
420-423 MHz...56.6 MHz
423-426 MHz...56.9 MHz
426-429 MHz...57.2 MHz
429-432 MHz...57.5 MHz
432-435 MHz...57.8 MHz
435-438 MHz...58.1 MHz
438-441 MHz...58.4 MHz
441-444 MHz...58.7 MHz
444-447 MHz...59.0 MHz
VHF MODELS...
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RC-96 repeater controller

Advanced Computer Controls, Inc. offers the RC-96 Repeater Controller. Remote programming lets the operator make changes to his repeater easily without a trip to the site. The '96 features autopatch and autodialer, with storage for 200 telephone numbers. The talking S-meter lets the user check signal strength into the repeater. The controller also supports pocket pagers, linking to other repeaters, and a bulletin board. It has high-quality synthesized voice with ACC's large, custom speech vocabulary.

The '96 has built-in keypad and indicators, with storage for 200 telephone numbers. The talking S-meter lets the user check signal strength into the repeater. The controller also supports pocket pagers, linking to other repeaters, and a bulletin board. It has high-quality synthesized voice with ACC's large, custom speech vocabulary.

The risk of lightning damage is minimized by a gas discharge tube across the phone line and transient suppressors on each transient suppressors on each I/O signal. Contact Advanced Computer Controls, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66046.

Circle 306 on Reader Service Card.

new rf/high-voltage adapters

Nemal Electronics International offers a new line of rf adapters for quick and reliable interconnection between different connector series. These adapters facilitate rapid interconnection of incompatible cables and equipment while maintaining low loss and VSWR.

The NE964 adapts a type HN jack to an SMA plug; the NE962 adapts an SMA jack to a type HN plug; and the NE970 adapts a type N jack to an LC plug. The NE866 is a BNC series bulkhead feedthrough with both isolation and hermetic seal.

Constructed of brass and plated silver/nickel, these adapters have Teflon® insulation and tolerate temperatures of -55°C to +195°C. Other specifications include an impedance of 50 ohms,

The Carolina Windom comes with a special dedicated matching unit, vertical radiator section, high power line isolator, No. 14 stranded antenna wire, and glass-filled insulators. The package includes CoaxSeal® and an illustrated manual.

The price is $75, complete and ready to install. For more information, contact Jim Thompson, W4THU, Radio Works, Box 6195, Portsmouth, Virginia 23703. A catalog offering a wide selection of wire antennas, parts and accessories is available on request.

Circle 307 on Reader Service Card.

MAXFAX™ and WEFAX

Kantronics has added a weather facsimile command, WEFAX, with EPROM update 2.8. This update is available for the KAM, KPC-4, KPC-2, KPC-1, and the KPC-2400. In addition, Kantronics introduces two programs to work with the KAM or KPCs, MAXFAX 64/128 for Commodores and MAXFAX-PC for PCs and compatibles. If you use a PC, the CGA (color graphics adapter) is required.

With MAXFAX, you can store the pixel bytes from the KAM or KPC directly in RAM to the screen, or from RAM to diskette for transport or to your graphics printer. An Epson graphics format such as the EPSON LX-80 is assumed. Each MAXFAX copy comes on diskette and costs $19.95. You can order from Kantronics, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66046.

Circle 308 on Reader Service Card.

The new 5D-FB 50-ohm coax, usable above 1200 MHz, is a solid copper center conductor with a PEF insulation, foil wrapped, TFE insulation, full braid, and black PVC weather jacket. N type connectors are available.

For information on the Tri-Band antennas, Triplexers, and 5D-FB coax contact NCG Company, 1275 N. Grove Street, Anaheim, California 92806.

Circle 305 on Reader Service Card.

The Radio Works has introduced the new Carolina Windom®, a high performance, 80-10 meter antenna system. While not a windom in the classic sense, its off-center feed system suggests the name. Fed with 50-ohm coax, it produces a low SWR across nearly all of the 75/80 meter band. Operation on 40-10 meters requires a transmatch.
frequency of 0.4 GHz, VSWR of 1.3, and voltage rating of 375/1500. The adapters offer electrical performance and construction to military specifications.

For details contact Nemal Electronics International Inc., 12240 NE 14th Ave., North Miami, Florida 33161.

Circle #309 on Reader Service Card.

AR-501 radio telegraph terminal

ACE Communications, Inc.'s model AR-501 is a triple-mode radio telegraph (CW) terminal for Amateur Radio operator, and short wave listeners.

The AR-501 performs as a CW decoder, CW trainer, and electronic keyer. Features include: automatic speed follow-up and threshold control, LED tuning indicator, 32 character LCD display, random code generator, and electronic keyer for both standard and iambic. Codes can be monitored in all three modes by internal speaker and printer through the parallel printer port.

It measures 4.5 x 6.25 x 2.25 inches and is powered by a 12 VDC source. The price of the AR-501 is $229.00 including ac power adapter and parts for hook up.

For more details contact ACE Communications, Inc., 22511 Aspian St., El Toro, California 92630-6321.

Circle #310 on Reader Service Card.

high-quality variable capacitors

Kilo-Tec announces the availability of the Nevada High-Power variable capacitors. They are capable of withstanding very high rf voltage up to 7.8 kV. These heavy-duty caps are suitable for high-power antenna matching units, power amplifiers, and transmitters.

There are two values: a 500 pF and a 250 pF. Approximate prices are $29 for the TC-250 and $40 for the TC-500. To order or receive a quote contact Kilo-Tec, PO Box 1001, Oak View, California 93022.

Circle #311 on Reader Service Card.
sporadic E season—1988

This is the second summer after the end of solar cycle 21. What kind of sporadic E conditions can we expect? In the period from May through September radiation from the nearly overhead sun generates high ion densities in the lower ionosphere that support short-skip propagation, including multiple short skips. The geomagnetic field clusters these ions into cloud-like patches known as sporadic-E ($E_s$). These patches form a thin layer of intense ionization in the E region about 60 miles up. A patch gives a strong, mirror-like signal reflection over skip distances of 600 to 1200 miles. Signals remain strong for about half an hour, up to a couple of hours after the onset of the first strong signal.

The frequency and magnitude of Sporadic E occurrences is a function of geographical location. The best locations for $E_s$ openings this summer are toward the equator and on either side of the geomagnetic equator. It’s especially good where the geomagnetic equator is furthest from the geographic equator. The Northern Hemisphere areas are: Southeast Asia (best) and the Mediterranean (next best) followed by South America in the Southern Hemisphere. These were shown graphically in this column last year on a contour map.

The highest frequency propagated by $E_s$ tends to occur at local noon. The $E_s$ patch is imbedded in the regular E layer and tends to track the E maximum ion density throughout the day, season, and sunspot cycle. During this summer expect about a 17 percent increase in the E layer as an $E_s$ base for higher maximum usable frequencies (MUF) over a 1200 mile hop. This increase gives the base MUFs of 47 to 53 MHz this year, so six meter openings should be more prevalent. Two meter openings may still be rare, especially this month; perhaps August will provide some. The highest probability of occurrence is near sunrise and again around sunset. These two $E_s$ characteristics affect short-skip openings differently. Openings on the higher-frequency bands occur around local noontime; the lower bands tend to have openings near sunrise and sunset. This occurrence characteristic is nearly constant over the sunspot cycle so there should be the same number of low to midlatitude $E_s$ openings in the next few years.

last-minute forecast

Expect the higher frequency DX bands to be very good during the first two weeks of June because of solar flux peaks and longer daytime hours. Both factors contribute to elevated MUFs during the evening at midlatitude locations. No single hop transsequatorial openings are expected but look for good sporadic E openings around noon toward the end of the month. Good nighttime DX conditions on the lower bands are expected during the last two weeks of the month, but they will be noticeably shorter in duration and noisier as northern tropical thunderstorm noise propagates toward us.Geomagnetic disturbances are anticipated from solar flares around the 5th, more probable on the 13th, and from coronal thinning on the 18th through 24th of the month. MUFs should decrease about 15 percent on east-west propagation paths on most days and probably 20 percent on those paths during disturbed conditions on the 13th. Signals should be 10 to 15 dB lower level and QSB will be noticed. Paths near the equator can expect 10 percent higher MUFs.

The moon will be full on June 29th and at perigee (its closest approach) on June 5th. Summer solstice is on the 21st at 0357 UTC. The Aquarid meteor shower starts about the 8th, peaks around the 28th, and lasts until about August 7th. The maximum radio-echo rate will be 34 per hour.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around noontime via short-skip $E_s$ propagation.

There will be long-skip conditions on ten meters in the afternoon during the peak times of the 27-day solar cycle. Otherwise, look to sporadic-E short-skip and multihop openings around
The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next higher band for possible openings.*
local noon for DX on this band. (Evening transoceanic openings usually don’t occur in the summertime.)

Twelve and fifteen meters, almost always open to some southern part of the world, will be the main daytime DX bands. Operate on 12 first, then move down to 15. DX is considered 5000 to 7000 miles on these bands. There may be some long, one-hop transoceanic propagation paths occurring early in the month.

Thirty, thirty, and forty meters will support DX propagation from most areas of the world during the daytime and into the evening hours most days. Forty meters joins this daytime DX group because of lower signal absorption, and therefore lower LF (lowest usable frequency) during this last unsunspot minimum year. DX on these bands may be either long-skip to 2500 miles or short-hop ES to 1250 miles per hour. There are many good hours of DXing ahead as the days get longer.

Thirty, forty, and sixty-one are all good for nighttime DX. Although the background thunderstorm noise is becoming noticeable, these bands are still quiet enough to provide good DX working conditions. Sidelobe propagation may be a contributing factor toward enhanced conditions at local sunset and will occur more often during the next two months.

ham radio
COMING EVENTS

Activities — "Places to go . . ." 

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE NOTIFY US IMMEDIATELY OF ANY HAMMEET LOCATION, CLASSES, EXHIBITIONS, FEA TURE ARTICLES IN YOUR LOCAL PUBLICATIONS. PLEASE CONTACT HAMMEET TO ORDER ADS AND THEY WILL BE Mailed TO YOU. 

HAMMEET'S 8th ANNUAL BUSINESS MEETING AND CONFERENCE WILL BE HELD AT THE HAMMEET HEADQUARTERS, 1101 CLAY STREET, NEW YORK, NY 10014, ON SATURDAY, MAY 18, 1985, AT 9:00 AM. THE MEETING WILL CONCL UDE WITH A BUSINESS MEETING AT 11:00 AM. 

OPERATING EVENTS

"Things to do . . ."

June 4 and 5: The Weine Institute of Northern Ohio (WIWIO) will be on the air with the summer session to commemorate the 50th anniversary of the VHF radio station in Cleveland, Ohio. The event will be held on Saturday, June 4, at 7 PM. 

JUNE 11: 11th Annual Hamfest sponsored by the Arizona Radio Club of Arizona. 

JUNE 13: 12th Annual Hamfest sponsored by the Amateur Radio Club of Texas. 

JUNE 14: 13th Annual Hamfest sponsored by the Amateur Radio Club of California. 

JUNE 15: 14th Annual Hamfest sponsored by the Amateur Radio Club of Nevada. 


JUNE 17: 16th Annual Hamfest sponsored by the Amateur Radio Club of Illinois. 

JUNE 18: 17th Annual Hamfest sponsored by the Amateur Radio Club of California. 

JUNE 19: 18th Annual Hamfest sponsored by the Amateur Radio Club of Illinois. 

JUNE 20: 19th Annual Hamfest sponsored by the Amateur Radio Club of California. 


JUNE 22: 21st Annual Hamfest sponsored by the Amateur Radio Club of California. 

JUNE 23: 22nd Annual Hamfest sponsored by the Amateur Radio Club of Illinois. 

JUNE 24: 23rd Annual Hamfest sponsored by the Amateur Radio Club of California. 

JUNE 25: 24th Annual Hamfest sponsored by the Amateur Radio Club of Illinois. 

JUNE 26: 25th Annual Hamfest sponsored by the Amateur Radio Club of California. 

JUNE 27: 26th Annual Hamfest sponsored by the Amateur Radio Club of Illinois. 

JUNE 28: 27th Annual Hamfest sponsored by the Amateur Radio Club of California. 

JUNE 29: 28th Annual Hamfest sponsored by the Amateur Radio Club of Illinois. 

JUNE 30: 29th Annual Hamfest sponsored by the Amateur Radio Club of California.
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Also available QSL CARD FILE DATA BASE.
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The CS64S is the most advanced control system available at any price. Total control of your ham shack/cub club station. Each user has individual access codes and privileges granted by the control op. All user’s logged to the disk & or printer. HI/L0 priority patch & security/monitor lock modes. Talking Packet BBS input soon.

**Super Comshack CS64S $349.95

$1.00 ship USA, incl. computer interface, disk, cables & manual (simplex version inc. on request).**

**SYSTEM OPTIONS**

- External Relay Control 3 DPDT relays

- 5 pin collector outputs. CS-85 $79.95

- EPM1M Auto boot cartridge customized with your system $195.99

- Beam control; speaks bearing and rotates beam 1 degree inc. $49.95

- Full 40 Channel memory $15.00

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- Dual scan inc. & offset/var resume

**Mini (Bear Cat) Computer Control FT-727R**

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**Model 202**

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today for more information
"Q" signals

There are times when we are reminded that not everyone is familiar or experienced in the language of Amateur Radio. I was following my usual practice of listening to the activity on a local repeater while on my way to work one morning, and encountered a lively discussion on the meaning of a "Q" signal.

The signal in question was "QRU," and each Amateur knew part of the answer and thought that the other was incorrect. I resisted the temptation to break in and enlighten them about the "true meaning and proper use," but instead waited to see what developed. Sure enough, the next morning, the pair got together again; one had looked it up and was now fully informed. He surprised the other Amateur by stating that they were both on the right track, but needed the whole story. As with most Q signals, QRU can be either a question or a statement. When followed by a question mark, it (naturally) becomes a question. Without the question mark, it is a statement or an answer to a question. The discussion and follow-up not only educated the two Amateurs directly involved, but was also helpful to the many ears tuned to that repeater on those two days. Further, the incident triggered a thought that I'm putting to use here — how many other Q signals are unknown or misunderstood by a large number of Amateurs?

why Q signals?

Everyone uses Q signals. Old-timers cringe when hearing Q signals used in voice communications. Their theory is that such signals were invented for CW use, and if you are talking, you should say the phrase instead of the abbreviation.

There was a time when, deeply involved in traffic handling on several CW nets around the country, I agreed with that philosophy. However, after several years of exposure to the voice (and digital) world, I can see the merits of using Q signals wherever they apply, on any mode of communications.

Q signals, and their early companions, "Z" signals, were developed as short-cuts in message-handling procedures in marine and commercial radio circuits. It certainly was easier and quicker for an operator to send "QRU?" instead of "Do you have any messages for me?" The answer, equally shortened, would be either "ORU" (I have nothing for you), or "OTC" (I have messages for you). Before you old-time traffic handlers jump on me, yes, I've tweaked the phrase a bit. OTC really stands for "I have... telegrams for you," but Amateurs are not in the business of sending telegrams. Anyway, the short Q signal reduced the amount of key-pounding, and to a busy commercial operator, this was a blessing. Amateurs, too, realized the advantage in both time and clarity in using abbreviations and operating signals, and adapted many of them to fit their operations. The "Z" signals served the same purpose in many commercial circuits, but for some reason never caught on with the Amateur fraternity — perhaps because ARRL (American Radio Relay League) publications listed and explained the use of Q signals. Also, it has been rumored that Z signals were proprietary to some network or service, but I've not been able to find a reference that proves this.

voice and digital usage

Everyone uses Q signals on voice operation from time to time. The old
Table 1. Common Amateur Q signals

<table>
<thead>
<tr>
<th>QRM</th>
<th>Is there interference on the frequency?</th>
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<tbody>
<tr>
<td>QRN</td>
<td>Is atmospheric noise (static) bothering you?</td>
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<tr>
<td>QRP</td>
<td>Shall I reduce power? (Seldom used by Amateurs as a question.)</td>
</tr>
<tr>
<td>QRS</td>
<td>Shall I send slower?</td>
</tr>
<tr>
<td>QRT</td>
<td>Shall I stop sending?</td>
</tr>
<tr>
<td>QRU</td>
<td>Do you have anything for me?</td>
</tr>
<tr>
<td>QRV</td>
<td>Are you ready?</td>
</tr>
<tr>
<td>QRX</td>
<td>Shall I wait?</td>
</tr>
<tr>
<td>QRB</td>
<td>Who is calling? (This is not a substitute for &quot;CO&quot;.)</td>
</tr>
<tr>
<td>QSB</td>
<td>Does my signal strength vary?</td>
</tr>
<tr>
<td>QSL</td>
<td>Do you acknowledge?</td>
</tr>
<tr>
<td>QSO</td>
<td>Are you in contact with ?????????? (Amateurs seldom use QSO as a query.)</td>
</tr>
<tr>
<td>QSY</td>
<td>Shall we change frequency ??????</td>
</tr>
<tr>
<td>QTH</td>
<td>What is your location?</td>
</tr>
</tbody>
</table>

There is interference on the frequency. Atmospheric noise (static) is bothering me.
Reduce power. (Most often used as a statement, as in "I am running QRP here!" meaning the power is only a few watts.)
Send slower.
Stop sending. (Usually used to mean the station is shutting down for the moment, as in "I'm going QRT for now.")
I have nothing for you.
I am ready. Wait (most often used as in "QRX 5 minutes.")."
Your signal strength varies.
I acknowledge. I am in contact with, or I have made contact with ....... . (More often used in referring to a contact between two Amateurs, as in "Thanks for the QSO, and 73 to you.")
Let's change frequency .
My location is ____________________________

Standard "QSL?" is used to mean several things: "Do you copy?", "Did you copy?", "Do you understand?", and so forth. The answering statement, "QSL" applies to all these questions and more.

When conditions are good, and the signals are "arm-chair copy" between the two stations, there's really no justification for using a voice Q signal, but habits don't get turned on or off according to band conditions. When conditions are poor, or there is abundant interference (there it is again -- the Q signal QRM applies), certainly the letter Q sets the listener up to expect two more letters that are pertinent to the situation, and it might be easier to understand "QSY up 3" than "Let's move up 3 kilohertz".

In digital communications, the need is not so much for overcoming interference or weak-signal conditions — packet and AMTOR systems handle that pretty well — but rather a way to reduce the keystrokes at the sending station. Not all packet and RTTY enthusiasts are expert typists, and a few 3-stroke Q signals that can take the place of a whole line of text are a blessing to both sender and receiver. (How often I've stared at a blank screen wondering if something was not working right, only to find the operator at the other end was "one-finger typing" the message.)

In summary, Q signals are both useful and permissible in any mode today. It will help Novices and higher-class licensees to feel more at home on the air if they know what Q signals to use and how to use them. Table 1 lists the most common signals in both their question and answer form. This is by no means a complete list — some, like "QTE?" (What is my true bearing in relation to you?!) would probably be hard to understand and elicit a "HUH?" (which, fortunately, needs no Q signal).

I have modified the original meaning of many of these signals a bit, to make them more compatible with current Amateur Radio usage. The original Q signals were developed for commercial and aircraft use, and the language was either more stilted or directly applicable to a specific situation. As they are wont to do, Amateurs have softened the language and slanted the meaning to fit their needs, which Table 1 reflects.

Amateur traffic nets, both CW and voice, have their own set of Q signals that help to speed up message handling and network management. Many are adaptations of more common signals, with the middle letter replaced by an "N," as in QNU, which is borrowed from QRU, meaning "I have no traffic for the net." Another net signal is QNX, meaning "You are excused from the net." A few minutes spent listening to some of the busier traffic nets on 80-meter CW, 75-meter phone, and a few 2-meter repeaters is a lesson in management and a discipline that gets things done efficiently. When you read the monthly message totals as reported in QST, you can see why.

There's another signal -- QST. It does not have a question as part of its definition. QST is an alerting call to all Amateurs, indicating that some important information is to follow. It can be used by anyone, and is often heard at the beginning of network announcements and 2-meter repeater emergency-practice sessions. You're undoubtedly familiar with its use before code practice and bulletin transmissions from W1AW, the ARRL Maxim Memorial Station in Newington, Connecticut, and on the cover of their magazine, QST, which is the official journal of the American Radio Relay League.

Q signals are a vital and interesting part of Amateur language, useful in conveying information quickly and showing that you are "with it" on the bands. They fit all modes of communication (yes, even Amateur Television — a snowy picture of a card that says QRX 5 in big letters will get its message across), and when both the sender and the receiver know the meaning of "QRM, QSY down 3," things work a lot smoother!

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Orders to US and Canada add 5% of total ($2 min., $10 max)
Florida residents add 5% sales tax. COD fee $2.
Yaesu's mini HTs.
The smallest, smartest, toughest radios. Anywhere.

Whether you're a Novice or Extra class operator, you're sure to appreciate the high power, durability and size of Yaesu's FT-23R Series mini-HTs.

To begin with, you'll find a model that's right on your wavelength. The 2-meter FT-23R. The 220-MHz FT-33R. Or the 440-MHz FT-73R.

Whichever you choose, you benefit from incredibly small packaging. Take a look at the actual size photo. Aluminum-alloy cases that prove themselves reliable in a one-meter drop test onto solid concrete. And moisture-resistant seals that really help keep the rain out.

But perhaps best of all, each radio blends sophisticated, microprocessor controlled performance with surprisingly simple operation. In fact, it takes only minutes to master all these features.


The FT-23R comes with a 72-volt, 2.5-watt battery pack. The FT-73R with a 72-volt, 2-watt pack. And the FT-33R with a powerful 12-volt, 5-watt pack.

You can choose the miniature 72-volt, 2-watt pack shown in the photo below. And all battery packs are interchangeable, too.

And consider these options: Dry cell battery case for 6 AA-size cells. Dry cell battery case for 6 AA-size cells. DC car adapter/charger. Programmable CTCSS (PL tone) encoder/decoder. DTMF keypad encoder. Mobile hanger bracket. External speaker/microphone. And more.

Check out the FT-23R Series at your Yaesu dealer today. Because although we can tell you about their incredible performance, toughness and small size, seeing is really believing.

Yaesu USA 1720 Edwards Road, Cerritos, CA 90701 (213) 404-2700. Repair Service: (213) 404-4884. Parts: (213) 404-4847.

Prices and specifications subject to change without notice. PL is a registered trademark of Motorola, Inc. FT-33R shown with optional FNB-9 battery pack.
Double Vision

TM-721A
Deluxe FM dual bander

The Kenwood TM-721A re-defines the original Kenwood “Dual Bander” concept. The wide range of innovative features includes a dual channel watch function, selectable full duplex operation, 30 memory channels, extended frequency coverage, large multi-color dual digital LCD displays, programmable scanning, and more with 45 watts of output on VHF and 35 watts on UHF. TM-721A—Truly the finest full-featured FM Dual Band mobile transceiver!

- Extended receiver range (138.000-173.995 MHz) on 2 meters; 70 cm coverage is 438.000-449.995 MHz. Specifications guaranteed on Amateur bands only. Two meter transmit range is 144-148 MHz. Modifiable for MARS/CAP. Permits required.
- 30 multi-function memory channels. 14 memory channels and one call channel for each band store frequency, repeater offset, CTCSS, and reverse. Channels “A” and “b” establish upper and lower limits for programmable band scan. Channels “C” and “d” store transmit and receive frequencies independently for “odd splits.”

Optional Accessories:
- RC-10 Multi-function handset/remote controller
- PS-430 Power supply
- TSU-6 CTCSS decode unit
- SW-100B Compact SWR power meter
- SW-200B Deluxe SWR power meter
- SWT-2 70 cm antenna tuner
- SP-40 Compact mobile speaker
- SP-50B Deluxe mobile speaker
- PG-2N DC cable
- PG-3B DC line noise filter
- MC-60A, MC-80, MC-65 Base station mic.
- MA-4000 Dual band mobile antenna (not supplied)
- MB-11 Mobile bracket
- MC-43S UP/DWN hand mic.
- MC-48B 16-key DTMF hand mic.
- Dual antenna ports.
- Full duplex operation.
- Programmable memory and band scanning, with memory channel lock-out and priority watch function.
- Each function key has a unique tone for positive feedback.
- Illuminated front panel controls and keys.
- Dimmer control.
- 16 key DTMF mic. included.
- Handset/remote control option (RC-10).
- Frequency (dial) lock.
- Supplied accessories: 16-key DTMF hand mic., mounting bracket, DC cable.


Complete service manuals are available for all Kenwood transceivers, and most accessories. Specifications, features, and prices are subject to change without notice or obligation.

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