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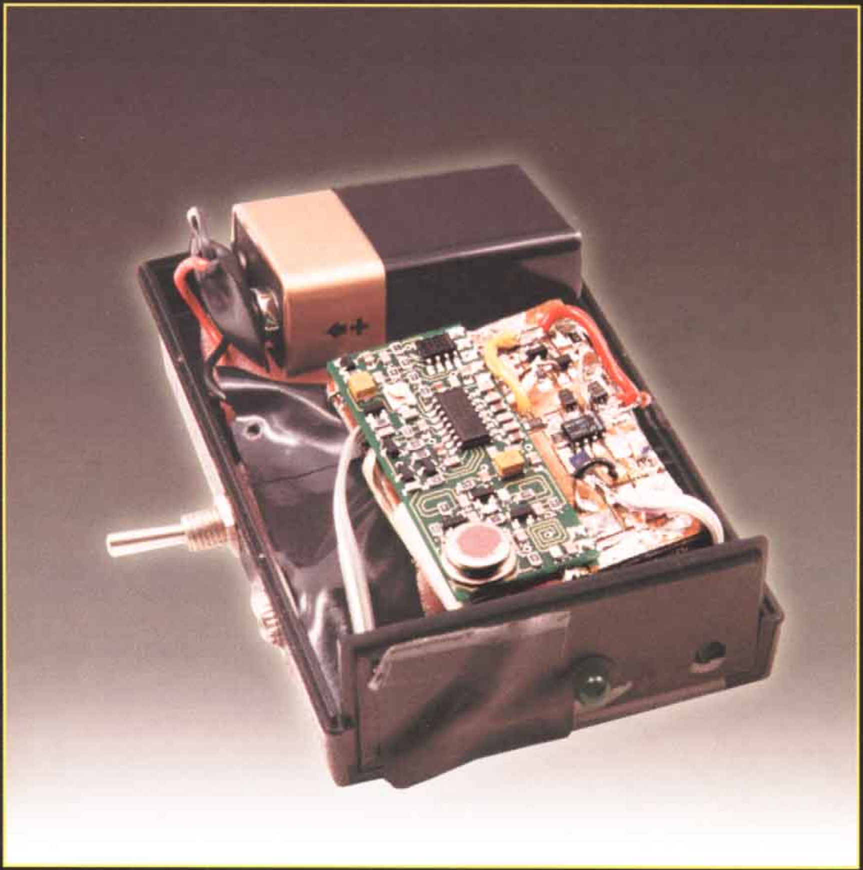
Forum for Communications Experimenters

November/December 2000



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QEX (ISSN: 0886-8093) is published bimonthly in January, March, May, July, September, and November by the American Radio Relay League, 225 Main Street, Newington CT 06111-1494. Yearly subscription rate to ARRL members is $22; nonmembers $34. Other rates are listed below. Periodicals postage paid at Hartford, CT and at additional mailing offices.

POSTMASTER: Form 3579 requested. Send address changes to: QEX, 225 Main St, Newington, CT 06111-1494

Issue No 203

Offices

225 Main St, Newington, CT 06111-1494 USA
Telephone: 860-594-0200
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Fax: 860-594-0259 (24 hour direct line)
e-mail: qex@arrl.org

Subscription rate for 6 issues:

In the US: ARRL Member $22, nonmember $34;

US, Canada and Mexico by First Class Mail: ARRL member $35, nonmember $47;

Elsewhere by Surface Mail (4-8 week delivery): ARRL member $27, nonmember $39;

Elsewhere by Airmail: ARRL member $55, nonmember $67.

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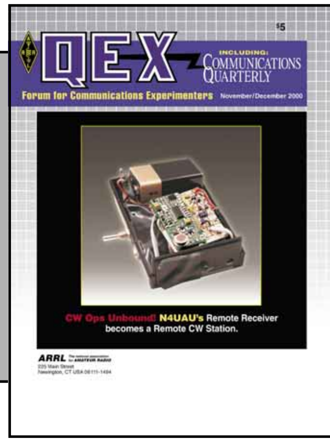

About the Cover

N4UAU's remote receiver from Part 2 of his series. Part 3 adds a keyer and multiplexing control; see p 37.

Features

Columns

Nov/Dec 2000 *QEX* Advertising Index

Empirically Speaking

In the telecommunications world, it seems that new technologies come along so rapidly that any one person has difficulty keeping up. As a group, radio amateurs should not have to worry about falling too far behind the power curve because we have some great magazines, Web sites and more communications gear than some large governments. What we see in the electronics trade press technically, though, is getting to be quite different from what we see in the amateur media. There may be many reasons for that, including a very different distribution of frequencies of interest to readers. For example, how many hams are working with 2.4-GHz gear in, say, New York City as opposed to the number of commercial links near that frequency?

Well, the commercial wizards are giving us the tools we need to jump into a lot of new technologies. Higher levels of circuit integration on chips and in subsystems are really making it easier for experimenters to try advanced technology. Amateur Radio has been and still is a wonderful proving ground for almost anything that potentially improves the way we communicate. *What* we communicate hasn't changed much in a while, but there may be new possibilities yet.

Some of you are undoubtedly wondering what happened to the next parts of two article series: John Stephensen, KD6OZH's "ATR-2000: A Homemade, High-Performance HF Transceiver" and my "Perceptual Transform Coding." We may have led you to believe they had disappeared forever, but the fact is that each author has had to take a hiatus to work on other projects. John indicates he will get cracking on his Part 3 soon. My Part 2 will appear sometime early next year. I want to have sample audio files (.WAV format) of coded and decoded voice available for downloading before we publish the article.

When I built my analog frequency compressor and coded a speech signal to reduce its bandwidth from 15-kHz to less than 4 kHz, the result was not quite what I expected. I thought the compressed signal would still be intelligible, but it is not. That may be a problem for Amateur Radio applications. I am still working on the decoder.

We have been lucky at *QEX* to get some outstanding pieces about state-of-the-art transceiver and antenna design, among other topics. Perhaps we should let you know some of what is coming in 2001, and make a call for articles on subjects that seem to need more coverage. First, we'd better tell you about what's inside our last issue of 2000.

In This Issue

Robert Brown, NM7M, has studied 160-meter propagation with great interest, and he gives us some results for a certain DX path on that band. Many effects that are found at those frequencies are not generally well understood. Some may be unknown to neophytes and others remain unexplained even by experts. Robert goes a fair way toward helping us understand them. R. P. Haviland's series on quad antennas continues (at last) with Part 3 and a treatment of multi-element designs. Tradeoffs among gain, beam width, impedance, front-to-side and front-to-back ratios are examined.

Rudy Severns, N6LF, takes a careful look at various types of antenna wire to see how the characteristics of different materials affect certain designs. He gives a concise explanation of skin effect and why it exists. I wasn't sure I remembered how to explain why it exists very well at all until I read Rudy's sidebar.

Warren Bruene, W5OLY, wrote a report almost four decades ago that, when combined with his updated prognosis, forms a very telling story of Amateur Radio. We publish them here in the hopes that future hams will read them and get a new set of bearings from the past. We find that new ideas and concepts are often rejected out of hand before they have a chance to come to fruition. Many do not bear fruit anyway, but it is interesting to measure our vision, both backward and forward in time.

Bill Sabin, W0IYH, discusses the properties and use of thermistors—resistors that change their value significantly over temperature. These

Continued on page 19

On the SSW Path and 160-Meter Propagation

How do the complexities of the ionosphere make 160-meter DX possible? Here's the story from an expert.

By Robert R. Brown, NM7M

It is well known that radio propagation takes place because of ionospheric refraction, reflection or scattering of RF waves. Of these three modes, refraction and reflection involve extremes of electron distributions. Thus, *refraction* takes place when the electron density varies slowly over a region having extent greater than a wavelength in all directions. *Reflection* involves distributions with strong gradients, the electron density changing rapidly in a region small compared with a wavelength. Scattering is between the two with regard to the distribution of ionization and often results when there is spatial structure in the ionization, as with field-aligned irregularities in auroral displays during magnetic activity.

The theory of electromagnetic-wave propagation by refraction goes back to Snell's Law, which in turn is based on Maxwell's Equations. It shows that signals deviate from straight lines in proportion to the electron-density gradients along a path. Thus, familiar earth-ionosphere hops in the vertical plane result from the vertical gradient in the distribution of ionization above the curved path, while lateral deviations from great-circle paths result from the gradients in electron density transverse (or perpendicular) to the great-circle direction.

Let's consider 160-metereter DXING, which is usually attempted during the hours of darkness. Then natural gradients of importance to propagation on that band are those below and above the nighttime E-region peak and around the sunrise/sunset portions of the terminator. In the D region, the ionization of electrons is caused by UV and X-rays in starlight, by galactic cosmic rays and by solar photons scattered into the dark hemisphere by the high atmosphere: the *geocorona*. As a result, VLF (3-30 kHz) signals are ducted back and forth between walls of a waveguide, whose walls are the ground and the lower D region.[1]

While wave reflections off the bottom of the D region may treated with a formalism similar to that used for metal mirrors, a complication

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[1]Notes appear on page 9.

arises because of the presence of the Earth's magnetic field. In particular, a rotation or shift of the plane of polarization may occur (see Note 1). For example, a wave incident at the bottom of the D region having its E vector in the vertical direction may, on reflection, still have a vertical component of polarization, but also a horizontal one, as well. Such *mode conversion*, as it is called, serves to change the wave polarization, particularly on long paths with many reflections.

Above the E-region peak, the vertical gradient at night is unique in that its sign or direction reverses across the bottom of an electron-density valley. The valley, shown in Fig 1, grows progressively deeper in the darkness, away from the terminator and contributes to stable ducting of signals in the vertical plane. In that same region, horizontal gradients in electron density are responsible for the horizontal deviation, or transverse skewing, of ray paths; but in contrast to the vertical gradient, the horizontal gradient is continuous in direction, with ionization always decreasing away from the terminator.

Unless specified otherwise, discussions of propagation assume wave refraction in an ionosphere that is parallel to the surface of the Earth; however, irregularities may be present that would tilt the ionospheric layers and reduce the scale of regions of constant electron density. That would break up the extent of coherent reradiation in the ionosphere and result in propagation closer to scattering, with waves going off in other directions than just the forward.

In addition, in disturbed, non-equilibrium conditions, other forms of ionization gradients may be present because of an influx of energetic electrons[2] during an aurora or because of their release from the Van Allen radiation belts[3] during major magnetic storms. Those gradients may be quite large, with horizontal structure as well as scales small compared to a wavelength and organized by the local magnetic field. In such circumstances, gradients may shift propagation from scattering to wave reflection by small, intensely ionized regions.

160-Meter Propagation

With regard to 160-metereter operations, transmitting antennas frequently use vertical polarization. Receiving antennas, on the other hand, are often chosen to focus on a limited range of directions instead of having the omni-directional features of verticals. Thus, the use of Beverage antennas has become very popular, as well as the use of smaller, directional antennas such those of the flag or pennant type.

In the last decade or so, considerable attention has been devoted to DXing on 160 meters. This has been heightened by the number of major DXpeditions to Asia that have included 160 meters, with XZ0A in Burma as a recent example. In this regard, a winter path from those regions has been singled out as quite productive for DXing from the USA, along the SSW sunrise terminator.

That differs markedly from the more conventional short path across the darkness at auroral and polar latitudes. An example is given in Fig 2, which shows that both the short path and the proposed path between the Far East and Midwest at the winter solstice. That type of path meets the terminator in the Indian Ocean area and then leaves it in the South Pacific area. It has received considerable attention[4] in DX circles, with the suggestion that 160-meter signals may be guided across great distances in darkness by the ionization inside the terminator. In addition, wave propagation in that direction has been considered as a form of long-path propagation, with signals going far greater distances than more-conventional short paths from DX targets.

That type of DX propagation, with signals guided for great distance along the terminator, has been widely accepted by DXers. While not said explicitly in so many words, the interpretation of the observations rests on wave refraction as a basis. But it has not been without criticism by the ionospheric community in the Amateur Radio circles because efforts using ray-tracing methods with the *PropLab Pro* program (Oler, 1994)[5] and both the CCIR and URSI model ionospheres[6] (IRI, 1990) have failed to verify its existence. In addition, it is not clear just what controls where signals become coupled to the terminator region nor how they become decoupled.

Failure of Wave Refraction

As noted above, terminator guidance of RF signals was studied initially by ray-tracing methods to see if it was possible to have 160-meter signals follow the direction of the terminator—say parallel to it but some distance away from it. While the possibilities are infinite, some tests showed that ray paths, starting close or parallel to the terminator would deviate away from it, going deeper into the dark hemisphere.

Those paths were obtained by using the *PropLab Pro* program, in three dimensions. That program solves the equation of motion that governs the advancement of a ray path, first introduced by Larmor in 1924:

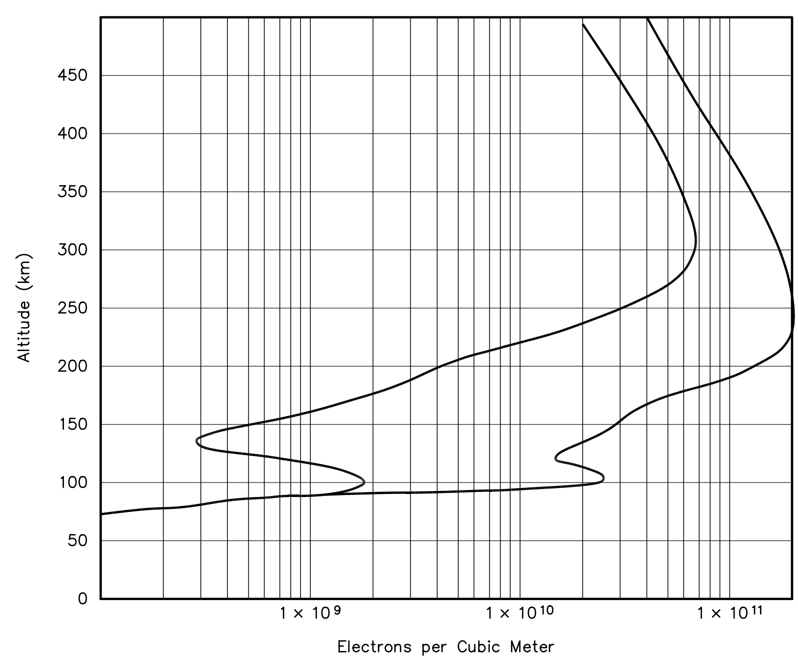

Fig 1—Electron density distributions for location near to (right) and far from (left) the terminator at night.

$$\frac{d\theta}{dS} = -\left(\frac{1}{n}\right)\frac{dn}{dl} \qquad \text{(Eq 1)}$$

This equation is the differential form of Snell's Law of refraction. It gives the rate of change of ray direction $d\theta/dS$ along the path S, in terms of the index of refraction, n, of the ionosphere and its spatial rate of change dn/dl in directions perpendicular to the path, in both the horizontal and vertical planes.

The search for a 160-meter path that advances horizontally by following the terminator resembles looking for a case of MF refraction in the lower ionosphere that is like HF propagation of Pederson rays in the high ionosphere. There, HF signals at the MUF may travel long distances and yet move along for awhile at a constant altitude, parallel to the Earth's surface. The HF case is possible around noon (see Note 1) as the electron density on the path goes through a maximum at the F-layer peak and gradients along the path are at a minimum. As a result, an unstable ray path forms just at the F-layer peak; however, the presence of any significant irregularity in electron density would disrupt its precarious equilibrium, because the gradients would not be in directions that could return to equilibrium.

In contrast to the preceding circumstances, an example of very stable wave propagation at lower altitudes is found in vertical ducting of MF signals in the density valley above the nighttime E region, shown in Fig 1. In this case, the gradients in the vertical plane are in opposite directions above and below the altitude of equilibrium of a ducted ray and in such directions as to return a ray toward equilibrium when perturbed.

With the MF case near the terminator, however, the horizontal variation of density does not go through any minimum; instead, it always decreases away from the terminator. So, there is only one type of gradient: one that refracts signals horizontally away from the ionization and only contributes to instability. Moreover, the rate of deviation on a signal path caused by density instability varies as the square of the wavelength, making the effect of a given gradient even greater at lower frequencies.

So at best, the ray-tracing search on 1.8 MHz amounted to looking for a location relative to the terminator where the horizontal deviation of a ray just matched the curvature of the terminator, but resulted in unstable propagation parallel to it. An estimate of the necessary curvature may be obtained by noting the changes in heading that go with a ray following the terminator half-way round the Earth: a change of p radians over a distance of 20,000 km or 1.57×10^{-4} radians/km.

With that curvature in mind and using the International Reference Ionospheres, the index of refraction and horizontal gradient of density were evaluated at a path altitude of 130 km in the midst of the terminator region in Fig 2. That rate of refraction would be achieved at about 300 km equator-ward of the terminator, as shown in Fig 3. Again, though, remember that would not be a stable path that could be maintained for any particular distance or time. That situation is extremely unstable since the electron-density gradient is always the same magnitude and from the same direction at any given

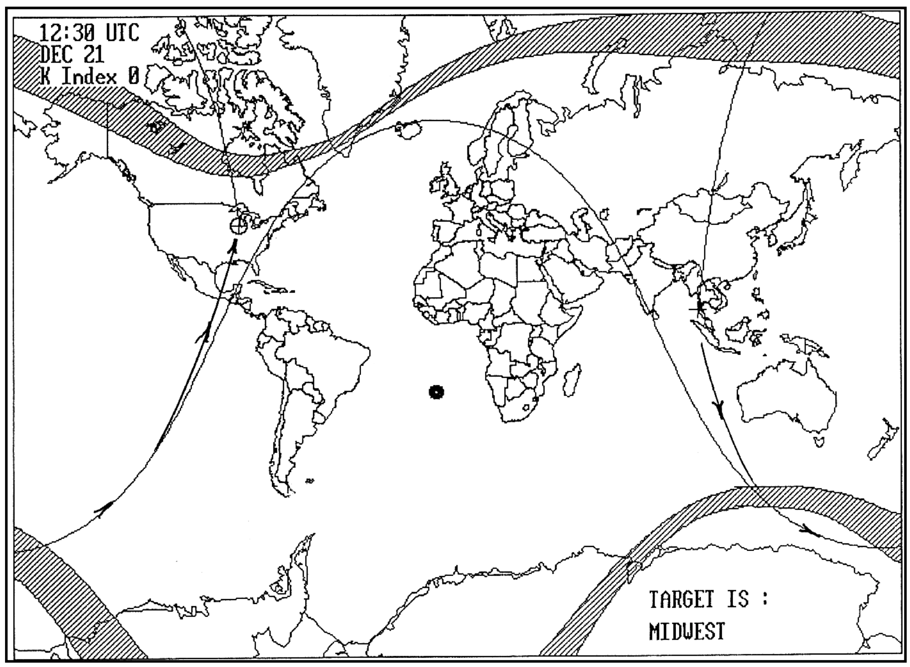

Fig 2—Short path across high latitudes from Burma to the Midwest at the winter solstice and a SSW path guided by the terminator.

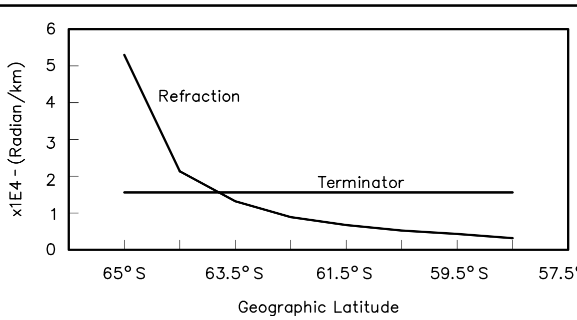

Fig 3—Rate of horizontal ray deviation by refraction in the local electron-density gradient at various latitudes, for SSW paths parallel to the terminator. The points are for the winter solstice and a longitude of 172°W. Note that each degree of latitude from the terminator corresponds to 111.11 km. The horizontal line is the rate of deviation of the terminator itself.

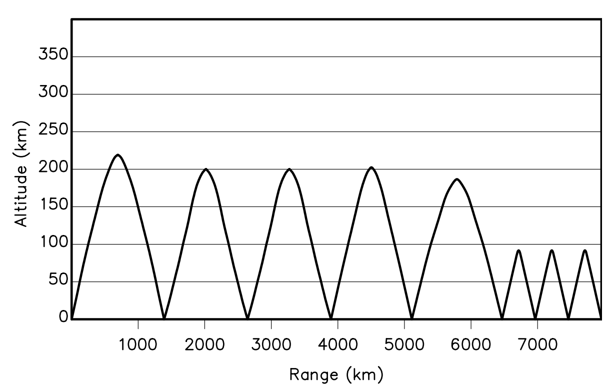

Fig 4—A two-dimensional ray trace of signals from the first part of the path in Fig 2 and launched at 20°.

distance along the terminator, and it does not reverse sign or direction about a minimum.

Ionospheric Absorption

Path stability in the face of disturbance is a matter related to ionospheric structure, but absorption relates to more-fundamental ionospheric processes, such as collisions of electrons with their surroundings. Like refraction, though, it varies with the square of the wavelength, or inversely with the square of frequency. That being the case, signals on the 160-meter band suffer the highest rate of absorption (in dB/km), particularly on E-hops. E-hops are always the case if a path nears the terminator.

The first leg of a path from Burma to the Midwest, shown in Fig 2, covers a distance of 8000 km to the terminator. Over at least 5000 km of that, the path consists essentially of F-hops, at least for radiation angles in the 15-25° range, which are typical of DX propagation. The remaining distance is in E-hops with heavy absorption. That is shown by the two-dimensional ray trace in Fig 4, for a launch angle of 20°.

Another view may be obtained by using the integral form of Snell's Law, as may be shown by a plot of the plasma frequencies in a vertical plane along the ionospheric path. In that formulation, signals are characterized by an equivalent vertical frequency (see Note 1) on entering the lower ionosphere. This is just like exploring the ionization distribution by vertical sounding. For a 1.8-MHz signal launched at 15°, the equivalent vertical frequency is 0.55 MHz; and in its vertical travel, the signal will never rise above the altitude of a 0.55-MHz iso-contour in the plasma-frequency plot. Similarly, signals launched at 25° have an equivalent vertical frequency of 0.82 MHz and are limited by their vertical excursions.

Fig 5 gives a plot of the transverse plasma frequency for the first 8000 km of the path from Burma to the terminator. Note that 1.8-MHz signals launched at 15° (on the left) will be limited to E-layer heights in going the last 3000 km to the terminator (on the right). Those launched at 25° will be limited similarly in going the last 1000 km. Those losses, along with any further travel, would reduce signals to unusable levels.

SSW Signals and an Alternative

The discussion above shows that signal propagation near the terminator is unstable when the ionosphere behaves according to current models. A general feature of those models is that ionization decreases steadily when going away from the terminator. That is another way of saying that there is no other recognized or accepted density distribution that would have a horizontal gradient in opposition to that of solar origin there. The question remains: How do signals propagate from DX and yet appear, around sunrise, to be coming from a SSW direction?

Since refraction by the known distribution fails and signals from the SSW are noted even in the absence of geophysical disturbance, wave *reflection*

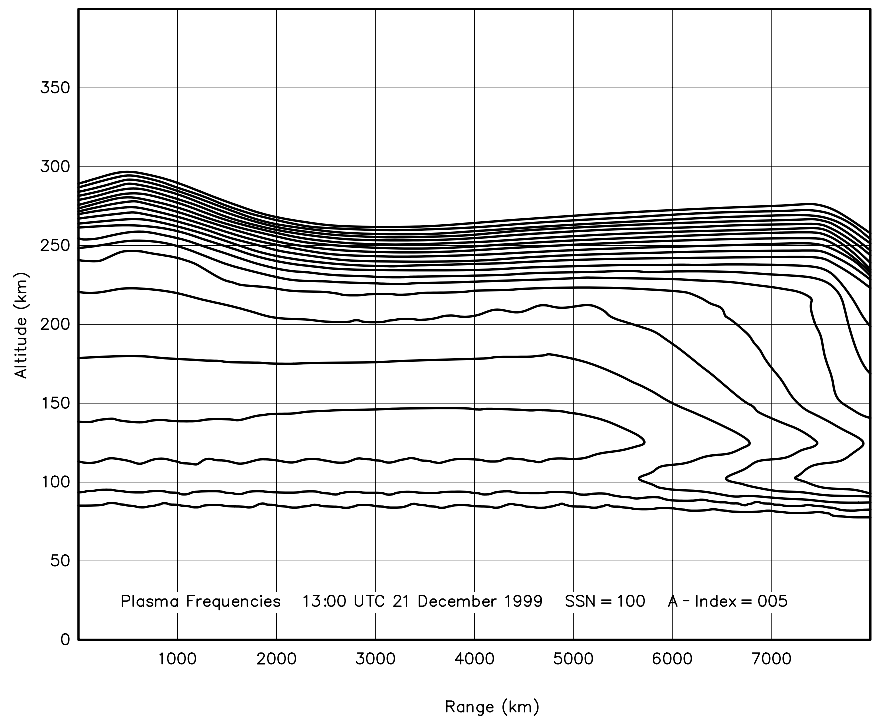

Fig 5—A transverse plasma-frequency plot for the first part of the path in Fig 2. This is essentially a cross-section of the ionosphere with the transmitter at the lower left and receiver at lower right.

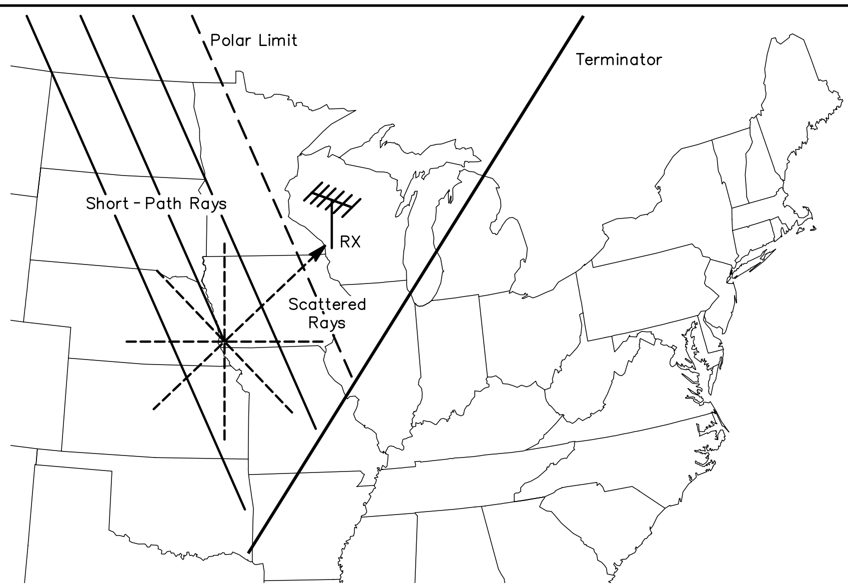

Fig 6—A representation of short-path rays (sold lines) which are scattered (heavy dashed lines) at high angles from one turbulent region. The lightly dashed ray shows the polar limit of the short-path rays and the heavy solid line shows the terminator. RX is the location of the receiver.

by steep gradients is ruled out and another mode must be sought. That leaves wave scattering as the only alternative. Fortunately, in contrast to refraction, scattering is capable of some large deviations in single-ray encounters, even with small scattering regions.[7]

The scattering region that serves to bring us 160-meter signals has its origin in systems that operate at sunrise. With turbulence, this region probably has quite a few small parcels of ionization—each less than a wavelength in extent—in the area off to the SSW, where the winter sun is already rising.

Those irregularities would carry ionization and perhaps reach the D and E regions to scatter incoming rays. Some of the scattered RF would reach the receiving station, as indicated in Fig 6, but would give the impression that the signals came from a false angle.

One way to study ionospheric irregularities during winter is by looking at turbulent aspects of the electrically neutral atmosphere. Motions of lower parts of the atmosphere and ionization are intimately connected as neutral constituents outnumber positive ions by far; with a high rate of neutral-ion collisions, ions are carried along by the motions of the neutrals (see Note 1). Electrons may follow positive ions because of electrostatic attraction.

Ozone Effects

It is possible to correlate the sunrise effects of atmospheric ozone with low-frequency propagation in the lower D region. There, the electron density gradient moves downward at dawn from about 95 km to 70 km, when solar UV begins to detach electrons from negative ions formed during the night (see Note 3). To detach these electrons, UV must reach the lower D region by passing through the ozone layer, which is somewhat opaque to UV. By noting the times of peak interference that result on suitable paths, it is possible to find when the gradient has lowered to about 80 km, midway between the limits for night and day.

Next, by noting times when the signal intensity is at minimum, one may use the data to find solar depression angles and heights of the ozone layer, when sunlight is just able to pass over the layer to reach the D region. For the case in Fig 7, the angle amounts to 3.3°; using 80 km as a reference, the height of the ozone layer is 69 km.

Now, the type of observations that give results as in Fig 7 have been carried out for well over two years using 55.5-kHz signals from NPG in Dixon, California, to Guemes Island, Washington. The basic data show how the height of the ozone layer varies throughout the year. The results sample the seasonal pattern: fairly stable across the summer months, but with considerable variation in height starting around the fall equinox, peaking around the winter solstice and then decaying toward the equinox again in the spring.

Within those variations, differences appear according to the basic weather pattern. Thus, Fig 8 shows the extremes of solar depression angles, week by week, throughout 1998 when the "El Niño" dominated the US. Fig 9 shows the extremes throughout 1999 when "La Niña" was over us. Please note that

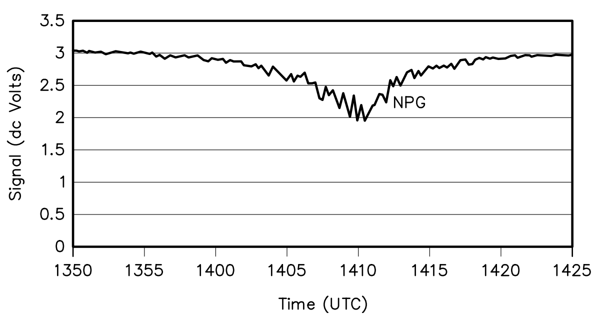

Fig 7—Slow time variation of NPG signals at sunrise.

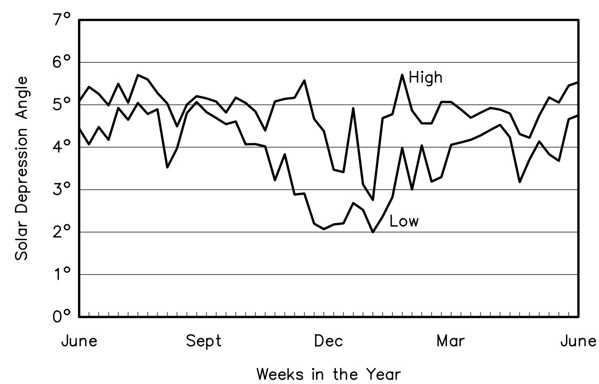

Fig 8—Solar depression angles for LF intensity minima centered on Dec 1998.

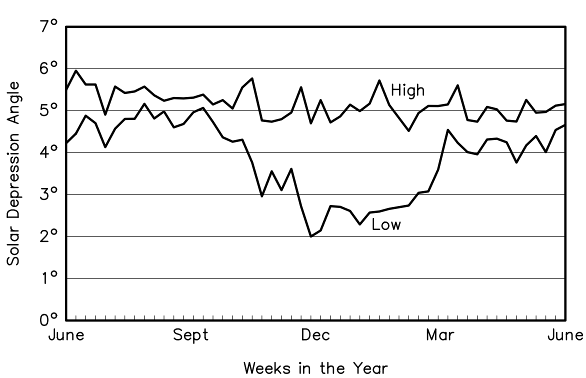

Fig 9—Solar depression angles for LF intensity minima centered on Dec 1999.

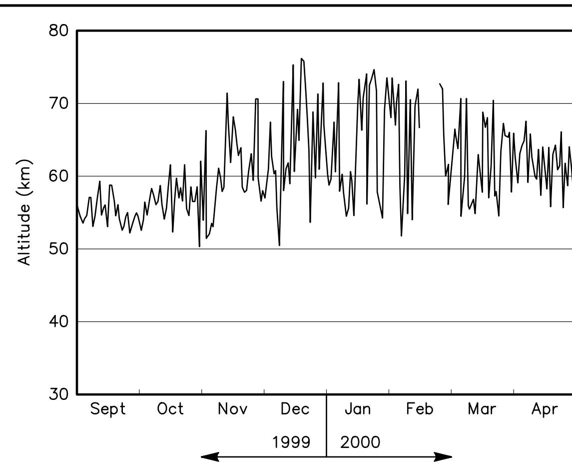

Fig 10—Heights of the ozone layer in winter, 1999.

"wrap-around" was used to center the figures on the winter months in 1999. Also note that the two sets of data are quite similar in their general features, but the El Niño pattern in 1998 shows far more activity in winter, with minimum depression angles separated less from maximum.

With the La Niña weather pattern, there were times when large-scale weather systems were infrequent. During one period, a large, high-pressure system over the Pacific Northwest lasted for almost a week. On that occasion, the ozone layer was thrust upward by about 20 km with the passage of the first front, then slowly it dropped downward over the remainder of the week.

In any event, Fig 10 shows that, day by day, the height of the ozone layer around sunrise in winter is quite variable, often rising or falling by 10-20 km. A theory of propagation of waves in a gravitational field[8] indicates that disturbances propagate, producing atmospheric displacements that increase in amplitude as they rise through regions of lower density. Changes seen in the ozone heights (50-70 km or so) point to even larger effects in the neutral atmosphere. Observed variability shows that the atmosphere is anything but calm; instead, this variability suggests that regions may have considerable horizontal structure and motion.

That spatial structure would be in the neutral regions, where ozone serves as a tracer, but it applies to ionization as well, as noted earlier. So, if the neutrals move up or down, east or west, north or south, ionization follows along and any 160-meter signals incident on the regions may then be propagated by wave scattering in the process. From that, it is only a small matter to have the short-path signals incident on the turbulent region off to the SW, with scattering then bringing signals to the receiver.

Discussion

Various degrees of detail were involved in the examining the question of whether 160-meter signals are guided along the direction of the termination. First, the ray-tracing method used not only current models of the ionosphere, like those from CCIR or URSI, but also a model of the earth's magnetic field, say the International Geomagnetic Reference Field.[9] In addition, the ray-tracing method goes through numerical integrations of the equations of motion for ray paths,[10] step-by-step, starting from the initial great-circle direction of the path. The results obtained by that method obviously depend also on the size and number of the steps taken in the numerical integration and the number of spatial dimensions used.

The next approach still involved using the equation of motion for a ray path but only for spot values of the rate of refraction or deviation of the ray path at locations near the terminator. That still involved the reference ionospheres and the geomagnetic field, but any questions about the uncertainties from the numerical integration methods were removed.

Lastly, the integral form of Snell's Law uses data from model ionospheres and shows locations that a ray path may reach through refraction, given its initial launch conditions. However, it does not deal with details along a path nor the signal intensity remaining after reaching any location. With the path under discussion, the integral form shows that any propagation or ray guidance that goes close to the terminator involves E hops and the heavy losses from the presence of low-lying ionization near that location.

Each of those arguments gave results that were against the idea of signals being guided by the terminator. Having made those points and noting that magnetically disturbed conditions are not suggested for the type of propagation, it falls to wave scattering to explain the observed propagation. That requires scattering centers and the turbulence at lower ionospheric altitudes in the winter months, as inferred from ozone recordings, provide the mechanism needed for propagation.

Having demonstrated guidance of signals along the terminator is not feasible for the distances involved, the question becomes "How does wave scattering contribute to the apparent propagation from the SSW." There, it was proposed that the path may be completed by signals being passed to the receiver by scattering of short-path signals that reach a nearby location: not the target station, but one in a turbulent condition resulting from sunrise.

To be more specific, for DX paths that go toward polar latitudes, there usually is a path that represents the high-latitude limit of propagation on a given day. That limit would be determined either by the flux of low-energy protons or solar electrons (polar drizzle) on the polar cap, or by field-line effects from the impact of the solar wind. The rest of the propagation paths from the DX transmitter will be within a broad swath of ray paths at lower latitudes and arrive in North America at more westerly longitudes. If those signals are to be heard in the US, they would have to be propagated by efficient modes, by either F hops, E-F hops or ducting. In any event, those paths would impact at ground level near the terminator as high-angle rays, because of the downward tilt of the F-layer in that region.[11]

However, the region will also be one in turbulence and signals reaching there will be scattered over a wide range of angles. Some scattered signals may reach sites that are east of the high-latitude limit but still in darkness, as shown in Fig 6. Therefore, without any signals coming in directly by short path, those signals would appear to represent DX propagation from the SSW. Nevertheless, other locations that lie within the swath of short-path rays coming from the north would not be aware that the signals also reached them by scattering.

With the advance of time, the sunrise terminator would move to the west. Being inclined to the north in winter, it would slowly cut off scattered signals from the SSW and move the scattering region northward, giving a "searchlight effect" for stations at fixed locations.

Conclusion

The present discussion has dealt with a SSW path from Asia that has been reported by a great many US DXers around sunrise in the winter. Given its direction, it is termed a "skewed path" and is even considered a form of "long-path" propagation. That interpretation, though, does not rest on any physical principle, only on its apparent geometry. When the physics of the path are examined, it is clear that the path fails in its effectiveness, being unstable in direction and subject to heavy, ionospheric absorption in the terminator region.

The alternate explanation offered here shows the path is not skewed to follow the terminator as a result of large-scale ionospheric structure; rather, it is skewed by local processes involving neutral constituents. Scattering is obtained toward the receiver by turbulence at sunrise in a nearby region of the atmosphere. Thus, the path is made up of two parts: one by refractions along a short path, and the other by scattering of short-path signals at high angles.

This article constitutes a rebuttal of an idea that is more talked about than documented, at least in any sort of detail. Aside from the article by Tippett, the rest of the story about the SSW path is found in anecdotal DX reports now archived in the "Top-Band Reflector," and not drawn together or published in any systematic form.

It is clear that atmospheric dynamics play an important role in creating scattering centers for 160-meter propagation and DXing. Discussion of this is new. Looking at the DX reports as they come in, the "fingerprints" of weather phenomena are found everywhere, from day-to-day variability of conditions to those DXpedition stations noting a "searchlight" effect, wherein their contacts advancing north to south with sunrise. DXers also find that the SSW path itself gradually moves toward the NW as sunrise passes.

Finally, there are reports of a SSE path being effective from the USA around sunset, although these are not well documented either. Those reports are far fewer in number than those of the SSW path, but they do exist. No discussion of those reports will be given here, as information is not available on scattering regions from the present, indirect methods. In the early evening hours, no ozone motion is evident, as there are practically no negative ions in the ionosphere at those times.

Degrees of turbulence at sunset and sunrise may be different according to latitude and variations observed here might not be the same as in other climes. This matter is left open to further discussion. Nevertheless, the linkage of ionization to neutral-atom motion is well known in physics and would certainly be present, no matter when or where.

Notes

[1]K. Davies, *Ionospheric Radio*, Peter Perigrinus, Ltd, London, 1990.

[2]R. C. Luetzelschwab, "Skewed Paths to Europe on the Low Bands," *CQ*, Aug 1999, pp 11-18.

[3]R. R. Brown, "Unusual Low-Frequency Signal Propagation at Sunrise," *Communications Quarterly*, Spring 1998, p 67.

[4]W. Tippett, "Long-Path and Skewed-Path Propagation on the Lower Frequencies," *1991 Proceeding of Fine Tuning*. Also in the *Top-Band Anthology*, Vol 1, Western Washington DX Club, 1998.

[5]C. Oler, "PROLAB PRO, High Frequency Ionospheric Signal Analyst," *Solar Terrestrial Dispatch*, Stirling, Alberta, Canada, 1994. For more information about the program, go to **http://solar.spacew.com/solar/proplab.html**.

[6]D. Bilitza, Ed., *International Reference Ionosphere* (IRI 90), National Space Data Center, Greenbelt, Maryland.

[7]J. Rottger, "On the Influence of Spread-F on HF Radio Systems," *AGARD Conference Proceedings, No. 173*, NATO, 1976.

[8]M. L. Salby, *Fundamentals of Atmospheric Physics*, Academic Press, New York, 1996.

[9]"International Geomagnetic Reference Field," *Utility Programs for Geomagnetic Fields*, National Geophysical Data Center, Boulder, CO, 1996.

[10]K. G. Budden, *The Propagation of Radio Waves*, Cambridge University Press, 1985.

[11]N. Hall-Patch, "Some Thoughts on Sunrise Enhancements," *Low-Band Anthology*, Western Washington DX Club, 1999.

Bob Brown was first licensed as W6PDN in 1937. He held KA6PTT, N7DGZ before his present call. He holds a doctorate in physics from the University of California. He has been an instructor of physics at Princeton University and a professor at the University of California at Berkeley. He retired in 1982.

Bob has done research on high-energy cosmic rays, solar proton and auroral electron bombardment of the polar atmosphere, ionospheric absorption at high latitudes and auroral X-rays in conjugate regions of the geomagnetic field. He has published over 80 scientific papers in refereed journals from 1944 through 1982 and a review article, "Electron Precipitation in the Auroral Zone," in Space Science Reviews (1966).

He has conducted several propagation columns in Amateur Radio publications: "Propagation and DX" in QRP Quarterly, "Over the Horizon" in The Canadian Amateur and "Propagation" in Worldradio. He has written Long-Path Propagation (published privately in 1992) and The Little Pistol's Guide to HF Propagation (Worldradio Books, 1995). Bob's current interest is the role of atmospheric effects in 160-meter propagation.

The Quad Antenna Revisited, Pt 3: Multi-Element Quads

Do you like quads? Do you like antenna modeling? Come explore HF quads having from 3 to 12 elements.

By R. P. Haviland, W4MB

The study of two-element quads in Part 2 of this series[1] showed that:

- Elements larger than self-resonance act as reflectors; they may produce main-lobe gains of up to about 8 dB and front-to-back (F/B) lobe ratios of 20 dB or so.
- Elements smaller than self-resonance act as directors; they can produce nearly as much main-lobe gain as the reflectors, but much lower F/B ratios.
- F/B ratio varies with element spacing.
- Points of maximum gain and maximum F/B ratio differ in frequency in all designs.
- Driven-element drive resistance is typically 150-200 Ω for good-performance designs, but tends to much lower values for maximum-gain designs.
- Matching to common transmitter output impedances is required in most designs, so there is no real benefit in designing the radiator for zero reactance. This leads to the concept of the "exciter quad," with reflector and radiator the same size, used as the basis of multi-element designs.

The idea of combining reflectors and radiators with radiators to improve performance appears to have originated with S. Uda in Japan shortly before WW II; however, the system was first described in the USA by H. Yagi,[2] so such antennas were first called "Yagis" and later "Yagi-Uda" antennas. All used dipoles as elements, but the principles work equally well for the loops of the quad family.

[1]Notes appear on page 18.

1035 Green Acres Cir, N
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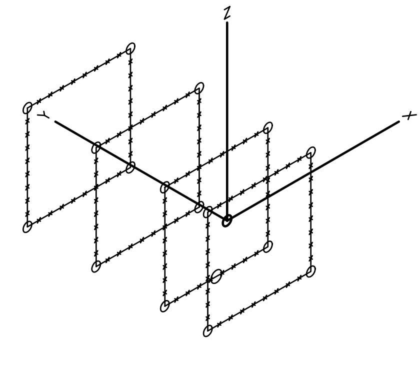

Fig 1—A typical four-element quad.

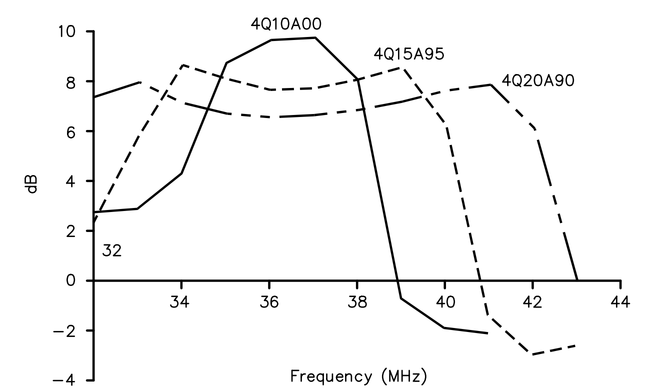

Fig 2—Gain of three versions of a four-element quad (4QUAD) with a reflector spacing of 0.15λ.

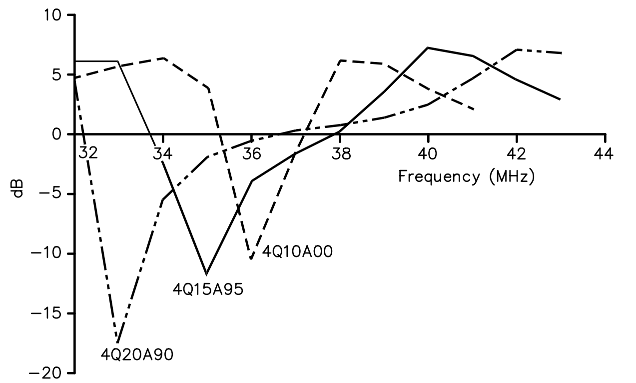

Fig 3—Back-lobe strength in the 4QUAD versions.

General Characteristics of Multi-Element Quads

The four-element quad illustrated in Fig 1 serves to illustrate the "degrees of freedom" in the design of multi-element quads that combine directors and reflectors and their resulting performance. In addition to providing a useful family of designs, these can serve as basis for still larger designs.

Fig 2 shows the variation of main-lobe gain with frequency for three combinations of reflector and director sizes. The family uses the pairing of reflector- and director-loop perimeters of 1.2-0.90, 1.15-0.95 and 1.10-1.00 λ; in each case, the radiator is the same size as the reflector and both directors are the same size. As is found to give good performance in two-element design, the radiator-reflector spacing is set at 0.15 λ and the radiator-director at 0.3 λ. The second director is 0.3 λ from the first. The direction of radiator to director is taken to be the "fore-lobe" or "gain" direction.

It is quite evident from the figure that both gain and bandwidth-of-gain are determined by the pair choice. The lower-frequency end is controlled by the reflector and the higher-frequency end by the director. Gain bandwidth is a little greater than the frequency separation of the parasitic elements. As the pair values approach each other, the peaks of around 6-8 dB (essentially the gain of a radiator and one parasitic element) move together, increasing gain overall.

The change in the back lobe with frequency is much different, as shown in Fig 3. The frequency of smallest back lobe is determined almost entirely by the reflector. At the extremes of the forward-lobe bandwidth, the back lobe can be as large or larger than the nominal lobe.

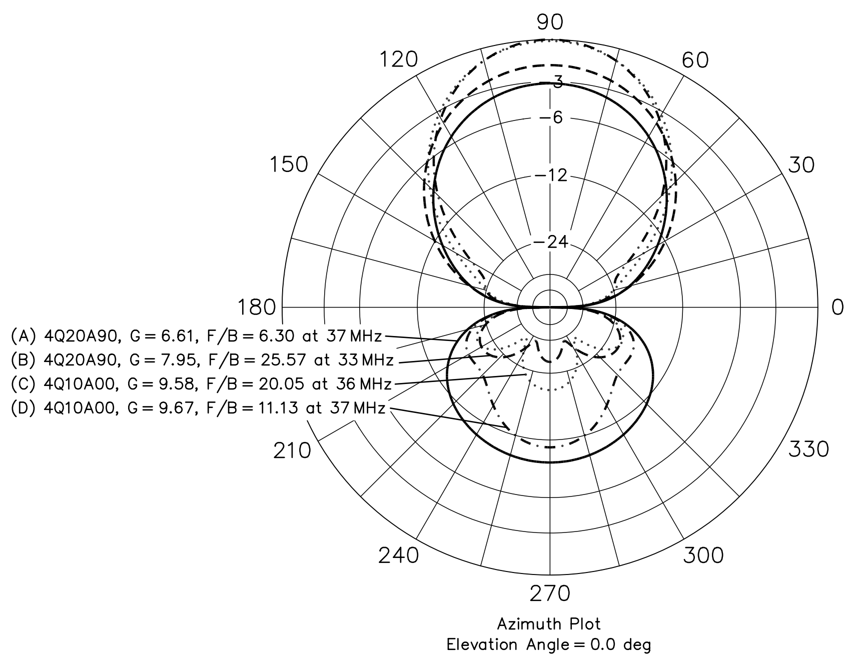

Fig 4—H-plane patterns for the 4QUAD versions: (A) maximum-bandwidth design at 37 MHz; (B) maximum-bandwidth design at 33 MHz; (C) maximum-gain design at 36 MHz; (D) maximum-gain design at 37 MHz.

Four examples of the H-plane horizontally polarized pattern are shown in Fig 4. Fig 4A is for the greatest-bandwidth type (4Q20A90), at 37 MHz. Gain exceeds one S-unit and the F/B ratio is such that the back lobe is essentially at isotropic-antenna level. The back lobe is much smaller at lower frequencies, with a F/B ratio of about $5^1/_2$ S-units at 33 MHz, as shown at 4B. This would be an acceptable antenna to cover both the 10- and 11-meter bands, or for all of 80 meters.

Fig 4C is for the highest-gain design shown (4Q10A00), at the frequency of best F/B ratio (36 MHz). Gain approaches two S-units and the F/B ratio essentially four S-units. This would be a good, high-performance antenna for a relatively narrow band, such as 20 meters; however, the back-lobe strength increases rapidly with frequency change, as shown at 4D (37 MHz).

Fig 5 gives a better picture of the way

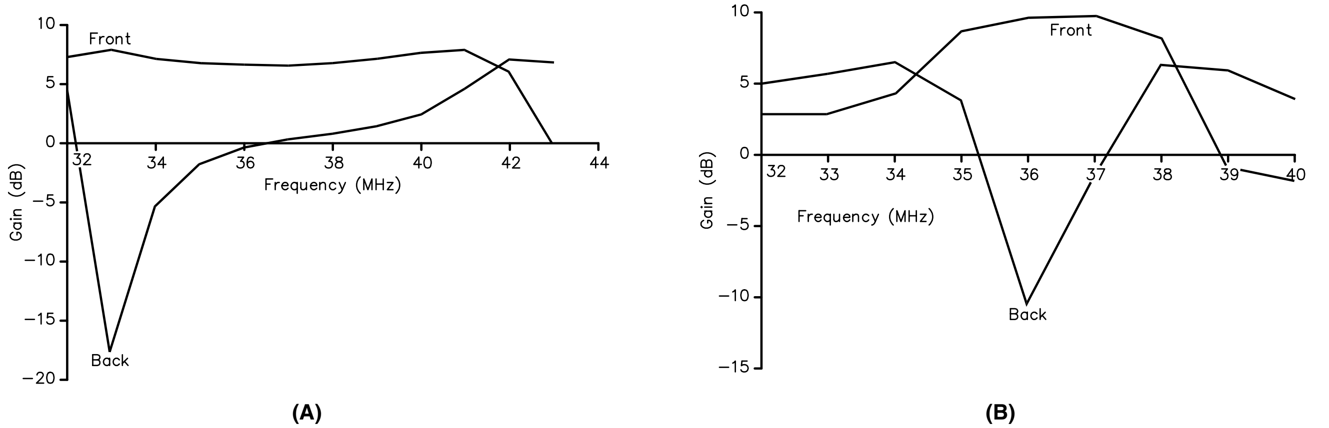

Fig 5—Comparison of front- and back-lobe gains: at A, maximum-bandwidth design 4Q20A90; at B, maximum-gain design 4Q10A00.

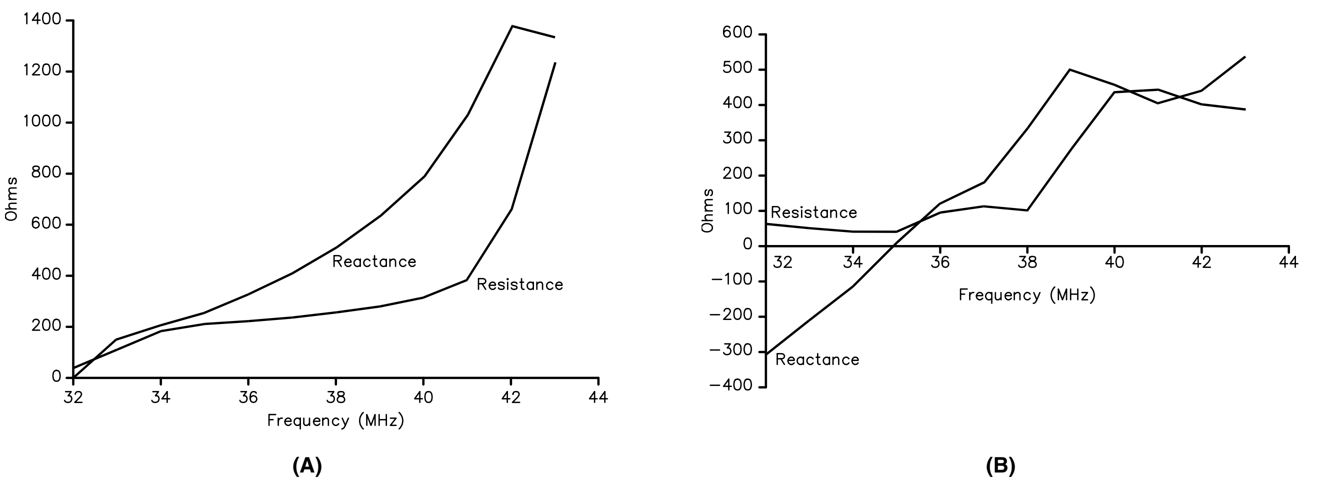

Fig 6—Drive impedances: (A) maximum-bandwidth design 4Q20A90; (B) maximum-gain design 4Q10A00.

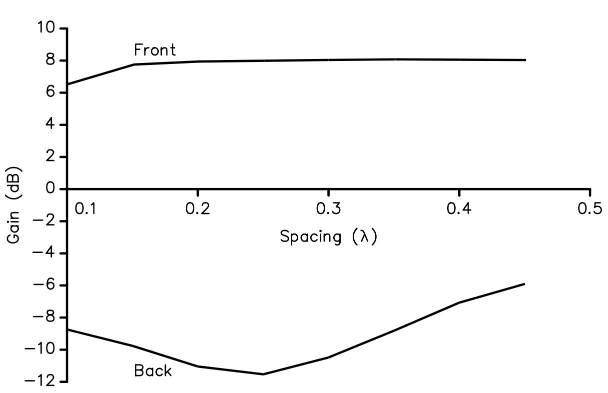

Fig 7—Effects of director spacing in a three-element quad (3QUAD).

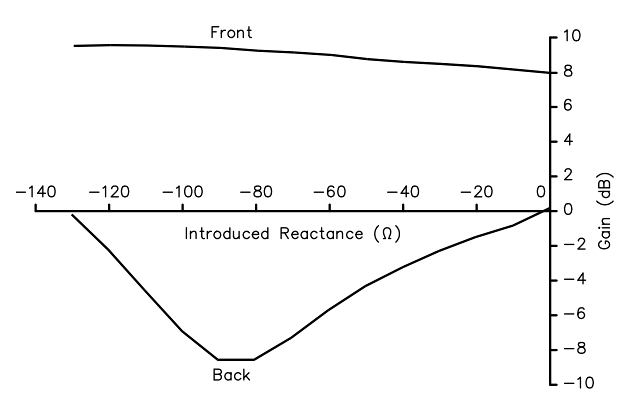

Fig 8—Effects of reflector tuning in 3QUAD.

the F/B ratio changes. The maximum-bandwidth design shown in Fig 5A has modest to poor F/B performance over much of its passband: It is good over only a narrow range near the low end. The maximum-gain type shows fair-to-good F/B performance over much of its passband.

Drive resistance of the maximum-gain and maximum-bandwidth type is shown in Fig 6. Over the range 34-40 MHz, the resistance of the maximum-bandwidth design is fairly constant, the reactance changes almost linearly with frequency. Both resistance and reactance vary widely in the maximum-gain design, even over the useful range of good gain. Single-frequency matching of either type is no problem, but a variable matching system (Transmatch) is needed to take advantage of the band-coverage possibilities. Multi-element fixed matching networks are possible, and would be required if the wide-band possibilities are to be used for wide-band signals.

Three-Element Quads

We may use the three-element quad to study the effects of varying element spacing and tuning on performance. The reference quad uses 0.15-λ and 0.3-λ spacing for reflector and director, 1.1 λ and 0.9 λ reflector and director peri-meter, and a wire-size (length/diameter) ratio of 10,000. The radiator is the same size as the reflector, and the frequency is 36 MHz.

Fig 7 shows the changes in lobe intensity as director spacing is varied. Gain is nearly constant, with maximum gain at spacings of 0.3-0.35 λ. The back lobe is smallest at a director spacing of 0.25 λ, but the variation is not great.

Rather than changing the size of elements when calculating performance, it is often convenient to "tune" the element by adding a series reactance, which is equivalent to introducing a tuning stub. Only a single entry in NEC is needed, instead of 16 changes in the element-dimension table. The element size-reactance curve of Part 1 (or Table 1 in this part) can then be used to calculate the equivalent change in element size.

Fig 8 shows the effects of reflector tuning (equivalent to changing length). Best F/B ratio occurs with 80-85 Ω of added capacitive reactance. This is equivalent to making the reflector 0.03 λ smaller. Maximum gain requires a still-smaller reflector, but the F/B ratio suffers.

Fig 9 shows the results of keeping the boom length constant at 0.45 λ, while changing the position of the radiator

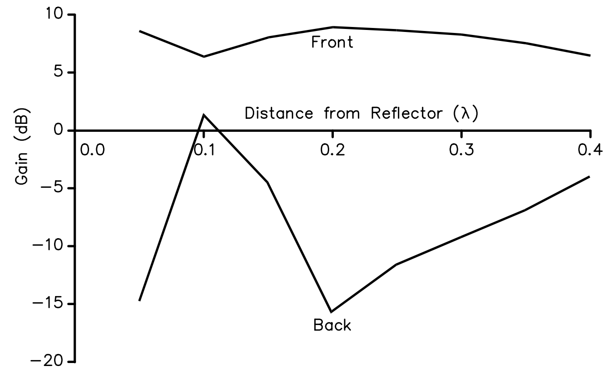

Fig 9—Effects of 3QUAD radiator location with a constant boom length.

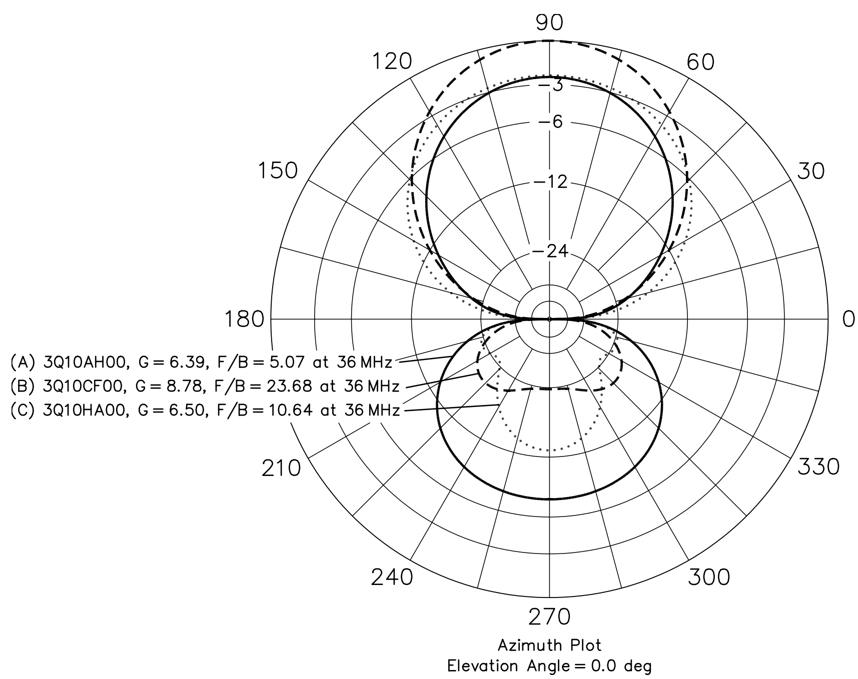

Fig 10—3QUAD H-plane patterns with a constant boom length: reflector-radiator spacing is 0.05 λ for A, 0.2 λ for B and 0.4 λ for C.

Table 1—Table of Reactance Change with Size Change

Conductor Size P/2A	*KB Change per 100 Ω of X*	*Reactance Change per 0.1 change in KB*
100	0.1108	90.25
200	0.0750	133.25
500	0.0570	175.50
1000	0.0484	206.25
3000	0.0391	255.75
10000	0.0360	305.75
30000	0.0282	354.25

Conductor size is element length/diameter.
KB change is in wavelengths.
Element reactance change is in ohms.
Values are valid for KB between 0.90 and 1.25.

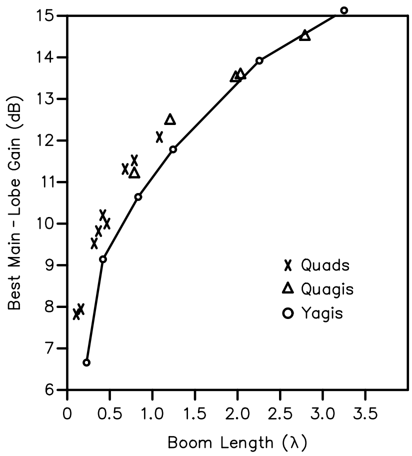

Fig 11—Expected maximum gain as a function of boom length.

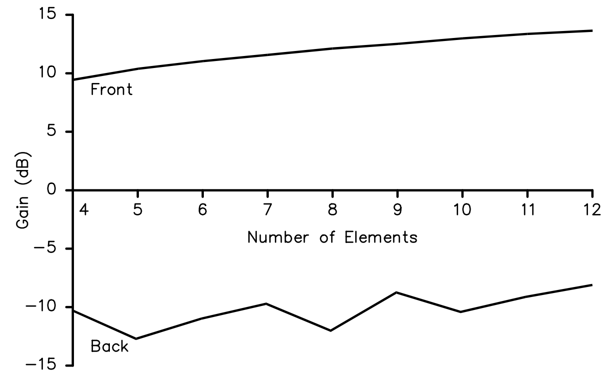

Fig 12—Lobe strengths for 4- to 12-element quads.

element. Both gain and F/B ratio are improved by moving the radiator toward the director by 0.05 λ, making the spacings 0.2 λ and 0.25 λ. Bandwidth does not change greatly, and drive-impedance changes are not great.

Fig 10 shows the pattern changes as the radiator location varies. The A pattern in Fig 10 is for a reflector spacing of 0.05 λ, B for 0.20 λ and C for 0.40 λ. The pattern is markedly better at B. Although detailed analyses are not included here, much of the adverse effects of a change in one parameter may be compensated by changing another parameter simultaneously. Nonetheless, the best balance among gain, F/B ratio, bandwidth and reasonable drive characteristics is at or near the dimensions shown at the start of this section.

Short-Boom Quads

Because of space problems (trees had priority over antennas), considerable computer and experimental time was spent on development of a short-boom, three-element 20-meter quad for use at W4MB. The boom length was limited to 20.5 feet. Results were mixed; some improvement in main-lobe gain was found. This was always accompanied by relatively poor F/B performance, though. This means, for example, that the evening long path to VK/ZL was accompanied by much stronger W6/W7 interference. It was found that a well-designed two-element quad gave better results overall. The moral: A good three-element quad for 20 meters needs a boom of 27 feet or so. Instead of trying to get high multi-element performance in a small space, look at fewer elements in the same space.

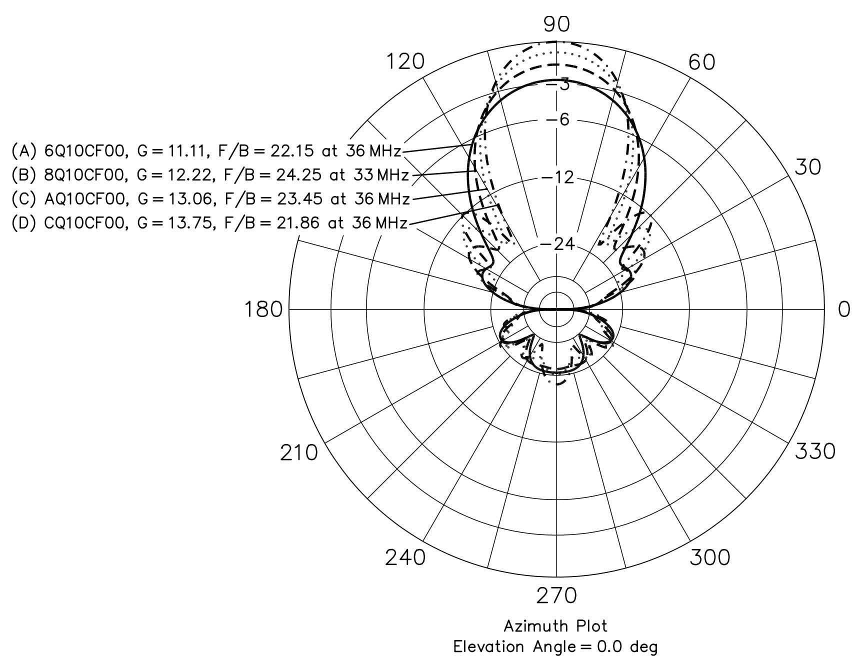

Fig 13—H-plane patterns for 6- to 12-element quads.

Big Quads

Extensive studies[3,4,5] have shown that the maximum gain of parasitic arrays is primarily determined by the length of the array, with a small effect by the number of elements. Fig 11 shows this effect. The solid line is taken from the comprehensive study of Yagi designs by Lawson. Individual points plotted are from my original study of quads and are *Mini-NEC* values. See also the curve in Reisert's article.[6]

Results of this study using NEC are plotted in Fig 12. The forward gain increases steadily, but the back-lobe size varies with the number of elements. If the spacing between elements were reduced, it would be found that both forward gain and back-lobe size vary with boom length. The variations are a consequence of reflections that occur at the ends of the array, which can either add or subtract, depending on array length (see the discussion in Lawson).

Horizontal-plane, horizontal-polarization patterns for four of these quads are shown in Figs 13A-D, being for 6, 8, 10 and 12 elements, respectively. This family shows the steady narrowing of the forward lobe that is required to increase gain. The back lobe stays at essentially the same level, but becomes more complex as the number of elements increases. Also noticeable is the growth of the first side lobe.

There is need for a standard way of comparing antennas with respect to interference-accepting characteristics. The simple F/B ratio (at 90° and 270° in these plots) may be misleading, since the 270° value can be a dimple in a larger lobe. This begins to be noticeable in Fig 4. A more complex method sometimes used includes:

- Main-lobe gain.
- Main-lobe beamwidth (at –3 dB, or sometimes –3 dB and –10 dB).
- Front to first side-lobe ratio.
- Average of all back lobes.

For the 12-element design, these values are 13.75 dB, 34°, 12 dB, and –25 dB, estimated.

The vertical-plane, H-polarized plot is shown in Fig 14, (A) for 6 elements and (B) for 12. The V-plane beamwidth is greater that of the H-plane. The vertically polarized component can be estimated as the lobe intensity at 0° and 180° in the H-plane plots, or 30 dB or more below the main lobe.

Fig 15 compares the lobe strengths for 8 and 12 elements. The bandwidth decreases with more elements: Since the added elements are directors, the bandwidth change is greater at higher frequencies.

Fig 16 shows the lobe structure of the 12-element quad for 34 (A), 36.5 (B) and 39 (C) MHz. At the highest and lowest frequencies, the forward gain is low and the back lobe large. As shown in Fig 15,

Why Revisit?

It has been nearly 20 years since I started to study the quad-antenna family in detail and just over 10 years since the first publication of my study results. In this time, antenna analysis has matured to give much-improved data; calculated values now come much closer to those found in tests. There has also been a great increase in computer capacity and speed: Evaluations that took 10 or 12 hours at the start of the study now take seconds. Some new techniques for transfer of antennas on "paper" to hardware have appeared that make it easier to get designed performance.

These factors made it seem desirable to revise the results of my studies, and the book *The Quad Antenna.* Many of the changes also appear in this series of articles.—*W4MB*

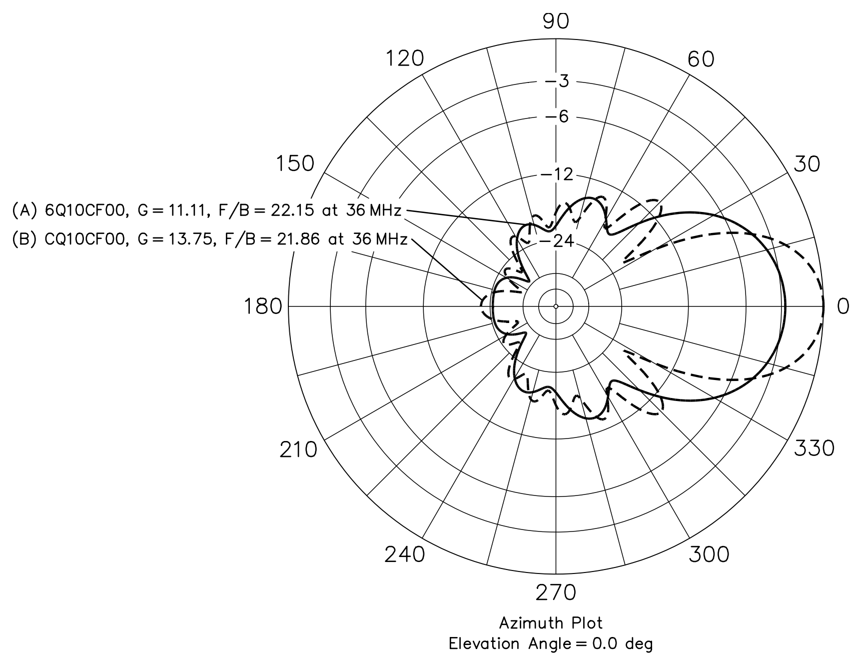

Fig 14—V-plane patterns, 6- and 12-element quads.

Note on Labels for Designs

If very much antenna work is being done, it is likely that a large number of computer analysis files will be generated. Unless these are well labeled, confusion results. For parasitic arrays, the I have adopted the following:

Label = *nTxxKLmm*

where:

n = number of elements (letters "a" through "f" are used to represent numbers 10 through 15)

T = Type of antenna or element, D-dipole, Q-quad, etc

xx = Reflector perimeter, (wavelengths – 1)

K = Reflector spacing, A-0.05, B-0.10, etc

L = Director spacing, A-0.05, B-0.10, etc

mm = Director perimeter, (wavelengths – 1)

A preceding or following letter indicates a variation, such as tapering of directors or variable spacings. The system conforms to the eight-letter needs of MSDOS and works with other computers.—*W4MB*

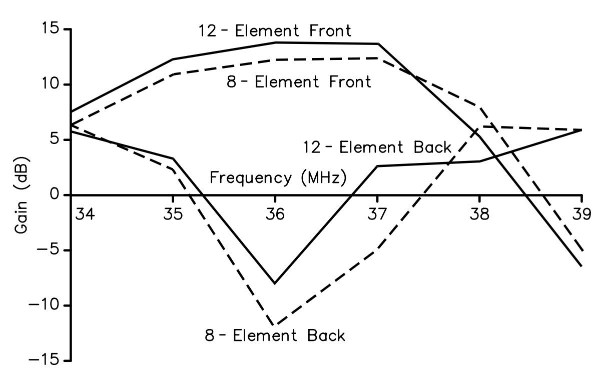

Fig 15—Lobe strengths for 8- and 12-element quads.

good gain and low F/B ratio occur in a relatively narrow bandwidth.

Current magnitude and phase are shown in Fig 17 for the current at the top of the elements. This is the reason that the driven element, No. 11, shows a phase of 180°. Drive at the bottom is the reference at 0°. Antennas this large act as travelling-wave antennas,[7] with the directors causing the wave velocity to be lower than in free space. In essence, the reflector, radiator and closest director form the exciter for the wave structure. While of interest, the theory of traveling-wave antennas doesn't lead to reasonable practical analysis, at least at present. It is much simpler to use NEC or one of its commercial versions for analytical work.

Optimizing Antennas

Usually, the word "optimum" means that the best value of some single performance variable is sought. It doesn't apply well to antennas, because a number of performance values are important. In antennas, "optimum" means the best balance among these. Further, because of the work involved, "optimum" in a practical sense means a reasonable approximation to this best balance.

In recent years, a formal method of looking for this balance has developed, called a "genetic algorithm" or GA.[8] It is actually a method of trial and error, coupled with a set of rules to select from a family of trial design values (an antenna in this case), plus another set of rules for evaluating the results of a trial and still another set for establishing another trial. Often, an element of chance is introduced to try non-obvious design possibilities. The actual calculation of performance may be done by any of the antenna-analysis programs available, typically with a core of NEC.

The general summary is that the technique works. There is nearly

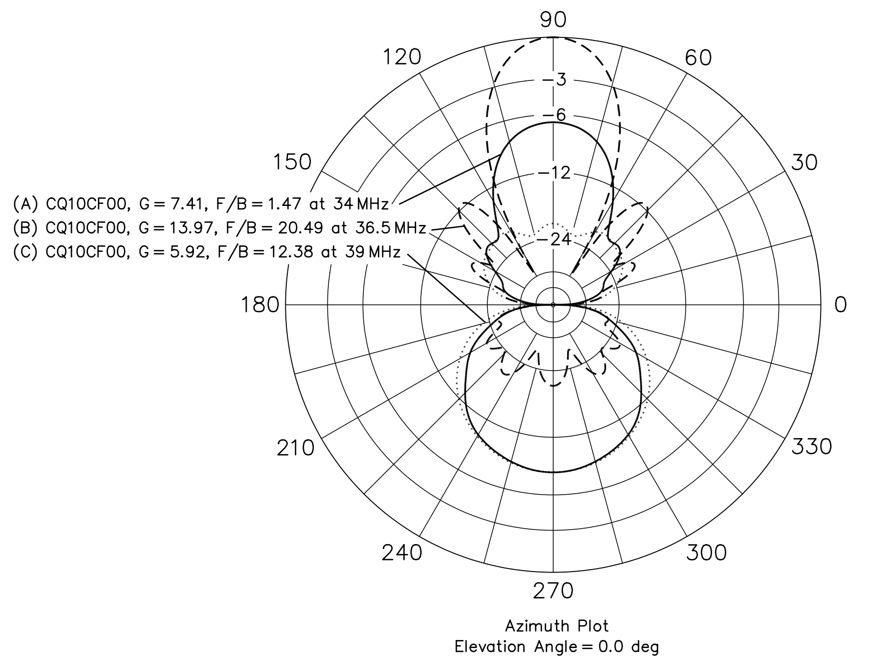

Fig 16—H-plane patterns for a 12-element quad at (A) 34 MHz, (B) 36.5 MHz, (C) 39 MHz.

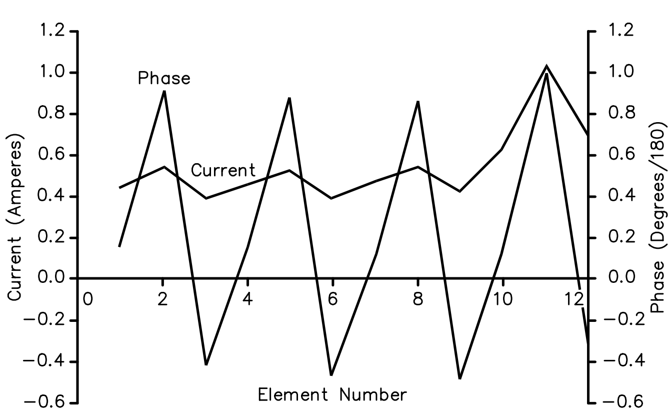

Fig 17—Element currents for a 12-element quad.

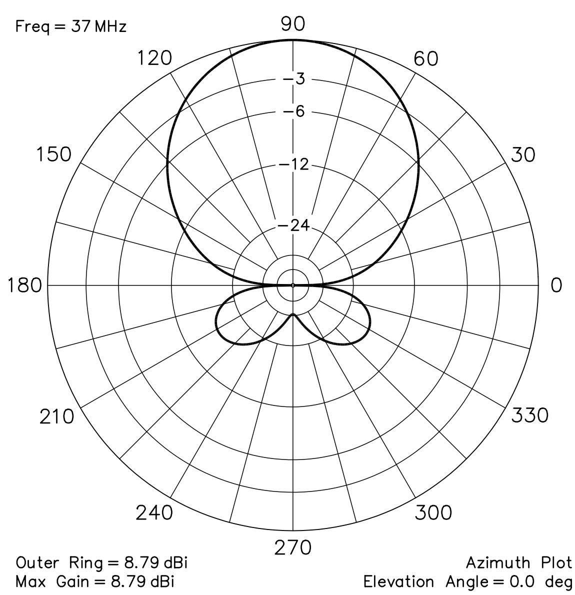

Fig 18—H-plane pattern for 4QUAD (4Q20A90), gain optimized.

always some increase in performance, at least with respect to the quantities included in the evaluation rules. There may also be some surprising designs of good performance, but there are some drawbacks. First, the "driver program" must be set up to contain the rules for selection and evaluation. This isn't a great problem when a prototype program for the antenna under study is available, but it is definitely work for a new type. When this is available and has been checked against known performance, a lot of computer time is required: It is common for a hundred individual antenna designs to be analyzed automatically.

Experience with the approach is building rapidly. A trial program, *YGO*, is available on the Internet for Yagi designs.[9] As far as I know, the quad family has not been investigated by this technique. For this series of extensions to quad studies, it was decided not to "go the GA route" for three reasons: One is the amount of work required. The second is that results seen, so far, indicate small improvements, not major ones. The third was that GA designs seem to be complex: For example, every element has a different size and spacing. In addition, it seems very difficult to set up an evaluation that reflects all goals of antenna design. The GA technique needs to be watched, and it may be desirable to revisit the conclusions of this paragraph later.

Some Optimization Tests

The following values were obtained by the old-fashioned method of varying one parameter to maximize one variable, then evaluating the effect on other variables. Fig 18 shows the result of changing the reflector of the four-element quad by adding 250 Ω of capacitive reactance. Compared to Fig 4A, gain has increased appreciably and the F/B ratio markedly. When this design is evaluated for other quantities, it is found that the wide-band characteristics of the starting antenna have been converted to a narrow-band design.

Fig 19 shows the effect of moving the reflector of the 12-element design of Figs 13 and 14. There is a small change in gain at the low-frequency end, expected because the reflector is the primary control of this part of the gain bandwidth. The wider spacing reduces F/B ratio. Possibly retuning the reflector would recover some of the lost F/B ratio, but this was not tried. Instead, tuning the original reflector was tried. As seen in Fig 20, a relatively small change does give better F/B, with an optimal value being shown. However, the result of such a small change indicates that the antenna performance would be sensitive to external influences such as the mast and guy

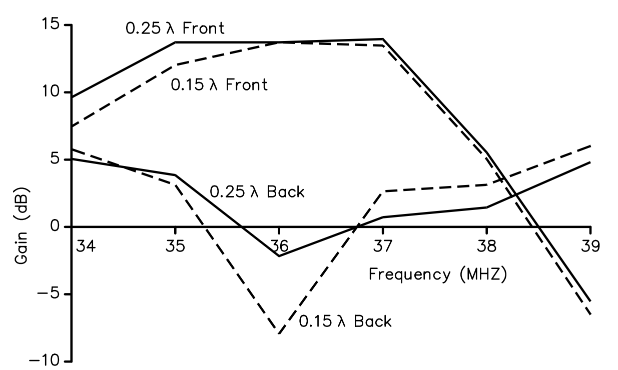

Fig 19—Effects of reflector spacing on a 12-element quad.

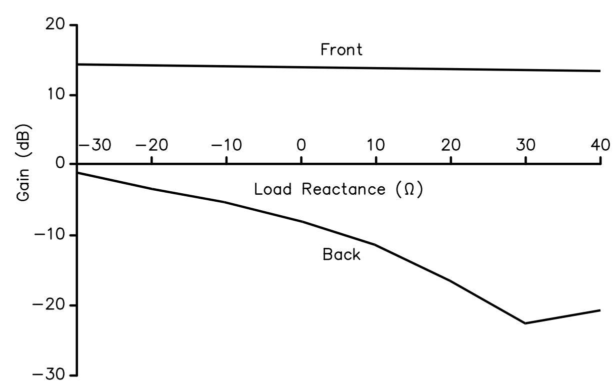

Fig 20—Effects of reflector tuning on a 12-element quad.

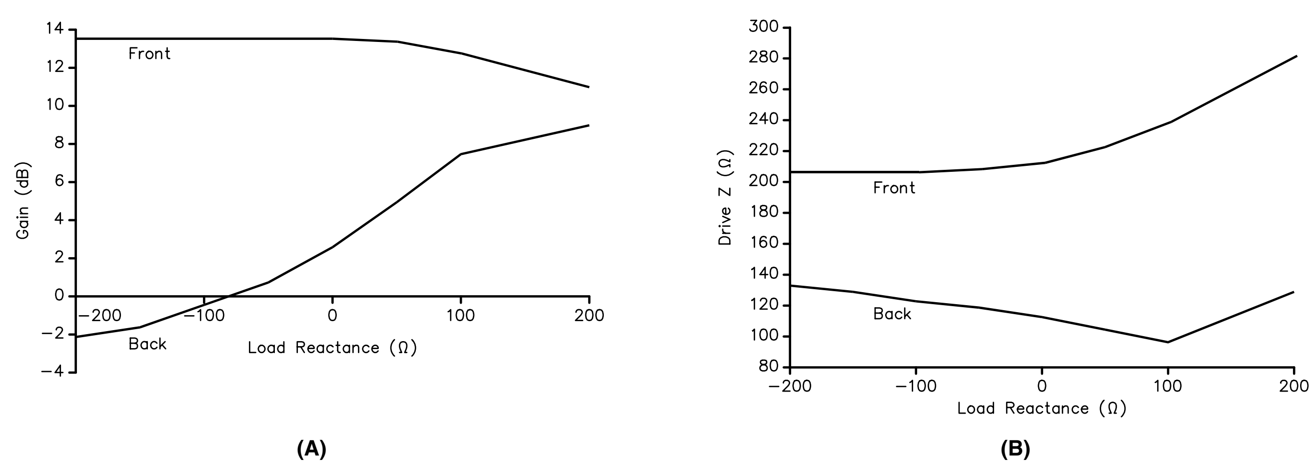

Fig 21—Effects of tuning the first director in a 12-element quad: (A) gain effect, (B) drive-impedance effects.

lines, or even nearby trees.

Fig 21 shows the result of tuning only the director farthest from the radiator. F/B can be improved, but the gain change is small where the F/B change occurs. There are minor changes in other characteristics, such as drive impedance as shown in Fig 21B. Lawson found that removing one director (perhaps during an ice storm) did not greatly affect performance of parasitic arrays: Total array length is the most important factor.

My point is that it may be easy to improve the value of a given performance variable, but it will not be easy to retain good performance of the several quantities that describe antenna performance overall. If you change a design parameter, be certain to look at effects on gain, F/B ratio, bandwidth and impedance. Also, think about practical consequences, such as proximity effects and construction problems.

That is not to say that improvements are not possible, particularly with respect to design values given here. Look at the total effect, not just at one variable.

Two-Reflector Designs

Several large quads described in the amateur literature have used two reflectors to improve F/B performance.[10] In my original quad study using *MiniNEC* for analysis, the possibilities of this were studied at some length. I found that some improvement in F/B was possible, but that it was only a few decibels. Therefore, use of two reflectors was not recommended.

Quite a different picture has emerged from studies using *NEC*. An example is shown Fig 22A, which shows the pattern with a single reflector and (B) that with an added reflector located 0.25 λ behind the first one. Gain increases about 0.6 dB and F/B ratio by over 10 dB. Drive resistance increases. Some variational studies indicate that the bandwidth characteristic of the antenna is not greatly changed.

Placing one reflector at 0.5 λ behind the director closest to the radiator, with the second one 0.4 λ behind this, also improved the gain and F/B ratio, but with less effect on the drive resistance.

Improvements found were not much greater than could be obtained by changing the size or location of a single reflector. The study was not exhaustive, however; different reflector-size/location combinations were not tried. At the moment, it appears that the increased boom length needed is better used to increase the number of directors. The matter needs more study.

Designs for Specific Bands

Table 2 shows some designs for specific bands. A primary factor in the choice of size was the practicality of construction—even two elements is a major undertaking on 80 meters. In this respect, these may be called "dream antennas," representing a small fraction of antennas used. It would perhaps be more realistic to scale the prototypes used to the next-higher amateur band; but then, it's nice to dream.

Specific values were developed by frequency scaling prototype designs already covered. Element size was then adjusted to give the same performance as the basic antenna, despite any change in scaled conductor size. The general performance expected is indi-

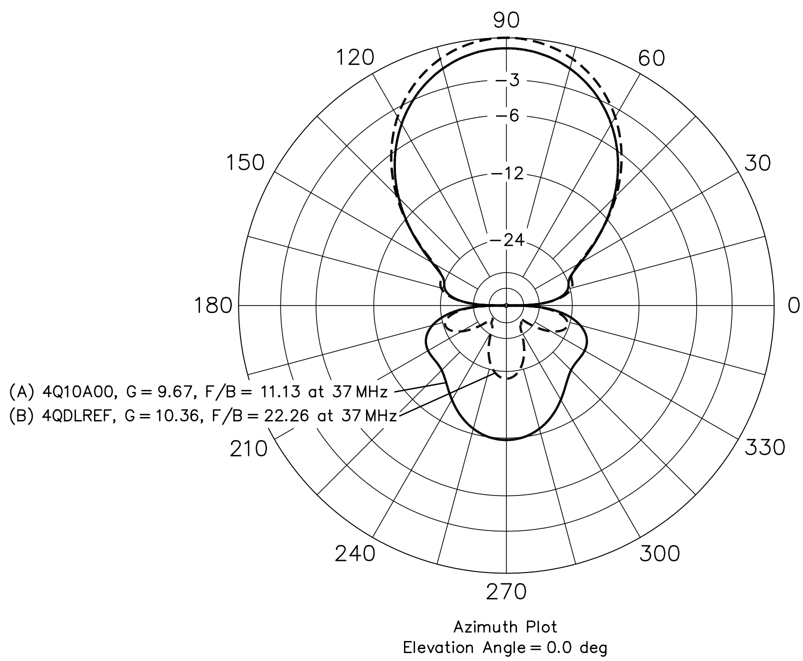

Fig 22—Dual-reflector effects in a four-element quad: (A) single reflector, (B) dual reflectors.

Table 2—Typical Large-Quad Design Values

Frequency (MHz)	*Prototype Name*	*Reflector ±Dimension*	*Spacing*	*Director ±Dimension*	*Spacing*	*Wire Size (AWG)*	*Gain (dB)*	*F/B (dB)*
7.150	3Q10CE00	18.02	19.66	16.38	39.33	14	8.5	28
14. 150	4Q05A00	9.13	10.43	8.70	20.87	16	10.8	20
21.250	4Q05A00	6.08	6.95	5.79	13.90	18	10.8	21
28.500	6Q10CF00	4.56	4.97	4.14	9.95	18	11.0	21
50.100	8Q10CF00	2.67	2.91	2.42	5.18	18	12.5	19
146. 000	AQ10CF00	0.92	1.00	0.83	2.00	0.25" OD	14.7	16

The "±" dimensions are the loop corners measured from the boom.
The first number in prototype name is the total number of elements.
The radiator is the same size as the reflector.

cated. All designs in this table should give good performance.

While the emphasis here has been on HF antennas, the information presented also applies at UHF and VHF, *providing that all dimensions (including the wire sizes) are scaled by the frequency ratios.* Of course, wire conductivity cannot be scaled, but the loss is usually negligible.

In Part 4 of this series, I'll examine the effects of ground on quads.

Notes

[1]R. Haviland, "The Quad Antenna Revisited," *Communications Quarterly*, Pt 1, Summer 1999, pp 43-73; Pt 2, Fall 1999, pp 65-85. A multitude of EZNEC2 description files used in this series are available for download from the ARRL Web **http://www.arrl.org/files/qex/**. Look for QUADS.ZIP.

[2]H. Yagi, "Beam Transmission of Ultra Short Waves," *Procedures of the IRE*, Vol. 16, June 1928.

[3]J. Lawson, *Yagi Antenna Design* (Newington: ARRL, 1986).

[4]J. Appel-Hansen, "The Loop Antenna with Director Arrays of Loops and Rods," IEEE *Transactions on Antennas and Propagation*, July 1972.

[5]R. Haviland, W4MB, *The Quad Antenna*, CQ Communications, Hicksville, NY, 1993-1996.

[6]J. Reisert, "Yagi-Uda Antenna Design," *Communications Quarterly*, Winter 1998.

[7]C. Walter, *Travelling Wave Antennas*, Dover, 1970.

[8]E. Altshuler, et al, "Yagi Antenna Design Using a Genetic Algorithm," *Communications Quarterly*, Winter 1998.

[9]You can download this package from the ARRL Web **http://www.arrl.org/files/qst-binaries/**. Look for YGO.TXT and YGO_INST.EXE.

[10]W. Orr, "Antennas," *CQ*, May 1979. The antenna described has six elements on a 52-foot boom with two reflectors, one radiator and three directors. All elements are equally spaced. For 10, 15 and 20 meters, the F/B is greater than 40 dB. No other dimensions are given.

R. P. Haviland was first licensed as W9CAK in 1931. He is a Fellow of the IEEE and has written 15 books and dozens of articles. R. P. is now retired. He worked as a project engineer on the first US rocket to reach outer space, which sent the first US radio transmission from beyond the atmosphere. He was also a founding director of the first commercial Communications Satellite Company.

EMPIRICALLY SPEAKING

Continued from page 2

components come in very handy for correcting transistor-bias variations from hot to cold, sensing airflow and compensating oscillators, as Bill ably demonstrates. Sam Ulbing, N4UAU, goes full duplex (well, almost) with his remote-control system. By sufficiently shortening equipment-switching times, information may be passed in both directions through multiplexing. This fine technique is seeing a lot of use in wireless networking.

Wouldn't it be nice if your antenna system automatically adjusted its pattern to maximize received signal-to-noise ratio and eliminate interference? That is precisely what adaptive beamformers do, and I submit a few basic thoughts about them. In increasingly hostile RF environments, beamformers could become necessary and standard on hand-held and mobile equipment. The analysis is based on ideal, isotropic radiators in free space. There is a bonus for transmitter-hunters.

In "RF" Zack Lau, W1VT describes a lightweight, QRP 4:1 balun.

What's Cooking?

Nate Sokal, WA1HQC, has written an update on class-E amplifiers. James Buckwalter, KF6SWC, worked with his Caltech friends to produce a switch-mode power supply that is the perfect companion for a class-E amp. We will present Warren Bruene, W5OLY's analysis of an ideal class of distortion-free tubes, intended to serve as a goal for tube manufacturers. Walt Maxwell, W2DU, has written us an article about the nature of power sources. Steve Best, VE9SRB, has come up with a derivation of the transmission-line equation starting from purely wave-mechanical arguments, thus proving again the equivalence of those two ways of looking at things.

We have a few articles on LF and MF propagation from Robert Brown, NM7M. Paolo Antoniazzi, IW2ACD, and Marco Arecco, IK2WAQ, have come through with a piece about making and measuring LF inductors, such as are often used for antenna loading. A nice collection of other antenna articles is on hand, including one from Valentin Trainotti, LU1ACM.

H. Paul Shuch, N6TX's, contribution is centered on the distributed processing used to search for extra-terrestrial radio signals. Bob Freeth, G4HFQ, has donated a treatise on how to make your sound card work for you in Amateur Radio applications. We have much interesting material on hand including articles on RF test equipment you can build, other construction projects and more.

What's Missing?

Software-defined radios or SDRs need our attention. An SDR is a radio that relies on DSP as much as possible and therefore contains hardware capable of many different modes and frequencies. That includes those not yet invented, because the software (firm-ware) may be rewritten to accommodate them, later. Alternatively, the software may employ adaptive algorithms that automatically detect new modes. SDRs represent a great chance for amateur experimenters to lead the way. The FCC is also excited about them and it is soliciting our input. The Amateur Radio Service is a perfect place to develop and test SDRs because —unlike commercial services—we may switch modes and frequencies at will, within certain restrictions. Unlike the Experimental Service, we need file no paperwork to start those operations.

Digital voice modes have been around since the public telephone network went digital. Until now, it has been a problem to attain the necessary data rates over radio below VHF. Recent breakthroughs in speech compression, though, have brought the technology within reach of typical low-speed modems and allowable HF bandwidths. We need to work together to adopt a standard for this modulation. It is allowed on virtually all our bands, and while its occupied bandwidth is limited, its baud rate is not. We must press to build these systems and test them, especially as they relieve congestion and interference.

In the field of antenna technology, several interesting prospects have popped up recently, including fractal antenna geometries and adaptive beamformers. DSP may also be combined with antenna arrays to combat multipath and other distortion. Academic research on these topics is hot: It may produce some surprises over the next 10 years or so. Articles in this area are certainly needed.

One reason more experimenters are not trying DSP and other digital methods is the scarcity of "freeware" for target computers. Look at the success of the DSP-10 (Bob Larkin, W7PUA, *QST*, Sep-Nov 1999) to see that fact. We aren't all programmers, but an understanding of software is important in this age. How about some more soft-ware projects? We will bring in a little more digital theory with regard to synthesizers and speech processing. We hope that others working in chiefly digital realms, such as wavelet transforms, would inform us about the state of their arts.—*73, Doug Smith, KF6DX,* **kf6dx@arrl.org**

Conductors for HF Antennas

Putting up an antenna for the low bands? What kind of wire will you use? This analysis may change your plans.

By Rudy Severns, N6LF

Most of us give little thought to the wire from which we fabricate antennas. Most of the time that's okay, but some antennas are quite sensitive to conductor loss. Then we need to think carefully about our choice of wire or other conductor. Recently, I have been building 160-meter wire arrays using hundreds of feet of wire in each. Some of the spans are over 600 feet, and they are attached to poles and trees that move in the wind. For this reason, I initially used #12 stranded Copperweld with PVC insulation. One of the antennas is a two-element, end-fire array—essentially a vertically polarized W8JK. It is a problem with any end-fire array that to obtain gain, the radiation resistance must be lowered by closely spacing the elements. In the case of a W8JK array, the impedance is in the range of 8 to 20Ω. As Krause pointed out in Reference 1, this makes the obtainable gain very sensitive to conductor resistance. The problem is particularly severe on 160 meters because the wire used is very long (over 700 feet in my array) and tubing is impractical.

The performance of the W8JK array was good, but I had a feeling that I could get much more from the antenna. This led me on a hunt to identify possible losses: to measure wire resistance, to analyze expected conductor losses, to finite-element model solid-copper and Copperweld (copper-clad steel) wire and to model the effects of wire losses on antenna performance. The results are interesting and give insight into appropriate conductor selection. It turns out my intuition was right, the conductor loss was high. The wire resistance was double the expected value, but the reason for that was a surprise.

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Conductors

Many types of wire, conductive strips and tubes can be and are used for antennas. The reference against which other wires are judged is solid #12 AWG, soft-drawn, bare copper. Other common choices are:

- seven-strand, hard-drawn copper
- solid #12 AWG Copperweld
- 19-strand Copperweld (#12 AWG)
- aluminum electric-fence wire, in various sizes

• Alumoweld (aluminum-clad steel, see Reference 2)
• #8 AWG aluminum clothesline
• aluminum tubing
• thin copper or aluminum strips
• stainless steel tubing
• towers and galvanized steel guy wires

Occasionally galvanized steel fence wire, stainless steel or copper plated steel electric fence wire is suggested for antennas. These are *very poor* choices, as I will show shortly. Table 1 lists the resistivity and conductivity for some common conductors. The values for steel are only approximate because they vary greatly with the exact composition and processing history.

Sometimes silver plating is suggested for conductors. The conductivity of silver is only 6% better than copper, but when the surface oxidizes, silver oxide is a much better conductor than copper oxide. We will not be considering silver conductors for the rest of this article, however.

Skin Effect

The resistance of wire at a given frequency depends on three things: size, electrical properties of the material (including surface corrosion!) and the

Table 1—Conductivity and Resistivity of Conductors

Material	*Conductivity (σ)* siemens/meter	*Resistivity (ρ)* ohm-cm
Silver	6.2×10^{7}	1.62×10^{-7}
Copper (annealed)	5.8×10^{7}	1.7241×10^{-6}
Aluminum (99.9%)	3.81×10^{7}	2.62×10^{-6}
Iron	1.03×10^{7}	9.71×10^{-6}
Low-carbon steel (AISI 1040)	0.5×10^{7}	20×10^{-6}
Stainless steel (AISI 304)	0.11×10^{7}	90×10^{-6}

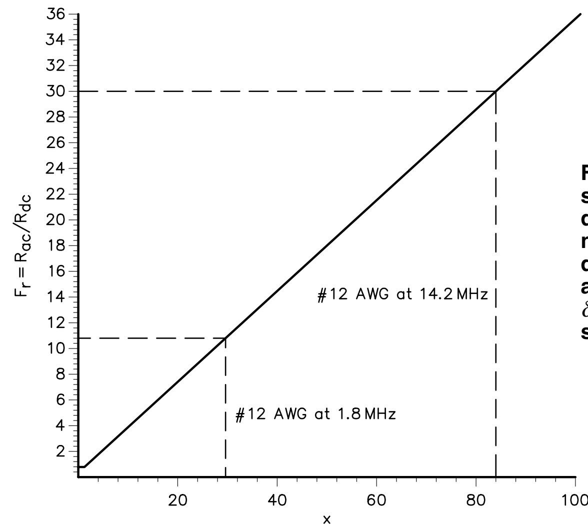

Fig 1—R_{ac}/R_{dc} ratio for solid round wire. Wire diameter (X) is normalized to the skin depth, , where *d* is the actual wire diameter and δ is the skin depth in the same units.

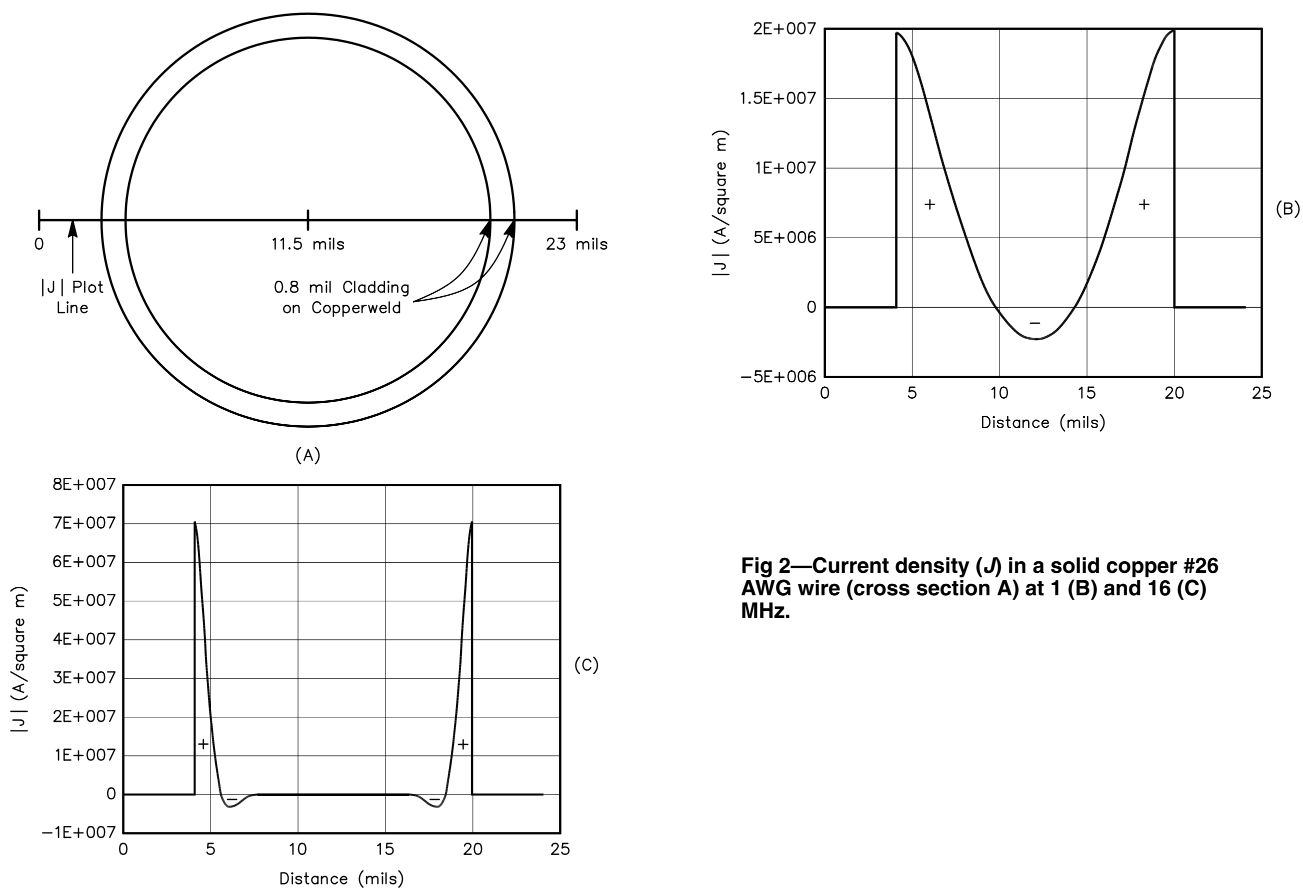

Fig 2—Current density (J) in a solid copper #26 AWG wire (cross section A) at 1 (B) and 16 (C) MHz.

resistance increase due to skin effect. Skin effect is the tendency for current to crowd to the outer perimeter of a conductor as frequency is increased. It is characterized by the depth at which the current density (J) has fallen to about 0.37 (1/e, where e=2.718). For good conductors, the skin depth (d) is expressed by:

$$\delta = \sqrt{\frac{1}{\pi\sigma\mu f}} \text{ meter} \qquad \text{(Eq 1)}$$

where:

δ = skin depth (meters)

μ = permeability = μ_r μ_o; μ_o= 4×10^{-7} H/m; μ_r = relative permeability

σ = conductivity in siemens/m (mho/m)

f = frequency (hertz)

For copper at room temperature:

$$\delta = \frac{2.602}{\sqrt{f_{MHz}}} \text{ mils} \qquad \text{(Eq 2)}$$

For f = 1.8 MHz, δ =1.94 mils. For f = 14.2 MHz, δ = 0.69 mils. The Appendix contains a graph of the relation between skin depth and frequency for copper at 20° and 100°C.

For round wire, the variation of R_{ac}/R_{dc} (F_r, or resistance factor) with normalized wire diameter $X = d/\delta\sqrt{2}$ is shown in Fig 1. The variable d is the wire diameter, in the same units as δ. The equation from which the graph is derived is given in the Appendix. For #12 AWG copper wire at 1.8 MHz, X = 29.5 and F_r = 10.8. For the same wire at 14.2 MHz, X = 83 and F_r = 30. This thirty-fold resistance increase at 20 meters is due to skin effect! It cannot be ignored on any amateur band.

I am fortunate to have access to finite-element modeling (FEM) CAD software that can directly calculate and graph current distribution and power loss in conductors such as solid copper wire or Copperweld, which is made of two different materials. The graphs in Figs 2 through 5 were generated using FEM software (see Reference 2).

Figs 2B and 2C give plots of the current density (J in A/m^2) along the line shown in Fig 2A, for solid #26 AWG copper wire (δ = 15.9 mils) at 1 and 16 MHz. The crowding of current to the outside perimeter of the wire and how crowding worsens as frequency increases is clearly shown. This is why the apparent resistance of the wire increases so much. At some points within the wire, the instantaneous current is actually flowing backwards (minus signs) due to the self-induced eddy currents that are the underlying phenomena responsible for skin effect. These currents must be balanced by more forward current (+) to keep the average current unchanged. That is, the same number of carriers must come out one end of the wire that you put in the other end. The net result is increased power dissipation for a given RMS current.

In Copperweld wire, the copper cladding on the outside of the wire is typically about 10% of the wire radius. For #26 AWG wire, the cladding thickness would be about 0.8 mils (0.0008 inches). Fig 3 graphs J for #26 AWG Copperweld. It is clear that the current is flowing only in the copper cladding; there is almost no current in the steel core. This is predicable from the skin-depth equation; δ is inversely proportional to the square root of the permeability. For steel, μ_r is highly variable, affected by the composition of the steel, the processing and even

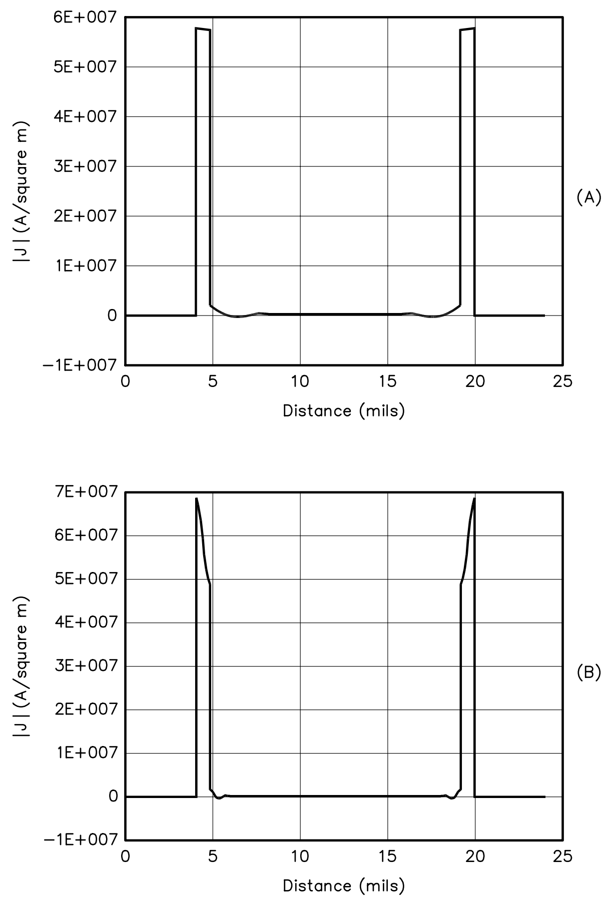

Fig 3—Current density in a #26 AWG Copperweld wire with 0.8-mil cladding at 1 (A) and 16 (B) MHz.

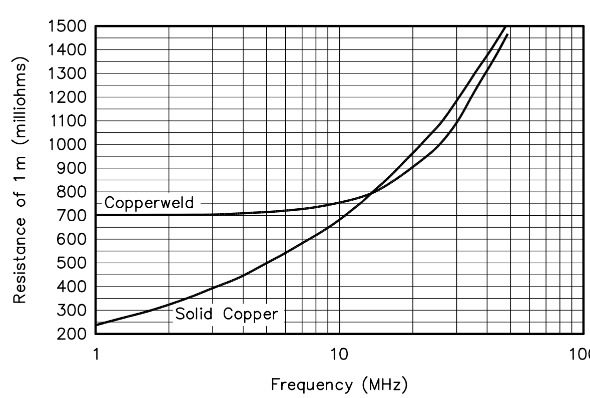

Fig 4—Resistance comparison of 1-meter lengths of #26 AWG solid copper and #26 AWG Copperweld with 0.8-mil cladding from 1 to 30 MHz. Derived from FEM modeling.

the current level. Losses can actually increase as the current increases because μ_r increases with flux density (B), reducing the skin depth and increasing R_{ac}. Thus, μ_r can be from 1000 to 10,000 or more, which means that the skin depth at 1 MHz and above is very small. Copperweld behaves very much like a tubular conductor. This can allow the conductor loss to actually be less or greater than a solid conductor of the same outside diameter, depending on the wall thickness and frequency.

A graph of R_{ac} for 1-meter lengths of #26 AWG solid copper and Copperweld (0.8-mil cladding) wires is given in Fig 4. Below about 14 MHz, the solid copper wire has less resistance. In fact at 2 MHz (160 meters), the Copperweld has more than twice the resistance of solid copper wire. This is simply because current in the Copperweld is crowded into a thin layer. The tube is too thin! Above 14 MHz, however, the tube has less resistance and the Copperweld is superior. Notice also that at low frequencies, the resistance of the Copperweld is nearly constant.

This can be explained from Fig 3, which shows that at low frequencies the current density is basically uniform and changing frequency doesn't change J much. As you reach the middle range of frequencies, current distribution in the tube is better than that in the solid wire and the loss is less. At some high frequency, current distribution in the tube will equal that in the solid wire (the core no longer matters) and its resistance will be the same. In Fig 4, the resistances begin to converge above 50 MHz. The resistances shown in Fig 5 for #12 AWG wires clearly illustrate the convergence at high frequencies. Thus, there is a region, depending on the cladding thickness, where Copperweld is superior to solid wire, but below this region, it is inferior!

Fig 5 is a graph of R_{ac} for 1-meter lengths of four different #12 AWG wires: solid copper, Copperweld with 4-mil and 2-mil cladding and an approximation for 19 strands of #26 Copperweld with 0.8-mil cladding. Again, we see the excess resistance for the 0.8-mil Copperweld at 160 meters, but now the crossover frequency with solid copper is just above 7 MHz. For 30 through 10 m, the stranded Copperweld is somewhat better (5-10%) than solid copper. Copperweld with 4-mil cladding (which is standard for solid #12 AWG Copperweld) is slightly better (≈5%) than solid wire on 160 meters and equal at higher frequencies. While the electrical properties are good and the wire is very strong and durable, the stiffness of Copperweld and its strong desire to remain coiled make it the devil's own invention to work with. Wear gloves and eye protection when working with it!

For 40 meters and up, stranded Copperweld is a good choice: It has low resistance, good strength and is reasonable to work with. For 80 and 160 meters however, the resistance is quite a bit higher and may be a problem for some antennas. Solid copper or Copperweld would be a better choice. In the case of iron fence wire, stainless steel wire or copper-plated steel electric-fence wire, the skin depth will be very small and the ac resistance very large. The copper plating on electric-fence wire is simply too thin to be of any help at HF.

We must also consider that the current distribution on all but the shortest antennas is not constant but nearly sinusoidal or a portion of a sinusoid. Because the losses are proportional to I^2R_{ac}, the loss will be different in different parts of the antenna. This can be accounted for by placing an equivalent resistance (R_{eq}) at the current loop, such that R_{eq} dissipates the same total power as the wire. The efficiency (η) of an antenna, taking into account only the radiation resistance (r_r) and the equivalent wire resistance, will be $\eta = r_r/(r_r+R_{eq})$. For $\lambda/2$ or $\lambda/4$ conductors with sinusoidal current distributions, $R_{eq} = R_{ac}/2$, where R_{ac} is the ac resistance for the entire wire length. A derivation of this result is given in the Appendix. For constant current distribution along the conductor, $R_{eq} = R_{ac}$.

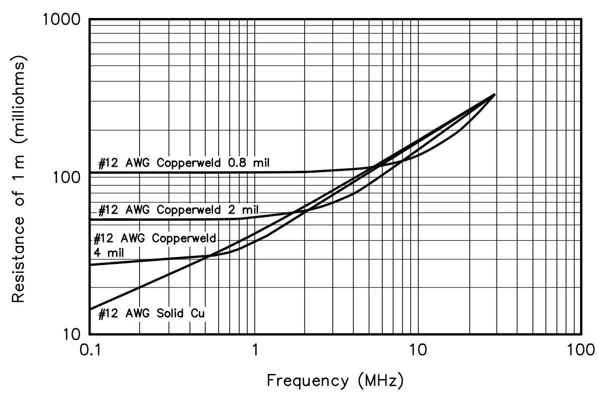

Fig 5—Resistance comparison of 1-meter lengths of #12 AWG solid copper, #12 AWG Copperweld with 4-mil cladding and an equivalent wire for 19-strand #26 AWG Copperweld with 0.8-mil cladding from 1 to 30 MHz. Derived from FEM modeling.

Table 2—Wire loss comparison for #12 wires.

	14.2 MHz Dipole		***1.85 MHz Dipole***		***1.85 MHz Ground-Plane***		***1.85 MHz W8JK***	
Conductor	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*
Perfect	2.14	0	2.14	0	5.27	0	5.93	0
Copper	2.09	–0.05	2.01	–0.13	5.10	–0.17	4.92	–1.01
19–strand Copperweld	2.09	–0.05	1.88	–0.26	4.81	–0.35	3.93	–2.0
Aluminum	2.07	–0.07	1.94	–0.20	5.01	–0.26	4.47	–1.46
Iron	–1.88	–4.02	–4.99	–7.13	–2.58	–7.85	–9.58	–15.5

Table 3—Wire loss comparison for #18 wires

	14.2 MHz Dipole		***1.85 MHz Dipole***		***1.85 MHz Ground–Plane***		***1.85 MHz W8JK***	
Conductor	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*	*Gain (dBi)*	*Loss (dB)*
Perfect	2.13	0	2.13	0	5.27	0	5.94	0
Copper	2.08	–0.05	1.99	–0.14	4.93	–0.34	4.06	–1.88
Aluminum	2.06	–0.07	1.92	–0.21	4.75	–0.52	3.29	–2.65

Effects of Wire Loss on Gain

Okay, so as frequency increases, the resistance of the wire increases and different conductors have more or less loss. *So what! Does it really matter?*

One way to get a handle on this question is to model some typical antennas and determine the effect of different wire sizes and materials on gain. You can also calculate R_{eq} and then calculate the efficiency of the antenna. This is done in the Appendix. Tables 2 and 3 show the results of modeling three different antennas using perfect, copper (Cu), aluminum (Al) and iron (Fe) conductors of two different sizes. I assumed a resistivity of 10^{-7} Ω-m and a relative permeability of 1000 for the iron wire. Steel wire could actually be worse (lower conductivity and higher permeability). The dipoles and the W8JK array are modeled in free space. The W8JK array has two λ/2 dipoles, spaced λ/8 apart and fed 180° out of phase. The ground-plane antenna has four radials, 10 feet above perfect ground.

The tables show several things of interest. First, for the same wire size, as frequency decreases the wire loss increases. This is because even though the wire resistance per-unit-length is decreasing ($1/\sqrt{f}$) the wire length is increasing (1/f). The net wire resistance increases as frequency decreases if the antenna length is scaled. This increase in wire loss can become important in low-band antennas. Second, except for the iron wire, the effect of wire loss and wire size is very small in dipole antennas. You can use copper or aluminum wire in fairly small sizes without compromising performance much. It is also clear that using iron fence wire is bad news.

The ground-plane antenna is more sensitive to wire characteristics than are the dipoles because of its lower impedance, but again the changes are small as long as copper or aluminum wire is used. The use of more radials will reduce wire loss.

The W8JK array, however, is very sensitive to wire size and material. With perfect conductors, the gain over a dipole is 3.8 dB. Using #18 AWG aluminum wire gives away most of that gain (–2.65 dB). Even with #12 AWG copper wire, there is still a loss of over 1 dB. In the W8JK, changing to a #6 AWG wire or two parallel, spaced #12 AWG wires reduces the wire loss to –0.53 dB.

Any low-impedance antennas, such as Yagis, end-fire arrays or short loaded verticals will be sensitive to wire size and conductivity. On 80 and 160 meters, many verticals are short and heavily loaded!

Flat-Strip Conductors

Up to this point, we have been considering round conductors. An alternative would be to use thin, flat conductors of either copper or aluminum. Fig 6A is a graph of R_{ac}/R_{dc} for thin, flat-strip conductors (see the Appendix for generating equation). For #12 AWG round copper wire at 1.8 MHz, $F_r \approx 11$. If we take the same wire and roll it out into a strip approximately 0.010×0.625 inches, the thickness of the strip in skin depths will be about 5. Looking at Fig 6 we see that for $X = 5$, $F_r = 2.4$, which is a factor of 4.6 lower than for the equivalent round wire. By dividing F_r by the corresponding values of X, we can create Fig 6B, which is a graph of resistance normalized to 1 Ω for a thickness of 1 skin depth. Notice that for $X < 1.5$, the $R_{ac} = R_{dc}$, but as foil thickness increases the resistance goes through a minimum at $X = \pi$ and then back up about 9% to level out at a constant R_{ac} regardless of the thickness. For $X < 1.5$, the current distribution in the conductor is almost uniform, so $R_{ac} = R_{dc}$. Above this point, the distribution in increasingly on the outer surfaces of the strip.

At high frequencies, all the current is on the outer perimeter of the strip so the thickness of the inside doesn't matter. Only the length of the perimeter counts. This is the same as for a round wire. The important difference between round and strip conductors is that for a round wire, you have to increase the diameter to reduce R_{ac}. This means you have a lot of unused copper (inside the wire) to buy. Strip or foil conductors can be kept thin and simply made wider to reduce R_{ac}. You put the extra copper to good use and in the end buy less.

Of course, there is the issue of increased wind area with a foil conductor. Foil also tends to "sail" and/or flutter in the wind, distorting the antenna shape and stressing the array. That is a downside! Putting a spiral twist in a foil conductor helps to keep it from flying around in the wind. I have found that 0.010×0.5- to 1-inch strip works pretty well and doesn't fly around or flutter too much. Unfortunately, copper and alu-

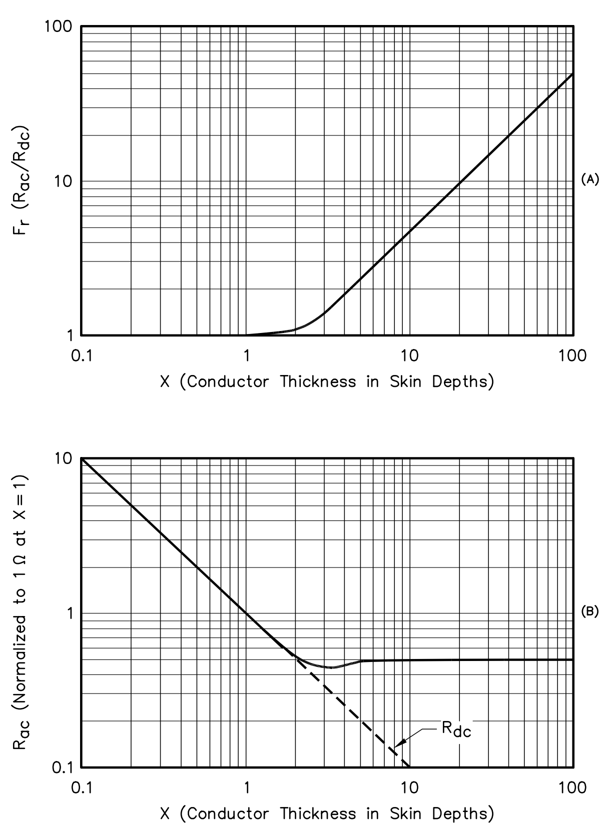

Fig 6—Resistance factor ($F = R_{ac}/R_{dc}$) for flat-strip conductors as a function of thickness in skin depths.

minum foils of appropriate sizes are not so readily available as round wire.

Composite Antenna Assemblies

In some cases, straight copper wire simply does not have the strength required, but the alternatives may have too much resistance. It is possible to compromise by using different conductors at different places in the antenna. Keep in mind that the losses are I^2R in nature. This means that the bulk of the losses occur in the high-current regions of the antenna. Fig 7 shows a 160-meter, two-element end-fire array mentioned earlier. The vertical portions have high current levels; they are made from 0.010×0.625-inch copper strip. The horizontal portions have much less current. The antenna is supported from the top between two poles 300 feet apart, so the upper wires have considerable stress. These are stranded Copperweld. The lower horizontal wires have very little stress; they are copper. The result is an antenna with minimum loss but strength where it is needed.

The 50-pF capacitors tune out the inductive reactance at the feedpoint. These must be high-voltage, high-current capacitors, which usually come in only a few standard sizes. The position of the capacitors and the lengths of the upper horizontal wires can be adjusted to give 450 Ω resistive at the feedpoint. That allows the use of 450-Ω ladder line as the feedline to ground level, where a 9:1 balun transforms to 50 Ω for the run back to the shack. Stub matching could be used instead.

Measurement of Wire Resistance

Theory and modeling are nice, but I wanted to make some actual measurements of wire resistance to confirm the modeling and calculations. Unless you have access to an impedance analyzer such as an HP4192 ($50,000 please!), this is not an easy measurement to make directly. After several false starts, I found it best to wind the wire into a large coil of well spaced turns and measure the Q on a Boonton 260A Q-meter. This gave reasonable results that are shown in Tables 4, 5 and 6. The values for resistance are probably not very precise, but the relative differences between different wires are clearly shown.

The coil is 17 turns (except for the #8 AWG aluminum wire which used 16 turns) spaced 1.5 wire diameters (with 1/8-inch Dacron rope) on a 4.2-inch ID PVC-pipe form. 4.5 inches long. The coil requires 19.5 feet of wire. The copper wire was #12 AWG, the antenna wire was 19 strands of #26 AWG Copperweld. (The Copperweld is nominally #13, but when I put a micrometer on the two wires, the sizes were not very different: 0.077 inches for the antenna wire, versus 0.082 inches for the solid copper wire. This results in only a 6% resistance difference.)

I took great care to make the two coils identical. They were both wound on the same form. Measurements were done with the same lead lengths and coil position relative to the Q-meter. A frequency counter was used to set the Q-meter frequency. The Q-meter zeros were carefully adjusted, and so on. The close values for the resonating capacitance show that the coils were very nearly identical.

Three things jump out at you from Tables 4 and 5:

- The Q for the coil made with stranded Copperweld is substantially lower than that of the solid copper wire coil.
- The variation in Q over frequency is different for each coil.
- The coil Qs begin to converge as the frequency is increased.

Remember that $Q = X_L/R_S$, where $X_L = 2\pi fL$ is the impedance, and R_s is the total series loss resistance.

In an antenna, we are interested in the resistance due to skin effect (R_{ac}), so we must separate the components of coil loss to get an estimate of the skin-effect loss. R_s has several components:

- skin effect in the conductor
- turn-to-turn and geometric proximity effects
- losses in the coil form
- loss in wire insulation
- radiation from the coil
- losses due to eddy currents in nearby conductors

Skin effect can be calculated quite accurately using the equation of Fig 1

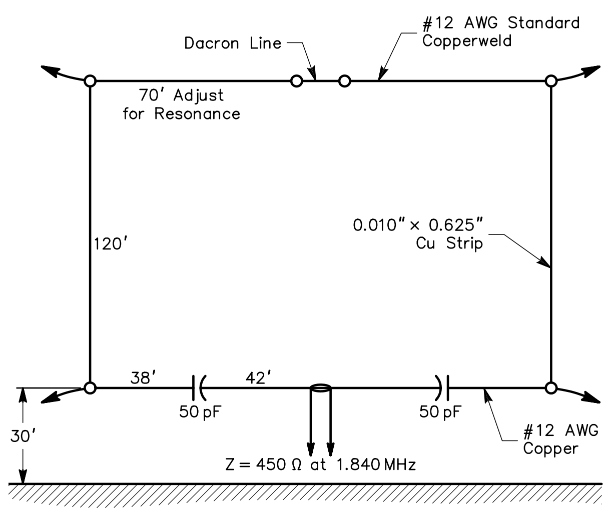

Fig 7—N6LF W8JK two-element vertical array using composite-wire construction.

Table 4 —#12 AWG bare solid copper wire test results using new wire

Frequency	*Resonating Capacitance*	*Measured Q*	X_L	R_S
1.8 MHz	359 pF	410	246.3 Ω	0.601 Ω
3.9 MHz	69 pF	360	591 Ω	1.64 Ω

Table 5—19-strand #26 AWG Copperweld test results using new wire

Frequency	*Resonating Capacitance*	*Measured Q*	X_L	R_S
1.587 MHz	460 pF	250	218 Ω	0.87 Ω
1.8 MHz	358 pF	270	247 Ω	0.92 Ω
3.9 MHz	68 pF	323	600 Ω	1.85 Ω

for a solid, round conductor. For 19.5 feet of #12 AWG copper wire at room temperature and 1.8 MHz:

R_{dc} = 0.031 Ω, from wire table (I measured the coil as 0.030 Ω on a bridge),

$R_{ac}/R_{dc} = F_r$ = 10.79, from Fig 1,

$R_{ac} = R_{dc} \times F_r$ = 0. 334 Ω

At 1.8 MHz, the total R_s in Table 4 is 0.601 Ω, which indicates an additional loss resistance of 0.267 Ω beyond the skin effect.

Given the close similarity between the two coils, we can estimate the stranded Copperweld coil resistance component due to skin effect to be:

$R_{skin} \approx R_s - 0.267 = 0.915 - 0.267 = 0.648$ Ω

This is 1.9 times the resistance of solid copper wire! This agrees rather well with the comparison in Fig 5 between 0.8-mil clad Copperweld and solid copper wires. In a dipole, I don't think this would matter but in a W8JK array, it's bad news.

Looking again at Fig 5, we would expect the skin-effect loss for the two types of wire to converge as we go higher in frequency, reflected in more similar Qs. This is what we see in Tables 4 and 5. We would also expect the Q of the Copperweld coil to decrease with frequency because X_L is decreasing, but R_s is not. For the solid wire coil, both X_L and R_s are decreasing, so Q is more stable.

Emboldened by these results, I wound coils using several other wires I had on hand or was able to scrounge from friends. The test results are given in Table 6. I threw in the iron fence wire just for kicks!

The differences in the 14 different wires tested are quite easy to see:

- The #12 AWG wire is better than #14 AWG
- New insulation has very little effect (but weathered insulation may not be so benign!)
- Oxidation of bare wire definitely reduces the Q. Both samples were only mildly oxidized. Longer exposure would have further reduced the Q.
- Stranded wire is inferior to solid
- Very fine stranding (168-strand sample) reduces the Q significantly
- For the same size wire, solid Copperweld is just as good as solid copper
- At least at low frequencies, stranded Copperweld is inferior to solid Copperweld and other copper wires, solid or stranded
- Iron fence wire is bad news!

I also wanted to verify the advantage of Copperweld wire implied by Fig 4. Using #14 AWG solid copper and solid Copperweld, I wound free-standing three-turn coils and then two coils on a ceramic coil form. The results are shown in Table 7.

In both cases, the Copperweld produced a coil with somewhat higher Q, as predicted by Figs 4 and 5. Remember that only part of R_s results from skin effect, so the difference between the two wires is diluted by other losses. The tests were run a number of times to be sure the differences were real and repeatable.

Aluminum Wire Connections

Aluminum wire has the advantages of very low cost and a better strength-to-weight ratio (≈3×) than copper. The reduced conductivity (σ) of aluminum can be accommodated by using a larger wire size. For an equal resistance, it will still weigh less than copper. Keep in mind we are talking about equal R_{ac} not R_{dc}! The difference arises because of skin effect, which is proportional to $1/\sqrt{\sigma}$. The skin depth will be greater in aluminum than in copper (at the same frequency) because of the lower conductivity. The lower weight and higher strength is helpful in long spans and may put off the need to use Copperweld conductors.

However, aluminum has one major disadvantage. Making a low resistance connection that will remain low during extended exposure to the elements is not a trivial exercise. It is very possible for a poor connection to introduce significant loss, especially if it is at a high-current point. There are also corrosion problems with connecting copper conductors to aluminum conductors.

Alumoweld Wire

In addition to Copperweld, aluminum-clad steel wire is available under the name Alumoweld. It is available in a variety of sizes, although the smallest size available is #12 AWG. It is also available as stranded wire and stranded guy wire equivalent to the galvanized wire used for guys. While it is very stiff—handling very much the same as Copperweld or steel wire—it has some advantages. In most atmospheres, it is much more resistant to corrosion than galvanized steel. It is electrolytically compatible with the aluminum tubing frequently used in antennas, so it can be used for support wires in aluminum antenna structures to avoid dissimilar-metal corrosion.

Towers and Supports

It is quite clear that iron fence wire is a very poor choice for antennas, but what about steel towers and the use of galvanized or stainless steel guy wires as antenna elements? In towers, the surface area is much larger than that

Table 6—Comparison of Q for coils made with various wires at 1.8 MHz

Wire description	*Q*
New #12 bare soft-drawn solid copper	410
New insulated solid #12	410
New insulated stranded #12 THWN	350
New insulated 19 strand #26 Copperweld	270
New #14 bare soft-drawn solid copper	353
New #14 bare solid Copperweld	360
New #14 bare stranded Copperweld	194
Oxidized #14 bare stranded Copperweld	162
New #14 bare 7/22 stranded hard-drawn copper	338
Oxidized #14 bare 7/22 stranded hard-drawn copper	300
New #14 168 strand superflex	225
#14 aluminum electric fence wire	260
#8 aluminum clothesline	360
#13 iron fence wire	25

Table 7—Coil Qs measured at 25 MHz

Coil form	*Wire*	*Measured Q*	X_L	R_S
Air	Copper	285	145 Ω	0.51 Ω
Air	Copperweld	310	138 Ω	0.45 Ω
Ceramic	Copper	266	186 Ω	0.70 Ω
Ceramic	Copperweld	282	193 Ω	0.68 Ω

Appendix

A. Skin depth in copper

Fig A is a graph of skin depth in copper as a function of frequency for two temperatures.

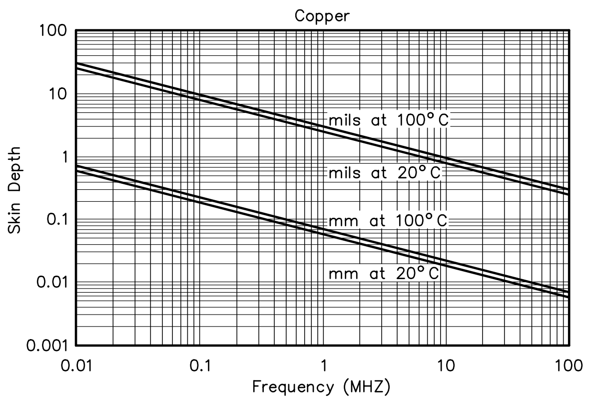

Fig A—Skin depth in copper at 20°C and 100°C; dimensions are in mils and millimeters.

B. R_{eq} Derivation

The current distribution in an antenna is usually a sinusoid or a portion thereof as indicated in Fig B. With the center as the origin:

$$I = I_{o(RMS)} \cos\left(\frac{\pi}{2}\right)\left(\frac{x}{l}\right) \qquad \text{(Eq A)}$$

Note that I_o, the current at $x = 0$, is RMS! The wire loss is:

$$\Delta P = \Delta R I^2 dx \qquad \text{(Eq B)}$$

Where ΔR is the resistance per unit length. The total power loss is then:

$$P = \int_a^b \Delta R I^2 dx = \Delta R I_o^2 \int_a^b \cos^2\left(\frac{\pi x}{2l}\right) dx$$

$$P = \left(\Delta R I_o^2\right)\left[\frac{l}{\pi}\sin\left(\frac{\pi x}{2l}\right)\cos\left(\frac{\pi x}{2l}\right) + \left(\frac{x}{2}\right)\right]_a^b \qquad \text{(Eq C)}$$

$$R_{eq} = \frac{P}{I_o^2} = \Delta R\left[\frac{l}{\pi}\sin\left(\frac{\pi x}{2l}\right)\cos\left(\frac{\pi x}{2l}\right) + \left(\frac{x}{2}\right)\right]_a^b$$

For $a = 0$ and $b = l$:

$$P = I_o^2\left(\frac{\Delta R l}{2}\right)$$

$$R_{eq} = \left(\frac{\Delta R l}{2}\right) \qquad \text{(Eq D)}$$

Where ΔRl is the total R_{ac} for the length of wire.

R_{eq} can be used directly to calculate the gain decrease due to conductor loss. The loss is simply the log of the efficiency:

$$loss = 10\log\left(\frac{r_r}{r_r + R_{eq}}\right) \text{dB} \qquad \text{(Eq E)}$$

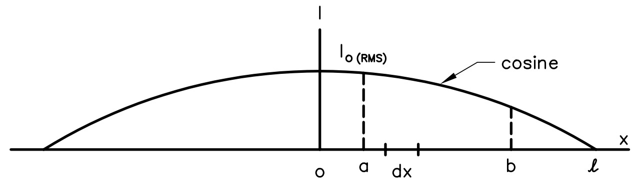

Fig B—Current distribution on an antenna wire and definition of equation quantities.

where r_r = radiation resistance. For a dipole in free space where r_r = 73 Ω, see Table 8.

The loss in gain by this calculation agrees with the gain loss in Table 2 that was derived using MOM (method of moments) in an antenna-modeling program.

C: Resistance Factor for Round Wire

$$F_r = \frac{R_{ac}}{R_{dc}} = \frac{\gamma}{2}\left[\frac{ber\gamma\, bei'\gamma - bei\gamma\, ber'\gamma}{ber'^2\gamma + bei'^2\gamma}\right]$$

$$\gamma = \frac{d}{\delta\sqrt{2}} \qquad \text{(Eq F)}$$

$\delta =$ skin depth, $d =$ wire diameter

where *ber* and *bei* are the real and complex parts of Bessel functions with complex arguments. They are often called Kelvin or Thompson functions. Most spreadsheet programs do not have these functions. Math programs like *Maple* or *Mathmatica* do have them. Fig 1 was done with Maple. It is possible to use series summation approximations that can be found in advanced math tables (see Reference 6).

D. Resistance Factor for Thin, Flat Foil where X = Thickness in Skin Depths

$$F_r = \frac{R_{ac}}{R_{dc}} = \frac{X}{2}\left[\frac{\sinh X + \sin X}{\cosh X - \cos X}\right] \qquad \text{(Eq G)}$$

This equation may readily be evaluated with a spreadsheet. Most spreadsheets have both circular and hyperbolic functions.

Table 8—Loss due to conductor resistance for dipoles using #12 AWG solid copper wire

Frequency (MHz)	δ (mils)	*X*	F_r	*L* (feet)	R_{eq} (Ω)	*Loss* (dB)
1.84	1.92	29.8	10.8	267	2.29	–0.13
3.75	1.35	42.5	15.3	131	1.59	–0.09
7.15	0.98	58.7	21.0	68.8	1.15	–0.07
14.2	0.69	82.7	29.5	34.6	0.81	–0.05

of a wire. Although the skin depth will be very small, the large surface area should help greatly. I would be more concerned with the joints between tower sections, particularly in high-current regions. This problem has been addressed by attaching copper wire jumpers across tower joints. The problem will be much worse in crank-up towers, where the sections have sliding joints between them.

I would be more concerned about loss in a steel tower if it were being used as part of an array with low impedances, especially if the tower is electrically short and heavily loaded. In that case, I would consider installing a collar at the top of the antenna and attaching several parallel copper wires in a cage around the tower from top to bottom. This way the copper is the conductor not the tower. This allows the tower to be grounded directly but still have the feed point open. If the collar were made significantly larger than the tower, then it would not only reduce loss but also increase bandwidth and reduce the loading necessary because of the larger effective diameter of the antenna.

Sometimes the guy wires on a tower or the rigging on a sailboat are used as antennas. Depending on the antenna, these can be very lossy and should be

Why is there Skin Effect?

When a time-varying current flows in a conductor, a time-varying magnetic field will be created around the conductor. A simple example is shown in Fig C. A current flowing in the wire creates a magnetic field around the wire as indicated. The direction of the magnetic field in relation to the current obeys the "right-hand rule"—that is, if the thumb of your right hand extends in the direction of positive current flow as shown, the magnetic field will curl around the wire in the same direction as your fingers.

Just as a current creates a magnetic field, a time-varying field, from some external or internal source, will induce a time-varying current in a conductor. This is called an "eddy" current and higher frequencies yield greater-amplitude eddy currents in a given conductor. The direction of the eddy current is such that its magnetic field opposes the inducing field.

We can see how these currents and fields create skin effect by examining Fig D. This is a section of a round wire carrying a current from one end to the other. This current is labeled "A." It is simply the net current flowing through the wire. This current creates a magnetic field both inside and outside the wire as indicated by the dashed lines "B." This field, in turn, creates an eddy current ("C") as shown.

Notice that near the center of the wire, the eddy current opposes the desired current, but on the outer part of the wire, the eddy current aids the desired current. If we look at a cross-section of the wire, we see that the current density near the center is reduced, but near the outside, the current density is increased. As frequency increases, less current flows on the inside of the wire and more flows near the outside surface. Of course, the *net* current stays the same, but it is crowded into a smaller and smaller portion of the wire's cross-sectional area.

The result is that the apparent resistance of the wire increases because we are using only a small portion the available copper area to carry current. This means that the loss for a given current will be higher. In copper at HF, the current is crowded into a layer of 2 mils, or less, in thickness. The rest of wire only provides mechanical support for the thin outer layer that conducts!

There is another way to look at skin effect. If you have a large sheet of conductor and you irradiate it with a electromagnetic wave perpendicular to the surface, the wave will penetrate the surface for some small distance. The amplitude of the wave decreases exponentially and the depth at which the amplitude has decreased to $1/e \approx 37\%$ ($e \approx 2.718$, the base of natural logarithms) is referred to as the penetration or skin depth (δ). Increasing frequency decreases δ.—*N6LF*.

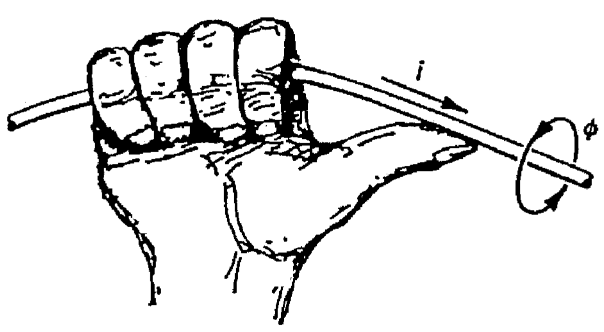

Fig C—The "right-hand rule" relates the direction of current flow to the magnetic field it produces.

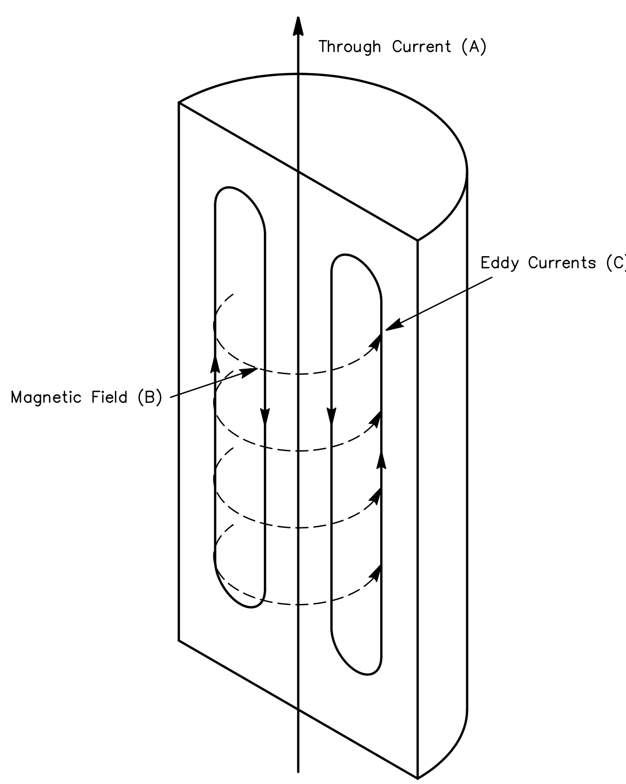

Fig D—Eddy currents in wire produce the skin effect. The through current (A) produces a magnetic field (B) that induces eddy currents (C). The eddy currents offset through current near the wire center and add to through current near the wire surface.

used with some caution. Note from Table 1 that the standard marine stainless steel (304) has a resistivity greater than 50 times that of copper. A number of years ago, I used an insulated backstay on my sailboat as a half-sloper, fed at the top and driven against the aluminum mast. To minimize the loss in the stainless steel backstay, I used a strip of copper (encased it in plastic tape to control corrosion) bent over the backstay in the form of a **U** for the distance between the two insulators. This proved very satisfactory during several years of cruising in temperate and tropical waters.

Stainless Steel and Mobile Antennas

Most mobile antennas are electrically short and heavily loaded, especially at and below 7 MHz. The result is very low radiation resistances. Because of its very high resistivity, stainless steel may not be a very good choice for these antennas despite the obvious mechanical and corrosion-resistance advantages. For example, consider an 8-foot center-loaded whip with a 0.5-inch diameter base section and a 0.125-inch diameter top section. The loss due to conductor resistance using stainless steel is 0.6 dB at 7.150 MHz, 1.3 dB at 3.8 MHz and 3.1 dB at 1.84 MHz. The use of stainless steel wire would result in losses very similar to steel fence wire.

Conclusions

For antennas with current-loop impedances above 35 Ω or so, any copper, Copperweld or aluminum wire in a variety of sizes will work just fine; however, for lower-impedance antennas, copper or Copperweld wire size #12 AWG or larger should be used. Copper or aluminum tubing is very effective for low-impedance antennas. For 80- and 160-meter antennas, the resistance of stranded Copperweld may be unacceptably high.

New insulation does not seem to affect loss, at least at 1.8 MHz, but surface oxidation does. Thin insulation should have only a very small effect on tuning but will suppress oxidation. This is a consideration for low-impedance antennas only.

By careful choice of conductor or combinations of conductors, considering both electrical and mechanical properties, it should be possible to keep the conductor loss low in almost any kind of antenna, with the possible exception of very small antennas.

Loose Ends

Despite the extensive discussion in this article, several subjects need more attention. I think the losses in steel towers need to be analyzed more closely. I also have not addressed losses from currents induced in guy and support wires. Usually these currents are small if the wire is short compared to $\lambda/2$, but steel wire can be quite lossy even with small currents. This subject needs some scrutiny. In searching through the literature, I found very little in the way of measurements or even discussion of antenna conductors. Books in the reference list contain some very useful tables, but if you know of any important articles I have missed please tell me.

Acknowledgment

I would like to express my thanks for the review and helpful comments provided by George Cutsgeorge, W2VJN. Mark Perrin, N7MQ, and Joe Brown, N7EZG, were very helpful in digging out great examples of "cruddy" wire for me to test. Tom Schiller, N6BT, told me about Alumoweld wire and related a number of hilarious anecdotes concerning antenna conductors.

Notes

1. J. Kraus, "Antenna Arrays with Closely Spaced Elements," *Proceedings of the IRE*, February 1940, pp-76-84, see Fig 5.
2. "Maxwell" by Ansoft Corporation, Four Station Square Ste 200, Pittsburgh, PA 15219-1119; tel 412-261-3200, fax 412-471-9427; e-mail **info@ansoft.com**; **http://www.ansoft.com**.
3. United States Alumoweld Company, Inc, 115 USAC Dr, Duncan, SC 29334; tel 800-342-8722, 864-848-1901, fax 864-848-1909; e-mail **sales@alumoweld.com**; **www.alumoweld.com**.
4. F. Terman, *Radio Engineers Handbook* (McGraw-Hill, 1943), pp 32-33.
5. D. Fink and W. Beaty, *Standard Handbook for Electrical Engineers*, 11th Ed., ISBN: 0070220050 (McGraw-Hill, 1999) Chapter 4.
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7. M. Abramowitz and I. Stegun, *Handbook of Mathematical Functions*, National Bureau of Standards republished by Dover Publications (New York, 1965), pp 379-385.
8. R. Straw, N6BV, Ed., *The ARRL Antenna Book*, 18th Ed. (Newington: ARRL, 1997), Chapter 20.
9. J. Devoldere, ON4UN, *Low-Band DXing*, 3rd Ed. (ARRL, 1999), pp 8-5 through 8-7.

A Prognosis of Amateur Radio Techniques

An absorbing view from the past —and a peek into the future.

By Warren Bruene, W5OLY

[This first part appeared as IRE District Conference paper DP 63-734, May 2, 1963. Now, let's step into the time machine. As we drift back 40 years many things common in 2000 fade into the distant future…—Ed.]

Part 1: A View from the Past

In order to avoid too much generality, I propose to treat this subject in three specific parts. First, we will discuss the current trends of station equipment design and operation, and extrapolate these trends into the near future. Second, we will discuss how we can make better use of the bands of frequencies we are privileged to use and a suggestion for again establishing a special-privilege class of amateur license. Third, we will turn loose our imagination a bit to predict what might be in store for us in the 1970s.

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Early in the history of radio, individual amateurs and experimenters often made very important contributions to advancing the state of the art. During the past few decades, however, the entire technological field of radio and electronics has been progressing at such an explosive rate that the individual experimenter's efforts have been dwarfed by the enormous output of industrial and educational institutions. The amateur can cull out this stream of new components and circuits and find some that can be put to excellent use.

Amateur Radio technology has progressed in a step fashion with each new major development that came along. Some of the more important of these were:

- vacuum-tube oscillators
- superhet receivers
- crystal oscillators
- class-B plate modulation
- tetrodes and pentodes
- beam antennas
- stable variable-frequency oscillators
- exploitation of VHF and UHF
- more-selective IF filters
- adoption of SSB

Others could be named, but the big one just now starting to have an enormous impact is that of semiconductors.

One easy way to predict the near-future developments is to look at the shortcomings of our present amateur station equipment and predict that these will be overcome. We must remember that some of the troubles result from improper operation. The solution here is to develop more foolproof equipment with automatic controls or develop more operator skill—or consideration for others.

Well, what are some of the shortcomings? The most important need today is for more efficient use of the electromagnetic spectrum allocated to our use. This simply means transmitting our messages so they occupy the least bandwidth for the least time and with the least interference to others.

On CW, minimum bandwidth occupancy can be achieved by stable oscillators and effective shaping of the transmitted pulses (dots and dashes). Chirps and key clicks should be virtually nonexistent. We need to adopt a standard method of measuring undesired radiation and then some standards that all transmitters should be expected to meet.

On phone, minimum bandwidth occupancy can be achieved by transmitting SSB only with transmitters employing good suppression of all emissions outside the necessary 2 to 2.5 kHz-wide voice band and reasonably low-distortion linear amplifiers. (See Fig 1.) The superiority of SSB is well established and its practicability is proven by its wide adoption. We should expect to see a cutoff date established in the near future for discontinuance of all other phone emissions in the HF bands. Amateurs themselves should take the lead in requesting such a regulation for their overall benefit.

Some standards for out-of-channel (3 kHz) emission suppression must be established so that all transmitters can be rated as good, acceptable or nonconforming. We should expect to see a regulation requiring all transmitters to employ ALC or its equivalent. The circuitry is simple and inexpensive, and it would minimize the adjacent-channel splatter or interference generated by overdriven amplifiers.

Most of the important undesired-emission measurements can be made on a strong off-the-air signal by a suitably equipped receiving station. "Official Test" stations could be designated for providing amateurs with measurements on their transmitted signals. A clear 10-kHz channel would be desirable for such measurements, and possibly designating a test channel for calling an Official Test station for purposes of emission measurement would be in order. Preferably, such testing should be done during hours when the band is not normally very busy. Amateurs with a strong interest in equipment could perform a very real service by equipping and operating such Official Test stations.

All operators can reduce their time on the air by simply learning to compose their messages to convey the desired information as briefly as possible. Some operators, particularly in the CW bands, are quite proficient in this, but a great many are not.

Fig 1 also illustrates the value of better receiver selectivity to reject interference from signals on adjacent channels. Selectivity is needed especially when several local stations are operating in the same band. Low cross-modulation and freedom from blocking are associated essential features of better receivers.

Channelization of Phone Bands

So much for cleaning up our signals, now let's look at some practices that might provide for more effective spectrum utilization. One idea that has been proposed, off and on, is the channelization of our bands. There really is no need for the ability to transmit on carrier frequencies between exact integrals of 1 kHz on phone. Practically all SSB phone signals are nominally 2-3 kHz wide so there still could be much overlap. Tuning could be simplified, however. The frequency-generating equipment in both transmitters and receivers could employ frequency-synthesizer and stabilized master-oscillator techniques. This is almost standard now in new commercial and military equipment. Two advantages could be realized. One is that much better frequency stability can be achieved easily since every carrier frequency is controlled by a single, standard crystal oscillator. The other advantage is that SSB would always sound natural since it would be relatively easy to keep within a few cycles per second of exact frequency. The process of tuning across the band would consist of stepping from one channel to the next.

The complexity and cost of suitable circuitry for the narrow amateur bands

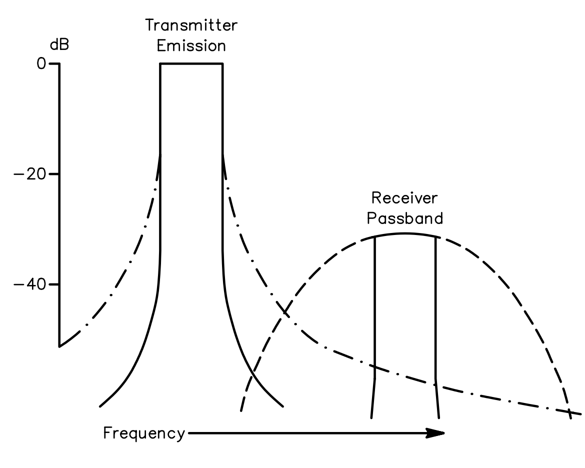

Fig 1—A graphic illustration of minimum bandwidth occupancy.

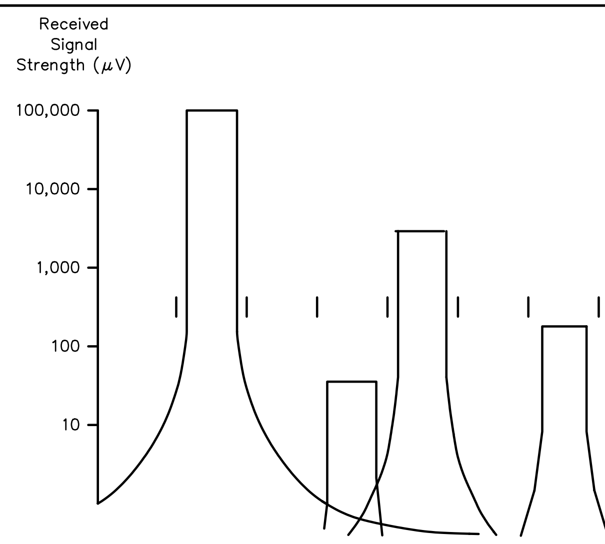

Fig 2—An example of 4-kHz phone and 200-Hz CW signal spacing.

is coming within reach now. Getting such a system started is the problem, since unless others had similar equipment capability, it would be of no value to any individual amateur.

There are some who argue that overlapping channel operation should not be permitted. For example, if SSB phone channels were assigned on a 4-kHz spacing, most interference between channels would be eliminated. (See Fig 2.) Except for extremely strong adjacent channels, the interference would be right on the same channel. If the interference is weak, you can work over it; if it is strong, you invite them into a round table, politely arrange to take turns, or move to another channel. By such means, every channel can be used as a good channel. This would help bring a little more order to the bedlam of interference often experienced now.

Equipment designed for 4-kHz channel spacings would be simpler and less expensive. Scanning over all channels in the band would be easier since there would be only 25 channels per 100 kHz of bandwidth. I suspect, however, that a compromise of 2-kHz channel spacings would provide better spectrum usage.

Code Transmission (CW and RTTY)

With the exception of a few RTTY stations, all amateur code transmission is by on-off keying of a carrier wave. A certain proficiency in sending and receiving the Morse code manually is a prerequisite to obtaining an amateur operator license. Many operators take pride in their CW proficiency and thoroughly enjoy this mode of communication. Their sending is clean, crisp and easy to copy at good rates of speed. Their messages are brief and concise; they truly make comparatively efficient use of the spectrum. Electronic keyers now being used are a very valuable aid for achieving more nearly perfect sending. Slow, sloppy sending combined with a broad signal easily can occupy more "kilohertz-minutes" of spectrum space for the same amount of message exchange as a phone station.

It has been my contention that the next major improvement in amateur communications must come in the area of coded transmission. The commercial industry has left the amateur far behind. In World War II, the superiority of FSK over on/off CW was proven. A system of 850 Hz between mark and space was adopted for radiotelegraph RTTY operation. This is used now by most amateur RTTY stations. Since commercial and military users have had to cram more traffic into their channels, they had to devise ways to do this. The most widely used system now transmits 16 narrow-band FSK channels within a normal 3-kHz voice channel. The channels are spaced 170 Hz apart starting with 425 Hz. The mark and space frequencies are ±42.5 Hz from the channel center frequency. Keying speeds of 60, 75 or 100 WPM in each of the 16 channels is used.

Another type of binary transmission uses phase-shift keying. A system of transmitting two RTTY channels on each of 20 tones within a 3-kHz voice band has found limited use. Thus, forty 100-WPM transmissions, or their equivalent in single-channel data, can be packed into the space used by one SSB voice channel.

Matched filtering (sometimes called *predicted wave* signaling) is another useful technique for increasing solid copy of weak signals. The transmitting and receiving equipment must be synchronized. The basic idea is that the detection system looks in both the mark and space channels during the period of time when the next character bit is to arrive. It then compares the coherent energy received in each to decide if it was a mark or a space. This technique makes it possible to copy signals that are almost obscured by noise to the human ear. If some related system were found advantageous to amateurs, the 60-Hz ac power system might be used to time the detector's gating periods. It even may be feasible to adapt matched filtering to manual keying using synchronized electronic keyers. The receiver detector would key an audio oscillator for manual copy of the Morse code. Noise on interfering signals would be eliminated, but errors might occur in the regenerated CW characters.

It is time for the amateurs to consider all of the techniques and systems employed in the communications industry and evolve one best suited for them. Once the optimum system is found, it should be voluntarily adopted as a standard by amateurs.

The ARRL should take the lead in this matter. They could define the basic objectives and then invite proposals from everyone including equipment manufacturers and all interested parties, even though they may not be licensed amateurs. The proposals would be evaluated and sifted down to a few for performance tests and evaluation of equipment cost and simplicity. The procedure could be patterned after that used by the FCC in finding and adopting a standard for FM stereo broadcasting.

Diversity

Amateurs have done nothing to overcome fading of CW signals in the HF bands. The use of diversity has been found essential for reliable circuits in the communications industry. (Note: This problem is overcome in phone transmission by the use of SSB.)

A signal received by two antennas spaced a few wavelengths apart seldom fades out at the same time on each antenna. The system based on this principle is called space diversity. This probably is not very usable by amateurs because most do not have enough real estate for one good antenna system, let alone two.

Frequency diversity uses the principle that two separate signals, spaced 1 kHz apart or so, seldom fade out at the same time either. This means transmitting two signals so the receiver can select the strongest one automatically. This system occupies twice the spectrum space but, even so, it may be more efficient in its use of spectrum in kilohertz-minutes than the present amateur practice of asking for repeats and sometimes even calling to re-establish contact. A standard diversity technique also should be derived out of the previously proposed search for a new amateur code modulation standard.

RTTY Code

It would be wise to examine several codes to find one that may be better for amateur use than the present RTTY code. The military soon will change to an 8-bit code, which has a parity check built into it so that if one bit of a character is received wrong, if prints an error symbol instead of an incorrect character. Perhaps this or another code would be useful for amateur RTTY use.

Antennas

Beam antennas have been found quite valuable. A great deal of attention has been paid to forward gain but the directivity is probably of greater value. Concentrating the transmitted signal in the direction of the other station reduces interference to others located in other directions. (See Fig 3.) Directional receiving antennas reject interfering signals arriving from directions other than the received signal. The result is more QSOs on the same channel.

An increased effort in achieving more antenna directivity is in order.

Forward gain should be secondary. Even an antenna with no forward gain but with high directivity would be valuable. It would dissipate the power that would be radiated in other than the desired direction and thus reduce interference to others. As a receiving antenna, it would reject many unwanted signals. Such an antenna might be a remotely tuned, high-Q arrangement that also could provide some selectivity ahead of a receiver. Quite probably a physically much smaller antenna would result.

Extra Privilege License

So far, we have discussed some modulation systems and techniques for more efficient use of our assigned portion of the spectrum. Now I'm going to propose a method of getting these improved practices into operation. The main new requirement is for somewhat better equipment. It makes little difference if the amateur build his own or buys it. The important thing is the characteristic of the transmitted signal. Therefore, I propose a special-class license for amateur stations (not operators) that meet certain prescribed standards. A small section of each band would be reserved for these special-class stations. To start with, 25 kHz would be enough, and I suggest it come from existing CW portions of the 3.5-, 7.0- and 14-MHz bands. Here would be a chance for the progressive amateur who maintains good equipment to get out of the present bedlam and meet others with progressive interests. These interests may lie in the fields of equipment technology or efficient traffic handling. Stations so equipped undoubtedly would be more valuable for emergency and disaster communications.

As more and more amateurs upgrade their stations to this special class, the band reserved for them could be widened. This could be done strictly based on effective band utilization. When the special stations achieve, for instance, 50-100% more communications per kilohertz of bandwidth, it would be time to give them more room. I do not propose to abolish present operation completely because there must be space that can serve as a training ground for those entering the amateur ranks and for those who, for various reasons, do not want to give up the practices they enjoy.

I have left a number of loose ends dangling in order to maintain better continuity in arriving at my main points. I hope to pick up some of these as we go on to talk about what amateur stations might consist of in the future.

Satellite Communication

Plans for OSCAR III are in the making. It will be an active satellite on the principle of Telstar, and it will operate in the 2-meter band. It is proposed simply to receive a 50-kHz band of frequencies and retransmit it in another 50-kHz portion of the band. (See Fig 4.) It will be a linear system with about 1 W PEP transmitter power. Thus, any type of transmitted signal can be relayed, such as FM, AM, SSB, CW and so forth. A continuous beacon signal also will be transmitted for locating and tracking the satellite. This will offer an opportunity for some advanced-level experimentation by amateurs and a means of getting long-distance communication at other than the HF bands.

Solid-State Devices

Transistors, diodes and a growing variety of solid-state devices are just beginning to make an impact on amateur equipment. Associated with these devices are all sorts of new circuit possibilities. Even computer technology will have an important impact.

The impressive blue glow of the 866A is now obsolete. Silicon rectifiers already have replaced tubes in power supplies almost completely. Electronic

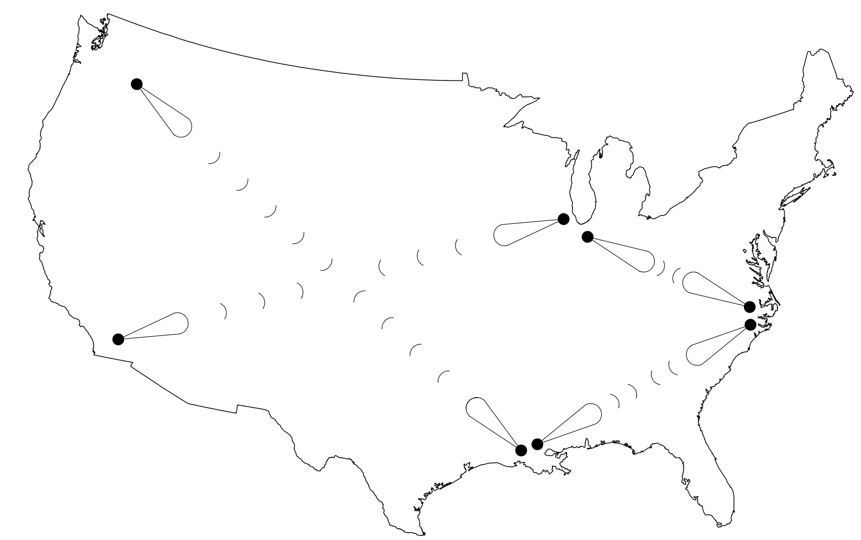

Fig 3—Examples of directivity (right) and "skip-help" (left) signal separation.

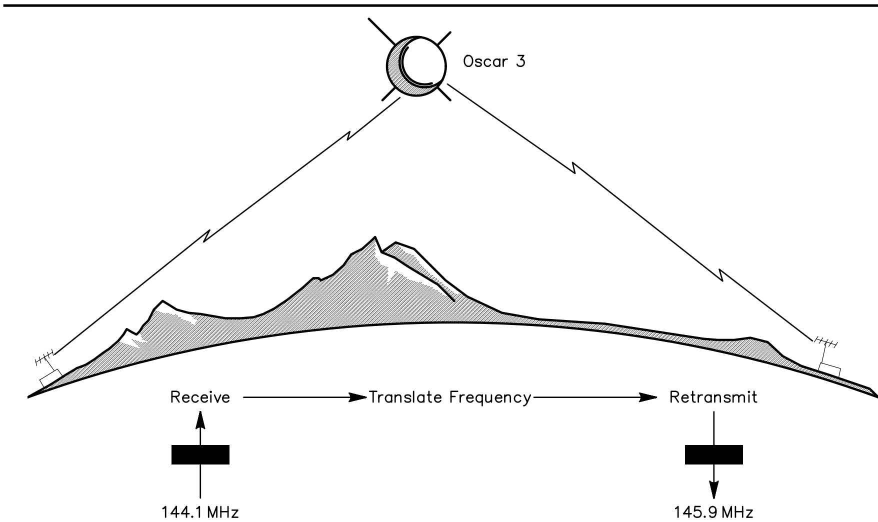

Fig 4—OSCAR 3, an early "active" (as opposed to passive reflectors or beacons) amateur satellite.

keyers are replacing the old mechanical "bug." They are becoming solid-state and employ logic circuits developed by the computer industry. Some are quite refined already, but we can look forward to their continued improvement, cost reductions and much wider usage.

The popularity of transceivers will continue to grow. The next major generation of equipment will use no vacuum tubes except perhaps in the power amplifier stage. Early attempts to replace tubes with transistors were only partly successful. The basic circuits must—and will—be changed to exploit the characteristics of transistors and other solid-state devices to better advantage. The previous shortcomings in the performance of transistor receivers and exciters can be overcome already and soon we can expect transistor equipment to top the performance of today's better tube equipment.

In a few more years, we can expect to see even the power amplifiers of transmitters transistorized. Transistor power amplifiers have been built that deliver kilowatts of power up to 2 or 3 MHz. They employ a new switching mode of operation that is much more efficient than class-C operation of regular transmitting tubes.

The power and frequency ratings of transistors are being increased at a rapid rate. Already a transistor has been produced than can deliver 700 W at 20 MHz! By sometime in the 1970s, the complete "transistorization" of transmitters up to the legal limit should be practical and commonplace.

Relays and switches will start to disappear. Except for those associated directly with operator controls, they will be replaced substantially by diodes or other solid-state switching. A commercial VHF airborne receiver already exists that is remotely tunable to many crystal-controlled frequencies. There is not a relay or switch in it, except the small remote-control head in the pilot's compartment.

We will see another major size reduction in equipment. In fact, soon it will reach the point where some smart human engineering must be employed to work out practical controls operable by human hands on the smaller panel space available.

Power consumption will be reduced greatly. The transmitter will be so much more efficient that complete battery-powered operation will be possible for 100-W transmitters.

Universal power supplies will be developed that provide more-convenient operation from 120-V ac, 12-V dc automobile systems or portable batteries. Size and weight will be reduced still more.

The above features made possible by solid-state devices will greatly improve the portability of amateur equipment. A large increase in mobile and portable operation will result.

RTTY Machines

In my previous discussion, I often referred to machine coding and decoding such as RTTY. For this to become more popular, we need a machine that is more of an adaptation of a regular typewriter. It should be capable of generating a coded signal from the keyboard and simultaneously transmitting the message. For reception, it would decode and type the received message. In addition, the machine should be useful as a conventional typewriter.

Use of regular type with both upper and lower case letters, as well as all standard typewriter symbols, would be much nicer than present RTTY. Some of the newer electric typewriters now available are designed for more simple electric control. It should not be too difficult to work out the details of the associated "little black box" that completes the coding and decoding needed for connecting to the transmitter and receiver. The practical design of the circuitry for this little black box will have a large bearing on the selection of the optimum new RTTY code standard for amateur use. This box would be an all solid-state device and not too complex for amateur use.

Facsimile and Slow-Scan TV

It is already possible to send pictures using slow-scan TV within a 3-kHz voice channel. I predict, however, that some kind of facsimile will find wider adoption. If it is of "wirephoto" quality, it could be used to transmit one's photograph. Maybe transmission of a caricature line drawing would be a good alternative.

A fax circuit would be very valuable in exchanging circuit diagrams and equipment construction details. I'm certain it would find many other uses also. One use might be the transmission of QSL cards, possibly even in color.

Computer Control

If we let our imagination run a bit to the "George Ramjet" age, we might visualize a computer-run station such as that shown in Fig 5. It would be completely automatic in operation. It would scan the channels for a CQ and automatically answer it after checking its memory to see if the calling station had been contacted recently. It could be programmed to exchange "canned" messages of the same level of intelligence as many of the routine CW exchanges today. Of course, it would print out the station log automatically. It would collect fax transmitted QSL cards from DX stations.

Wouldn't a rig like that be a real "lash-up" for a sweepstakes contest? Can you imagine what the bands would

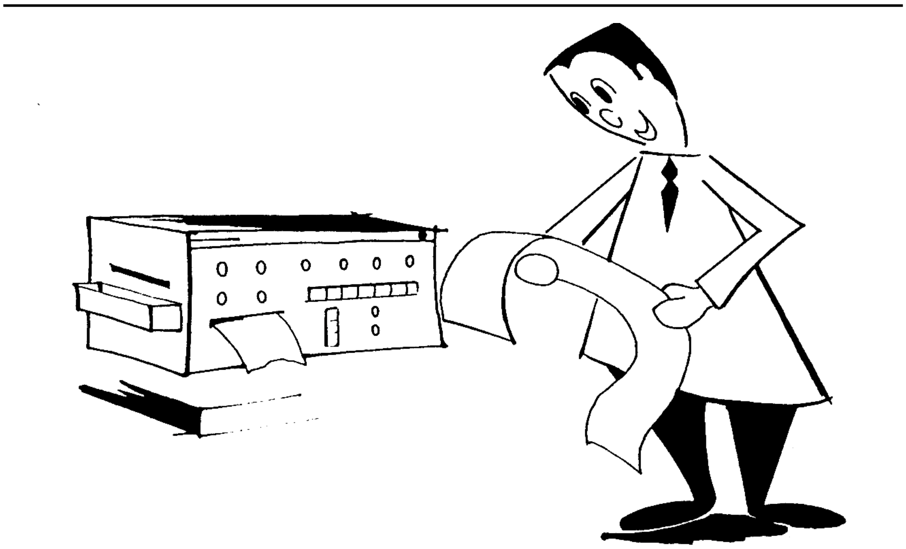

Fig 5—Ham George Ramjet checks how much DX his new automatic, computer controlled ham station worked during the night.

sound like with a country full of machines madly trying to outdo each other? I'd bet it wouldn't be long before someone demanded that such a computer control have an operator's license. All this may sound a bit fantastic, but it is all easily within the realm of today's technology. The only thing fantastic would be the price tag!

Conclusion (Part 1)

With the exception of George Ramjet's amateur station, everything discussed may become quite commonplace quite soon. I have not attempted to explore the possibilities of new breakthroughs. Everything is based on the assimilation and utilization of today's advanced technology.

Before concluding, however, we should remind ourselves that our amateur bands are a public resource and that our use of them should serve the public good overall as well as our own enjoyment. In general, we can say that our avocation is a good training ground and a means of interesting many youngsters in the fields of electronic science and technology. In addition, by maintaining stations with good, solid communications capability, amateurs can provide, with some organization, a very effective service to their community when needed. The anticipated widespread use of these things we see coming soon will contribute to greater enjoyment of our hobby by more people as well as equipping us to be of more public service.

Part 2: An Update

[Now, forward to May, 2000.—Ed]

While rereading my previous prognosis of 37 years ago, I noted that quite a few of the things predicted have happened. Recall that was 19 years before the IBM PC was introduced. Witness the following:

- solid-state transceivers
- smaller and more portable equipment
- narrow-band FSK
- coherent CW
- PSK
- satellite communications
- slow-scan TV
- tuned-loop antennas
- stable frequency synthesizers
- substantial channelization of SSB on 1-kHz steps
- computer control

Thank goodness, we don't have a country full of automatic stations trying to outdo each other—although, with the computer aids contesters use we are getting close on contest weekends.

The technology advances to increase spectrum utilization have been rather modest, perhaps because the number of hams on the air hasn't increased very much. Back then, we didn't dream of the wide use of PCs and the Internet.

Commercial and military use of HF has matured and microwave, satellite, fiber optics and cell phones have taken over most of the exploding need for increased communication capacity. Cell phones are now the fastest-growing communication means. HF has been largely relegated to a backup mode for mobile and long-distance communications. It takes a fair amount of operator skill and antenna knowledge—skills that many hams have.

Thus, hams can fulfill an important backup to other forms of communication in times of disaster as well as for civic functions such as parades, walks, marathons and other events. Some hams debate radio technology over the Internet, probably because they can post messages that others can access at any time.

Computers and Computer Technology

It may surprise readers to know the Collins Radio Company developed a computer-controlled, 250-kW shortwave broadcast transmitter back in 1966, 15 years before the IBM PC. One computer system controlled three transmitters. The entire day's schedule was programmed into the computer. About 15 seconds before a move to a new frequency, the computer would shut the transmitter down. It would then switch antennas, command a servo system to tune the transmitter to the new frequency, switch audio lines (perhaps for a different language) and bring the transmitter up to full power. The system was ready to receive program material—all within 15 seconds. There were no meters on the transmitter since the computer monitored all "meter readings." Computer memory was very precious because it consisted of many small magnetic toroids threaded in a wire matrix. One core was required for each bit of memory storage. The whole system cost approximately half a million dollars—the equivalent of several million today. Now, some 34 years later, hams can buy a computer-controlled transceiver off the shelf.

Digital signal processors (DSPs) based upon computer technology will rapidly replace the modulators and demodulators in ham transceivers. The digital filters can have steeper "skirts" without the spike of time delay near the band edges that tends to cause "ringing." Furthermore, bandwidth can be easily altered.

Modulators and demodulators will be developed that may be software-programmed for "any" kind of modulation. Developing a new modulation scheme, for example, will be done with software instead of hardware. PSK31 is an example of what is coming.

I expect to see a special-purpose computer designed especially for amateur station use. It will be assembled mostly from standard PC parts with some added boards. These boards may contain A/D and D/A converters, frequency synthesizers and DSPs. Receiver front ends and, of course, the RF power amplifiers will remain external. A built-in spectrum analyzer will allow observation of propagation conditions, nearby signals and the transmitter's signal. A special computer operating system will be developed that will be much simpler than *Windows* or even DOS, but that will have the basics needed with room for expansion. It will contain simple word-processing and printer-control capability. It will have better immunity from conducted and radiated RF than today's machines.

Better quality SSB speech will be optional. Instead of just one narrow filter bandwidth and response shape (with pre-emphasis) for optimum intelligibility, wider filter passbands will be selectable. In addition, the passband shape will be standardized so the receiver can flatten it back out, restoring naturalness, while retaining an improved SNR.

Accurate receiver tuning will remove the "duck" sound of off-tune SSB. A means of detecting the frequency error from the speech signal itself has already been developed.[1] It could be used to automatically adjust the receiver tuning, or to indicate the direction and amount of error so that the operator could quickly and accurately adjust the receiver's tuning manually.

A much easier way is to simply increase adoption of channelization of SSB on 1-kHz steps. Then, the receiver synthesizer tuning can be set for 1-kHz steps to accurately step from one channel to the next. Transceivers should be kept adjusted to WWV within a couple of hertz.

There will be wider acceptance of

[1]R. Dick, "Tune SSB Automatically," *QEX*, Jan/Feb 1999, pp 9-18.

machine coding and decoding. I would prefer the ASCII code that computers use and a choice of three speeds: for slow typists, fast typists (that closely follows the typing) and transmission of stored messages or documents. In addition, I would prefer MSK for a constant-envelope signal and a faster bit rate (within the same bandwidth), even with the extra bits required for synchronization.

I expect to see a much more efficient means of generating and amplifying SSB signals. DC power input should not exceed 150% of the RMS RF power output. I expect to see greater use of tuned-loop antennas. They will automatically track receiver tuning and be always ready for transmitting. Their narrow bandwidth will reduce receiver overloading and cross-modulation from signals of very strong nearby stations. Such antennas could be combined with a vacuum-tube linear amplifier that uses a separate broadband filter (output network) for each band. This would eliminate all manual transmitter and antenna tuning.

As a former designer of high-performance HF linear amplifiers and developer of RF-feedback circuits for IMD reduction, it distresses me that ham amplifiers have worse IMD today than they had 40 years ago. I realize that transistors may require different techniques to decrease distortion, but there is no doubt in my mind that much more can be done than is being done. I hope there will be more progress in this area in the near future.

I would like to see a simple, low-cost device developed to measure the adjacent-channel emissions of ham transmitters. It could also measure the performance of strong signals received over the air. The FCC has rules limiting these "spurious" emissions outside the necessary bandwidth. This instrument would measure how well we comply.

The popularity of QRP will continue to grow for two reasons. One is to reduce RFI in congested neighborhoods. The other is that the transmitters and receivers are simple enough for hams to revive the art of building their own—from kits, if not from scratch. The use of digital modulation schemes using PSK or MSK (such as PSK31) will allow lots of good, solid QSOs with QRP power, even with hidden antennas.

I do not expect that there will be much interest in spread-spectrum on HF, but I wouldn't be surprised to see hams modifying digital cell phones for use on some of the UHF bands.

Conclusion

Although the uses of HF have substantially matured, the technology will continue to advance. Hams will have many opportunities to experiment and achieve better and more enjoyable communication. They will keep up with the state of the art and continue to be in a position to serve their communities in emergencies.

Warren Bruene has been licensed since 1935 with calls of W9TTK, W0TTK and W5OLY. Three widely used circuits he originated are tetrode neutralization, RF feedback to improve linearity and a directional-wattmeter circuit. The wattmeter circuit, published in April 1959 QST*, is the basis for most wattmeters used by hams today. In addition, he has been grated 22 patents. He co-authored* Single Sideband Principles and Circuits *(McGraw-Hill, 1964) and authored the chapter on "High Power Linear Amplifiers" in* Single Sideband Principles and Systems *(McGraw-Hill, 1987),also in* HF Radio Systems and Circuits *(Noble Publishing, 1995) as well as single chapters in several engineering handbooks. He is a graduate of Iowa State University, a member of ARRL and a Life Fellow in the IEEE. His Fellow citation was for "advancing SSB radio communications." He spent 44 years with Collins Radio (Rockwell), where he designed the Collins 30K-1 amateur transmitter, the 30S-1 linear amplifier and many commercial, military and broadcast transmitters with power outputs ranging from 500 W to 250 kW for frequencies from VLF to UHF. These transmitters included many "firsts" in RF power-amplifier design, output-network design and automatic tuning systems. He spent another six years with Electrospace Systems before retiring to part-time consulting. He received the "Engineer of the Year" award from the Preston Trail Chapter, Texas Society of Professional Engineers in 1975. In 1982, he received an "Engineer of the Year" award from Rockwell International. He is listed in* Who's Who in Engineering *and* Who's Who in America.

A Simple UHF Remote-Control System: Pt 3

Just for fun: So far, these projects have involved only one-way communication. Now, let's look at two-way communication to offer you many hours of radio fun.

By Sam Ulbing, N4UAU

When I told my friends on the CW net that I was working on a remote keyer, they laughed and said they figured I just wanted to be able to QNI without having to get out of bed. Not a bad idea, but that meant more design work. The keyer in Part 1 [1] would only send; I needed to be within hearing distance of the rig to copy incoming code. The remote speaker of Part 2 [2] could not work with the remote keyer of Part 1, because they operate on the same frequency and the RM transmitter is always transmitting. I could use one or the other but not both at the same time. A transmitter on a different frequency would help; but the only ones available were out of the ham band or on 433.92 MHz. The answer: multiplex the transmitter and receiver. By using a microprocessor to control the transmitters and receivers, we can do two-way communication on the same frequency.

This meant setting some priorities. I want to listen to everything from the HF rig but to send whenever I feel the urge, which meant the HF side should have priority to send, within some limits. Recall the ham who keeps his PTT down and never lets it up to listen. If I didn't design the circuit right, the UHF transmitter at the HF rig would do the same thing, and the keyer could never get a word in edgewise. On the other hand, if the rig-audio transmitter shuts down occasionally, I could get annoying gaps in the signal copy. Fortunately, solving this dilemma with CW is fairly easy.

At 50 WPM—very fast CW—a dot lasts 24,000 µs, and it is followed by a space lasting for 24,000 µs.[3] From the Linx data sheets, the transmitter can be turned off in 10 µs and the receiver takes only about 40 µs to respond to a signal from the remote keyer. The microprocessor, with a color-burst crystal, can do an average instruction in 6 µs, so there was plenty of time for the microprocessor to turn off the transmitter,[4] switch the antenna to the receiver, turn it on and listen for a signal from the remote transmitter during a space. With no remote signal, the transmitter continues sending HF

[1]Notes appear on page 44.

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audio to the remote receiver. If a signal from the keyer is detected, the HF audio stays off, giving the remote key priority.

The circuit stays in the keyer mode until there is no keyer signal for one second, then it resumes sending audio from the HF rig. Because CW is so "slow," even at 50 WPM, it is possible to do all this with no effect on the CW. I can listen to everything that comes in and still send to the HF rig whenever I want to. Of course, if the guy at the other end doesn't stop sending, he won't hear what I am sending but that's always been a problem. My program also allows for static that is so bad that there are never any spaces (which means that you wouldn't be copying any intelligible code). If no space is detected in the HF audio within 10 seconds, it will stop sending and listen for a remote signal before continuing. At least it lets you send, "Sri no cpi qrn qrt, 73."

Base Station

Fig 1 shows a block diagram of the "base station" located at the HF rig. Figs 2, 3, 4 and 5 [5] show the circuit of the base station. It is more complex than those of the first two projects.

There are several reasons for this. The microcontroller has to control three areas: the T/R switch, the output- and the input-signal routing. The fact that the signal from the rig has noise also adds to the complexity. The CW signal

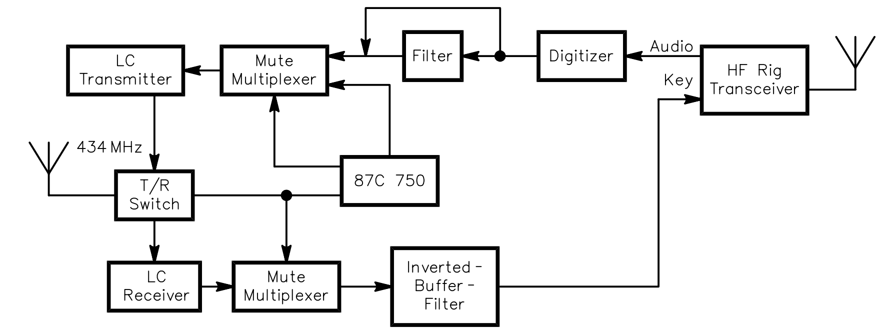


Fig 1—A system block diagram for two-way keying/audio communication on a single frequency.

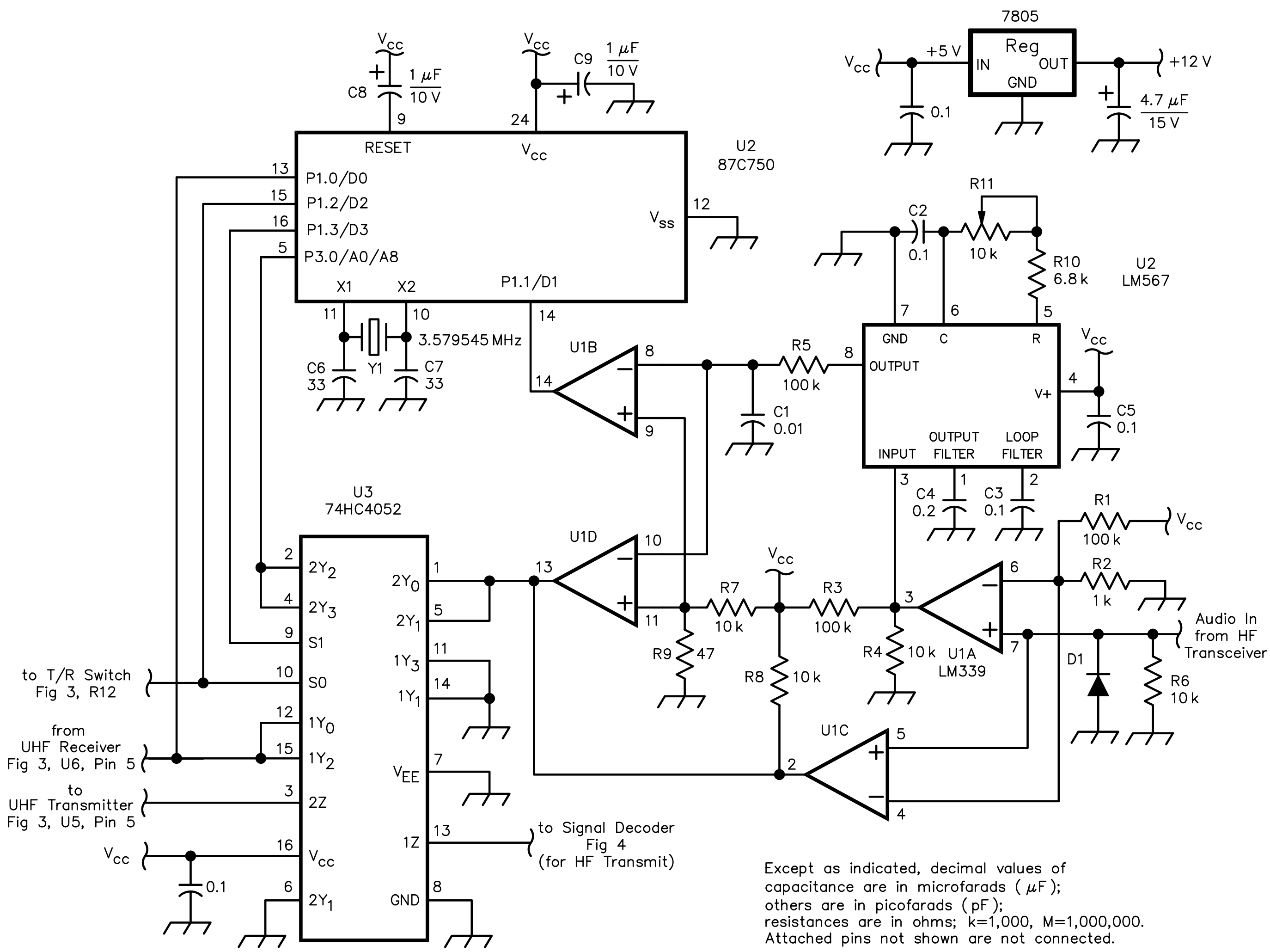


Fig 2—Schematic of the base station: from the HF audio to UHF transmitter, including the transmit/receive multiplier section.

from the HF rig is not just a "carrier-present/carrier-absent" signal. There is noise of varying levels across the receiver passband. On a bad day, the noise can be as strong, or stronger, than the desired signal. On good days, the noise level is very low—but noise is still there. The first design challenge is to hold the HF-audio transmitter's data input (U5, pin 2) high when CW is present but keep it low when only noise is present.

As you might guess, it is impossible to eliminate all the noise, but the circuit I used with "Improved Uncle Albert's Keyer" has shown itself to work very well for this purpose.[6] It is essentially the circuit I used in Part 1, with a few modifications. Recall that the CW tone we listen to on an HF rig is in fact an audio tone, usually about 750 Hz, that is switched on and off. This audio signal is routed through U1A and U2 for filtering.

This circuit eliminates noise in two ways. U2 contains a bandpass filter that passes 750-Hz signals and rejects others. With a bandwidth of about 100 Hz, U2 eliminates a lot of the noise. R1 and R2, at the input of U1A, reduce noise inside the passband. An audio signal must be greater than the threshold value set by them (they work as a voltage divider) to trigger the comparator, sending the signal to U2. A threshold value around 50 mV worked well before, so I used it again.

Passband noise rejection is altered by using both the volume control and the RF-gain control on your HF rig. Set them so that most noise is below the threshold, while the signal is above it. I have found that adjusting the RF gain can yield significant improvements in noisier conditions. U1A also conditions the signal for processing by U2. For best operation, U2 requires a signal of constant magnitude. The output of U1A is a constant-amplitude square wave with its magnitude set by R3 and R4 at about 0.5 V. U1D buffers the output of U2 and then routes it to the multiplexer input. U1B routes the signal to the microcontroller so that it will know when a CW signal is present. The purpose of U1C is described below.

U3 is a dual-switch multiplexer used by pins 15 and 16 of the microcontroller to control the flow of the HF and remote signals. When no remote signal is present and an HF signal is present, pin 3 is connected to pins 1 and 5 and the modified audio from the HF is sent to the LC transmitter. At the same time pin 13 is connected to pins 11 and 14 (ground) to prevent the signal from being retransmitted by the HF rig. When the ID is sent every 10 minutes, pin 3 is connected to 4 and 5 while pin 13 remains connected to ground. To retransmit the signal from the remote keyer, pin 13 is connected to pins 12 and 15 and pin 3 is connected to 2 and 4.

The T/R switch I used is shown in Fig 3. It is based on the design shown in Fig 6A, but you could use any of the others for your version. Mine has the advantages of drawing no power when receiving and little current when transmitting, but it is bulky because it requires a λ/4 phasing line between the antenna and receiver.

When in the transmit mode, pin 15 of the microcontroller goes high turning on Q1, Q2 and Q3, which places –5 V at the junction of R18, C15 and L1. The negative voltage forward biases D2 and

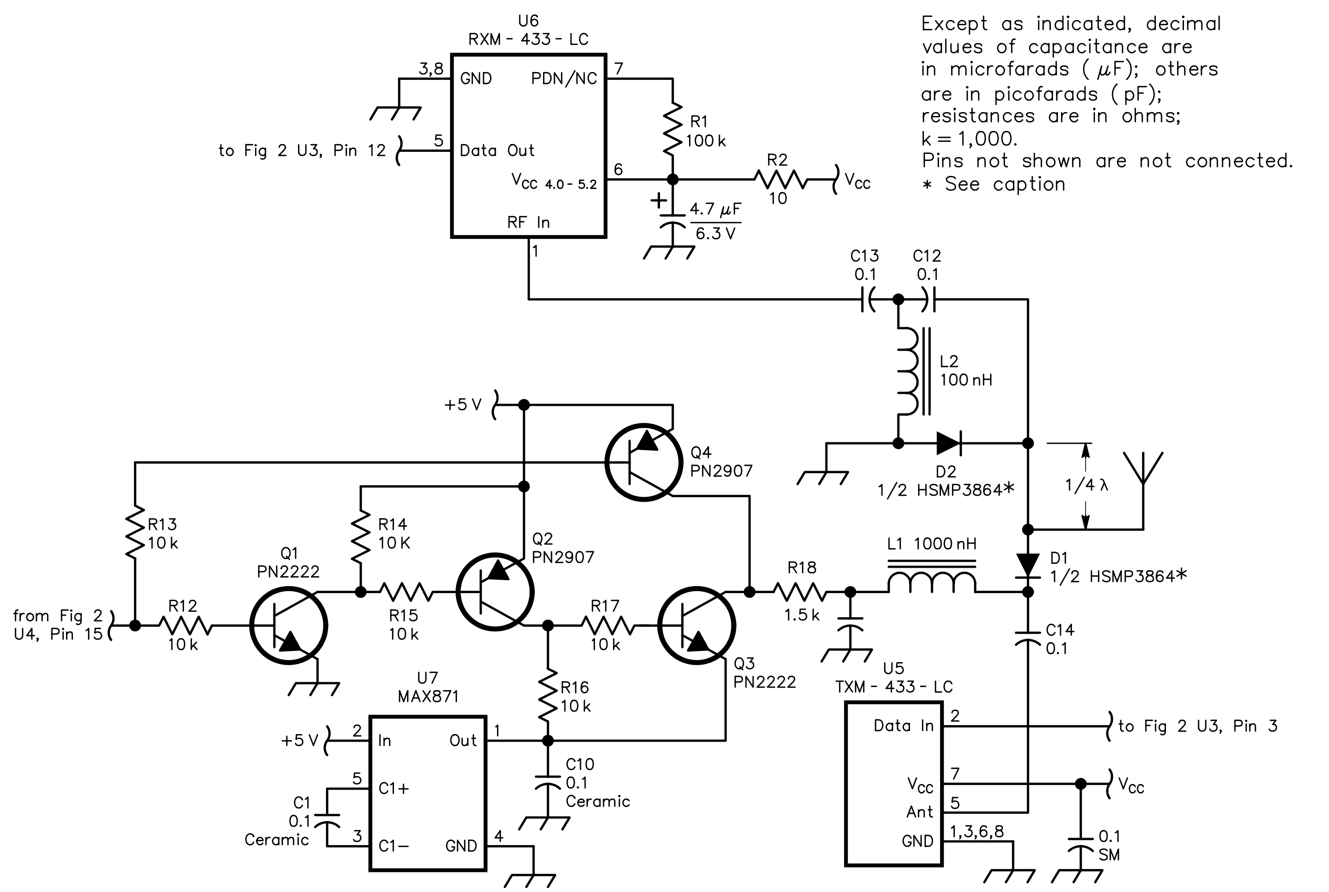


Fig 3—The filtered-audio-driven T/R switch circuitry. D1 and D2 are not two halves of the same HSMP3864; use one-half each of two separate ICs.

D1. Because D1 is forward biased, the transmitter is connected to the antenna and because D2 is forward biased, the input to the receiver is connected to ground protecting it from overload. The λ/4 line between D2 and the antenna makes the short look at the receiver input look like an open circuit at the antenna base. Thus, the transmitter sees only the antenna impedance.[7]

While receiving, pin 15 of U4 is at 0 V. Q1, Q2 and Q3 are off and Q4 is on, putting +5 V on the switch output, which reverse biases the diodes. The transmitter is isolated and the receiver is connected to the antenna. Notice that it takes three transistors to control the negative voltage. They are necessary for 0 to +5 V signal to control a –5 V line.

In Part 1, we saw that it is necessary to send modulated CW to the remote receiver in order to avoid noise. Rather than have the microcontroller send CW modulated by a fixed-frequency, this circuit modulates the CW by the actual input-modulation frequency. Input to the LC transmitter comes from two comparators, U1C and U1D. If I had used only the output of U1D, the signal would be pure CW, not modulated CW. The output of U1C varies at the frequency of the HF audio signal. By ANDing these outputs, we send the actual input CW that has been filtered to reduce noise. This approach provides significant benefits over sending the

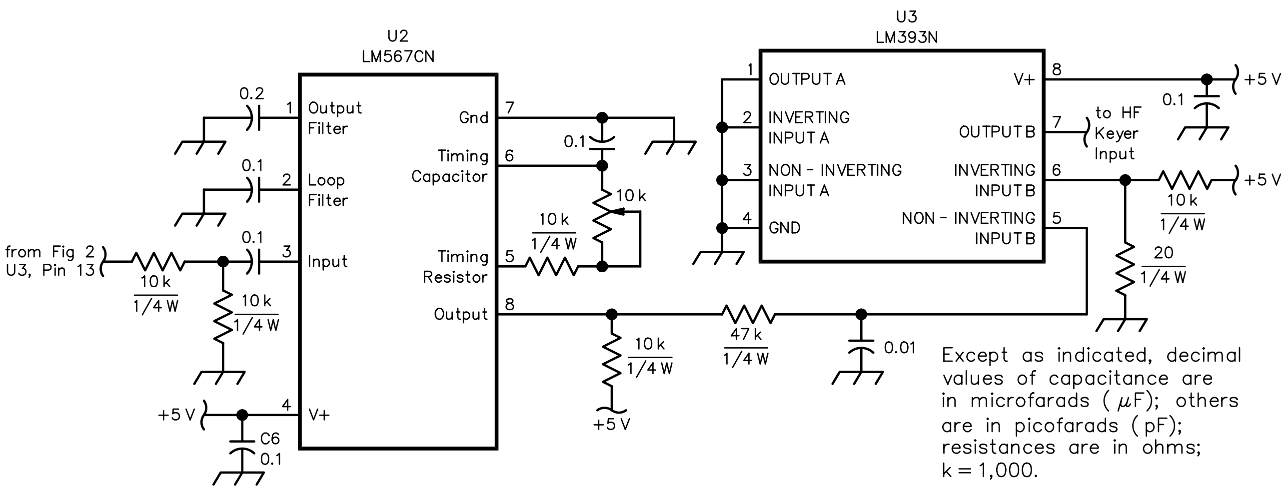


Fig 4—Schematic of the base station: The filter and buffer sections decode 700 Hz CW from the UHF receiver for transmission on HF (based on Fig 5 from Part 1, see Note 1).

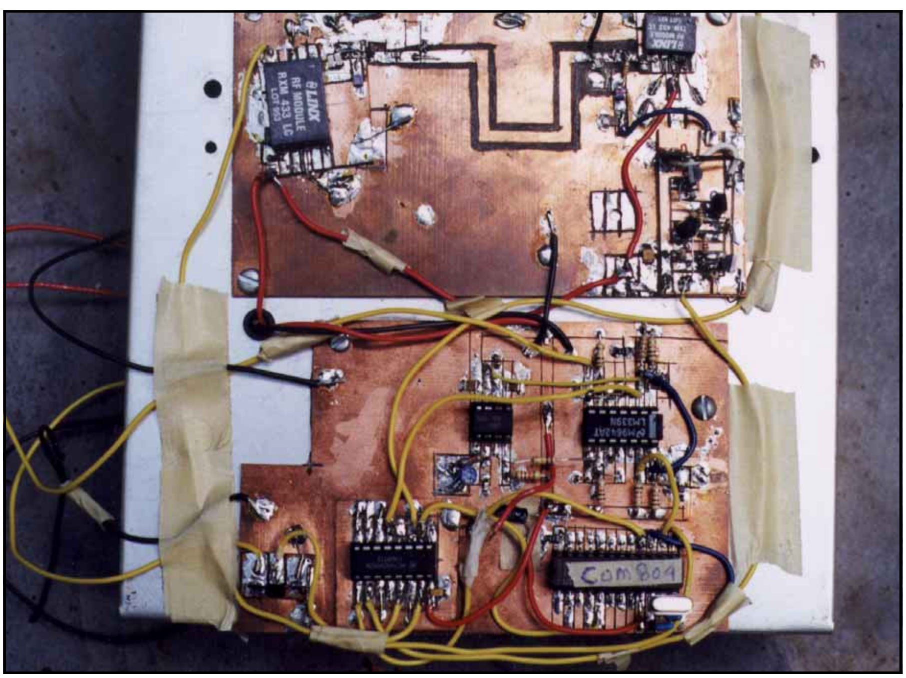
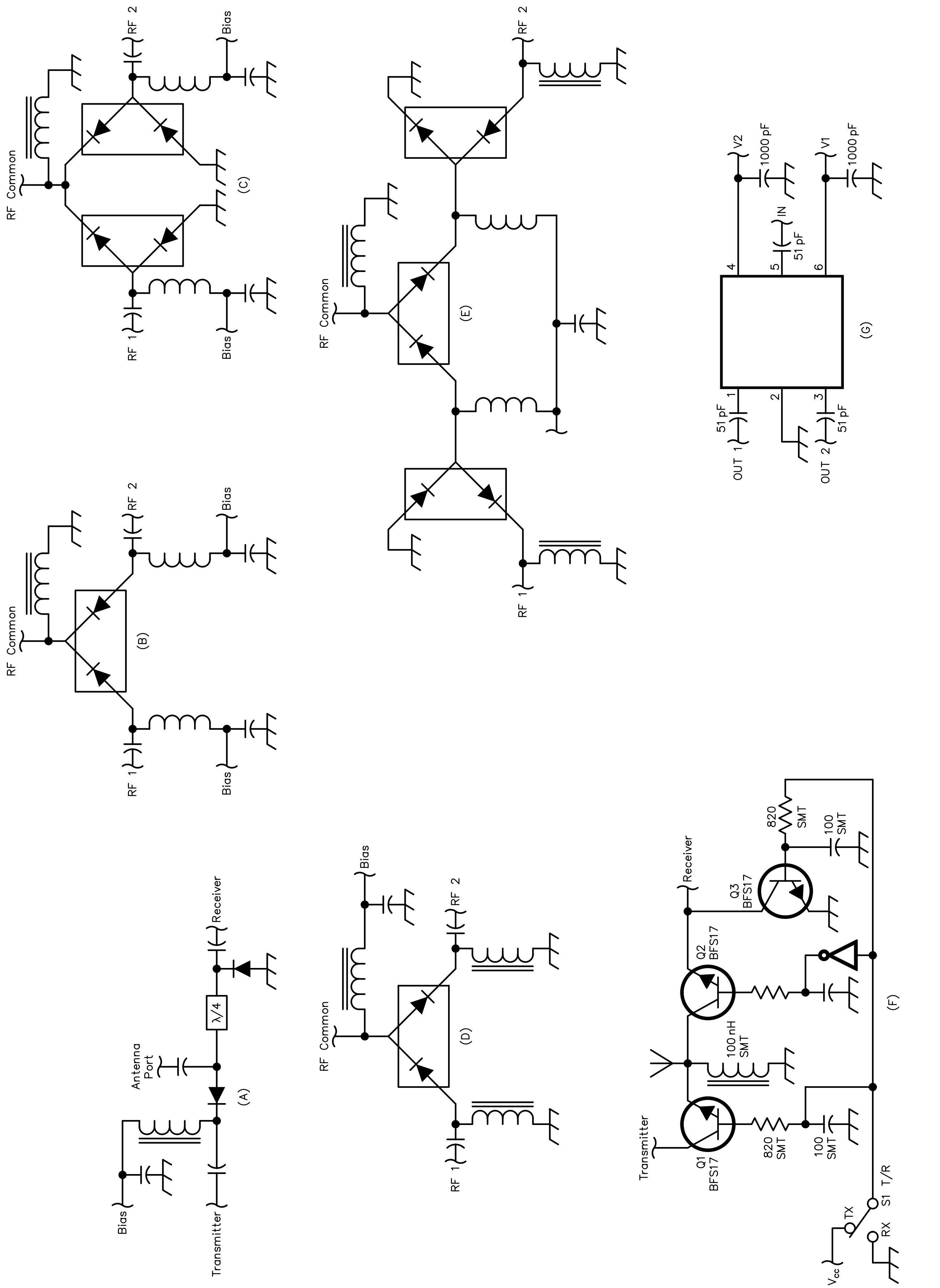


Fig 5—The two-way communicator base station. It looks messy because it is in a constant state of experimentation. The T/R circuit is on the upper board; the trace with the U-shaped bend is the λ/4 microstrip phasing line. The transmitter, receiver and multiplexer were first mounted to small pieces of PC board and these were connected to the main board. This method allows easy experimenting with SM parts.

Fig 6—Some UHF T/R switch schematics. A through E are based on Hewlett-Packard PIN diodes. (A) A low-current T/R switch. (B) Simple SPDT switch using only positive current. (C) High-isolation SPDT switch, dual bias. (D) Switch using both positive and negative bias currents. (E) Very high isolation SPDT switch, dual bias. (F) is an RF Monolithic design. (G) is from the California Eastern Laboratories UPG155 IC datasheet. V1 and V2 are control voltages (high = 3, low = 0 and V2 = –V1). PIN diodes present a pure resistance to RF that can be varied over a very wide range (1 to 10 kΩ) by applying a dc control current. When the control current is switched on and off, a PIN diode makes a good T/R switch. Its small size, high switching speed and freedom from parasitic effects make it a fine choice for broadband signal control.


RF Common
RF 1
RF 2
Bias
(A)
(B)
(C)
(D)
(E)
(F)
(G)
Antenna Port
Transmitter
Receiver
λ/4
Q1 BFS17
Q2 BFS17
Q3 BFS17
100 nH SMT
820 SMT
100 SMT
S1 T/R
TX
RX
V1
V2
IN
OUT 1
OUT 2
1000 pF
51 pF
1
2
3
4
5
6

filtered CW remodulated at a fixed frequency, because it sends more information to the remote station.

To understand why, think about what happens to the signal once it is retransmitted by the LC transmitter. Any HF signal that is greater than the threshold of U1 and within the passband of U2 will be transmitted. The transmitted signal will be on or off. There is no difference in volume with an on-off system; all signals appear to be equally strong. When you hear the CW on the remote speaker you will not be able to tell if the signal is just over the threshold or 40 dB over S9—that information has been lost. This means that any static that is just over the threshold set by R1 and R2 will be as loud as a strong CW signal.

By modulating this signal at the actual frequency of the input signal instead of a fixed frequency, you can get more information about the input signal. Because the CW filter has a 100-Hz bandwidth, it will pass signals from 700-800 Hz. This allows you to distinguish a 750-Hz signal from one at 800 Hz. With this information, you will know if the signal is centered or on the edge of the passband, which can help you tune the HF rig or determine if there is an interfering signal nearby.

The Remote Station

See Figs 7 and 8.[8] The receiver output (pin 5 of U3) is sent to three-pole, two-position multiplexer U4, part of the T/R circuit. This switch allows the microprocessor to change the speaker between the receiver and the keyer's sidetone and switch the antenna between the receiver and transmitter. In receive mode, pin 14 is connected to pin 13, which has the received audio. Pin 15 is connected to pin 1, which is at Vcc; and pin 4 is connected to pin 3, which is at ground.[9] The voltages on

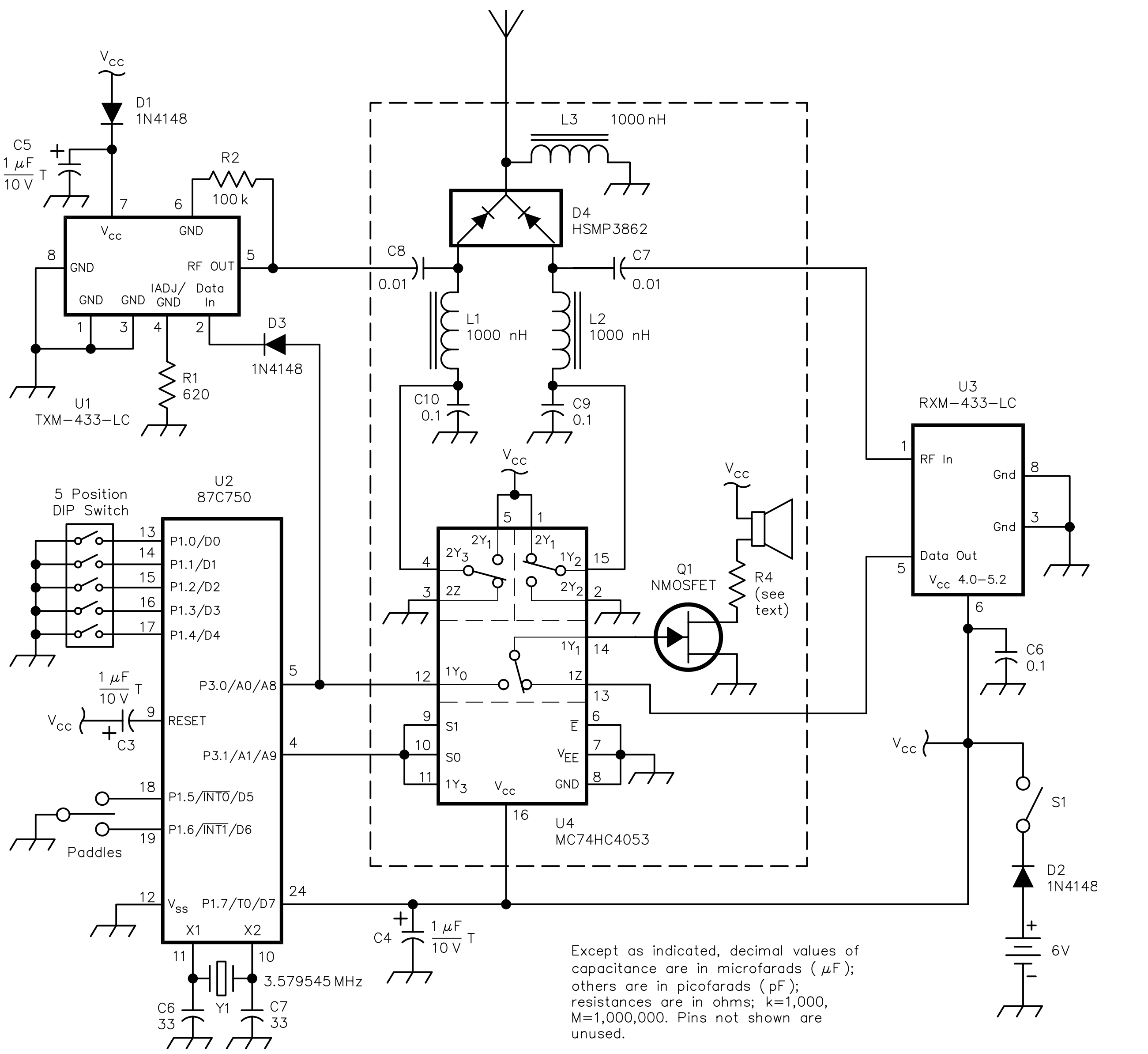


Fig 7—Schematic of the remote system. The T/R circuitry bounded by dotted lines.

pins 15 and 4 forward bias the diode between the antenna and the receiver input and reverse bias the diode between the antenna and the transmitter, thus connecting the receiver and isolating the transmitter. While transmitting, the sidetone is connected to the speaker and the diodes are biased in the other direction.

Because the receiver and audio amplifier Q1 increase the power consumption, a larger-capacity battery is needed than for the keyer in Part 1. Four AA or AAA batteries would be ideal but they are large. A 9-V battery and a 5-V regulator would also work. The use of small headphones instead of a speaker can substantially reduce power consumption; and the addition of a resistor, R4, can also help to reduce power. These circuits have worked well in my tests, but growing accustomed to a constant-amplitude audio signal takes some time.

Ideas for Making a 50-Ω Feed Line and T/R Switches

While a single receiver or transmitter circuit can have the antenna located right at the output pin, this is not possible when both modules are in use. For best use, the antenna needs to be connected to the module through a 50-Ω line. Small coax will work, but it has significant loss at these frequencies. A 50-Ω microstrip is more elegant when the feed line is on the PC board (see Fig 9).[10] I made mine with a Dremel tool, by grinding away enough copper to reduce the edge-capacitance effects.

T/R switches serve two functions. When the transmitter is off, the transmitter branch should look like an infinite impedance to the receiver; if is not, the receiver antenna circuit will not be 50 Ω. When the transmitter is active, it is important not to overdrive the receiver input. In addition, the receiver input needs to look like an open circuit to the transmitter for best match. Poor T/R design can cause a significant reduction in performance.

The easiest way to make a T/R switch might be to not make one at all but rather use two separate antennas. I did this and it worked. Of course, each antenna has an effect on the other; and if they are very close, the transmitted signal heard by the receiver could be very strong.

Unfortunately, I could find very little information on low-power UHF antenna switches in the amateur literature, and most of the ones I found were mechanical. While a mechanical switch would probably work, it seemed a bit "low-tech" for this project. I wanted a solid-state switch. It seems to me that the area of UHF, solid-state T/R switches is one in which a person with RF skills and measurement equipment could contribute a lot to Amateur Radio literature. Until then, here is what I have learned.

Doug DeMaw described some solid-state T/R switches in *Solid State Design for the Radio Amateur* and *The QRP Notebook*. The simplest one consists of two diodes (connected "back to back") at the receiver input to limit the input voltage. I tried this and it worked, but I had three concerns.

The data sheet for the Linx modules indicates an absolute-maximum allowable input to the receiver of 0 dBm. Since the transmitter output power could be 4 dBm at 5 V, this could damage the receiver. Coupled to a 50-Ω antenna, 0 dBm (1 mW) implies a peak of 0.3 V (0.22 V RMS). A simple diode circuit would not protect the receiver input since they allow about 0.6 V. Also, this circuit does not isolate the transmitter from the receiver when in the receive mode, causing impedance mismatches. Finally, ordinary small-signal diode capacitance can cause further signal deterioration, especially at these higher frequencies.

I found several ideas for UHF T/R circuits at the HP Web site,[11] one at the RF Monolithics Web site[12] and recently one at the California Eastern Labs Web page.[13] These are shown in Figs 6A-F. I built the designs in Figs 6A and 6B because they were the first ones I discovered, but I have left my circuits in prototype form so I can try some of the others. Both 6F and 6G[14] look much simpler to construct. I was rather disappointed when buil-ding 6A. It seemed simple but with the circuitry for the control voltage, it took up a lot of board space and needed many components.

General Observations on Working with UHF

I find it fun to play with UHF and I have tried a number of variations on the projects discussed here. Keep in mind that I have no equipment capable of

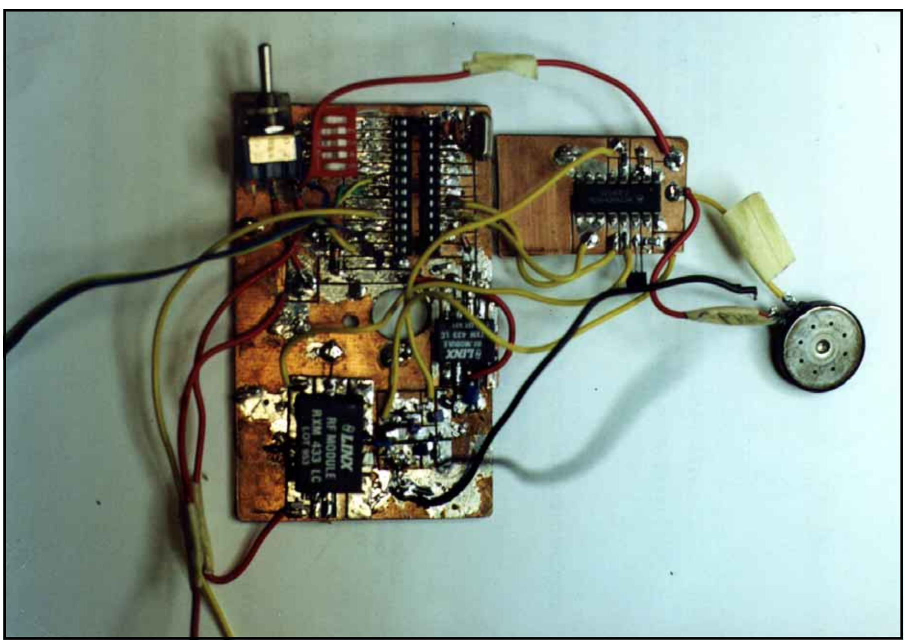


Fig 8—The remote station on the same board that holds the transmitter of part 1 (plus an add-on for the multiplexer). Making use of SM parts and using both sides of the PC board, this will fit easily on the board, but the present layout works and allows easy experimenting. Notice the Fig 6B T/R switch in the lower right. It is on a separate board and can easily be removed to test other T/R configurations.

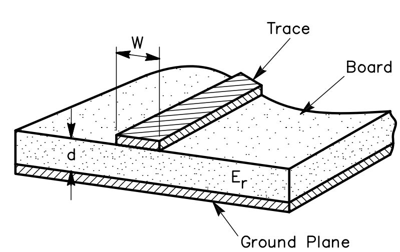


Fig 9—Microstripline construction. FR-4 PC board 0.062 inches thick has an ε in the 4-4.5 range. At 440 MHz, a 3-mm-wide trace has Z_0=50 Ω.

measuring anything at 400 MHz—except, of course, the signal-strength meter section of Part 2. I have a DVM and a 60-MHz ’scope that was useful when developing the lower-frequency portions of the circuits, but nothing to measure SWR, high-frequency impedances, etc. Consequently, my projects are truly amateur in the classic sense. I build them; if they work well enough to do the job, they are a success. If not, I try another design. The ones I have discussed have been successful. It is likely that you will find ways to make them work better. If you do so, please let me know or write an article for *QST* or *QEX*.

Common problems include leaving insufficient space around the antenna. I noted some proximity effects when the keyer was too near a computer. Digital circuits can degrade receiver performance. A microcontroller in a circuit can be a source of noise, although I have not noticed any in mine. If so, you might need to move or shield it.

I discovered that operation of the HF transmitter on certain frequencies interferes with the UHF circuits either via its antenna or via the leads to the rest of the circuit. As always with high-frequency circuits, use plenty of ceramic and tantalum bypass capacitors. Of course, surface mounted parts give best results. Route the ground traces of the digital circuits away from the antenna inputs. I recommend grounding the antenna through an inductor (about 100 nH) to prevent static discharges from damaging the input circuits.

Other Things You Might want to Try

A wireless intercom: I built a 434-MHz personal communicator using the LMC6482 as a pre-amplifier for an electret microphone as input to the transmitter of Part 2. I am presently working on a wireless intercom circuit using the LM4830, which has an audio amplifier, mic pre-amplifier and digital volume control, all on a single chip. I have thought of using project 2 as a VOX inter-motorcycle communicator, so both hands remain on the handlebars.

I wonder if it is possible to build an SSB version of the two-way CW circuit. Because RM modules do not shut off the carrier between data elements, an RM two-way system would have different design considerations. By periodically stopping transmission, you introduce annoying breaks in the input signal. Because the RM circuits need the power input disconnected to stop transmission, T/R switching will take longer. On the other hand, the receiver signal would be much easier to copy than the on/off signal of the two-way LC circuit. This looks like an interesting area to explore.

Remote HF radio control: This final part offers is a basis for some interesting alternative designs. Consider if the transmitter sent modulated CW and the receiver had demodulators at two different frequencies, say 800 and 2000 Hz. It should be easy—with some microprocessor coding—to key the HF rig with the 800-Hz signal and decode the 2000-Hz signals to remotely control the HF volume, frequency, etc.

A garage door opener: The intended purpose of the Linx modules is for digital control. In fact, the datasheets show a “typical” circuit using Holtek encoders and decoders.[15] These encoders and decoders send a series of short and long pulses (dots and dashes). The receiver decoder determines if the coded address is its address and, if so, it sets a switch. Morse code could serve as the amateur version of this. Send your call sign to open the garage door. You can bet no one else is going to be able to figure out what code you’re using and break in! If you do experiment with using these circuits for control, just remember your transmitter has to send your call sign at least every 10 minutes. Other design ideas include a wireless modem or a remote robotic control system.

The opportunities to experiment using the newly available UHF modules are enormous. We amateurs have some very valuable spectrum and fewer restrictions than commercial users. I hope you will join me in making use of these unique opportunities.

Notes

[1] S. Ulbing, “A Simple UHF Remote-Control System: Pt 1,” *QEX*, Jul/Aug 2000 pp 32-39.

[2] S. Ulbing, “A Simple UHF Remote-Control System: Pt 2,” *QEX*, Sep/Oct 2000 pp 33-40.

[3] Unless you are using weighted code.

[4] Actually the transmitter is already off for a space. It is just necessary to wait for its signal to quiet down.

[5] This circuit looks messy because I am using it as a test bed to try different T/R switches and other designs. Although messy, it works well.

[6] S. Ulbing, N4UAU, “The New and Improved Uncle Albert’s Keyer,” *The 1996 ARRL Handbook,* pp 22.21-22.24. The original Uncle Albert’s Keyer appeared in *QST* (Jan 1994, pp 42-44).

[7] HP Application Note 1067 discusses this T/R switch in detail.

[8] The empty socket is for the microcontroller, which has been removed for reprogramming.

[9] The output filter is not needed for the remote receiver as the ac coupled audio amplifier and our ears do the filtering.

[10] For details of microstripline construction see the discussion in Chapter 6 of *The ARRL UHF/Microwave Experimenter’s Manual* (Newington, Connecticut: ARRL, 1997). Order No 3126, $20. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Check out the full ARRL publications line at **http://www.arrl.org/catalog/**.

[11] Some of the application notes I found useful are ANs 922, 929, 951-1, 951-2, 1048, 1049, 1054, 1067, 1072.

[12] Their *HX/RX Designers Guide* has lots of useful information. Look at **http://www.rfm.com/hxrx/hxrx.htm**.

[13] The UPG155 datasheet is at **http://www.cel.com/pdf/datasheets/u153.pdf**.

[14] Since writing this article, I have tested a T/R switch using the circuit of Fig 6G. It works well and is easier to build than the other ones I describe—if you can work with a package that is only 2.0×1.25 mm with three pins on each side.

[15] Information on these ICs is available at the Holtek Web site: **www.holtek.com**. □□

Thermistors in Homebrew Projects

Thermistors are inexpensive, readily available components. Better yet, they can greatly enhance the performance of your projects. Learn how to put them to work for you.

By William E. Sabin, W0IYH

Thermistors are interesting compnents that Amateurs can use to enhance their projects. Variations in circuit temperature that affect gain, distortion and control functions such as receiver AGC or transmitter ALC can be compensated. Dangers of self-destruction of overheated power transistors can be greatly reduced. Oscillator drift can be greatly reduced. This article discusses thermistor properties and shows three examples of how they can improve project performance. We will see first that some easy mathematics improves the understanding.

Mathematics of Thermistors

A thermistor is a small bit of intrinsic (undoped) semiconductor material between two wire leads. As temperature increases, the number of liberated hole-electron pairs increases exponentially, causing the resistance to decrease exponentially. This exponential nature is seen in the resistance equation:

$$R(T) = R(T0) \bullet e^{-\beta\left(\frac{1}{T0}-\frac{1}{T}\right)} \qquad \text{(Eq 1)}$$

where T is some temperature in Kelvins and $T0$ is a reference temperature, usually 298 K (25°C), at which the manufacturer specifies $R(T0)$. The constant β is experimentally determined by measuring resistance at various temperatures and finding the value of β that best agrees with the measurements. A simple way to get an approximate value of β (this is usually all we need in ham-gear design) is to make two measurements, at room temperature, say T = 25 °C (298 K) and $T0 \approx$ 100°C (373 K) in boiling water. Suppose the resistances are 10 kΩ and 938 Ω. Eq 1 is solved for β:

$$\beta = \frac{\ln\left[\frac{R(T)}{R(T0)}\right]}{\frac{1}{T}-\frac{1}{T0}} = \frac{\ln\left(\frac{938}{10000}\right)}{\frac{1}{373}-\frac{1}{298}} \approx 3507 \qquad \text{(Eq 2)}$$

Usually, the exact value of temperature is not as important as the ability to maintain that temperature. A better estimate of β, if needed, can by achieved by a linear regression method using the program *THERMIST.BAS*, downloadable from the *ARRL QEX* Web site.[1] This program takes the logarithm of both sides of Eq 1, which provides a linear relationship between log($R(T)$)

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Cedar Rapids, IA 52403
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[1]You can download this package from the ARRL Web http://www.arrl.org/qexfiles/. Look for THERMIST.ZIP.

and (1/*T0* – 1/*T*). Five equally spaced data points are input to get the slope of the line, which is β.

The following examples illustrate a few of the main ideas of thermistor-circuit design that can be employed in a number of similar situations.

MOSFET Power-Transistor Protector

It is common practice to compensate the temperature sensitivity of power transistors. In bipolar (BJT) transistors, thermal runaway occurs because the dc current gain increases as the transistor gets hotter. The runaway condition is less likely to occur in MOSFET transistors, but with excessive drain dissipation or inadequate cooling the junction temperature will increase until its maximum allowable value is exceeded. A common procedure is to mount a diode or thermistor on the heat sink close to the transistors, so that the bias adjustment tracks the flange temperature. References 1 and 2 give detailed discussions of this.

A special problem occurs when a sudden large increase in transistor dissipation occurs. The flange temperature changes rather slowly because of the thermal capacitance (heat storage) of the heat sink. However, the junction temperature rises much more rapidly and can rise above the maximum limit before the correction circuit has a chance to function. If a thermistor is attached to the ceramic case with a small drop of epoxy as shown in Fig 1 it will respond much more rapidly (shorter time constant) and *may* (not guaranteed) save the transistor. The circuit (Reference 3) of Fig 2 detects a case temperature of about 93°C and completely turns off the FETs until the case temperature drops about 0.3°C, at which time the FETs are turned on again. A red LED on the front panel warns of an over-temperature condition that requires attention.

In Fig 2, a 4.7 V Zener (D1), *R1* (metal film) and *R*th (thermistor) are a voltage divider with an output of 0.6 V.

If this voltage decreases slightly (because the resistance of *R*th decreases slightly) Q1 starts to come out of saturation, Q2 quickly goes into saturation and the gate voltage of the FET goes to a low value, turning it off. At the same time, the 20-mA LED (RS 276-307) lights. The voltage divider equation is:

$$0.6 = 4.7 \bullet \left(\frac{R\text{th}}{R1 + R\text{th}} \right) \qquad \text{(Eq 3)}$$

If we solve this for *R1*, we get:

$$R1 = 6.831 \bullet R\text{th} \qquad \text{(Eq 4)}$$

If the value of *R*th is known at some temperature, then *R1* is the value that activates the circuit at that temperature. An interesting feature is that the voltage across the thermistor never exceeds about 0.6 V, and this greatly reduces the self-heating of the thermistor, which could otherwise cause a substantial error in the circuit behavior. The temperature variations of Q1 and D1 are small sources of error, so this simple circuit should be placed in a cool location, not directly on the heat sink.

The RadioShack precision thermistor (RS 271-110) is rated at 10 kΩ ±1% at 25 °C. It comes with a calibration chart from –50°C to +110°C that can be used to get an approximate resistance at some temperature. We are most interested in FET case temperatures in the range of 70°C to 100°C. It is assumed that very close temperature knowledge is not needed, but in order to be sure that the thermistor is okay, I measured its resistance at 20°C (68°F) and in boiling water (≈100°C). The circuit of

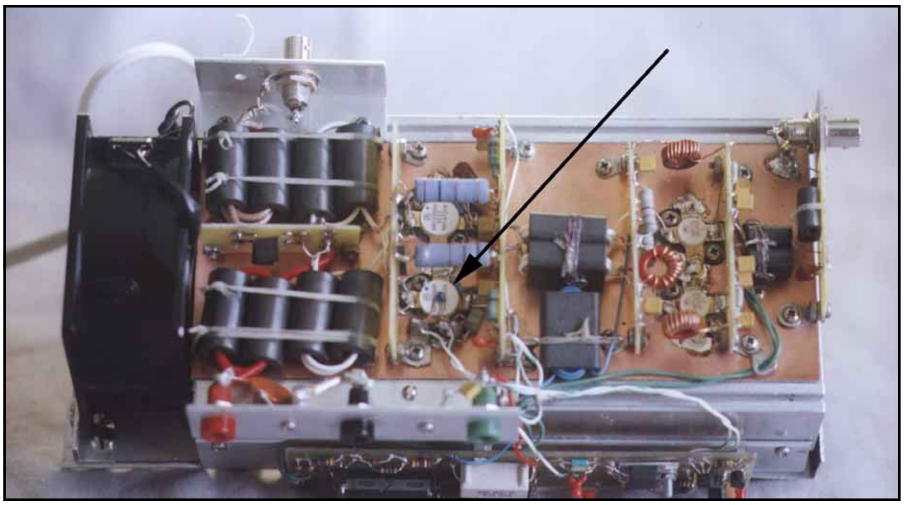

Fig 1—The thermistor attached to power MOSFET with a drop of epoxy.

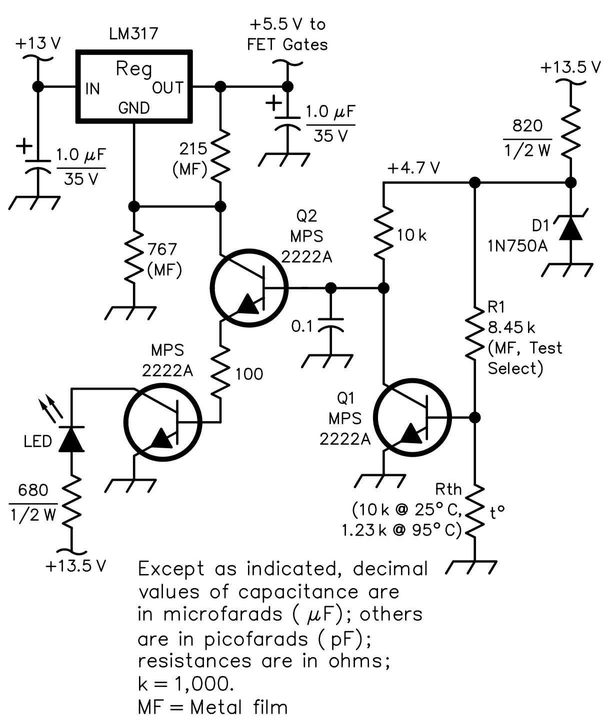

Fig 2—Schematic of a MOSFET temperature-protection circuit.

Fig 2 is intended for 50°C or greater when the 10-kΩ thermistor is used. At lower temperatures, the circuit of Fig 3 would be better.

The following procedure was used to get the desired temperature control:

1. The MRF150 MOSFET has a maximum allowed junction temperature of 200°C. The thermal resistance θ_{JC} from junction to case is 0.6°C/W.
2. The maximum expected dissipation of the FET in normal operation is 110 W.
3. I selected a case temperature of 93°C. This makes the junction temperature 93 + (0.6) (110) = 159°C, which is a safe 41°C below max.

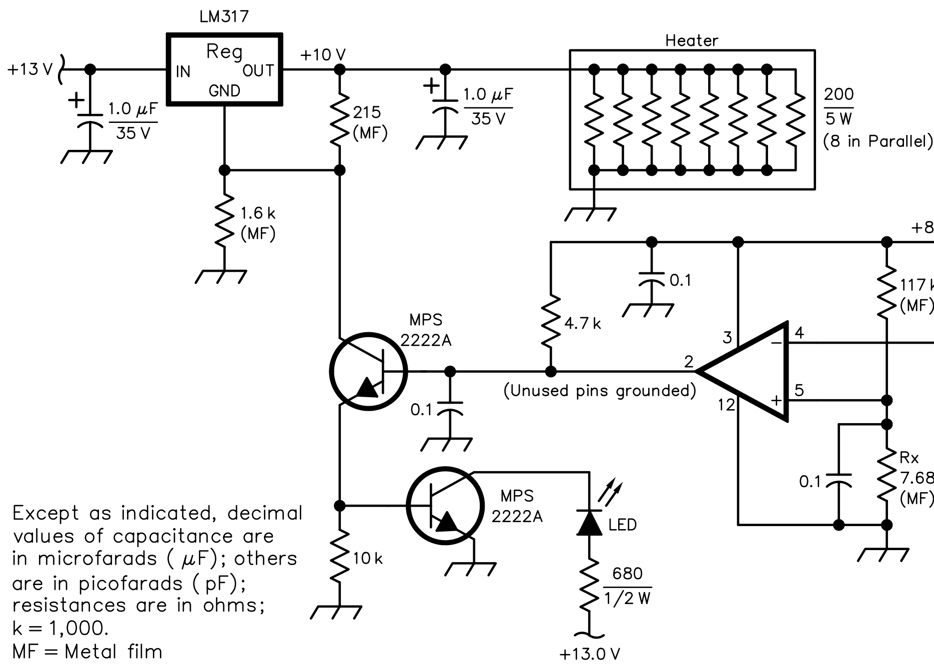

Fig 3—A VFO temperature-controller schematic. The eight 200-Ω resistors labeled "Heater" serve to warm the VFO enclosure

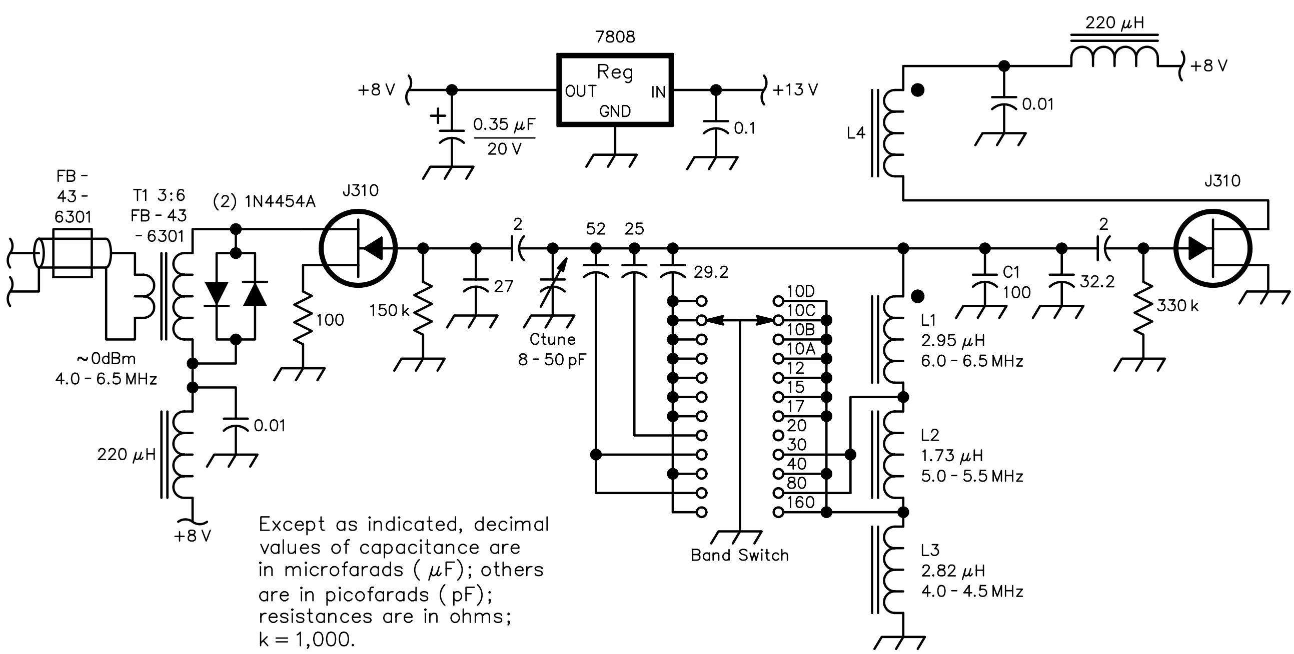

Fig 4—Schematic of a three-band VFO.
L1—22 t #24 AWG, wide spaced on a T68-7 core
L2—17 t #24 AWG, wide spaced on a T68-7 core
L3—22 t #24 AWG, wide spaced on a T68-7 core
L4—2 t #24 AWG link, close wound on L1 core
SW1—Electroswitch D4C0312N
Ctune—Hammarlund RMC-50-S

4. The FET has a rating of 300 W maximum dissipation at a case temperature of 25°C, derated at 1.71 W/°C. At 93°C case temperature, the maximum allowed dissipation is 300 – 1.71 (93 – 25) = 184 W. The safety margin at that temperature is 184 – 110 = 74 W.
5. A very simple way to determine the correct value of *R1* is to put the thermistor in 93°C water (let it stabilize) and adjust *R1* so that the circuit toggles. I found that 8450 Ω was the nearest standard value for a metal-film resistor. At 93°C, the measured value of the thermistor was about 1230 Ω.

Measurements of the circuit sensitivity determine the temperature values at which the circuit toggles on and off. I replaced the thermistor with a resistor decade box and measured resistances of 1224 Ω and 1234 Ω. Solve Eq 1 for the *T* that corresponds to each value of *R*:

$$T = \frac{\beta}{\ln\left(\frac{R}{R0}\right) + \frac{\beta}{T0}} \qquad \text{(Eq 5)}$$

which is easy to perform with a handheld calculator or math program such as *MathCAD*. Using the two values of *R*, I found a temperature range of about 0.3°C.

A Temperature Controlled VFO

The circuit of Fig 3 is used to control the temperature of the three-band VFO shown in Fig 4, inside a thermally insulated enclosure. The Wheatstone bridge circuit with an LM339 comparator as a null detector is more sensitive and less temperature dependent than Fig 2. The ’339 works quite well at a level of 0.5 V at each of its two inputs. This circuit is preferable at lower temperatures, such as 30 to 35°C, where the thermistor resistance is in the 8 k to 7 kΩ range.

I use eight 200 Ω, 5 W metal-oxide resistors at the output of the LM317. The total heat applied is 4 W to maintain 33°C, and the resistors are placed so that their heat is distributed uniformly: Half are placed near the bottom and half near the top. The thermistor is mounted in the center of the box (see Fig 5), close to the tuned circuit and in physical contact with the oscillator ground-plane surface, using a small drop of epoxy.

The enclosure is homemade from sheet aluminum and angle stock. It is large enough that it has almost no effect on frequency. The outside is lined with 1/4-inch plexiglass and the inside surfaces are lined with 1/4-inch Styrofoam sheets. The tuning and bandswitch shafts are thermally insulated from the outside world, using plastic shaft couplers. Plexiglass blocks attach the box to the front panel. Electrical grounding is via an RF choke (for dc) and several 0.01 µF capacitors (for RF).

The temperature at the thermistor location is maintained within 0.1°C, as determined by thermistor resistance measurements. Using the method of the previous example to get the temperature range and knowing the frequency drift versus temperature coefficient of the VFO in parts per million (PPM), the frequency change can be found as follows:

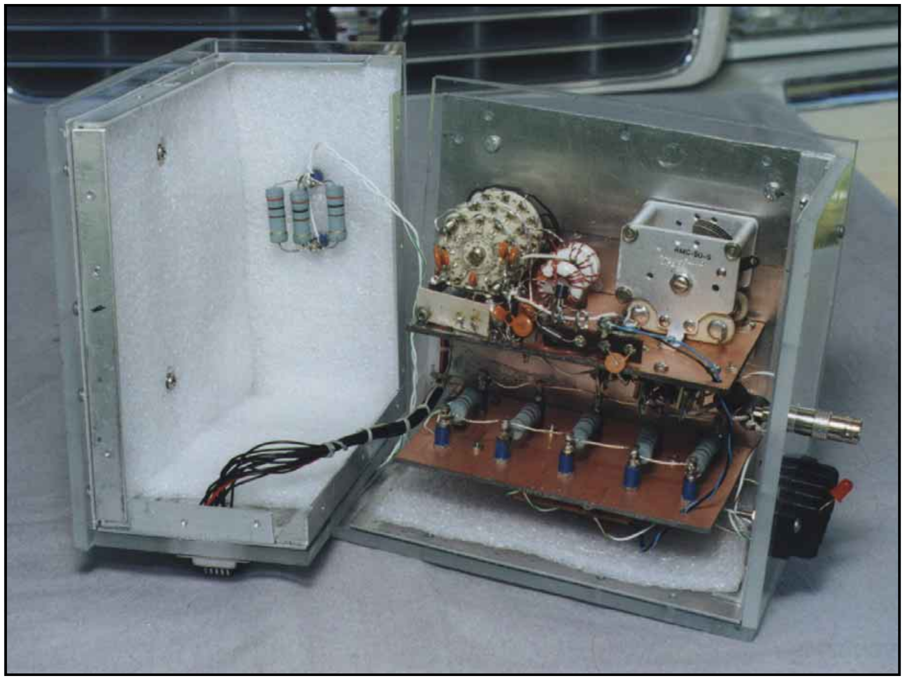

Fig 5—The three-band VFO with temperature control. Five of the eight 5-W resistors are mounted on a circuit board to the right of and slightly below center. The other three 5-W resistors are wired together slightly to the left and above center. The thermistor is not visible in this photo.

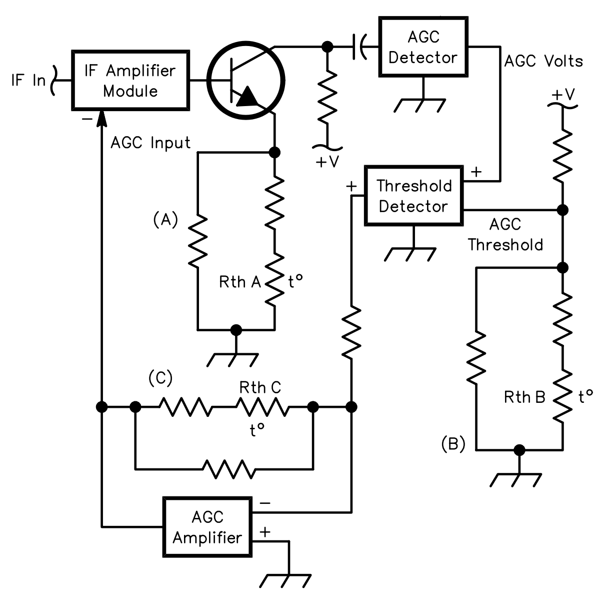

Fig 6—Temperature compensation of IF-amplifier and AGC circuitry: (A) IF amplifier, (B) AGC threshold, (C) AGC gain. Each compensation circuit uses two resistors and a thermistor (RthA, RthB or RthC).

$$\Delta F = F_{VFO} \bullet PPM \bullet \Delta T \qquad \text{(Eq 6)}$$

Plugging in some numbers encountered for a 4.0 to 6.5 MHz VFO with PPM = –500/°C and Delta T = 0.1, before temperature compensation,

$$\Delta F = 6{,}000{,}000 \bullet \frac{-500}{1{,}000{,}000} \bullet 0.1 = -300 \text{ Hz} \qquad \text{(Eq 7)}$$

which indicates that some improvement is needed, in particular the temperature coefficient. If the enclosure is massive or well insulated, the rate of temperature change can be slowed down. When a small temperature range of 0.1°C was maintained, the VFO temperature coefficient was improved by a factor of about seven with a negative-temperature-coefficient capacitor, C1. Inexpensive polystyrene capacitors, for example Mallory type SX, have a well-controlled negative (–120 PPM/°C) temperature coefficient that is intended to offset the positive temperature coefficient of inductors. (Caution: To prevent damage, use pliers on the leads as a heat sink when soldering.) No other temperature compensation was needed (probably fortuitously).

Temperature-compensating capacitors are available from Surplus Sales of Nebraska (**www.surplussales.com**). Other suitable fixed ceramic capacitors are Vishay 561 series, type 10TC NP0, which I have found are excellent. No variable trimmer caps are needed because each VFO band has a 25 kHz margin at each end and a calibrated analog dial is not used. The two-turn feedback winding in the J310 drain satisfies all three coils. The three inductors use Carbonyl-TH, T68-7 cores (white paint), which are claimed by Amidon to be the most temperature-stable mixture at normal room temperatures.

The oscillator finally turned out to be very slightly overcompensated. Over the 0.1°C range, the frequency varies ±20 Hz or less, with a period of about five minutes. Superimposed is a very slow drift of average frequency that is due to settling of component values, including possibly that of the RadioShack thermistor. These gradual changes became negligible after a few days of continuous operation.

One problem that is virtually eliminated by maintaining a constant temperature is the "retrace" effect on cores and capacitors. Because of "retrace," components subject to a substantial temperature transient of some kind may take several hours to recover their previous L and C values. The thermistor may also show a retrace effect.

Reference 4 shows ways to perform the temperature compensation operation and gives further references. An especially good method is to toggle the value of *Rx* (Fig 3) slightly so that a variation of ±0.5°C is created inside the VFO enclosure, and then do the temperature compensation. Because of the small average power dissipation in the VFO plus controller, it is economical to let the VFO run continuously so that initial warm-up drift (measured less than 1 kHz) and retrace are avoided.

If the VFO is mixed with crystal frequencies and then bandpass filtered (the "mix-master" approach) the final local oscillator (LO) can be quite stable and very clean spectrally. This is especially so if the crystals (±20 PPM/°C *maximum*) are temperature compensated, temperature controlled or even phase-locked to a reference (see Reference 5). An LO frequency counter (see Reference 6) offset by the IF is a very simple and excellent way to read the actual RF signal frequency to within ±50 Hz, if the reference crystal is of high quality and perodically adjusted to WWV. The frequency stability is quite adequate for HF SSB/CW, which are the primary applications for this equipment. Listening tests confirm that in SSB speech, slow frequency changes of ±50 Hz are hardly noticed. The reason for the three bands of the VFO (4.0-4.5, 5.0-5.5, 6.0-6.5 MHz) is to minimize spurious mixer products due to harmonic intermodulation that can slip through the LO bandpass filters.

Gain and AGC control

Thermistors are used to stabilize, or to vary in a controlled manner, the gain of an amplifier or an on-off threshold. Fig 6 sketches three examples. Fig 6A is the final IF amplifier stage that is part of an IF amplifier module. The thermistor compensates the gain variation of the module. Figs 6B and 6C are part of the receiver AGC circuit. The thermistors compensate for variations in AGC slope (dB/V) and vary the AGC threshold.

In all cases, use a resistor decade box to determine the resistance-temperature correlation and plot an RT curve. Then look for a resistor-thermistor circuit that seems reasonable. A *MathCAD* or *Excel* worksheet is then an elegant way to get the component values experimentally by comparing the thermistor-circuit temperature curve with the desired RT curve. Usually, a close approximation is good enough, and perfection is not justified. (Catalogs offer a wide assortment of thermistors for these projects.) The method suggested in Fig 6, one thermistor and two resistors, is a simple combination that gives a good approximation to the desired RT curve. In many cases, one of the resistors can be deleted. Reference 7 shows other useful and interesting thermistor applications.

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1. N. Dye and H. Granberg, *Radio Frequency Transistors*, Chapter 4, (Boston: Butterworth-Heinemann, 1993).
2. H. Granberg, "Wideband RF Power Amplifier," *RF Design*, Feb 1988. Also *Motorola RF Applications Report* AR313 (1993), p 424.
3. W. Sabin, W0IYH, "A 100-W MOSFET HF Amplifier," *QEX* Nov/Dec 1999. See also "Letters to the Editor," *QEX*, Mar/Apr 2000, pp 60-61.
4. *ARRL Handbook* (Newington: ARRL) 1995 to 2000 editions, Chapter 14. Order No. 1832, $32. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Check out the full ARRL publications line at **http://www.arrl.org/catalog/**.
5. See Chapter 17 of the *ARRL Handbook* for a description of the Ten-Tec Omni VI Plus transceiver.
6. Radio Adventures Co, RR4 Box 240, Summit Dr, Franklin, PA 16323; tel 814-437-5355, fax 814-437-5432; **information@radioadv.com**; **http://www.radioadv.com**. Model BK-172. For a description, see the 1999/2000 *ARRL Handbook,* Chapter 26.
7. P. Horowitz and W. Hill, *Art of Electronics*, Second Edition (New York: Cambridge University Press, 1989).

Introduction to Adaptive Beamforming

What is the difference between an antenna and a filter? Maybe not so much as most of us think. Let's consider antennas that might reject interfering signals.

By Doug Smith, KF6DX

Several books and articles have discussed using adaptive filters as a way to build systems that either eliminate or enhance some time- or frequency-domain property of an input signal.[1] That is the basis for most embedded noise-reduction systems today. *Adaptive beamforming* extends the same techniques to the spatial domain. When it comes to antenna arrays, the goal is a certain radiation pattern that maximizes the received signal and eliminates unwanted signals. Arrays achieve their directivity by phasing of various elements. Placing element phase under DSP control opens some very interesting possibilities indeed.

In adaptive filtering, coefficients are adjusted on the fly according to the degree of correlation of the input data to past copies of itself. This correlation forces convergence to a set of filter coefficients that maximize (or minimize) output energy. We now extend the idea of auto-correlation to the spatial distribution of radio signals.

Adaptive antenna arrays may be devised that automatically cohere; that is, they may be made to steer their patterns to produce signal cancellation (nulls) in the direction of undesired signals, even without knowledge beforehand of their direction of arrival. Algorithms have been invented that separate strong signals from weak signals when their directions of arrival are different. Other algorithms may distinguish among signals based on their distance from the array; a system may be constructed that is sensitive to nearby signals and insensitive to distant signals, or *vice versa*.

Still other signal characteristics may be used to steer an array. Transient signals may be rejected in favor of steady or stationary signals. The possibilities are virtually unlimited. Such systems are likely to be significant in the design of *software-defined* radios. To begin, we will restrict our discussion to arrays that null directional interference.

Take the simple example of a two-antenna array as in Fig 1. Two identical, omnidirectional antennas (*omnis*) feed a summation network; one of the antennas is routed through an adaptive FIR filter, though, that

[1]Notes appear on page 55.

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conditions its signal. The unconditioned signal is called the *primary* signal and the other, the *reference* signal. Imagine that two radio signals are received: a desired signal and some interference. The desired signal and the interference are coming from two different directions. Both antennas receive both signals, but being at separate locations, their outputs are not precisely identical but are time-related functions of one another. Imagine a single signal arriving from azimuth angle θ. When the signal at the primary antenna is:

$$S_{PRI} = A\cos\omega_0 t \qquad \text{(Eq 1)}$$

the signal arrives at the reference antenna earlier by an amount of time:

$$\delta = \frac{l\sin\theta}{c} \qquad \text{(Eq 2)}$$

where l is the antenna separation and c is the propagation speed. The signal at the reference antenna is therefore:

$$S_{\mathrm{REF}} = A\cos(\omega_0 t + \omega_0\delta) \qquad \text{(Eq 3)}$$

The radiation pattern of the array may now be steered by altering the filter coefficients. Time of arrival is translated into direction of arrival by virtue of the array's architecture.

Upon convergence, the conditioned reference signal contains an interference component that is very nearly a match to the interference component of the primary signal and the interference is nulled in the summation. The system's output contains little interference, therefore, but still contains the desired signal, just as described in the case of adaptive noise reduction. This

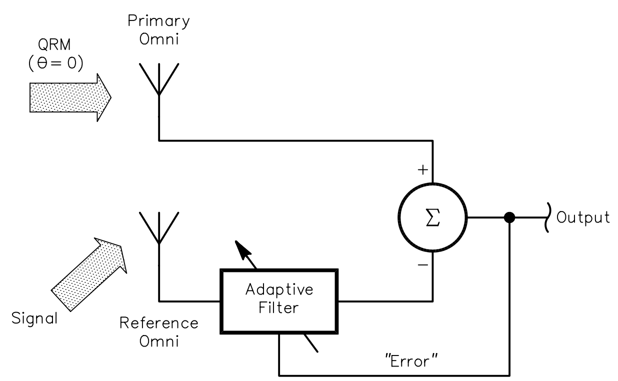

Fig 1—A block diagram of a simple, two-antenna adaptive array.

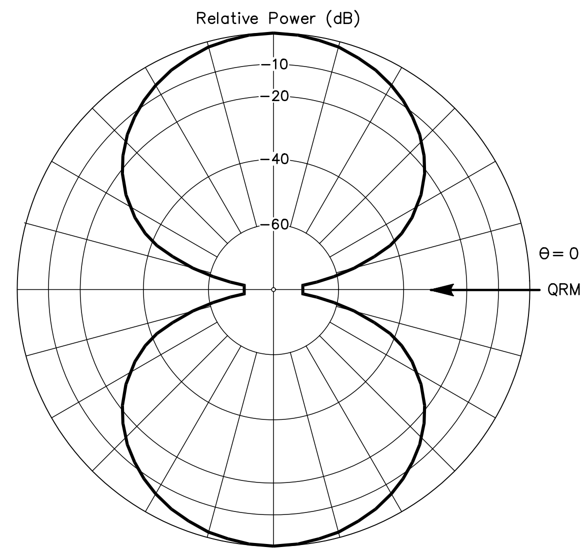

Fig 2—A typical radiation pattern of a simple, two-antenna adaptive array. (See Note 3.)

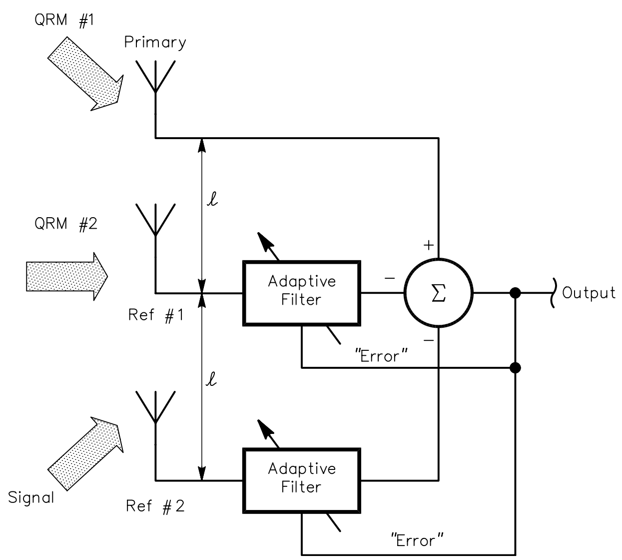

Fig 3—A block diagram of an adaptive array with two reference antennas.

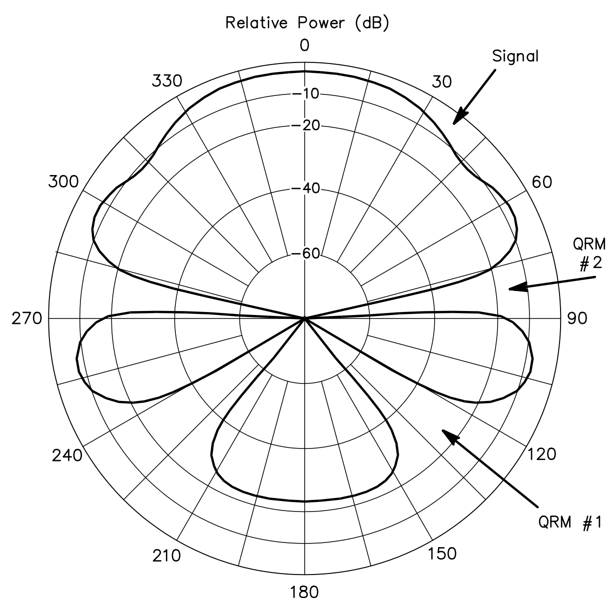

Fig 4—A radiation pattern of a two-reference adaptive array. (See Note 3.)

arrangement was originally studied by Howells in the late 1950s and later developed by Howells and Applebaum.[2]

In actual practice, primary and reference antennas usually feed separate receivers for amplification, selectivity and detection. The receivers add noise that may have significant impact on performance. Note that for this system to work properly, the interference must be strong compared to the desired signal, since the filter coefficients are

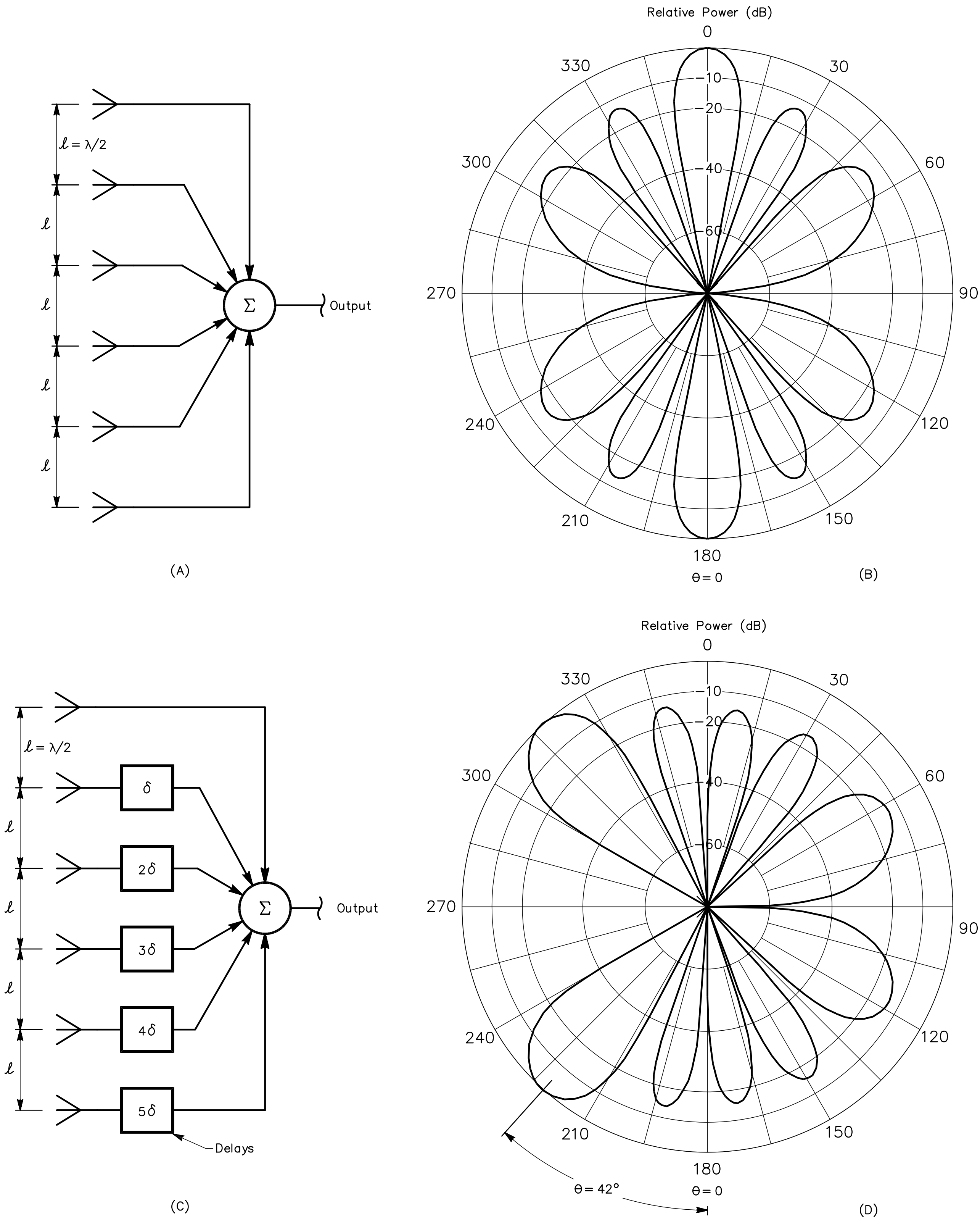

Fig 5—(A) A six-antenna linear array. (B) Its radiation pattern *without* delays inserted in the signal paths. (See Note 3.) (C) Steering of the array by placing fixed delays in the signal paths. (D) The radiation pattern of a manually steered array.

determined almost exclusively by the interference. The desired signal will not be nulled if it is strong compared to the receiver noise. This set of conditions is commonly found in the field.

We are interested in the radiation pattern of the converged "side-lobe *canceler*" described above. It is obviously bidirectional, since signals arriving from direction $\pi-\theta$ produce the same situation. Widrow and Stearns (see Reference 2) have shown that the nulls it forms have depth proportional to the strength of the undesired signal. Just when the interference gets stronger, the null improves! The pattern is always similar to a figure eight as shown in Fig 2.[3]

Under certain conditions, the desired signal may be so much stronger than the receiver noise that it tends to cancel itself. It is tempting to add artificial noise to the input to prevent this, but a better way to deal with this is the leaky LMS algorithm, discussed in Reference 1. It is restated here in vector form:

$$H_{k+1} = \gamma H_k + 2\mu e_k x_k \tag{Eq 4}$$

where H is the coefficients vector (the set of coefficients), X is the input data vector, and γ is chosen to be a positive constant less than unity. As mentioned before, values of γ greater than one may be tried to force the equivalent noise power downward, but the leaky LMS algorithm is only conditionally stable in this case.

Perhaps a more practical situation occurs in the presence of multiple interference sources. To cope with it, more than one reference omni must be used. Multiple nulls may be formed in this way. See Fig 3. Both conditioned reference signals are subtracted in the final summation. In this case, each antenna is distance l from its neighbor, although spacings may be varied to achieve different goals. This system may produce two sets of nulls, as shown in Fig 4. Note that each FIR adaptive filter may have as few as two coefficients.[4]

If more than two *interferers* are expected, more reference omnis may be added. When the number of interferers exceeds the number of reference omnis, the system converges to the solution that minimizes output energy. That is a complex function of all input variables.

Manual Steering with a Pilot Signal

It is not hard to see that the pattern of an adaptive array may be manually steered by artificially placing a pilot signal in the direction of the desired null or lobe. In one type of adaptive *beamformer*, this manual control may be retained along with the ability to adaptively null interference. The Howells-Applebaum side-lobe *canceler* is not preset to any particular "look" direction in the absence of signals; the pilot-signal adaptive beamformer relaxes to a predetermined directivity pattern in the absence of signals other than the pilot.

As array complexity grows, adaptive elements may shrink to have single weights—simple delay lines—to control radiation pattern. Fig 5A shows a six-omni array with $l = \lambda/2$ and Fig 5B its radiation pattern. When fixed delays are inserted in the signal paths, the array may be steered to some extent, as shown in the example of Fig 5C. The main lobe is now centered on angle:

$$\theta = \sin^{-1}\left(\frac{\lambda_0 \omega_0 \delta}{2\pi l t}\right) = \sin^{-1}\left(\frac{c\delta}{l}\right) \tag{Eq 5}$$

Sensitivity is maximized at this angle because the incident wave produces conditioned signals that are in phase with one another to add in the summation. As an example, when δ=250 ns and l=150 m, θ=30°. Notice that this pattern is independent of frequency so long as the omnis themselves are broadband, isotropic dipoles. The system may be considered a pilot-steering system, as the weights (delays) are manually set.

Spatial Architecture

To continue this line of thought, we must examine how antennas should be arranged to achieve some particular advantage in an adaptive beamformer. In practice, it may be found that placement does not matter much, since an adaptive system will take advantage of whatever spatial diversity it is given. Instinct would tell us that bigger is better. Still, an understanding of practical ways to control element phase is obviously needed and we have yet to examine narrow-band solutions to the adaptive beamformer problem.

One solution presents itself in analytic-signal form, much as seen in the discussion of modulation in Reference 3. Let us assume we are working with narrow-band signals and that each antenna's signal is conditioned by a Hilbert transformer with weighting,

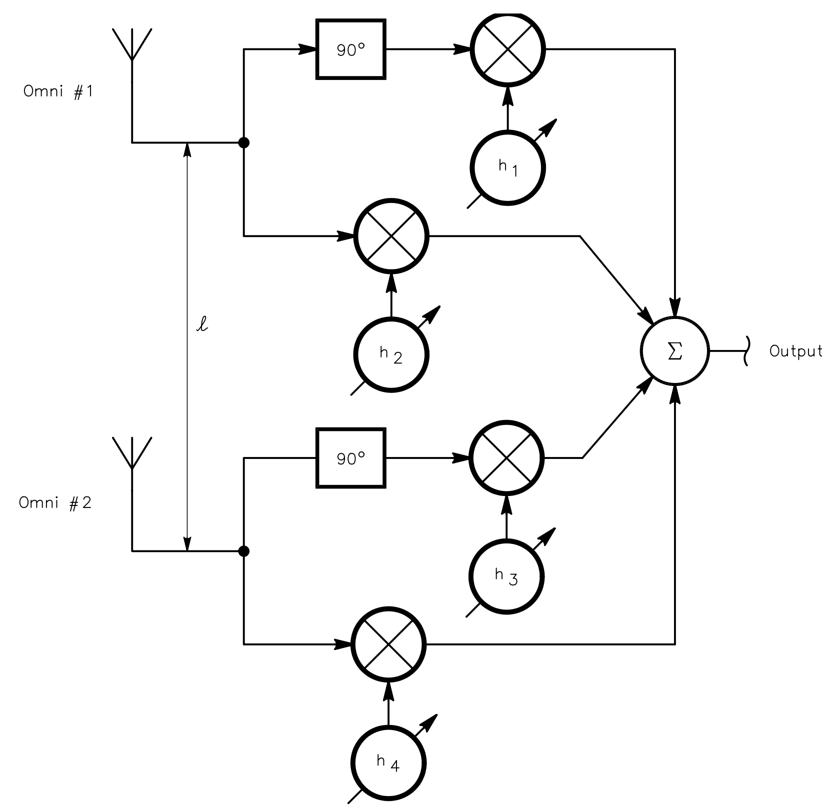

Fig 6—An antenna array with Hilbert transformers as the adaptive elements.

as shown in Fig 6. Each signal is weighted by a complex factor:

$$H = A(\cos\phi + j\sin\phi) = Ae^{j\phi} \quad \text{(Eq 6)}$$

where

$$\phi = \tan^{-1}\left(\frac{h_2}{h_1}\right) \quad \text{(Eq 7)}$$

Now, any phase angle ϕ may be generated by adjusting the two coefficients h_1 and h_2. The absolute magnitude of throughput gain is just:

$$A = \sqrt{h_1^2 + h_2^2} \quad \text{(Eq 8)}$$

We made the input signal narrowband so that we could minimize the number of weights. Broadband signals could be handled by using a Hilbert transformer with many weights in each leg (a long, analytic FIR filter pair). It might even be possible to design analytic filters that control the phase of different signals separately within the passband.

The efficacy of the pilot-signal system extends only to the similarity of the pilot signal and the desired signal. That is, the pilot signal injected at the receiver site is designed to have characteristics that resemble those of the desired signal. Further, the presence of a pilot signal in the array output may render the output unusable. This drawback has led to the development of algorithms that switch the pilot signal on only when adaptation is performed, then the coefficients are "frozen" and the pilot signal switched off for normal operation.

The Griffiths Beamformer

L. J. Griffiths' algorithm is a take-off on the LMS algorithm.[5] It may be used with advantage where knowledge exists beforehand about the correlation between the desired response, d_k, and the filter coefficients H_k. To do it, we have to reformulate the LMS algorithm a bit:

$$\begin{aligned} H_{k+1} &= H_k + 2\mu e_k X_k \\ &= H_k + 2\mu(d_k - y_k)X_k \quad \text{(Eq 9)} \\ &= H_k + 2\mu d_k X_k - 2\mu y_k X_k \end{aligned}$$

Now substitute the average value $E[d_k X_k] = S$ for its instantaneous value above and the result is Griffith's:

$$H_{k+1} = H_k + 2\mu(S - y_k X_k) \quad \text{(Eq 10)}$$

Now S is fixed and the thing operates without a desired response input, d_k. It converges on the least-mean-squares solution like the LMS algorithm, but to put it to work, one must have knowledge of the desired signal's direction of arrival, its autocorrelation function, and the array geometry. As against that, the pilot-signal algorithm does not need to know the look direction or array geometry since the pilot signal could be transmitted remotely.

Frost's Adaptive Beamformer

Both the above-described systems use "fuzzy reasoning" to place restraints on performance so they don't go wild. They are useful in imposing predetermined conditions on that performance, so weak desired signals can do little to alter the patterns they produce. In the design of O. L. Frost,[6] a "hard" constraint is placed on the look direction. With this restraint, the sensitivity in the look direction is fixed without regard to the strength of desired signals from this direction.

A block diagram of a Frost beamformer is shown in Fig 7A. Fixed steering delays are again used ahead of tapped delay lines. Now this is

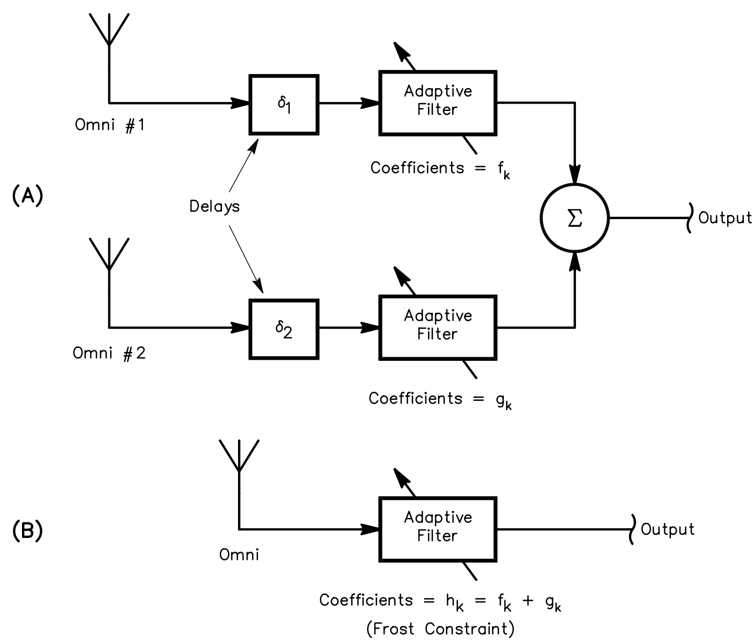

Fig 7—(A) A Frost beamformer. (B) Equivalent circuit of a Frost beamformer.

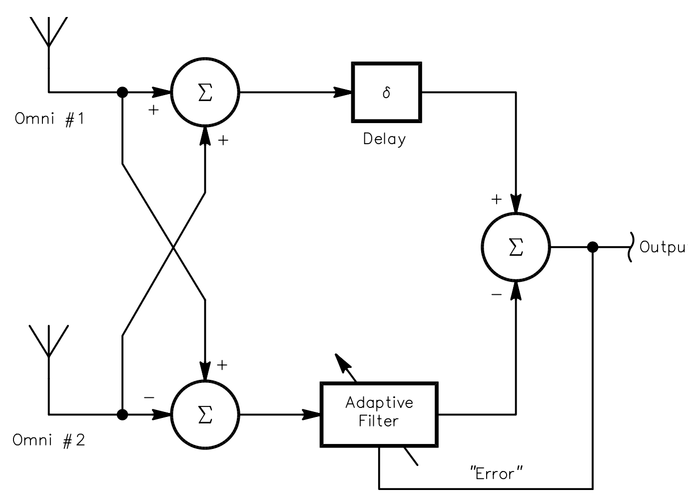

Fig 8—An adaptive array with super-resolution.

shown equivalent to the system of Fig 7B, where the summation is carried out as if the entire antenna-processing array were a single processor. Each of the weights in this single processor is simply the sum of the corresponding weights of each individual processor at Fig 7A. After the fixed weights of this single processor are found for a specified look direction, the weights of the adaptive processors may be varied as long as the "Frost constraint" is maintained. This way, the system obeys a fixed transfer function and filters the desired signals arriving from the look direction. It also minimizes output power by adapting itself to eliminate interference from other directions.

In the absence of the Frost constraint, and were the output power minimized, all the weights would go to zero and the output would disappear. The Frost constraint forces linear combinations of weights to be equal to certain constants. This causes the number of degrees of freedom to be less than the number of adaptive weights by the number of constraints placed. "Degrees of freedom," in this case, means the number of adaptive nulls that may be maintained.

Distortion in Adaptive Beamformers and "Super-Resolution"

We may define distortions of both the desired signal and of the radiation pattern. As for the signal, distortion is produced by rapid variations in the adaptive weights and this is to be minimized in the steady state. In many cases, noise in the weights can have an effect on the output. While this seems contrary to information in the References about DSP filters, we are now discussing non-Wiener behavior. Partial interference cancellation is a normal result and ideal characteristics sought are seldom achieved exactly.

Steering constraints have a large effect on output. If one of the conditions is that the desired signal have narrow bandwidth, then the worst interference that might be generated would also have narrow bandwidth. On the other hand, if the desired signal is broad and the interference sinusoidal, adaptive algorithms will try to modulate the interference so that it cancels some of the desired signal at the interference frequency. These are effects with which every user of LMS-algorithm noise reduction is familiar.

Steady-state radiation patterns seem to depend as much on adaptation constants as input-signal fluctuation. Angular resolution of a regular antenna array is limited by the well-known Rayleigh criterion[7] for diffraction, also expressed by Fresnel in another form. A first cut at the 3-dB beamwidth of an array is:

$$Beamwidth = \frac{\lambda}{d} \text{ radians} \qquad \text{(Eq 11)}$$

where λ is wavelength and d is the aperture diameter. When a signal is received with a high SNR, further improvement in resolution is possible through a concept developed by W. F. Gabriel in the late 1970s.[8]

Since antenna nulls are always sharper than lobes, bearings may be obtained more accurately by seeking a null. Radio foxhunters have known this for a long time. Exact information about direction of arrival may be used to turn a sharp null into a sharp beam. The nulled output is simply subtracted from the signal received on a separate omni antenna, as shown in Fig 8, with weighting. The result is super-resolution. This system has seen little use in amateur circles, but it is expected to be a major part of software-defined radios that optimize performance in various increasingly hostile RF environments.

Notes

1D. Smith, KF6DX, "Signals, Samples and Stuff, Pt 3," *QEX*, Jul/Aug 1998.

2B. Widrow and S. Stearns, *Adaptive Signal Processing*, Prentice-Hall, Englewood Cliffs, NJ, 1985.

3Notice that these plots assume that all the "omnis" in the discussion are isotropic. That's why the patterns seem to have infinite nulls. The plots were generated with a *QBASIC* program the author wrote 10 years ago. Therefore, readers who want to duplicate the work must "crank" the antenna arrays into their simulation programs. They will likely get plots slightly more realistic, but similar to those shown here.

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7Lord Rayleigh (J. W. Strutt), "On the theory of optical images, with special reference to the microscope," *Philosophical Magazine, Part B* Vol 42, No 5, 1896 (ISSN 1364-2812).

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RF

By Zack Lau, W1VT

A Miniature HF 50:200 Ω Balun

This 50:200-Ω balun was designed for serious QRPers who want a lightweight HF balun using commonly available parts. It should handle 3 W on 20 meters, more than enough for "milliwatt" types who never run more than 1 W. While there are designs with less loss, rarely do they use popular parts found in most junk boxes. This balun uses the FT-37-43; perhaps the most popular ferrite toroid used in amateur construction projects.

The design is shown in Fig 1. First, a 4:1 unbalanced to unbalanced transformer steps up the impedance to 200 Ω. Next, a 1:1 transformer provides the desired unbalanced-to-balanced function. For simplicity, both transformers use 10 turns of #28 AWG enameled wire, bifilar wound. Bifilar transformer winding is quite simple if you know how. There is also a clever way to wind the transformers—if you want to eliminate a solder joint—that I'll cover later.

First, wind each winding side by side with no crossovers (see Fig 2). Next, strip the insulation from the two "center" conductors closest to each other and verify with an ohmmeter that they are different windings. Finally, solder them together.

The balun is even easier—the input wires go in together and they come out together. Most of the time it doesn't matter if you swap the output wires—so it isn't necessary to keep track of the wires. If you are feeding a phased array, however, phasing does matter. Swapping the wires may result in an unexpected 180° phase shift.

The clever way to wind the transformers is to use one 12-inch length and two six-inch lengths of wire, instead of four six-inch lengths. This way, after you wind the first bifilar winding, you have enough wire to wind the second core, with the longer piece of wire. I'd recommend winding the two six-inch lengths of #28 AWG enameled wire first, on separate cores. Then wind the 12-inch length on both cores. This is shown in Fig 2—the 12-inch wire is black. The six-inch wires are red, and should have a lighter shade in a black and white picture. Alternately, you can slip some thin Teflon tubing over a wire before it is spliced to the other core. The tubing can then be slid over the splice. With a bit of skill or luck,

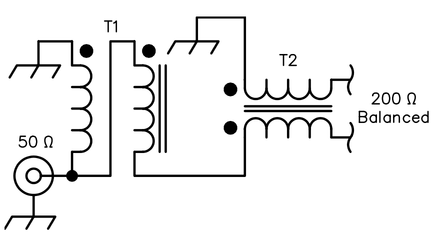

Fig 1—Schematic of a low power 50:200-Ω balun.
T1, T2—10 turns #28 AWG enameled wire bifilar wound on a FT-37-43 toroid.

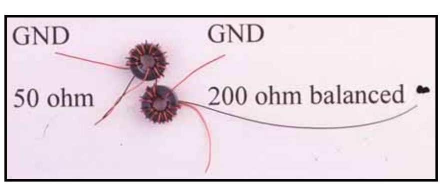

Fig 2—Details of the FT-37-43 toroid core windings.

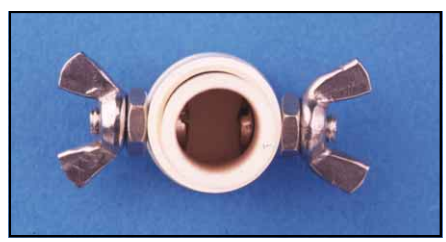

Fig 3—#10 Hardware installed in ½-inch CPVC pipe.

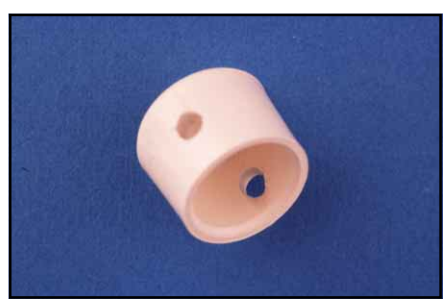

Fig 4—A machined CPVC pipe cap showing the carved hole (lower).

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the splice will be just the right thickness to hold the tubing snugly in place. I often use 18-gauge lightweight spaghetti from Small Parts.[1]

I've found that 1/2-inch CPVC tubing is just big enough to house this balun. The tough fit wasn't the ferrite cores, but the screw terminals. Number 8-32×1/2 screws just barely fit inside the pipe caps—I needed to angle one of the mounting holes so that the second screw would fit. With a mounting hole properly modified, even #10-32×1/2 screws are useable, as shown in Fig 3. Fortunately, CPVC is soft and easy to carve with a sharp knife. This is shown in Fig 4. Notice the lower hole: Plastic is carved away so it doesn't obscure the hole. I put on a pair of safety goggles and used a hobby knife to do the carving.

Choosing the location of the holes is a tradeoff between having enough clearance for the screw head and having a reasonably flat surface for the lockwasher. I made my holes about 0.2 inches from the interior flat surface of the cap. I put the cap on a section of 1/2-inch pipe and drilled them together. The pipe can be easily held in a vise, unlike the awkward shape of the cap. Unlike copper caps, the walls of the cap vary in thickness significantly. The holes in the pipe provide reference marks for milling notches for the screw heads. To get the pipe to fit around the screw heads, I milled 0.33-inch wide notches.

Mounting the UG-1094A/U connector was also a challenge. I decided the pipe cap was too small to properly tighten the 3/8×32 mounting nut. Instead, I first used a 3/8-inch bullet-shaped drill bit to cut a relief for the connector—just a little more than 50 mils should work fine. Otherwise, the connector won't screw in flush against the pipe cap. Next, I drilled a Q-sized hole (0.3320 inch) and tapped a 3/8×32 thread for the BNC connector. This extra-fine tap may be difficult to find—you may need to get it from a machine shop supply. I applied some Loctite thread-locker to the connector and firmly threaded the BNC connector into the pipe cap. When installing threaded BNC sockets, it helps to plug in an old BNC plug for protection. It also provides a knurled surface designed for gripping firmly with fingers.

I used solder lugs wrapped around the screw heads to attach the balun wires. As shown in Fig 5, bending them in a **U** maximized clearance for the short section of pipe cap. You might imagine installing them facing downward and then bending them so the tabs face upwards. I found it necessary to bend them before installation, however, it is too difficult to bend them in place. I soldered the balun to the solder lugs first. Then I slid the tubing over the balun, and firmly attached the pipe cap.

The final tricky part is soldering the wires to the BNC connector. I used a 60-W temperature controlled iron to solder the wire directly to the body of the connector, after first tinning the connector and wire. This is why I used the longer connector—the shorter UG-1094/U would be more challenging to solder. Finally, I soldered the other wire to the center conductor. I really doesn't matter which is the ground side and which is hot, so you may find it convenient to swap wires when making these final two connections, depending on the wire lengths.

This may seem like a lot of work to machine the three CPVC parts shown in Fig 6—it is. It is much easier to make a larger balun, perhaps using 3/4 or 1-inch plastic tubing and end caps. The

[1]Notes appear on page 59.

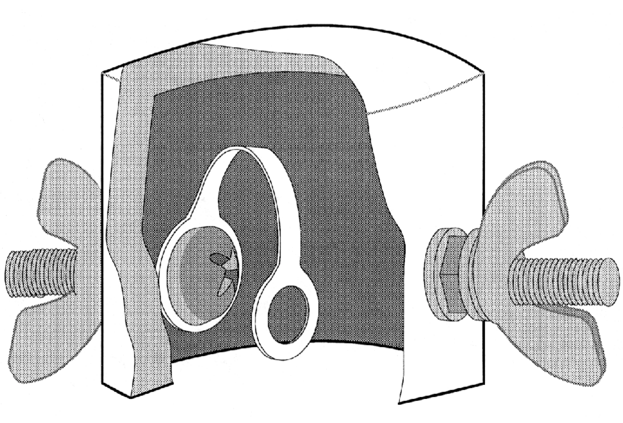

Fig 5—The arrangement of the screw terminals and solder lugs.

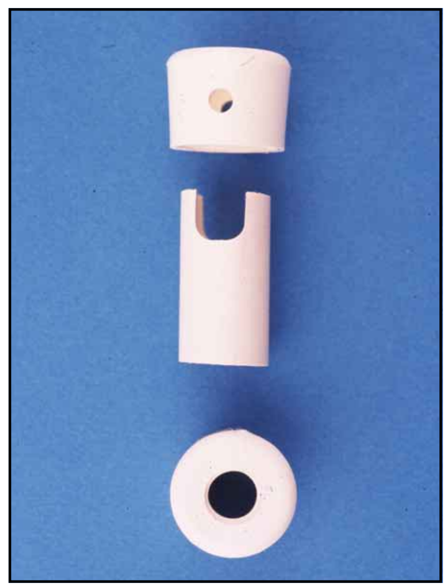

Fig 6 —CPVC pipe parts prepared to build the balun.

Table 1—Measured balun performance

Frequency MHz	*Transformer loss (dB)*	*1:4 Balun loss (dB)*	*Unbalanced Fig 3A (dB)*	*Unbalanced Fig 3B (dB)*	*Low Z loss (dB)*	*High Z loss (dB)*	*Balun RL (dB)*
1.8	0.29	0.66	1.06	1.06	0.18	1.31	22
3.5	0.21	0.43	0.68	0.67	0.17	0.75	27
7	0.16	0.32	0.47	0.47	0.24	0.53	33
10	0.15	0.29	0.42	0.42	0.38	0.51	34
14	0.15	0.27	0.38	0.39	0.67	0.54	31
28	0.15	0.29	0.38	0.39	1.34	1.31	24
50	0.18	0.39	0.41	0.43	2.32	2.76	18
60	0.19	0.39	0.46	0.48	2.38	3.38	17

result wouldn't be as small or light, however.

The performance is quite good when properly terminated, considering the low cost and lightweight of this design. The balun weighs just 0.070 ounces without the CPVC parts. The loss from 3.5 to 60 MHz is under 0.43 dB, for over 90% efficiency. The return loss is better than 24 dB from 3.5 to 28 MHz, degrading to 19 dB at 50 MHz, when terminated directly in a $^1/_4$-W 200-Ω carbon-composition resistor. The results shown in Table 1 are for the packaged balun in Fig 7. Not surprisingly, the extra lead length for the screw terminals degraded 6-meter performance slightly. Thus, this should work well on all HF bands when terminated in a 200-Ω load. It may even be useful on 6 meters.

The insertion loss was measured using a pair of baluns back to back, as shown in Fig 8. The loss was also measured with points A and B alternately grounded. Ideally, the loss wouldn't change. The RF signal generator was a Marconi 2041. The power meter was a HP 437B with an 8482A sensor. The return loss was measured with a Marconi signal generator, a Mini-Circuits ZFDC-20-5 directional coupler and a HP 8563E spectrum analyzer, as shown in Fig 9. The measurement data is shown in Table 1. The numbers shown represent half the total loss measured, since most people want to know what the loss is for a single balun or transformer. The accuracy of the HP power meter and sensor is ±0.02 dB, enough to justify the precision shown in the table.

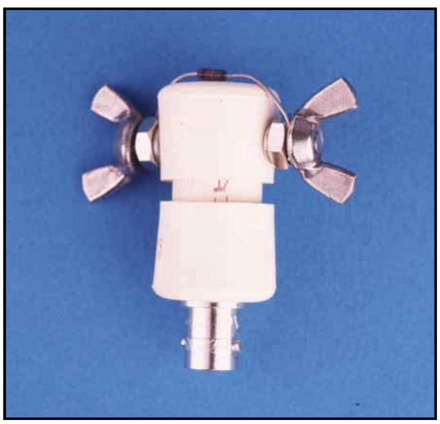

Fig 7 —The completed balun with a 200-Ω $^1/_4$-W carbon-composition resistor used for return-loss measurements. The balun is in a typical mounting position, with the load resistor across its top.

The last two columns are an attempt to characterize balun performance when the unbalanced input impedance is either 12.5 or 200 Ω. The impedance step-down was measured by simply swapping the input and output connections. The impedance step-up was a little more difficult—I measured the balun along with yet another 4:1 bifilar transformer—10 turns of #28 AWG on an FT-37-43 core. I subtracted the loss of the extra transformer for the table entries. The balun works quite well from 7 to 14 MHz, showing no additional loss. However, high-frequency performance was significantly degraded at both high and low impedances. The impedance step-down actually enhanced low-frequency performance. This is the result of the transformers operating more efficiently—while the wires used in the balun are short enough to avoid causing an impedance mismatch.

The loss can be used to estimate power handling. As described on page 2-21 of *QRP Power*,[2] the power loss required to raise the temperature of a toroid core is:

$$P_{loss}(\text{mW}) = (\text{surface area in cm}^2)(\Delta T)^{1.2} \quad \text{(Eq 1)}$$

An FT-37 core has a surface area of 5.7 cm^2, which means that 0.27 W are required to raise its temperature 25°C. If the core has a loss of 0.4 dB, 9% is lost as heat. Thus, the power handling capability is 0.27 Watts/0.09 or 3 W.

While amateur operation is intermittent, rather than 100% duty cycle, it may not be safe to significantly increase the rating if the cores are sealed in a plastic tube that prevents heat dissipation. On the other hand, a little wind will significantly increase the power handling—just ask anyone who has tried to solder outdoors on a windy day.

Further research is needed to determine the power handling capability on 6 meters. While the insertion loss is nearly 3 dB, much of it is the result of impedance mismatching—2.76 dB. It is likely that a tuner will allow the balun to handle more than 0.5 W, but exactly how much is unclear.

You need to be very careful when using baluns with phased arrays. A properly designed balun has outputs that are virtually indistinguishable from each other, except for relative phase between the input and output. This can be disastrous—an unexpected 180° phase shift can turn a high-gain array into a low-gain nightmare. Why doesn't it work? An incorrectly phased balun or antenna is an all too common answer. Thus, it is important to identify and mark balun output ports intended

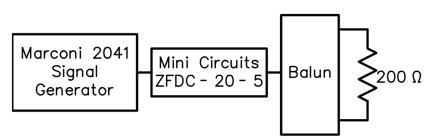

Fig 9—Schematic of the setup used to measure return loss of the balun when terminated in a 200-Ω carbon-composition resistor.

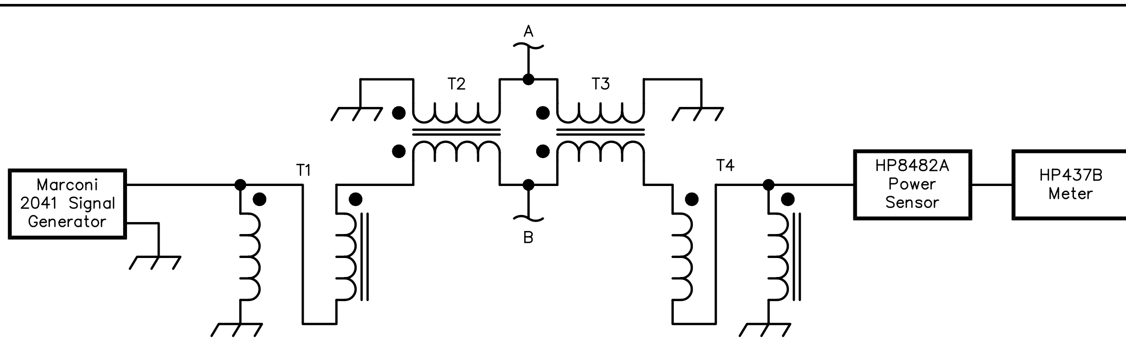

Fig 8—Schematic of the power-loss test setup using a signal generator and power meter.

for use in phased arrays, so they can be used properly. Antennas also have phase—it makes a significant difference if you flip yagis over.

Theoretically, it may be possible to construct identical baluns if you use colored windings and pay careful attention to detail. Realistically, most people need to measure constructed baluns and properly identify the output terminals. Fortunately, this is quite easy to measure, if you have the appropriate phasing harness. After all, if an in-phase harness works properly, any output terminal will be either in-phase or 180° out of phase relative to any other terminal. Thus, you can just connect a resistor of the appropriate wattage across any two terminals and see whether it heats up. If it does, the terminals are out of phase. If the resistor is stone cold, the terminals are in phase. I'd use the same resistance used to terminate the baluns properly. Don't touch the resistors while RF power is applied, always turn off the transmitter first.

There are two caveats to this simple test procedure. First, the resistor needs to be of low inductance. If you use a wirewound resistor, it may act as a RF choke, remaining stone cold no matter how it is hooked up. Thus, I'd make sure that the resistor heats up properly when connected to terminals that are suppose to be out of phase—don't just look at the in-phase connections.

The second is the influence of SWR-foldback circuitry. Theoretically, it may be possible for an aggressive foldback circuitry to reduce power enough to make the heating effect too small for conclusive results. Fortunately, such transmitters usually have an SWR or power indicator that will indicate the SWR change. In this case, a rise in SWR or drop in power should correspond to the out-of-phase connection. The in-phase connection should not change the power or SWR.

Notes

[1]Small Parts Inc, 13980 NW 58th Ct, PO Box 4650, Miami Lakes, FL 33014-0650; tel 800-220-4242 (Orders), 305-557-7955 (Customer service), fax 800-423-9009; e-mail **smlparts@smallparts.com**; Web site **http://www.smallparts.com/**.

[2]J. Kleinman, N1BKE, and Z. Lau, W1VT, Eds. *QRP Power,* (Newington, Connecticut: ARRL 1996; ISBN: 0-87259-561-7) Order No. 5617, $12. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Check out the full ARRL publications line at **http://www.arrl.org/shop/**.

Letters to the Editor

A High-Performance Homebrew Transceiver, Corrections and Improvements

In Part 3 (Nov/Dec 1999), on page 50, I suggested running the +7 dBm mixers with only +4 dBm of LO injection. This idea came from the manufacturer's manual in regard to operation "where dynamic range is not important." I had thought, since these were merely LO premixers handling no receive or transmit signals, that dynamic range was not a factor. However, even in this case there are spurious mixer products, which can be reduced using the higher LO level. IMD in the PTO mixer is of particular concern when the dual-receive function is enabled, when two input frequencies are applied. I have raised the levels in both the BFO and PTO mixers to +7 dBm, with a noticeable improvement. The modification required an additional MAV-1 amplifier ahead of the MAV-11 in Fig 5 for the PTO mixer. The LO level for this mixer is now adjusted by a gimmick capacitor used in place of the 1 pF coupling capacitor in the 34.285 MHz filter shown in Fig 4. (This is a very sharp, single-frequency filter using under-coupled resonators; the capacitor consists merely of a short, heavy wire connected to one coil, and positioned near the other coil.) For the BFO mixer in Fig 4, it was sufficient to adjust the resistor between the MO and the MAV-11. Further improvement in mixer performance is clearly possible; see Chapter 6 in Wes Hayward, W7ZOI's, *Introduction to Radio Frequency Design*.[1]

The bias resistor for each MAV-11 was reduced to 270 Ω. I also found that the RFCs in the supply leads to the MMICs could be eliminated with little change in gain and a probable reduction in any tendency to parasitics or stray radiation. In Part 4 (Jan/Feb 2000), on page 48, I explained that the 47 Ω resistors in the LO lines to the product detector (Fig 4) and balanced modulator (Fig 9) were included to enable measurement of LO power. This was a bad idea; it increases the required LO power, worsens the spurs caused by stray LO energy, and still does not provide an accurate measurement. The usual practice is to disconnect the LO from the doubly balanced mixer (DBM), connect a 51 Ω resistor from the LO amplifier output to ground and measure the level with a scope. The trimpots previously used to adjust LO levels have been eliminated; the levels are now adjusted by selecting components in the BFO amplifier (Fig 3). Thanks to George Cutsogeorge, W2VJN, for all these corrections and suggestions concerning DBM operation.

Some builders may have difficulty locating surplus Signal/One crystal filters for the IF board at 8815 kHz. Those found at flea markets are up to 30 years old and of questionable quality. (I was lucky to find a few good filters in a large junk-box supply.) An easy substitute would be new filters from International Radio, 13620 Tyee Rd, Umpqua, OR 97486; tel 541-459-5623, fax 541 459 5632; e-mail **INRAD@rosenet.net**; **www.qth.com/inrad/**. The IR filters designed for Kenwood radios at 8830 kHz would work splendidly. The filter matching circuits will need a slight change, the two BFO crystals will need replacing, and the 25-kHz-bandwidth noise-blanker filter will need to be homebrewed. No other circuits in the radio will require modification. The reason for this flexibility is the premixing scheme (described in Part 1, Mar/Apr 1999) in which the tunable BFO is used to obtain IF-shift operation, but is not a factor in determining the transceiver frequency.

Another correction applies to the frequency counter in Part 5 (Mar/Apr 2000). In the caption to Fig 13, I twice referred to an IC as type 7473; it should say 7493. Thanks to Ulf Edlund, SM3CUX, for pointing out this error.—*Mark Mandelkern, K5AM, 5259 Singer Rd, Las Cruces, NM 88005;* **k5am@zianet.com**.

A Simple, Rapid and Precise Method of Finding True North Using GPS

GPS receivers have become ubiquitous in the ham-radio community because of their well-known abilities to provide exact location and altitude information. When used in conjunction with a simple plumb line, though, they also possess the ability to precisely and quickly provide the exact direction of true north at any location during daylight. This avoids any dependence on magnetic compasses or deviation tables; they also avoid errors caused by nearby ferromagnetic structures or dc-powered equipment.

This ability comes from the sunrise-sunset readings available on a GPS. For any given location, sunrise and sunset times vary during the year; for a given time of year, they vary with latitude and longitude of the observer.

For any given day and location, though, sunrise and sunset times may be used to determine the exact time of local noon. That is the time when the Sun is highest in the sky. It is also midpoint time between sunrise and sunset readings, added to the sunrise time. It is the exact time when a plumb-line shadow will fall on a precise north-south line.

Determining local noon is quite easy. For example, at my location on September 7, local sunrise and sunset are 12:11 and 00:48 GMT, respectively. This translates to 07:11 AM and 7:48 PM, CDT. The time between sunrise and sunset is 12 hours 37 minutes or 757 minutes. Dividing this number by two and adding it to the sunrise time yields local noon. This calculates to 12:11 GMT plus 6 hours 18.5 minutes or 18:29.5 GMT (13:29.5 CDT).

This is almost 1½ hours past 12 PM on local clocks, but don't be surprised. The variation at your location depends on (1) your position in your time zone and (2) whether daylight-savings time is in effect.

If the Sun is not visible at local noon, you can still use this method. Mark the plumb-line shadow at whatever time the Sun cooperates and record the exact time when you mark the line. Correct the azimuth angle of the shadow line by the amount your time record is different from local noon. The Sun moves about 15° per hour, or 0.25° per minute. In northerly latitudes, the shadow moves almost this much. Calculate the angle the Sun moved before or after local noon, then redraw the shadow line using a protractor. This system is often considerably more accurate than can be obtained with a magnetic compass.

Microwave enthusiasts will find this system particularly useful because most hilltops don't have true-north references. Dish aiming requires exact azimuths.—*Robert Templin, PE, W5OE, 605 Robin Dale Dr, Austin, TX 78734.*

Hi Robert,

Thanks for that and I would add this: Get your measurement as close to local noon as you can to avoid errors caused by variations in latitude and time of year. Near the equator and the

[1] Order No 4920, $30. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Check out the full ARRL publications line at **http://www.arrl.org/shop/**.

North Pole, errors are maximized. A vertical antenna mast may be used in place of a plumb line is most cases—73, Doug, KF6DX

New Super-Regenerative Circuits for Amateur VHF and UHF Experimentation (Sep/Oct 2000, to Charles Kitchin, N1TEV)

Congratulations and thank you for one of the most informative, interesting and practical homebrew articles I have ever read in any electronics publication. Surely this is one of those articles that make my *QEX* subscription so very worthwhile.

When I finish my current project (converting a GE MASTR II 450 MHz receiver to 222 MHz for a local repeater) I look forward to the fun of trying some of the concepts presented in your article. Adequate NBFM reception on a regenerative receiver—Who woulda thought it! Three "Atta-Boys" to you and the others who worked this out.

Overall, the article certainly illustrates that those old-timers often had very good ideas and that revisiting some of them using modern components and ideas can yield surprisingly good results. No response is necessary or expected. Just wanted to express my thanks.—*Tom McKee, K4ZAD, 104 Water Leaf Ln, Cary, NC 27511-9728;* **mckee@deltacomm.com**

Upcoming Conferences

Satellites and Education Conference XIV, 7-9 March, 2001

The Satellites and Education Conference is an annual conference focused on introducing educators to a wide range of learning opportunities available through satellites. It provides a forum for innovative educators to share their ideas with others for implementation in the classroom. Individual workshops and lectures hosted by speakers who are leaders in their field are provided over a three-day period in the spring. Workshop sessions accommodate all educators from elementary school teachers to university professors. Plenary-session speakers have included leaders in government, education, professional associations, media and industry. In the past, we have had an interactive audio-conference with Canterbury, England, and interactive sessions with Oklahoma State University. These sessions have enabled attendees to become familiar with distance-learning opportunities as they utilize communication satellites. Vendors also exhibit the latest low-cost satellite tracking equipment available to educational institutions.

For more information, visit the conference Web site at **http://www.sated.org/eceos/ttsc.htm** or contact Nancy McIntyre, West Chester University, 189 Schmucker Science, West Chester, PA 19383; tel 610-436-2393, fax 610-436-3045; e-mail **nmcintyre@wcupa.edu**.

Southeastern VHF Society Call for Papers

The Southeastern VHF Society will host its fourth annual conference on April 20-21, 2001, in Nashville, Tennessee, at the Holiday Inn Brentwood. This is the *first call* for presentations to be made at the 2001 conference and papers to be published in the conference *Proceedings*. Papers should be submitted to Dick Hanson, K5AND, (7540 Williamsburg Dr, Cumming, GA 30041; tel 770-844-7002, fax 770-889-8297; e-mail **k5and@ga.prestige.net**) for review by February 20, 2001. *This is the drop-dead date.* Papers may be submitted in hard copy or, preferably, on diskette in MS Word 7 format. We can also convert Word Perfect files; pictures are best in black and white. Be sure to number figures, graphs, drawings and pictures so that we can match them up with the references in the body of the article.

Next Issue in QEX/Communications Quarterly

We've had trouble getting in contact with M. A. Chapman, KI6BP, to discuss details of two articles he submitted to *Communications Quarterly* that we'd like to publish. If anyone knows him, please let him know we are trying to reach him and that we hope all is well.

To start 2001, we have a piece from Nathan Sokal, WA1HQC, and Richard Redl updating us on class-E amplifiers. To go with that, James Buckwalter, KF6SWC, and his Caltech friends have put together a keyed, switch-mode power supply that looks like just the thing for powering those amplifiers.

Steve Best, VE9SRB brings us a derivation of the transmission-line equation that begins with purely wave-mechanical arguments. We think this proof brings insight to what happens on transmission lines and shows that more than one viewpoint can be correct. Robert Brown, NM7M, con-tinues his series on LF propagation by examining the effects of atmospheric ozone. It looks like R. P. Haviland, W4MB's next piece on quad antennas will also appear.

2000 Index

Issues 198 through 203
(All dates are 2000 unless otherwise indicated.)

Tech Notes (Bertini)

Upcoming Conferences

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United States Postal Service

Statement of Ownership, Management, and Circulation

1. Publication Title	2. Publication Number	3. Filing Date
QEX	0886-8093	September 27, 2000
4. Issue Frequency	5. Number of Issues Published Annually	6. Annual Subscription Price
Bimonthly in: Jan, Mar, May, Jul, Sep, Nov	6	$22.00

7. Complete Mailing Address of Known Office of Publication *(Not printer) (Street, city, county, state, and ZIP+4)*
225 Main St., Newington, Hartford County, CT 06111-1494
Contact Person: Debra Jahnke
Telephone: (860) 594-0297

8. Complete Mailing Address of Headquarters or General Business Office of Publisher *(Not printer)*
225 Main St., Newington, Hartford County, CT 06111-1494

9. Full Names and Complete Mailing Addresses of Publisher, Editor, and Managing Editor *(Do not leave blank)*
Publisher *(Name and complete mailing address)*
David Sumner, 225 Main Street, Newington, CT 06111-1494
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13. Publication Title: QEX
14. Issue Date for Circulation Data Below: Sep/Oct 99 - Jul/Aug 99 Sep/Oct 00

15. Extent and Nature of Circulation	Average No. Copies Each Issue During Preceding 12 Months	No. Copies of Single Issue Published Nearest to Filing Date
a. Total Number of Copies *(Net press run)*	7,934	9,401
b. Paid and/or Requested Circulation (1) Paid/Requested Outside-County Mail Subscriptions Stated on Form 3541. *(Include advertiser's proof and exchange copies)*	6,046	5,963
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c. Total Paid and/or Requested Circulation *[Sum of 15b. (1), (2),(3),and (4)]*	7,711	7,617
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h. Copies not Distributed	55	1,697
i. Total *(Sum of 15g. and h.)*	7,934	9,401
j. Percent Paid and/or Requested Circulation *(15c. divided by 15g. times 100)*	97.87%	98.87%

16. Publication of Statement of Ownership
☒ Publication required. Will be printed in the Nov/Dec 200 issue of this publication. ☐ Publication not required.

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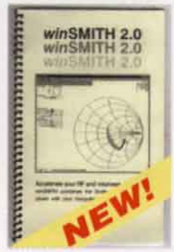

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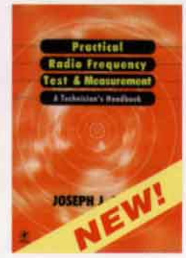

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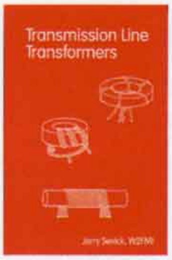

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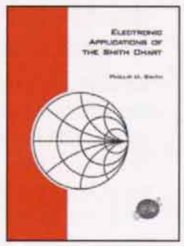

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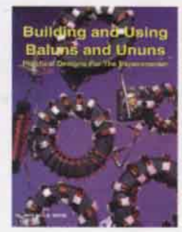

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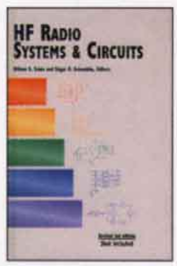

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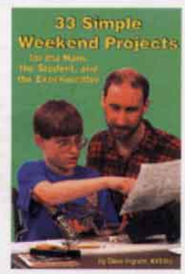

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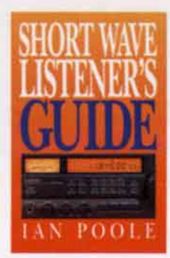

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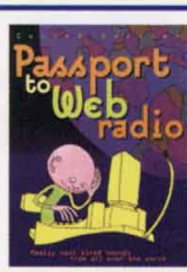

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