QEX

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March/April 1999

Forum for Communications Experimenters

The Heart of a 160 to 2-Meter Homebrew Station

QEX: Forum for
Communications Experimenters
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QEX

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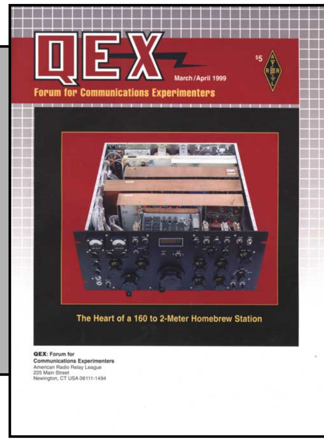

A view into the top of K5AM's homebrew transceiver.
Read about it on page 16.

Features

Columns

Mar/Apr 1999 *QEX* Advertising Index

Empirically Speaking

Many great inventions and discoveries have flowed from unique methods of reasoning. Euclid, Kepler, Newton, Edison, Einstein and Hawking, to name a few, achieved majestic results by thinking in ways few had previously imagined. Sometimes, the process begins simply by questioning preconceived or outdated notions of reality. We get ideas by examining the work of others, then we build on it. Some good independent thinking is found on the pages of this issue.

Peter Martinez, G3PLD, has been doing some thinking about digital communications again, and has written "PSK31: A New Radio-Teletype Mode" in *RadCom*. We hope to bring you some of this exciting new material soon.

Our article flow and diversity has been good, but we aren't awash in submissions yet. I know many of you are experimenting with antenna-analysis software, LF and VLF propagation, extremely weak signal work, digital audio and so on. How about contributing what you've discovered? *QEX* is the ideal forum for the presentation of both well-founded and developing concepts. It is a medium for the exchange of information about systems with results, but also for proposals, advice and learning. A lively "Letters" column confirms that.

In This *QEX*

Ex-Editor Rudy Severns, N6LF, has been working on antennas, both at his computer and in his yard. He exhibits a thorough and engaging analysis of his top-band vertical. The capabilities and pitfalls of antenna modeling software are carefully included. It's a great example of the best use of what's at hand, and it's "largely" a success!

Remote control of transceivers is a hot topic these days. Ken Beals, WK6F, realizes "real-time" control over a link at 10 GHz, where plenty of bandwidth is available. He tackles the combination of control and audio signals and addresses other not-so-obvious requirements.

Mark Mandelkern, K5AM, presents Part 1 of a complete, high-performance transceiver for the homebrewer. This ambitious project deserves close attention to its goal of top performance. Look at the August 1993 and October 1995 *QEX* issues for more information about some of the subsystems. Frank Heemstra, KT3J, gives an interesting picture of how coupled tuned circuits interact and evaluates the analogy to active filters. Some of his data were obtained from actual measurements using a chart recorder. Frank offers to provide details of the instrumentation techniques on request.

In Europe, receiver IMD is of prime concern to radio amateurs. Bombarded by very strong international broadcast signals, receivers with poor high-end performance are virtually useless. Detlef Rohde, DL7IY, understands this problem, and comes to us with a front-end design that survives the attack. Although the situation isn't unique to digital receivers, dynamic range is one of the first hurdles for DDC receiver architects.

Eric Nichols, KL7AJ, has some notes on an out-of-the-ordinary technique. This one got us thinking—and we publish it with the hope it will provoke more thought and discussion. In an article from the 1998 Southeastern VHF Society Conference, Chairman Jim Worsham, W4KXY, converts the venerable Dentron MLA2500 amplifier to 6-meter operation with a minimum of redesign. If you don't want to build from scratch and want to make more contacts when the band is "half-open," this might be the ticket.

Stu Bonney, K5PB, concludes his two-part series on wind-load standards for Yagis. I was surprised to learn that many hams with large antenna structures haven't investigated survivability. Time spent on this is a good investment! Bill Sabin, W0IYH, is designing with MOSFETs, so he contributes a 24 to 40 V, 8-A power supply. The subject gets his usual meticulous treatment—like many things, it's not as straightforward as it might seem at first.

Zack Lau, W1VT, describes a preamplifier switched by a transfer relay and looks at the losses in wet N connectors.—*73, Doug Smith, KF6DX,* **kf6dx@arrl.org**

Another Way to Look at Vertical Antennas

What's the difference between a dipole and a vertical? Maybe not as much as you think. Come along and try another point of view.

By Rudy Severns, N6LF

The grounded vertical is one of the earliest radio antennas, well known to Marconi and widely used today by amateurs, particularly for 80 and 160 meters. VHF verticals with "ground planes" are also popular. Traditionally, ground has been viewed as an integral part of the antenna—in effect supplying the "missing" part of the antenna, since, at low frequencies at least, the vertical portion of the antenna is usually less than λ/2. Even when the antenna is not grounded, but raised above ground, we still use the terms "elevated ground system," "counterpoise ground," "ground plane" and so on. In this view, we retain the concept that ground is an integral part of the antenna and an ungrounded vertical must have some structure that replaces the "real" ground. While this conceptual framework has served us well for over 100 years, it tends to limit our thinking to more traditional solutions. A change in viewpoint exposes useful variations, better suited for particular applications.

The traditional view, stemming largely from the work of Brown, Lewis and Epstein[1] in the 1930s, is that a λ/4 vertical, with a ground system of 100 or more long radials, is the ideal—anything else is an inferior compromise.

Recent work,[2,3] using primarily *NEC* modeling, has indicated that elevated ground systems with only 4 to 8 λ/4 radials can be very competitive with the more-traditional 120-buried-radial antenna, although that is the subject of some controversy, due to the difficulties experienced with experimental verification. There is even the heresy that radials as short as λ/8 may be only marginally less effective than full λ/4 radials and have significant practical advantages. Elevated-radial systems have their own drawbacks, such as (1) nonuniform radial currents,[4] which lead to asymmetrical patterns and perhaps increased loss, and (2) the need for an isolation choke at the feed point. A network of wires, arranged in a circle λ/2 in diameter and suspended above ground, may be more trouble than simply burying the wires. There has been considerable

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[1]Notes appear on page 8.

discussion—regarding traditional λ/4 radials used in elevated ground systems—as to whether these are a poor choice or not and whether other arrangements may be superior.[4, 5, 6, 7, 8]

Because most amateurs are severely limited by available space and the cost of towers and extensive ground systems, the traditional buried-radial or even the elevated λ/4-radial systems are frequently infeasible. What is needed is a wide range of other choices for the antenna structure from which to choose the best compromise for a given situation. Obviously, the final design should sacrifice as little performance as possible.

An alternate way to look at verticals has been suggested by Moxon (see Note 5) and others:

1. The antenna is a shortened (less than λ/2) vertical dipole with loading. The loading may be symmetrical or asymmetrical, lumped or distributed, inductive or capacitive, or a combination of all of these. Usually, the loading contributes little to the radiation, although some loading structures may radiate.

2. Ground is *not* part of the antenna. However, the interaction between ground and the antenna—and the loss in the ground—must certainly be taken into account. This includes both near and far fields.

This view can the maintained even when a portion (or all!) of the antenna is buried.

At first glance, this seems a trivial conceptual change. Nonetheless, looking at a vertical as a short, loaded dipole in proximity to ground—rather than as a grounded monopole—opens possibilities not usually considered with the more traditional point of view. For example, with a full λ/4 vertical, one would not normally consider adding a top hat for loading. However, in so doing, the diameter of an elevated ground system at the base of the antenna can be drastically reduced, seemingly out of proportion to the size of the top loading hat. This can be a very real advantage by reducing the footprint of the antenna. A shortened, horizontal dipole antenna with a hat at each end is very well known; it draws little comment. Nevertheless, vertically orienting the antenna and manipulating the end-loading devices to suit the application is not so common—although the antennas are conceptually identical!

Loaded Dipoles in Free Space

One of the simplest ways to resonate a shortened dipole (less than λ/2) is to add capacitive elements or "hats" at the ends, as shown in Fig 1. As indicated, the feed point may be anywhere along the radiating portion of the antenna. Fig 1 shows symmetrical end loading. Fig 2 shows extreme asymmetrical loading, where only one capacitive loading structure is used. This is, of course, the familiar ground-plane antenna being viewed as an asymmetrical dipole. Actual antennas can vary between these two extremes, since they incorporate various sizes and geometries of loading hats to suit particular applications.

When the vertical portion of the antenna, h, is less than λ/4, top loading is commonly employed. However, top loading is usually not considered when $h \geq \lambda/4$. This may be due to our past view that we need an extensive set of buried radials, or equivalently, an elevated system of λ/4 radials. For a λ/4 vertical, the diameter of the radial system will be ≈λ/2, changing only slowly as the number of radials is varied. On the other hand, if we lengthen the vertical section beyond λ/4, add some top loading or even some inductive loading, the diameter of the bottom radial structure drops rapidly.

A simple example illustrating this point is given in Figs 3 and 4. Fig 3 shows an asymmetrical λ/4 dipole with two radials (L1 and L2) at each end. L2 is varied from zero to 22.3 feet, and L1 is readjusted, as needed, to resonate the antenna at 3.790 MHz.

Clearly, adding even a small amount of top loading (L2) greatly reduces the length of the bottom radials (L1), and consequently the land area required

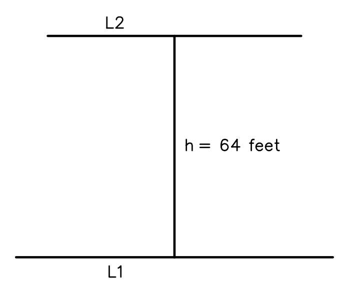

Fig 3—Asymmetric two-radial dipole. F_R = 3.790 MHz.

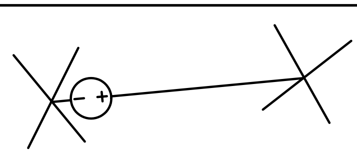

Fig 1—Short loaded dipole.

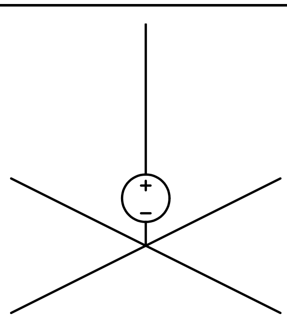

Fig 2—Asymmetrical dipole.

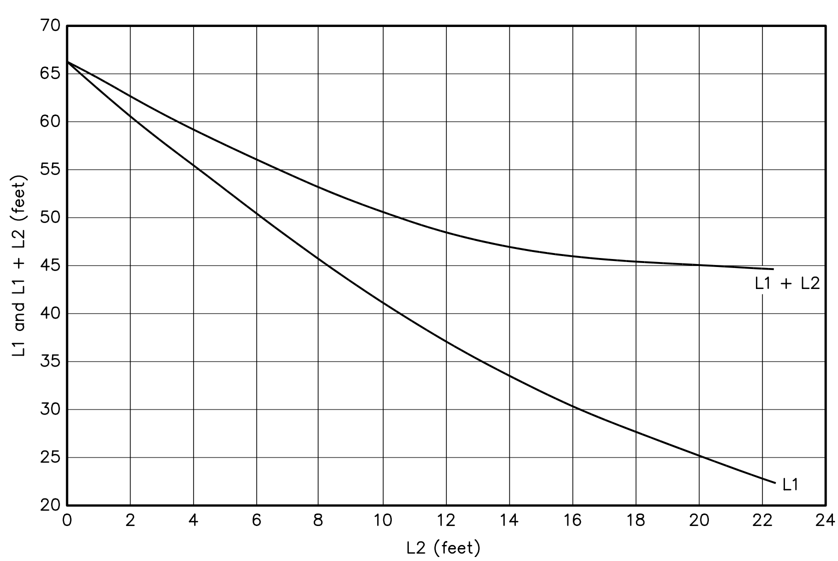

Fig 4—Effect of top loading on radial length.

for installation. This is a matter of considerable practical importance to those with restricted space in which to erect an antenna. With somewhat more complex loading elements, the footprint can be reduced even further.

In addition to greatly reducing the length of the radials, a number of other things happen during the above exercise:

1. With only two radials and no top loading, the radiation pattern varies with azimuth by about 0.7 dB, making the pattern slightly oval. This pattern asymmetry essentially disappears as the radials (L1) are shortened with top loading (L2).

2. When placed over ground, the currents in individual λ/4 radials are rarely equal. This can lead to asymmetric patterns and increased loss. The current asymmetry rapidly decreases as the radials are shortened.

3. The peak gain, and the angle at which it occurs, changes relatively little as top loading is added and bottom radials shortened while keeping the vertical section the same length.

4. Small amounts of inductive loading could also be used to supplement or even replace the top loading. As long as the vertical section is close to λ/4, the radial lengths can be reduced to λ/8 without seriously increasing losses.

Modeling Issues

The realization that everything—from the length of the radiator to the type and distribution of loading—is a potential variable that may be adjusted to achieve specific ends, is a very liberating idea. Unfortunately, it brings its own set of problems. Which variations are best for a given application? A multitude of questions arise when judging any particular variation.

The large number of possibilities and questions cannot be dealt with analytically, at least beyond an elementary level. The only practical way to deal with the many variables is to systematically explore the possibilities with *NEC*, *MININEC* or other CAD modeling software. Yet, even that is not a simple matter. Each modeling program has particular strengths and weaknesses that affect its use for this problem. The bottom portion of a vertical for 80 or 160 meters is usually very close to ground (less than 0.05 λ). For these applications, the modeling software should implement the Norton-Sommerfeld ground and properly model the current distribution in the lower part of the antenna as modified by induced ground currents. Only *NEC 2* and *4* do this. Of course, if the lower part of the antenna is buried in the ground, only *NEC 4* is suitable.

Loading structures may consist of a web of wires with multiple wires at each junction, perhaps of different diameters, and with small angles (less than 90°) between adjacent wires attached to the same node. *MININEC*-based software can model multiple acute angles if segment tapering is used, but if many wires are used in the structure, the number of segments becomes quite large. *MININEC* Broadcast Professional, using a different segment-current distribution, does an even better job without the need for tapering. However, both of these programs do not model the interaction properly for very low antennas over real ground. *NEC 2* can model the ground effects correctly, but may not handle the multiple small angles properly, especially if different diameter conductors are connected. *NEC 4* is much better in this respect, but is not widely used by amateurs because of its expense.

Real grounds are frequently stratified beginning only a few feet down. On 160 meters, the skin depth is of the order of 15 to 20 feet, and it is common to have several layers with different electrical properties over that distance. Even in homogeneous ground, the effect of rain and subsequent drying creates a varying conductivity profile. None of the presently available software addresses this problem. The validity of *NEC 2* or *4* modeling for ground has been questioned because of differences between experimental measurements and predictions made by modeling. This is a critical issue. If NEC is fundamentally deficient with regard to ground modeling, then the comparisons to date between buried-radial and elevated-radial systems are invalid. That includes the work reported in this article! On the other hand, NEC modeling may be fine, but the problem lies with the highly variable nature of real ground. This is particularly so down to depths of 15 to 20 feet, which cannot be simulated with NEC, but that could greatly modify experimental results. Some support for this view comes from experimental work at higher frequencies. There the skin depth is much less, and modeling predictions are in much better agreement with experiments.

The presently available software, while a remarkable achievement, is not totally satisfactory to fully exploit the possibilities. The suggested point of view brings this out. A great deal of care must be used when modeling a vertical with a complex loading system near ground.

A Design Example

The advantages of employing a different conceptual approach can be illustrated using the 160-meter vertical used at N6LF, where an effective antenna was built on a very difficult site at low cost.

The site is on a narrow ridge—approximately 60 feet wide at the top—in a forest. There is no possibility of installing an extensive buried radial system because of the dense forest, heavy underbrush, steep slopes and very large old-growth stumps. Even an elevated system of normal size, about 260 feet in diameter, is not practical.

A support for the antenna was constructed from three Douglas fir trees, fastened together to form an **A** frame (see the sidebar "A Large **A**-Frame Mast, Inexpensively" for details). This resulted in a support 135 feet high. Allowing eight feet from the bottom of the antenna to ground and a few feet of slack at the top for sway in high winds, the final vertical length is 120 feet—very close to λ/4. Because the antenna is located over 700 feet from the shack, 75 Ω Hardline coax (a freebee from the local CATV company) is used for the transmission line. The antenna was designed to have a 75-Ω feed-point impedance to match the transmission line. The feed-point impedance at the junction of the lower hat and the vertical wire was manipulated by adjusting the relative sizes of the bottom-hat and top-loading wires. Alternately, I could have used a larger hat on the bottom and moved the feed point up into the vertical part of the antenna, but this was not done because of the limited space available for the bottom hat. I also tried some inductive loading at the base and at the junction of the top-loading wires. Relatively small amounts of inductive loading—with very little additional loss—would further reduce the size of either or both of the capacitive loading elements. I did not keep any inductive loading because sufficient space was available for the arrangement shown.

The final antenna is shown in Fig 5. There are four radials at the bottom, connected by a skirt wire at the ends. The diameter of this bottom-loading structure is only 40 feet, compared with 260 feet for normal λ/4 radials. Two sloping wires are used for loading at the top. A sloped top hat may not be optimal when compared to horizontal wires: The radiation resistance is somewhat lower. Nevertheless, this arrangement is very simple and allows the antenna to be tuned by changing the angle of the wires with the vertical portion of the antenna. This can be done from ground level by shifting the attachment points for the guy lines supporting the sloping wires.

Christman's comparison (see Note 2) between a 120-buried-radial vertical and an elevated four-radial vertical (both with $h = \lambda/4$) indicates that the gain and radiation-pattern differences between the antennas are quite small: 0.35 dB for peak gain, 1° for peak gain angle. Because the difference is so small, I have chosen to use the four-radial elevated antenna as the reference antenna, since it is much easier to model than a complete 120-buried-radial antenna.

Using *NEC4D* for modeling, radiation-patterns for a four-radial ground-plane antenna and this antenna were compared. The result is presented in Fig 6. The model assumes ground of average electrical characteristics under the antenna ($\sigma = 0.005$ S/m; $\varepsilon = 13$). The wire used was #13 copper, and its loss was included in the modeling. The price paid for drastically reducing the diameter of the bottom loading structure is a peak-gain reduction of 0.5 dB. This is a fair trade for dramatically easing the installation of the lower loading element because 0.5 dB will probably not be detectable in actual operation. In the real world, where full-size ($\lambda/4$) radials very likely have varying currents (see Notes 4 and 8), the smaller antenna may not, in fact, be inferior at all. In this particular example, full-size radials would need to zigzag down a steep hillside at various angles. It is very doubtful they would have been any better than the small hat that was adopted.

Any antenna with an elevated radial system needs an isolation choke (common-mode choke, or *balun*, if you prefer) on the transmission line near the feed point. One effect of moving the loading from the bottom to the top of the antenna is to increase the potential between the feed point and ground. This requires more inductance in the isolation choke to properly decouple the transmission line. For this application, I happened to have a roll of $^1/_2$-inch Hardline. The roll was about two feet in diameter, so I simply expanded it into a coil three feet long and two feet in diameter with a simple wood framework to hold it in place. Fig 7 is a photo of this king-sized decoupling choke.

The result was a choke with 350 µH of inductance (4 kΩ at 1.840 MHz). When this value of inductance was placed in the model with a buried transmission line, there was still some interaction; resonance was displaced downward. This was also found true on the actual antenna. This illustrates one of the drawbacks of very small bottom-loading structures: A choke with enough inductance to avoid interaction may not be practical, at least on 160 meters. Since the current in the choke is relatively small, additional losses due to ground currents will not be very large. The Q of the choke, however, must be high to limit losses in the choke itself.

The monster balun shown here is extreme and not required. A much smaller choke could be used. The large structure was used because it was actually very convenient with the materials on hand.

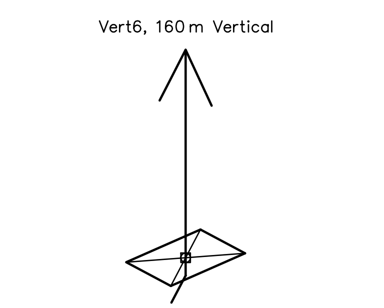

Fig 5—Antenna configuration.

Fig 7—Rudy and the "small" decoupling choke.

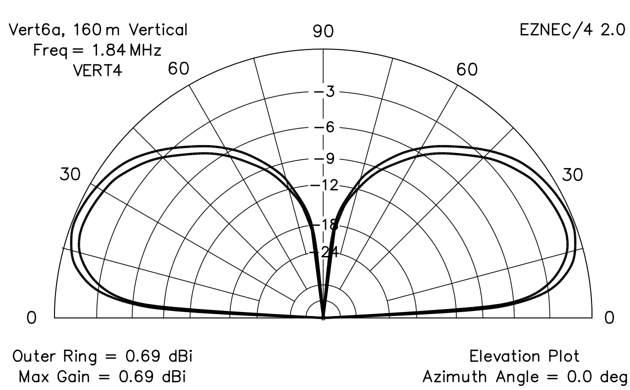

Fig 6—Comparative radiation pattern.

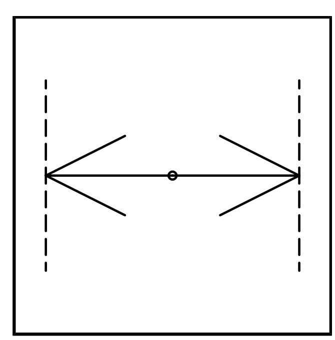

Fig 8—Flat versus drooping loading wires.

A Large A-Frame Mast, Inexpensively

A λ/4 vertical is about 70 feet tall on 80 meters, and 130 feet on 160 meters. Getting this height with a tower can be expensive. I needed a less-expensive alternative. In the Pacific Northwest, fir trees with heights greater than 100 feet are common, and can usually be purchased locally and inexpensively if they are not already growing on your property. In the southeastern US, there are extensive pine forests which, while not typically as tall as the firs, can be used in the same way. I have many tall Douglas Fir trees on my property, so I selected three of them, two with 12-inch diameter bases and one of about 8 inches. I trimmed the top off the two larger trees at a point where they were about five inches thick. This gave me two poles approximately 80 feet long. Since I was only going to support a wire vertical, I topped the smaller tree at a point where it was roughly two inches thick. This gave me a pole 60 feet long. I was trying to have the cross-sectional area at the top of each large pole roughly equal to the area at the base of the smaller pole when they were overlapped.

The next step was to drag the poles to the antenna site and assemble the A-frame shown in Fig 9:

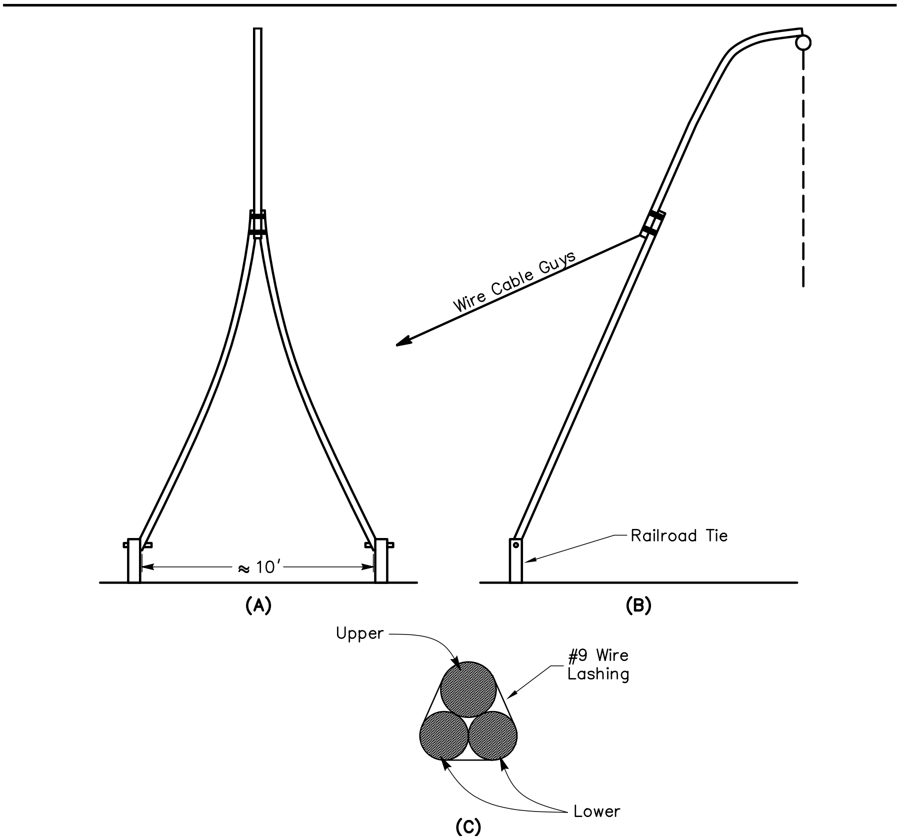

Fig 9

1. I bought a large, used railroad tie and cut it in half at the middle of its length. I then buried each half vertically with about 18 inches above the ground to form a pivot post. I placed the posts about 10 feet apart.

2. I placed the two large poles, side-by-side, midway between the two posts.

3. I placed the smaller pole on top of the two large poles—overlapping by about five feet—and lashed the three poles together using #9 galvanized smooth iron fence wire as indicated in Fig 9C. To begin the lashing, I stapled the end of the wire; as I applied each turn, I tightened it with a claw hammer. After 15 turns or so, I stapled the free end.

4. I then spread the butt ends of the large poles out to the pivot posts. [Did you use a team of mules, or just your burly "pecs"?—*Ed.*] This spreading tightened the lashings very nicely (!) so that the three poles were solidly connected.

5. I wanted to raise and lower the A frame at will and keep the pole ends away from soil contact (rot!). Therefore, I created a pivot at each post by drilling a 2-inch-diameter hole through the post and pole butt. I then inserted a length of 1.5-inch galvanized iron water pipe as the shaft for the pivot. To keep the pipe from slipping out, I put a pipe cap on each end as a retainer.

6. The next step was to attach *two* halyards (one spare, just in case!) to the top of the mast. I used two small pulley blocks—the kind typically used on sailboats—and then rove a length of black, sun-resistant, 3/8-inch Dacron line through each block. The lines were long enough to form a continuous loop reaching the ground, so I could hoist or recover the antenna at will.

7. Finally I erected the **A** frame. In my case, I used a nearby tree as a gin pole (suitably guyed!!) along with three steel blocks and a long length of wire rope. Hoisting power was supplied by a small tractor. I took great care because of the forces involved. The initial lift required a pull of over 1000 pounds and the **A** frame weighs over a ton. (Green trees are heavy!) If I were more patient, I could have allowed the trees to dry out (months!), which would have greatly reduced the weight.

I choose not to raise the mast to a vertical position because I wanted the antenna and the loading structures to stand clear of the mast and any guys. As shown in Fig 9B, I left frame tilted about 15° from vertical and bent the top over like a fishing pole, so it is even farther out from the base. The green pole bent relatively easily, and the bend became permanent when the wood dried out.

I used two wire-cable back-guys, anchored at the junction of the poles, to hold the mast in place. Although the weight of the mast makes it unlikely it would blow over towards the guys, I use the spare halyard as a guy from the top of the mast in the opposite direction to the wire guys. This arrangement minimizes conductors in the near field of the antenna.

The cost of the entire exercise was less than $75, and I expect to get many years of use from the mast. Of course, I had the trees, the tractor and the hoisting tackle, which kept the cost very low.

Table 1—Antenna Comparison at 3.510 MHz

Antenna	*h (ft)*	*L1 = L2 (ft)*	Z_{middle} Ω	Z_{end} Ω	*Peak gain (dBi)*	*Peak angle °*	*Wire loss (dB)*	*2:1 BW (kHz)*
λ/2	137	0	91	>5000	+0.30	16	0.08	270
Lazy-H	120	4.4	96	1096	+0.28	17	0.07	280
Lazy-H	100	10.4	94	384	+0.12	19	0.07	280
Lazy-H	80	17.4	81.3	180	-0.06	20	0.08	260
Lazy-H	69.8	21.6	71.2	127	-0.07	21	0.09	240
Lazy-H	60	26.3	59.7	90.9	-0.15	22	0.10	200
Lazy-H	40	38.3	33.7	40.8	-0.38	24	0.16	140
Lazy-H	30	45.6	21.5	23.8	-0.59	25	0.23	100
λ/4 (2 radials)	69.8	—	—	38.8	+0.11 by-0.39	22	0.15	200
λ/4 (4 radials)	69.8	—	—	35.7	+0.21	22	0.13	175

More Modeling

In the process of developing this antenna, a great deal of additional modeling was performed to explore the effect on performance of different loading arrangements. One of the more interesting variations was a symmetrically loaded, two-radial antenna called a Lazy-H vertical (see Note 6). This antenna is intended to be supported between two trees. The antenna is identical to that shown in Fig 3, except that L1 = L2. Table 1 gives a comparison between a full λ/2 vertical, a λ/4 ground-plane with two and four radials and the Lazy-H with different values of h (height of the vertical portion) varying from 120 down to 30 feet. Note that the λ/4 Lazy-H is within 0.3 dB of the four-radial λ/4 vertical and has greater bandwidth. If two supports are available, the Lazy-H is much easier to fabricate than the four-radial version, and has significant size in only two dimensions instead of three. I assumed #13 copper wire and average ground for the models. Z_{end} is the impedance at the junction of the vertical section's lower end and the lower radials. The bottom of all the antennas is assumed 10 feet above ground.

In the 160-meter example given earlier, the top loading structure was simply a pair of drooping wires led to anchor points near ground. The question arises as to the comparison between flat configurations, like that shown for the Lazy-H and the drooping-wire alternative. This question can be quickly answered by modeling an end-loaded dipole in free space with two different configurations as shown in Fig 8. The modeling shows that the drooping wires must be lengthened to achieve resonance, the radiation resistance is significantly lower with drooping wires and the far-field pattern is essentially the same. From a practical point of view, the use of drooping wires greatly simplifies the structure, and has very little effect on the far-field pattern. It may reduce the efficiency of the antenna if the radiation resistance is lowered too much, however. This is the kind of trade-off information critical to a new design.

In general, modeling this class of antennas shows that peak gain and peak-gain angle primarily determined by ground characteristics and the height of the vertical radiator, h. The loading means has only a second-order effect on the radiation pattern. A variety of loading arrangements can satisfy a particular situation with little loss of performance—as long as we keep the radiation resistance high enough to control losses.

Conclusions

This article has advocated a different conceptual view of vertical antennas: They can be viewed as loaded dipoles close to ground. Changing the point of view makes it easier to recognize the wide range of options available for configuring a high-performance vertical to meet the needs of a particular site and set of limitations. To assess the many options, we need the help of software. Unfortunately, no available software package provides the desired computational capabilities. Users of any antenna modeling software should be very careful when setting up the model and interpreting results.

Acknowledgement

In addition to the referenced papers, other workers in this field have pointed out the advantages of the point of view presented here. This idea is certainly not the author's creation, although I wholeheartedly endorse it. Moxon's work deserves careful reading. I am indebted to Dr. L. B. Cebik, W4RNL; Dick Weber, K5IU, and Grant Bingeman, KM5KG, for their comments and support.

Notes

[1]Brown, Lewis and Epstein, "Ground Systems as a Factor in Antenna Efficiency," *IRE Proceedings*, June 1937, Vol 25, No. 6, pp 753-787.

[2]A. Christman, "Elevated Vertical Antennas for the Low Bands: Varying the Height and Number of Radials," *ARRL Antenna Compendium*, Vol 5, 1996, pp 11-18.

[3]A. Christman, "Elevated Vertical Antenna Systems," *QST*, Aug 1988, pp 35-42.

[4]R. Weber, "Optimal Elevated Radial Vertical Antennas," *Communications Quarterly*, Spring 1997, pp 9-27.

[5]L. Moxon, "Ground Planes, Radial Systems and Asymmetric Dipoles," *ARRL Antenna Compendium*, Vol 3, 1992, pp 19-27.

[6]R. Severns, "The Lazy-H Vertical Antenna," *Communications Quarterly*, Spring 1997, pp 31-40.

[7]J. Belrose, "Elevated Radial Wire Systems For Vertically-Polarized Ground-Plane Type Antennas," Part 1, *Communications Quarterly*, Winter 1998, pp 29-40; Part 2, Spring 1998, pp 45-61.

[8]D. Weber, "Technical Conversations," *Communications Quarterly*, Spring 1998, pp 5-7 and 98-100.

A 10 GHz Remote-Control System for HF Transceivers

Have you ever wondered what that computer-control port on the back of your new transceiver is for? This system provides the building blocks for putting it to good use.

By Ken Beals, WK6F

One of the most interesting changes in Amateur Radio equipment in recent years has been the inclusion of computer control interfaces. These control ports allow varying degrees of transceiver control via a personal computer (PC). While one might question what good such an interface does when the radio and the computer are sitting in front of the operator, this type of control interface opens a whole new range of remote-control possibilities. With the advent of "front-panel-less" transceivers like the Kachina 505DSP, remote control becomes easier still. Several articles in past years have shown some practical systems for controlling these computer-controlled HF transceivers over a remote link. One system used a PC to control a station over a telephone line,[1] while others have used hand-held UHF transceivers to provide wireless remote control.

One thing that seemed to be lacking in the systems I saw was the true feeling of "real-time" control. Pushing buttons on an H-T just isn't the same as using the main tuning knob on an HF transceiver. Telephone control had some limitations due to the relatively low data rates possible, although this is becoming less of an issue now with higher speed modems. I wanted a system that would allow transceiver remote control using a high-speed data channel and the necessary audio channel. The system should also be flexible enough to work with any radio/software combination. If this could be done over a wireless link, that's even better. Since my interest was in controlling a remote station—hopefully on a hilltop—from home, a point-to-point microwave auxiliary link with large bandwidth seemed like a good solution. There seemed to be nothing readily available.

About the time I began looking for such a system in the summer of 1994, I was finishing my BSEE program at California State University, Chico. One of the program requirements was to construct a project that illustrated what I had learned in the preceding years. I thought about what would be

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[1]Notes appear on page 15.

required to implement a remote control system, and remote control seemed like an interesting project. This article describes the basic system that resulted from that work. The system remotely controls any RS-232-equipped transceiver via a 10-GHz microwave link, while simultaneously providing the uplink and downlink audio channels.

Let me say right up front that I don't expect anyone to run out and duplicate these units exactly. This article is an overview of the system rather than a nuts-and-bolts description. Although complete schematics and board layouts are available, my goal is more to elicit comments and suggestions from others interested in remotely controlling amateur stations. With the new RF radiation limits, as well as ongoing efforts of communities to limit outdoor antennas, the option of moving the HF part of the station to a remote site is becoming more attractive for many amateurs.

System Overview

This system provides the two basic things necessary to control a transceiver remotely: a data channel for control and an audio channel for both receive and transmit audio. The data and audio channels are independent, and both channels are active at all times. The audio channel is full duplex, with both the uplink and downlink active at the same time; the data channel is a half-duplex connection capable of 115,600 bit/s, with only one direction active at a time. Having the control and audio channels active simultaneously allows you to control the transceiver while receiving or transmitting, just as you would with the radio sitting in front of you. Fig 1 shows the basic idea. What we have is a wireless equivalent of RS-232, microphone and speaker cables, all bundled together.

One other requirement that I placed on the system is simplicity of operation, which is apparent in the finished units (see Fig 2). The system needed to be completely self contained and free from the need to twiddle with any controls once it was up and running. This was especially important in the case of the unit to be placed in the remote location. Most good hilltops are a bit difficult to access, and the intended 30 to 40-mile range meant making it as operator-independent as possible.

Although both units are shown together here, one unit would obviously be placed in the remote location and one in the operating position. The units are nearly identical, with slight differences—such as the speaker volume control on the operating-position unit—made of necessity. The front panel of each unit includes two LEDs that indicate power and the equivalent of the "data carrier detect" on a modem. In addition, a 10-segment LED bar graph display gives a quick indication of signal strength between the units.

The rear-panel connectors are also slightly different, to allow for the different input and output signals (Fig 3). In addition to the connectors for the microphone and speaker, the serial port and the externally mounted 10-GHz signal sources and antennas, there is a standard modular-telephone handset connector. This allows connection of a telephone handset so the system may also be used as a full-duplex voice-communication system. The handset can also be used as a substitute for the microphone and speaker when operating the transceiver. All link adjustments are internal, and once set, should not require further adjustment during normal operation.

The 10-GHz signals are generated by Gunnplexer transceivers mounted directly on surplus 48-cm-diameter dishes and connected to the station or control point via two RG-58 cables. One cable carries the transmit signal to the Gunnplexer, while the other carries the received signal as well as the dc to power the Gunnplexer. The received signal on the cable is in the 60-MHz range, so for reasonably short cable runs, the use of extremely low-loss cable is unnecessary. For those readers who are unfamiliar with these cute little microwave devices, an abbreviated explanation of Gunnplexer operation will be presented later.

Block Diagrams

Figs 4 and 5 show the block diagram of the two units. Although their basic functions are the same, the input and output signals for each unit required slightly different circuits. The data channel is implemented using frequency shift keying (FSK) on the 10-GHz carrier. The audio channel is provided by first digitizing the analog audio signal from either the microphone or receiver and using FSK to modulate a 4 to 5-MHz carrier. That signal is then used to frequency modulate the 10-GHz carrier.

The signals are recovered at the other end by first using the Gunnplexer to convert the microwave signal to a composite signal in the 56 to 61-MHz range. Standard single-chip FSK demodulator devices recover the digital data stream for each channel. The data channel signals are converted to RS-232 voltage levels and sent to either the transceiver or the computer. The digital audio signals are converted back to analog, then either amplified to speaker level—in the case of receiver audio—or shifted to

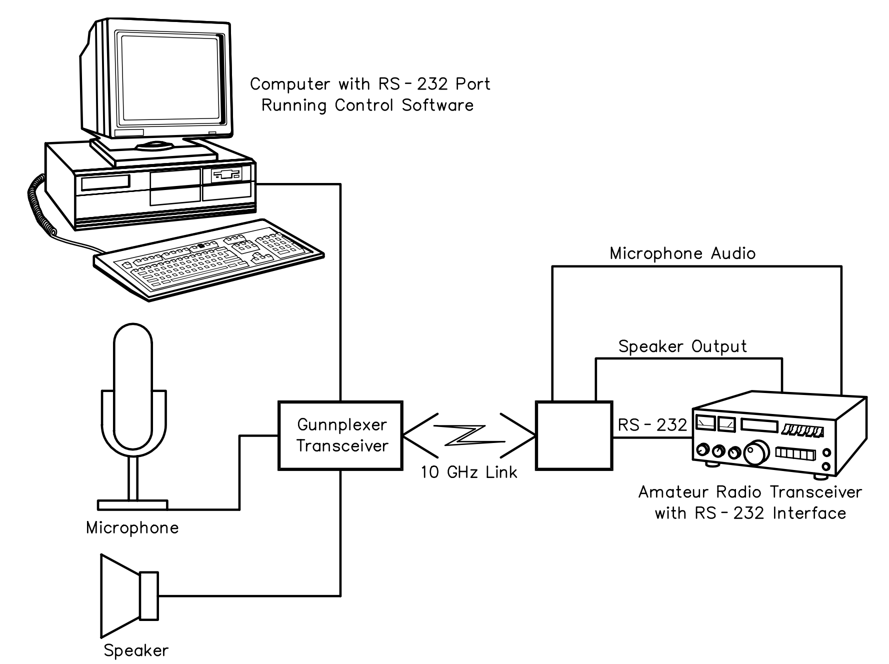

Fig 1—System functional diagram.

Fig 2—Front view of the remote (left) and control (right) 10-GHz transceivers.

Fig 3—Rear view of the remote (left) and control (right) 10-GHz transceivers.

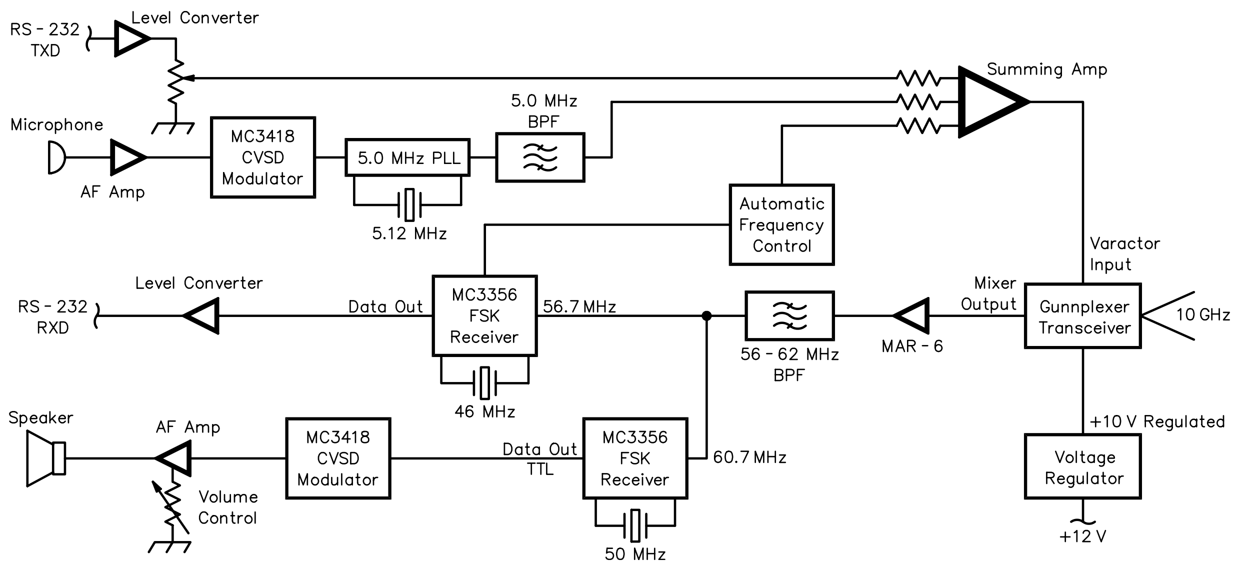

Fig 4—The control transceiver block diagram.

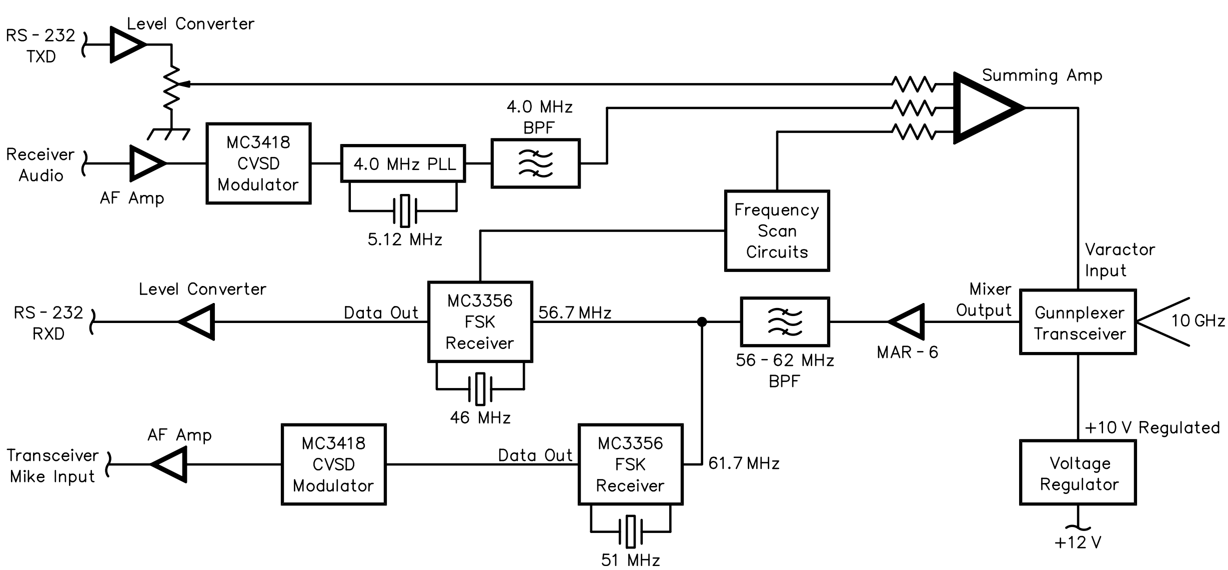

Fig 5—Remote transceiver block diagram.

the level and impedance required by the transmit audio input on the rear of the transceiver.

Detailed Signal Flow

The signal flow from one end of the link to the other is essentially the same whether we're talking about the uplink or the downlink. The signals can be followed using Figs 4 and 5.

When the controlling PC sends a command, the RS-232 signal enters the control-link transmitter through the rear connector and is immediately converted to TTL levels by the MC1489 data-transceiver chip. This signal is then attenuated by an adjustable voltage divider, which allows adjustment of the frequency deviation of the FSK signal. This signal is buffered and combined with the audio signal, as well as a frequency-control voltage, to be described later. This composite signal feeds a relatively high power op amp, an LM6313, originally designed to drive transmission lines. The op-amp output is applied to the 10-GHz Gunnplexer frequency-control input.

The audio signal enters the system either through the microphone connector on the rear panel of the control-site unit, or via the line-audio input jack at the remote site. The signal is buffered and its level is adjusted by a generic LM324 op amp. A two-pole active low-pass filter cuts off frequencies above about 4 kHz. This filter prevents *aliasing* by limiting the input frequencies to a value less than half the sampling frequency.

Analog to Digital the Easy Way

I mentioned before that the analog audio signals are digitized before they're used to modulate the carriers. There are a number of ways to accomplish this, but many of the systems—especially those used today by the telephone system—do not lend themselves to simple asynchronous transmission schemes like the one used here. In many commercial telephone systems, a device samples the audio amplitude, converts it to a number between 0 and 255 and presents the sample as an 8-bit digital word. The word must then be converted to a stream of serial bits, sent out, and then recovered in a way that assures positive determination of the beginning and end of each word in the serial bit stream.

A simpler approach is known as *delta modulation*. This scheme doesn't transmit the absolute level of each sample, but merely a signal proportional to the difference between the current and previous samples. If the sample is greater than the last sample, we transmit a "1;" if the sample is less than the previous sample, we transmit a "0." This process works well as long as the signal doesn't change too much between samples. Fig 6 shows a sine wave along with the digital output of the encoder, as well as the reconstructed audio waveform after decoding.[2] The analog signal is reconstructed by integrating the encoded bit stream. Now it doesn't matter when you start receiving the bits, or even if you miss one once in a while, since the incremental change in the signal level is small from one sample to the next.

The basic system can be improved by allowing the integrator to also change the slope of the ramp adaptively, so it can follow rapidly changing signals as well as slowly changing ones. This technique is known as *continuously variable-slope delta modulation* (CVSD). Motorola has implemented this system, with both the encoder and decoder, into the MC3418. All you need is a clock and a filter to limit the input signal frequency, and you have digital audio on a chip. Since this device was developed for limited-bandwidth telephone applications, it's a good fit with the typical narrow-bandwidth audio produced by a transceiver.

The ADC sample clock runs at 62.5 kHz, so the audio varies little between sample times. It is possible to use a much slower sample clock, and thus reduce the bandwidth, but I found that the recovered audio quality could be improved somewhat by using this higher sampling rate. The overall audio channel quality could be further improved by increasing the bandwidth even more, but since both receive and transmit audio are processed by the transceiver's IF, the quality would probably not change very much at the listener's ear.

Modulation

To separate the audio-channel data stream from the control signals on the auxiliary 10-GHz link, a relatively low-frequency subcarrier is used in a form of frequency-division multiplexing. The system uses a phase-locked loop (PLL) that generates a 4 or 5 MHz subcarrier—depending on whether it is the uplink or downlink—which is frequency-shift keyed by the digital signal from the ADC. Fig 7 shows a simplified block diagram of the circuit. The modulated subcarrier is then mixed with the data channel signal and sent to the Gunnplexer frequency-control input.

Generate 10 GHz without Tears

The microwave signal is generated with a pair of MA/COM Gunnplexer transceivers. These relatively low-cost devices take the headache out of microwave-signal generation and recovery. There have been many construction articles that detail the operation of these units, so I won't go into much detail here. The units provide a 10-GHz signal source via a Gunn diode oscillator, a frequency proportional to an applied control voltage and an IF-output signal. The IF signal is generated by a device called a *ferrite circulator*, which functions essentially as a mixer. The IF is the difference between the frequency of the incoming signal and that of a Gunn diode oscillator. The Gunnplexers are rugged devices that require only a single supply voltage of +10.0 V

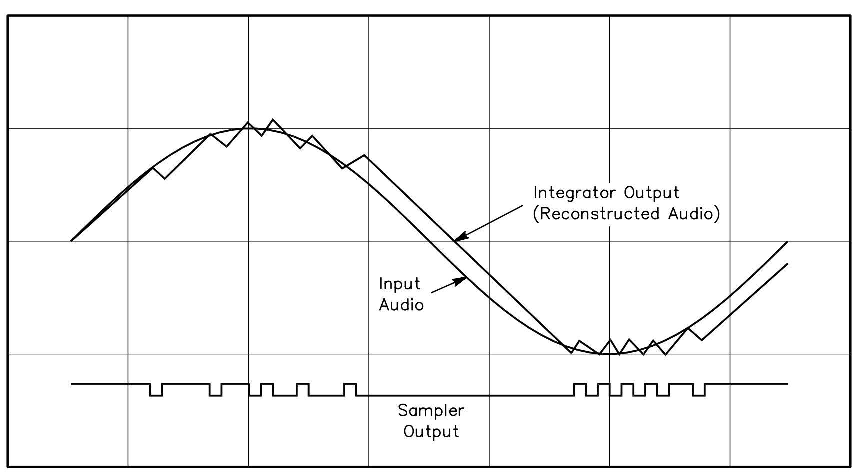

Fig 6—CVSD coding/decoding diagram.

dc. Their design allows the Gunnplexer to be mounted directly to a dish antenna at the proper feed point, eliminating the need for a feed line that would be necessary if the signal source was inside the shack.

The genius of the Gunnplexer is the double use of the Gunn oscillator in a single unit. It not only provides the transmitted signal, but also serves as the first LO in the receiver chain. By tuning the oscillators to slightly different frequencies, the received IF can be set to any convenient value. Many projects use regular FM broadcast radios as the IF strip, since separating the Gunnplexers by 100 MHz or so puts the first IF in the FM broadcast band.

Fig 8 shows the receive-signal path. In this case, the Gunnplexers are separated by 56.7 MHz. The received signal produced by the Gunnplexer is routed to a Motorola receiver chip (MC3362) with a LO frequency of 46.0 MHz, which results in a 10.7 MHz IF signal. This simplifies the design, since this is a very common IF frequency. It allows the use of commonly available ceramic filters to set the passband width of the receiver—280 kHz in this case. The audio channels are offset by an additional 4.0 or 5.0 MHz (uplink or downlink), so an identical receiver circuit with a 50.0 or 51.0 MHz LO will recover the digital audio signal. The Motorola MC3356 device is an FSK receiver on a chip, with oscillator, mixer, discriminator and data-recovery circuits all included. It also provides a squelch circuit and discriminator outputs, which are critical in dealing with frequency-stability issues present when using the Gunnplexer.

As shown in Fig 9, each Gunnplexer is mounted in an aluminum box, along with the PC board. The board holds a receiving preamp (Mini Circuits MAR-6), a toroidal matching transformer for the IF output and the +10 V supply for the Gunn diode. The feed horn is mounted directly to the "business" end of the Gunn oscillator cavity. The entire assembly is mounted to a support arm that extends from the center of the dish. The dishes are a couple of surplus wireless-cable dishes that I found for a good price. The set up is very similar to that shown in N6GN's high-speed data link project in the *ARRL Handbook*.[3]

The Problem of Frequency Stability

One problem with the Gunnplexer is its relatively poor frequency stability over temperature. The manufacturer gives a figure of 350 kHz / °C, which, if we think in terms of the HF bands, is like going from the CW to the phone band every time the temperature changes 1°C! Even with the wide bandwidth used here, it wouldn't take much of a temperature change to move the received signal clear out of the passband. Obviously, we need a circuit that will automatically compensate for changes in temperature.

Fortunately, we have the tools we need to address the frequency stability problem. We can adjust the frequency by changing the dc voltage on the frequency-control input to the Gunnplexer. We can track the frequency shift by looking at the discriminator output of the receiver chip, which is a linear function of the received signal frequency. All we need do is feed back the discriminator's dc output voltage to the Gunnplexer frequency control input, and we can force the IF to stay the same, no matter how much—within reason—the microwave oscillators change frequency. This method is shown in MA/COM's applications notes that come with the Gunnplexer. I also used some additional ideas shown by Messrs. Bellantoni and Powell in their article about the 24-GHz voice system they built, which allows for automatic signal acquisition and tracking.[4]

The frequency-control loop works like this: When the units are first powered on, one unit scans a 35-MHz range looking for a signal. When the receiver finds a signal, the squelch opens and turns on the front-panel data-carrier-detect LED, which in turn, after a slight delay, stops the scan. The other unit now looks at the filtered discriminator voltage, and changes the dc level going to the Gunnplexer varactor by just enough to keep the signal in the center of the passband. If the drift is so much that the receiver cannot compensate and loses the signal, the discriminator volt-

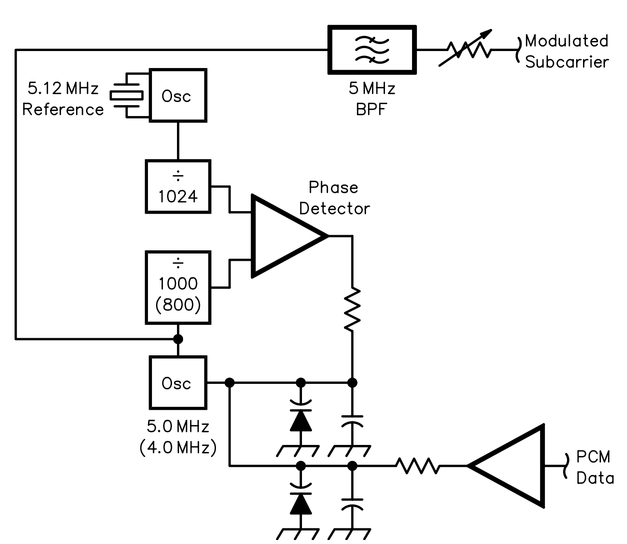

Fig 7—Audio-channel carrier generation/modulation.

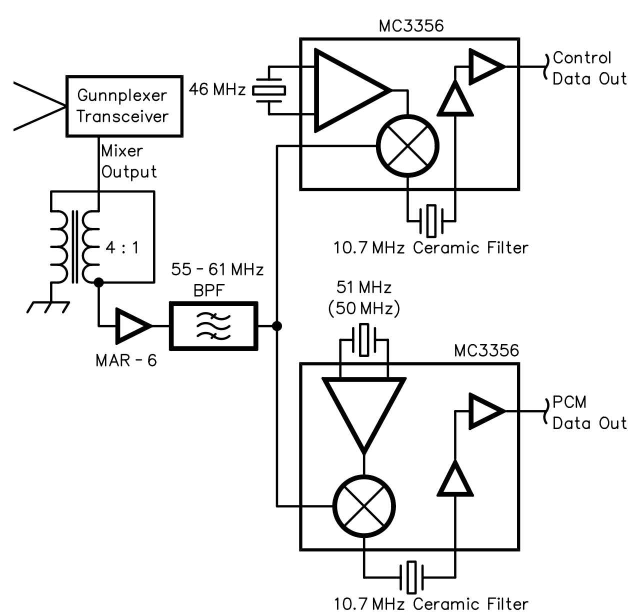

Fig 8—Received-signal path.

age will return to the center of its range, consequently shifts the transmit frequency to the center of its range. The other unit will lose the signal, go back to scanning until it finds the signal again. This process takes less than one second, and since the audio channel is squelched when no signal is present, it is only just noticeable that it happened at all. Although a command on the data channel would probably be lost, the control software could be set up to confirm all commands as received and executed, and to initiate retries for lost commands.

One other disadvantage of the Gunnplexer is that it "hears" its own transmitted signal as well as the desired signal from the other end of the link. Changing the Gunn oscillator frequency in transmit also changes the first LO in the receive path. It is necessary to sample the transmitted data signal and offset the change in discriminator voltage caused by the receiver hearing the transmitter. This ensures that the frequency-control system isn't trying to follow the FSK. This is easily done by using the data channel's digital signal to cancel any change in the discriminator voltage that occurs as a result of transmit-frequency changes before applying it to the Gunnplexer frequency-control line. Now that we have a complete frequency-control signal, it is combined with the data and audio channel signals and sent to the frequency-control input of the Gunnplexer.

Putting It All in a Box (or Two)

As with any project, once all the circuits are designed and functioning, packaging is the remaining challenge. Since this was going to be a one-of-a-kind project, I considered just wiring everything up on one big piece of perfboard, perhaps even using the tried-and-true art of dead-bug construction. Nevertheless, since this was going to be the culmination of my 3½ years of full-time school attendance, I decided on something a little more professional. Fig 10 shows an internal view of the control-site transceiver. All circuitry was assembled on custom-etched circuit boards, and the boards enclosed in metal enclosures I found at a local surplus house. Each major subassembly, with the exception of the power supply, was built on a small board that plugs into a large "mother-board" mounted on the lower half of the enclosure. As shown in Fig 11, this allowed each board to be raised out of the unit on an extender card for easy servicing. As we all know, it may work on the breadboard, but when the final circuits are assembled, nothing works exactly the way you intended. All of the signals on the back plane are either dc or relatively low-frequency signals, so extender boards do not affect the operation of the units. The high-frequency signals are handled by the two small coaxial lines to the rear panel, one for th received signal, and the other for the

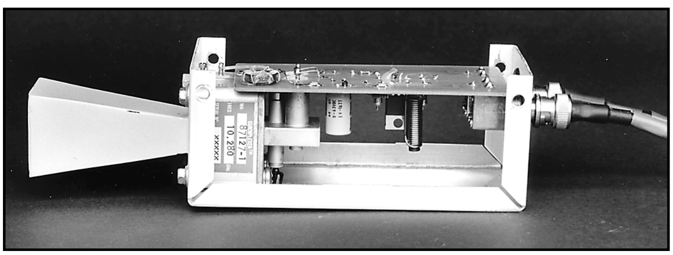

Fig 9—The antenna feed assembly includes a Gunnplexer and feed horn, along with a board that holds a receiving preamp, IF matching transformer and the +10 V supply.

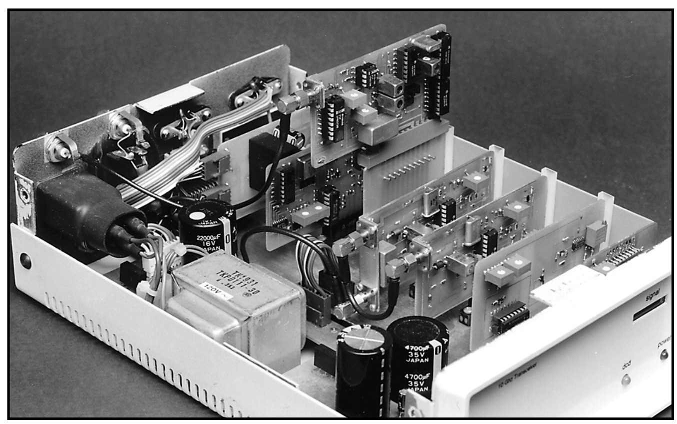

Fig 11—Extender boards permit easy servicing when required.

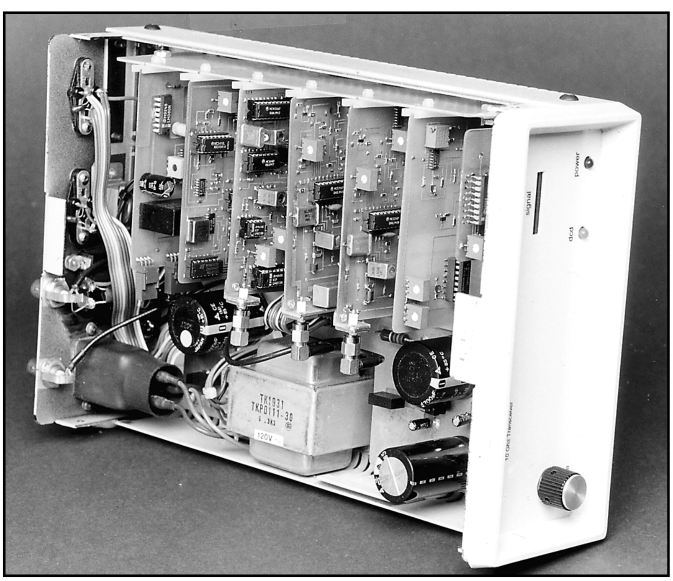

Fig 10—An inside view of the control transceiver with the top removed.

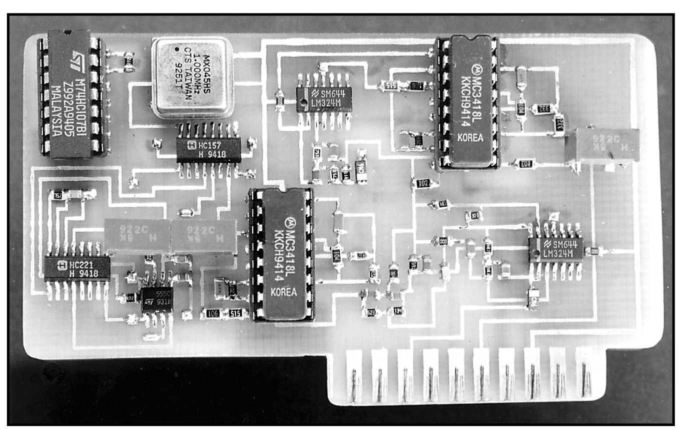

Fig 12—A close-up of the ADC board. Note the mix of surface-mount and traditional components used. Some of the boards have components on both sides.

Gunnplexer frequency-control signal.

Since the size of the enclosures was already determined, I had a limited amount of space with which to work. I decided early on that the use of surface-mount components was going to be necessary to fit everything. Unfortunately, not all the devices were available in the SOIC small-outline packages. In addition, components such as filters and coils were also easiest to find in traditional packages. Some of these components had to be adapted to surface-mount techniques, which usually meant bending the leads 90° to form "feet" that could be soldered to pads on the board. The majority of the adjustable inductors had to be hand wound to a particular value, and I had the luxury of a HP 4195A Network/Spectrum Analyzer to measure the values of these inductors as they were constructed.

Fig 12 is a close-up of the ADC board. It reflects the mix of surface-mount and traditional components used. For a few of the boards, it was also necessary to mount components to both sides, and the use of surface-mount components made the layouts much simpler. The one job I didn't feel like tackling was that of plating the via holes that connect the two sides of the board together. I opted for the inelegant but simple solution of soldering short wires to both sides of the board at each via. This method was also used to tie the ground planes together.

The boards were laid out using the PaintBrush application that is present in Windows. Although there are many good PCB layout packages available, I found PaintBrush easy to use and flexible in laying out the double-sided boards. The layouts were printed using a laser printer onto special iron-on sheets. Once the iron-on sheets are peeled off, the printer toner remains on the board, forming the resist. This method has been described in the pages of *QST*, and I found it quick and easy to produce good quality boards. I initially had trouble aligning the two sides of the board, and opted to etch each side in a separate process. It took a little longer, but with the small size of some of the pads I found it easier to get the two sides lined up accurately enough for the via holes.

Does It Work?

Unfortunately, I didn't have the facilities to do any testing over long distances. The best I could do was set up the dishes on temporary stands a few hundred yards apart. Everything worked as I expected, and the received signal levels indicated that the receiver limiters were at maximum. Path-loss calculations predict that the system should easily work over a 40-mile line-of-sight path, with sufficient signal-to-noise (SNR) ratios to provide low bit-error rates.

I used a Yaesu FT-747 and the *RigWindows* software from MFJ to test the basic operation of my system and found this combination quite easy to use. This particular software lets you use horizontal mouse movements to change the VFO frequencies. (A trackball is a good approximation to a tuning dial.) I also hooked up a digital encoder to the mouse port. When mounted on the front panel of a box with a large knob, it provided the real-time feel that I was after. With eyes closed, it feels like sitting right in front of a radio.

Future Enhancements

Since any HF station is generally more than just a transceiver, a number of improvements and additions are needed. By using another PC and suitable software at the remote site, control codes could be filtered appropriately and sent to any number of suitably equipped devices connected to the PC. These could include amplifiers, rotator controls, antenna switches, etc.

Of course, the most important addition at this point is the addition of a watchdog timer that would detect failure in the control link and kill the transmitter if necessary [as required by 47 CFR 97.213(b)—*Ed.*] Since there is no error detection or correction on the data channel, there is always the possibility that a garbled command could put the transceiver in an unknown state. Protection could easily be implemented in software on a PC at the remote site.

This is one of those projects that will probably never be "finished." With modular construction, it is relatively easy to make changes to particular circuits without disturbing circuits that are working well. Of course, software can always be upgraded to enhance the operability of the system without *any* hardware changes.

A Word of Thanks

As mentioned earlier, this is a "Senior Project" leading to my BSEE degree. I would like to thank everyone in the Electrical and Computer Engineering Department at California State University, Chico—where I attended school 20 years later than I should have—for all their understanding and support. A note of special thanks to Professors Dr Harold Petersen, Dr Ben-Dau Tseng, Dr Philip Hoff and Dr Richard Bednar who took the extra time necessary to help me work through the myriad of details in such a project. They helped to make my late return to school an enjoyable and rewarding experience.

Now if I can just find a good hilltop . . .

Notes

[1]L. Amodeo, and J. Schultz, "Computer Remote Control of an Amateur Radio Station," *QST*, Nov 1991, pp 25-30.

[2]*Motorola Telecommunications Device Data*, 1992.

[3]R. Schetgen, KU7G, Ed., "A 2-Mbit/s Microwave Data Link," *The 1994 ARRL Handbook for Radio Amateurs*, (Newington: ARRL, 1993) pp 32-4 to 32-11.

[4]J. Bellantoni and S. Powell, "A Communications System using Gunnplexer Transceivers," *Communications Quarterly*, Fall 1990, pp 9-15.

Ken Beals, WK6F, was first licensed in 1967 as WN6VFJ, upgrading shortly thereafter to WB6VFJ. He was most active in the late 1960s and early 1970s in the HF contesting arena from various multioperator stations in Southern California, as well as working for a local ham-radio business (W6HX) installing radio towers and antennas. After spending 18 years at Xerox Corporation as a technician, he decided in 1991 to pursue a BSEE degree (full time) and graduated from California State University, Chico in 1994. He is presently employed by Hewlett-Packard Company as a Manufacturing Engineer supporting HP's RF Network Analyzer production line. Ken looks forward to again becoming active in Amateur Radio; he also enjoys flying as a private pilot.

A High-Performance Homebrew Transceiver: Part 1

Here is a general description of a transceiver built without regard to size or complexity—the only goal is optimum performance for DX and contests.

By Mark Mandelkern, K5AM

Ham radio affords an opportunity for a variety of fascinating experiences. For many hams, foremost among these is operating with equipment built in a home workshop. Traditionally, hams have built equipment that outperforms factory gear. Today, modern methods allow the production of miniaturized, microprocessor-dependent units that are difficult for radio amateurs to duplicate. However, in regard to basic performance and operating features, the home workshop can often produce gear that outclasses factory equipment.

This article begins a description of a homebrew transceiver built for serious DX work and contest operating. It was designed without compromise regarding performance and basic operating features, but includes no modern computer-related features. The goal was to build a radio that would outperform any available factory-built radio, regardless of price.

In 1948, I began homebrewing my first station, W9ECV, in Wisconsin. By 1990, everything in the shack was homebrew except the transceiver. Work began on this high-performance homebrew transceiver—it was completed three years later. Two portions of the radio, perhaps the most innovative, have already been described in *QEX*: The high-performance AGC system and the non-crunching noise blanker.[1, 2] A description of the completely homebrew K5AM station recently appeared in *NCJ*.[3]

Basic attitudes toward homebrewing acquired in the 1940s have been retained. Homebrew means "built-from-scratch," and often newly designed. I've tried to keep up with some of the latest devices and techniques, but this transceiver uses no microprocessor, no synthesizer and no phase-locked loops. No phase noise, no spurs, and no birdies! This, of course, involves limitations: No memories, no instant frequency jumps, no computer control, no DSP.

For ordinary DX work, nonetheless, these features are not missed. For contesting, however, here is the disclaimer: World-class contesters would not find this radio acceptable. They often want to operate two radios simultaneously, with computer control. They

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[1]Notes appear on page 24.

want frequencies read from the radio and entered into the computer to create band maps, and for logging purposes. In other words, they want bells and whistles *ad infinitum*. So I must hedge on the contesting claim in the subtitle of this article; this radio is designed for the sort of contesting that I do. This includes single-operator DX contests, domestic HF contests and VHF contests on several bands with transverters. Results have been gratifying. A number of section and area awards have been won with this homebrew radio. I have operated the best of the current factory radios and—in many respects—find them lacking in comparison. The bottom line is that this homebrew transceiver has certain performance characteristics that surpass those of the best factory gear. This more than compensates for the lack of bells and whistles, at least for my style of operating. As for backing up the basic performance claim, I'll give data in a following article about alignment and measurements.

This radio was built for daily use in a station with a heavy operating schedule. It was not built as an experimental platform. Experimenting with each circuit, trying at each stage to obtain results surpassing all previously published circuits, would have meant that the project would run to decades, rather than years, and would likely never have been completed. Besides, this builder has no professional training and lacks the expertise for such a project; I merely put together circuits, already optimized by experts, to produce a complete operating unit. Individual circuits were chosen from the available literature at the time of construction.

This article describes only the general plan of the transceiver, with emphasis on design considerations and a discussion of features required for serious DX work and contesting. Subsequent articles will give circuit details.

Conversions: Fewer are Better

The craze for multiple conversions began in the 1950s. At that time, there were some good reasons. A single-conversion receiver with a 455 kHz IF had intolerable images on the 10-meter band. Thus, a first conversion to about 3 MHz was a great improvement. In addition, it was difficult to obtain very sharp selectivity at 455 kHz, so a third conversion to about 50 kHz was helpful. Eventually, we saw advertisements for quadruple-conversion radios.

The advantages of multiple conversions came at the cost of increased IMD. Mixers have noise figures equal to their conversion loss, and this loss must be compensated by gain stages ahead of the mixer. The result is very high signal levels at the last mixer before the high-selectivity filter. It is this last mixer that mainly determines the close-in dynamic range of the receiver. The current hype about strong front-ends is misleading. The crunch is at the last mixer [for nearby signals—*Ed*]. Even more misleading is the hype about strong preamps for VHF and UHF. A preamp cannot improve the IMD performance of a receiver, it can only degrade it; the more gain in the preamp, the more trouble down the line.

The situation is different today than in the 1950s. Excellent crystal filters for both SSB and CW are available at about 9 MHz. With a first up-conversion and a second conversion down to 9 MHz, superb image rejection and selectivity are easily obtained.

This transceiver tunes the bands in 1-MHz segments, with a variable first IF tuning 40 to 39 MHz. The tuning is "reversed" because the injection is on the high side; eg, at 68 MHz for the 28 to 29 MHz band. The result is a very high-performance radio with only two conversions. The high-side injection method results in virtually no spurious responses; it is used on all bands. There is no difference between the 160 and 2-meter bands, as far as mixing scheme is concerned. The arithmetic is easy: Simply add 40. Injection is at 41.8 MHz for the 160-meter band and at 184 MHz for 2 meters. All the front-end oscillators use third or fifth-overtone crystals, well known for their low phase noise. For 2 meters, the 92-MHz oscillator is followed by a balanced doubler. (In this article, the author uses the term "front-end" to indicate both the receive front-end and the transmitting circuitry along with a control panel for a particular frequency range: HF, 6 meters or 2 meters.—*Ed.*)

I didn't invent this conversion scheme. It is taken from the Signal One CX7, a radio that appeared in 1969, and that had dozens of ground-breaking innovations. Few hams today know that the CX7 is the granddaddy of all present-day radios. I operated, repaired and modified my CX7 for 20 years before designing my homebrew transceiver. Did I learn anything from the CX7? Well, when it first arrived in my shack, all I could say was: What are those funny little things with three legs?

Basis for Design

Many ideas were taken from the CX7 besides the frequency-mixing scheme. Notably the RF speech clipping, the transmitter driver circuits and the PA bias and ALC circuit. In addition, some surplus parts were used from basket-case CX7 radios found at flea markets. The crystal filters, precision-machined mechanical parts (bearings, etc) for the PTOs, the panel escutcheon for the frequency counter, the conduction-cooled 8072 PA tube, the anode clamp and heat sink, and a few other miscellaneous small parts were salvaged from various dismantled assemblies.

The goal in building my own radio was to obtain improvements in performance, operating convenience and new features. The features I had used for 20 years and liked, I retained. Even the panel layout bears some resemblance to the CX7. This led one friend to ask: "Is your radio a CX7 clone?" Definitely not! Is every radio with a 455 kHz IF a clone of the first one? Although the frequency-mixing scheme in this radio is the same as in the CX7, virtually every circuit is newly designed. At the same time, my debt to the CX7 designers is enormous. It was essential to keep the same IFs for two reasons: First, I planned to use surplus 9-MHz CX7 crystal filters, which are excellent, if selected from a batch. Second, transverters that fed directly into the CX7 40 MHz first IF had already been built.

Here are some features of the K5AM homebrew transceiver:

- Balanced JFET and balanced MOSFET mixers
- Careful gain distribution
- High dynamic range
- Non-crunching noise blanker (see Note 2)
- High-performance, no-pop, no-click hang AGC circuit (see Note 1)
- Super-sensitive integrating squelch for SSB and CW (mainly for 6-meter DX)[1,4]
- Complete TTL logic control. Eg, the mode switch has seven leads, not dozens.
- Quick, one-button PTO switching
- Relay-switched crystal filters
- A sharp CW filter at the IF output (in addition to one at the input)
- Electronic attenuators for all audio level controls. This reduces hum problems.
- 60-Hz filter in receiver audio
- High-pass filter in transmitter audio to eliminate externally induced hum.
- Pulse tuning circuit for safe, easy external amplifier tuning; set for 33% duty cycle[5]
- Sharp CW shaping. No dit delay after key closure.

- RIT—separate knobs for each PTO
- One-button second-PTO monitoring
- Automatic (optional) transverter switching tied to PTO switching. This is mainly for instant 6 to 2-meter switching during VHF contests. It is also useful for instant switching between 144.200 MHz and 3.818 MHz (75-meter liaison) during meteor showers.
- Clear, sharp audio—the audio output module uses a class-A, push-pull circuit

Tuning the First IF

This radio differs from nearly all current designs—the first IF circuits are tunable. One advantage applies to VHF DX operators. The basic 40 to 39 MHz tunable transceiver forms an excellent foundation for the attachment of VHF transverters. For the 6-meter band, the LO injection is at 90, 91, 92 or 93 MHz, from four separate crystal oscillators. This high-side injection yields no detectable spurious responses. The overall conversion total is merely two. This is in sharp contrast with common practice, where 50 MHz is converted to 28 MHz; LO injection at 22 MHz allows the possibility of unwanted spurs. In such an arrangement, the overall conversion total can be as high as five, with the resulting high possibility of spurious responses, spurious emissions and IMD.

Tuning the first IF means that the circuits following the first mixer have a bandwidth of 1 MHz and thus the second mixer must be strong enough to handle any signals within this range. This is in contrast with current production radios that use a fixed-frequency first IF and crystal filter at about 70 MHz, to limit the spectrum of signals within the first IF strip. These filters are usually about 20 kHz wide, with poor shape factors and ultimate-rejection characteristics, compared to filters at 9 MHz. During lab tests using two-tone spacing of 20 kHz, such receivers may demonstrate excellent dynamic range. Nevertheless—with today's crowded band conditions—one cannot hope to find a clear 20-kHz-wide segment in which to operate. Published reviews do not address this problem.

This radio, on the other hand, while not employing a first IF filter, does have good dynamic range performance with respect to adjacent-channel signals, as well as more distant signals. This performance relates more closely to real operating conditions.

Another problem arises in some current production receivers. These develop AGC in the first IF, ahead of the sharp crystal filters, and apply AGC voltage to the first IF strip and front-end. This may prevent IMD in the first IF strip, but gain is reduced and weak signals may be lost. Receiver sensitivity is reduced in the presence of nearby strong signals that lie outside the second-IF passband. Hence, front-end AGC and gain reduction may be unacceptable for some weak-signal work.

In this radio, AGC is applied only to the IF strip at 9 MHz, after the crystal filters. This arrangement allows the IF gain to be controlled with no loss of sensitivity in the front-end. The mixers are built to handle strong signals, and the sharp 9-MHz crystal filter following the second mixer effectively keeps off-channel signals out of the AGC circuits. The front-end runs wide open at full sensitivity, the best arrangement for weak signals. On the lower HF bands, it is sometimes pru-

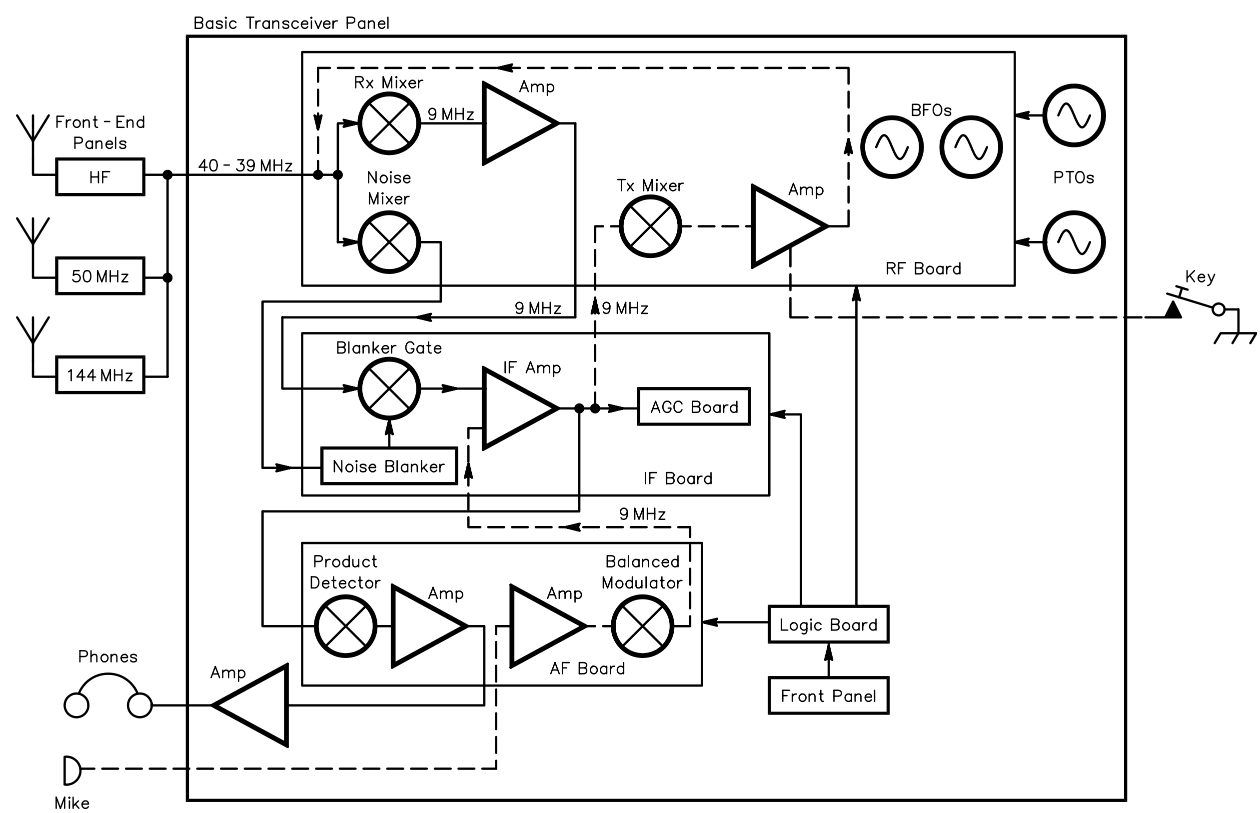

Fig 1—K5AM homebrew transceiver simplified block diagram. This diagram includes the basic transceiver panel, tuning 40 to 39 MHz, and the three front-end sections, for HF, 6 meters and 2 meters. Additional front-end sections, for other VHF or UHF bands, can be easily added. The dashed lines indicate transmit signal paths.

dent, although not always necessary, to use the front-end attenuator.

Block Diagram

Fig 1 shows the main sections of the transceiver. Its most unusual feature is that it is built on separate panels; one for the basic 40 to 39 MHz transceiver, and one for the 200 W HF front-end. These two items were completed, after three years work, in 1992 and together form an HF transceiver in the usual sense.

However, the transceiver covers all ham bands from 160 meters to 2 meters, so we need a brief description of how the transceiver fits into the complete station. The 6 and 2-meter front-end sections were built a few years earlier. These are often called transverters, but in this case they have 40 to 39 MHz outputs; they do not translate the VHF bands to *any* amateur band. They are integral parts of the transceiver, functioning exactly like the HF front-end. Now there are 11 ham bands from 160 through 2 meters. It takes four panels, and we have only 2 W on 6 and 2 meters, not a very acceptable power level for an 11-band transceiver. So, we now add 100 to 200 W 6 and 2-meter amplifiers. One 6-meter amplifier was built in 1951, when I was still in high school. So, I finally have a complete 160 to 2-meter homebrew transceiver at about 200 W, on only six panels, which took only about 40 years to build! There are also homebrew 1.5-kW amplifiers covering all bands.[5,6,7] A block diagram of the entire station is shown in the *NCJ* article (see Note 3).

Compactness and miniaturization have obviously not been prime goals for this project. On the other hand, this style provides a good deal of flexibility. For example, many homebrewers concentrate on VHF/UHF SSB/CW DX operating. Building the basic transceiver separately allows one to add whatever VHF/UHF front-end sections are desired. The 40-MHz IF, with high-side injection, works very well on the 6 and 2-meter bands, and should work well on higher frequency bands.

Receiver Gain Distribution

Receiver gain distribution is a crucial factor in obtaining high dynamic range. In this radio, the IF strip—after the sharp filters—was designed with the highest practical gain. This allows the signals at the mixer to remain at a low level. Gain before the sharp filters increases signal levels at the last mixer, and so reduces dynamic range. Gain after the filters has no detrimental effect on receiver performance, just as turning up the audio gain in a large room does no harm. This is one reason why a receiver with minimal conversions, and thus less need for front-end gain, has the best potential for superior performance. The most serious limitation on putting most of the gain after the filters is the possibility of BFO signal leakage into the 9-MHz IF strip. The IF strip in this radio has the unusually high gain of 107 dB, and operates at the unusually low signal-input level of –119 dBm—or –128 dBm with the 200 Hz CW filter. This requires exceptional filtering and shielding of the IF strip, the BFO circuits and the power supply leads. The construction methods are briefly described below.

Frequency Mixing Scheme and IF Shift

Fig 2 shows the premixing scheme. The 3 to 4-MHz VFO is converted to 31 to 30 MHz LO injection power at the mixer, and thus 40 to 39-MHz signals are converted to 9 MHz. Most significant is the way the BFO frequency is subtracted, resulting in an IF shift, an essential operating feature. In this way, the BFO frequency does not affect the receive frequency.

With this premixing method, the receiver frequency is the suppressed-carrier frequency for SSB and the zero-beat frequency for CW. If a 14.010 MHz CW signal is tuned for a 500-Hz audio tone, the receiver will read 009.5. An advantage of this method is that if tuned to zero-beat, the receiver reads the actual signal frequency. A more important advantage appears in VHF DX work, where operators frequently shift between SSB and CW—without changing

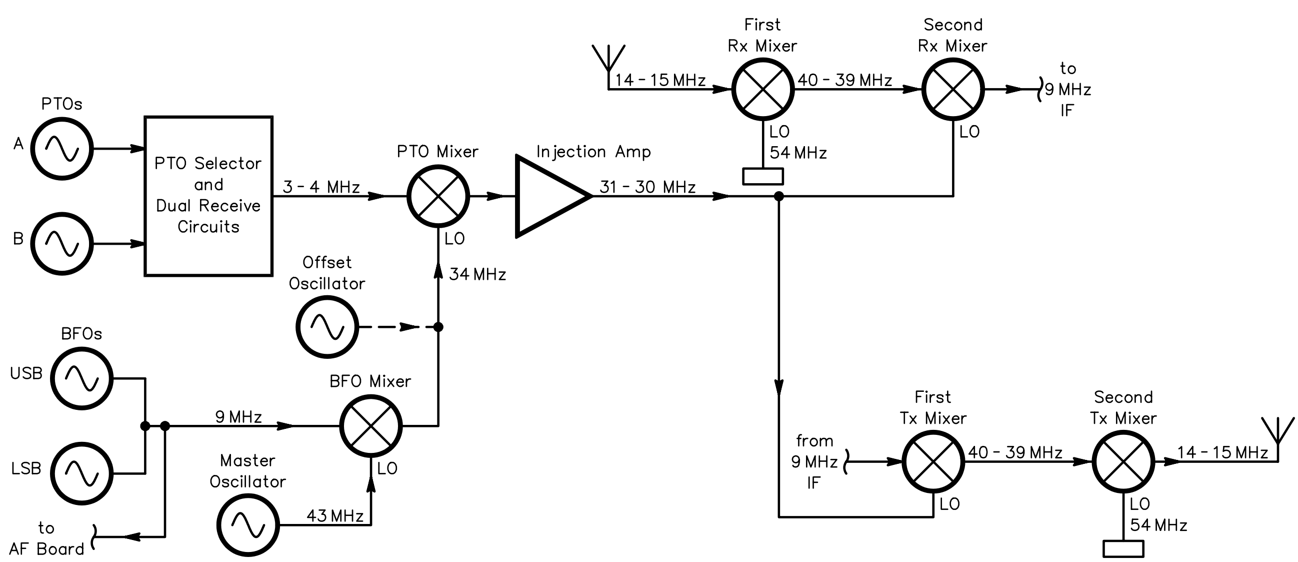

Fig 2—Simplified frequency-mixing scheme. Premixing the low-frequency VFO up to the injection frequencies required by the mixer would be simple enough. The main feature here, however, is the mixing of the BFO frequency to obtain IF-shift operation, a necessity for any radio designed for serious DX or contest work. The result is that the BFOs can be tuned without changing the receive frequency. Credit for this mixing scheme is due to the CX7 designers. Rather than actual frequencies, round numbers are used in this diagram; this allows the basic idea of the mixing scheme to be shown without lots of ugly decimals. See text for details.

frequency—during a single contact. With some “modern” radios, a frequency shift occurs when changing modes, or changing CW filters, often resulting in a lost contact. With the system used here, the CW signal is transmitted within the SSB channel.

For transmitting CW, the master oscillator is disabled, and the offset oscillator is used for injection at the PTO mixer. The BFO still provides the signal for the CW carrier gate on the AF board. Thus the offset oscillator can be used to vary the transmit frequency. Using the round numbers from Fig 2, setting the offset oscillator to 33.9995 MHz (via the front panel **CW OFFSET** knob) will result in a 500-Hz offset. The operator will hear a 500-Hz tone when the offset monitor button is pressed.

This frequency mixing method is exactly as in the CX7, although in Fig 2, round numbers have been used in order to show the basic idea without lots of ugly decimals. The actual VFO frequency range is 3.1 to 4.1 MHz, eliminating an obvious spur on the entire 80-meter band. To accommodate this shift, the master oscillator runs at 43.1 MHz. The actual crystal filter center frequency is 8815 kHz, and the actual BFO frequencies are 8816.5 kHz and 8813.5 kHz. Any crystal filters in the 9 MHz range with corresponding BFO crystals may be used, with no circuit change. The actual frequencies in Fig 2 are now easily calculated; in USB and CW modes the BFO is at 8.8165 MHz, the BFO mixer has output at 34.2835 MHz, and the PTO mixer output, for LO injection, has the range 31.1835 to 30.1835 MHz.

The functioning of the IF shift feature can be seen in Fig 2. When receiving USB signals, for example, the BFO frequency is above the crystal filter passband, because of the high-side injection and the resulting inversion in the front-end. If the BFO frequency is decreased somewhat using the **IF SHIFT** control on the front panel, it will be closer to the crystal filter passband, resulting in a lower-frequency audio passband. At the same time, the output frequency of the BFO mixer will increase, causing the PTO mixer output frequency (the injection frequency) to also increase, in the same amount. This causes the frequency of the signal in the IF strip to decrease, so it mixes with the BFO to produce the same audio tones as before. Thus, the receiver is still precisely tuned to the station. One can work out the simple formula for the frequency of the receiver audio output, using the signal frequency, the transmitted tone frequency and all the oscillator frequencies. The BFO frequency appears twice, with opposite signs, and cancels out.

Front-Panel Controls

While not having the bells and whistles that a microprocessor-controlled radio might have, this radio does have some features that, as least for this operator, beat current factory radios in operating convenience. I do have a few late-model radios up at my Horse Mountain VHF contest station (at 7900 feet) and do know that memories and other features are useful and fun, and that one can become accustomed to menu-driven controls. Nevertheless, I find that the traditional panel controls on my homebrew radio at home allow quicker and easier operation in the heat of battle, such as during a rare one-minute DX opening on 160 meters, or during a contest.

The front panel is shown in Fig 3. The template in Fig 4 shows the controls and labels clearly. Some of the control features are discussed below.

The dual PTOs have separate, large tuning knobs and separate, large RIT knobs. This feature is very desirable; in some factory radios, the RIT, once activated for one VFO, also affects the other VFO, where it is not wanted. The separate RIT circuits are turned on by simply pulling the knobs.

The PTO control switch (**A/B**) is a three-position, black Bakelite bat handle lever switch, directly beneath the digital frequency display. The right position selects PTO A, left selects PTO B. Center position selects split; receive on A, transmit on B. What about the opposite split? I never use it. It’s best to acquire a fixed habit in this regard and stick to it; then there is less chance of transmitting on the wrong frequency. If desired, the opposite split may be obtained using the monitor switch.

For monitoring the B channel (with counter read-out), there is a momentary push button directly beneath the digital frequency display, and a lever switch at the lower left. I thought that having the button in the middle would be most convenient, but toward dawn after a long night on the 160-meter band, the lower switch is used more often. All the lever switches on the front panel are old Switchcraft models with black Bakelite levers. They’re still found on the surplus market. They are manufactured in either momentary or fixed styles, but the fixed version can be used either way: Lean gently on the lever for momentary action, push harder to make it hold. For example, the **KEY** switch can actually be used to send CW in an emergency, or pushed fully down to hold for a steady carrier.

When the lower B-channel monitoring switch is pushed up, it lets you set the PTO B frequency while listening to PTO A. A panel control sets the amount of **CW OFFSET**, and an audio-monitor momentary push button to hear the offset. The controls are arranged so that the button can be pushed—and the **CW OFFSET** knob turned—with one hand. Another knob sets the **OFFSET LEVEL** of the tone in the headphones. The transmit-sidetone frequency is independent of the offset, and is set internally, with a level control on the front panel. The **DUAL RX** feature is turned on and off by simply pulling that knob; the knob adjusts the balance.

It is universally agreed that the best speech processing method is RF clip-

LISA MANDLEKERN

Fig 3—Front panel of the basic 40 to 39-MHz transceiver. Four of the operating features are enabled by means not apparent in the photo. The A RIT, B RIT, BLANKER, and DUAL RX knobs are all attached to potentiometers with ganged push-pull switches. Pulling the knob out actuates the function.

ping. This requires a second SSB crystal filter, but the extra DX countries and contest points make it well worth the trouble. Separate **MIKE**-gain and **CLIPPING** controls are provided on the front panel. After changing mikes, one calibrates the radio for the new mike by simply turning the **CLIPPING** control fully ccw and adjusting the **MIKE** control for minute indications on the **CLIP** meter (at clipping threshold). Then the **CLIPPING** control is advanced to the desired level. Microphone calibration ensures the best gain distribution in the transmit audio circuits, the proper audio level at the balanced modulator, the best transmit-audio quality and optimal carrier suppression. The **CLIP** meter circuits are adjusted so that full-scale on the meter represents 20 dB of RF compression; one-quarter scale represents the normal level of 6 dB compression.

In addition to the front-panel **MIKE** jack (which I never use), there is a high-level speech input jack on the rear panel, which receives audio from the station audio-distribution system and digital voice recorder.[8] There is no **KEY** jack on the front panel; I don't like a clutter of cables on the operating bench. There are **KEY**, **KEYER** and **FSK** jacks on the rear panel.

The radio has two analog meters on the panel. In receive, one meter reads **SIGNAL** level. When the squelch circuit is in use, the other meter reads integrator voltage (**SQUELCH**); this allows quick and easy squelch-level adjustment. In transmit, one meter reads **ALC** voltage. In SSB mode, the other meter reads RF-**CLIP** level. Constant monitoring of these two meters ensures good signal quality.

Full break-in CW operation (QSK) is included for HF: no dit shortening no lag and break-in ability up to 50 WPM. Semi-break-in (**SQSK**) is available for all bands; the delay is set by a knob on the panel. A three-position panel switch chooses **QSK**, **SQSK** or neither. A Curtis keyer chip is included, with a **SPEED** control on the panel. The CW-waveform make and break times have separate internal adjustments.

The **TUNE** switch provides **PULSE** tuning (described earlier) if pushed up or a steady carrier if pushed down.

The **AF GAIN** (AFG) knob on the front panel also functions when using an external audio DSP filter. This is much more convenient than dealing with the AF-gain control on the external unit. For this purpose, audio jacks on the rear panel provide connections to the line-level input/output jacks on the DSP unit. Also included are special amplifier and attenuator circuits to set (and forget) the proper drive level to an external unit, and to equalize the DSP on/off audio levels in the radio.

In lieu of the RF gain (RFG) control found on almost all receivers, this transceiver uses an **IF Gain** (IFG) control. This IFG control lowers the gain only of the IF strip, leaving all stages ahead of the sharp crystal filters running at full gain. This preserves full sensitivity for weak-signal work. The main use of this IFG control is for "AGC threshold" operation. For this reason, the circuit is arranged so that using the IFG control does not cause the S-meter to read upwards. AGC-threshold operation is very effective with extremely weak 160-meter DX signals and with EME (earth-moon-earth, moonbounce) signals. Even the best AGC system is not as good as a well-trained ear.

In this radio, the AGC threshold is about 10 dB above the MDS (minimum-discernible signal) level. Available gain is sufficient so that ambient antenna noise activates the AGC system. For AGC-threshold operation, the **IF Gain** is reduced so that weak signals and noise are a few decibels below the AGC threshold. In effect, the AGC threshold is raised; there is no decrease in sensitivity. The ear can now hear the weak signal in the noise. The AGC system is prevented from reducing the receiver gain at every little static crash. On the other hand, large static crashes and loud signals will activate the AGC system and protect the operator's ears. This is impor-

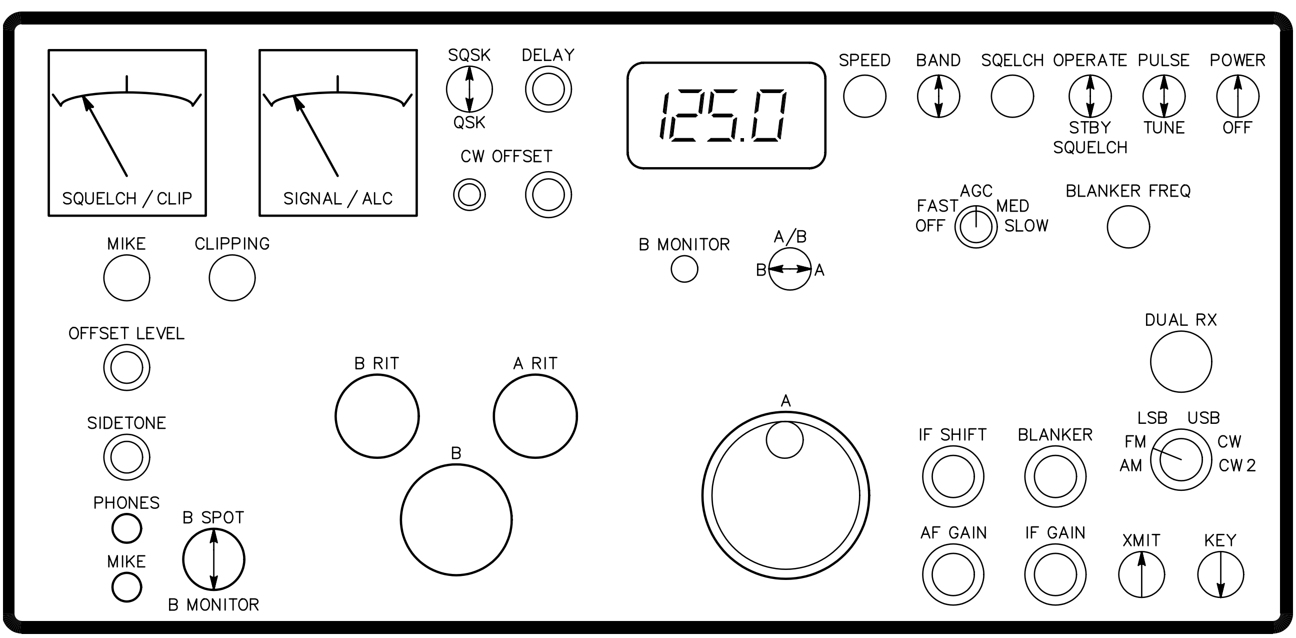

Fig 4—Front panel layout. This shows the transceiver controls and their labels. The digital readout counts the PTO outputs directly, and shows only the kilohertz part of the transceiver frequency. The megahertz band is indicated by band switches on the front-end panels. For details of the operating features, see text.

tant for hams who wish to work DX for many, many years. This operating method is safe, in contrast to the all-too-common practice of turning off the AGC.

Most radios suffer from two RFG control problems: Most RFG controls cause the S-meter to read upwards, making it very difficult to set the correct AGC-threshold operation level. Also, most RFG controls reduce the front-end gain; this reduces receiver sensitivity and causes signals to be lost. The IFG control in this radio raises the AGC threshold; the S-meter correctly indicates signal levels relative to the set threshold, and receiver sensitivity is not reduced.

The **STBY/OPERATE** switch is combined with the **SQUELCH** switch. **OPERATE** is up, **STBY** center, and **SQUELCH** down. Thus switching to **SQUELCH** automatically switches to **STBY**, and disables the transmitter. This is the simple one-touch procedure for leaving the radio on an HF DX operation frequency or a VHF DX calling frequency and going back to the workbench.

The **SQUELCH** level control is a separate knob, not linked to the **SQUELCH** switch, so it does not have to be reset each time the squelch is turned on or off.

There is one last, very important feature: There are no concentric dual knobs!

Construction and Layout

Fig 5 shows the basic transceiver, with four main circuit boards. Each board is hard-wired to the radio, with a 12-inch-long harness. This allows the boards to be lifted out for testing while in operation. Sub-boards are mounted on hinged spacers, for immediate access. The PTOs are mounted to the front panel, which can be removed in seconds, with no knobs or couplings to remove. The rear panel, which carries the power supply, receiver audio-output module and input/output jacks, can be removed in seconds.

All the gear in the shack is built on standard 19-inch black rack panels. The basic 40-MHz transceiver is built on a 8.75-inch-high panel. The 200 W HF front-end is on a 7-inch panel. The 2-W 6 and 2-meter front-ends are each built on a 3.5-inch panel.

The need for exceptional filtering and shielding was mentioned above in the section under "Receiver Gain Distribution." To exemplify my techniques, the three signal boards can be used. Each measures about 7.5×15×2 inches and is constructed using double-sided circuit-board material. A floor is soldered-in, forming a 1.5-inch-high living space for the circuits, and a 0.5-inch-deep space below for filtering. All power and control leads go through soldered-in feedthrough capacitors in the floor, and then through additional filters in the "basement." Individual stages are separated by soldered-in walls. Signal leads pass through tiny windows in the walls. As required, some compartments are fitted with shielded ceilings.

Some of the surplus parts used in this design may not be readily available. If readers want to duplicate the radio, perhaps only in part, any of the currently available replacement crystal filters in the 9-MHz region could be used. Because of the unique premixing scheme, no circuit changes will be needed; only the two BFO crystals will need to be selected accordingly. The VFO problem will always present an interesting challenge.

The block diagram in Fig 1 reflects the physical layout into four main boards called LOGIC, RF, IF and AF. Smaller assemblies are mounted on the front and rear panels. The four boards, the panels and the HF front-end formed the six main phases of the project. A total of three years was scheduled, six months for each phase. I started with a rough overall plan, but not a complete plan for each phase; no schematics at first, only input/output specifications. Typically, the six months work for each phase were divided into four months of designing and two month's of building. General descriptions of each of the six main parts of the radio are given below, with emphasis on design criteria and operating features. Subsequent articles will deal with circuit details.

Logic Board

No one who has replaced the mode switch in a traditional radio has fond memories of the experience. With up to six modes, including several CW filter choices and numerous circuits to control, there can be dozens of leads. After fighting mode switches for decades, I used modern logic ICs to put an end to this nightmare.

There are even better reasons for using logic control. One is crystal filter switching. Switching filters directly at the mode switch is problematic in that it requires the filters to be somewhat exposed. In contrast, logic control with miniature relays allows careful shielding and greatly improved ultimate rejection. That is, there is reduced "blow-by"—the signals that leak around the filter.

Logic-mode switching also offers mechanical advantages that are important for the experimenter. The mechanical linkages and couplings that connect a traditional mode switch to the front panel are eliminated. I


LISA MANDLEKERN

Fig 5—A view from above.

used no such linkages in this radio. A further advantage is the ease with which one can make modifications to the various control functions, simply by adding a few TTL gates.

Controlling six modes, the mode switch used here has merely six TTL-level leads, and a ground. The logic board converts the TTL signals to whatever levels are required to control the various circuits. In addition to the mode-switch signals, some panel controls (**PULSE / TUNE**, **STBY**, **A / B**, **DUAL RX**, **B MONITOR** and **B SPOT**) are converted to TTL signals that are combined in the logic chips for control of various circuits. This makes it very easy to implement any desired function. For example, one touch of the **TUNE** button automatically shifts the radio into CW, silences the mike, closes the PTT and KEY lines, disables the sidetone and shifts the carrier into the clear channel you have found.

The PTOs are also controlled by the logic board. The single, three-position PTO lever switch has only two leads and a ground, instead of a 16-pole, three-push-button, 24-wire assembly, plus an 8-pole, 16-wire concentric rotary knob obstacle course, as in the CX7. PTO control for flexibility is of special concern. The radio has two PTOs and dual-receive capability. In addition, it provides instant one-button monitoring of the second channel, for split frequency DX operations. (Dual receive is not suitable for extremely weak, barely readable DX signals.) In addition, there is provision for spotting the transmit frequency (as for 40-meter DX SSB work) while receiving. The logic board must select which PTO to read on the frequency counter in any given situation. For all this PTO and counter control, it was expedient to go beyond the simple TTL gate chips and use the larger data selectors, the 74151, as logic-function generators. Each variable is assigned a symbol, the logic function that gives the desired result is written and the chip is wired accordingly.

RF Board

This board contains the balanced JFET receive mixer that converts the 40 to 39-MHz signals to 9 MHz, and the balanced MOSFET transmit mixer that does the reverse. It also contains the two 9-MHz BFOs, the premixing circuits that enable the IF-shift circuit, the dual-receive circuits and the front-end of the tunable noise blanker. Premixing does allow the possibility of spurious responses or birdies, so there is very extensive filtering and shielding.

This board includes the oscillator that enables the adjustable CW offset. Also included is the mixer that mixes the offset oscillator with the normal output of the BFO mixer, to produce the audio tone for headphone monitoring of the offset.

IF Board

The IF strip functions in both receive and transmit modes, using six MOSFET stages, and several additional switching and buffering stages. The board includes the crystal filters, the RF clipping (and clipping metering) circuits and the AGC circuits.

For SSB, there are matched 2.4-kHz filters at the input and at the output of the IF strip. The result is a 2.0-kHz passband. The second filter is essential, because the RF clipping takes place within the IF strip and produces distortion products. This is an unavoidable effect of RF clipping. Because of the greatly increased effectiveness of the SSB signal, it is well worth the effort and cost of the second filter. The gain of the IF strip is reduced greatly—to a fixed level—during SSB transmissions. Clipping occurs at a fixed level within the strip. The amount of clipping is adjusted by varying the gain of a transmit stage leading to the IF strip. The transmit drive level is set on the front panel of each front-end module. There is no transmit output level control on the transceiver's panel—it would have to be readjusted when switching between front-ends.

RF speech clipping requires careful monitoring. A meter provides more definite indications than does headphone feedback. The meter indicates the actual amount of clipping, and alerts the operator to changes caused by fatigue, over-enthusiasm or changes in mike position. The RF clipper itself is not expensive—two diodes at a total cost of 10 cents—but associated circuitry is required for convenient panel adjustment, reliable operation and accurate metering. **CLIP**-meter calibration is independent of mike-input levels or the setting of the **MIKE** control.

For CW, narrow IF filters can be switched in. There is a 200-Hz filter at the IF strip input, and a matched 250-Hz filter at the output. The second filter removes excess noise developed in the IF strip. Many radios exhibit a disturbing behavior when a sharp CW filter is switched in. Their gain drops noticeably. In this radio, an extra stage of amplification ahead of the sharp CW filter compensates for the loss. When the sharp filter is switched in, the S-meter holds steady, and there is no change in the sound of the received CW signal, except that the interfering signals are gone, and the noise-level is reduced.

The AGC system has been fully described in a previous article (see Note 1). The AGC system can make or break an otherwise good radio. Poor attack performance—with clicks and thumps—can cause operator fatigue, a crucial factor in all-night DX operating or all-weekend contesting. Poor decay performance can cause excessive receiver recovery time after a strong signal ceases transmitting, preventing a weak signal from being heard. Even worse, poor decay characteristics can sometimes cause excessive receiver recovery delay after every transmission, preventing reception at precisely the most important time. To avoid these problems, hang AGC circuits with carefully controlled timing are required.

AF Board

This board contains the product detector, balanced modulator, sidetone oscillator and low-level AF circuits. The need to isolate the BFO and the IF strip has been noted above. Hence, the BFO signal from the RF board is routed to the AF board at a low level, then amplified and fed to three BFO gates. Two of these feed the product detector and balanced modulator; the third is the carrier gate for CW.

The station microphone audio-distribution system includes a one-stage, high-pass filter, mainly to eliminate hum introduced by an external digital voice keyer (see Note 8). The AF board includes two stages of high-pass filtering, effectively eliminating any residual hum.

PTOs, Counter and Power Supply

The PTOs are straightforward, each with four stages of buffering. To avoid any possible spurious emissions during split operation, the PTOs are powered on and off at each TR transition. This permits full break-in operation with no chirp. The counter is simple, using 7400 series TTL chips. It reads to 100 Hz, which I find adequate.

The power supply may seem overdesigned, but there is a reason. It results from years of experience with radios in which transient pulses travel between stages, and boards, by way of the power-supply circuits, or even a common power transformer. The elimination of these transients is crucial for proper AGC performance. The result is four separate regulated supplies, with four separate small transformers. These power only the basic transceiver, with its 200-µW exciter output. Locating the 200-W HF power amplifier on a separate panel keeps the rest of the transceiver cool and stable. The four separate small transformers, rather than being a problem, yield advantages in acquisition, mounting, space fitting and cost over a four-secondary transformer.

The four supplies include ±18 V for the main boards and +8 V for the logic board and counter. The fourth supply is a separate +18 V supply for the receiver audio-output module, since this stage is often a serious offender in producing power-supply transients. Each of the four main boards and the counter then have individual on-board regulators for +15, –15 or +5 V, as needed. This double regulation avoids all transient problems.

HF Panel

The little-known 8072 external-anode tetrode—manufactured by RCA among others—is a gem. Only half the size of a 6146, it easily delivers a linear 200 W, even up on the 2-meter band, where I also use one. It is rated to 500 MHz. Being conduction-cooled, it needs no noisy forced-air cooling, but does require a special anode clamp and heat sink, salvaged from a junked CX7 found at a flea market. Although it is not necessary, a small inaudible muffin fan is attached to the heat sink. The fan comes on only when transmitting and has a timer to keep it going for one minute after each transmission. This means that it runs continuously during a contest. A new 8072 is expensive. I use a total of eight of these tubes on various frequencies, but they were all obtained as used surplus at very little cost. With adequate protection circuits, they seem to last forever.

The CX7 was the first ham transceiver produced with so-called "broadband PA tuning". In fact, the PA used a pi-L circuit on the seven 1-MHz bands, switching banks of internally adjustable tune and load trimmer capacitors. This is merely a fixed-tuning arrangement. It works well enough on the lower bands, but the poor L/C ratio on 10 meters makes coverage of an entire 1 MHz segment difficult. In addition, coil-turn shorting and toroid-core losses result in reduced output on 15 meters. In addition, a set of manual controls was provided for use when desired.

This homebrew radio uses a variation on this theme. Ten 1-MHz bands were to be covered. The lower seven bands use seven separate fixed-tuned pi or pi-L networks, with no provision for manual control. For the three upper HF bands (24, 28 and 29 MHz) there is one manually tuned pi-network with front-panel controls and no provision for fixed-tuned operation. This arrangement suits my operating habits ideally. It improves performance by avoiding shorted turns on a single tank coil, and using more-appropriate components and better L/C ratios. For contesting, the panel controls are tuned for 28 MHz, and the result is equivalent to fixed-tuned operation on all bands.

A drive control on the panel of the HF front-end is adjusted for the correct amount of ALC compression—measured at the 8072 grid—or for proper drive level to an external 1.5-kW amplifier. ALC also runs from each driver and each kilowatt amplifier back to the corresponding front-end panel, with ALC metering at the transceiver. There has been much written about distortion caused by ALC circuits. It is true, but applies only to improperly designed and improperly adjusted ALC circuits. This radio applies ALC control voltage to dual-gate MOSFETs, as is common. An IMD problem can occur if too much gain reduction is attempted by varying the bias of a MOSFET. An extreme case occurs when ALC is improperly used to reduce the gain of a radio to drive a low-input-level transverter. At the K5AM station, the gain of each front-end panel is adjusted to obtain 3-dB ALC compression, a moderate amount. A single MOSFET may be able to handle a 3-dB gain reduction. To ensure the cleanest signal possible, however, ALC voltage is applied to three cascaded MOSFET stages, so that each reduces the gain by only 1 dB.

The drive control could be used to reduce the HF panel output to 5 W for QRP work, which presents refreshing challenges. Using the drive control, however, involves the usual touchy adjustment problem and band-change inconvenience. The HF front-end section has a separate **QRP** level knob, and a switch to enable it. The knob adjusts an output detector that feeds the ALC circuits. The drive control is used to obtain the correct amount of ALC compression, read directly on the transceiver panel. It is not as critical on CW as on SSB, but too much ALC compression will distort the CW waveform.

The **QRP** switch is part of the **AMP** (amplifier) switch, with three positions: **LP** (low power), **BF** (barefoot, meaning transceiver only) and **PA** (external amplifier enabled). The 6 and 2-meter front-ends have similar switches. The entire station is controlled from the front-end panels conveniently located near the operator. On HF, the three positions instantly provide 5 W, 200 W or 1500 W output. The **LP** position can be used with the level set for 150 W for low-power contesting.

There is a panel switch for a second receive antenna, which is a necessity on 160 meters and useful on other bands, as well. The receiver circuits are fully protected—using reed relays—against RF energy picked-up by the auxiliary antenna while transmitting on the main antenna. A jack on the rear panel allows the auxiliary receive antenna relay to be controlled by a button on the desk, or by a foot switch.

Summary

This article gives a general description of a high-performance homebrew transceiver built with regard only to the best performance. It shows what can be accomplished without microprocessors, synthesizers and PLLs.

Notes

[1]M. Mandelkern, K5AM, "A High-Performance AGC System for Home-Brew Transceivers," *QEX*, Oct 1995, pp 12-22.

[2]M. Mandelkern, K5AM, "Evasive Noise Blanking," *QEX*, Aug 1993, pp 3-6.

[3]M. Mandelkern, K5AM, "A Homebrew Contest Station," *NCJ*, July/Aug 1998, pp 12-13.

[4]M. Mandelkern, K5AM, "A Sensitive Integrating Squelch," *QST*, Aug 1988, pp. 27-29.

[5]M. Mandelkern, K5AM, "Design Notes for 'A Luxury Linear' Amplifier," *QEX*, Nov 1996, pp 13-20.

[6]M. Mandelkern, K5AM, "A Low-Drive, High-Power All-Band Tetrode Linear Amplifier," *CQ*, July 1990, pp 60-65.

[7]M. Mandelkern, K5AM, "A Luxury Linear," *QEX*, May 1996, pp 3-12.

[8]M. Mandelkern, K5AM, "The AMSAFID: An Automatic Microphone Switcher Amplifier Filter Integrator Distributor," *QST*, Nov 1995, pp 47-49. □□

A Double-Tuned Active Filter with Interactive Coupling

You can combine active filters in a way that maintains their passband width and yet steepens their skirts. Come learn how.

By Frank W. Heemstra, KT3J

Those who design or experiment with audio frequency electronic circuits often have need of a simple active band-pass filter that provides more selectivity than can be obtained with a single-pole filter of appropriate bandwidth. In order to satisfy this requirement, designers often contemplate cascading two or more second-order filters. Indeed, my old *ARRL Handbook*[1] shows a CW filter consisting of two multiple-feedback filters in cascade. For simplicity of design and construction, they are identical circuits with the same resonance frequencies.

[1]Notes appear on page 29.

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Cascading the filters increases the selectivity, but it also narrows the passband. To compensate for the bandwidth reduction, we must either lower the Q or use more-difficult stagger-tuning techniques.

This, however, is just the opposite of the way double-tuned inductance-capacitance (LC) filters behave. When a pair of tuned LC circuits is coupled, the passband broadens, rather than narrows, and it cuts off more sharply at the band edges. With critical coupling, the bandwidth is approximately 1.4 times that of a single-tuned circuit, and the passband is nearly flat.

This difference in behavior is a result of the reciprocal nature of coupling. In LC electrical wave filters, electrical energy surges back and forth between adjacent sections. Through such interaction, the behavior of each section of a filter is influenced by all other sections and by the loading at both filter ends.

A Coupling Phenomenon

An insight into the nature of interactive coupling and the fundamental role it plays in filter design can be gained by observing the behavior of a coupled pair of high-Q resonant circuits, or "resonators," when one of them is shock excited[2] into free oscillation. Each of the resonant circuits is tuned to the same frequency, f_0. If either of the circuits is shock excited while there is no coupling, the wave pattern of the damped oscillations is the familiar exponential decay shown in Fig 1.

However, when the pair of tuned circuits is closely coupled and one of them is shocked, the wave patterns of the voltages on the coupled resonators reveal an entirely new characteristic, as shown in Fig 2.

The wave patterns show that after the initial shock, the shocked resonator immediately begins to transfer its stored energy to the coupled resonator. Since the amplitude of its oscillations goes to zero, all of the energy is transferred. The rising and falling amplitudes show that the entire store of electrical energy is passed back and forth between the two "tank" circuits as it dissipates through resistive losses. After each transference, the energy always returns in reversed phase.

Because of the phase reversals at the nodal points, each of the patterns in Fig 2 may be recognized as a composite signal that resembles the double-sideband output of a balanced modulator. Each pattern is, indeed, the sum of the instantaneous voltages of two equal-amplitude oscillations with a small separation in frequency. Thus, in each of the tuned circuits, the damped oscillation occurs at two frequencies simultaneously.

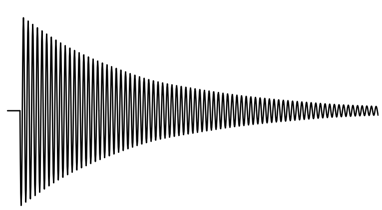


Fig 1—Damped oscillation wave pattern of a shock-excited resonant circuit having a *Q* of 80. The *Q* can be found from the pattern by multiplying the amplitude half-life—in periods of oscillation—by 4.53.

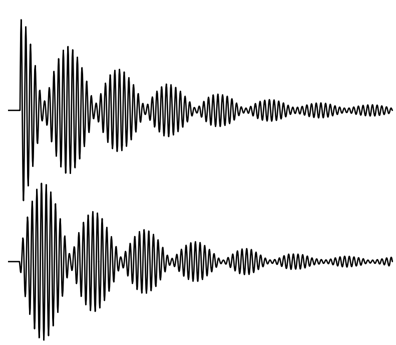


Fig 2—Free-oscillation wave patterns of the voltages on a coupled pair of resonant circuits, each having a *Q* of 80, following an initial shock to the circuit shown by the top pattern. The coefficient of coupling is 0.1.

The phase difference in the modulation envelopes of the two patterns shows that the oscillations have the same phase in both circuits at one of the two frequencies, and they have opposite phases at the other frequency. The difference frequency is the rate at which the energy can be transferred back and forth between the circuits, and therefore, depends on the amount of coupling.

Obviously, these two frequencies, or "normal modes" of free oscillation, are only possible because of the ability of the coupled circuits to exchange stored energy. Tuned stages that are cascaded, where signal energy flows in only one direction, can never behave this way. Tuned circuits with interactive coupling lose their individuality and behave as a unified system with characteristics not possessed individually by either of them.

Double-Tuned Filters

As it relates to filter design, the effect of coupling on the tuned pair of circuits is that it splits the single resonance of each circuit into a double resonance corresponding to the two "normal mode" frequencies of free oscillation. These resonances are the two peaks that appear in the passband response of a double-tuned band-pass filter when the coupling exceeds a critical amount. By increasing the frequency separation, the increased coupling broadens the passband, while leaving the slope of the skirts unaffected.

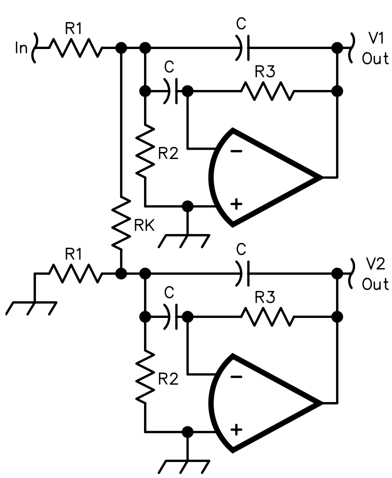
The amount of coupling is specified by a coefficient of coupling, K. This parameter, and the loaded Q of the tuned circuits, entirely determine the shape of the selectivity curve and relative bandwidth of the filter. *Reference Data for Radio Engineers*[3] contains useful data and formulae for double-tuned filters, including a family of normalized selectivity curves with their associated design parameters.

Coupled Active Filters

Although the conventional wisdom in active-filter design seems to be totally ignorant of it, interactive coupling techniques are also applicable to active filters. Coupling the two previously mentioned multiple-feedback filters is amazingly simple. The schematic diagram in Fig 3 shows how a pair of these filters can be coupled to form a double-tuned filter. The coupling takes place through the single resistor, R_k, joining the active tuned circuits.

The coupled active circuit in Fig 3 is the mathematical analogy of the double-tuned LC filter in Fig 4. The voltages at the outputs of its opamps must satisfy equations that are identical to those derived from network analysis of the voltages in the LC-filter tuned circuits. Thus, the active double-tuned filter is an "analog" circuit in the true and original meaning of the word.

The LC circuit in Fig 4 is an "ideal" filter. It is not practical because L_k must be nearly a pure inductance with a relatively high value. Of course, the reason for using active filters is to accomplish what would be impractical with LC filters. My active circuit duplicates the theoretical behavior of the idealized LC filter in every way.



$$\omega_0 = \frac{1}{\sqrt{T_3 \bullet T_0}}$$

$$Q = \frac{1}{2}\sqrt{\frac{T_3}{T_0}}$$

$$K = \frac{R0}{R_k}$$

Where:

$$T_3 \equiv R3 \bullet C$$

$$T_0 \equiv R0 \bullet C$$

$$\frac{1}{R0} \equiv \frac{1}{R1} + \frac{1}{R2} + \frac{1}{R_k}$$

Fig 3—Schematic diagram of a coupled pair of multiple-feedback, second-order active band-pass filters. Interactive coupling takes place through the mutual resistance, *RK*, joining the circuits.

Analogies: Reactive Coupling

The interaction through R_k is mathematically identical to the coupling interaction through the mutual inductance L_k in the LC filter. Thus, the connection of R_k is analogous to reactive coupling.

As regards the phase response: In conformity with reactive coupling, the output voltages of the active tuned circuits have a 90° phase difference at the resonance frequency, f_0. As for reciprocity, it is apparent from the symmetry of the circuit diagram in Fig 3 that the two multiple-feedback filters have a reciprocal relationship to each other. Neither of them has a preference as an input or output stage. This ability to propagate a signal in either direction is a necessary condition for the coupling interaction. The unused input of the filter selected as the output stage must be grounded to avoid shifting its resonant frequency or altering its Q.

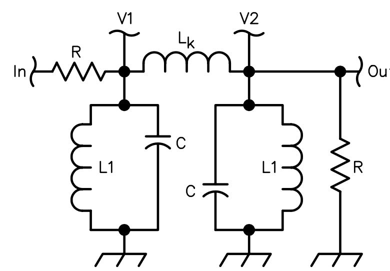
Selectivity Curves and Matching Loss

The selectivity curves of the analog filter are characteristic of double-tuned filters and conform to theory. The three measured curves in Fig 5 are a parametric family, with KQ as the parameter. The Q is fixed at 7.6. In the flat-topped response of curve A, the tuned circuits are critically coupled ($KQ = 1$). In the double-humped curves B and C, they are over-coupled, with KQ = 1.5 and 2, respectively.

In the symmetrical LC filter of Fig 4, there is a 6-dB input-to-output voltage-matching loss at the frequencies of peak response. This loss is accurately simulated by the analog circuit if the multiple-feedback filters have unity gain.

Curve Symmetry

The selectivity curves of double-tuned filters with reactive coupling are not symmetrical on a logarithmic frequency scale, as single-tuned and cascaded filters are. The reason for this is that the mutual reactance is frequency-dependent. The inductive reac-



$$\omega_0 = \frac{1}{\sqrt{C \bullet L0}}$$

$$Q = R\sqrt{\frac{C}{L0}}$$

$$K = \frac{L0}{L_k}$$

Where:

$$\frac{1}{L0} \equiv \frac{1}{L1} + \frac{1}{L_k}$$

Fig 4—An idealized, double-tuned LC filter that is mathematically equivalent to the coupled, active-filter circuit in Fig 3. The active filter duplicates the theoretical behavior of this LC filter in every way.

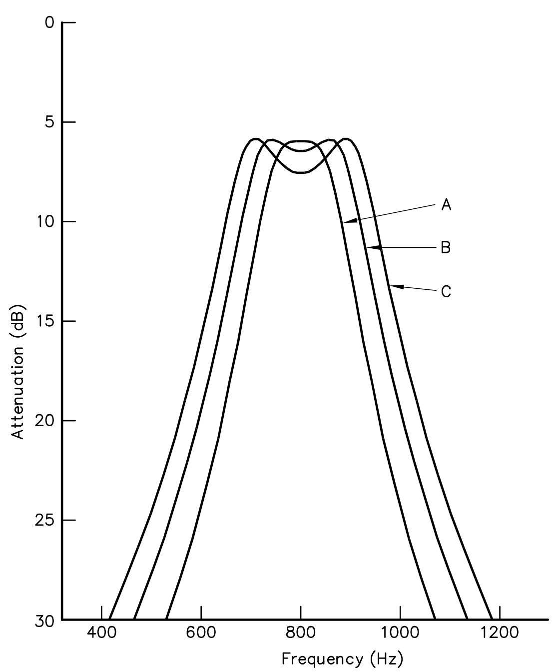


Fig 5—Measured parametric family of near-resonance selectivity curves for the active double-tuned filter in Fig 3. The parameter, *KQ*, is 1, 1.5 and 2 in curves A, B and C, respectively. The resonant frequency, f_0, and *Q* are the same in all three curves.

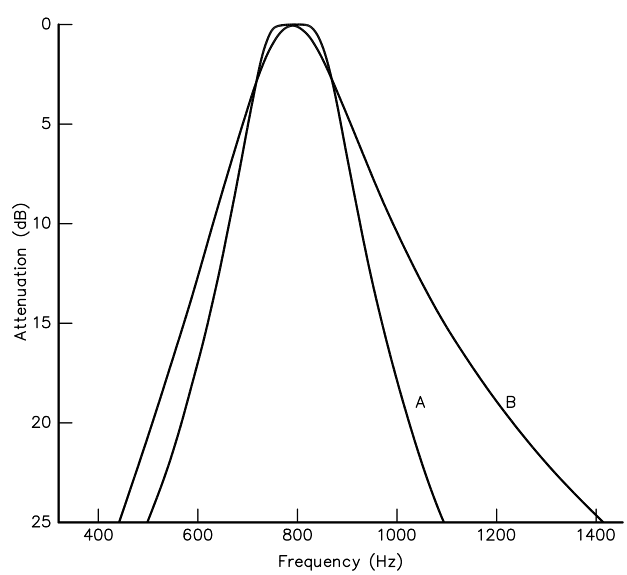


Fig 6—A comparison of the selectivity curves of two fourth-order active band-pass filters, having equal bandwidths and band-center frequencies. Curve A is the response of the double-tuned active filter in Fig 3, and curve B is that of a cascaded pair of multiple-feedback filters.

tance of L_k, which couples the tuned circuits in Fig 4, causes the high-frequency side of the curve to attenuate more rapidly, causing it to be nearly symmetrical on a linear frequency scale. Because of analogous coupling in the active circuit, the curves in Fig 5 are also symmetrical. Cascaded filters cannot duplicate this characteristic.

For high-frequency narrow-band filters, the distinction between these two kinds of symmetry makes little difference. With audio frequency filters, however, where the bandwidth is not small relative to the band-center frequency, there is a significant difference in the attenuation at frequencies above the passband. Fig 6 compares the curve for my double-tuned filter with that of a cascaded pair of multiple-feedback filters. The parameter KQ is 1.1 in the double-tuned filter. For a fair comparison, the cascaded filter was designed to have the same 3-dB bandwidth and band-center frequency as the coupled filter. On a logarithmic frequency scale, the curve for the cascaded filter is symmetrical.

Design: Coupling Coefficient

In the active double-tuned filter, the coefficient of coupling is equal to the resistance ratio, $R0 / R_k$, where $R0$ is the resistance of $R1$, $R2$, and R_k in parallel. The required value of R_k for a given coefficient, K, may be obtained from the formula:

$$R_k = \left(\frac{1-K}{K}\right)\left(\frac{R1 \bullet R2}{R1+R2}\right) \quad \text{(Eq 1)}$$

The important parameter, KQ, however, turns out to be one-half the ratio of the reactance, X_C, of the capacitors at resonance to the mutual resistance, R_k. Therefore, for a given parameter, KQ, the required resistance of R_k is determined entirely by the value of C and the resonant frequency, f_0, regardless of the Q.

Loading Effect

Each of the circuits is loaded by R_k in exactly the same way as if it were in parallel with $R2$. Consequently, both the resonance frequency, f_0, and the Q of the circuits are increased by the connection. If existing circuits are coupled without modification, f_0 and Q in both circuits will increase by the fractional amount:

$$\frac{\Delta f}{f} = \frac{\Delta Q}{Q} = \frac{1}{\sqrt{1-K}} - 1 \quad \text{(Eq 2)}$$

For example, if R_k is calculated for $K = 0.2$, its connection will cause f_0 and Q to increase by the fractional amount 0.118, or 11.8%.

This shift in f_0 and Q, which results from the connection of the mutual resistance, R_k, is another interesting analogy. As expressed above in terms of the coupling coefficient, K, the shift is precisely the same as occurs in the tuned LC circuits in Fig 4 when the mutual inductance, L_k, is connected.

Design Modification

Since the loading effect of R_k is equivalent to putting R_k in parallel with $R2$, existing circuits can be compensated with a modified value of $R2$. The required values for R_k and $R2$ can be calculated from Eqs 5, 7 and 8 of my complete design procedure, outlined below.

My schematic diagram obviously assumes positive and negative supply voltages. Biasing of the noninverting inputs for operation from a single supply is properly left to the designer's own ingenuity.

Design Procedure

It is not necessary to use precision components for this filter. My design and alignment procedures achieve excellent performance and accurate frequency alignment with standard 5% component values.

Part of the procedure is that, after each of the successive calculations of $R3$, R_k and $R1$, standard 5% resistance values closest to those calculated should be selected and their nominal values used in all subsequent calculations. $R0$ is not an actual resistor used in the circuit. It is a calculated value used only for the subsequent calculation of $R2$.

Specified are f_0, Q, C, KQ and $GAIN$. Calculate:

$$X_C = \frac{1}{2\pi f_0 C} \quad \text{(Eq 3)}$$

$$R3 = 2Q\, X_C \quad \text{(Eq 4)}$$

$$R_k = \frac{X_C}{2KQ} \quad \text{(Eq 5)}$$

$$R1 = \frac{R3}{2 \bullet GAIN} \quad \text{(Eq 6)}$$

$$R0 = \frac{X_C^2}{R3} \quad \text{(Eq 7)}$$

$$R2 = \frac{1}{\frac{1}{R0} - \frac{1}{R_k} - \frac{1}{R1}} \quad \text{(Eq 8)}$$

Frequency Alignment

The criterion for good performance of this filter—just as that for LC band-pass filters—is accurate frequency alignment of the tuned stages. Fine tuning the active filters to the specified resonant frequency also assures that the calculations of $R3$, R_k, and $R0$ are correct for the specified Q and K.

Using the alignment procedure described below, each of the two single-tuned circuits should be tuned to the resonant frequency, f_0, within a small fraction of $f_0 / 2Q$.

The Q and the coupling coefficient, K, are not critical. It is evident from the curves in Fig 5 that large errors in these parameters are tolerable.

Step 1: To align the resonant frequencies, it is first necessary to use a standard resistance value for $R2$ that is slightly higher than the calculated value. A permanent space should be provided on the circuit board for adding a resistor in parallel with $R2$.

Step 2: The next step in the procedure is to accurately measure the resonance frequency of each stage without coupling. For this measurement, the coupling resistor, R_k, must be temporarily connected in parallel with $R2$ in the circuit under test.

Fig 7 shows the diagram of an auxiliary test circuit that I have found convenient for measuring the resonant frequency. The measurement is fast and accurate. The limiting amplifier supplies positive feedback to the filter, causing it to oscillate at its resonant frequency. The response can be read with a frequency counter. The limiting amplifier should provide a loop gain slightly greater than unity.

Lacking a counter, other means can be used to find the frequency for which the filter's input-to-output phase difference is exactly 180°. A calibrated audio signal generator and an oscilloscope with an X-Y mode would be useful.

Step 3: Since the initial value of $R2$ was selected to be higher than the cal-

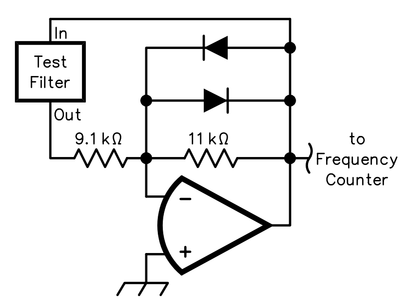


Fig 7—Diagram of auxiliary test circuit used to measure the resonant frequency of inverting, single-pole, band-pass filters. The limiting amplifier provides positive feedback to make the circuit oscillate at its resonant frequency.

culated value, the frequency measured in Step 2 will be lower than the specified frequency, f_0. This measured frequency, f_1, and the previously calculated value, $R0$, are used in the equation below to calculate a resistance value, R_p:

$$R_p = \frac{R0}{1-\left(\frac{f_1}{f_0}\right)^2} \qquad \text{(Eq 9)}$$

Step 4: The calculated value of R_p above is the $R2$ parallel resistance necessary to increase the resonant frequency of the circuit to the specified frequency, f_0. Place a standard resistance closest to this calculated value of RP in parallel with $R2$ and check the resonant frequency again. It will usually be within one or two hertz of the specified frequency on the first try.

Notes

[1]M. Wilson, K1RO, Ed., *1986 ARRL Handbook* (Newington: ARRL, 1985), p 29-5.

[2]The shock comprises suddenly storing electrical energy in the circuit through a step change of the current in the inductor or the voltage on the capacitor. It may be applied by magnetic induction as in an automobile ignition system, or by suddenly dumping a charge on the capacitor.

[3]*Reference Data for Radio Engineers*, 4th edition, International Telephone and Telegraph Corp, 1956; "Simple Bandpass Design," pp 241-246.

Frank took up Amateur Radio as a technical-interest hobby in 1983 after retirement from the US Naval Research Laboratory, where he spent 30 years as a research physicist. He received his MS in physics from Iowa State College in 1953.

At NRL he worked under the leadership of Chester L. Buchanan, W3DZZ, (who invented the antenna trap; QST, Mar 1955, pp 22-23, 130) in the field of deep-ocean technology and developed expertise in underwater electroacoustics and magnetometry. Frank participated in many deep-sea search and surveillance missions, including the successful searches, led by Buchannan, for the Navy's lost nuclear submarines, Thresher *and* Scorpion.

CW is Frank's favorite mode, because it's fun. For his own edification, he enjoys electronic experimentation and instrumentation. He is especially interested in applications of circuits made to satisfy time-dependent differential equations describing the behavior of known physical systems. (In the bygone days of analogue computing, such circuits were known as "differential analyzers.")

A Low-Distortion Receiver Front End for Direct-Conversion and DSP Receivers

An experiment with Analog Devices' AD831 mixer IC yields a front end with an impressive +30 dBm third-order intercept.

By Detlef Rohde, DL7IY

Introduction

In this article, I describe a low-distortion receiver front end that has a high third-order intercept point (+30 dBm) and is very useful for both direct-conversion (D-C) and DSP receivers. It includes a broadband phase shifter to create the in-phase and quadrature (I and Q) injection signals for phasing-method SSB and CW reception. Conversion to baseband or to an IF are both possible. Audio or IF DSP may then be used.

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Motivation

When constructing my KK7B-like shortwave transceiver a couple of years ago, I followed Rick's suggestions[1, 2] and used high-level passive mixers. I experimented with various networks to get the needed phase shift of 90° for the I and Q mixers. I found these networks to have less bandwidth than I desired. After doing some *SUPERCOMPACT* simulations of broadband phase shifters, which were cascaded hybrids like those R. E. Fisher, W2CQH, introduced years ago,[3, 4] I found the phase splitting technique to be rather complicated when there is a need to cover more than one octave. In 1995, Mini-Circuits introduced a broadband 90° phase splitter that covers nearly the entire shortwave range (3-30 MHz). Its phase imbalance is specified to be no more than 3.5°.

As shown by the calculations of Byron Blanchard, N1EKV,[5] this is too large for good sideband suppression in D-C systems. On the other hand, it is good enough when accompanied by DSP phase tweaking in a following IF stage. Unfortunately, the price of this device is high; my decision not to use it was easy.

When first setting up my DCTRX (direct-conversion transceiver) in 1993, I was fascinated by the wide spectrum of experimental options for it, all of which seemed to need techniques that were to be developed in the next decade! My first VFO for 14 MHz was analog. Because of its

[1]Notes appear on page 33.

poor stability, lack of precise tuning and readout ability, I constructed a DDS VFO with the then-new AD7008 chip from Analog Devices. This project, which I called DDS1, forced me to step into programming of computers using various high-level languages (BASIC, Pascal, Borland Delphi). I wrote about the project in the July and August 1995 issues of the German ham-radio magazine *FUNKAMATEUR*.[6] See also my program *DDSWIN* at the TAPR download site on the WWW. Unfortunately, the AD7008 has no quadrature output. So, I looked for another way to get more than one octave of phase-split signals from my LO.

The simple RC phase-shift network used by KK7B and others is well known by experimenters. Rick suggested that I use a hard-limiting amplifier after this simple network in order to overcome the drawbacks of having amplitude balance at only one frequency. I picked up this nice idea, and constructed my broadband phase shifter, called BPS1.[7] This unit has become part of my DDS-tuned DCTRX during the past two years. For the next step, I decided to add DSP to my rig. Analog Devices came out with their EZ-KIT Lite, at a ham radio price. I wanted to expand my *DDSWIN* platform to have DSP-program download capability. Since I use the PC's parallel port to tune my DDS, it seemed easiest to use a serial port for simultaneous DSP functions. Shortly after I began to create a program for the serial port, I realized that I had to get back to basics; I had much to learn about my new ADSP 2181 kit, and *Windows* port programming, too!

After more than one year of "software soldering," I was at the break-even point: My new version of *DDSWIN* was capable of downloading DSP programs. I could even see how Johan Forrer's (KC7WW) audio signal processor[8] could work with my rig. The results were looking quite good; they encouraged me to do more experiments with IF-DSP later. Now I was free to do some other experiments that had been on my back burner, such as improving the large-signal performance of my rig by using doubly balanced *active* mixers.

Introducing the AD831

Analog Devices' AD831 low-distortion mixer[9] is a very interesting device. Its +24 dBm third-order intercept point (IP3) is better than most amateur gear on the market today. Its built-in limiting amplifier for the LO allows oscillator injection levels as low as –10 dBm. The limiting amplifier simplifies the phase-shifting technique for the LO as well.

When constructing the BPS1, I used high-speed CMOS gates at the output of my DDS, which is no longer necessary with the AD831. The low-pass-filtered sine-wave output of my DDS1 has a level that is ideal for this application. Feeding two mixers via a simple Fisher hybrid (not the cascaded one) is, from my experience, the best way to get good amplitude and phase balance over a wide range of frequencies. The

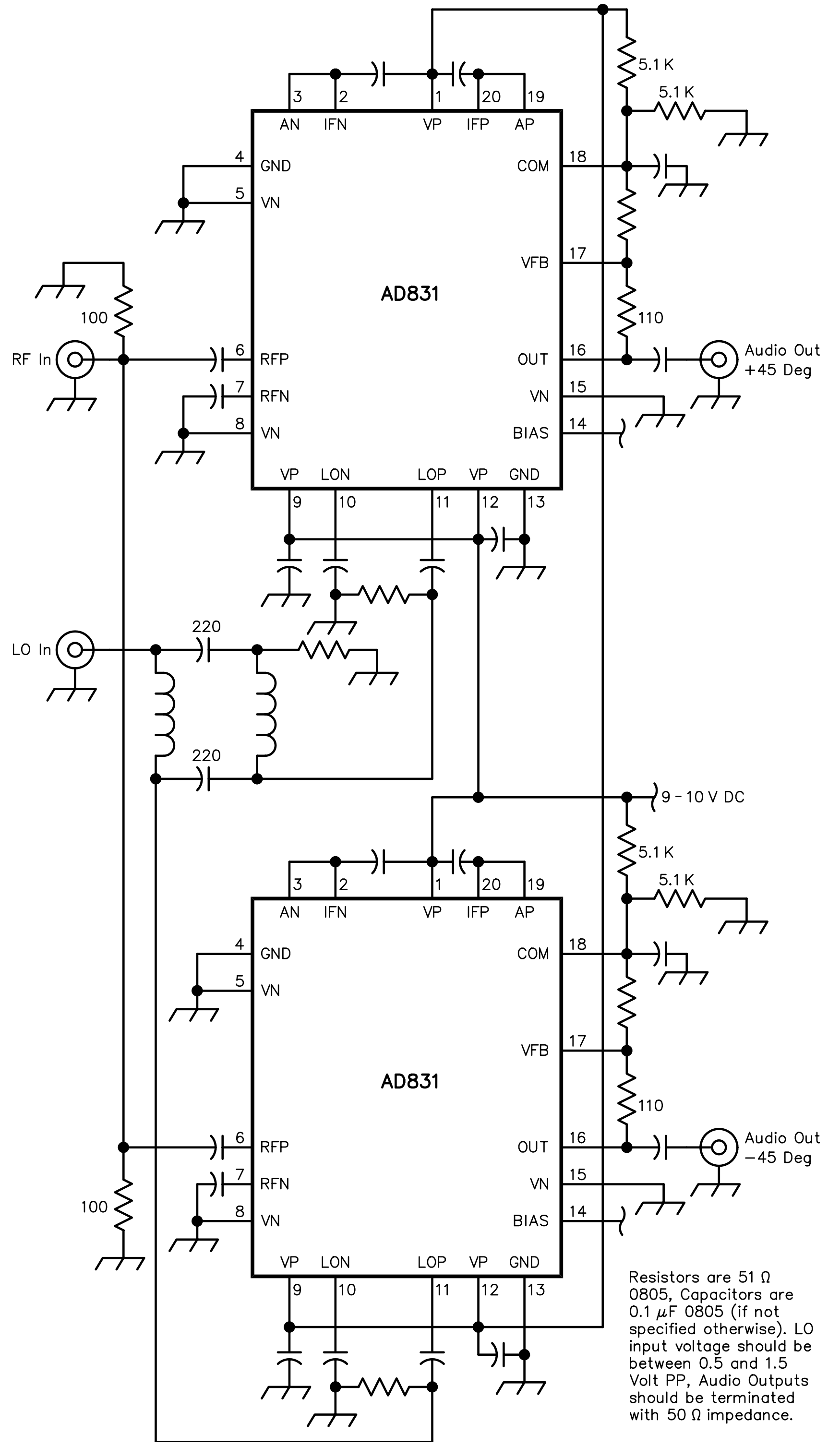

Fig 1—High-level active mixer schematic.

above-mentioned RC network will do the job too, but I found the limiting amplifier in the AD831 less effective against amplitude variations than my BPS1 limiter. The mixer's integrated, low-noise output amplifier allows a setup with no mixer insertion loss.

This amplifier can drive low-impedance loads, as is known from the R2 concept by KK7B. The use of two mixers, parallel connected at the inputs and without an extra splitting device, simplifies the construction and increases the large-signal capability. There is a drawback that should not be overlooked: The quiescent current of the device must be >100 mA for best IMD performance. This is compensated somewhat by the power saved at the LO ports. Because my supply voltage is normally 13.5 V, I decided to use the mixers with a single supply, which is possible with +9 to +10 V. When looking at the data sheet of the AD831 (Rev B, page 11), the seasoned experimenter will see some mistakes, which I have corrected in the circuit diagram (Fig 1). For my application, I have set the output RC low-pass filter's cutoff at approximately 100 kHz by using 100 nF feedback capacitors.

The Experimental Setup

When I have an idea, I prefer the quickest way to its realization, and seldom find time to make a fine PC layout. So it was with this little project of testing the AD831 in an I-Q mixer setup. I found a surplus 1×2 inch enclosure at my work, and a piece of experimenter's board. It has ground on one side and copper eyes on the other, with a 0.1-inch grid. This is fine for soldering SMD capacitors and resistors of size 0805 or 0603 in between the eyes. See Fig 2 for a close-up of the PC board's bottom side.

With the idea in mind of using the mixers in a streamlined PCB layout later, I decided to use PLCC 20 sockets so I could easily remove the chips. While this is surely not critical at HF, I have not tried it at VHF or UHF, where operation of these mixers is also possible. Since the mixers were getting warm, I mounted the PC board in the enclosure so that the brass cover was in good thermal contact with the top of the chips. Fig 3 shows a top view of the experiment.

When constructing the Fisher hybrid, I did not give much attention to the exact inductances of the twisted-pair transformer windings. With a ferrite core from my junk box, I later measured 3.2 µH for one of the twisted-wire strands wound on the core. The capacitors were off-the-shelf SMD 0805 220 pF units. During the phase measurements, I found these capacitors relatively noncritical. They had little effect on the achieved 90° phase difference. This was *not* the case when using the RC network. Phase tweaking could be done by adding a small trimmer to ground on one of the LO inputs.

Measurements and Results

It's difficult to make precise phase measurements without sophisticated, high-priced equipment. One must account for several possible measurement errors, which may occur because of the various phase delays in the measurement setup. For coarse alignment, an oscilloscope may be used, with two channels in X-Y mode. A phase balance of 90° and good amplitude balance produces a perfect circle on the screen. (See Fig 4.) Since we need phase accuracy to within 1°, the best fine-alignment method is a test with a real receiver, where one can switch between sidebands while injecting a strong carrier into the RF input. During tweaking, the power level of this carrier should be high enough to hear the suppressed opposite sideband when it is down 40 dB or more.

Because we are interested only in the mixer's output signals, measurement of fine phase errors at the LO inputs is not worthwhile. The results would depend strongly on the quality of measurement equipment available—higher frequencies = higher price!

I did all my measurements with a relatively low-priced oscilloscope and my Multitone Test Generator MTG 1 (described in *QEX* in 1994).[10] I also used this generator to measure the IP3 performance. Its RF output consists of two –7 dBm tones, 20 kHz apart, in the 80-meter band (3560, 3580 kHz). I measured the level of the third-order intermodulation products at frequencies of 3540 and 3600 kHz. The distortion product was measured at the speaker output of my R2-like receiver, with no AGC and at an audio level of about 100 mV RMS. I applied the generator signals to the mixer's input without attenuation and found that an attenuation of 70 dB was necessary in order to achieve the same 100 mV output when observing one of the two injected frequencies. That means a calculated IP3 near +30 dBm, which is a very good value! One would seldom find amateur equipment of this performance level.

As always with D-C receivers, one may see unexpected outputs when very strong signals are applied. Excess phase noise on the LO injection is often responsible. From my DDS1, I found the two –7 dBm tones to represent the threshold of objectionable hum and noise caused by reciprocal mixing.

Switching between several measurement frequencies is easy with the

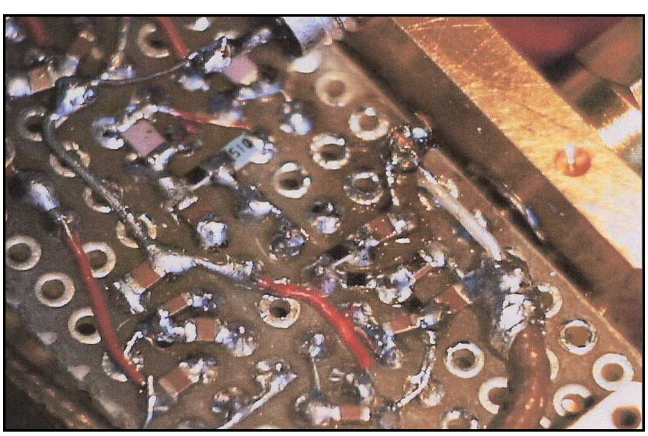

Fig 2—A view of the breadboard underside shows chip components installed.

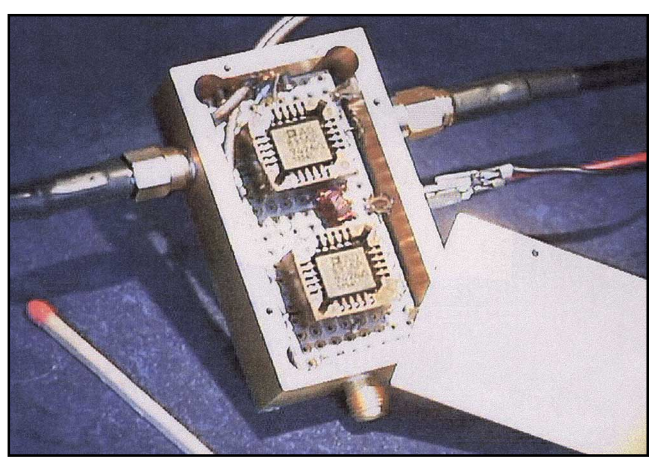

Fig 3—A top view of the breadboard in its enclosure.

DDSWIN platform, since it allows memorization of frequencies and descriptive text as well. What's the frequency range of the unit without phase readjustment? It has been proven that coverage of at least four HF amateur bands is possible. I tried it on 160, 80, 40, 30 and 20 meters with good suppression of the unwanted sideband.

Conclusion

The set up of a low-distortion I-Q mixer has been demonstrated. Its outstanding performance makes it a useful device for both D-C and IF-DSP receivers. It allows simplified construction of high-quality modern receivers with little alignment and low cost. I have not tested the mixer as a modulator, but this may certainly be tried also. The above-mentioned PC program *DDSWIN* is available. Contact me for it by e-mail.

Acknowledgments

Thanks to Rick, KK7B, and Johan, KC7WW, for their fruitful contributions and discussions during the design of my latest projects. I am also very grateful to Dwight K. Elvey of Santa Cruz, California, who helped me so much to understand my DSP better. He is a very highly skilled programmer and a patient advisor, too—I proved it by exchanging countless e-mails with him! Thanks finally to my wife Erika, DJ8AA, for missing me while I spent many hours in my radio shack.

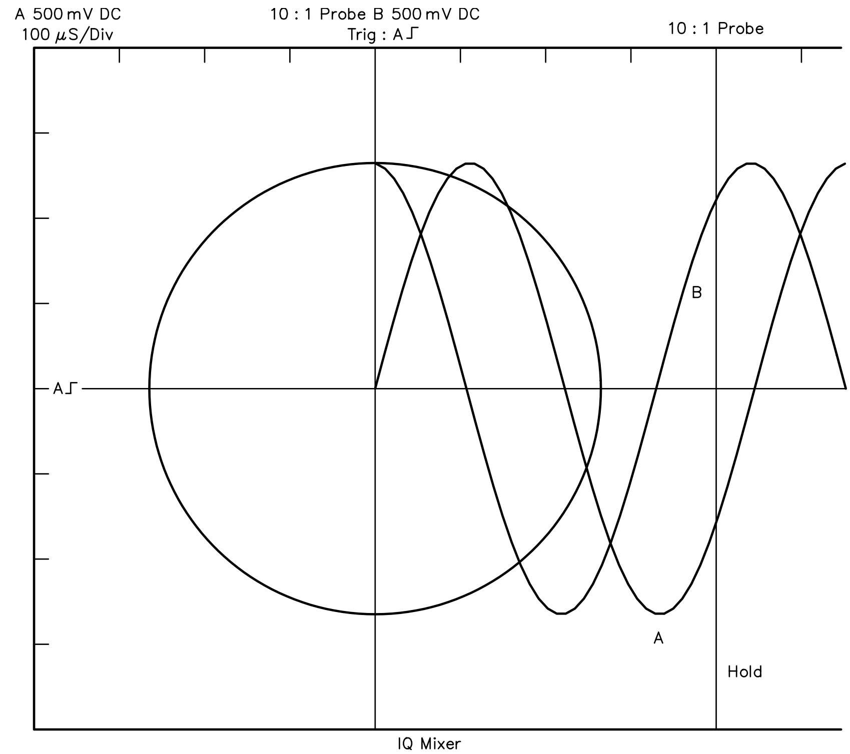

Fig 4—X-Y oscilloscope display of quadrature-injection signals.

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Detlef is 57 years old and has been a licensed amateur since 1959, with the call sign DL7IY. His present job is developing optical to millimeter-wave (60 GHz, V-Band) converters for mobile communication purposes at Heinrich Hertz Institute, Berlin, where he has been employed as a technical staff member since 1986. He holds a degree in electrical engineering from the Technische Fachhochschule Berlin (1983) and a master degree of radio and television servicing (1968). He has also been active in the education of electronic professionals and nonprofessionals for many years.

Polarization Modulation: A Technique Worth Investigating

Come for a walk on the wild side with KL7AJ, who proposes we impress information on a signal by altering antenna polarity. Can a frequency simultaneously carry multiple signals of differing polarity?

By Eric P. Nichols, KL7AJ

Only a few characteristics of a radio wave can be manipulated in order to convey intelligence. When we speak of modulating a carrier, we generally think in terms of amplitude or frequency modulation and their various flavors, such as single-sideband and frequency-shift keying. Phase modulation can be interchanged —for all practical purposes—with frequency modulation, so we can discount that as a separate technique.

There are other properties of a radio signal, however, that can be intentionally modulated with information. One of these is the polarization of the wave. When we do (infrequently) think of the polarization of a radio signal, it is only to match our receiving and transmitting antennas' polarization for maximum sensitivity. We don't look for any real information in the polarization because, most of the time, there isn't any.

Polarization modulation falls into that odd category of modulations known as *spatial modulation*. Spatially modulated signals are unusual in that the information is transferred by virtue of some physical orientation of the receiving antenna relative to the transmitting antenna

For clarity, let's examine a familiar form of spatial modulation, Doppler shift. Suppose that a perfectly stable oscillator—one with no inherent modulation of any kind—is on a satellite. We know—from theory and experience—that if we measure the oscillator frequency from an earthbound receiver, we will detect a shift in frequency over time. The frequency is higher as the satellite (oscillator) moves toward us, and lower as it passes away from us. However, if we were to ride in another satellite at exactly the same velocity and direction as the first, we detect no frequency shift at all.

In other words, the frequency modulation is caused only by the motion of the transmitter relative to the receiver. We could generate frequency modulation—in concept—by putting a rock-solid oscillator on a railroad car, and moving the car back and forth on the tracks at an audio frequency. This is not a practical means of generating FM! Of course, if you were to ride along with your receiver on another railroad car coupled to the first one, you would not be able to detect any FM, although you'd probably develop a bad case of motion sickness. (Since the satellite or train is following a circular path, some modulation would be detectable because the two stations are, in effect, rotating around a common center point, and because they are a separated by a finite distance. See also my "More Spatial Diversity?" sidebar—*Ed.*)

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Now let's apply this same principle to the concept of polarization. Imagine, for a moment, a dipole antenna mounted on a shaft of an electric motor, spinning at 60,000 RPM, or 1000 Hz, a nice audio frequency. Let's feed some *unmodulated* RF into the antenna (through a pair of slip rings to avoid tangling the feed line). Now, at some distant location, let's build a similar spinning antenna—also with slip rings, so as not to wind ourselves around the thing—and run the feed line to the antenna terminals of a receiver. Then let's orient our motor shafts so they're on the same axis. Now, if we were to spin the antennas at the same speed and direction (and with a phase shift proportional to the propagation time—*Ed.*), we would detect only a steady carrier at the receiver. If we were to stop the rotation of either the receive or transmit antenna, however, we would find our signal at the receiver to be amplitude modulated at 1000 hertz. It would be, very nearly, a perfect sine wave, since the attenuation of a pair of dipoles is very nearly a sinusoidal function of their relative polarizations. This is so as long as they're on the same axis. Off-axis polarization is an entirely different—and important—matter that we'll discuss in detail later. Therefore, it is evident that, in a rather impractical manner, we have generated an AM signal in one frame of reference that is undetectable by another receiver in our rotating frame of reference. This may generate some intriguing possibilities for high-security communications.

Now the security aspects are not relevant for radio amateurs, but spectrum conservation is, and amateurs can do many experiments using various aspects of polarization modulation toward this end. Fortunately for us radio amateurs—and for sane people as well—we need not rely on motors and slip rings to generate polarization modulation, for the polarization of a signal can be shifted electronically.

The Mathematics of Polarization

It's good to have a solid foundation in the mathematical principles of polarization. Not that we can't make some fascinating discoveries by the seat of our pants with a hot soldering iron, but we can avoid reinventing the wheel by trying things that are easily proven impossible. For those readers familiar with antenna modeling using *NEC*, *MININEC* or the like, none of the following should be a surprise. Nonetheless, a good review of the basics is in order.

For the sake of simplicity, the remaining discussion will only refer to the E-plane characteristics of a radio

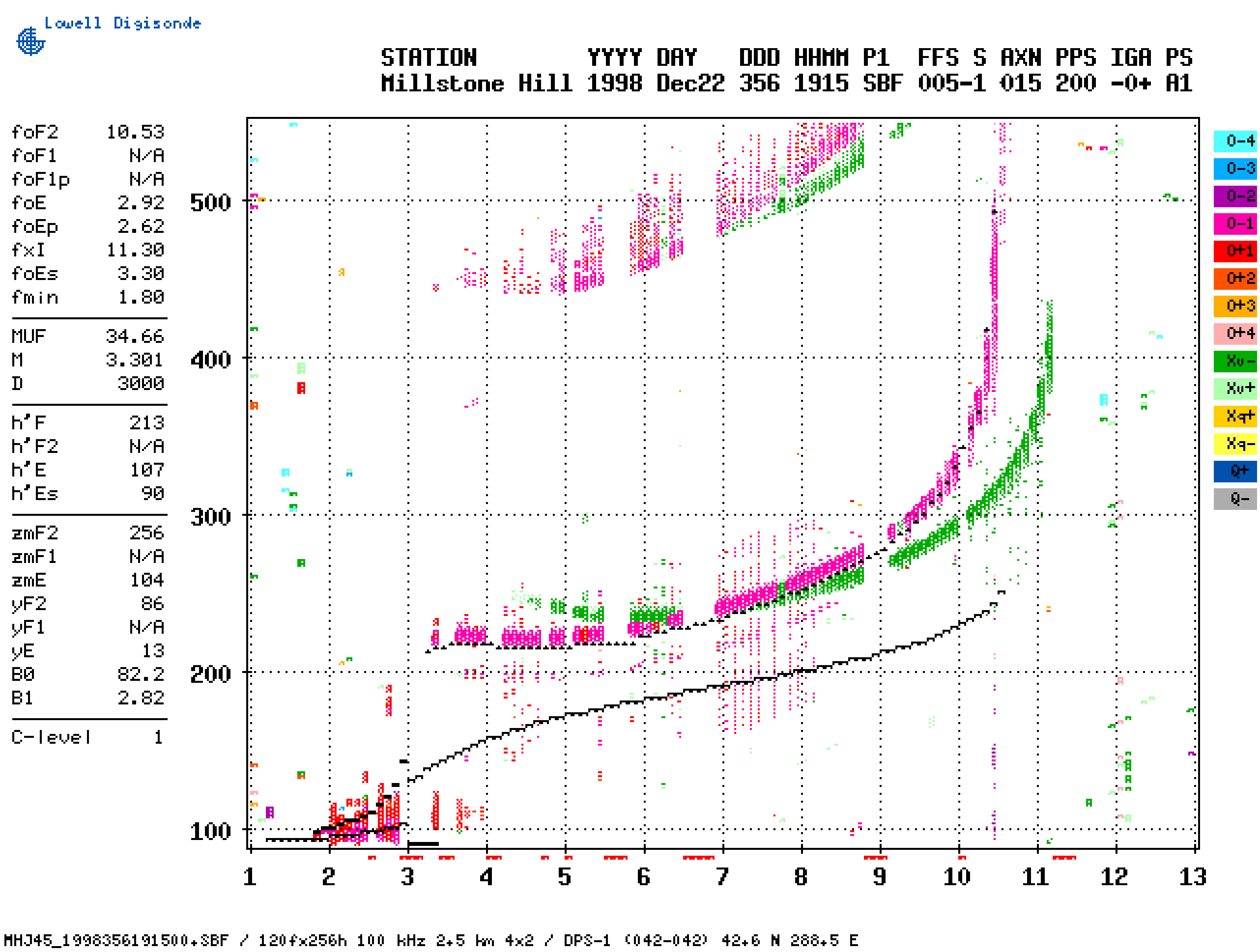

Fig 1—A Digisonde ionogram downloaded from **http://digisonde.haystack.edu/scripts/latest.exe?** about 1915 UTC on Dec 22, 1998. The green areas of the original color image have been circled to distinguish them from the red areas in this B&W image. The existence of both red and green components indicate the presence of both right- and left-hand-circular-polarization signal components.

wave. In addition, for the sake of this discussion, the E-plane (electric-field plane) will always have the same orientation as the physical antenna elements, which certainly makes things easier to visualize. For all practical purposes, we can apply this to the H-plane, or magnetic-field plane, just by rotating it 90°. I know this is not rigorously true, because of the nuances of electromagnetic field theory, but you're not too likely to encounter the nuances in the ham shack!

Let's return to our simple pair of receiving dipoles. Fortunately, the gain of an antenna can be expressed independently of its polarization properties. A dipole is as good as a "death ray" when it comes to polarization analysis and modulation techniques.

In theory, if we cross-polarize transmitting and receiving dipoles (again, in the coaxial case), we should be able to achieve a perfect null. In other words, there should be an orientation where we can expect no transfer of transmitter power to the receiver's antenna terminals. We have to assume the antennas are far enough apart to be in each other's "far field," that is, where there is almost no mutual inductance. Assuming that the dipoles are infin-itely thin and perfectly conducting (that would be a good trick!), as we rotate the antennas toward identical polarization, the transfer of power should increase as the sine of the rotation angle. This on-axis rotation gives us the most dramatic change of power transfer relative to polarization angle. In other words, the greatest polarization sensitivity occurs for on-axis rotation. This sinusoidal function is the best that you can achieve. Contrary to some theories floating around, *absolutely nothing* can be done to make any antenna more sensitive to polarization rotation than a sinusoidal function.

However, a lot can be done to *decrease* the polarization sensitivity of a transmit-receive antenna pair. In fact, as we move off-axis we find that the polarization sensitivity continually decreases, and is anything but sinusoidal as we perform "broadside" polarization rotation. To clarify what I mean by broadside polarization rotation, imagine two vertically polarized Yagi's. Not very high-gain jobs, though; we want to be able to receive a fair amount of radiation from the sides. Yagi number 1, the transmitting antenna, is aiming north. Yagi number 2 is located to the west of Yagi 1, and aiming east toward the side of Yagi 1. Now, let's rotate the polarization of Yagi 1 towards horizontal, but still aiming north. We will see very little change in the received signal strength in Yagi 2, until Yagi 1 is very nearly horizontal, at which point the signal strength will take a nosedive. Again, this off-axis pattern can be demonstrated in *MININEC* quite handsomely. It's kind of a bloated figure-eight affair, with some very sharp nulls near 90°.

Now, as long as we avoid these nulls and their immediate vicinity, we find that we have nearly no polarization sensitivity off the sides. Let's assume that, instead of just spinning our transmitting antenna merrily around with our 60,000 RPM motors, we just vary it a few degrees. Again, this is much easier done electrically than mechanically. We will have the greatest polarization shift with Yagi 1 on-axis, since we're near the peak of our sine-wave function. Broadside, however, with Yagi 1 aimed north again, we have almost no sensitivity to polarization rotation. What it looks like from the receiver's standpoint is that "down the barrel" of the antenna, our pol-mod (polarization modulation) signal has more modulation than it does off the sides. Can you see how this can help suppress interference off our sidelobes? They will have less modulation, and hence less bandwidth, than the on-axis signal. You can't demodulate much of anything coming off the sides. You just get carrier! I see some interesting possibilities already. You don't need a high-gain antenna to observe this property; I used the Yagi's for demonstration purposes only. A dipole will show identical spatial properties.

Practical Hardware

"That's all fine on paper," you say. "But how do I actually *do* anything with this?" Good question. So, how *do* we achieve polarization modulation in the real world of ham radio?

The simplest means of achieving polarization modulation—which includes everything from a few degrees to full-blown mark-space cross polarization—is to use a pair of crossed dipoles, fed with different values of power. As long as there is no phase shift between the vertical and horizontal components, the result is linear plane polarization. (I'll share a few words on circular and elliptical polarization later.) Equal power fed into the vertical and hori-zontal elements of a crossed dipole will result in 45°-angle polarization. It will be received equally poorly on either vertical or horizontal antennas. For low-power operation, voltage-controlled attenuators can be inserted in the feed lines and driven with audio signals (in the proper phases) to achieve small degrees of polarization modulation.

For RTTY, you can use PIN diodes to switch mark signals to the horizontal element and spaces to the vertical element. Of course, the frequency and power level you're using will dictate the actual hardware. These are just suggestions. At microwave frequencies—where there is more room for experimentation—you have a lot more latitude in the techniques. At 10 GHz, you might physically rotate a Gunnplexer at audio frequencies with a good (large) stepper motor. (No, I haven't actually tried this.) I've yet to fully investigate the effects of centrifugal force on radio signals.

Running Circles Around Interference

So far, we've dealt mainly with linear, or plane, polarization. Circular polarization can be adapted to pol-mod techniques as well. We might achieve extra elbow room on the ham bands through frequency re-use, by means of polarization diversity. This technique is common in geostationary satellites, where we have V and H channels.

Isolation between signals at lower frequencies can be improved by means of right- and left-hand circular polarization. Because of conditions over most terrestrial paths, achieving perfect isolation between V-pol and H-pol signals is nearly impossible using only crossed dipoles. Right- and left-hand circular polarization, however, cancels most of the odd, random reflections that upset the polarization discrimination with plane antennas.

Along with pure right- and left-hand circular, however, there are all degrees of freedom in between the two, in the form of elliptical polarization. We can take crossed dipoles, as mentioned above, and generate a 45°-plane field with equal power, in-phase RF signals. If we shift the relative phasing (at an audio frequency) instead of shifting power ratios, we can achieve narrow-band modulation that oscillates between slightly right-hand elliptical, or slightly left-hand elliptical polarization. The resulting signals can be easily separated using either right- or left-hand-circular receive antennas, but guess what: You can't receive any of the modulation with a plane-polarized receive antenna! [In the correct plane—*Ed.*] As you can see, the possibilities are manifold. Not only can we re-use a frequency using

More Spatial Diversity?

While polarization modulation might seem a bit off the beaten path, it is presented here in the spirit of stimulating discussion about new and different things. Readers considering the use of techniques described may be wondering what emission designator to use. I believe such "pol-mod" emissions must be considered AM, when the signal fed to the antenna is unmodulated carrier. This is because regardless of the transmitting antenna(s) physical arrangement, the transmitted fields are amplitude modulated with respect to any frame of reference not changing with the modulation. The occupied bandwidth of the actual fields is commensurate with AM. A special case occurs when the polarization is rotating at the carrierfrequency, which is simply circular polarization. Occupied bandwidth is not increased in this case.

As for Doppler shift, we don't normally regard the emission as being frequency-modulated just because the source is in motion. A satellite's motion is not intended to communicate information any more than the aging of a transmitter's crystal frequency reference. What emission designator describes a particular signal depends as much on the intent of the operator as on the means of bandwidth generation. A very "chirpy" or "clicky" CW signal, however, might represent a case wherein the intent is overwhelmed by the interference caused.

Eric's train, with its exhausted crew, creates an FM signal with deviation limited by the velocity at which the train can move. I'd like to propose another of these "thought experiments."

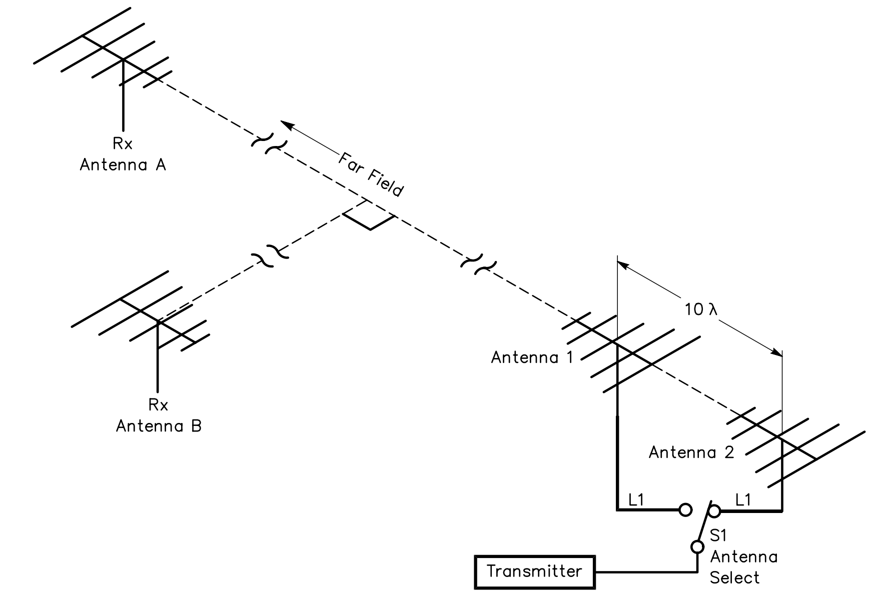

Fig A—Another example of spatial diversity? Decide for yourself.

Suppose we place two transmitting antennas on a line in the direction of intended propagation, separated by 10 λ (see Fig A). The antennas are connected by equal lengths of transmission line to a single-pole, double-throw switch at the midpoint. The switch's common is fed by a transmitter, which supplies only an unmodulated carrier. We toggle the switch at a rate equal to 0.1 of the carrier frequency.

Were the switching system perfect, distant receiver "A," on the transmitting antennas' axis, would receive an on-off (CW) signal with almost twice the field strength of that produced by a single antenna. This is because on axis, the field from one antenna arrives at the other's location in time to reinforce its field. Distant receiver "B," on a line orthogonal to the transmitting antennas' axis, would sense almost no change in field strength. A beam is formed, and the occupied bandwidth is dependent on the angle made with the receiver. This is, I think, another example of spatial diversity. What do *you* think?—*Editor*

horizontal and vertical plane polarization, we can re-use the same frequencies with right-hand *and* left-hand circular polarization. Therefore, we can get four times the bang per buck just by the judicious use of polarization.

Gimme an X, Gimme an O: What's that spell? Radio

I'd like to offer a few comments about the use of circular polarization on the HF bands. Yes, you read me right. Because of what I consider a conspiracy of ignorance, there has been nothing in the ham literature about the advantages of circular polarization for HF-skywave propagation.

Whenever you launch a plane-polarized wave into the ionosphere—where there is a magnetic field—the wave splits into two circularly polarized modes: the ordinary, "O," mode and the extraordinary, "X," mode. These two counter-rotating waves take entirely different paths through the ionosphere. This is not saturnine metaphysics. Every Ionosonde since the 1930s has shown this phenomenon quite clearly. If you look at any of the dozens of Digisonde ionograms available on the Web, notice the red and the green plots. The red is the O mode, and the green is the X mode. It's quite revealing. Fig 1 is an example. The point is that only one of these waves, in all likelihood, does you much good, since the two propagation modes come back to earth at completely diverse locations, especially in high-latitude regions. Why not take advantage of the different modes by intentionally launching either an O-mode signal (clockwise C-pol, in the northern hemisphere), or and X-mode signal (counterclockwise C-pol in the northern hemisphere)? You'll save half your power and reduce QRM at the same time.

Conclusion

As the title of this article suggests, polarization modulation is not the answer to everything. It's only another technique we should investigate to fulfill our collective mission to advance the state of the radio art. Our under populated microwave bands are an ideal place to experiment with this, but certainly not the only place. Frequency re-use schemes on the lower bands can go a long way toward making life better for those of us in the trenches. Attention to the polarization of our signals is one simple way to do this. □□

A Dentron MLA2500 HF to Six-Meter Conversion

Another MLA2500 with dead 8875s finds salvation— and is reborn for the magic band at 50 MHz.

By Jim Worsham, W4KXY

Editor's Note: This article also appears in The Proceedings of the 1998 Southeastern VHF Society Conference.

There are a limited number of six-meter amplifiers commercially available on the market today, and they tend to be rather expensive when purchased new. As interest in this band increases, one source of economical power on six meters—and I believe on other VHF bands as well—is older, unused HF amplifiers. This paper describes the conversion of a Dentron MLA2500 HF amplifier to six meters. It is not the one and only way to do such a conversion, and is not a construction article, per se. It is only meant to describe the one way to do such a conversion that satisfied my unique requirements. I hope that this paper will help stimulate other amateurs to start thinking about how they too might give that old HF amplifier in the closet a new life on the VHF bands.

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Background

After several months of operating on six meters with 100 W from a brick amplifier, I decided that the time had come for more power on the band. When six meters is open, you can work just about anybody with 10 W. When it is not open, it takes more power to get people's attention and hold a frequency during a contest.

I started researching amplifiers for six meters by studying published designs, looking at commercially available amplifiers and talking to local amplifier experts. I quickly concluded that high cost made a new commercial amplifier out of the question. There was a chance that I might find one used, but it has been my experience that those who have such amplifiers keep them. Another option was to build an amplifier for six meters completely from scratch, using one of several published designs. This was initially attractive, but I was concerned about the metal fabrication involved and the difficulty in finding many of the parts for such a project. It seemed a bit much for someone who had never even owned an amplifier before. In December 1996, I came across an old Dentron MLA2500 HF amplifier for sale at a good price and started thinking about converting it to six meters.

I checked out the MLA2500 and found it to be in good condition with one exception, the 8875 tubes. The amplifier clearly had been run hard: One of the tubes had dark oxidation on its anode, indicating that it had been run much hotter than it should have been at sometime in the past. Replacing the 8875s was unacceptable because of their high cost ($800 for a new pair).

A couple of articles have been written about replacing the 8875s in an MLA2500 with other tubes, such as the Svetlana 4CX800A or 4CX400A.[1, 2] I studied those articles with great interest, but was discouraged by the major circuit modifications required to supply the proper filament, grid and screen voltages for these tetrodes. I felt that replacing the 8875s with another triode was the best way to go for me, since it would be possible to use the existing power supply and metering circuits in the MLA2500 with few, if any, changes.

After some discussions with Dick Hanson, K5AND, I decided that the 3CX800A7 was the best choice. There were several factors driving this decision. Surplus 3CX800A7s or pulls were available at reasonable prices. The 3CX800A7 requires relatively little drive power in a grounded-grid configuration, which meant I could drive it directly with the 20 W from my transverter and expect to get 750 to 800 W out. Several published designs use the 3CX800A7 at VHF and even UHF.[3, 4, 5] Finally, Dick agreed to help me with the project, and he has quite a bit of experience building VHF amplifiers using the 3CX800A7.

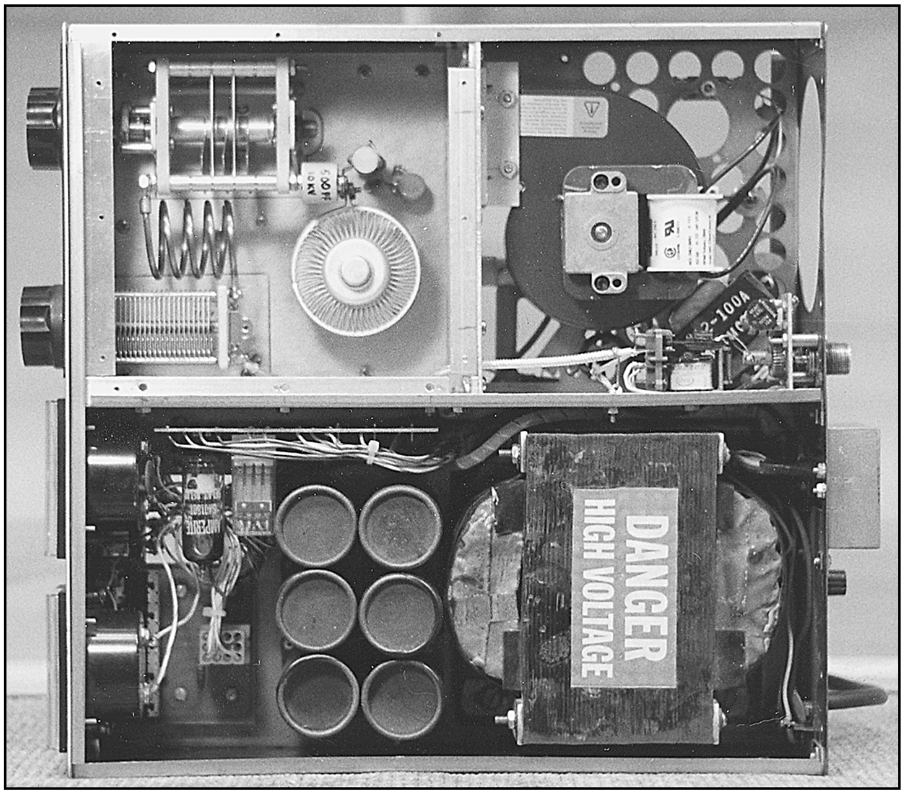

Fig 1—A top view of the MLA2500 chassis. The lower half houses the power supply, metering and control circuitry. The upper half is the RF deck. The TR relay, RF input connector, RF output connector, relay-control connector, wattmeter PC board and associated wiring can all be seen in the lower right-hand corner of the RF deck.

The Conversion

The top of the MLA2500 chassis is divided into two halves, as shown in Fig 1. The lower half houses the power supply, metering and control circuitry. This half is mostly left untouched. The upper half is the RF deck. The first step in the conversion process is to completely strip the RF deck, with the exception of the TR relay, RF input connector, RF output connector, relay-control connector, wattmeter PC board and associated wiring. These parts can all be seen in the lower right-hand corner of the RF deck.

The RF deck is divided into two halves by an aluminum plate mounted with aluminum angle stock. Another aluminum plate is mounted to the MLA2500 chassis floor in the front half of the RF deck. The combination of these two plates forms a somewhat airtight compartment for the 3CX800A7 and output-tank circuitry in the front half of the RF deck. This compartment is pressurized by a blower mounted in the rear half of the RF deck. The 8875 filament requires 6.3 V, while the 3CX800A7 requires 13.5 V. As a result,

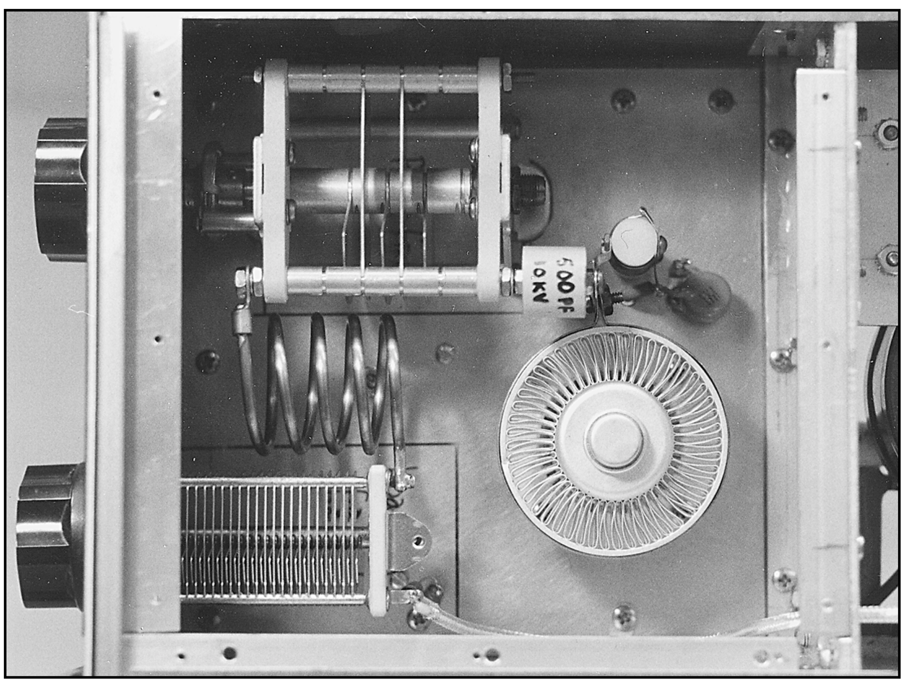

Fig 2—A close-up view of the 3CX800A7 and output-tank circuitry. The 3CX800A7 is in the lower right-hand corner. Plate choke RFC2 is just above it. The output tank is a low-pass pi circuit. The plate TUNE capacitor is in the upper-left. The plate LOAD capacitor is in the lower-left. Inductor L4 is between them. RF output is routed from the LOAD capacitor through a short piece of Teflon coax along the bottom and out of the compartment at the lower right, to the TR relay.

[1]Notes appear on page 41.

a filament transformer has been added in the rear half of the RF deck, next to the blower and TR relay. Both the filament transformer and blower require 120 V ac, which is obtained from one half of the plate transformer's primary.

Fig 2 is a close-up view of the 3CX800A7 and output-tank circuitry. You can see the 3CX800A7 in the lower right-hand corner of the compartment. Just above it is the plate choke RFC2, as well as bypass and output-coupling capacitors. The plate choke consists of 40 turns of #22 enameled wire close wound on a $^1/_2$×2-inch-long Teflon rod. The connection to the anode of the 3CX800A7 is made with a homemade $^1/_4$-inch-wide copper strap wrapped around the anode and held in place with a single bolt and nut. In the upper-left corner of the compartment is the plate **TUNE** variable capacitor. In the lower-left corner of the compartment is the plate **LOAD** variable capacitor. Between these two variable capacitors is inductor L4 of the pi-configuration output-tank circuit. This inductor consists of 4$^1/_2$ turns of #6 copper wire, 1$^1/_4$ inches ID and 1$^3/_4$ inches long. The output from the tank circuit is routed through a short piece of Teflon coaxial cable from the plate-**LOAD** variable capacitor, along the bottom, out of the compartment and finally to the TR relay.

Fig 3 shows the top cover for the front half of the RF deck. A chimney made of high-temperature silicon rubber is held in place with a Teflon collar and fits over the 3CX800A7's anode cooler. This assembly directs air through the cooler and out the top of the compartment. A screen mesh—which is not visible in the photograph—covers the air outlet and provides RF shielding.

Fig 4 shows the bottom of the MLA2500 chassis. As was the case above the chassis, the lower half of the under-chassis space contains mostly power-supply circuitry, and most of that is left untouched. The upper half of the under-chassis space contains the cathode and filament circuitry.

Fig 5 is a close-up view of the upper half of the chassis bottom. To the left of the 3CX800A7's socket is the filament choke RFC3, which is wrapped in fiberglass tape to prevent shorting. The filament choke consists of 16 bifilar turns of #16 enameled wire close wound on a $^1/_2$×2-inches-long phenolic rod. The choke is connected through two series-connected resistors (0.9 Ω total) to the filament transformer. These resistors set the filament voltage measured at the 3CX800A7's

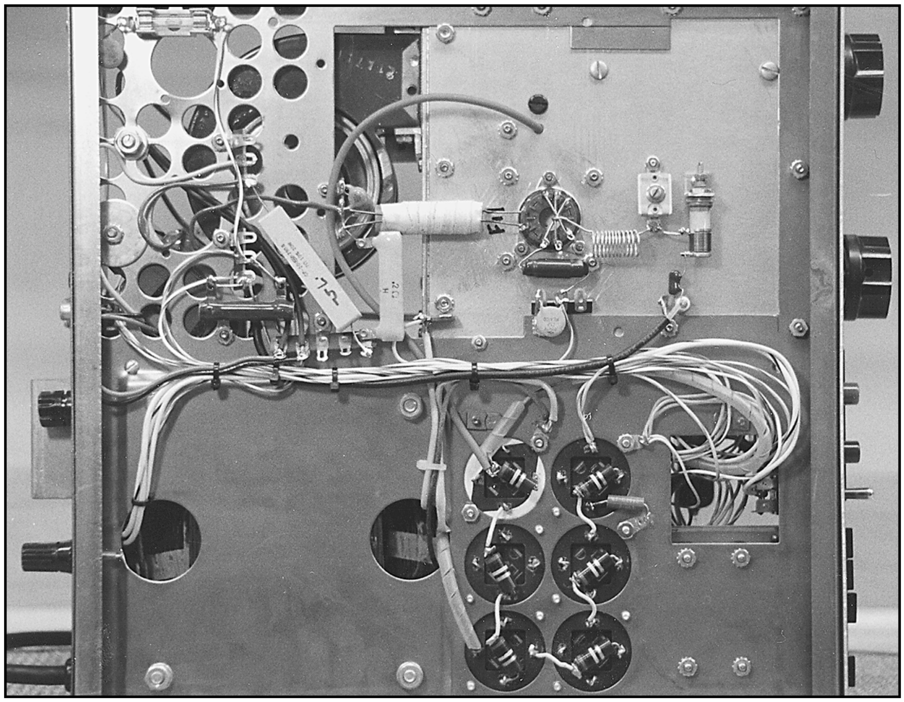

Fig 4—The bottom side of the MLA2500 chassis. The lower compartment contains mostly power-supply circuitry. The upper compartment contains the cathode and filament circuitry.

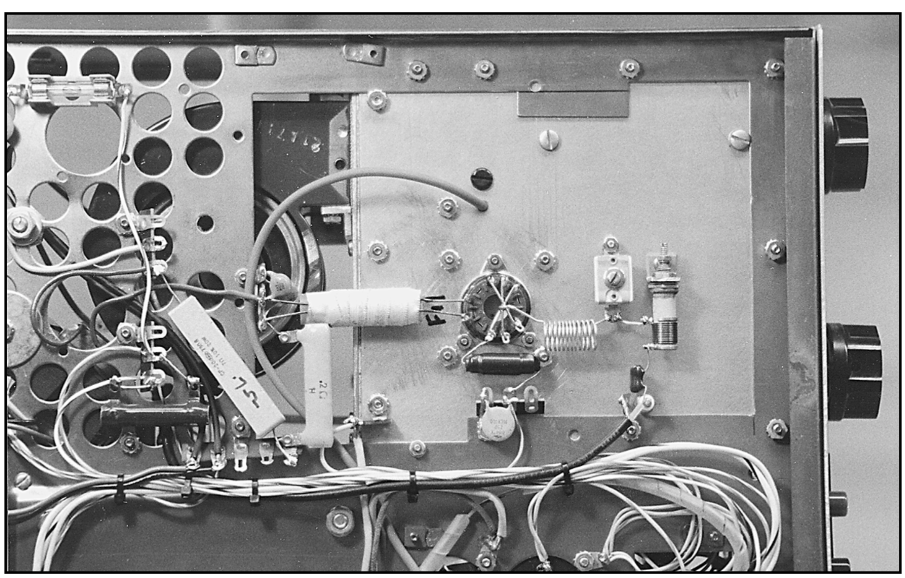

Fig 5—A close-up view, under the chassis, cathode and filament circuits. To the left of the tube socket is the filament choke RFC3, which connects through two large resistors to the filament transformer.

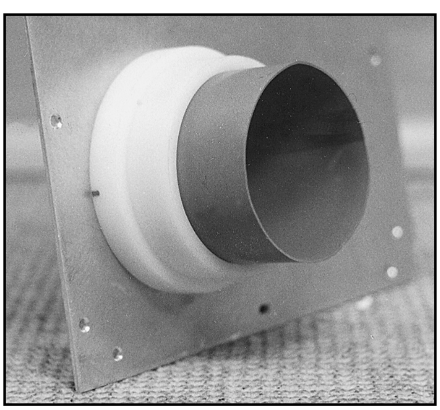

Fig 3—The modified top cover for the front half of the RF deck with the rubber chimney and Teflon collar in place.

Fig 6—The front panel of the converted MLA2500. The left side has been covered with a thin, black aluminum plate. Note the new power switch between the TRANSMIT light and the STAND BY switch.

socket as close as reasonably possible to 13.5 V. This also has the added benefit of limiting the 3CX800A7's filament-inrush current. Setting the filament voltage as close as possible to its specified value and limiting the filament inrush current helps to insure a long life for the 3CX800A7.

Below the socket is an RF choke connected to the cathode. (It's labeled 7 µH, Z-50.) The choke is in series with a 20-kΩ resistor—located just below the 0.7-Ω filament resistor—a 1.5-A fuse and a 6.8-V Zener diode, located in the upper-left corner of the figure. The 20-kΩ resistor has a set of contacts from the control relay across it. When the amplifier is unkeyed, the contacts are open, letting the 20-kΩ resistor bias the 3CX800A7 off. When the amplifier is keyed, the contacts close, allowing the Zener diode to bias the 3CX800A7 on.

To the right of the socket is the cathode input matching circuit. This circuit consists of two inductors and a capacitor in a **T** configuration. The capacitor and one of the inductors are variable so they can be adjusted for minimum input SWR. Fixed inductor L5 is 12 turns of #16 wire, 3/8 inch ID, 7/8 inch long. Variable inductor L6 is 12 turns of #18 enameled wire close wound on a 3/8-inch ceramic form with a red slug. RF from the transmitter is routed from the TR relay through a short piece of RG-58 coax to the input matching circuit.

Fig 6 shows the front panel of the converted MLA2500. The left side of the front panel has been covered with a thin aluminum plate—painted black—to cover all remaining holes in the front panel. The power switch was moved from its old location in the lower-left corner to a location between the transmit light and the standby switch.

One additional part of this conversion project that cannot be seen in the figures involves the time-delay relay. The original time-delay relay had a 75-second delay (Amperite 6NO75T). This delay is too short for the 3CX800A7 to warm up properly. The original relay was replaced with one having a 180-second delay (Amperite 6NO180T).

Fig 7 is a schematic diagram of the MLA2500 after the conversion to six meters has been completed. Fig 8 is a copy of the original MLA2500 schematic for comparison.

Initial Tune-Up and Results

The amplifier was initially tuned up by installing a directional wattmeter between a six-meter transmitter (capable of supplying 20 W) and the amplifier. The output of the amplifier was connected to a dummy load through another directional wattmeter. After a three-minute warm up, the amplifier was keyed with no drive and a zero-signal plate current of approximately 45 mA was verified. With one or two watts drive from the transmitter, the **TUNE** and **LOAD** capacitors were adjusted for maximum output. The **TUNE** and **LOAD** capacitors were made to tune properly at approximately half mesh by spreading or compressing turns of L4 in the output tank circuit. (Turn the juice off and discharge the plate before adjusting, please!—*Ed.*) The variable capacitor and inductor (L6) in the input-matching circuit were then adjusted for minimum input SWR. The drive power was then increased in steps to 20 W while keeping the **TUNE** and **LOAD** capacitors adjusted for maximum output. After the amplifier was tuned up with 20 W of drive, the input matching was touched up one last time for minimum input SWR. The final operating parameters observed for the amplifier are shown in Table 1.

Table 1—Amplifier Operating Conditions

Plate Voltage (under load):	2100 V
Idle Current (no drive):	45 mA
Plate Current (CW):	600 mA
Drive Power (CW):	20 W
Output Power (CW):	800 W
Efficiency (calculated):	63%

Note that the pi output tank circuit used in this amplifier may not provide enough harmonic attenuation to meet FCC spectral-purity requirements. I always use this amplifier with a low-pass filter, Industrial Communications Engineers (ICE) Model 426.

Conclusion

I have described the conversion of a Dentron MLA2500 HF amplifier for six-meter operation. The conversion involved building a new RF deck in the amplifier's existing chassis using a 3CX800A7 tube. The existing power supply, control and metering circuitry were mostly unchanged. It is my belief that using older HF amplifiers as the basis for building VHF amplifiers has many advantages, particularly for those who do not want to build a new amplifier completely from scratch. In closing, I would like to thank Dick Hanson, K5AND, and Pat Stein of Command Technologies for their help with this project.

Notes

[1]G. T. Daughters, AB6YL, "New Life for Dentron MLA2500s," *QST*, May 1996, pp 45-48.

[2]B. N. "Bob" Alper, W4OIW/6, "4CX400A Russian Tubes for the MLA2500 Amplifier," *Communications Quarterly*, Summer 1996, p 29.

[3]R. Schetgen, KU7G, Ed., *The ARRL Handbook for Radio Amateurs 1995*, 72nd ed. (Newington: ARRL, 1994), pp 13.48-13.58. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to **pubsales@arrl.org**. Check out the full ARRL publications line on the World Wide Web at **http://www.arrl.org/catalog**.

[4]S. Powlishen, K1FO, "A 3CX800A7 Amplifier for 432 MHz," *The ARRL UHF/Microwave Projects Manual* (Newington: ARRL, 1994), pp 8-12 through 8-28.

[5]C. Patterson, WA3HMK and Dick Hanson, N4HSM, "Conversion of Commander II, 2 to 6 Meters," *Proceedings of the 1997 Southeastern VHF Society Conference* (Newington: ARRL, 1997), pp 111-116.

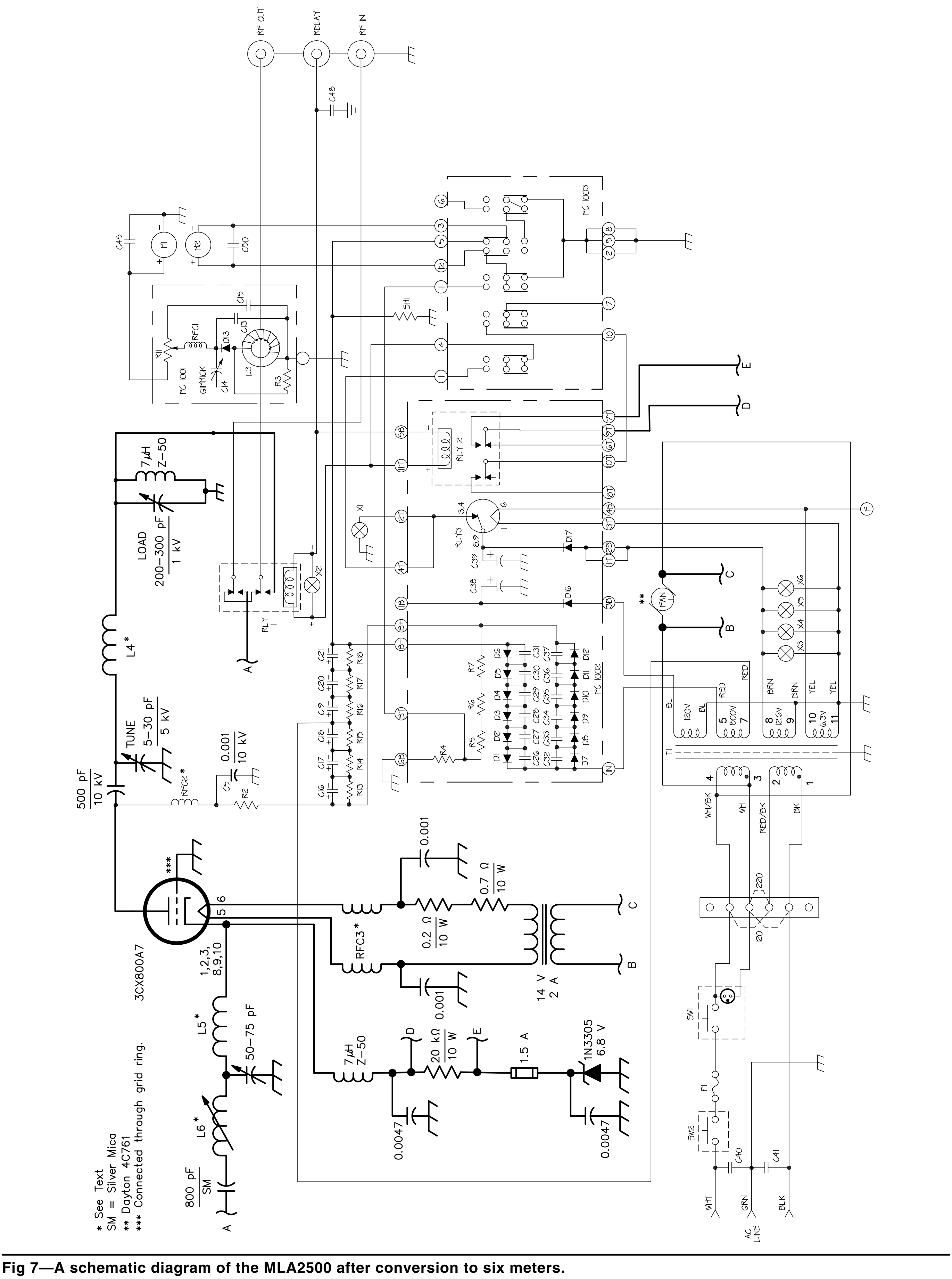

Fig 7—A schematic diagram of the MLA2500 after conversion to six meters.

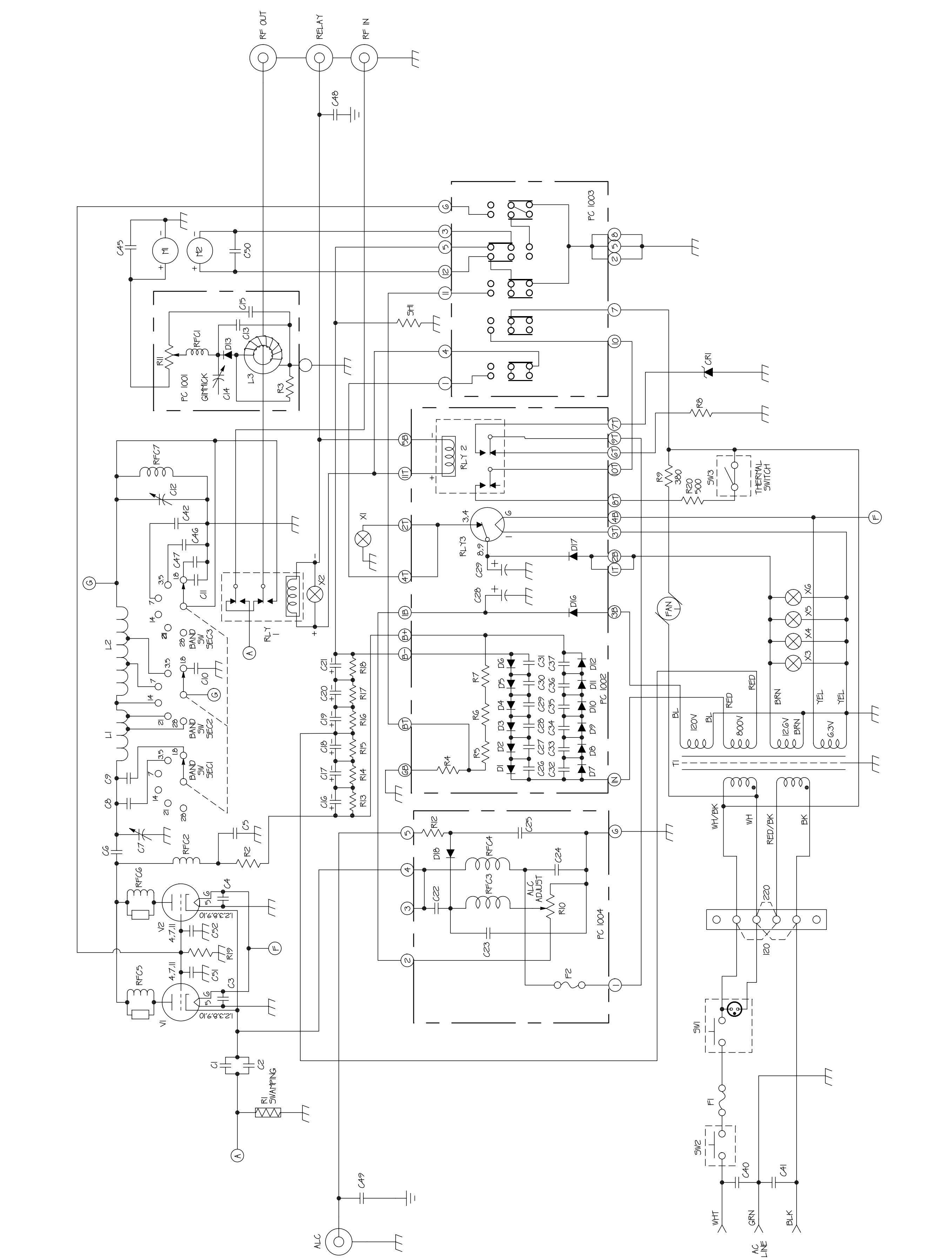

Fig 8—A copy of the original MLA2500 schematic for comparison.

Practical Application of Wind-Load Standards to Yagi Antennas: Part 2

Now let's look at practical formulas and work an example. We're on the road to wind-tough antenna-mast assemblies.

By Stuart E. Bonney, K5PB

Unless you are the adventurous type, the prospect of frequently climbing your antenna tower and spending hours strapped to the top of it may not meet your definition of fun. Most of us prefer to plan and install our beam antenna system once. We want to do it right and enjoy the fruits of these strenuous labors in confidence that the system will survive the inevitable storms and high winds that come our way.

The first of this two-part series reviewed the evolution of EIA standards applicable to Yagi-type antennas, discussed the physical and mathematical bases for wind-stress analysis, and described improved methods for determining and specifying wind loads.[1] Our emphasis from this point forward is primarily practical, concentrating on applications.

The current standard, TIA/EIA-222-F,[2] contains basic wind-velocity data as well as equations for calculating the effects of antenna height and wind gusts. Let's now examine how to apply this information to your own site, evaluate the antenna you plan to use, determine the wind loads it will put on your mast/tower and select a suitable mast. I've included tables and shortcuts to simplify the design process. To help you apply this material more easily, we will also look at several practical examples.

How Much Wind?

This may seem a facetious question, but it is indeed relevant. In brief, the answer depends on where you live and the height of your antenna. The answer is also related to how wind velocities are determined, which is a bit more complex than it might seem at first.

Wind data are collected and processed primarily by NOAA and its National Weather Service. Since wind is subject to extremely localized and short-term influences, velocities are averaged over large areas and long time periods to arrive at a basic maximum for a given geographical area. The US standard is based on the fastest mile, which is the velocity averaged over the time required for a single particle in a wind stream to travel one mile. At 60 mi/h, this is one minute. These velocities are specified at a standard height of 33 feet (10 meters) in open terrain, and they represent a basic maximum velocity that, statistically, will be exceeded once in 50 years.

In 222-F, these data are listed by state and county. For the Atlantic and Gulf coastal areas, a factor is included for hurricanes. Of course, if you ever take a direct hit from a big hurricane, statistical averages probably will be of small comfort. In addition, these data do not apply to tornadic winds.

The following is a gross summary of 222-F wind data, which actually run to nearly 33 pages of listings. For large interior areas of the US, 100 miles or more inland, basic maximum wind velocity is 70 mi/h. In windy areas of the upper Midwest and Plains states, it ranges from 80 to 90 mi/h, which applies also to Atlantic and Gulf coastal areas less than 100 miles inland but not

[1]Notes appear on page 49.

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Richardson, TX 75080

right on the coast. For most coastal counties from North Carolina to Texas, and a few coastal counties in the Pacific Northwest, it is 100 mi/h. For most California counties, it is either 70 or 75 mi/h, except for mountain regions in the north where it is 80 mi/h. As might be expected, south Florida and the Keys have the highest numbers, ranging from 100 to as much as 120 mi/h. These numbers tend to be slightly lower than in 222-C, mitigating somewhat the 222-C gust factor problem described in Part 1, although for some areas they are actually higher.

Determining basic maximum wind velocity for your location is the first step in site planning, and 222-E or F are the best sources. However, finding a copy may not be easy. Since the data are based on US weather records, a local or nearby National Weather Service office or airport may be able to provide data for your county or area. Local ordinances or building codes may also specify a figure. If you must resort to one of these alternate sources, look or ask for the 50-year basic maximum wind velocity. (*Never* design for winds less than your local building officials recommend. For legal and safety reasons, building officials are the ultimate authority for all structural decisions. If there is ever a problem, you want the Building Department on your side.—*Ed.*)

Height and Wind-Gust Corrections

Air is a slightly viscous fluid, meaning that it is subject to frictional effects. Thus, wind close to the ground is slowed by contact with the ground itself, and even more by surface obstructions such as trees and buildings. Neglecting for the moment any turbulence caused by surface roughness, this effect gradually diminishes with increasing height. At 900 to 1000 feet above ground in open country, it has virtually disappeared. Starting at our standard height of 33 feet, we see this effect as a steady, quantifiable increase in wind velocity with increasing height.

Surface obstructions and localized weather phenomena result in eddies and vortices that we perceive as short duration lulls and gusts in the wind. These effects are greatest at ground level and diminish slowly as we gain altitude, although they can extend to well over 1000 feet. Since basic wind-velocity figures are smoothed by the relatively long measurement periods, they do not indicate the magnitude of these short-term variations. We must account for gusts because they result in peak winds that an antenna system must be able to withstand.

In 222-F, the height factor is designated K_z and is expressed as:

$$K_z = \left(\frac{z}{33}\right)^{\frac{2}{7}} \qquad \text{(Eq 1)}$$

where z = antenna height above ground, in feet.

The minimum value of K_z is 1 for heights of 33 feet or less. The maximum is 2.58, which occurs at about 900 feet. For urban and heavily forested areas, the height factor expressed by Eq 1 overcompensates slightly for effective height and, thus, is conservative. For hilltop sites, figure your antenna height above the tower base and add the elevation above surrounding terrain. For example, if your tower is 100 feet tall and is on a hill that is 75 feet above average nearby terrain, use a height of 175 feet. If your antenna is mounted on the roof of a multistory building, use total distance to ground level as the height.

Bear in mind that for hilltop locations, adding the height above surrounding terrain is only a simplified rule of thumb. Long slopes can cause an accelerating effect beyond the mere increase in height, and it may be wise to add an empirical factor to compensate for such effects.

The gust factor is designated G_h and is expressed as follows:

$$G_h = 0.65 + \frac{0.60}{\left(\frac{h}{33}\right)^{\frac{1}{7}}} \qquad \text{(Eq 2)}$$

where h = antenna height above ground in feet.

The maximum value of G_h is 1.25, which is for antenna heights of 33 feet or less. The minimum value is 1, which applies to heights of about 1400 feet and above. Few of us will need to be very concerned about the minimum.

Again, keep the following in mind, especially if you live in a hilly or mountainous area or near high-rise buildings: Local topography greatly influences peak winds, and 222-F methods for determining anticipated velocities are most useful in establishing a nominal baseline. You may need to adjust that figure upward to account for unusual local conditions. A modest boost also could be justified as a cushion against the statistical averages.

Adjusting Basic Wind Velocity for Height and Gust Factors

As given in 222-F, the height and gust factors are simply additional terms (multipliers) in the equation for calculating total wind force. They are not included in the wind-velocity term, which represents the basic wind velocity described earlier. A more direct method is to adjust this velocity with a simple correction factor C_v calculated with the following equation. Note that we use the square root of the product of K_z and G_h, since total velocity is squared when computing actual wind force:

$$C_V = \sqrt{K_z \times G_h} \qquad \text{(Eq 3)}$$

To reduce the number crunching required, Equations 1, 2 and 3 were used to compute wind-correction factors for several heights. The smallest value is 1.12, for heights of 33 feet or less. Other values are shown in Table 1. To illustrate their use, let's assume that we wish to install an antenna at 100 feet in an area of northern Texas where basic wind velocity is 70 mi/h. From the table, the correction factor is 1.26. Therefore, the peak wind velocity for this location and height is 70 × 1.26 = 88 mi/h.

Similarly, for areas where basic wind velocity is 80 mi/h, and for an antenna height of 75 feet on a hill 50 feet above average terrain, we find the peak velocity to be 104 mi/h. For a stack of multiple antennas, use the height of the highest antenna. Corrections for heights between table entries can be interpolated. If you wish to include an additional velocity margin as mentioned earlier, use a table entry one or two steps higher than your actual height.

Peak velocity is the most important wind parameter for your installation, because it is a wind peak, even if only momentary, that can result in damage to your antenna. The wind-survival rating, if available, for your choice of antenna is normally a peak velocity. If a rating is not published, ask; it may encourage manufacturers who do not supply this information to do so in the

Table 1—Corrections to Basic Wind Velocity for Height and Gusts

Antenna Hgt. (ft)	*50*	*75*	*100*	*125*	*150*	*200*	*250*	*300*
Correction	1.17	1.22	1.26	1.30	1.32	1.37	1.40	1.43

future. Bear in mind as well that a peak rating based on RS-222-C with an assumed gust factor is probably erroneous.

The antenna wind-survival rating should be not less than the peak wind velocity for your location and installed height. Should you worry if the rating and peak velocity are nearly equal? Consider this: If a drag factor of 1.1 were to be used for elements and booms, which the data from several sources indicate is more accurate than 1.2, computed wind loads would be about 8% lower. This equates to an increase of about 4% in wind-survival velocity. In addition, antenna elements deflect in high winds. This, they can shed 10% or more of their wind loads, which translates to an additional 4 or 5% of wind-survival velocity. (A drag coefficient of 1.2 is still recommended for practical reasons, such as code compliance. We will use that figure in the following examples.)

Therefore, an antenna rated at, say, 90 mi/h under 222-F criteria probably will survive at least 95 mi/h. Looking at it in terms of the loads transferred to mast and tower, the actual load at 90 mi/h is no greater than the calculated load for 85 mi/h or slightly less. These numbers are in the right directions, but the margins are not large, so it is best not to skimp. There is always some degree of uncertainty about peak wind velocities, despite the best predictions. Notice that the first part of this analysis does not apply to antennas rated under 222-C, since the effective drag factor in that standard is already somewhat less than 1.1, as discussed in Part 1.

If you live in an urban, 70-mi/h basic wind area and your goal is a tribander at 60 feet, you will probably be safe with nearly any antenna from a reputable manufacturer. When big beams and tall towers on hilltops are your game, however, taking chances on the numbers can prove costly. If you build your own antennas, use the peak velocity for your site and installed height as minimums. Designing, building or modifying antennas for peak wind velocities of 100 mi/h or more has been shown quite achievable and affordable.[3] In large areas of the country, such designs will provide good wind-survival margins.

Evaluating Antenna Wind-Load Areas

Wind-load areas specified by manufacturers for their antennas often are based on 222-C flat-plate-equivalent criteria. Some also use the "square-root-of-the-sum-of-the-squares" method to arrive at a total-effective-area figure for the complete antenna. If it is not clear how the wind area has been specified, ask the manufacturer for details.

Weber[4] has shown that the square-root-of-the-sum-of-the-squares method is incorrect. Compared to the application of cross-flow principles, this method overstates real wind loads. For antennas with two-inch booms not more than about 25 feet long, where total-element area is usually 2.5 to 3 times boom area, the error is less than 15%. When element and boom areas are more nearly equal, as with wide-spaced beams on long, three-inch booms, the error runs as much as 40%. Here it is well worth the effort to determine accurate wind surface areas.

If wind-load areas are not specified, or the method used to specify them cannot be determined with confidence, it is possible to estimate them with acceptable accuracy. Table 2 lists typical projected areas for reflectors and directors used in Yagi antennas for 20, 15 and 10 meters. Two ranges are shown for each band; one for low taper rates, and one for high taper rates. The low-taper figures apply mainly to elements with tip-section diameters of $^3/_4$ inch or larger, which are rare in modern designs. High-taper designs generally use smaller tip diameters: $^7/_{16}$ or $^1/_2$ inch, sometimes $^5/_8$ inch for 20-meter elements. These are the most common sizes because they result in lower wind loads and greater survivability ratings. Boom areas are usually easier to calculate because most manufacturers provide boom dimensions in their literature.

Here's how to determine *WSA:* For elements, calculate projected area (length × OD, in square feet) for each section of each element, then total these numbers. If an element has traps, each trap typically adds about 0.15 ft^2 to the projected area. Trapped elements are shorter, but most of the shortening is in the tip section, which has the least effect on total area. If you do not have actual element dimensions, use the appropriate values from Table 2. As a shortcut, calculate the projected area of the reflector only, then multiply that area by the number of elements. Errors will be on the conservative side. Add areas for any traps after you have done this. Multiply total projected area for all cylindrical elements by a drag factor of 1.2; the result is the total *WSA* for the elements.

To find *WSA* for the boom, calculate its projected area, as above, and multiply by 1.2. For example, *WSA* for a 3-inch by 40-foot boom is 12.0 square feet. Add to this any flat-plate areas such as a boom-to-mast mount. Since the aspect ratio of these areas usually is 7 or less, multiply the projected area by a drag coefficient of 1.4. The sum is total *WSA* for the boom assembly. Element end areas and other mounting hardware usually can be safely ignored.

Determining Total Antenna Wind Load

Based on cross-flow concepts, the area used to determine wind loads for a complete antenna is simply the larger of total element *WSA* or boom *WSA*. To illustrate, let's use a popular four-element 20-meter mono-band antenna, the hy-gain 204BA. The measured projected area for the reflector is about 2.5 square feet, and for the second director is about 2.2 square feet. Taking the average for four elements and including an appropriate drag factor yields the total *WSA* for the elements. Therefore, *WSA* = 2.35(4)(1.2) = 11.3 square feet.

The boom is 26 feet long by two inches in diameter. Calculated *WSA* is 5.2 square feet. The boom-to-mast fitting, with a projected area of 0.3 square feet and an aspect ratio less than 7, has a *WSA* of about 0.4 square feet. Adding these figures, we have a net *WSA* of about 5.6 square feet for the boom assembly. Since this is smaller than the element *WSA*, it is the latter by itself that determines effective *WSA* for the complete antenna.

Once *WSA* is known, the actual wind load at any velocity can be calculated.

Table 2—Typical Projected Areas for Elements (Ft^2)

Element	*20 Meters*	*15 Meters*	*10 Meters*
Reflector - LT	2.8 - 3.0	1.7 - 1.8	1.0 - 1.2
Reflector - HT	2.4 - 2.7	1.4 - 1.5	0.8 - 0.9
Directors - LT	2.6 - 2.8	1.5 - 1.6	0.9 - 1.0
Directors - HT	2.2 - 2.5	1.3 - 1.4	0.7 - 0.8

Note: LT = low taper; HT = high taper

Normally you use peak wind velocity for your installed antenna height as previously determined. Wind load is then calculated using the following equation (repeated from Part 1 for convenience).

$$F = 0.00256V^2 WSA \qquad \text{(Eq 4)}$$

or, alternatively,

$$F = V^2 (WSA)/390 \qquad \text{(Eq 5)}$$

where:
- F = Wind load force, in pounds
- V = Peak site wind velocity, in mi/h
- WSA = Antenna wind surface area, in square feet.

As an example, let's assume that your peak site wind velocity is 90 mi/h, and you plan to install the 204BA beam from the preceding example. As described, total WSA for the four elements is larger than WSA for the boom and totals 11.3 square feet. Using Eq 5, the peak wind-load force, F, is $90^2(11.3)/390=235$ lbs.

Mast Forces and Strength Requirements

Mast selection probably receives less thought than many other aspects of the typical installation. If you mount only a tribander a foot or two above the top of your tower, nearly any reasonable approach to mast selection will suffice. If you want a longer mast to increase antenna height, or if you contemplate any kind of multiple-antenna stack, mast loads go up rapidly and demand more than casual treatment.

Mast loads consist mainly of transferred antenna wind loads. To a lesser degree, they depend on direct wind loads on the mast itself. These loads create a bending force, or more precisely, a moment, which the mast must withstand without bending permanently. A moment is force acting through a distance; in this case, from an antenna to the top of the tower thrust bearing:

$$M = FD \qquad \text{(Eq 6)}$$

where:
- M = Moment; in foot-pounds
- F = Force, in pounds
- D = Distance, in feet, over which force acts.

Mast strength is a function of the material's rated yield strength and its section modulus, which is a parameter related to shape and thickness. These determine the maximum moment a mast will withstand. If exceeded, a permanent bend will result. Table 3 lists several popular mast materials and indicates the maximum bending moments each will handle.

For tubing and pipe sizes not listed in the table, you can calculate maximum allowable moments using the following equations. First, calculate the section modulus (SM) based on outside (OD) and inside (ID) diameters in inches:

(for cylinders)

$$SM = \left(0.098OD^4 - ID^4\right)/OD \qquad \text{(Eq 7)}$$

Then look up the rated yield point (YP) for the material in an appropriate source, or use values from Table 3, if applicable. The maximum allowable moment, $M_{\max}$ is found as follows:

$$M_{\max} = YP(SM)/12 \qquad \text{(Eq 8)}$$

Since commercial yield-point ratings for aluminum are averages rather than minimums, some engineers prefer to use more-conservative ratings that are based on military, rather than commercial, standards. A common figure for 6061-T6, for instance, is 35 kpsi rather than the 40 kpsi shown in Table 3. The use of commercial standards is defensible on at least two grounds. First, antenna loads calculated with 222-F tend to be slightly overstated, as previously discussed. Second, it is good practice to allow a reasonable margin of unused capacity when selecting masts. In a practical sense, therefore, the question depends on how you prefer to allocate load margins at various points in the system.

Calculating Mast Loads

Fig 1 illustrates a mast supporting two Yagi antennas, seen broadside to the boom and end-on to the elements. Wind is broadside to the elements. The mast bending moment in this example actually consists of three forces and three distances: the wind forces F_1 on antenna 1, F_2 on antenna 2, and F_m on the mast itself, and the distances D_1, D_2 and D_m. These moments are exerted at the point where the mast emerges from the typical welded sleeve or thrust bearing at the top of the tower. The total moment is the sum of the individual moments. The general case can be expressed as follows:

$$M_{tot} = F_1(D_1) + F_2(D_2) \cdots + F_m(D_m) \qquad \text{(Eq 9)}$$

where:
- M_{tot} = total moment on mast, in foot pounds

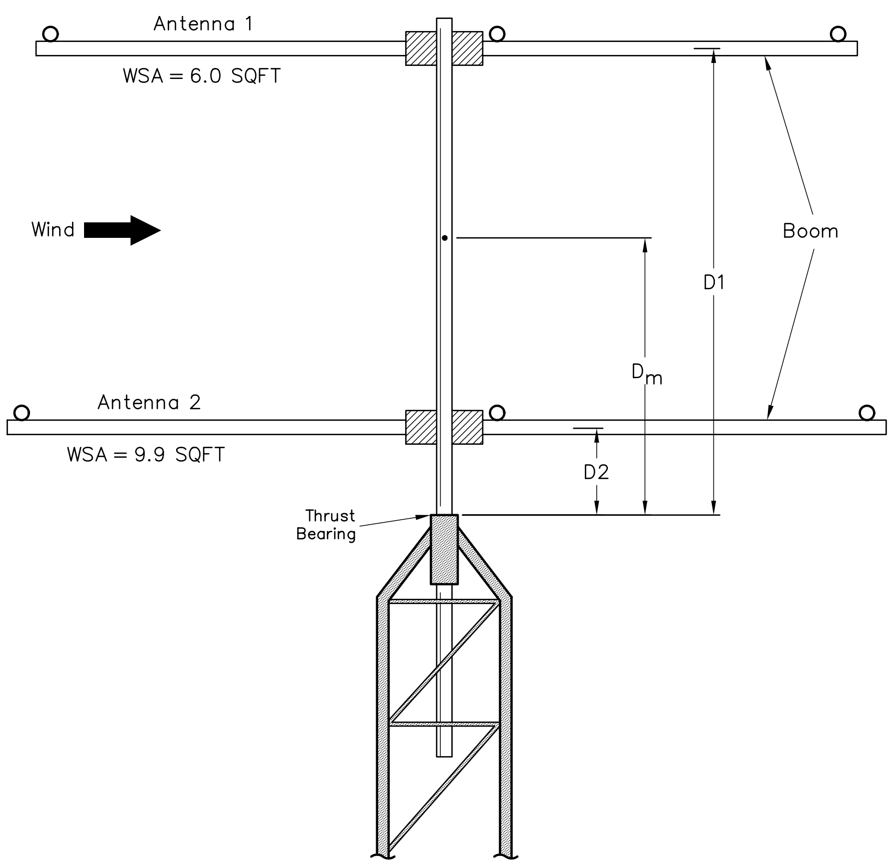

Fig 1—Typical mast bending moments due to wind forces.

F_1, F_2, etc. = individual wind forces (lbs) exerted on attached antennas
D_1, D_2, etc. = distance (ft) of antenna above mast bearing
F_m = wind force (lbs) on mast itself
D_m = one-half of the mast length (ft) above the tower bearing

To illustrate how these moments are calculated, let's assume that antenna 1 has a total element *WSA* of 6 square feet and is mounted 7.5 feet above the tower bearing. Using Eq 5 and assuming a peak wind velocity of 100 mi/h, the wind load for this antenna is 154 pounds. Multiply this figure by the distance, 7.5 feet, to determine the mast bending moment: 1155 foot-pounds. Similarly, let's assume that antenna 2 has an element *WSA* of 9.9 square feet and is mounted 1.5 feet above the tower bearing. Its corresponding wind load is 253 pounds, and the moment is 380 foot-pounds.

We will assume the mast itself extends a total of 8 feet beyond the bearing and is 2 inches in diameter. Its projected area is 8 (2 / 12) = 1.3 square feet. With a drag coefficient of 1.2, *WSA* is about 1.6 square feet. Since the wind force acts over the entire length of the mast above the bearing, D_m is the average or one half of this length. Thus, the wind load is 41 pounds and the moment is 41(4) or 164 foot-pounds.

The total mast bending moment at a wind velocity of 100 mi/h, therefore, is 1155+380+164 = 1699 foot-pounds. Checking Table 3, we find that a mast made of 6061-T6 aluminum tubing with a 2-inch OD and 0.250 wall thickness is adequate, although its overload margin is only about 5%. We could also choose a 1020 DOM carbon steel mast of 2-inch diameter and 0.120-inch wall. This would give a more comfortable additional load capacity of about 15%.

The force transferred to the tower is simply the sum of individual wind loads, not the moments. In our example, this is 154 + 253 + 41 = 448 pounds, which does not include any allowances for rotator and mast below the bearing.

Before moving on, a few other points are worth noting. First, notice that antenna 1, although smaller than 2, contributes much more to total mast load because of its longer moment arm. Therefore, spacing is a trade-off. Closer spacing reduces mast loads, but electrical performance suffers if it is too close. In my experience, the spacing in that example is adequate for HF antennas with moderate boom lengths. I've observed some interaction as slight shifts in resonance and the frequency of maximum front to back ratio, compared to an isolated antenna.

Second, changing out a bent mast in a stack of antennas is usually no small task, so including a bit of insurance here may be worthwhile. This can often be done merely by going up another step in wall thickness. A larger mast diameter increases strength even faster, if your tower bearing can handle it, although this also increases the wind load slightly. Finally, don't forget to check mast loading with booms broadside to the wind, especially if you use long, large-diameter booms.

Tips For Mast Selection

An aluminum mast is a good choice if properly selected and applied. While not as strong as steel of equal size, it weighs only about one-third as much and is much easier to handle at the top of a tower. Its cost is comparable to that of the better grades of steel mast. For corrosion resistance and strength, only seamless 6061-T6 tubing or pipe is recommended. A good mast bearing is important with aluminum. The steel sleeve used in some tower top sections can induce abrasion wear at the most critical point on a mast.

Steel, especially in the high-strength grades, is often an obvious choice for stacks of medium to large arrays. Hot-dip galvanizing, inside and out, is desirable to prevent rust, and many steels suitable for antenna masts can be galvanized without loss of strength. Steel tubing and pipe are available in a wide range of carbon content, alloys, fabrication techniques and heat treatments. The best practice is to deal with a reputable supplier. Then ask if published yield strengths are certified and by whom.

Ordinary 1½-inch (1.900 OD) galvanized steel water pipe has been an old standby for mast material, and many hams swear that it works just fine. It's cheap and readily available, but using it can be risky for anything more than a single HF beam mounted not more than a foot or two above the mast bearing. As commonly sold, its strength and quality are largely unknown, so why risk expensive antennas just to save a few dollars on a mast?

Dealing With Ice Loads

If you live in the upper South or southern plains (where I live) or along parts of the eastern seaboard, you know that icing in winter is common, although big ice storms are infrequent. Most serious for antennas is freezing rain, where the precipitation falls as rain but temperatures at ground level freeze it on contact. Atmospheric conditions under which this occurs are limited, but significant ice buildups can occur when they persist.

The results are twofold. First, the buildup of ice adds to wind surface area, increasing wind loads. Second, the antenna members must bear the added weight of the ice, which in solid form weighs about 56 pounds per cubic foot. There are generally two ways of dealing with this. An antenna can be designed and rated for full ice-loaded conditions, or for a maximum wind survival rating under dry conditions with appropriate derating for ice loads. Manufactured amateur antennas are generally rated without ice, and some do not have wind ratings at all.

Table 3—Typical Mast Materials, Sizes and Strengths

Aluminum Pipe 6061-T6	*OD (inches)*	*Wall (inches)*	*Yield (psi)*	*Section Modulus*	*Maximum Moment (ft lb)*
1½ Schedule 40	1.900	0.145	39 k	0.325	1058
1½ Schedule 80	1.900	0.200	39 k	0.411	1336
Aluminum Tubing 6061-T6	*OD (inches)*	*Wall (inches)*	*Yield (psi)*	*Section Modulus*	*Maximum Moment (ft lb)*
2-Inch	2.000	0.120	40 k	0.314	1047
2-Inch	2.000	0.188	40 k	0.443	1477
2-Inch	2.000	0.250	40 k	0.536	1787
3-Inch	3.000	0.250	40 k	1.370	4567
Steel Tubing (Type)	*OD (inches)*	*Wall (inches)*	*Yield (psi)*	*Section Modulus*	*Maximum Moment (ft lb)*
1020 DOM Carbon	2.000	0.120	75 k	0.314	1963
1026 DOM Carbon	2.000	0.188	83 k	0.443	3064
1026 DOM Carbon	2.000	0.250	83 k	0.536	3707
4130 HT Chr. Moly	2.000	0.250	110 k	0.536	4914
4130 HT Chr. Moly	2.000	0.375	110 k	0.664	6087

Note: DOM = drawn over mandrel, HT = heat treated

Although 222-F does not specify a particular thickness of ice, it recommends a maximum of 0.5 inches of radial ice, which is consistent with other standards. In Annex A, the standard also notes that peak winds seldom occur simultaneously with extreme ice loading, and that basic wind loads can be reduced under icing conditions to 75% of normal values (equal to 87% of peak wind velocity). Rough calculations suggest that typical amateur Yagis would survive only 50 to 70% of normal peak velocity under these conditions. However, this does not necessarily mean that these designs are deficient. In most areas, this may be sufficient for periods of many years, even decades, due to the infrequency of such extreme conditions.

In more than 30 years in the ice belt, I have found that severe icing conditions in my locale are more likely to occur with nearly dead-still to moderate winds. A little wind has actually been helpful; it kicks off antenna vibrations and flexing that result in ice shedding, so that large buildups tend not to occur. As evidence, the ground beneath actually becomes littered with small semi-cylindrical ice shards. This does not mean that large buildups and high winds never occur together. We all have seen dramatic photographs of icy, broken antennas, but other factors can be involved as well. Examinations of local antenna failures due to ice storms sometimes have indicated that construction was below par, or that metal fatigue may have contributed. North Texas is well populated with hams and beam antennas, and affords numerous opportunities to observe such failures. Actual failure rates, however, have been low.

Certainly, it is possible to design and build antennas capable of surviving 100-mi/h winds with half-inch ice loads if cost is no object. Whether it makes economic sense in light of probabilities is the real question. For amateur antennas, it seems quite reasonable to design for dry wind loads, choosing design velocities in excess of expected peak local wind velocities. This allows for good overload margins balanced by acceptable costs. With good construction, the result will be robust antennas with survival qualities consistent with the conditions most likely to occur.

Some Notes On Cross-Flow Principles

Cross-flow principles of fluid dynamics are described in a book by Dr. S. F. Hoerner, a German scientist who emigrated to the US after World War 2. His book, *Fluid Dynamic Drag,* is difficult to find although often cited. The concepts it covers, described in Dick Weber's 1993 article (see Note 4), cast new light on observations I had made over several years.

My homebrewed beams for 15 and 20 meters are made of 6061-T6 aluminum and are stacked at about 55 feet. When I built them in the early '70s, I gave more thought to electrical design than wind loads. The elements use $^{7}/_{8}$-inch tubing for tip sections. The mast is 6061-T6 pipe, which is more limber than a steel mast of the same size.

I soon found that antennas and mast both whipped severely in the fierce spring storms and high winds common to northern Texas. Thereafter, I rotated the antennas during storms to a heading where they seemed to ride easier and mast deflection was reduced. Invariably this occurred about 40(to 50(off the primary wind direction. This defense often required further rotation as the wind veered with the passage of a front, tending to confirm the easing effect.

These repeated observations were at odds with the notion that the net effective wind area of a Yagi is the square root of the sum of the squares of boom area and total element area. By this theory, the wind force on my antennas should peak at some angle off the wind. My eyes consistently told me, however, that the force then was *not more, but less.*

A few years later I wrote a PC program to model wind stress. I found that my 20-meter beam had a computed wind survival rating of only about 80 mi/h, mainly because of its "fat" tips. Yet, this antenna has survived undamaged for 25 years, despite storms that uprooted nearby trees or broke off large limbs. I attribute this largely to having taken advantage of the cross-flow effect.

Conclusions and Acknowledgments

As we have seen, methods for determining antenna wind loads based on obsolete standards leave much to be desired. At best, they have often resulted in confusion and inconsistent application. Improved standards and methods are available. They are more detailed, precise and easier to apply. Their universal adoption would be a major step toward common specification methods for antenna manufacturers, providing a level field for all. For purchasers, the ultimate benefits are confidence in the numbers, simplified antenna wind-load evaluation and a comprehensive basis for installation planning.

The practical methods described here in Part 2 can be used to determine the peak wind velocity for your location and installed antenna height. They can also help you calculate antenna wind loads and select a suitable mast. If you have previously felt uncertain about how to do all this, I hope that this material encourages you and helps you achieve not only good results, but also the satisfaction of knowing you did it yourself.

Thanks to Dick Weber, K5IU, who provided vigorous discussions as well as numerous useful suggestions. Thanks also to Warren Bruene, W5OLY and Dr Tim Bratton, K5RA, for manuscript reviews and comments.

Notes

[1]S. E. Bonney, K5PB, "Practical Application of Wind Load Standards to Yagi Antennas, Part 1," *QEX*, Jan/Feb 1999, pp 46-50.

[2]ANSI/TIA/EIA-222-F, *Structural Standards for Steel Antenna Towers and Antenna Supporting Structures*, Electronic Industries Association, Washington, DC, March 1996. TIA/EIA distribute their standards through Global Engineering Documents. You can order the standard from Global at 800-854-7179 (US and Canada) or 303-397-7956 (outside US and Canada). Global also has a Web page at **http://global.ihs.com**. The standard is 121 pages long and costs $88. The parts applicable to Yagi anatennas are included in this article.

[3]David B. Leeson, W6NL (W6QHS), *Physical Design of Yagi Antennas* (Newington: ARRL, 1992).

[4]D. Weber, K5IU, "Determination of Yagi Wind Loads Using the 'Cross Flow Principle,'" *Communications Quarterly*, Spring 1993.

Power Supply for a MOSFET Power Amplifier

Moderate-voltage FETs are becoming common and inexpensive. They offer inexpensive watts, but 40 to 50-V supplies are rare. Here's one designed for a 120-W amplifier.

By William E. Sabin, W0IYH

As part of a 120-W output MOSFET broadband amplifier project—with good intermodulation-distortion specifications—for the HF (1.8 to 29.7 MHz) bands, I needed a 40 V, 8 A (modifiable to 50 V, 7 A) regulated supply with over-current, over-voltage and short-circuit protection. The amplifier output-stage MOSFETs need to be well protected from damage. The power-supply design helps with this important task.

I could not use the conventional approaches that are common in 12 and 24-V supplies because the widely used regulator chips and power transistors do not have sufficient voltage ratings for a 40-V (or 50-V) supply.[1] Therefore, the circuit of Fig 1 evolved, which satisfies all requirements for my application. This circuit is very similar to a simple discrete circuit that has been around for many years (for example, see *The 1971 ARRL Handbook*), but some refinements make it an interesting subject for discussion and appropriate for this application.

The power transformer was specially designed and built for me by Ronald Williams, W9YZ,[2] and has the model number AV-574. This transformer is available from him for $90, including UPS ground delivery in the continental US. It has primary taps for 112/117/122 V, 60 Hz and a 36 V, 8 A secondary. The rectified dc is a little more than 52 V at no load. I was not able to find a transformer with these specs in any of the catalogs available to me. This transformer is excellent, and I highly recommend it. A transformer used with a capacitor-input filter should—in principle—have a higher rating than 8 A RMS for an 8 A dc output. Nevertheless, the temperature rise for this transformer is quite reasonable for typical Amateur Radio applications, where over-design isn't normally needed.

Although the pair of MOSFETs that I use are rated for 50-V dc operation, I wanted to use 40 V for an extra margin of protection. I can easily get the 120-W power output with good distortion products at this supply voltage. This no-tune amplifier can be used

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[1]Notes appear on page 54.

either "barefoot," or to drive a legal-limit, grounded-grid PA.

Circuit Discussion

In Fig 1, Q1 drives the bases of Q2-Q5 in parallel. The output voltage (24 to 40 V) is divided down by R1 through R3 and sent to the base of Q6. This voltage is compared with the highly regulated and ripple-free 17-V reference voltage supplied to the emitter of Q6 by the LM317. The collector of Q6 controls the base of Q1, thus closing the regulator loop. The LM317 input cannot handle the full 50-V supply, so D7, a 5-W Zener, drops 22 V.

Current limiting is provided by Q7. When the current exceeds 8 A, the voltage drop across R4 and R5 makes Q7 conduct. That pulls the base of Q1 down, limiting the current through Q2-Q5 to no more than about 8 A, which also reduces the output voltage. D2 and D3 limit the transient base current of Q7 to a safe value. This circuit also employs a "fold-back" feature. A constant current of 6 mA flows through R6 and R7. This current is determined by Q8, R8 and the 17-V reference. If the output falls below 17 V or so, D4 stops conducting. The short-circuit current drops quickly to about 4 A, because there is no longer any voltage drop across R6 and R7. The 6 mA constant current assures that the current limiting and fold-back operate the same at 24 V as they do at 40 V.

There is a good reason for the 24 to 40-V adjustment pot. When designing a PA it is necessary to run the voltage up and down at many frequencies, to look for oscillations or other peculiari-

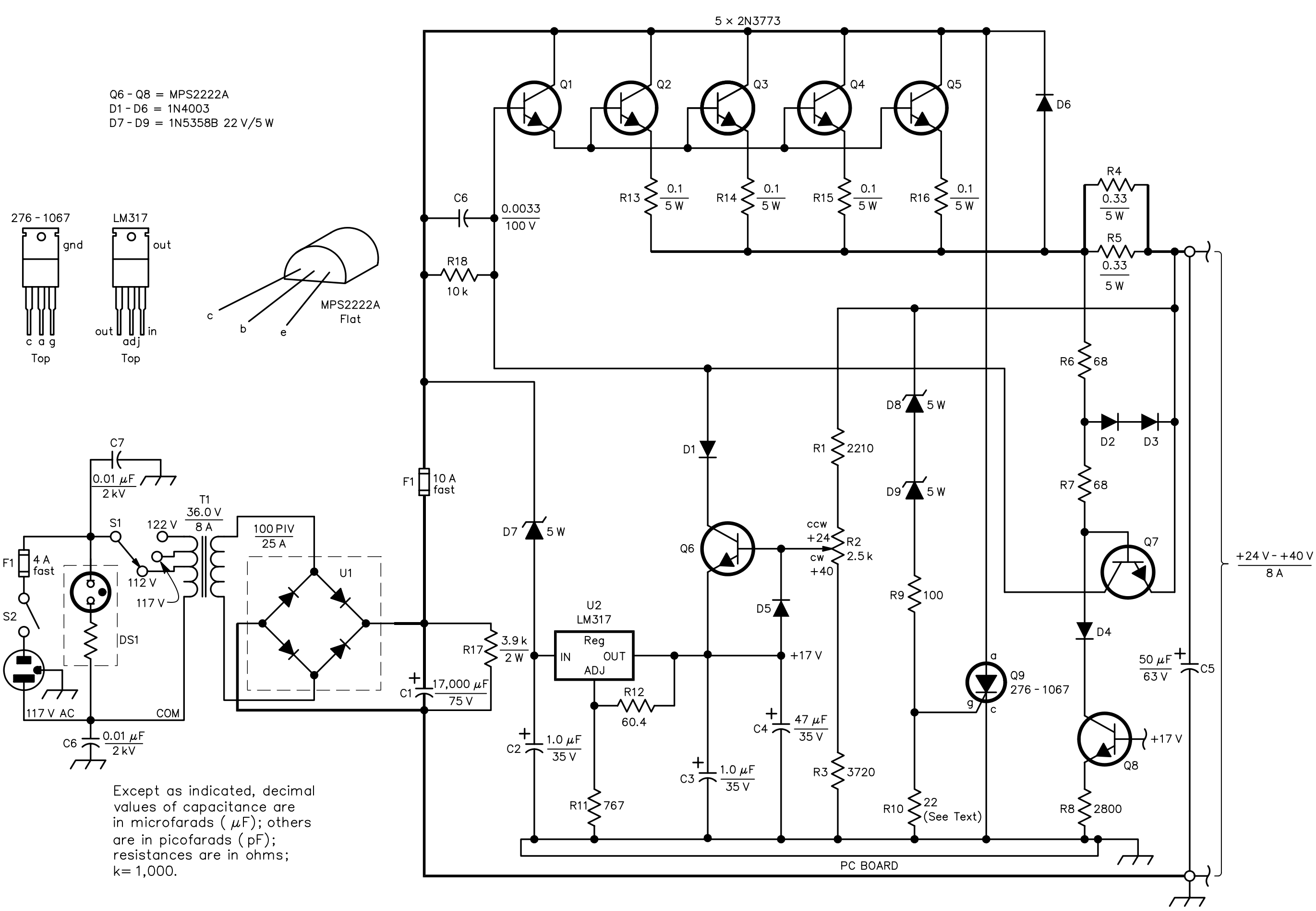

Fig 1—Schematic diagram. Unless otherwise specified, use 1/4 W, 5%-tolerance carbon composition or film resistors. Equivalent parts may be substituted for those shown here. RS indicates RadioShack part numbers.

C1—17000 µF, 70 V (Sprague 36DX)
C2, C3—1.0 µF, 35 V (RS 272-1434)
C4—47 µF, 35 V (RS 272-1027)
C5—50 µF, 63 V (Multicomp MCLV series)
C6—0.0033 µF, 100 V
C7, C8—0.01µF, 2 kV (RS 272-160)
D1-D6—1N4003 (RS 276-1102)
D7-D9—1N5358B 22 V, 5 W
DS1—red neon lamp (RS 272-712)
F1—4 A fast-blow fuse (RS 270-1010)
F2—10 A fast-blow fuse (RS 270-1015)
Fuse Holder (2, RS 270-367)
Q1-Q5—2N3773
Q6-Q8—MPS2222A (RS 276-2009)
Q9—SCR (RS 276-1067)
R1—2210 Ω metal film, 1%
R2—2.5 kΩ (Clarostat RV4 series or equiv)
R3—3720 Ω metal film, 1%
R4, R5—0.33 Ω, 5 W wire wound, 5%
R6, R7—68 Ω, 1/4 W, 5%
R8—2800 Ω metal film, 1%
R9—100 Ω, 1/4 W
R10—22 Ω, 1/4 W, test-select (see text)
R11—767 Ω metal film, 1%
R12—60.4 Ω metal film, 1%
R13-R16—0.1 Ω, 5 W wire wound
R17—3.9 kΩ, 2 W
S1—SP3T switch non-shorting
S2—SPST switch 120 V, 10 A (RS 275-690)
T1—Avatar AV-574 (see Note 2)
U1—25 A, 100 V PIV (Multicomp GBPC2501 or equiv)
U2—LM317 regulator; use mounting kit (RS 276-1373)
Heat sink—Wakefield 441K; use insulator kits (RS 276-1371)
Heat sink—Wakefield 621A; use nsulator kit (RS 276-1371)
Heat sink grease (RS 276-1372)

ties.[3, 4] On a new design, we want to bring the voltage up gradually to look for anything that might zap or overheat the MOSFETs.

A problem occurs in the collector of Q6 if D1 is not present. If the output voltage falls below 17 V, the reference voltage feeds through D5 to Q6's base, and then through the now-forward-biased base-collector junction. The collector is then clamped to about 17 V, thus defeating the short-circuit protection. This causes damage to one or more of the transistors (Q1 through Q7) so D1 is a crucial component. D4 is important for the same reason. D5 prevents reverse biasing of Q6's emitter-base junction, which is not recommended by transistor manufacturers. (The emitter-base reverse breakdown voltage is quite low, around 5 V dc.—*Ed.*) This power supply is a "beast," but the MPS2222As are not, so they are well protected.

One reason for using the unusually high 17-V reference is that it reduces the amount of voltage division in R1 through R3. This helps to increase the regulator loop gain. The 17-V reference also reduces the collector-emitter voltage of Q6 and Q8 and the dissipation in U2. Extremely tight regulation is not needed in this application, and the circuit is more than adequate. Fluctuations in output voltage due to the PA's speech-dependent load variations are almost entirely regulated out by the wide-bandwidth regulator loop. C5 and the combination of R4-R5 with the output resistance of Q2 through Q5 help to establish the open-loop bandwidth, but C5 is not a good high- frequency bypass, so C6 prevents oscillations.

The components Q9, D8, D9 and R9-R10 protect against excessive output voltage. If this voltage exceeds about 47 V for any reason, the SCR blows the 10 A fuse. All dc circuitry is thereby very quickly disconnected. R10's value is test-selected to assure the 47-V protection level. At the same time, D6 provides a path to ground through Q9 that very quickly discharges C5 and any other capacitors in the PA. D6 also prevents any inadvertent reverse conduction of Q1 through Q5, which might damage them.

The story regarding Q1 through Q5 is interesting. The widely used 2N3055 has a 60-V breakdown value, so it is not quite good enough to achieve a safe margin. The 2N3773 has a 140-V breakdown and costs only about $1 more. Of particular concern is the safe-operating area (SOAR) of these transistors. Using data sheets downloaded from Motorola's **http://www.mot.com**, we get the simplified dc SOAR diagram of Fig 2. At 50 V V_{CE}, corresponding to a short-circuit load, we do not want a continuous 2 A to flow in each of the four transistors (shown as point X). This would correspond to an 8 A short-circuit output, if there is no fold-back. The fold-back reduces the current to about 1 A per transistor, shown as point Y, which is a lot safer. This requirement regarding SOAR is real and must be treated with respect. Also, the heat sink must be able to hold the short-circuit case temperature rise in Q2 through Q5 to about 60°C,[5] which degrades the SOAR diagram significantly, as learned from the data sheets. The fold-back is very helpful in this respect also. The fold-back also keeps the transformer cool if a long-term short circuit occurs at the output. The question arises: Why not use a smaller

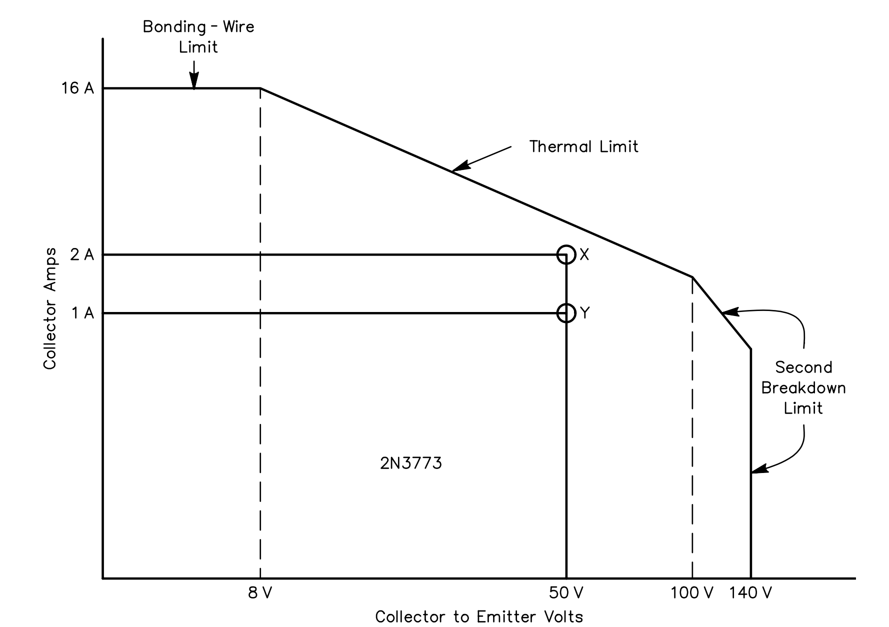

Fig 2—2N3772 SOAR diagram.

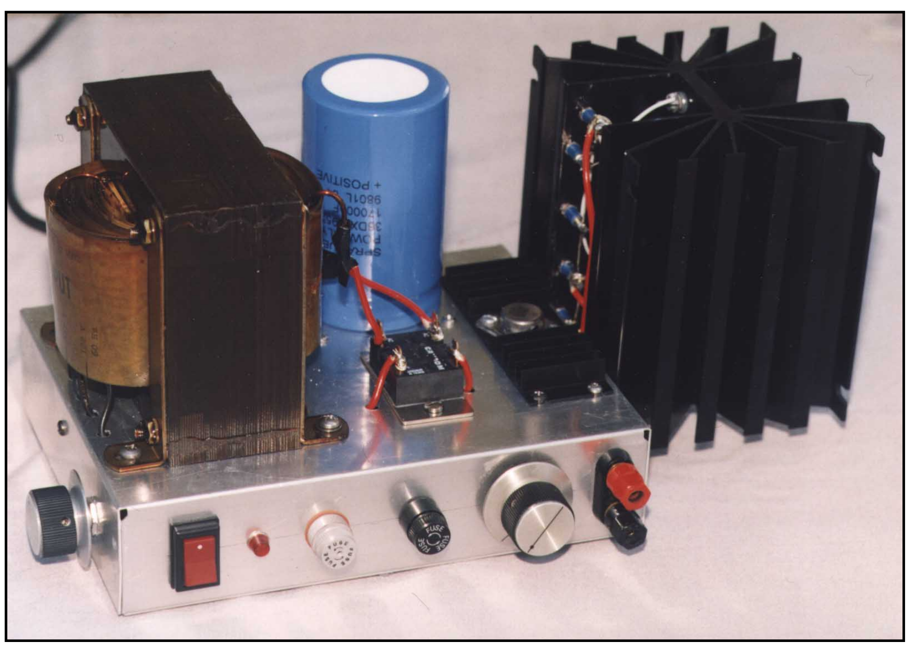

Fig 3—Outside view of the power supply. The primary-winding-tap switch (optional) is on the left side; it should never be "hot-switched." The heat sink is oriented so that air can circulate vertically through it. The ac fuse is to the left of the dc fuse.

heat sink and reduce the short circuit current even more? I decided that in any 24 V, 8 A experiments that I might do, where the heat sink would dissipate (52 – 24) × 8 = 224 W (56 W per transistor), the larger heat sink would be advisable. This is one of the penalties of the series regulator.

Construction

Figs 3 and 4 show the construction details. A 9×7×2-inch chassis is used. A secure bottom cover with rubber feet is advisable to prevent accidents. A PC board contains the low-level circuitry, as suggested in the schematic. I attached it to two of the #10-32 transformer mounting screws, but other mounting methods are possible. The ground plane should be solidly grounded. It is important that a wire from F2 connect *directly* to the anode of Q9 (*not* through a trace on the PC board). A trace cannot handle the large current that blows the fuse. This is shown in Fig 4.

A PC board is available from FAR Circuits.[6] An optional rotary switch selects the primary tap on T1, but this switch should never be "hot-switched," to prevent damage to the switch contacts. C1 stores a lot of energy—(53 V)^2 (17000 µF) (0.5) = 23.9 joules. That energy can make a very loud spark and do damage if suddenly discharged. Note also that the bridge rectifier must have a 100-V rating. To quickly discharge C1 after turning off the power, apply a short at the output terminals that limits the discharge current to 4 A. C1 is supported by its terminal screws on a 2½×2½ inch piece of RadioShack perfboard or other insulating material. Both terminals are isolated from ground by a pair of large holes in the chassis.

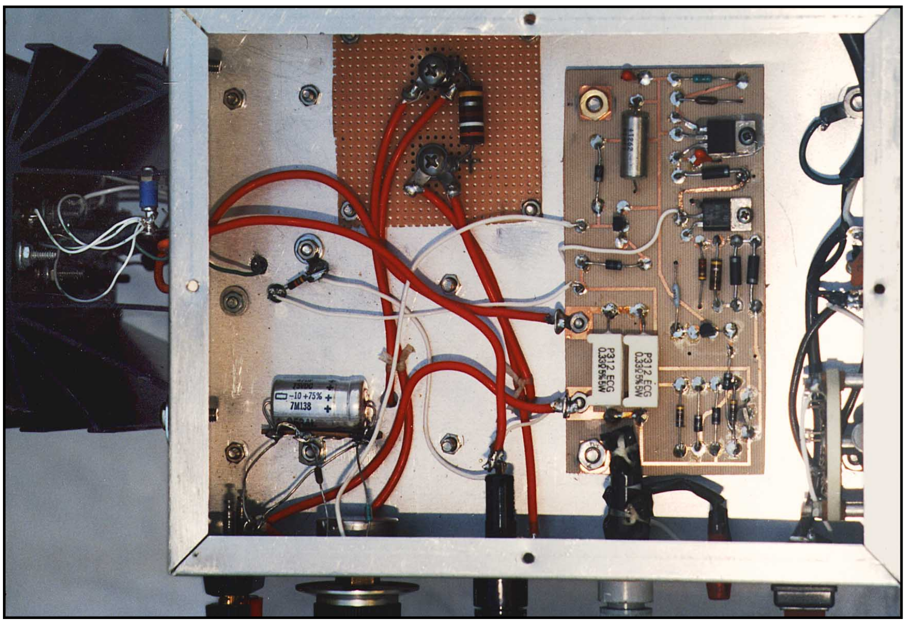

Fig 4—Under-chassis view. The PC board is mounted on two of the transformer's #10-32 mounting screws. Notice the white wire from the anode of Q9 (center pin, lower TO-220 case) across the PC board and underneath it directly to the dc fuse holder (important). R18 is connected at Q1, base to collector (left side of photo). R4 and R5 (large white blocks at lower left of PC board) are slightly elevated from the PC board.

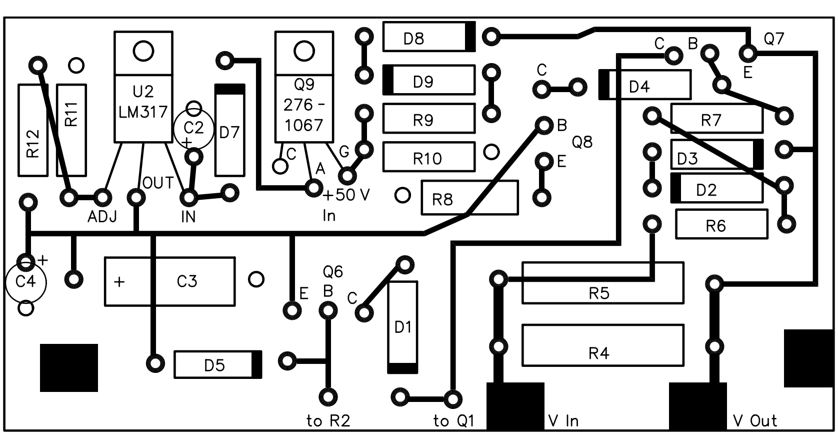

Fig 5—This part-placement diagram helps identify the parts in Fig 4.

Metal-film 1% resistors are recommended for R1, R3, R8, R11 and R12. R2 should be a high quality, 2-W pot. R6, R7 and R18 should be checked and selected to be sure they are within 5% of each other, or they may be 1% metal film.

The heat sink for Q2 through Q5 is the Wakefield 441K (use mica washers). Anything smaller than this is not recommended for the reasons mentioned previously. R13 through R16 and D6 are also mounted on this heat sink, using terminal strips or standoff insulators. Note that the heat sink is oriented so that convection carries air up through the bottom. Q1 mounts on a Wakefield 621A, also with a mica washer. Number 12 stranded wire is used for the heavy lines shown in Fig 1. Fig 4 shows how I attached these wires to the PC board using solder lugs and #4 hardware. R4 and R5 should be mounted above the board about ⅛ inch because they get warm. In particular, the ripple current of C1 should not be allowed to flow through the chassis, but must be isolated with a single-point ground at the negative binding post, as shown in Fig 1. C5 connects across the binding posts, and R1 is connected very close to the positive binding post.

The three-wire line cord has the green wire tied to the chassis (*do it!*). A strain relief is mandatory for this cord. All exposed 120-V wiring should be covered to prevent personal injury. The 50 to 55-V dc in this supply is enough voltage to pose a hazard. We are all accustomed to sticking our fingers into low-voltage transistor circuitry, but it is a bad idea at this voltage level.

Here's a test-select procedure for R10. Connect a 10 kΩ resistor across R3 so that the output voltage can be increased beyond 47 V by adjusting R2. Select R10 so that F2 does not blow below the 47-V level. Then remove the 10 kΩ resistor. An adjustable pot for R10 (that can become misadjusted) is not advisable.

Be sure to check everything very

carefully with an ohmmeter before turning the thing on. A gradual start-up using a variable transformer is a good idea, if one is available. Monitor the output voltage as you bring it up. Make sure that R2 is working okay. The next step is to gradually increase the loading beyond 8 A and verify that the current limiting works. Then apply a short circuit at the output to verify the 4 A value.

Variations for 50 V

The approach described here can be modified for 30 to 50-V output. Avatar (see Note 2) can supply a model AV-576 transformer rated for 45 V, 7 A. C1 will need to be 10,000 µF, 100 V; 70 V is not quite safe enough. Zener diodes D7 and D8 (but not D9) should have a 30-V rating (1N5363B). The reference voltage should be raised to 25 V (R11 = 1180 Ω, 1%) so that Q6 and Q8 will still be well within their voltage ratings. For a 30 to 50-V output range, R1 = 1.25 kΩ and R3 = 3.75 kΩ are correct if the reference is 25 V and if R2 is 2.5 kΩ. Q9 should operate at about 55 V.

Notes

[1]There is a TI TL783C high-voltage adjustable regulator chip, but it is very difficult to procure for some reason that I could not determine. Wait a year or more.

[2]Avatar Magnetics, 240 Tamara Tr, Indianapolis, IN, 46217. Contact by letter.

[3]Dye and Granberg, *Radio Frequency Transistors*, Butterworth-Heinemann, 1993.

[4]Sabin and Schoenike, *Single-Sideband Systems and Circuits,* McGraw-Hill, 1987/1995; or *HF Radio Systems and Circuits*, Noble Publishing (Crestone), 1998.

[5]See any modern *ARRL Handbook* power-supply chapter for a discussion of heat-sink requirements.

[6]FAR Circuits sells PC boards for this project for $8 plus $1.50 shipping/handling. They accept credit-card payment for an additional $3. Their address is: FAR Circuits, 18N640 Field Ct, Dundee, IL 60118; tel/fax 847-836-9148; **www.cl.ais.net/farcir/**. ☐☐

RF

By Zack Lau, W1VT

Adding an Amplifier with a Transfer Switch

I recently added a linear 10-W amplifier to my 2304-MHz station and found the exercise a bit more complicated than expected. The simple textbook diagrams don't include all the little details involved. I did a bit of work with a dual-trace 'scope to make everything work the way I wanted it to—instead of just wiring it up and hoping for the best.

The first problem is getting the 2.3-GHz amplifier and switch—neither is particularly easy to find, especially if you insist on running everything from 12 V. Many relays and high-power transistors for this band prefer 28-V supplies. I found a like-new latching 12 V Dow-Key model 412 transfer relay at a flea market many years ago—it even had the indicator contacts. The amplifier was made using some exotic Fujitsu GaAs FETs found at a Microwave Update flea market. VHF and Microwave conferences are often good sources of hard-to-find parts—sellers will often make an effort to bring the exotic stuff if they think there will be many buyers. While the bias supply does require a negative voltage, that is easily obtained with a '555 timer chip and a few inexpensive parts.

The impedances of the unmatched FETs were very low, making them tough to match with low loss. I ended up using leaded chip capacitors and copper foil to match the devices. I also mounted the devices on aluminum-backed Rogers Duroid, which minimized losses between the circuit board ground planes and the source/mounting flange of the transistor. Fig 1 shows the FLL 100 transistor mounted between 50-Ω microstriplines. I used #2-56 stainless-steel screws to make short electrical connections to the aluminum ground plane.

The first exercise with the relay was determining exactly what the indicator contacts would do—could I use them as an interlock? Not really, as observations with a 'scope didn't indicate any deliberate sequencing. Ideally, the indicator would reliably indicate actual contact closure. A few milliseconds make a difference when dealing with an extremely fast and expensive solid-state device. I decided

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it was necessary to make an actual sequencer, in order to power-up the amplifier only when the RF transfer-switch contacts were closed. Otherwise, the amplifier might be powered-up while the input and output are connected together by the transfer relay, as shown in Fig 2A. I wouldn't count on the input of the transistor being able to handle 10 W of RF without damage.

Since the relay featured latching contacts, which only need a short pulse to switch states, I used a sequencer based on the National Semiconductor LM3914 bar-graph chip. This chip features a dot mode, which can be used to toggle the two relay coils with short pulses. It can also be set up as a bar graph, which is how most sequencers are wired. The schematic is shown in Fig 3. With a 10 µF timing capacitor for C3, a receive-to-transmit transition results in 15 and 30 millisecond pulses to the receive and transmit coils, respectively. On a transmit-to-receive transition, there are 20 and 60 millisecond pulses to the transmit and receive coils, respectively. The first pulses are actually redundant, since the relay is already in that state. An elegant circuit might omit them, but I decided this simple circuit was "close enough." The length of the pulses can be increased by increasing the value

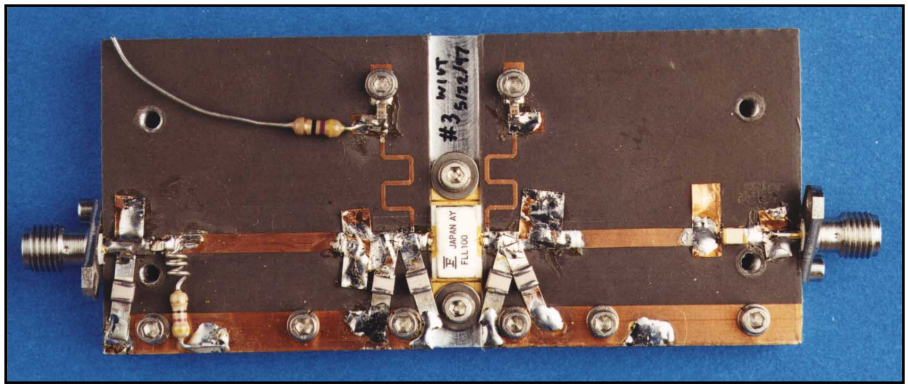


Fig 1—A photo of the amplifier circuit board.

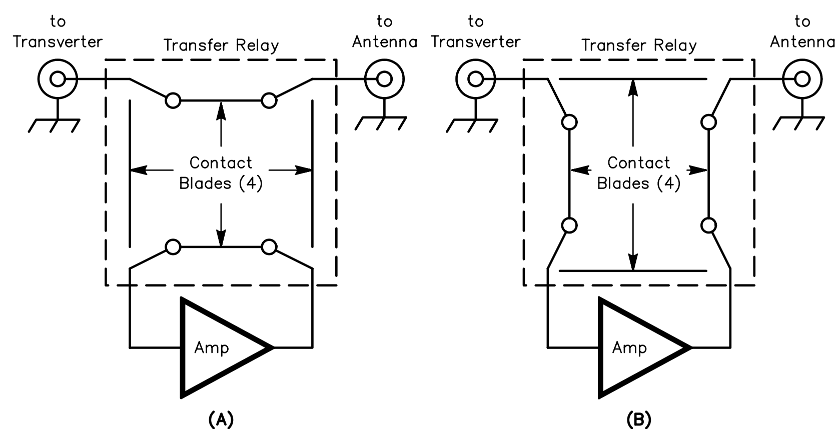


Fig 2—(A) Amplifier bypassed with a transfer relay. (B) Switching in an amplifier with a transfer relay.

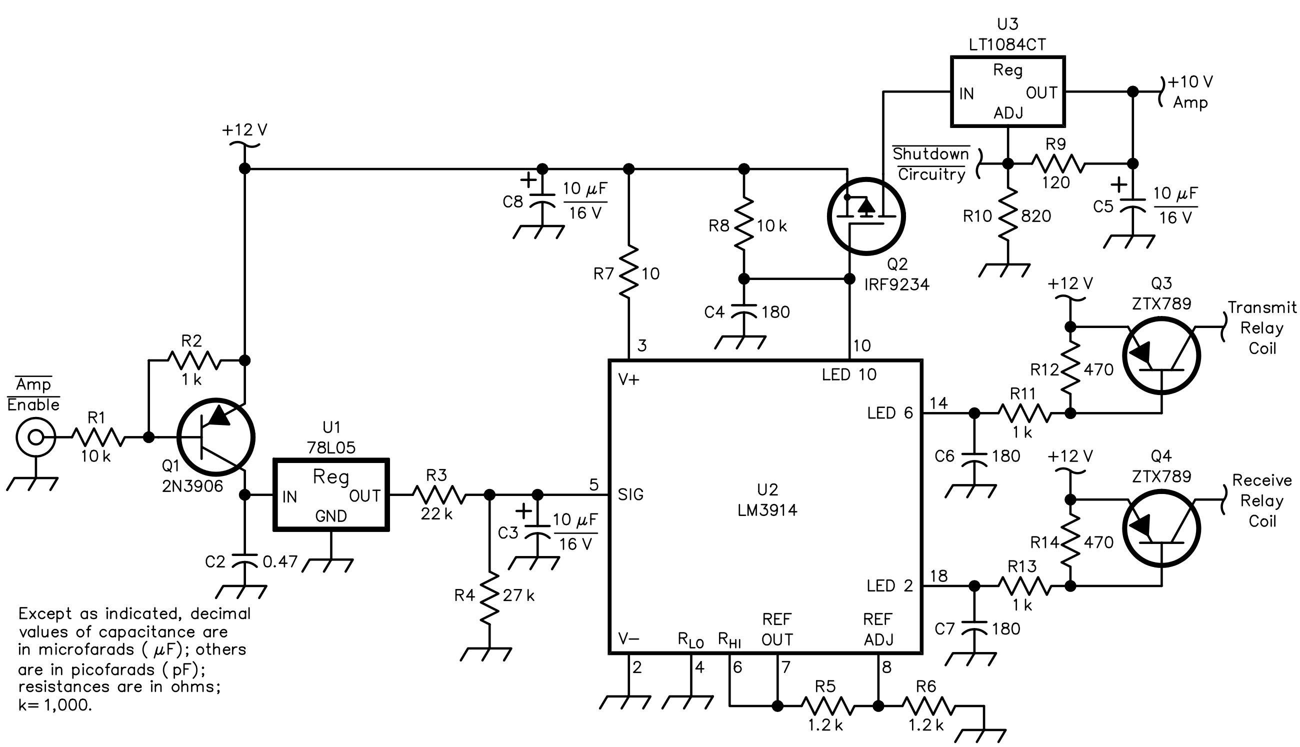


Fig 3—Transfer relay and amplifier sequencer.

of C3. However, if it takes too long for the relay to switch, the transverter might supply RF before the amplifier is ready. It may be necessary to increase the delays used in the transverter, to better accommodate the amplifier. A better solution may be to interlock the amplifier with the transverter, so RF is applied only after the amplifier is switched and biased properly. This could be done using the **+10 V AMP** signal to enable the transmit RF.

I'd actually prefer a fail-safe relay in this application. I'd wire it to bypass the amplifier unless power is applied. High-quality relays that work well can be hard to find, however. It makes sense to use whatever is available right now, instead of waiting to find the "perfect" parts.

Loss of a Wet N-Connector Junction

Ever wonder how much a little water in a coax connector can affect your signal? I did an experiment to find out. First, I measured the insertion loss of the dry coax, and then with a little water inside the connector junction (UG-21B/U and UG-23B/U). Then I added a lot of water. Next, I shook the water out of the connectors as best I could and remeasured the insertion loss. I then soaked both connectors, open ends down, inside a jar of water for 10 minutes and measured the insertion loss after mating the connectors. The connectors were finally shaken dry for a final set of reference measurements. Table 1 and Fig 4 show the results. The signal generator was a Marconi Instruments 2041 and the power meter was a Hewlett-Packard HP 435B/8481A.

Water can be a real problem—even a little bit can result in huge performance degradation at 2304 MHz—over 11 dB! Even at 2 meters, a wet connector can noticeably degrade signals. According to a study by Vern O Knudsen, the minimum-perceptible increase in loudness is about 0.4 dB.[1,2] The definition of 1 dB as the minimum-perceptible increase appears to be a "factoid," based on the lack of supporting studies.

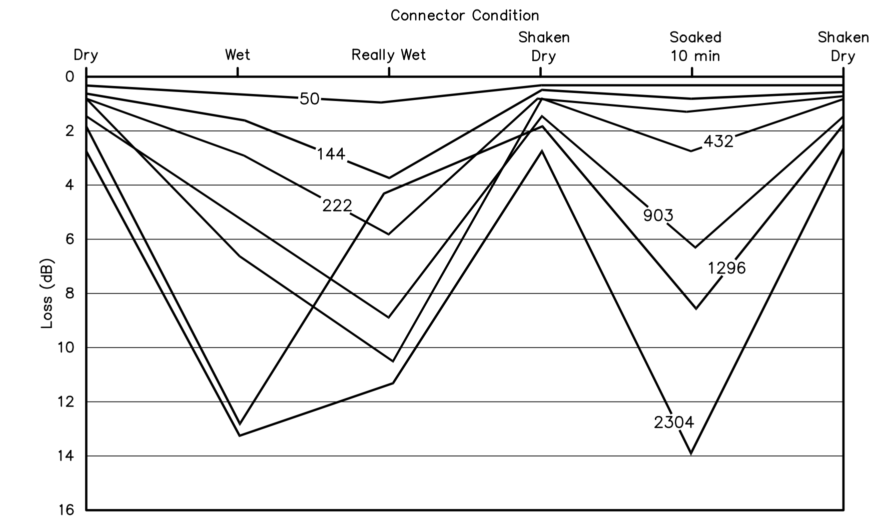


Fig 4—A chart of loss incurred in dry and a range of wet conditions.

Table 1—Insertion Loss of RG-213 Cables with a Wet N-Connector Junction

Cable lengths are 5′ 11″ and 9′ 10″ for a total of 15′ 9″ of RG-213 coax.

f (MHz)	*Dry(dB)*	*Wet (dB)*	*Really wet (dB)*	*Shaken Dry (dB)*	*Soaked 10 min (dB)*	*Shaken Dry (dB)*
50	0.2	0.5	0.9	0.2	0.2	0.2
144	0.4	1.5	3.6	0.4	0.6	0.4
222	0.6	2.7	5.9	0.6	1.0	0.6
432	0.7	6.6	10.4	0.7	2.7	0.7
903	1.3	—	9.0	1.3	6.1	1.3
1296	1.6	12.9	4.6	1.7	8.4	1.6
2304	2.6	13.2	11.2	2.7	14.0	2.6

Notes

[1]L. Stewart, *Acoustics*, D. Van Nostrand, 1930, pp 224-225.

[2]Knudsen, *Physics Review*, 21, 84 (1923). Study referenced in Stewart's book.

Next Issue in QEX

Among other features in the next *QEX*, we round out our collection of 6-meter power amplifiers with a unit from Richard Frey, K4XU. This time, it's solid-state, push-pull, class AB—an improved version of the author's earlier efforts using a pair of MOSFETs in the inexpensive TO-247 plastic package. Dick takes us through the design in systematic fashion and provides a parts list and PC-board layout.

Ron Barker, G4JNH, details his new work on the popular gamma match. He gives a brief history of this impedance transforming system and explains how widely accepted models of it didn't give good results at his location. His rigorous analysis exhibits unique insight into the operation of this renowned antenna feed system.

In addition, from Europe, where receiver dynamic range is critical, Ulrich Graf, DK4SX, addresses some requirements for the next generation of Amateur Radio receivers. He calls for significantly better performance than is generally available today and points manufacturers in a slightly different direction from that in which many are currently moving. He provides some interesting ideas and discussion in our forum.

Upcoming Technical Conferences

"Four Days in May" '99 QRP-ARCI Conference

QRP Amateur Radio Club, International (QRP-ARCI) proudly announces the fourth annual "Four Days In May" (FDIM) QRP conference commencing Thursday, May 13, 1999—the first of four festive days of 1999 Dayton Hamvention activities. Mark your calendar for this extra bonus day and register early for this not-to-be-missed QRP event of 1999. Amateur Radio QRP presentations, workshops and demonstrations will be the focus of the full-day Thursday QRP Symposium to be held at QRP ARCI headquarters—the Days Inn Dayton South. Papers to be presented include:

- "Vertical Antenna Design and Analysis," by L. B. Cebik, W4RNL
- "Constructing QRP Equipment," by Rev. George Dobbs, G3RJV
- "Design of a DSP-based Coherent CW Transceiver," by George Heron, N2APB
- "QRP Construction Tools and Tricks," by Dick Pascoe, G0BPS
- "Mixer Madness," by Clark Fishman, WA2UNN
- "PIC-based Rainbow SWR Bridge/Tuner," by Joe Everhart, N2CX
- "When Signals Go Wrong—Distortion Demystified," by Dave Benson, NN1G

FDIM QRP Symposium Registration

Registration for the Thursday, May 13, 1999 FDIM QRP Symposium is $10 if prepaid by May 1. Please send your $10 registration fee (US check, money order, international money order) made out to "QRP ARCI" and a SASE by May 1, 1999, to Philip Specht, K4PQC, 925 Saddle Ridge, Roswell, GA 30076 USA.

Awards Banquet Registration

The banquet will take place on Friday May 14, 1999. The ticket fee is $25, paid by a US check, money order or international money order made out to "QRP ARCI." Please send your payment and an SASE to Scott Rosenfeld, N7JI, (QRP ARCI Banquet Tickets, 2250 Paterson St 50, Eugene, OR 97405-2988, USA) by May 1, 1999.

QRP ARCI FDIM Headquarters

The Days Inn Dayton South (DIDS) will be the 1999 FDIM QRP headquarters. Hank Kohl, K8DD, has arranged a special block of reduced-rate rooms to be held at the hotel at $72/night (plus tax) with as many occupants as desired. Hank can be reached at QRP-ARCI Rooms, 1640 Henry, Port Huron, MI 48060-2523, USA. Full details of FDIM can be found on the QRP ARCI Homepage at **http://www.qrparci.org/fdim99.html**.

1999 Southeastern VHF Society Technical Conference

The third-annual Southeastern VHF Society conference will be held April 9th and 10th, 1999. We will again have preamp noise-figure testing, antenna-gain measurements, the technical program, a flea market, vendor displays, a family program activity, presentation of the K4UHF award, the banquet and door prizes.

The conference location is the Atlanta Marriott Northwest in Marietta, Georgia. The hotel and conference center is conveniently located northwest of Atlanta, Georgia, at exit 110 of I-75, Windy Hill Rd. The hotel rate will be $69 per room for single or double occupancy, if you make reservations before March 18, 1999 and mention the conference. For reservations, call Marriott at 1-800-228-9290.

The technical program will include presentations on EME, noise-figure measurements, linear amplifiers, VHF contest roving, transverters and interfacing among others. The banquet speaker will be Joel Harrison, W5ZN, ARRL Vice President and avid VHFer.

Antenna-gain measurements (2m and up, maximum of two per band, per person) and preamp noise-figure testing (50MHz to 24 GHz) will be done at the conference location before the technical program begins—so no one has to miss the presentations. Please register antennas and preamps in advance to ensure a testing time slot.

Please visit our Web site **http://www.svhfs.org/svhfs/**, e-mail to **k4sz@stc.net** or write SVHFS, PO Box 1255, Cornelia, GA 30531, for further information about the conference and a registration form, which includes a sign-up area for activities.

Letters to the Editor

A Concise Calculation Method for Pi-L Networks

WA6BAN:

◊ I just finished reading the Sept/Oct '98 issue of *QEX*. The article by Dr. Lickfeld, DL3FM, seems to have the wrong formula in paragraph 4, page 48. It should read:

$$R_1 = \frac{nV_p}{I_p} \tag{Eq 1}$$

Could this possibly be a translation error? I have never heard of n being anything but 2. The class-AB calculation would show much too low a load resistance for full power output. This is the second time R_1 has been published wrong lately. *Communications Quarterly* never corrected the 4CX1600 amplifier by W6FR.[1] I can send you my typical design calculations using a 4CX1600B tetrode amplifier if you would like them.—*73, William H. Sayer, WA6BAN, 25219 W Posey Dr, Hemet, CA 92544*

DL3FM:

◊ I received the copy of WA6BAN's letter, comprising a bad review of my article in the Sep/Oct '98 *QEX*. I am grateful to the letter's author. It will, I hope, help settle certain disagreements. I am busy working on a Pi-L amplifier, using principles corresponding closely to those in the article. This amplifier is proving to be a very difficult system. Much to my regret, I have not had the opportunity to see W6FR's article in *Communications Quarterly*, mentioned in the letter. I would be grateful to you if you would be so kind as to arrange that I get a copy of the design calculations using a 4CX1600B as W6BAN mentions.—*73, Prof. Dr. Karl G. Lickfeld, DL3FM, Rombecker Weg 71, D-45470 Mülheim an der Ruhr 1, Germany*

KF6DX:

◊ The formula appearing in Dr. Lickfeld's article is the same as shown in Chapter 13 of *The ARRL Handbook*:

$$R_o = \frac{V_p}{nI_p} \tag{Eq 2}$$

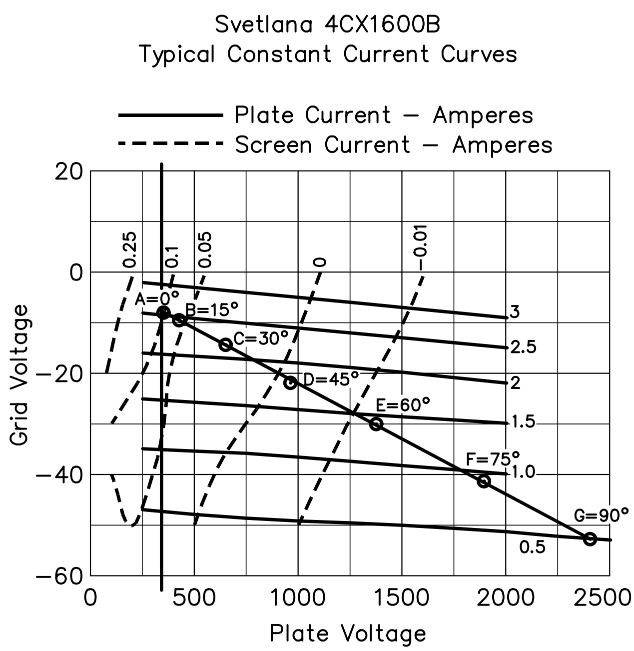
and the constant, n, is indeed selected according to the class of operation. It approximates the RMS-to-dc current ratio of the plate current. Since the conduction angle changes with class, it makes sense that the optimum load resistance also changes.[2] For comparison, below is the procedure for a 4CX1600B as outlined by WA6BAN. Many of the equations used can be found in W5FD's article "New and Improved Formulas for the Design of Pi and Pi-L Networks" in the Aug 1983 *QST*. Also, see "A Note on Pi-L Networks" in the Dec 1983 *QEX*. Some may think that Mr. Sayer's Eq 6 differs from the old-method equation for X_{C2} contained in a sidebar on page 13.6 of recent *Handbooks*. The type of analysis given on Mr. Sayer's Fig 1 stems from E. L. Chaffee's work in the 1930s.[3] The Eimac Performance Calculator (Eimac Division, Varian, San Carlos, CA; **www.eimac.com**) simplifies the determination of the points plotted on the constant-current curves. —*Doug Smith, KF6DX, QEX Editor;* **dsmith@arrl.org**

WA6BAN's Design Procedure for a 4CX1600B:

1. As the control grid in this tetrode has limited dissipation, we will run with little or no grid current—class AB.
2. An idle plate current of about 200 mA is a good starting point.
3. The plate-voltage excursion should not go below the dc screen voltage.
4. Draw the load line on the constant-current curves (see Fig 1). [Chaffee analysis.—*Ed.*]
5. Calculate the power output:

$$P_o = \frac{\left(\frac{V_{p\,pk}}{\sqrt{2}}\right)\left(\frac{I_{p\,pk}}{\sqrt{2}}\right)}{2} \tag{Eq 3}$$

As the instantaneous plate voltage and current are peak values, they



$$I_{dc} = \frac{0.5A + B + C + D + E + F}{12}$$

$$= \frac{1.3 + 2.4 + 2.2 + 1.8 + 1.4 + 0.8 + 0.5}{12} = 0.867\ A$$

$$Peak\ Fund = \frac{2.6 + 4.632 + 3.8 + 2.54 + 1.41 + 0.41}{12} = 1.283\ A$$

$$Power\ Output = \frac{1.283 \times 2100}{2} = 1347\ W$$

Fig 1—A Chaffey analysis of a 4CX1600B.

must be divided by $\sqrt{2}$ to get RMS. As this is a single-ended amplifier, divide by 2.

6. Calculate the load resistance desired by the tube:

$$R_o = \frac{V_{\mathrm{ppk}}^2}{2P_o} \qquad \text{(Eq 4)}$$

7. Next, calculate the reactance of the tuning capacitor:

$$X_{C1} = \frac{R_o}{Q_L} \qquad \text{(Eq 5)}$$

where $Q_L \approx 12$.

8. Calculate the reactance of the pi output capacitance:

$$X_{C2} = R_2 \sqrt{\frac{\frac{R_o}{R_2}}{Q_L^2 + 1 - \frac{R_o}{R_2}}} \qquad \text{(Eq 6)}$$

where $R_2 = 50\ \Omega$.

9. To get some idea for component requirements, calculate the tank-circuit circulating current. Change the parallel resistance to series:

$$Q_2 = \frac{X_{C2}}{R_{series}}$$

$$R_{series} = \frac{R_{parallel}}{Q_L^2 + 1} \qquad \text{(Eq 7)}$$

$$I_{RMS} = \sqrt{\frac{P_o}{R_{series}}}$$

10. Using the Eimac calculator, check the power output and the peak output current. I check:

$$R_o = \frac{2P_o}{I_{\mathrm{RMS}}^2} \qquad \text{(Eq 8)}$$

The figures should be somewhere in the ballpark. You may try changing some parameters, but I wouldn't allow any grid current and—despite some claims—I would not allow the plate voltage swing to go below screen voltage.

11. The average amateur builder won't have the correct parts—I did not for my 160-meter amplifier design using the Russian 4CX1600B. I use no screen bypass, so the plate-to-screen is one high-current supply, and the screen-to-cathode is a high-current supply for the overall high voltage.

My example is as follows: Plate-to-screen = 2260 V; screen-to-cathode = 297 V; grid bias = 56 V for $I_p = 0.2$ A.

The other calculations follow:

$$P_o = \frac{\left(\frac{2557-350}{\sqrt{2}}\right)\left(\frac{2.62-0.3}{\sqrt{2}}\right)}{2} \approx 1280\ \mathrm{W} \qquad \text{(Eq 9)}$$

$$R_o = \frac{2200^2}{(2)(1280)} \approx 1890\ \Omega \qquad \text{(Eq 10)}$$

$$X_{C1} = \frac{1}{(2\pi)\left(1.9\times10^6\right)\left(527\times10^{-12}\right)} \approx 159\ \Omega \qquad \text{(Eq 11)}$$

$$Q_L = \frac{1890}{X_{C1}} \approx 11.8 \qquad \text{(Eq 12)}$$

which is pretty close to 12.

$$X_{C2} = 50\sqrt{\frac{\frac{1890}{50}}{141.38-37.8}} \approx 30.89\ \Omega \qquad \text{(Eq 13)}$$

$$C2 = \frac{1}{(2\pi)\left(1.9\times10^6\right)(30.89)} \approx 2712\ pF \qquad \text{(Eq 14)}$$

Note: I had to add an **L** section because I only had a 2000 pF vacuum capacitor. The circulating current in the output network is found by:

$$I_{\mathrm{RMS}} = \sqrt{\frac{(P_o)\left(Q_2^2+1\right)}{50}} \approx 9.625\ A \qquad \text{(Eq 15)}$$

I used a two-tone test and had 1200 W with no apparent distortion. The Eimac calculator shows a little more power output and a lower R_o. However, Eimac gives $R_o = 1900\ \Omega$ for the 4CX1600B in a typical example. *—73, William Sayer, WA6BAN*

[1]M. Gonsior, W6FR, "Power on a Budget," *Communications Quarterly,* Winter 1995, pp 55-64.

[2]This equation entered the *Handbook* in the 1972 edition, having first appeared in an article by Irv Hoff, W6FFC ("Pi and Pi-L Network Design for Amplifiers," *QST,* Dec 1971, pp 34-37). Mr Hoff gives no reference for the equation, but states that it approximates the ac plate impedance, rather than the dc plate resistance that is commonly used. Hence, this equation is completely different from methods meant to determine dc plate-load resistance. Mssrs Lickfeld and Gonsior (see Note 1) have cited and used the equation correctly.

[3]"Simplified Harmonic Analysis," Rev. of Sci. Instr., Vol. 7, October 1936. The procedure is described in a current resource: *Single Sideband Systems and Circuits,* by W. Sabin and E. Schoenike, editors (McGraw-Hill: New York, 1995) in Chapter 14.

A Pair of 3CX800s for 6 Meters

◊ Congratulations on your design article on the 50-MHz amplifier in the Jan/Feb '99 *QEX*. I used your data to calculate the circulating RF current on your triode amplifier. My 4CX1600B amplifier at 50 MHz showed a circulating current of 13.22 A at 1280 W. I didn't have the constant-current curves for the 3CX800A7, so I couldn't calculate the peak RF plate current with the Eimac calculator. Thanks for doing a very good job on the amplifier design and testing.*—73, William Sayer, WA6BAN*

Handy Coil Winder; Compact Mobile Tuner

◊ When I find errors in the first three schematics I looked at in this (Jan/Feb '99) *QEX*, I wonder if the move away from a monthly publication was worth it. I'm not referring to the more complex schematics—I'm not well versed enough in what the designer had in mind to "follow the flow." I do know, however, a '7806 regulator (p 52) is not a '7608. From the caption on p 29, all taps on the tuner's inductors are on inductor L2. If Fig 1 is correct on page 29, then Fig 2 on the next page must be incorrect with respect to switch S2B. *QEX* is supposed to be a more technically oriented magazine than, say, *QST*, but these errors make me wonder. I suggest that perhaps you have someone run the schematic on a *PSPICE*-type program to see what it really does and make any necessary corrections to the schematic before publication. Waiting two months for a correction is obviously too long, and querying the author is undoubtedly an inconvenience to him. I know this doesn't happen too often, but if I were really itching to build a project from a bad schematic, it would certainly be disappointing if it didn't work. More so, if I went to the expense of having a PC board made for the project. Finally, my öersonal feeling about editing is that editors should work with the writers concerning clarity, relevance and the elements of style. Explanations at the beginning or end of an article, as you had on page 9, are acceptable. Editorial comments inside the text of someone else's article are distracting and sometimes seemingly demeaning. Unfortunately, these in-text editorial notes seem only to crop up in electronics-related magazines.

However, you are doing a good job overall. I just miss the good old days of 35 and 50-cent *QEX* cover prices.—*Very 73, C. H. Stewart, KD5DL, PO Box 181, Duncan, OK 73534*

◊ You are right about those errors you caught. We regret them, and our policy is to correct them when noted. Unlike *QST*, *QEX* is purely a forum for experimenters to share their work with each other. As such, the authors take some of the responsibility for the exactness of the material. We don't have the resources to build or simulate every design, nor would most readers want us to do so. We make every effort to ensure accuracy, but sometimes, things get by us. In any case, I think you'll find that authors are quite receptive to comments or inquiries about their work. Don't be afraid to write them, as you've written us! We work with authors to present their articles with clarity and impact. We try to avoid making editorial remarks inside their text unless it is necessary and expeditious.

Finally, do you not think *QEX* has made significant progress since 12 years ago, when the cover price was $1.75?—*Doug Smith, KF6DX,* QEX *Editor;* **dsmith@arrl.org**

About the *QEX* Web Site

◊ Today, I tried three times to subscribe to *QEX* via the Internet. Since my typing is not the greatest, I was sent back to a new form to be filled out again. Why can't you save the data already entered, as is done with some other Web pages, and save us all the reentering time? Sorry I lost patience, but I have to get on with other details for today. Happy New Year—*Bill Harris, W7KXB, 4410 E University Dr 104 304, Mesa, AZ 85205*

Your comments have been passed on to our Webmaster. I suspect the secure nature of the link has something to do with this.—Doug Smith, KF6DX, QEX *Editor;* **dsmith@arrl.org**

Tune SSB Automatically

◊ I really enjoy what *QEX* has become and look forward to its appearance in my mail! That said, it's on to the comments. I would have preferred a reference to the real software on the FTP site rather than spending 5.5 pages (*QEX* Jan/Feb '99, pages 13-18) on a listing of software titled: "Do not attempt to create working software from . . ." *QEX* pages are too valuable! More block diagramming or flowcharting of the software would have been more relevant. In the "Cheap Sweep" article, pp 54-55, I would consider it appropriate for the editor to remind readers that a sweep waveform can be derived easily electronically—with a '555 chip or counter and digital-to-analog converter, for instance. Continuous-rotation servo pots aren't all that cheap or available to many—not to mention stepper motors and drivers. (Yes, I do have some ball-bearing "wobbulator" capacitors in my stash of WW2 mechanically driven sweepers, but my stash goes back 50 years.) Whereas the author refers to other alternatives, he doesn't spell them out; and for the young guys who are learning, this would be good to do. In the Coil Winder (p 51) article, the motor is appropriately used and necessary! I use a small lathe with a magnet actuating a reed switch every turn—it actuates an electromechanical counter and is very simple; no electronics, just magnetic electromechanical. Anyhow, thanks so much for the diversity of interesting material. There are many ways to wind a coil or vary a frequency or write software (skinning cats is no longer allowed), and certainly part of our inspiration is seeing what others have done and considering what we would do! When I was 13 years old, I built a 6C4-807 transmitter exactly like the magazine article—my ham friends laughed at me—they'd never seen anyone *not* change the design somehow. It stung then and reminded me to ad lib. Best wishes for continued growth and success!—*73, S. Premena, AJ0J, POB 1038, Boulder, CO 80306-1038*

◊ Mr. Dick's software is available on the Internet site. I realize that most of us aren't familiar with this type of code, but the author made multiple references to it in his text. We thought it necessary to have it at hand for those interested. The idea behind "The Cheap Sweep" is that, for those without counters, timers, or ADCs, an alternative is available. A note about the unique nature of this sweeper might have been pertinent, but *QEX* articles stand mainly on their own "two legs." K0VXM's design is undoubtedly something many have thought about over the years. I heard from a few who belittled the use of a coil winder, but I know from personal experience that they are extremely useful—especially for those switching power supply transformers—when hundreds of turns are involved. I'm happy you liked the article.—*Doug Smith, KF6DX,* QEX *Editor;* **dsmith@arrl.org**

Standard Application Programming Interfaces

◊ This is a great idea whose time should have come long ago. I never heard of VA3LGD before reading the article (Jan/Feb '99, pages 19-21), but he has definitely zeroed in on an important matter. I hope the idea gets off the ground quickly and flies high!—*Bob Heath, W8BZ/VE3ADX, PO Box 467, Sault Ste Marie, MI 49783*

◊ The author has made an excellent case for our community to get started on the task of the easy interfacing of Amateur Radio equipment and computers. This standardization effort has been delayed far too long. Delay will only make the task more difficult to accomplish. Lawrence is to be lauded for his pushing and *QEX* for providing the exposure. However, I find that most of his case is focused on a Personal Computer Interface with the MicroSoft *Windows* concept as an illustration. This is shortsighted for the future that we cannot predict, but can expect. I suggest that the first focus be on all of our communication equipment and the external digital world. This includes not only the Personal Computer but, for example, the small processors such as Basic Stamp Modules and PIC Emulators. We seem to have come full circle in some 20 years to the days of the '4004, but we do not need to interface-program these new devices with binary or assembly development systems as we did in the 1970s. My next focus would then be on the radio and the PC as the author suggests in his example. This could be followed with other equipment interfaces. If properly accomplished, many new amateur enterprises would be developed, as has been the case for the PC. Ham applications would increase 10 to a hundredfold and the costs would decrease. I look forward to the day when all of our communication, control and test equipment can also be "Plug and Play."—*Warren L. Dowler, KE6LEA, 526 Camillo St, Sierra Madre, CA 91024-1402;* **dowler_w@compuserve.com**

◊ I agree that this idea is very important. That the League should lead the way in this area is crucial to its success, in my opinion. Any volunteers? Contact me directly via e-mail or Pony Express (see page 2 for US-mail addresses) and I will play referee until you get the ball rolling.—*Doug Smith, KF6DX,* QEX *Editor;* **dsmith@arrl.org**

Phase-Noise Measurement

◊ I am getting all kind of nice e-mail comments on my article. One of them by Stuart Rumley, KI6QP, mentioned that there is an error in the schematic in Fig 15. He is right and I apologize for not seeing it during the article review. In the schematic, there is a direct connection between the base of the BC149C transistor and the 400-µF decoupling capacitor. Instead of the direct connection, there must be a 68-kΩ resistor.—*73, Jos F. M. van der List, PA0JOZ, Fluitkruid 20, 2201 SM Noordwijk, Netherlands;* **jvdrlist@gironet.nl** □□

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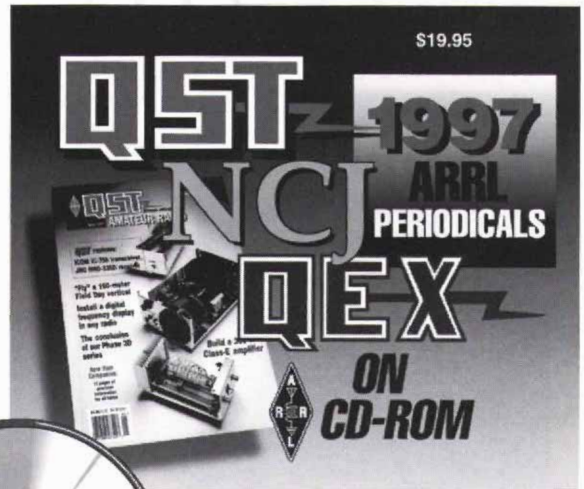

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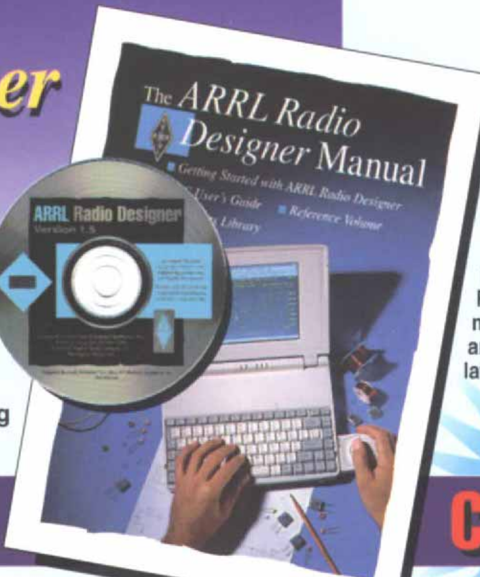

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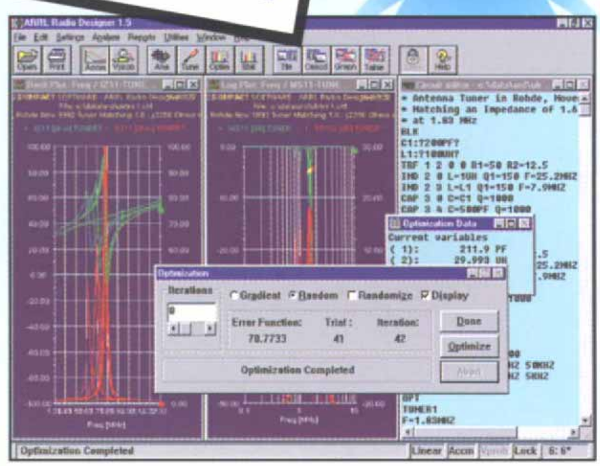

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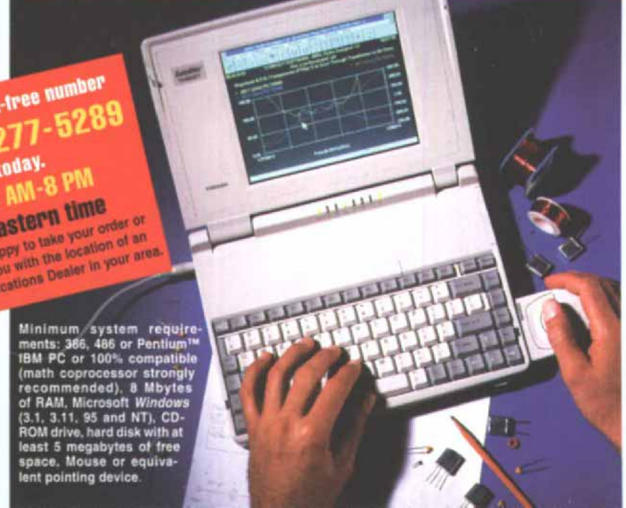

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