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THE JOURNAL OF COMMUNICATIONS TECHNOLOGY

Summer 1995

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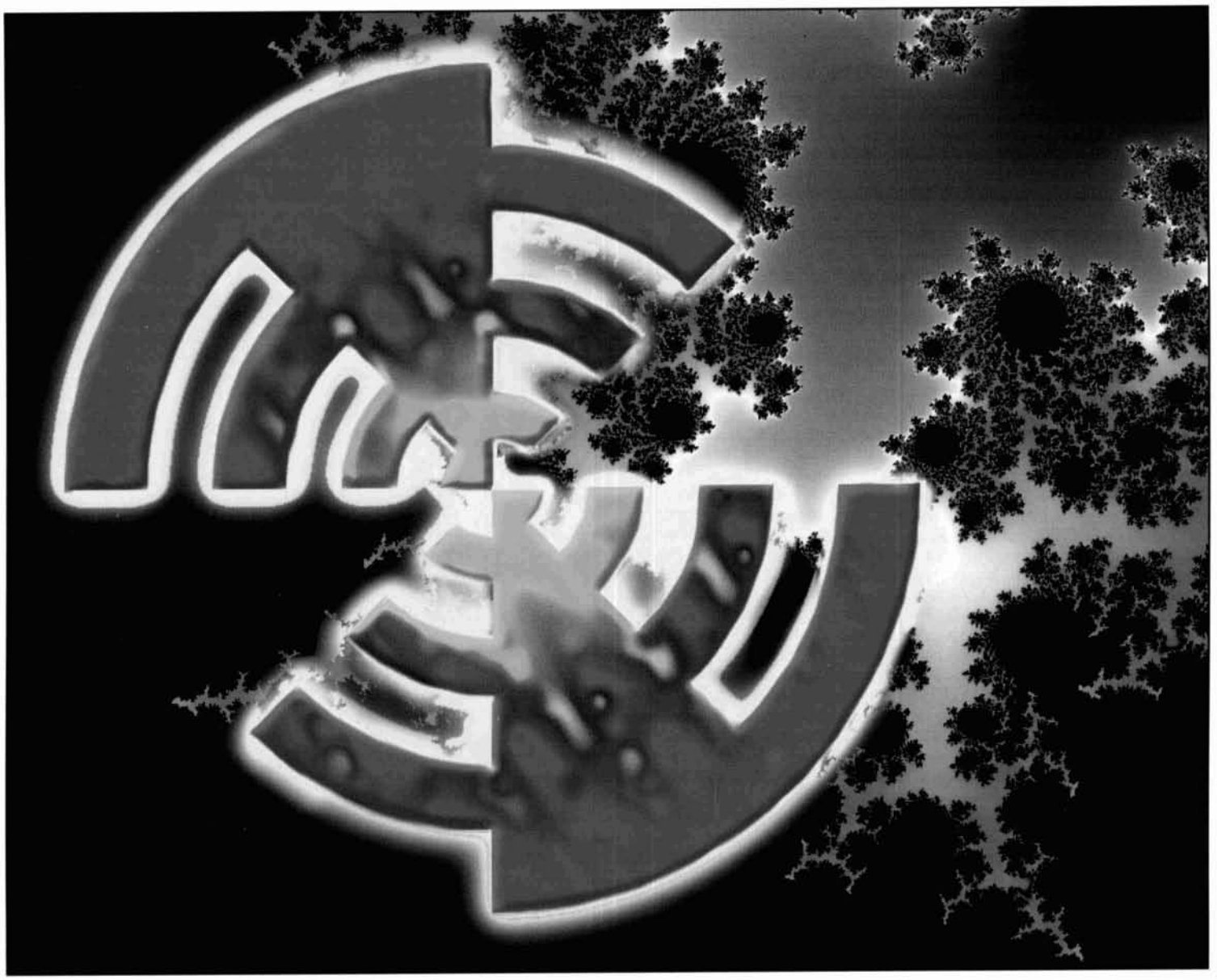

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- A Few Words About DSP
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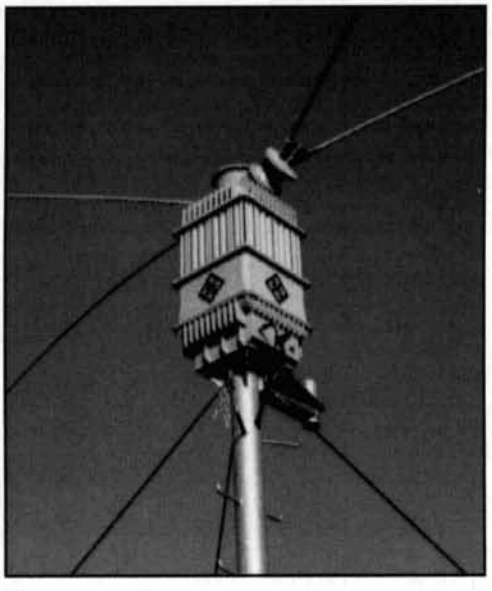

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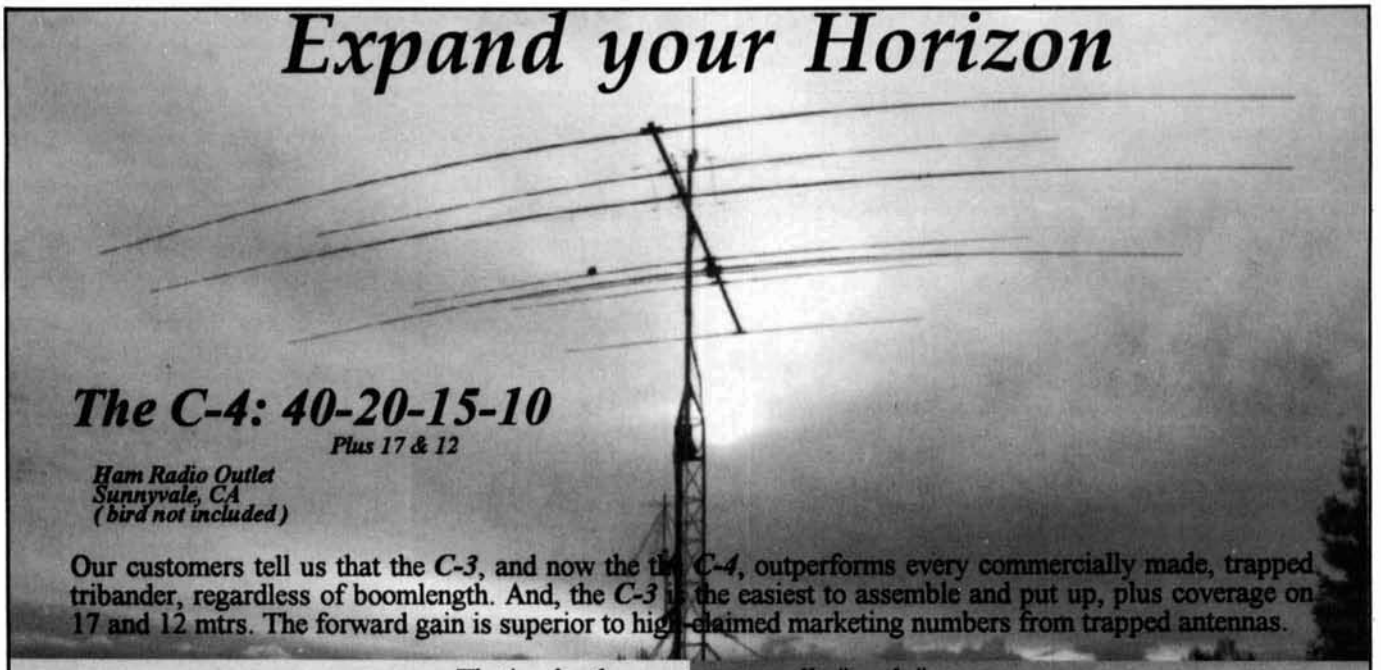
On the cover: Most of us, when we think of antennas, focus on such Euclidean shapes as rectangles, triangles, and squares. In this issue, N1IR would like us to think fractally—as our cover by Bryan Bergeron, NU1N, reflects. For all the details, see "Fractal Antennas" beginning on page 7.

Expand your Horizon



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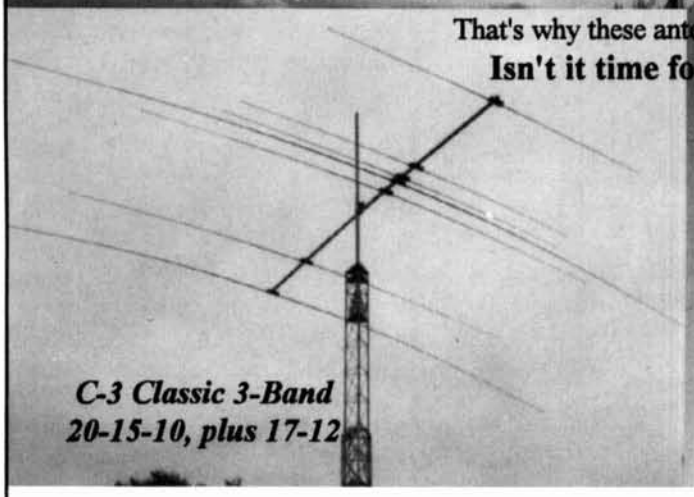

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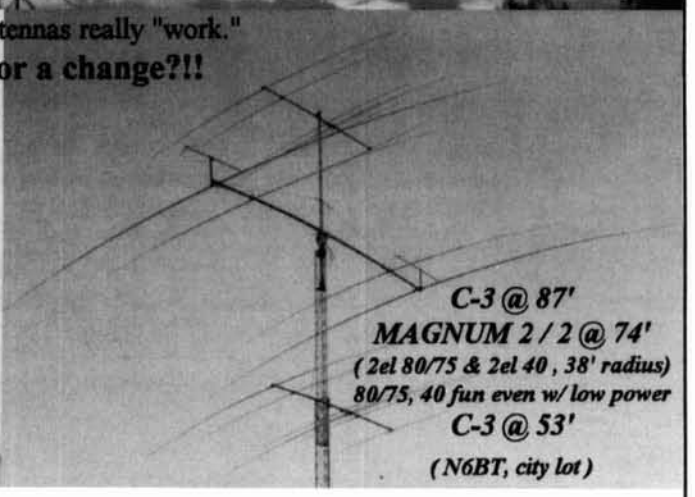

C-3 @ 87'
MAGNUM 2 / 2 @ 74'
(2el 80/75 & 2el 40 , 38' radius)
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C-3 @ 53'
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EDITORIAL

The Fractal Antenna: a new challenge for antenna experimenters

Not every radio amateur will build a receiver, transmitter, power supply, or any of the other accouterments of the "complete" ham station; however, most have tried their hand at putting up an antenna. These creations range from wires launched over the limbs of trees with bows and arrows, rocks, or fishing poles, to elaborate stacked arrays with dozens of elements. Some DXers I know have incredible mobile antenna setups mounted atop their vans, which are rotated either by mechanical means or the "armstrong method." Still other avid VHFers travel to rare grid locators with satellite dishes in tow.

On the more aesthetic side, my husband's Uncle John, W1ENE, has a yard full of hardware—all elegantly landscaped with birdbaths and flower beds. Joe Reisert, W1JR—a bird lover—adorns the elements of his antennas with bird feeders to attract his fine feathered friends. He's even had birds try to nest amongst the elements. At our QTH, tucked beside a small lake in New Hampshire, K1BQT and I have a "featherweight 6-meter beam" (see this issue's "Tech Notes" beginning on page 5), plus 75 and 20 meter dipoles, a 2-meter vertical, a 220 vertical, one for the aircraft band, and a 6-meter vertical. As a matter of fact, there's so much antenna activity around here that I'm constantly harvesting pieces of black electrical tape, feedline, aluminum tubing, and rope from the lawn.

Because of the ham's eternal search for the perfect antenna, a whole industry has grown up around antenna design and construction. Instead of the old "cut and try" method, we employ SWR analyzers, noise bridges, and the increasingly popular antenna optimization programs. These tools, in addition to the overwhelming number of designs found in the amateur radio literature, help us homebrew antennas tailor-made for our locations.

Everyone has a favorite antenna: verticals, dipoles, Zepps, Quads—you name it, we build them and extol their virtues to all who will listen. However, I imagine that the majority of us have never given a thought to building a fractal antenna—billed as a very small, highly efficient antenna whose design is based on fractal geometry. A goof-ball idea? Perhaps. It's something I'd never encountered before I received "the manuscript" in the mail. Inside was a photo of a "fractal quad"—a construction that looked more like a Coptic cross than an antenna. The author (Chip Cohen, N1IR) claimed it really resonated, and I was intrigued enough to pass his work along to you, his peers, for comment. For more details, turn to "Fractal Antennas: Part 1" beginning on page 7.

Now, Chip is well aware that there will be skeptics, perhaps many of them, who think his idea is crazy. All he wants is a chance to put forth his theories and share his ideas with other interested hams. Chip would also like to encourage as many of you as possible to construct one of his designs and report back to him with your findings. In Part 1, Chip presents his fractal quad. Those who prefer other antenna types might want to hold off until they check out some of the designs presented in Part 2.

Who knows, maybe Chip has something here, or maybe he doesn't. That's what he's daring us to find out. After all, our hobby has always been about experimenting: venturing to try the unproven, the unusual, the seemingly impossible. Bob Lewis, W2EBS, once told me that every antenna article should run with the caveat that what works for the author might not work for the reader. On the flip side, he said that as long as the reader had received his warning, we should go ahead and run all kinds of antenna articles. I'm taking his advice.

Okay, I know you're all itching to get started; so, open to page 7 and begin reading. I'm challenging you to build one of these little fractals and report back with your results. Are you up to the challenge?

Terry Littlefield, KA1STC
Editor

P.S. By the way, if you just can't "do" the fractal, consider building G2AJV's toroidal antenna (page 23) or K1BQT's "featherweight" beam, instead!

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A publication of
CQ Communications, Inc.
76 North Broadway
Hicksville, NY 11801-USA

Editorial Offices: P.O. Box 465, Barrington, NH 03825. Telephone/FAX: (603) 664-2515.
Business Offices: 76 North Broadway, Hicksville, NY 11801. Telephone: (516) 681-2922. FAX: (516) 681-2926.

Communications Quarterly is published four times a year (quarterly) by *CQ Communications, Inc. Communications Quarterly* is the philosophical successor of *Ham Radio Magazine* founded by T.H. "Skip" Tenney, Jr., W1NLB and James R. Fiske, W1HR. Subscription prices (all in U.S. Dollars): Domestic—one year $29.95; two years $56.95. Canada/Mexico—one year $34; two years $65. Foreign—one year $39.95; two years $76.95. Foreign Air Mail—one year $80; two years $155.

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Second-class postage paid at Hicksville, NY and additional mailing offices.

Postmaster: Please send change of address to Communications Quarterly, CQ Communications, Inc., 76 North Broadway, Hicksville, NY 11801. ISSN 1053-9344.

Printed in U.S.A.

TECH NOTES

Summer Projects

Here are a couple of summertime ham activities for those of you who are tired of the usual fishing and camping trips, visits to Disney World, or oceanside clambakes on Cape Cod. Six meters has been hot lately, boosted by sporadic-E. Here's a nifty little antenna that K1BQT uses to get out of his New Hampshire QTH and still stay cool.

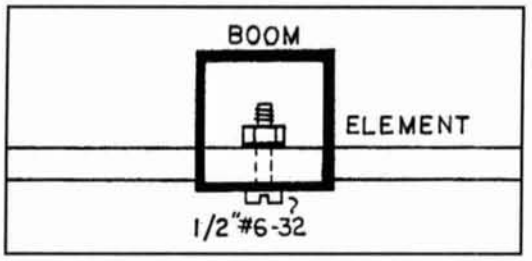

Figure 1. Element mounting detail.

A Featherweight 6-meter Beam

Rick Littlefield, K1BQT

With the advent of 6-meter operation in Europe, 6-meter SSB is experiencing something of a rebirth in the United States. Many imported HF rigs now include 6-meter coverage, and domestic companies like Ten-Tec and MFJ are introducing new products for the 50-MHz band. About a year ago, I got involved in this trend and needed an antenna to get on the air. Unfortunately, my home QTH is on a lake surrounded tall trees and hills—not prime VHF real estate! The featherweight beam described here evolved as part of a scheme to erect an antenna as high as possible using only an unguyed chimney mount for support. You probably won't find this approach to antenna raising in anyone's antenna handbook, but it certainly seemed like an interesting challenge to me!

Antenna design

Lacking an antenna optimizer program, I searched for a cookbook solution to establish a "baseline" for my short-boom Yagi. The 23rd edition of Bill Orr's *Radio Handbook* yielded a simple 3-element equal-spaced array on a 6-foot boom (page 25–15). This provided appro-

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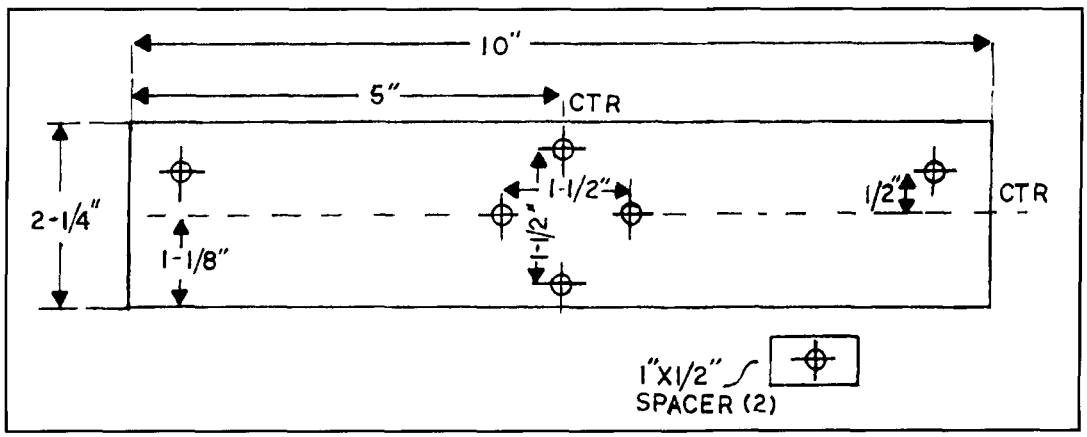

Figure 2. Center block drilling pattern.

priate element lengths, so I scaled them to my target frequency and corrected k for 5/16-inch elements. However, I decided to do some experimenting and substitute a hairpin match* for the Delta match called for in Orr's design.

The theory behind the hairpin-matched driven element is simple. First, you shorten a standard-length radiator slightly to introduce capacitive reactance. Then you split the element to provide a balanced feedpoint, and add a coil or hairpin across that point to introduce inductive reactance. Working together, these artificially introduced C and L elements form an L network—a familiar circuit used in solid-state transmitters and antenna tuners for impedance transformation. In this particular application, the objective is to transform the low driven-element impedance (typically 20 ohms or less) to a more useful value; usually 50 ohms to match coax line or 200 ohms to match the balanced side of a 4:1 balun.

Armed with these basics, I set out to build my antenna. The game plan was simple—to use the corrected element lengths for my reflector and director, then determine exact dimensions for the driven element and hairpin experimentally after the antenna was constructed. The resulting project is strictly a low-budget, kitchen-table concoction. No fancy tooling or special materials are required.

Construction details

The boom is made from a 6-foot length of 1 x 1 x 0.047-inch square aluminum stock, available at most hardware stores under the "MR" label. To prepare the boom, first mark its midpoint—36-inches from either end—to locate where the driven element center block will be mounted. Next, place marks—on center-line—at 32-1/2 and 34-1/4 inches from one end. Drill two 5/16-inch holes completely through the boom at these locations. These holes, spaced 1-3/4 inches apart, will accept a standard TV-type U-bolt assembly for mounting the antenna to its mast. On this same surface, drill a 5/16-inch element-mounting hole 1/2-inch from each end of the boom. Drill these holes off-center, as close as possible to the edge of the aluminum stock (see **Figure 1**). Now, cut the reflector and director elements from 5/16-inch thinwall aluminum tubing stock. Make the reflector 117-1/2 inches (58-3/4 inches either side of center) and the director 107 inches (53-1/2 inches either side of center). Scribe the mid-point of each to provide a centering mark for use during final assembly. The antenna's driven element is split in the mid-

(Continued on page 98)

* Thomas Cefalo, Jr., WA1SPI, "The Hairpin Match," *Communications Quarterly*, Summer 1994, page 49.

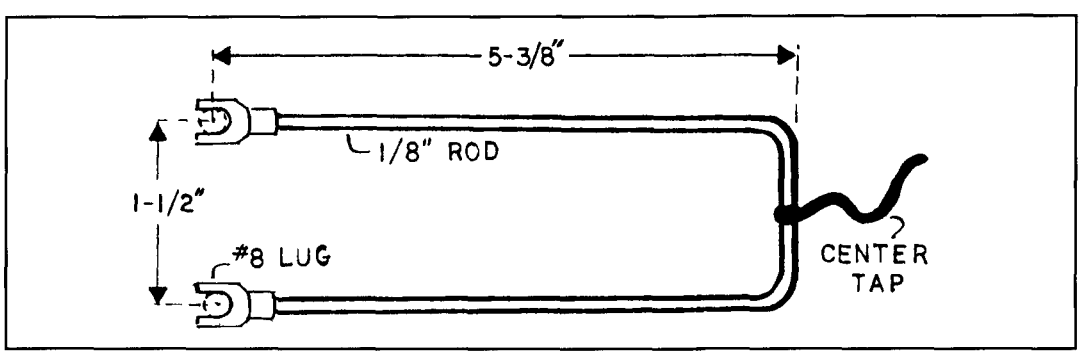

Figure 3. Hairpin match.

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Fractal Antennas Part 1

By Nathan "Chip" Cohen, N1IR

2 Ledgewood Place
Belmont, MA 02178

Summary: An article describing fractal geometry and how it applies to antenna design. Fractal quad designs are given as examples.

Summer 1995 issue, pages 7-22.

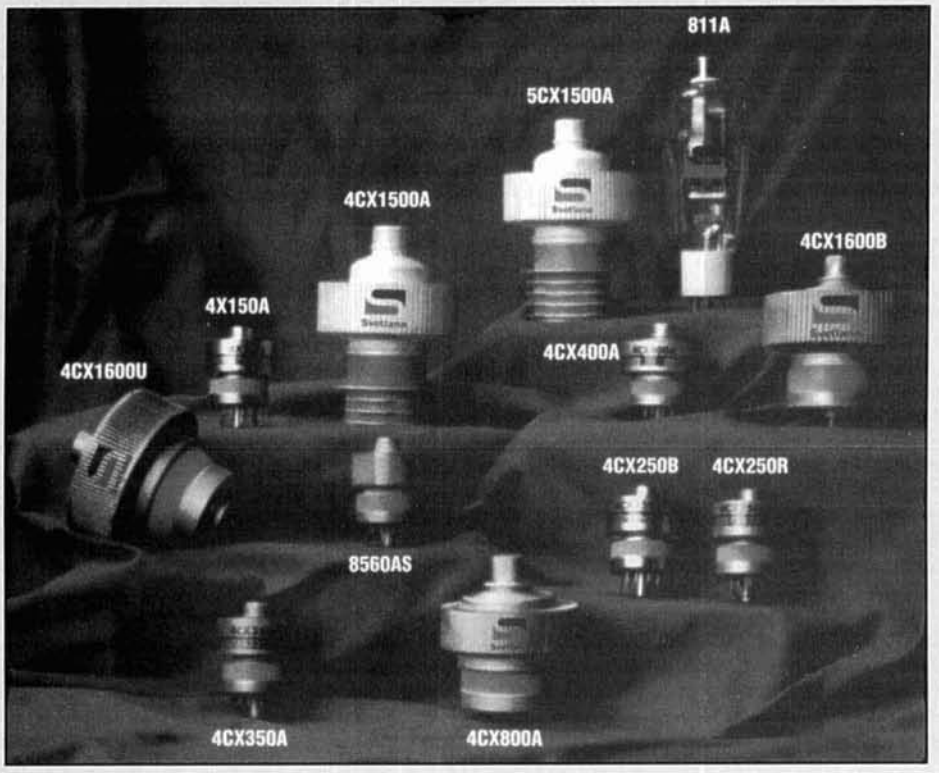
Please contact the author for additional information.

PRODUCT INFORMATION

New Power Tube Literature from Svetlana

Svetlana Electron Devices, Inc. has published a new synopsis of Svetlana power tubes for amateur radio. The list includes tubes which are generally available in Russia but are new to the Western market. It also includes many exact drop-in replacements for the Western tubes appropriate for amateur radio linear amplifiers. Major operating characteristics and maximum ratings are also provided.

Svetlana also publishes a list of plug-compatible drop-in replacement products for broadcast and industrial service. These lists may be obtained, at no cost, by contacting either of their offices. Their headquarters office address is: Svetlana Electron Devices, Inc., 8200 South Memorial Parkway, Huntsville, AL 35802; phone 205-239-6900; or 205-882-1344; fax 205-880-8077; toll free 800-239-6900. The Marketing & Engineering office address is: Svetlana Electron Devices, Inc., 3000 Alpine Road, Portola Valley, CA 94028-7582; phone 415-233-0429; fax 415-233-0439; toll free 800-5-SVETLANA (800-578-3852).



Roger C. Jennison, *G2AJV*
"Wildwood," Nackington,
Canterbury, U.K.

THE G2AJV ANTENNA

And Maxwell's displacement current

We expect toroidal coils to retain the magnetic field within their structure. However, we may sometimes forget that when a toroidal coil is fed with an alternating current, there is a considerable electric induction field threading through the torus and the space around it. With an air-cored torus, this electric induction field extends for a considerable distance. It even may be detected beyond a range of one wavelength from the torus, where the wavelength is dictated by the frequency of the alternating current.

The nature of the electric induction field is precisely the same as that which occurs between the plates of an air-spaced capacitor. However in this case, it comes about indirectly because of the circulation of the magnetic field in the toroidal coil. James Clerk Maxwell called this varying electric field a displacement current because early capacitors, such as Leyden jars, usually contained a dielectric that showed a charge displacement when placed in an electric field. However Maxwell went on to show, theoretically, that although space was apparently empty, it nevertheless had a dielectric constant, or permitivity. Therefore the concept of the displacement current is still applicable—even if there is apparently no physical polarization of strictly empty space. The displacement current generated by the alternating conduction current in a toroidal coil is used in ordinary electrical systems when we thread any

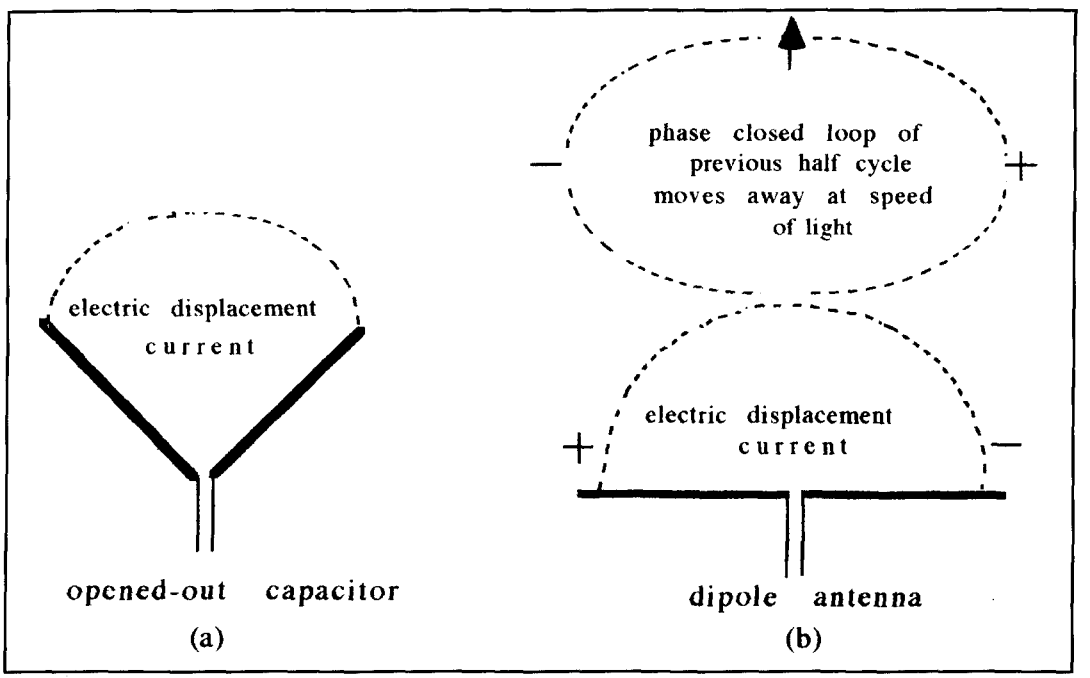


Figure 1. (A) The electric displacement, shown only as a dotted line for a single value, completing the circuit between the ends of an opened-out end-fed capacitor. (B) A similar drawing for a dipole antenna that also shows a closed loop of displacement current from the previous half cycle, now untethered, moving away at the speed of light.

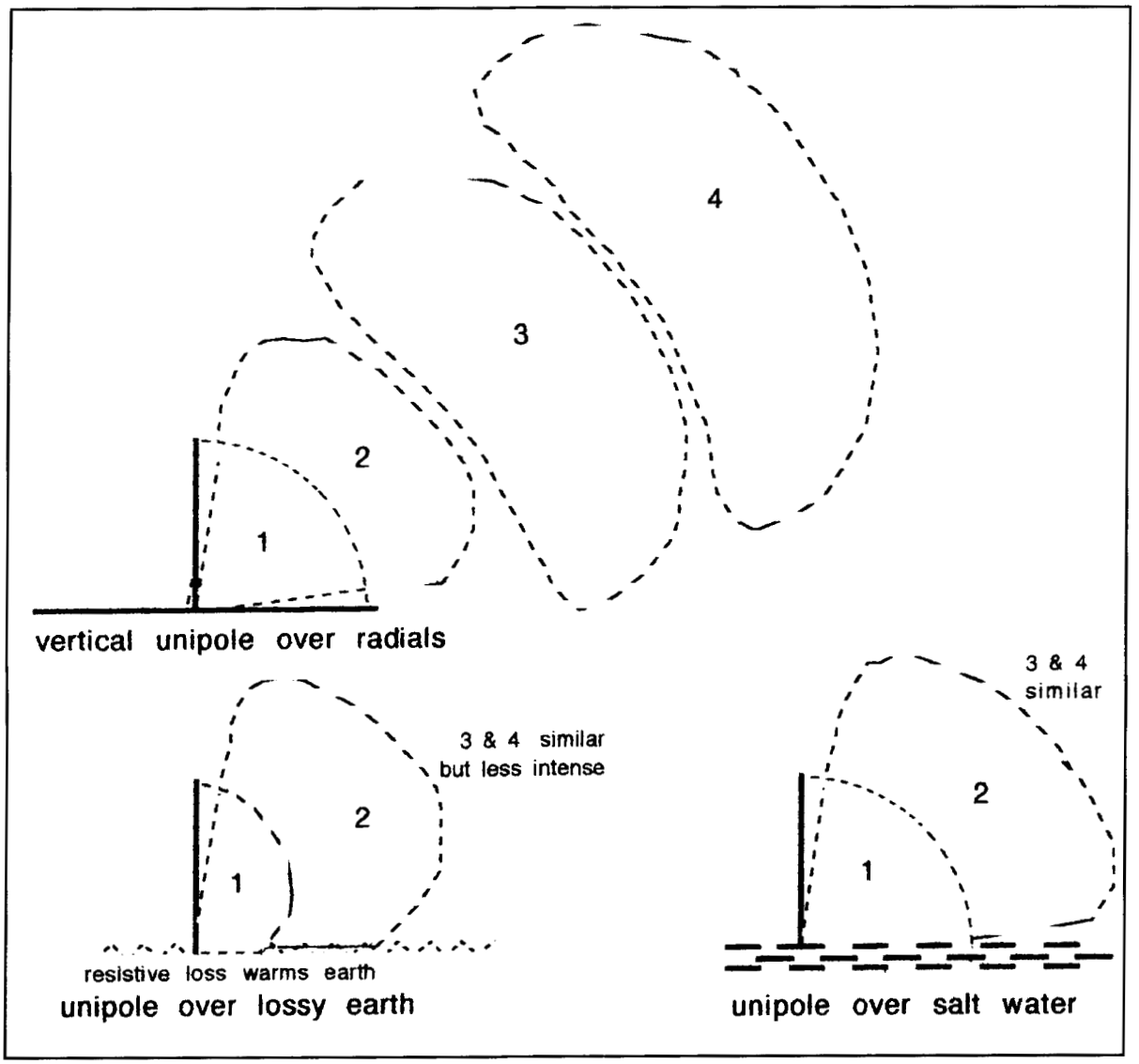


Figure 2. The stages in the development of the radiation field from the local induction field of a resonant unipole antenna in various situations.

length of a single wire through the torus. The displacement current is induced into the wire, giving rise to a conduction current. This conduction current may be detected if there is a load between the two ends of the wire, so the system acts as a transformer.

The radiation process

When we connect a transmitter to a wire dipole, we also obtain a displacement current that appears as an electrical induction field emanating at a right angle to the wire on each side of the center. Looping through the surrounding space onto the other side of this displacement current completes the current circuit (see **Figure 1**). There is also a magnetic induction field that is not trapped within the structure, as in a torus, but appears in the form of closed circles in the space around the wires. We know from our experience that signals are radiated from this system. People have been doing it ever since the first experiments of Hertz and Oliver Lodge. But how and why are the signals radiated?

A decade or two before the first experiments, James Clerk Maxwell predicted that a component of a local oscillating electromagnetic field would not be retained in the manner of the local induction field, but would radiate into space at a speed given by the ratio of the electric and magnetic units. He showed that this ratio was so close to the speed of visible light that the two must surely be the same.[1]

Maxwell started with four equations, three of which were already known. The fourth was his

own. Nowadays, we tend to refer to all four equations as "Maxwell's Equations" because they are the major mathematical formulae relating electromagnetism. In fact, we can reduce the ingredients to just two equations because the remaining two are always identically equal to zero in a propagating wave. The fact that they *are* zero provides important information about what the wave *does not* have. It does not have electric charge (Div**E** = 0) and it does not have free magnetic poles (Div**B** = 0). Just remember that and you can forget these first two equations—provided the wave is propagating.

The two equations that give rise to the radiative component are Curl**E** = –d**B**/dt and Curl**H** = **j**+d**D**/dt. Translating the first of these from vector algebra into ordinary language, it states that the electric field curls around and that this produces a varying magnetic field. The second equation is Maxwell's own. It states that the magnetic field curls around, and that this curling can produce two types of current: the ordinary conduction current, signified by **j** (if conductors happen to be present), and a varying electrical displacement or *displacement current*, signified by d**D**/dt. As we have seen, a displacement current is one that flows through the dielectric in a capacitor. Maxwell showed theoretically that, although space was apparently empty, it nevertheless had a dielectric constant or permitivity. This means that the concept of displacement current is still applicable even if there is no physical polarization of strictly empty space. When a wave is propagating in empty space, there are no conductors. As a result, there is no conduction current and the second equation becomes simply Curl**H** = d**D**/dt. This may be written Curl**H** = d(ϵ_0**E**)/dt, which mirrors the first equation as Curl**E** = –d**B**/dt may be written Curl**E** = –d(μ_0**H**)/dt where ϵ_0 and μ_0 are the permitivity and permeability of empty space, respectively.

Therefore, the propagation of an electromagnetic wave through space depends on the mutual relationship exhibited between the almost symmetrical equations for a curling magnetic field

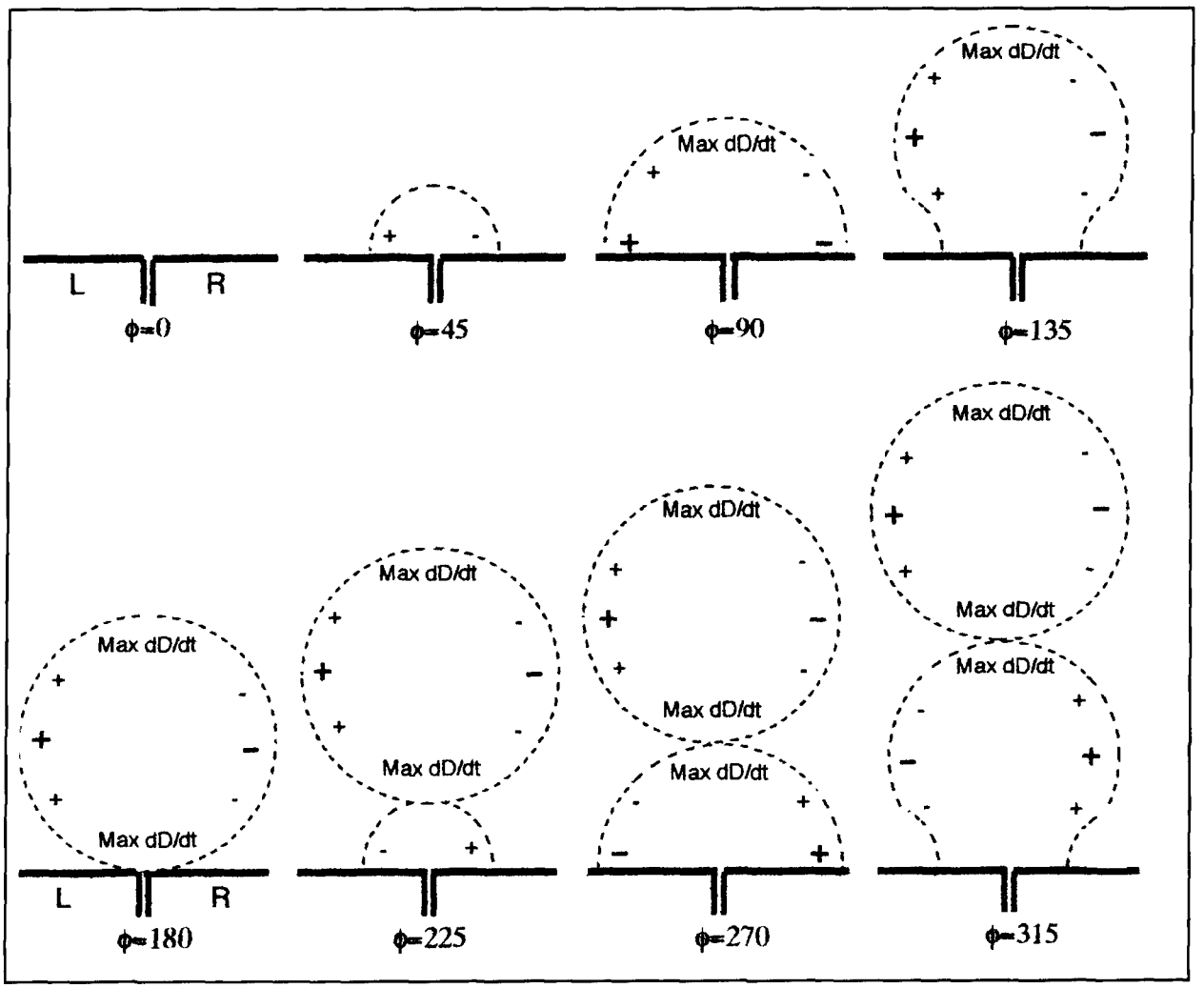


Figure 3. Detail of the formation of a single quantum of radiation from a resonant dipole antenna.

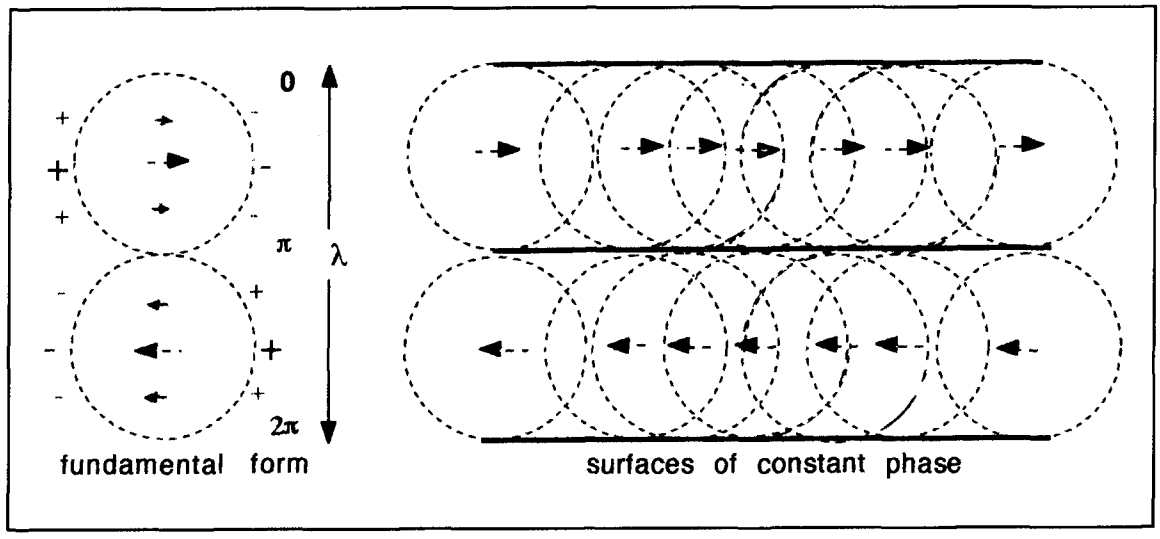


Figure 4. The formation of wave fronts from coherent photons. The photons merge into common envelopes with oppositely directed displacement currents every half cycle.

and a curling electric field. It is interesting to note that the right-hand side of the equation for the latter looks exactly like a magnetic current, as distinct from an electric current (see box).

The resulting electric and magnetic fields of the propagating wave system are mutually at right angles to one another and in the same phase—unlike the typical standing wave system on a dipole antenna. It becomes a little confusing to try to portray the electric and magnetic currents on the same diagram, so henceforth I will confine the discussion to the electric displacement current.

Figure 2 shows practical examples of displacement current launched from a unipole antenna under various circumstances. When the antenna is isolated, the induction field grows outwards reaching the tips of the conductors in stage 1. It then balloons out to stage 2, when the input wave form decreases, and finally crosses over to the opposite polarity. The process repeats with stage 1 in the opposite polarity, and the original stage 2 is nipped off to form a closed loop that moves to position 3, and so on. I have also used the remaining illustrations on the same diagram to depict the situation over a semi-infinite conducting surface—the sea, and over lossy earth. These diagrams will be useful later.

In many accounts of electromagnetic waves, one gains the misleading impression that the wave fronts at great distances from the transmitter represent entirely separate electric fields of alternately opposite polarity and no mutual connection, which stretch in straight lines until they peter out in the far, far distance. We shall shortly see that this impression is not true; the radiation always consists of precisely determined loops, the leading edges of which determine the wave fronts.

Photons by the score

Now we come to the big question. How can these fields propagate forever through empty space if they carry no batteries, charges, or other apparent generators? The answer is really very simple.[2] They form complete *phase-closed* loops, the phase-length of which is always one wavelength—irrespective of the length of the wave front. This may seem impossible, but it isn't.

Figure 3 shows the development of a minimal signal transmitted from a simple dipole antenna. The Φ represents the phase angle of the input at each stage of the process. Note that the displacement current always completes the circuit in the correct phase while it is attached to the antenna; however, the upper ring of displacement current that has become detached and moves off into space must itself become phase closed. Incidentally, because it is only a rate of change of the field vectors, the displacement current moves off simultaneously from both the positive and negative electrodes. Unlike a conduction current, it does not require current carriers to dictate the direction of the flow.

Now look at **Figure 4**. On the left it shows an elementary pair of phased-closed loops, with opposite polarity, coupled together to form a radio frequency photon. On the right, it shows a coherent set of photons that combine to form extended loops with negligible phase change along the wave front. The phase around the long loops ultimately closes around the end of the final photon. These displacement current

Maxwell's Equations for Waves in Free Space

Div**D** = 0 *(meaning that there are no free charges in the wave)*
Div**B** = 0 *(meaning that there are no free magnetic poles)*
Where Div *is an abbreviation for divergence and therefore these equations simply mean that there is no spreading of the field in all directions as there could be from a point.*
D *is just the permitivity,* ε_0*, times the electric field,* **E**;
B *is just the permeability,* μ_0*, times the magnetic field,* **H**.

The following two equations give rise to the propagating wave:*
Curl**H** = d**D**/dt = d(ε_0**E**)/dt
(A curling magnetic field gives an electric displacement current.)
Curl**E** = –d**B**/dt = –d(μ_0**H**)/dt
(A curling electric field gives a magnetic "displacement" current.)

*Strictly, these are partial derivatives and should be written with a curly d, but my typewriter will only print an ordinary lower case d.

loops carry within themselves the resonant energy of the wave system that is "frozen" into the system while it is traveling at the speed of light, but which is released on encountering a receiving antenna or other obstacle.

You will have noticed that it is the closure of the electrical circuit via the displacement current that completes the current circuit when you excite an antenna. That is where some of our troubles may lie. Any extensive resistance in the local induction field, such as the earth in the lower left diagram in **Figure 2**, can absorb quite a lot of the power. The remainder is propagated to distant parts in the elongated loops of the weakened radiation field, which spread wider and wider until they break up into individual photons.

At very great distances from the transmitter, far out amidst the stars, some of the radiation transmitted 10 years ago from your 100-watt transmitter still lingers. However, because the radiation is always diverging as it proceeds (the inverse square law), the photons at very great distances no longer overlap their fellows on either side, but proceed independently at the same speed. Nevertheless, each photon remains phase closed and nothing can interfere with this state unless it collides with matter that can absorb its energy. This is an interesting point. Individual photons are not weakened by the inverse square law, for their phase must always be closed and they cannot shrink; they just become separated further from their original neighbors. This means that one day, in about ten thousand million years, a little of the signal that you are now transmitting will still be extant and heading for the ends of the universe.

It may seem strange that we are discussing photons in a radio context, but radio astronomers are very much at home with the

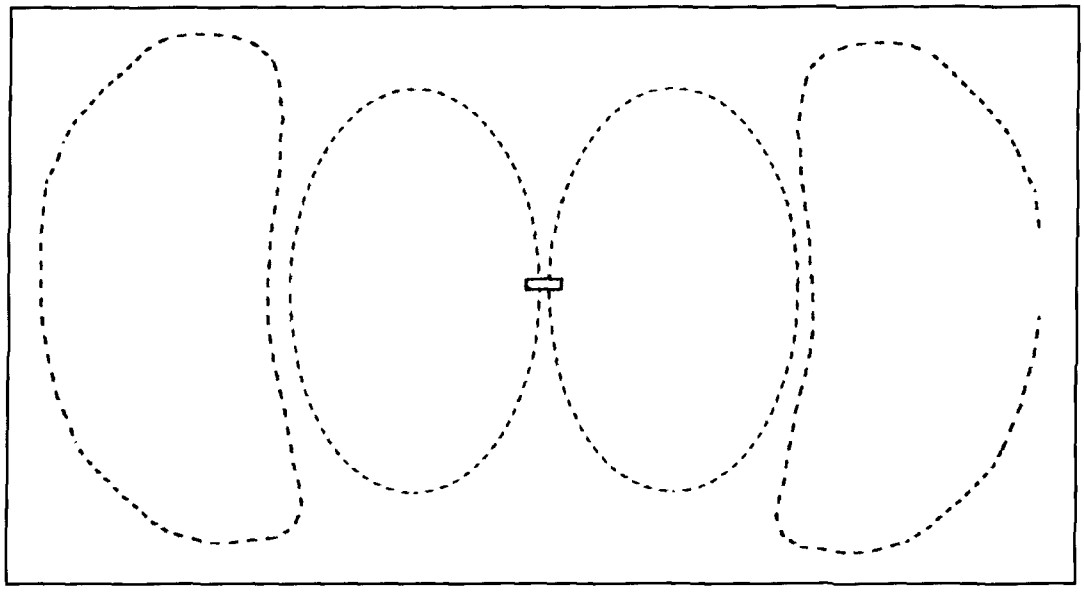


Figure 5. The radiation from a simple toroidal antenna. The current emerges parallel to the axis and the radiation is equatorial, at a right angle to the axis.

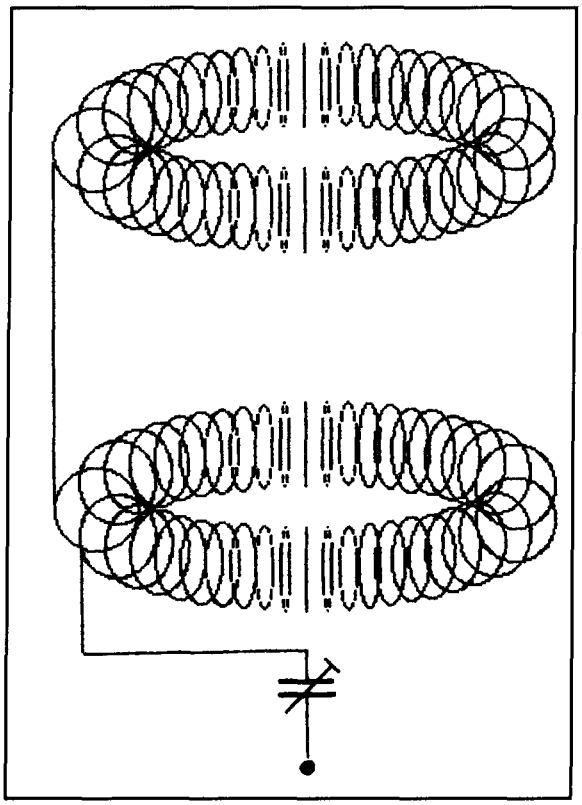


Figure 6. A cascaded toroidal antenna for mobile operation. The preset capacitor at the base is used for trimming the resonance in situ.

concept. They have been receiving them from distant regions of the universe for many decades. Incidentally, it has been shown that when a photon arrives on an antenna or other suitable target it is momentarily equivalent to a virtual electron-positron pair spaced apart by half a wavelength.[2]

Toroidal antennas

From our brief account of Maxwell's equations, you will have noticed that it should be possible to launch a signal from either an electric or a magnetic displacement current. A typical antenna uses both—although short linear dipoles are primarily electric. As we have seen, the electric induction field is much more prone to loss because of the presence of local electrically lossy materials such as the earth; this deprives the radiation field of much of its energy and gives a false indication of the significance of the standing wave ratio. Magnetically lossy materials are less common, so the loops of magnetic displacement can survive much closer to the ground. That is why electrically screened magnetic loop antennas can operate in these conditions. Nevertheless, antennas that rely entirely upon the generation of electrical displacement current have their uses. This displacement current is greedy for anything that will help to dissipate its energy. If all else fails, it will form associated loops of magnetic current and radiate into space—just as we saw from Maxwell's equations. This can be encouraged by the provision of a *good* conducting surface or wires in the immediate locality, as indicated in **Figure 2**. We shall return to this aspect later.

Many years ago I realized that toroidal coils ought to radiate, and I demonstrated this to my students. I originally called the devices Maxwell antennas because the essential principle was embodied in his equation Curl**H** = d**D**/dt. All I did was to develop the Maxwell principle into practical radiating systems, but the editors of *Radio Communication* bestowed my callsign upon them.[3] It also transpires that James Corum, K1AON, had independently observed that toroids could radiate. I understand that he took out a patent on some applications, although we appear to have differed slightly in our interpretation of the underlying principle. The loops of displacement current emanating from an isolated toroid are indicated in **Figure 5**. Incidentally, trying a ferrite-cored toroid is of little use. The local permeability inhibits radiation, as is apparent from the term Curl**H**, and not Curl**B**, in the displacement current equation.

Air-cored toroidal coils produce an intense electrical displacement current and, remarkably, are relatively easy to match to a 50-ohm cable. These coils are also exceedingly compact and may be a godsend to those who have nowhere to erect an antenna of standard size. Unfortunately, contrary to what one might expect, they are primarily electric and, unless very loosely wound, are not magnetic radiators. As a result, one must beware of local absorption—although loose electric coupling to a good local conductor, even if it is close to, but not touching, the ground, can be an advantage. On the positive side once again, although the bandwidth of the resonance is quite narrow, such coils will continue to radiate efficiently over a much wider band on the low frequency side of resonance and various simple A.T.U.s can be used to retain a good match. If they are well designed, air-core toroidal coils will handle quite high powers. This can compensate somewhat for their radiation efficiency which, on the HF bands, does not quite compete with that of a good half-wave dipole at reasonable height.

In the loft of my house, I have toroidal dipoles operating on 80, 17, and 15 meters, and I have had numerous transatlantic 'phone QSOs on all these bands. I have also had transatlantic 'phone QSOs on 15 and 20 meters with unipolar double toroids only 10 inches high sitting on the washing machine and fed barefoot from the

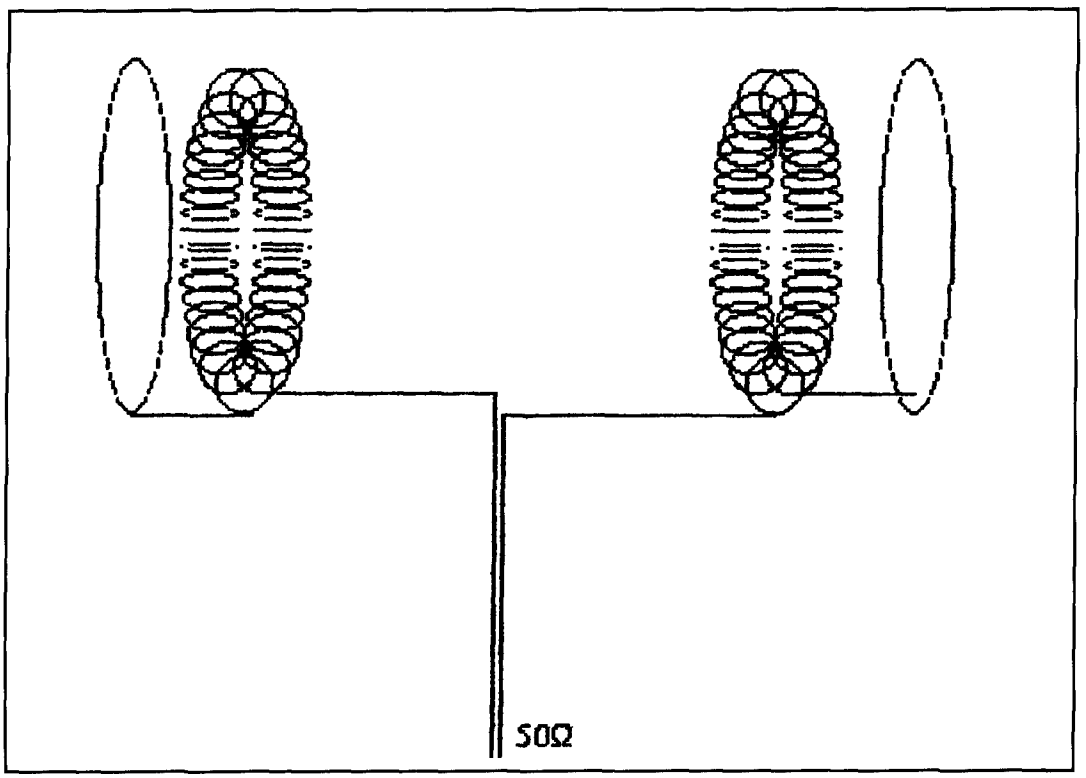


Figure 7. A pair of toroids in dipole configuration. The end plates resonate the coils, but also affect the matching.

transmitter two feet away in the kitchen on the ground floor of my house. So if you have thought that moving into confined premises might bring an end to your amateur radio activity, all is not lost. You can still operate, albeit with somewhat reduced efficiency; and that may partly be recouped by adding a linear amplifier. On 160 meters, I have had a little more trouble due to losses in the very extensive induction field. This might well be less troublesome with seaborne operation, but I have not tried that. On the other hand, the efficiency is considerably improved at VHF. This is partly because the induction field is more confined. For mobile use, the vehicle provides a good ground and less lossy environment. The mobile versions of these antennas are very satisfactory. They have the advantage of being so diminutive that one does not have to lower them when driving into the garage, and vandals do not associate them with expensive radio equipment. Actually most people think that they are a new form of lamp!

Some practical antennas*

A typical mobile antenna for 2-meter operation is shown in **Figure 6**. It is made from rigid copper wire and is self-supporting, with the bottom end of the capacitor connected directly into a standard plug. The final turn of the top coil is open circuit, unless you have used a length of wire that is too short. In this case you can terminate it in a small metal plate, no bigger than a postage stamp. The total length of wire, including the straight sections, is approximately half a wavelength. This is more important than the diameter of the turns on the coils, which can be approximately half an inch.

Wind the coils as straight solenoids on a $^{7}/_{16}$ inch or slightly smaller mandrel. They will probably expand slightly when released. Remove them from the mandrel and then pull them round, as neatly as you can, to form a torus. The trimming capacitor (about 2 to 10 pF) permits fine tuning for resonance in the center of the band—irrespective of the particular vehicle or rig—provided that the antenna socket mount is well bonded to, or has a high capacitance to, the roof of the vehicle. There is no connection from the antenna to the outer conductor of the cable. The antenna is about two and a half inches high, but should be contained in a thin-walled plastic tube to protect it from the elements and vandals. In the U.K., the well-known store "Woolworth" stocks small microwave "Keepers" beakers that have waterproofed sealed lids and are excellent for the purpose. One mounts the plug on the outside of the lid and feeds a wire to the capacitor carrying the coils through a small hole in the center

*The designs pictured and discussed in this article are copyright 1993: Emeritus Professor R.C. Jennison, "Wildwood," Nackington, Canterbury, CT4 7AY, U.K. (This does not apply to do-it-yourself construction.)

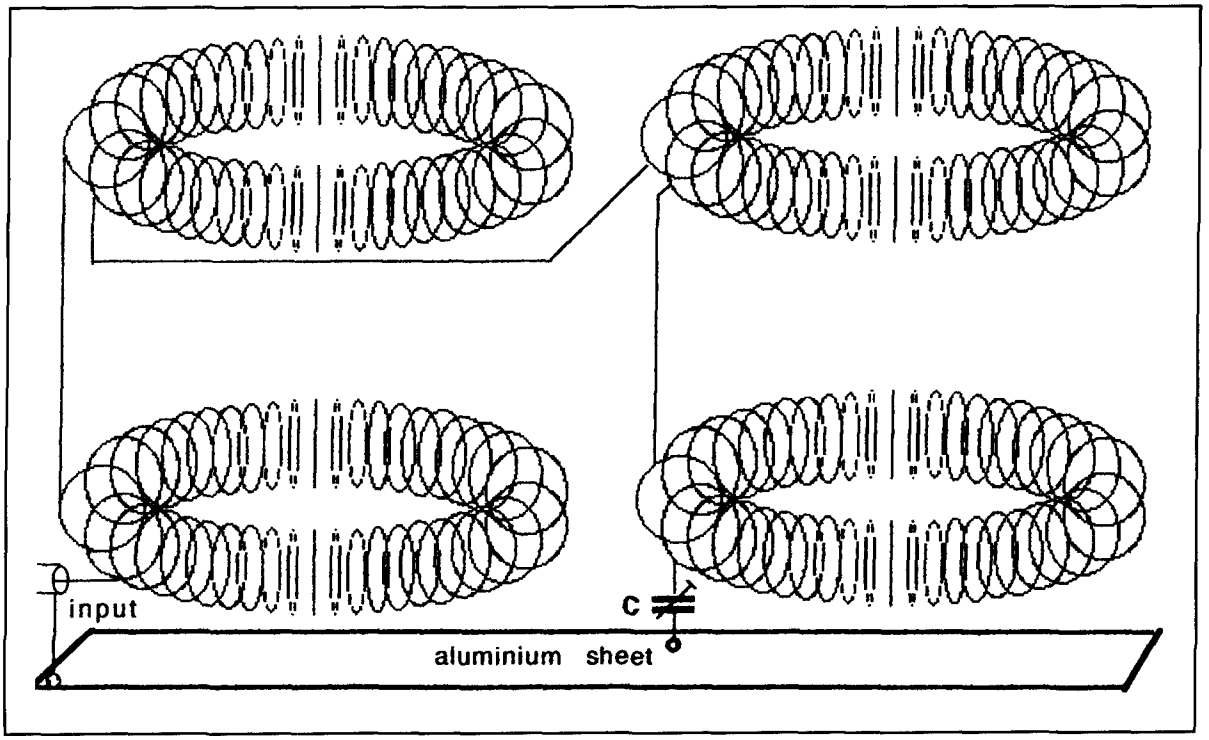


Figure 8. An experimental folded dipole.

of the lid. The inverted beaker is then snapped in place over the top. Trim for resonance when the antenna is finally installed on the vehicle. Allow for the fact that the cover, when replaced, will probably detune it about 2 MHz to the low side. If your construction is a little floppy, you can fit a little pad of soft foam plastic between the top coil and the beaker. Gauge the thickness carefully to avoid compression of the toroidal system.

The antenna may be plugged into a magnetic or a gutter mount, but the distance above the metal roof of the vehicle affects the impedance and the plastic beaker lowers the resonance by about 2 MHz. These VHF antennas are vertically polarized, and if properly trimmed in situ, they should provide a very good match to the transmitter and their radiation will rival a standard whip.

Precisely the same principles of construction,* appropriately scaled, may be used for mobile antennas on most of the HF bands and for some indoor applications where an appropriate ground surface may be available (e.g., my experiences with the washing machine). I am not usually a mobile operator on HF, but I decided to try the version in **Figure 6** and I have had 'phone contacts with Arizona using 14 and 21-MHz types when on the road. Peter Dodd, G3LDO, the technical editor of *Radio Communication*, has also had some success in this area.[5] There is a slight problem in the physical support for these larger versions. It is tempting to mount them on a central support, such as one and a half or two-inch diameter plastic pipe for the 14, 18, and 21-MHz versions, but the field at the center can be quite intense and many plastics can be quite lossy, ultimately incurring flash-over and the risk of fire. It is safer to mount the system from the outside and I leave that to your ingenuity.

There are a variety of alternative arrangements for toroidal radiators. Perhaps the most obvious is the balanced dipole. These work quite efficiently in the 30 to 3.5 MHz range when mounted in the loft, but are less effective close to the ground. They can be achieved with pairs of cascaded coils, similar to **Figure 6**, but duplicated on the opposite side of the feed. Alternatively, one can use the simple arrangement in **Figure 7**. The two free connections of the toroidal coils are each connected to metal plates about 6 inches from the coils. The size and spacing of these plates will have a marked effect on the resonance and the terminal impedance and you will have to experiment to optimize the performance in your own QTH. If you get it right, the cable will be perfectly matched. The resonance of this center-fed system is often sharper than that of the end fed variety, which was designed for operating very close to the ground; however, you should be able to get a good match without any extra antenna tuning components.

The basic polar diagram of the horizontal dipole is a "figure-eight" pattern, and the elec-

tric polarization is roughly parallel to the common axis of the toroids. In view of the intense local field, it is sometimes advantageous to feed the antenna via a balun—and you may like to experiment with the optimum transformer ratio. On most bands a 1:1 transformation is optimal, but the feed impedance is somewhat dependent on the relationship between the precise number of turns and the size and spacing of the end plates. I have had a number of 80-meter transatlantic 'phone QSOs with this configuration in the attic.

My antennas on 15, 17, and 20 meters use the same configuration and are wound with two-inch diameter coils of hard drawn copper wire formed into toroids. The 15-meter version has 18 turns on each coil and the 17 and 20-meter versions have 21 and 30 turns, respectively. I recommend that you add at least one extra turn to these figures if you build your own; it is much easier to snip off a turn than to add an extra one if you just miss the band. The inner diameter of the toroids is about one and a half inches; this is not critical. The tuning plates are about five and a half inches, six and a half inches, and eight inches in diameter, respectively. If they are too small, just add a bit of aluminum foil to increase the absolute capacity of the plates. A useful dodge is to form each plate from either three or four thin aluminum discs drilled and loosely bolted together about three quarters of the distance along a diameter. These may be opened out to form a clover leaf of appropriate size. The coils may be spaced as far apart as you wish. These dimensions should serve as a guide for the other bands. You will need rather large coils for 80 meters. I have made a variety of dipoles for this band using the same 2-inch diameter format. With 100 turns one can get a good match, but the end plates are rather large and clumsy—typically eighteen or more inches in diameter. With 150 turns the plates are relatively small, but the feed impedance goes up, and I have had to use a 4:1 balun. Nevertheless, I have been using the latter model as my standard 80-meter antenna for some years, despite the fact that I have acres of garden!

"Top band" (160 meters) is another story. I have already mentioned the extensive induction field and earth losses at this frequency. It is easy to get a good match to 50 ohms, but most of the power slightly warms the ground in the surrounding acre and you can't detect where it is going other than by comparing the local and far field components. The cascaded toroid system shown in **Figure 6** does work, but results are disappointing. I have tried many other configurations, such as the folded cascade system shown in **Figure 8**. This arrangement was conceived to increase the intrinsic radiation resistance and thereby, hopefully, decrease the ground loss. Feeding it between the plate and the lower toroid, it still matches quite well into 50 ohms if the toroid is only an inch or so above the plate, but the radiated power still leaves much to be desired. Nevertheless, both of these configurations will benefit considerably from the virtual earth technique that I later applied to a single toroid, as described further on in this section.

In my early experiments, I had noticed that the intense local field at the lower frequencies induced large currents on the outside of the coaxial feeder cable. This effect was inconsequential with the mobile antennae and washing machine systems, as the length of exposed feeder was trivial. However, in the more remotely connected systems it interfered with my field measurements, so I kept the feeder cable as short as possible and/or reduced the effect by using a balun. I also noted that, on top band, the antennas were often much more lively if the end of the cable at the antenna was not connected to the local earth beneath the antenna. The induced current emanating from the toroid onto the outer sheathing of the cable was clearly encouraging the radiative component by completing the return path with a good conductor and thereby depriving the earth of a little bit of warmth.

To investigate this phenomenon, I ultimately reverted to two of my early single-toroid antennas and experimented on how to turn the effect

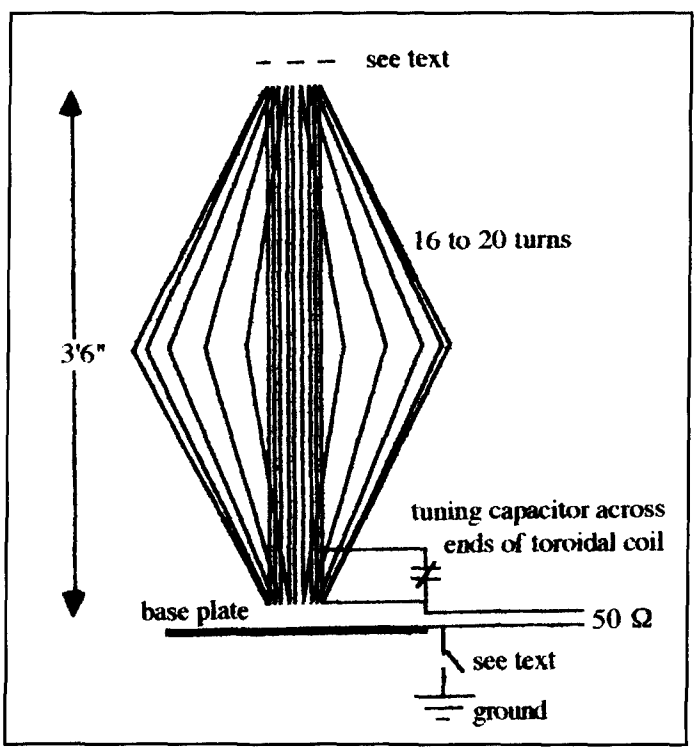


Figure 9. An experimental, elongated, "top-band" toroid in the shape of a fat cigar. This was used to investigate the decoupling of the ground from the near field.

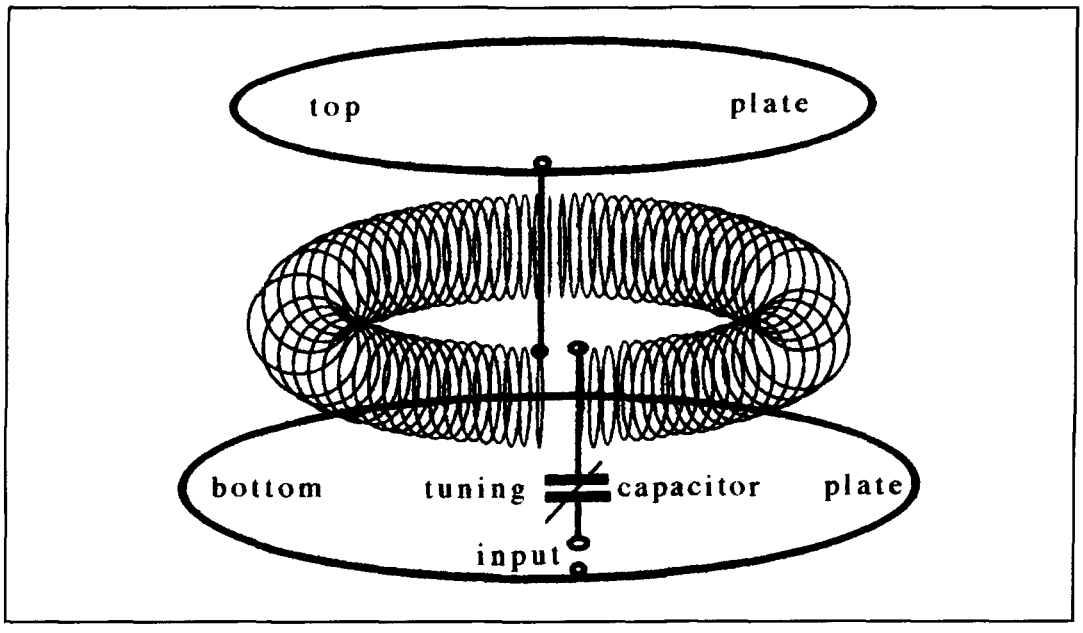


Figure 10. A compact toroidal antenna that proved to be quite successful on "top band" when fitted with a simple virtual earth over lossy ground.

to advantage on top band. Quarter-wavelength radial ground wires would defeat the object of an antenna that otherwise could be placed in any corner of a building; but any conductor in the vicinity, and the termination of that conductor, may have a considerable effect on the radiation efficiency. Loose coupling to the top slightly increases the radiation and detunes the system, but doesn't avoid the absorption in the ground. The real secret lies at the base of the toroid.

One of my old toroids is shaped like a fat cigar. It is about 3 feet 6 inches tall and the main winding is tuned by a vacuum capacitor to reduce the risk of flash-over. It was originally fed from a well-insulated two-turn primary winding that gave a good match to 50 ohms, but was not easy to adjust if one made variations to the system. I decided to try end-feeding it between one side of the tuned circuit and a metal plate that could be connected directly to the outside of the coaxial feed, as shown in **Figure 9**. This provided a simple and easily adjustable system, but the radiative property was dependent on where the connection to ground was made.

One option was to avoid grounding the outer of the coax at the antenna. Another was to induce a signal from the opposite end of the antenna onto a separate wire running close to the coaxial feeder, acting as sort of a stub. I tried the trick of loose coupling into the local field by connecting this separate wire to a very small aluminum plate about six inches above the coil, shown dashed in **Figure 9**. Although the wire ran horizontally, quite close to the feeder, against a wall and near the ground, tests showed that this relatively compact system, when properly matched, was almost as good as a 28-foot high half-wave resonant dipole. The gain was greatest with a wire running a few inches away from the coaxial cable and terminated to ground at the transmitter, but this could just be a function of the local layout and conditions. I now understand James Corum, K1AON, has used the technique of strongly exciting conducting elements from toroids; e.g., by immersing them through the toroid. No doubt he can provide further information from his experiences, although I do not think he has tried exciting the toroid from the space between the winding and a conducting plate or using the loose top coupling outlined in this paragraph. Incidentally, the loose top coupling trick can be tried on other designs that I have discussed. If the coupling is very loose it retains the main power house in the toroid, so the resonant frequency and the matching are only slightly affected, but more of the energy is converted to the radiation field. It also hardly changes the overall size of the system. If you are going to have a cable to connect to your antenna, you may as well have another wire running nearby. However, a better trick is probably the virtual earth that I shall describe shortly.

I decided to go back to basics. The top coupling method seemed to be a little complicated and the cigar-shaped toroid was rather clumsy. I decided that it might be more of a challenge to see how small one could make the main radiating element while still achieving efficient radiation. I therefore reverted to a single toroidal coil consisting of about 45 meters of

3.5 mm copper wire wound solenoidally with a lathe on a three inch mandrel and then bent round a rigid foam plastic core, nine inches in diameter, to form a torus having an outer diameter of 16 inches. The torus was sandwiched between two clover leaves, each formed from three 20-inch diameter aluminum discs, with rigid foam plastic spacers separating the coil from the plates. The upper and lower spacings were three inches and one inch, respectively. These spacings affect both the resonant frequency and the feed impedance, and the lower plate should be spread somewhat larger than the upper plate.

In its final configuration, the whole assembly measures 7-1/2 inches tall (about one thousandth of a wavelength) and 30 inches wide. The width may be reduced by winding more turns on the coil, but this will also affect the matching and the spacings must be adjusted. The dimensions that I have quoted are convenient for future mobile or marine operation, where a low profile may be required; although the main object of the exercise was to investigate means of improving the radiation efficiency in the absence of the relatively large conducting surface provided by an automobile or a large stretch of water.

As quarter-wavelength radial ground wires were out of the question, I resolved to try a virtual earth counterpoise technique, and with this I had some success. I used an arbitrary length of wire (about 25 feet) connected to the lower plate of the toroid assembly and terminated in a metal plate 2 feet by 3 feet. The counterpoise plate should not touch the ground, but may be insulated from it with a spacing of about one foot. It may be placed in any convenient position and the wire does not have to be straight. It is, however, necessary to experiment for the optimum performance at a particular site. Similarly the lower plate of the toroidal assembly should be kept a short distance off the ground in order to reduce the earth current. Mine is tucked away on a convenient shelf at a height of 7 feet. The antenna is fed at 50 ohms via a high-voltage capacitor to one end of the toroidal coil. The other end of the coax connects to the lower plate, as shown in **Figure 10**. The bandwidth of this antenna is approximately 15 Kc/s, but it will still function efficiently a long way off tune. Nevertheless, there is a simple way of tuning it remotely by varying the input capacitor. I use a 200-pF variable capacitor rotated by a very small 6-volt geared motor that may be driven either way with push buttons in the shack; the tuning range is from 2.25 to 1.5 MHz and the SWR never exceeds 1.5:1. You may have to experiment to emulate this performance in local conditions.

Conclusion

I recommend these antennas for mobile VHF systems where low garages and vandals may present a problem for whip antennas. I also commend them to anyone who is deprived of the space to erect an external antenna on any of the HF bands, but I do not claim that the HF versions are an equal substitute for a good external antenna. They are not difficult to make, but do not be disillusioned if they do not work straight away. They usually require some adjustment to tune them and to obtain optimum matching. For this purpose, an instrument such as the MFJ-259 is an absolute godsend. One can also use a fluorescent tube to assist rough tuning, but do not be misled into thinking that a brightly glowing tube necessarily means that

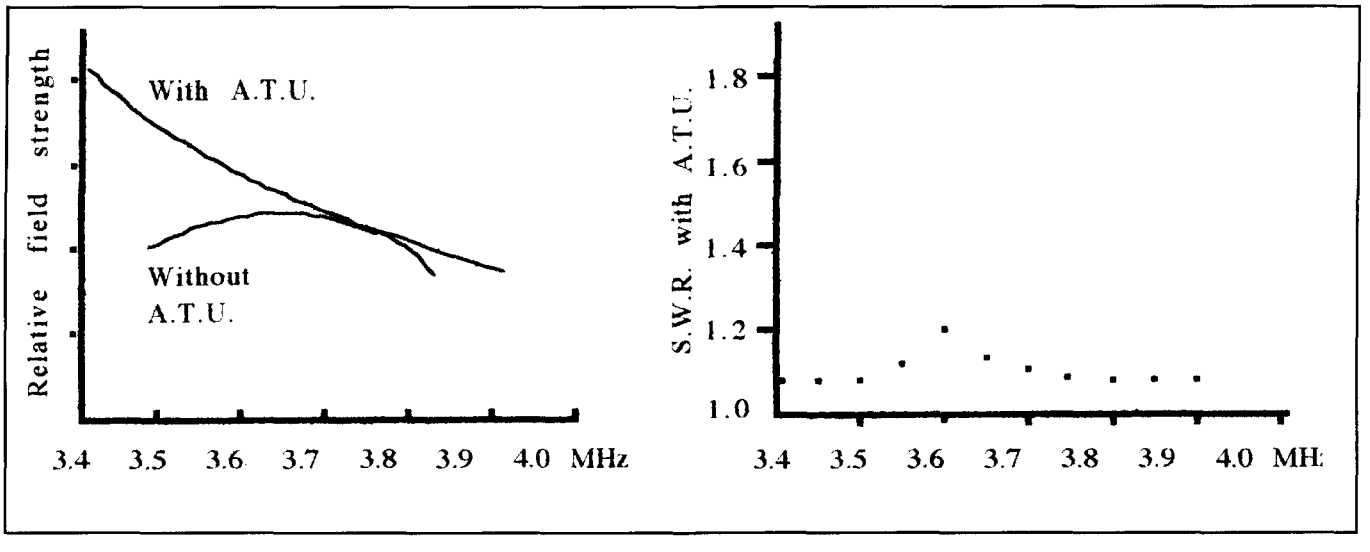


Figure 11. The relative field strength and standing wave ratio of a cascaded toroid on a flower bed a few feet from the transmitter.

you are radiating lots of power; the current driving the tube is from the local induction field.

Once you have one working, you will find that the 2-meter version will cover the whole band if centered on 145 MHz. On the HF bands, it often pays to set the resonance towards the high frequency end of the band and use an A.T.U. to match it at lower frequencies. This is because one only has to deliver a high current into the toroid; however, above resonance it starts to act as a choke, the radiation gets weaker, and it ultimately goes into other modes. Measurements on an 80-meter double toroid sitting on a flower bed just outside the shack window are shown in **Figure 11**. Note the increased field strength away from resonance at the lower frequencies.

For readers with an experimental turn of mind, there is still much research to be done on these systems and many variants and improvements are possible. I have not had the opportunity to try marine operation and this could be quite interesting. One final warning: beware of RF burns, RF heating, lossy dielectrics, and the risk of fire. The local fields are very intense. ■

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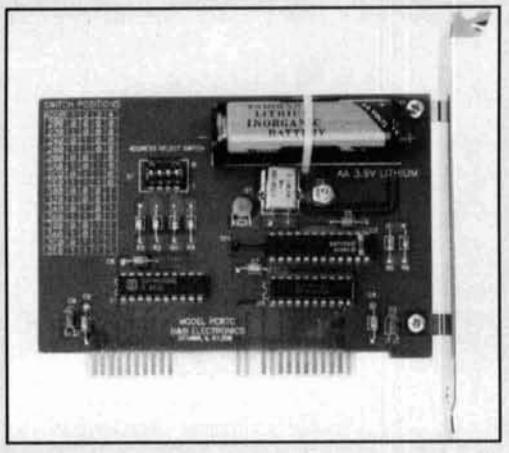
PRODUCT INFORMATION

B&B Electronics' Time Clock for PCs

B&B Electronics Manufacturing Co. has released a temperature stable, real time clock for PCs. The Model PCRTC replaces the clock found in most PCs.

Based on a highly stable oscillator circuit, the PCRTC's clock uses a software driver to update the DOS clock every minute. This allows the user and all applications to access the correct time without using any special commands or function calls. The 8-bit (short slot) PCRTC card, which does not require IRQ lines, reduces your computer's clock error to 15 seconds/month, or less than 5 seconds/month if operating in a temperature stable environment. When power is off a 3.6V lithium batter maintains the clock. Features include a battery status function that lets you replace a low battery before time is corrupted, and a register that helps you log and determine the time of last power loss or reset.



The PCRTC is priced at $129.95 and includes the software driver, a timesetting utility, and a simple command library on a 3.5" floppy disk. To receive more information about the PCRTC and B&B Electronics' Serial Communication and Control Equipment catalog contact: B&B Electronics Manufacturing Co., 707 Dayton Road, P.O. Box 1040, Ottawa, IL 61350; phone 815-434-0846; 24-hour fax 815-434-7094; 24 hour BBS 815-434-2927; Internet: catrqst@b&b-elec.com

1 Watt Power Chip Resistors Are Smallest Yet

Philips Components has announced the smallest chip resistors able to dissipate 1 watt. The PRC201 range is housed in a new 1218 package, having the same dimensions as standard 1812 devices, but with terminations on the longer side. The benefits of the design include improved strength and reduced stress of the solder joint, and more efficient heat dissipation.

The PRC201 surface mount power resistors have tolerances of five percent, and come in a wide range of values between 1Ω and 1MΩ. one percent tolerances are available upon request. The devices are supplied in blister tape, and are placed usingexisting surface-mounting equipment.

Price per unit is 8¢ for average quantities. For more information, contact Ron Cearns, Marketing Manager, Philips Components by calling 407-881-3257.

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ORBITAL ANALYSIS BY SLEIGHT OF HAND

A review of manual, intuitive satellite tracking techniques

Modern digital computers, along with low-cost, high-power software, have been indispensable tools in the quest to communicate via orbiting satellite. Tracking software has become so ubiquitous that a whole generation of enthusiasts is unaware that there once was another way to track the early OSCARs. While I'm not opposed to progress, I am of the opinion that the magical computer tends to separate its user from the basics of the underlying application. Through a review of manual, intuitive satellite tracking techniques, we can gain a more thorough understanding of the rather simple forces of nature underlying those mystical Keplerian Elements.

Introduction

There is a curious object on display in my classroom—a pancake-flat Plexiglas™ disk about a half meter in diameter. A North polar projection map of the earth appears on one side of the disk, and a South polar projection appears on the other. Sandwiching this pre-Columbian globe are two transparent overlays, pinned to the poles with a robust grommet so as to rotate freely. On the rotating panels, great arcing lines with cryptic notations are inscribed in grease pencil. The object looks very much its part—an icon to the high priest of satellite navigation.

You probably know this device as an Oscar-Locator, a graphical aid to satellite tracking. This granddaddy of all Oscar-Locators was in fact the original OSCAR 1 orbital plotting board, preserved throughout these countless generations of amateur radio satellites. We called it a Satelabe then—a word derived from Astrolabe—the navigational instrument that guided explorers across un-charted oceans from ancient times until the development of the sextant in the 18th Century. Even though my students would relegate it, along with my Osborne 1 computer and log-log-decitrig slide rule, to the musty, dusty shelves of a technology museum, I keep it for the most utilitarian of reasons—because it still works.

The early days

The microcomputer is today as indispensable a part of the world of satellite communications as the mini-HT is to amateur radio. Starting with Dr. Tom Clark's legendary *BASIC Orbits* program and continuing up through the present software array with its dazzling high-resolution graphics, we all have tools at our disposal, the likes of which NASA could only dream about during the days of Apollo. Yet in those eras of antiquity, BC (Before the Computer), a handful of dedicated visionaries managed to conceive, construct, and connive into orbit, the world's first non-Government sponsored artificial Earth satellite. They did so with tools that today would be considered laughably crude, but they did it. They left us AMSAT as their legacy. They left us their Satelabe, as a reminder.

I was not really one of the original OSCAR cadre, although I sat at their feet in awe. As a

high school student sitting quietly in the back of the room, watching my heroes concoct their minor miracle, I said to myself: "Some day, when I grow up, this is what I want to do." And someday, when I grow up, I will.*

In the meantime, I've been privileged to teach satellite communications to a whole generation of technologists. Only their way, it's hard to tell where the computer ends and the technician begins. Armed with their silicon ephemera, they manipulate Keplerian Elements to the twelfth significant figure, produce orbital predictions in double precision, and haven't the foggiest idea what they all mean.

An old tool

That's where the Satelabe comes in. By using this old tool, and applying mathematical concepts no more advanced than those available to the ancients, one can visualize the balance of forces that hold a satellite in its orbit. With purely manual techniques, one can perform orbital analysis to far less accuracy, but with far greater clarity than can be accomplished by digital computer.

That's also what this article is all about. In it, I'll review a few of the basics of the satellite orbit, armed with nothing but a pencil and a pocket calculator (a few years ago, I would have said slide rule). Although I doubt I'll ever wean you from the MegaTrak 1000 program running on your 100-MHz 80686 with the gigabyte hard drive, I hope to remind those of you who might have forgotten, what's really going on behind the zillion-pixel screen, back in the land where the ones and zeros cavort in wild abandon.

Circles and ellipses: an eccentric view

A good place to start our description of the orbits of communications satellites is with a review of the most historically significant controversy in astronomy—the nature of our solar system. Probably the first to propose a heliocentric theory of the universe was the Greek astronomer Aristarchus of Samos (circa 310-230 BCE). His view received little attention from the ancients, who favored a geocentric scheme, as later formalized by the Greco-Egyptian mathematician Claudius Ptolemaeus in the 2nd century CE. In Ptolemy's system, simple circular motion was used to describe the motion of all visible celestial bodies around the earth. The apparent retrograde motion of the planets was explained by a rather complicated system of epicycles on deferents.** Still, predictions of celestial events based upon the Ptolemaic system were crude at best.

The great Polish astronomer Nicholas Copernicus may well have been trying to smooth out some of the inconsistencies in the Ptolemaic system's prediction of eclipses. He was probably the first astronomer since Aristarchus to propose anew a heliocentric solar system (*De Revlutionibus Orbius Coelestium*, 1543). Although he placed the Sun in its rightful place, his system still retained a variant on epicycles to explain retrograde motion. Circular orbits, it seemed, were tricky things.

Tycho Brahe was a renowned fence sitter. Unable to decide the relative merits of the Ptolemic and Copernican systems for himself, he created one of his own, incorporating elements of each. This Danish astronomer saw an immobile earth around which the sun revolved, with the other five known planets then revolving around the Sun. To support his model, he recorded a lifetime of observations of the planets, the Moon, and Supernova 1572. His excellent observations, published posthumously by Kepler (in *Tabulae Rudolphinae*, 1627) led to the breakthrough that eliminated the inconsistencies in all earlier solar system models. Indirectly, it also gave us our most powerful tools for satellite orbital analysis—the Keplerian Elements.

Professor of mathematics at Graz, Johannes Kepler proposed in 1609 (in *Astronomia Nova*) that planetary motion could be described not by circles, but by ellipses. He showed that any planet (and by extrapolation, satellite) must orbit in an ellipse with its primary at one focus. His laws of motion further described the change of orbital velocity required throughout a stable elliptical orbit, and explained (*Harmonice Mundi*, 1619) the interrelationship between the size of the ellipse and the orbital period. Kepler's elliptical orbits made possible a full understanding of not only the heliocentric solar system, but three and a half centuries later, the orbit of your favorite OSCAR.

It is important to note here that a circle is simply a special case of an ellipse, one whose eccentricity equals zero.** Since our earliest artificial earth satellites (Sputnik I, OSCAR I, etc.) all had low eccentricities, we tended to analyze their orbits as circular. That's where the early OscarLocator came in. It turns out that the motion (relative to an observer on the ground) of

*Okay, I exaggerate my modesty slightly. I have had a minor role in all of this, rising through the ranks of Project OSCAR volunteers to eventually become Technical Director, and then Chairman of the Board, a position I held until 1990.

**To visualize this rather convoluted scheme, think of the popular carnival ride called the Tilt-a-Whirl. Each rider (planet) sits on a cart. The cart sits on a platform that describes a circular orbit (deferent), while whipping around in its own circular sub-orbit (epicycle).

**Eccentricity is a measure of the "lopsidedness" of any ellipse, on a range of 0 to 1. Values approaching 0 represent a round, and those approaching 1 represent a flat shape.

a satellite in circular orbit can be easily described as an arc on a polar projection map. The Satelabe is such an arc on a map. With it, we can not only extrapolate any orbit into the future for prediction of Acquisition of Signal (AOS) and Loss of Signal (LOS) at a ground location, but we can also project required antenna bearings (azimuth and elevation) for a given pass, and predict mutual satellite visibility for two ground stations. Useful, no?

More complicated orbits (such as the highly eccentric Phase III and Molniya satellites) defy such simple graphical analysis because their ground tracks tend to corkscrew, and do not necessarily repeat over time. It is for such orbits that tracking software really shines. However, we still have a good number of useful satellites in nearly circular orbits (PACSATS, MicroSats, MIR, space shuttles, various weather satellites, and of course the most useful Clarke, or geosynchronous, orbit). All but the latter (which stands motionless in the sky, from our terrestrial vantage point) can benefit from Satelabe analysis.

The ideal orbit

We will begin our analysis of satellite orbits by making quite a few simplifying assumptions. For starters, let's consider an artificial satellite in perfectly circular orbit (that is, eccentricity of zero) around a perfectly spherical earth of uniform density, with no atmosphere. Of course, our planet is both lumpy and oblate (wider at the equator than at the poles, by about 0.1 percent or so); however, ignoring these details simplifies orbital analysis. We will correct for reality later. Similarly, let's turn an n-body problem (one which includes the gravitational effects of the moon, sun, planets, and stars) into a much simpler 2-body problem, by pretending that nothing exists but our satellite orbiting such a perfect planet.

For this simple and ideal case, only two influences determine the motion of the satellite around its primary: the force of gravity (pulling the satellite toward Earth), and the pseudo-force of the inertia (pulling the satellite away from Earth). Further, for a stable orbit (one that does not change over time), these two forces must be balanced, in exact equilibrium. Because we can define both gravity and inertia mathematically, and set the two equal to each other, we produce what I call the Basic Orbital Equation. I'll spare you the derivation (it appears in the literature) and cut to the result:

$$\frac{mV^2}{r} = \frac{GMm}{r^2} \qquad (1)$$

In **Equation 1**, little **m** stands for the satellite's mass, big **M** for the mass of the earth, **V** for the satellite's velocity, **r** for the radius of its orbit, and **G** for Newton's universal gravitational constant. The left-hand expression relates to inertia, and the right-hand side to gravity. As I said, a stable orbit requires that the two be equal. What's interesting about this equation is that it can be simplified. The **r** in the left-hand denominator can cancel one of the **r**'s in the right. The **m**'s in the two numerators can cancel. This leaves us with:

$$V^2 = \frac{GM}{r} \qquad (2)$$

which is not only simpler, but allows us to draw some interesting conclusions. For one thing, you'll see that **Equation 2** makes no reference at all to the mass of the satellite. The orbital behavior of any satellite appears *independent* of its mass!* Periodically, my students try to sell me on the notion that the mass of the satellite is absent from **Equation 2** because it's negligible relative to that of the earth, thus can be ignored. Negligible though it may be, this is not the reason. You can see from the above that, when equating inertia to gravity, the satellite's mass *cancels*.

The next interesting thing we can learn from this equation is that orbital radius varies inversely with the *square* of velocity. So you see that as you move a satellite closer to earth, it moves faster. And (because the relationship is a square), if you place it really close to earth, it moves *really* fast. This is consistent with our observations. The space shuttle, only a couple of hundred kilometers up, zips right along, orbiting our planet in about an hour and a half. Our natural satellite, the Moon, is closer to 400,000 km up. It meanders across the sky, taking a whole moonth—er, month—to orbit. Also, at an intermediate distance of 36,000 km, satellites in the so-called Clarke** or geosynchronous orbit move at moderate speeds, orbiting the Earth in exactly 24 hours.

Notice now what appears in the numerator at the right of **Equation 2**. Contrary to popular belief, the **GM** product is not a Chevy. Rather, it is a constant for all satellites orbiting the earth. If we know this value (and we do: 4 x 10^{14} m^2/s^3), we can easily calculate the velocity for a satellite orbiting at any radius, or the orbital radial that would correspond to any velocity. Try this, for example: What is the

*The *launch* of a satellite into its given orbit, on the other hand, is highly mass dependent. The more massive the satellite, the more thrust is necessary to insert it into its orbit. You have to kick the football hard, so it will not only sail *high* enough to clear the goal post, but will also be moving *fast* enough not to fall short of the goal line. That is why, for example, the launch for a heavy Phase III D is so much more costly than that for a MicroSat. Did you ever wonder how we kick a Phase III football into its intended orbit? Why with an Apogee Kick Motor, of course!

**Named for Arthur Charles Clarke, communications engineer and science fiction author who first proposed this orbit in a 1945 *Wireless World* article.

fastest possible velocity for a satellite orbiting our ideal earth?

Well, we know from **Equation 2** that the lower the orbit, the faster the velocity. So, what is the lowest possible orbital radius? If we ignore atmospheric drag, trees, mountains, and tall buildings, and rule out subterranean orbits (the tunnel isn't finished yet), that would be the radius of the Earth—about 6370 km. Plugging this **r**, and the **GM** product, into **Equation 2**, we come up with a speed of about 7900 meters per second, or 17,700 mph. So now we know the top speed for your satellite, or my motorcycle.

But what about Evel Knievel's motorcycle? Rumor has it his rocket-powered monster can exceed the speed we just computed! What will happen to him if it does? Maximum orbital velocity is also called escape velocity. Anything that exceeds this speed will break free of earth's gravity (or more properly, will find inertia exceeding the force of gravity). So, in excess of 17,700 mph, Mr. Knievel will find himself flying free of Earth's influence, departing into outer space. (Some say he's already there.) In fact, any interplanetary spacecraft must be accelerated to beyond escape velocity.

Our next task is to determine orbital period—the time required for our satellite to circumnavigate the Earth exactly once. If we know orbital radius, we also know circumference, or the distance the satellite travels in one orbit. Because a circle contains 2π radians, the distance traveled in one orbit equals $\mathbf{2\pi R}$. At the velocity calculated with **Equation 2**, orbital period is: $\mathbf{t = d/v}$, so:

$$t = 2\pi (r^3 / GM)^{1/2} \qquad \textbf{(3)}$$

We can now do what Clarke did 50 years ago—compute the required orbital altitude to achieve geosynchronicity. This requires us to rearrange **Equation 3** to solve for **r**—an operation that is a bit awkward, but not an algebraic impossibility. Let's see now, if I did this right:

$$r = (GM\, t^2 / 4\pi^2)^{1/3} \qquad \textbf{(4)}$$

Plugging in the orbital period required to synchronize with the Earth's rotation (24 hours, or 86,400 seconds), and our old friend the Chevy—er, **GM** product, the above gives us an orbital radius of about 42,290 km. Subtracting the Earth's 6370 km radius leaves us with a satellite 35,920 km above our planet. Within the constraint of our simplifying assumptions and round-off errors, that's exactly where Clarke said it should be.

Orbital increment defined

Let's return for a minute to this business of a spinning Earth. The Earth's rotation means that if a satellite orbits our planet exactly once, it will *not* necessarily come back to rest above the same point on Earth from which it started. Consider, for example, a polar orbit with a two-hour period. Say our satellite crosses the equator northbound (at a right angle), zips over the North Pole, crosses the equator southbound (again at a right angle), slips under the South Pole, and then returns northbound to the equator again. While all this has been happening, the earth has been spinning eastward, about 1000 mph worth at the equator. Now, if we note the point on the Earth over which the first northbound equator crossing took place (we call this point the orbit's *ascending node*), and then note where the next ascending node occurs, the second ascending node point is going to be about 2000 miles west of the first.

Incidentally, the vector sum of those two circular motions (the satellite's orbit and the Earth's rotation) describes a sinusoid. This is the source of those sine-wavy ground track lines on the flat map at Mission Control, which you've seen on the TV news. Now, the distance along the equator between two successive ascending notes is called *orbital increment*, or westward progression, and is measured, not in miles, but in degrees of longitude on that same flat map in Houston. Since our planet spins 360 degrees every 24 hours (more or less), the earth has spun about two twenty-fourths of 360 degrees, or about 30 degrees, during the course of a two-hour orbit. This means the two successive ascending nodes are 30 degrees apart; the orbital increment is 30 degrees.

But wait, there's an easier way. If we continue to consider our ideal, two-body problem, ignore the motion of the earth and its retinue of satellites about the sun, and accept our simplified view of a perfectly spherical earth of uniform density, then orbital increment is simple to estimate. Increment (in degrees) will equal exactly the orbital period (in minutes) divided by four! This can be readily proven by dimensional analysis, but trust me.

With the Satelabe, we use orbital increment to plot successive orbits. Say we know the equatorial crossing longitude in the ascending node for a given orbit. We lay the edge of our OscarLocator cursor over that longitude on the plotter, and the curved cursor shows the path the spacecraft will take over the Earth for the succeeding half orbit. To forecast the next orbit, we estimate increment by dividing period by four, add increment to the previous crossing longitude, move the cursor, and start again. An example: During a recent space shuttle mission, the orbital period was exactly 1 hour, 30.4 minutes. That's 90.4 minutes, which, divided by four, gives us an orbital increment of 22.6

degrees. On one orbit of that mission, the shuttle crossed the equator northbound at a longitude of 72.6 degrees west, at exactly 2145:30 Zulu. At exactly 2315:54 Zulu, 90.4 minutes later, the shuttle would cross the equator northbound again, at a latitude of 72.6 degrees + 22.6 degrees of increment, which equals about 95.2 degrees west.

How far up was that space shuttle? An orbital period of 90.4 minutes equates to 5424 seconds. From **Equation 4**, the orbital radius equals 6680 km. Subtracting the Earth's 6370 km radius, we see that the shuttle is only 310 km up. No wonder the overhead SAREX signals are so strong!

We still occasionally find equatorial crossing times for various ham satellites published in the AMSAT literature. If you know what ascending node value will bring the satellite overhead at your location, you can extrapolate from any given crossing time using the above method to estimate your next AOS.

And now, reality intrudes

The above computations work well for our ideal, circular orbit; but what happens when we consider the more general case of Kepler's famous ellipse? Even in elliptical orbit, gravity and inertia must always be in equilibrium. Because the distance between the satellite and its primary varies along the elliptical path, the force of gravity is ever changing. This requires a like change in inertia throughout the orbit, which is only possible if the satellite *speeds up and slows down.*

In fact, the orbital velocity of a satellite in elliptical orbit does indeed vary, from maximum at *perigee* (the point of the orbit that brings the satellite closest to the earth) to minimum at *apogee* (the point of farthest separation between satellite and earth). We can readily compute *apogee radius* as the sum of apogee height plus the Earth's radius. We can insert this value into **Equation 2** to determine the satellite's velocity at apogee. In a similar fashion, we can use *perigee radius* to compute the satellite's velocity at perigee. The mean orbital velocity will be somewhere between those two values, although we would need to apply some calculus to determine an exact value.

Since the distance between earth and elliptical satellite is ever changing, we can't directly apply **Equation 3** to determine orbital period. Also, although calculus gives us an exact solution, here's a simple first-order approximation. Compute period per **Equation 3** for a circular orbit with radius equal to your satellite's apogee. Do the same for perigee. Your satellite's actual orbital period is roughly between these two values. For example, a satellite in *synchronous transfer orbit* has its apogee at Clarke altitudes (the resulting period, for a circular orbit, would be 24 hours), and its perigee at space shuttle altitudes (corresponding period an hour and a half). The midpoint between these two values is just under 13 hours, which comes close to transfer orbit period.

Remember our myth of a uniform spherical earth? I guess you know by now that it simply isn't true. After all, our planet is a spinning body. Four billion years of spin have made the earth oblate—wider at the equator than across the poles. This happens to people in middle age. The more we spin, the wider we get at the equator. We're not obese, just oblate.

The same is true of our neighboring planets, the sun, and the stars in general. The inertia of a spinning body slings some of its matter outward, and makes it bulge. Even if we start with a perfectly circular orbit, this satellite bulge will cause the force of gravity to vary as the satellite circles the Earth. Because inertia and gravity must balance, the satellite will speed up and slow down, making its orbit wobble. We call this orbital wobble *oblateness precession*, and we can quantify it, although you don't really want to see the equation.

Well, okay, if you insist:

$$\Theta = 9.964 * [(R_e + h) / R_e]^{(-7/2)} * [1 - e^2]^{-2} \quad \text{Cos } i \qquad \textbf{(5)}$$

See what I mean? The important thing at this point is not to compute oblateness precession, but to recognize its effect—which is to make the satellite slip a little to the East with every orbit. So there goes our nice, simple estimate of orbital increment, right out the window. This explains, in part, the cumulative error in extrapolating equator crossings from a satellite's orbital period. In the short term, the estimates are acceptable for communications. However over days or weeks, a new ascending node observation becomes necessary in order to obtain acceptable OscarLocator results.

The next simplification to disprove is the notion of a stationary earth. Remember that our planet is moving around the sun at a rate of (360 degrees/year)/(365.242 days/year) = 0.985673 degrees/Day. Consequently, if we performed a simplified orbital analysis based upon the assumption of a stationary earth, the satellite would accumulate a westward error of almost a degree per day, when measured with respect to the Earth's surface.

For communications purposes, we *must* describe the satellite's orbit with respect to the earth's surface—that's where we are! Unfortunately, at almost a degree per day, it doesn't take many days for our accumulated OscarLocator error to become substantial. This

is why, in practice, we try to obtain a new equatorial crossing point every couple of days. The Satelabe user gets these from computer-generated orbital prediction tables, which take into account the *sidereal precession.*

Did you happen to notice that the errors from sidereal and oblateness precession accumulate in opposite directions, one toward the East and the other toward the West? This means there is a possibility that we can design an orbit where the two effects will cancel, and we can achieve a reasonable approximation of our ideal orbit. Such an orbit does indeed exist, and it's called *heliosynchronous*, or sun-synchronous.

Satellites in sun-synchronous orbit will trace a ground track that repeats from day to day. This is most useful for earth resource assessment, and is used by NASA for many of its environmental and weather satellites. However for communications satellites, the sun-synchronous orbit has a number of advantages beyond simplicity of orbital calculation. One is that the spacecraft can be placed in perpetual sunlight—a useful feature for systems that derive their electrical power from the sun. Another is that a satellite in a properly designed sun-synchronous orbit can, over time, provide communications access to all points on the surface of the Earth. This is handy in packet store-and-forward applications. AMSAT's early Phase II satellites were all in heliosynchronous orbit, as are some of the newer MicroSats.

The final orbital simplification, the two-body assumption, is the most difficult to dispel. Our solar system has been described as being composed of the Sun, Jupiter, and assorted debris. All of it, even the debris, tugs on our satellite. I know of no simple algorithm for dispensing with the gravitational effects on an Earth satellite's orbit of the Sun, Moon, and planets. The computations are so complex as to defy manual solution. Here, then, is an area where computerized orbital analysis is not only justified, but really comes into its own.

Describing the orbit: size and shape

We can't use a computer to analyze an orbit unless we have a means for describing that orbit mathematically. The standard description of orbits used in the current generation of AMSAT tracking programs is the *modified Keplerian element set*, a collection of numbers that allows us to extrapolate the satellite's motion over time. The balance of this article is devoted to describing the Keplerian elements in conceptual, rather than computational, terms.

We've already seen how we can define the "roundness" of an orbit by its *eccentricity*, abbreviated **e**, on a scale of zero to one. Eccentricity, the first of our Keplerian elements, actually describes the shape of an ellipse quite completely. For current ham satellites, orbital eccentricities range from a low of approximately 0.00076 (UOSAT OSCAR 22) to a high of 0.72 (AMSAT OSCAR 13). That's a range of three orders of magnitude, and encompasses orbits from almost perfectly circular to very highly elongated.

Of course, the number of different orbits that could share the same eccentricity is infinite. Once having described the *shape* of our orbit, we next need to quantify its *size*. This can be accomplished in several different ways. The size of a given ellipse can be described by measuring the distance across it in the longest direction (Major Axis) or in the shortest direction (Minor Axis). These two axes intersect at right angles, midway between the two foci of the ellipse. We sometimes define orbital size in terms of *half* these values (Semi-Major Axis or Semi-Minor Axis). Similarly, apogee and perigee radii (the sum of which equals Major Axis) will define the size of the ellipse. We could also use orbital distance, the "circumference" of the ellipse.

Another way to define the orbit's size is in terms not of dimensions, but rather time. Here we would invoke the elliptical equivalent of **Equation 3**. Because the larger the ellipse, the longer the orbital period, if we know the period and eccentricity, we can determine all the critical dimensions of the ellipse. This is *almost* how we define orbital size when using satellite tracking software, but not quite.

Consider that a sine wave can be described either by its *frequency* (in cycles per second) or, alternatively, by its *period* (in seconds per cycle). The latter is what we measure on an oscilloscope and is readily converted to the former, because the two are reciprocals. In like fashion, the reciprocal of orbital period is orbital frequency, or mean motion. As a Keplerian Element, mean motion is abbreviated **N**, and is typically measured in orbits per day. Our current crop of satellites have mean motions ranging from 2.059 orbits per day (AMSAT OSCAR 10) up to 14.69 orbits per day (UOSAT OSCAR 11). The higher the mean motion, the shorter the orbital period, and the smaller the orbit.

Eccentricity **e** and mean motion **N** adequately define the size and shape of an orbit. It now remains for us to define the orientation of the orbit in space, and the location of the satellite within the orbit for any given point in time.

Describing the orbit: roll, pitch, and yaw

Just as aircraft require three angles (roll, pitch,

and yaw) to describe their orientation with respect to the Earth, so can we locate our elliptical orbit in three-dimensional space by developing three appropriate Keplerian Elements. The "roll" parameter refers to the angle between the satellite's orbit and the Earth's equator, measured at the ascending node. This angle is called inclination, is measured in degrees, and is abbreviated ***i***. An inclination of zero degrees means the satellite's orbit always remains over the equator, with the satellite moving eastward. Clarke orbits are an example of 0 degrees inclination. Another possible equatorial orbit, with the satellite moving westward, would have a 180 degree inclination. A polar orbit has an inclination of exactly 90 degrees.

Any inclination between 0 and 90 degrees defines a *prograde orbit*. This means that the satellite's horizontal motion is in the same direction as the Earth's rotation. Any inclination between 90 and 180 degrees makes for a *retrograde orbit*, with the satellite's lateral motion opposite the direction of the Earth's rotation. Prograde orbits are easier to achieve, because the eastward rotation of the Earth gives us some free thrust at launch time. In fact, the closer to the equator we move our launch site, the more of this free thrust is available. This is why the European Space Agency maintains its launch facility at Kouru, Guiana, practically on the equator.*

A retrograde orbit, with an inclination slightly above 90 degrees, is required to achieve sun-synchronicity. This is because ***Cos i*** in **Equation 5** needs to be negative to give oblateness precession its proper direction to overcome sidereal precession. As a result, launch to heliosynchronous orbit is most readily achieved from extreme northern or southern launch sites, where there is less of the Earth's eastward rotational velocity to overcome.

Currently, our most nearly polar prograde satellites, AMSAT OSCAR 21 and Radio Sputnik 12/13, have inclinations on the order of 83 degrees. Our most nearly polar retrograde satellites, OSCARs 14, 16, 17, 18, 19, 20, 22, 25, 26, and 27, all have inclinations on the order of 98 degrees. (See, I told you heliosynchronous orbits are popular.) Our most equatorial ham satellite OSCAR 10, currently has about a 27-degree inclination.

For our "pitch" parameter, we locate the orbit's perigee point in degrees with respect to the earth's equator. Consider that perigee of an orbit, like the nose of aircraft, can be oriented up or down. An *Argument of Perigee* (abbreviated **w**) of 0 to 180 degrees indicates perigee in the Southern Hemisphere, with **w** = 90 degrees meaning perigee occurred over the South Pole. Values of **w** between 180 and 360 degrees mean that perigee occurred in the Northern Hemisphere, with **w** = 270 degrees placing perigee over the North Pole.

Our reference for Argument of Perigee was the Earth's equator. However in order to define "yaw," we need an external celestial longitudinal reference. The First Point of Aries, a fixed point in space by which we define the beginning of spring, is such a reference. If we draw a line from this reference point to the center of the Earth, and another line from the center of the Earth to the orbit's ascending node (northbound equator crossing), then the angle between these lines is our third orbital attitude parameter. This Keplerian element is called *Right Ascension of the Ascending Node* (RAAN), and is generally abbreviated Ω. Of course, since our planet is both spinning and orbiting the sun, Ω is not stable, but varies over time.

The Keplerian Elements Ω, **w**, and i give us three degrees of freedom, by which we can completely describe the orientation or our elliptical orbit in three-dimensional space. This leaves us with but one problem yet to solve: Just where within that orbit is our satellite at any given moment?

Describing the orbit: location, location, and location

Before we can say where a satellite is *now*, we have to know where it was *then*. But when is *then*? We need to make an observation and clearly document exactly when that observation was made. Epoch Time, **T**, is the exact time at which a positional reading was taken. It is normally entered into tracking software as year, day of year, and fraction of a day since 0000:00 hours. Looking up **T** for an OSCAR in a recent orbital elements printout, I see a value of: 95159.69677441. This is interpreted as Year = 1995, Day = 159 (that's June 9th), and Time = (0.69677441 * 24 hours), which comes out to about 16 hours, 43 minutes, 21.3 seconds. Don't worry, the tracking software will do the conversion for you.

So much for the *when*. Now how about the *where*? At Epoch **T**, we locate the satellite in its orbit. We use perigee as our reference point. If we divide the orbit into 360 increments that are equal by *time* (not equally spaced in position), we can measure how long since perigee, on a scale of 0 to 359. It's important to note that, even though we measure this parameter on a scale of 0 to 359 (and even call the unit "degrees"), this is *not* an angular measure of the satellite's position. Rather, mean anomaly (abbreviated **M**) tells us what time fraction of an

*As the old Greek philosopher Will Rogers said, "There ain't no free launch . . . but a little extra thrust reduces the cost."

orbit has occurred since the last perigee, as of Epoch **T**. Since the other Keplerian elements give us (indirectly) orbital period, we can then compute exactly where the satellite was in space at **T**, and where it will be at any future time.

Finally, we have assumed our satellites to be traveling in a vacuum, but as you know, we live at the bottom of a gaseous ocean. This stuff (mostly nitrogen, a little oxygen, and various trace elements), which we call air, stays pretty much near the surface—but some of it does extend into space. To compensate for atmospheric drag, we often plug a figure called decay rate (n dot) into our software. This tells by how many revolutions per day our mean motion will actually change. It's usually pretty close to zero.

Conclusions

For ease of operation, there's nothing to beat a good GUI-driven, high-resolution, math co-processor supported satellite tracking program. After all, communicating with an orbiting satellite can be equated to downing the soaring eagle with a hurled ping pong ball, while blindfolded. The computer can at the very least help to strip the blindfold away. And, yes, I admit that my lab on campus boasts a fairly capable PC, on which my students crunch ephemera using a number of AMSAT tracking programs.

However on the wall above the computer, hangs that original OSCAR I Satelabe of which I spoke earlier. It occupies a place of honor, as it were, an anachronism of inestimable value. The humble Satelabe not only helps the students (and their Professor, as well) visualize the intricacies of the elliptical orbit; but as we reach for the stars, it also serves as a reminder of where we started, how far we have come . . . and just how much further the journey may yet carry us. ■

PRODUCT INFORMATION

Multipurpose CAD Program for the IBM PC

CIRCAD Version 3.6 is a complete schematic, PCB editor, multilayer autorouter to Gerber film design system. It runs on any IBM PC (or clone) 286+ machine with 640K of RAM and EGA, VGA display. A mouse (3 button is preferred). PC/MSDOS 3.3+ or DRDOS 5.0+ required. Program features file compatibility with OrCAD, Tango, Protel, AutoCAD (DXF), Gerber and others. It supports HPGL, Houston Instruments plotters and all PCL laser printers as well as color inkjets and 24 pin Epson dot matrix printers. Written entirely in Assembler language for speed. Hypertext help system includes user manual on-line. With CIRCAD you can work on schematics and PCB files simultaneously. Real preemptive multi-tasking enables you to send one file to the plotter (or laser) while editing another file. Does NOT require Windows or OS2 to run.

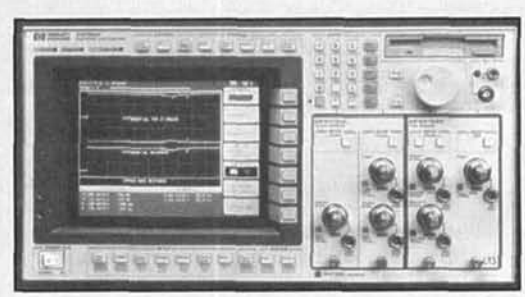
Special student and ham price. By ordering the program through Dover research, students and ham radio operators can purchase the full CIRCAD package for $299 (US); regularly priced at $995. (Full technical support is NOT provided for this special price.) For more information, call Dover Research Corporation at 612-492-3913.

HP Unveils Highest-Bandwidth Two-Channel Scope Module

Hewlett-Packard Co. announced that it has pumped up the bandwidth for its HP 54750A modular oscilloscope two-and-a-half times to 50 GHz. Fifty GHz is the highest bandwidth for a two-channel scope module available on the market. HP has also added TDR (time-domain reflectometry) capabilities that enable designers of high-speed circuits to match impedance and prevent faulty timing and clocking. These new capabilities are available as plug-in modules.

The dual-channel HP 54752A and single-channel HP 54752B 50-GHz plug-in modules for the oscilloscope enable users to capture and measure fast signals from radar, radio frequency and microwave applications as well as 10-gigabyte telecommunications signals. In addition, the HP 54753A TDR and HP 54754A differential TDR plug-in modules allow designers of fast circuits, printed circuit boards and computer backplanes to view the impedance characteristics and perturbations of their target systems, including the impedance of stripline traces, semiconductor packages, cables and connectors.

These plug-in modules and the oscilloscope mainframe are available now with delivery expected about four weeks from receipt of order. For more information, contact Hewlett-Packard Co., Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059; or call 1-800-452-4844 ext. 9274.



Axel Stark
Head of the
Antenna Development Group
Rohde & Schwarz, Germany

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PROPAGATION OF ELECTROMAGNETIC WAVES

A refresher course on wave propagation and the implications for the application of antennas

On hearing the words "electromagnetic-wave propagation," many will automatically think of research institutes or universities. Perhaps they will even have thoughts of science for its own sake. However, as can clearly be seen from the block diagram in **Figure 1**, which incidentally is valid for all radiocommunication, electromagnetic-wave propagation has considerable practical importance. It is impossible to say anything about radiocommunication as an entity without including the propagation medium in one's calcula-

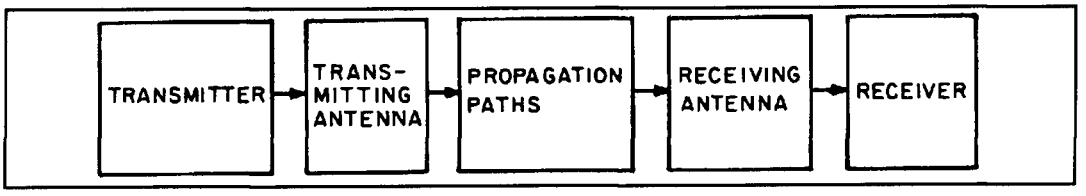

Figure 1. Block diagram of radiocommunication link.

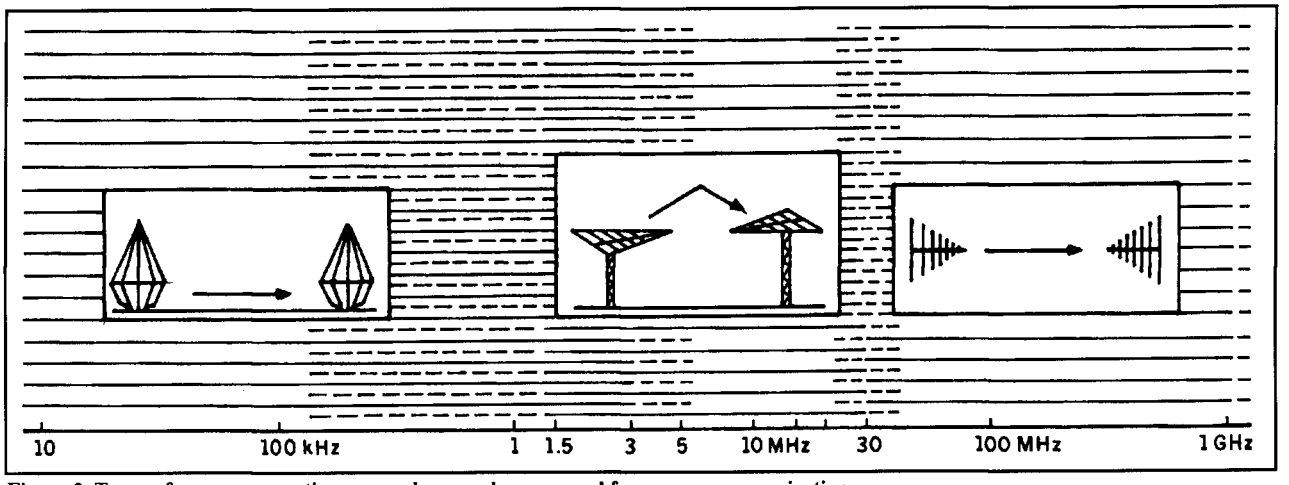

Figure 2. Types of wave propagation: ground-wave, skywave, and free-space communication.

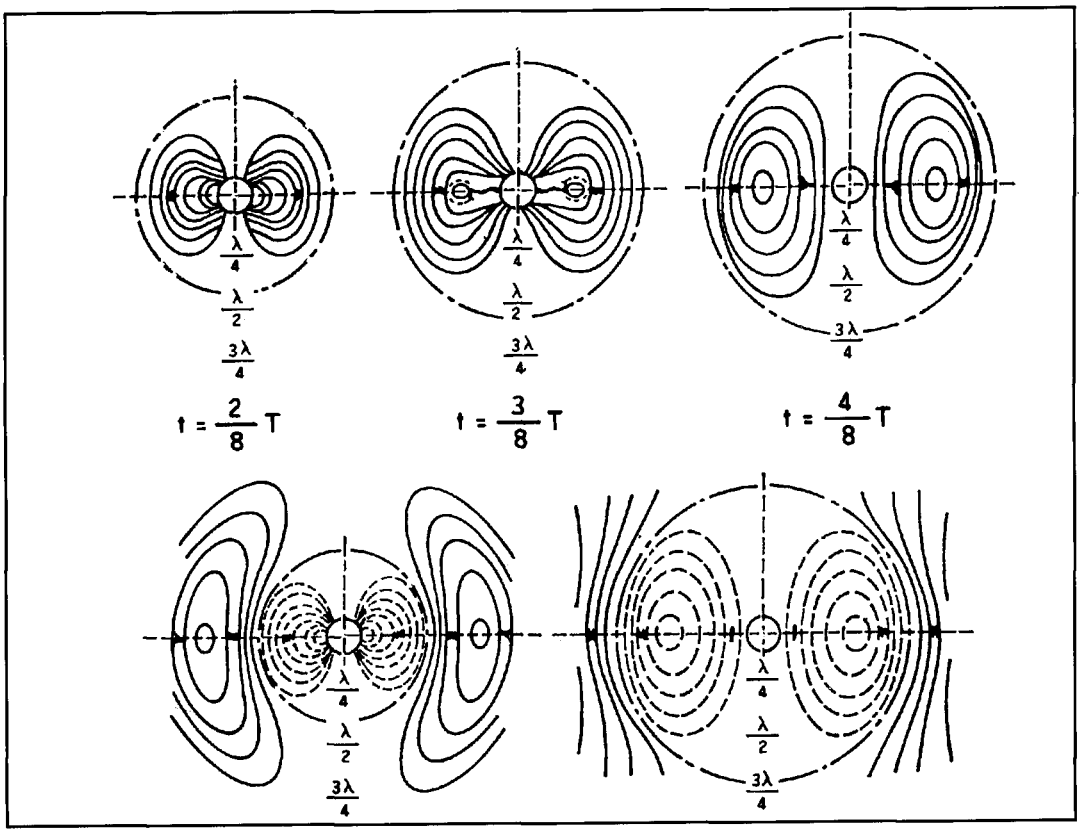

Figure 3. Radiation field of dipole.

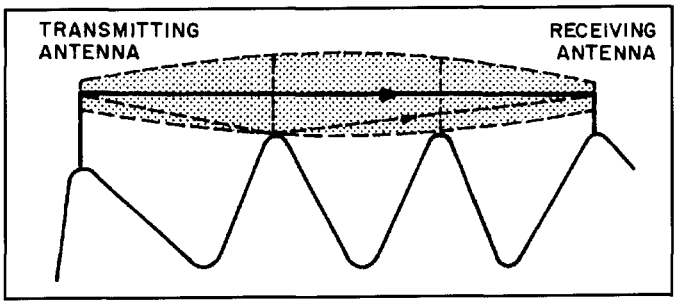

Figure 4. First Fresnel ellipsoid.

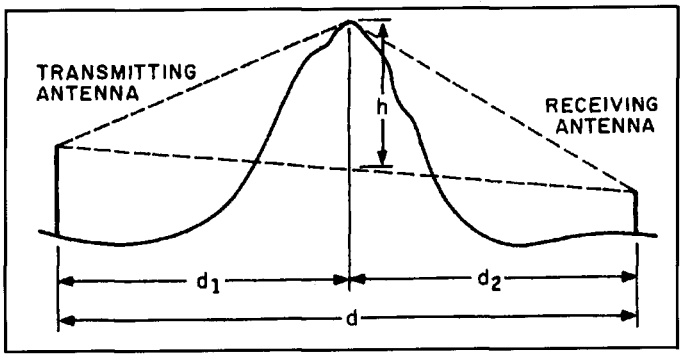

Figure 5. Obstacle between transmitter and receiver.

tions as the link between the transmitting and receiving ends. Only then is it possible to undertake proper optimization between communication reliability and the cost of the radio equipment. This is true for both a service radio link and the parasitic reception used in radiolocation.

Inclusion of the transmission path in calculations when designing a system has, until now, often been standard practice only in the planning of radio and television broadcast coverage, elaborate systems, and those bordering on the horizons of technology. This is mainly due to the peculiarities of the interface between the antenna and propagating medium. High-frequency interfaces between items of equipment can often be described adequately by the characteristic impedance and the permissible VSWR. With a transmitter, the power to be transmitted must also be taken into consideration. At the receiving end, additional parameters include the noise figure and the gain.

The interface between the antenna and propagation medium must be described by completely different parameters. Radiation pattern and antenna gain are among the most important expressions used to describe the interaction between antenna and propagating medium.[1] Depending on the frequency band concerned, wave propagation provides optimization criteria and often limits as well. This refresher course will remind the reader of important facts concerning wave propagation and indicate the implications for the application of antennas.

Figure 2 summarizes the main forms of electromagnetic-wave propagation in current use and the typical frequency ranges of each type of propagation.

Free space and tropospheric propagation usually occur above 20 MHz for terrestrial radio links. The *ground wave* is chiefly used under 5 MHz and can be derived from free-space propagation. Roughly speaking, the lower the frequency the better the ground-wave propagation. One of the more important applications in professional radiocommunication is the use of *skywave links via the ionosphere* in the frequency range 1.5 to 30 MHz. Telecommunication satellites have only partially been able to take over the functions of the shortwave band. Calculation capabilities have led to considerable advances, and they now allow a shortwave radio link to be treated and optimized as a complete system. High-angle transmission plays an important role in skywave communication, and the associated wave propagation makes particular demands on the characteristics of an antenna.

Propagation in free space

Figure 3 shows, for an oscillation of period T over a number of moments in time, how the field of radiation builds up around a dipole, frees itself from the antenna, and is launched as a wave into space.[3]

The *free-space field strength* E_o of a halfwave dipole in the far-field region is:

$$E_0 = 222 \cdot \frac{\sqrt{P}}{d} \frac{mV}{m} \qquad (1)$$

where:

P = power in kW fed to the antenna
d = distance from the transmitter in km

If the antenna has a gain, G, referred to the isotropic radiator then:

$$E_0 = 173 \cdot \frac{\sqrt{P \cdot G}}{d} \frac{mV}{m} \qquad (2)$$

Free-space propagation occurs when radiocommunication is established between two antennas situated well away from the surface of the earth and reflecting or absorbing objects. The transmitter and receiver must be small in comparison to the antennas and incorporated in the antenna feed points, so interference from connecting cables is avoided.

Listing these conditions shows that genuine free-space propagation between two antennas rarely occurs. In practice, it is usually sufficient to keep the first Fresnel zone free of obstructions, as shown in **Figure 4**. The first Fresnel zone is defined as the surface of an ellipsoid of revolution, with the transmitting and receiving antennas at the focal points at which a reflected wave has an indirect path half a wavelength longer than the direct path between two antennas. If an obstacle protrudes into the first Fresnel ellipsoid, the effects of *shadowing and diffraction* need to be investigated. Because these conditions occur frequently, we'll consider the example of a single high obstacle. The field strength, E, is obtained in the region of the obstacle with reference to the free-space field strength E_o:[4]

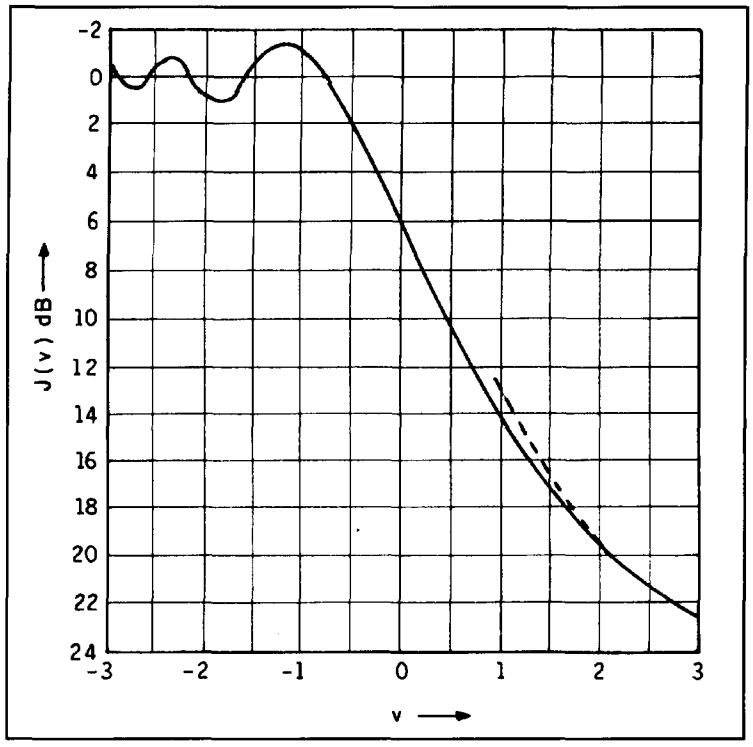

Figure 6. Diffraction loss relative to obstacle in transmission path given by dimensionless parameter v. Dashed line: approximation $J(v) = 3 + 20 \log v$.

$$\frac{E}{E_o} = \frac{1}{\sqrt{2}} \cdot \sqrt{[1/2 - C(v)]^2 + [1/2 - S(v)]^2} \qquad (3)$$

where $C(v)$ and $S(v)$ are the two Fresnel integrals:

$$C(v) = \int_0^v \cos \frac{\pi u^2}{2} \, du,$$

$$S(v) = \int_0^v \sin \frac{\pi u^2}{2} \, du. \qquad (4)$$

The dimensionless parameter v is defined by the height h and position d_1, d_2 of the obstruction shown in **Figure 5,** as well as by the wavelength λ:

$$v = h \cdot \sqrt{\frac{2}{\lambda}\left(\frac{1}{d_1} + \frac{1}{d_2}\right)}$$

$$= h \cdot \sqrt{\frac{2}{\lambda} \frac{d_1 + d_2}{d_1 d_2}} \qquad (5)$$

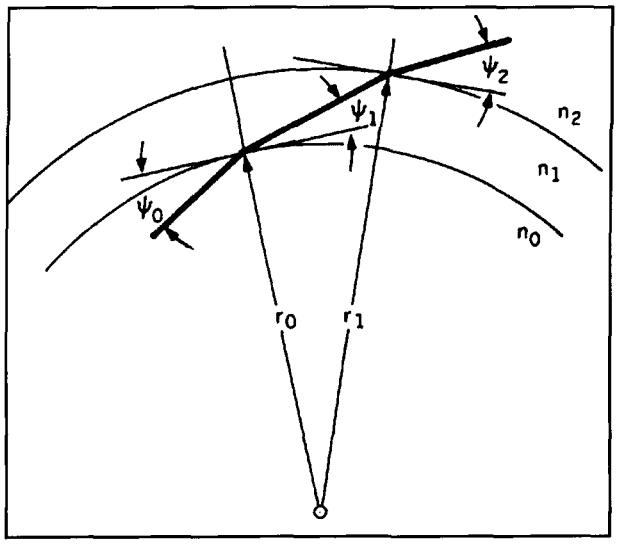

Figure 7. Refraction effect of layered troposphere; n_i = refractive index of air, r_i = distance from center of earth.

Figure 6 shows the transmission loss by diffraction (see Math 1 below)

$$J(v) = -20 \log \frac{E}{E_o}$$

from CCIR Report 715,[5] in the difficult to compute range of v. If v is greater than 1, an approximation can be used:

$$J(v) = 13 + 20 \log v \qquad (6)$$

The error compared to exact calculation is ≤1 dB.

Tropospheric propagation

What at first glance could be considered as free-space propagation is usually wave propagation in the troposphere near the surface of the earth. The dielectric constants and the refractive index of air generally decrease with increasing height above the ground. **Figure 7** shows how a wavefront is refracted by layers of different dielectric constants.[3] Due to the typical decrease in the refractive index of air with height, the resulting *bending back to the Earth's surface* leads to a range which is considerably larger than that assuming straight-line propagation (**Figure 8**). Severe refraction due to exceptionally large reductions in the refractive index with height is one cause of occasionally extended over-the-horizon ranges.

The curved radiation paths in the troposphere are not suitable for graphical representation. A useful aid are diagrams similar to **Figure 9**, in which the radius of the earth is corrected by mathematical transforms so that the radiation path becomes a straight line. For the so-called standard atmosphere, the radius of the earth is increased by a factor of k = 4/3.

The *height of the transmitting antenna* above its surroundings has considerable effect on the field strength obtained. If antennas are not

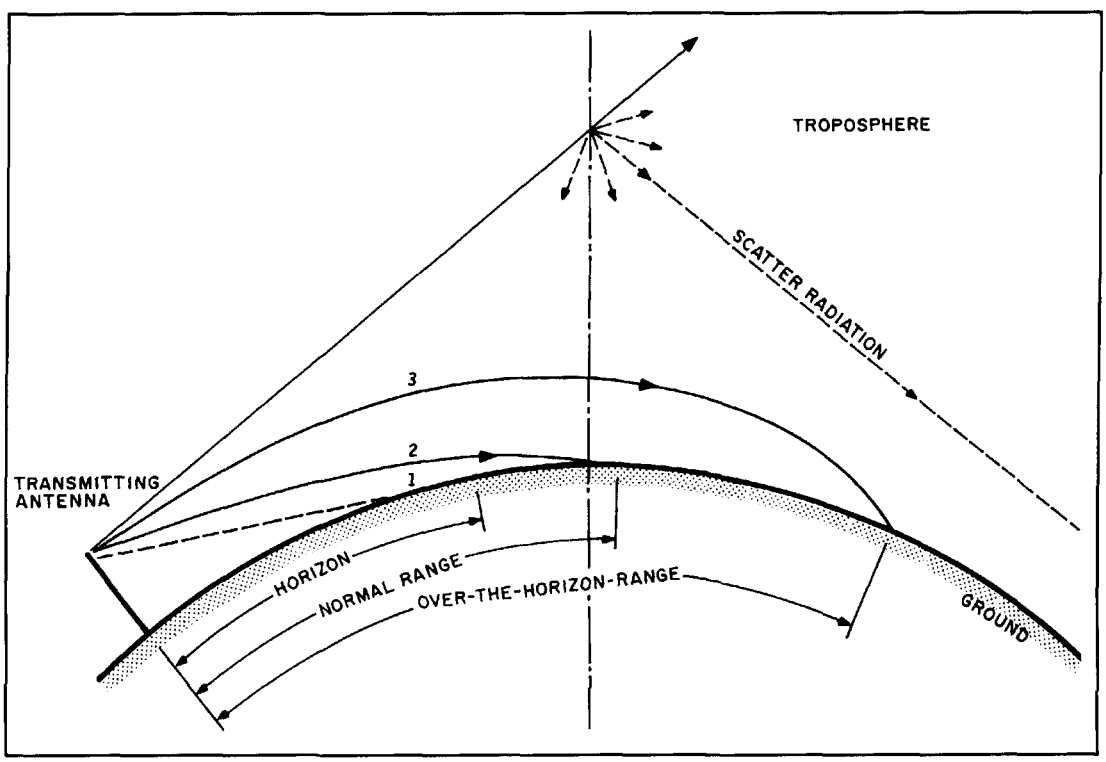

Figure 8. Range of electromagnetic waves with straight-line propagation (curve 1), with normal refraction (curve 2) and especially severe refraction (curve 3).

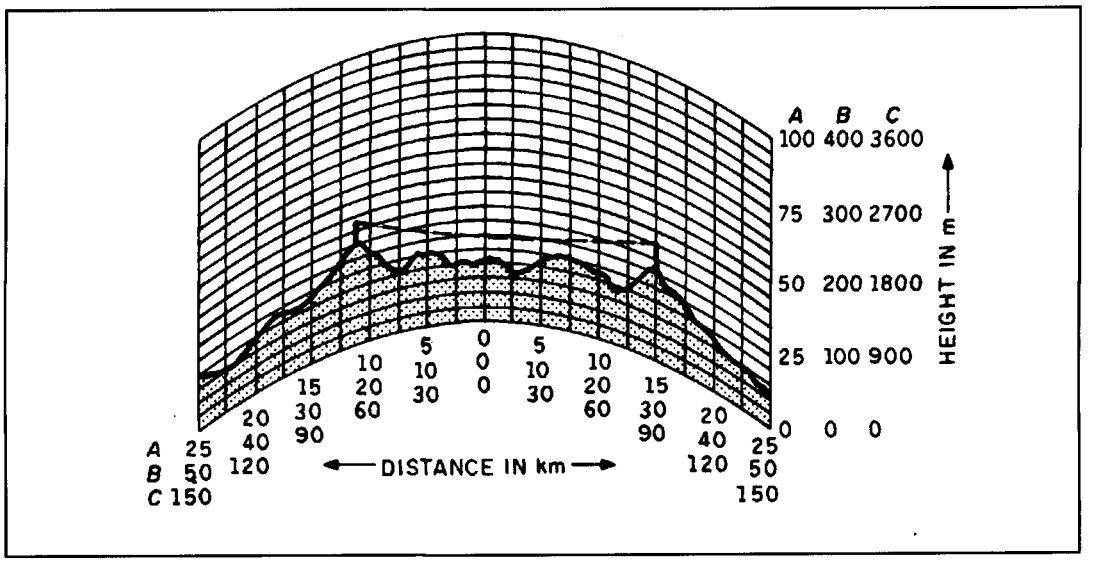

Figure 9. Diagram for straight-line transmission path with standard atmosphere.

erected very high above the ground, **Equations 1** and **2** for free-space approximation yield results that are far too high. Because evaluation by calculation is very time consuming, it is advisable to use the propagation curves given in CCIR Recommendation 370.[5] For example, **Figure 10** shows the field-strength values to be expected in the VHF band with radiated power of 1 kW ERP[1] at a height of 10 m above the ground (with the height of the transmitting antenna as the variable parameter).

The free-space field-strength curve is shown on the graph as a boundary value, and is only achieved when the antenna is very high above the ground. The curves presented on the graph are valid 50 percent of the time and for 50 percent of the locations over land, as well as over the North and Mediterranean Seas. Diagrams for other percentages of time and correction factors for location probabilities can be obtained from **Reference 5**, along with the corresponding graphs for the frequency range 450 to 1000 MHz.

The *irregularity of terrain* has a considerable effect on the average field strength that can be achieved. The curves in **Figure 10** are valid for a terrain of average irregularity in which height differences greater or less than $\Delta h = 50$ m occur in only 10 percent of all cases (**Figure 11**). Over terrain with lower differences in height, higher values of field strength are obtained. **Figure 12** is from CCIR Recommendation 370 and shows the attenuation correction factors for a wide range of height differences over land. The effect of terrain irregularity depends on the distance from the transmitter and decreases with longer distances. Note also, that when the correction factors are used with **Figure 10**, they still yield

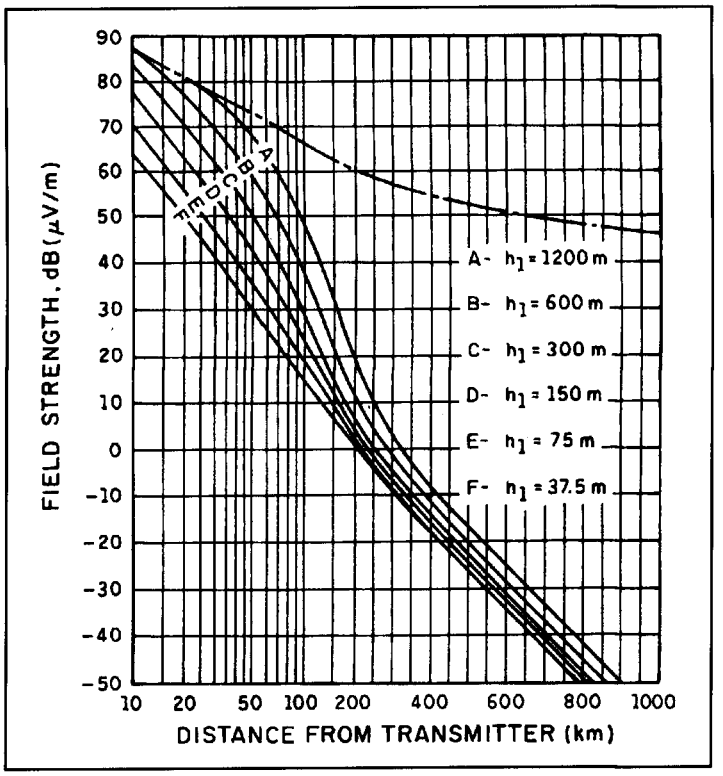

Figure 10. Field strength in VHF band at height of 10 m above ground with radiated power of 1 kW ERP for different heights h_1 of transmitting antenna (solid lines). Dot-dash line shows field strength in free space.

only average values for 50 percent of the locations, and not for every site on land; particular shadowing and diffraction problems must be treated separately.

The troposphere is constantly undergoing *changes with respect to time*. Information about the attainable field strength can therefore only be statistically based. Field strengths several orders of magnitude above the average values

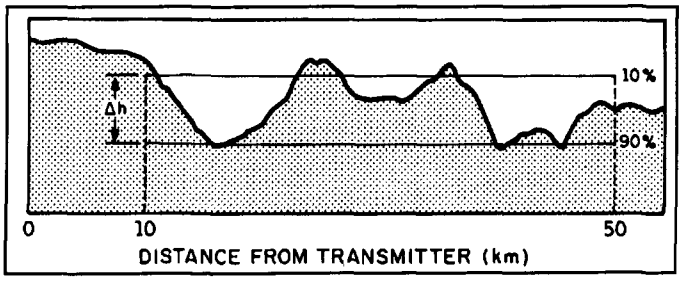

Figure 11. Definition of terrain irregularity. Average irregularity exists when differences in height are more or less than Δh = 50 m in only 10 percent of cases.

given in **Figure 10** may occur, particularly at longer distances from the transmitter. **Figure 13** gives an example of propagation in the region of the Mediterranean Sea. The field strengths for 10, 5, and 1 percent of the time are considerable above the average value (50 percent), which is due partly to the scatter and duct effects mentioned later. These graphs provide decision aids in optimizing and planning systems not only for broadcast coverage by radio and television, but also for radiolocation.

From the propagation curves in **Figure 13** and other relevant characteristics of the receiving equipment, data can be derived to help in the *choice of receiving antenna*. **Figure 14** shows the gain required by a receiving antenna in the Mediterranean region for reception of trans-horizon VHF signals over the given distances with respect to the percentage of time on the abscissa. For instance, if reception for 10 percent of the time over very large distances is considered a success, **Figure 14** shows that an omnidirectional 100-watt transmitter can be received with a simple omnidirectional receiving antenna—with a gain of about 2 dB—at a distance of 400 km.

Photos A and **B** show antenna systems from Rohde & Schwarz. The type of omnidirectional receiving antenna shown can be mounted on the masts used here—even along with transmitting antennas. This is possible because of the exceptional isolation between the individual radiating elements.

If a receiving antenna with a gain of about 15 dB (such as the combination of log-periodic antennas shown in **Photo C**) is used, **Figure 14** indicates that for 10 percent of the time a 100-watt transmitter can be received at a distance of more than 600 km. Due to the higher gain of the receiving antenna, reception over 400 km is possible 30 percent of the time. With a 30-dB high-gain receiving antenna, like the model shown in **Photo D**, transmitters of the type mentioned above can be received 10 percent of the time over distances of 1000 km. Transmitters at a range of 600 km can be received approximately 30 percent of the time and transmitters 400 km distant can be registered nearly continuously.

Taking into account the *frequency-dependent criteria*—for instance the first Fresnel zone, the field-strength curves, and the corrections for terrain irregularities—the techniques of field-strength evaluation described can, in principle, be used up into the upper UHF band. At still higher frequencies, additional propagation attenuation caused by rain and fog must be taken into consideration, as well as narrowband absorption effects due to resonant components in the air. **Figure 15** shows this additional attenuation according to CCIR Report 719, referred to a propagation path of 1 km.[5]

I have not yet mentioned two special forms of tropospheric-wave propagation. In *scatter transmission*, the scattering of the radiated waves by collision with inhomogeneities in the atmosphere is used to provide over-the-horizon communication (**Figure 16**). The scattering usually occurs in all directions, so only a very small portion of the transmitted power reaches

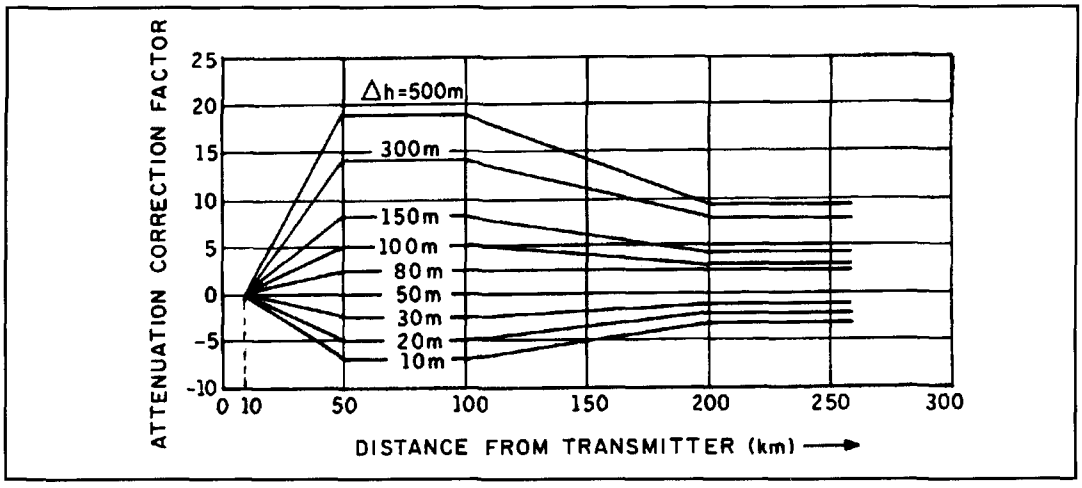

Figure 12. Attenuation correction factors for field strength with various differences in height of terrain.

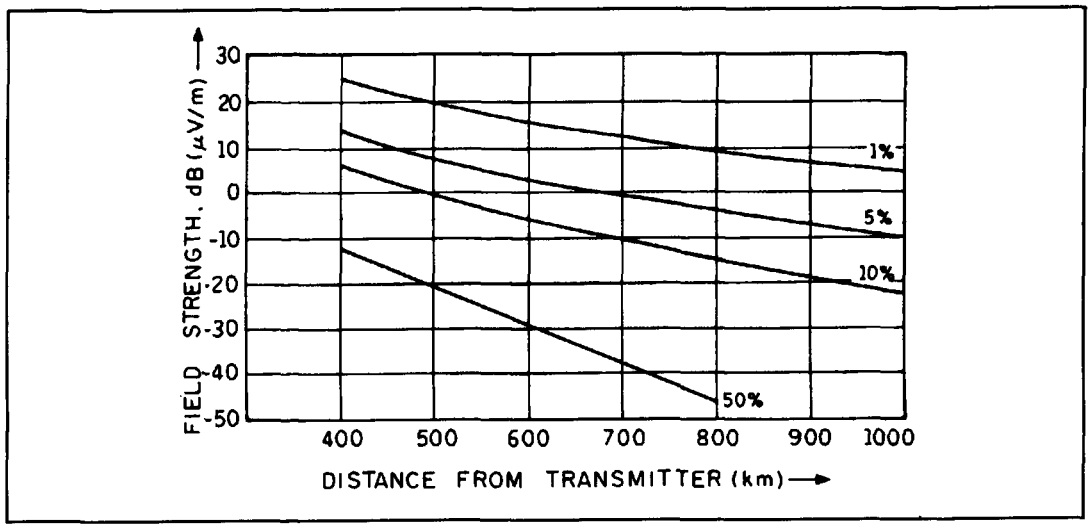

Figure 13. Example of achievable field strength for 50, 10, 5, and 1 percent of transmission time at h_1 = 300 m (transmitting antenna), h_2 = 10 m (receiving antenna) and 1 kW ERP.

the receiving site; high antenna gain enables the transmitter power required to be kept within reasonable limits. Propagation data are available in CCIR Recommendation 238.[5]

Duct propagation also enables exceptionally long ranges to be achieved. In this case, the waves are fed as if through a waveguide (**Figure 17**) whose "walls" are formed by sharp changes in the dielectric constant of the propagation medium. Duct propagation occurs only sporadically. Until now, it has been less used systemically than, say, scatter propagation. Data relevant to duct propagation are available in CCIR Report 718.[5]

Ground-wave propagation

Propagation of ground waves across an ideally conducting surface can be derived directly from propagation in free space. If, with a dipole in free space as shown in **Figure 3**, an ideally conducting surface is arranged at the feedpoint of an antenna and perpendicular to its axis, one obtains the characteristic field pattern of a vertical antenna. How a wave frees itself from an antenna can be shown very clearly in this case.[6]

Wave separation

Figure 18 shows an electrically short antenna with an antenna height $H \leq \lambda/10$, which occurs frequently in practice with the low frequencies of ground-wave propagation. In **Figure 18**, a) shows a moment in time at the start of period T; the charge on the antenna is initially low. Drawing b) shows the moment of highest charge reached after a quarter period, t = T/4. For the remaining diagrams, the dashed

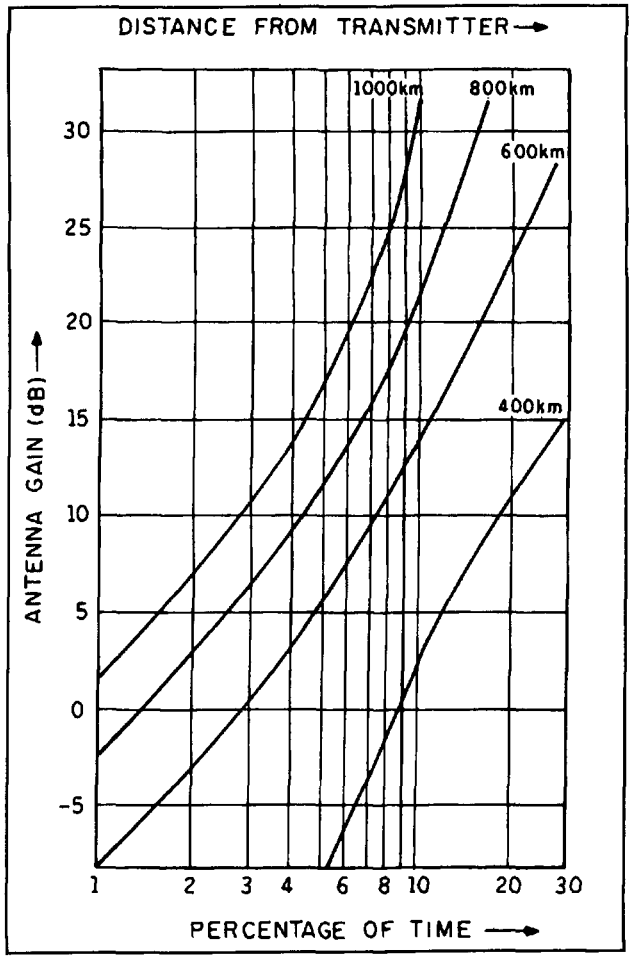

Figure 14. Gain required for receiving antenna for various percentages of time and distances from transmitter.

boundary line of electric flux is important because all lines of electric flux within this limit return to the source in the second quarter period c), such as flux line A with charge 1. The flux lines external to the boundary line of

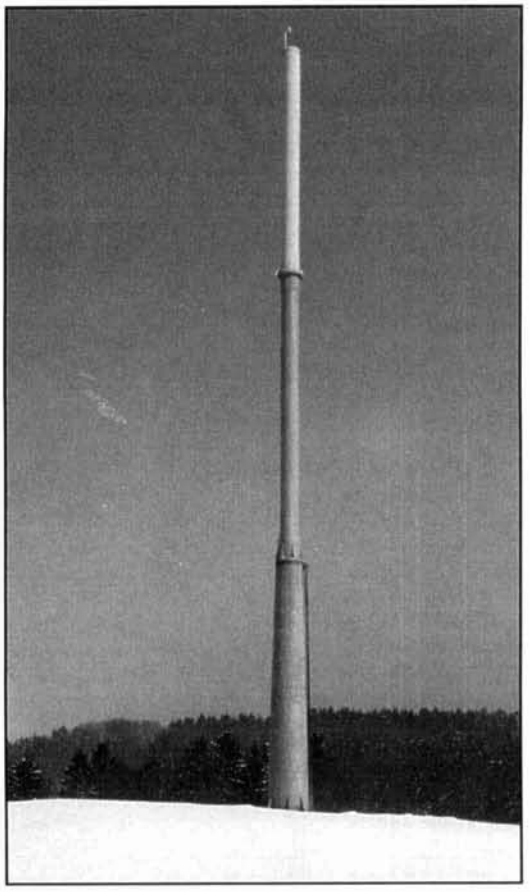

Photo A. Antenna System HK.53 from Rohde & Schwarz for VHF, UHF, or VHF + UHF.

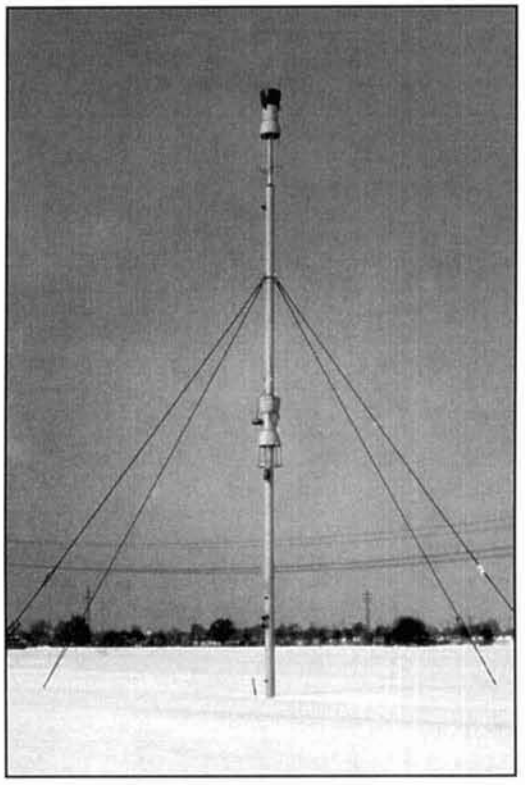

Photo B. The AK 214 antenna for VHF and UHF.

the flux—for instance flux line B—behave completely differently. On the conducting surface, charge 2 moves away from the antenna while the corresponding charge on the antenna itself moves in the direction of the feedpoint.

To understand the deformation of field line B in d), the magnetic components must be taken into consideration. Wherever charges move, that is to say wherever current flows, a magnetic field is automatically built up together with the associated magnetic flux Φ In addition, with fields varying with time, there is also a magnetic flux varying with time, or $d\Phi/dt$. In c) and d) the maximum flux of $d\Phi/dt$ is at the leading edge of the wave, where it forces the flux line to bulge upwards. Because charge 1 moves towards the feedpoint of the antenna while charge 2 moves away from the antenna, a point of zero change and zero current is formed between the two charges. In the region of this zero-current point, the magnetic flux and its time-varying component $d\Phi/dt$ are also very low. At this point, the effect of varying magnetic flux causing the outward bulge further away from the antenna is no longer present. Consequently, flux line B "falls" to the conducting surface.

In e) flux line B has made contact with the conducting surface; the line divides and charges 3 and 4 are obtained. Flux line B and the associated charge 3 move back towards the feedpoint as flux line A did previously. Electric-flux-line section B2 and charge 4 are now separated from the antenna and travel off into free space. Diagram f) shows, for a moment somewhat later in the second quarter period, how the next flux line C "falls" into the low field region between B1 and B2, then makes contact with the conducting surface and divides in the same way as flux line B. In g), C1 has nearly returned to the feedpoint of the antenna. B2, C2, and all other components of the lines of electric flux external to the limit line in b) leave the antenna radially at the speed of light.

In the following half period, the antenna is charged again. Because of the opposing polarity of the source voltage, the signs of the charges and the resulting direction of the lines of electric flux thus defined are changed over; otherwise the procedure is the same as described for the first half period. The newly formed field packet "pushes" the already existing packets from the previous half periods out into space (**Figure 19**).

To summarize: the source supplies energy to the system in the first and third quarter periods of oscillation, and reaches its maximum at $t = T/4$ and $t = 3T/4$ of the field when the charge on the antenna is at its greatest. Some of the energy flows back into the source as reflected power in the second and fourth quarter periods. The

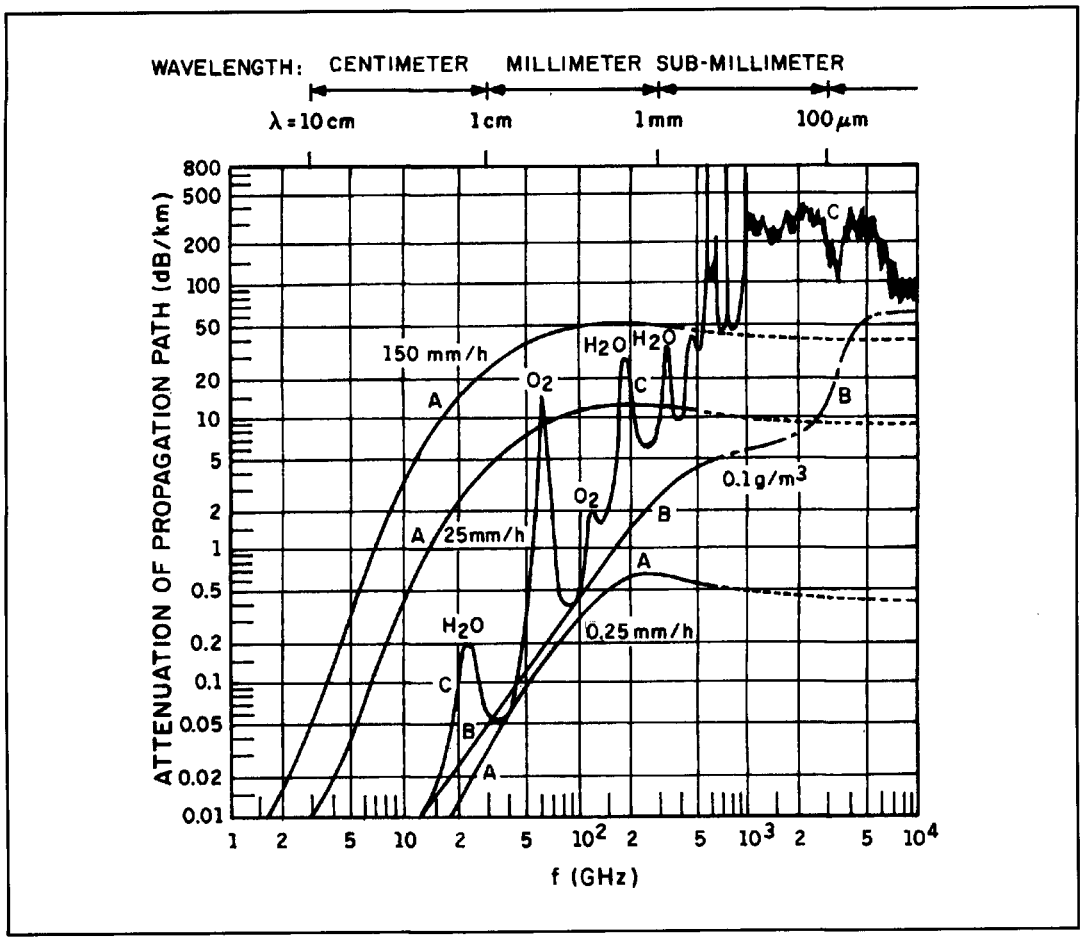

Figure 15. Additional field-strength attenuation at very high frequencies caused by (A) rain, (B) fog, (C) molecular absorption by water vapor and oxygen in air.

remaining energy forms the field components of the radiated wave, and is the active output energy. The ratio of reflected power to active power depends very much on the size of the antenna in relation to wavelength, to the shape of the radiating elements, and in particular on the feedpoint construction. This is expressed in the input impedance of the antenna.

As the figures show, the procedure of wave separation is very closely associated with active and reactive currents in the area immediately surrounding the antenna. If propagation does not occur over a well-conducting surface such as sea water, but over the poorly conducting real-earth surface instead, an earth net provides a means of reducing the loss caused by current heating of the ground and of approaching wave separation in the ideal case with a highly conducting surface.

Field-strength determination

For the propagation of ground waves over an *ideally conducting surface*, the far-field strength E_0 can be roughly calculated with a similar, simple rule-of-thumb formula as for free-space propagation:

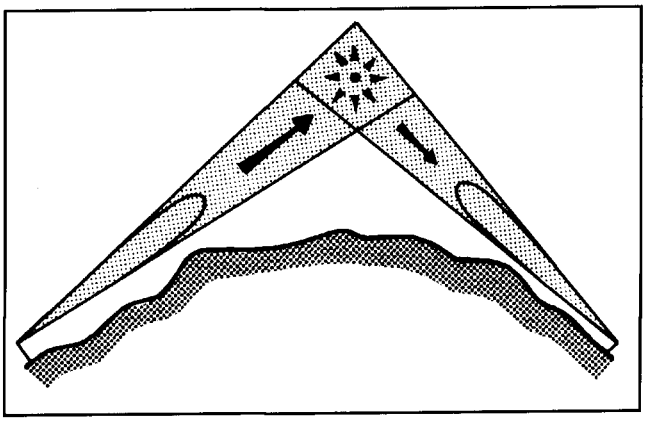

Figure 16. Scatter propagation.

$$E_o = 300 \cdot \frac{\sqrt{P}}{d} \quad \frac{mV}{m} \qquad (7)$$

where:

P = power in kW fed to the antenna
d = distance in km

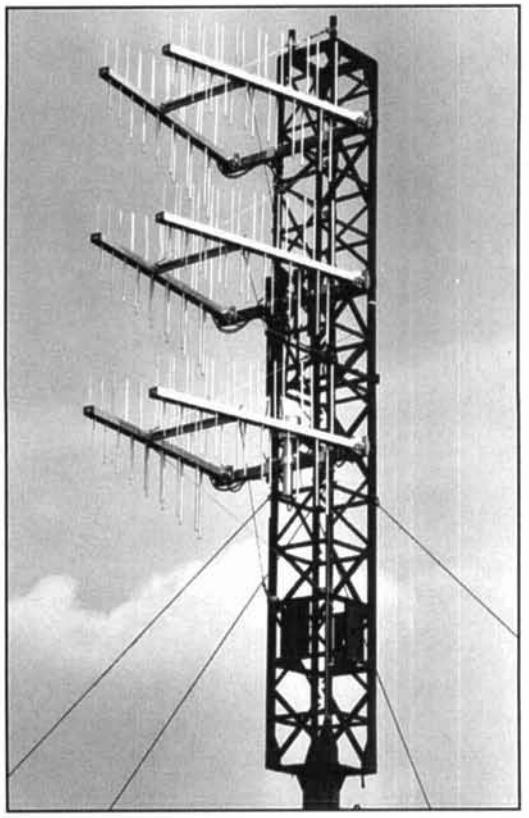

Photo C. Array of log-periodic antennas for reception with gain of about 15 dB.

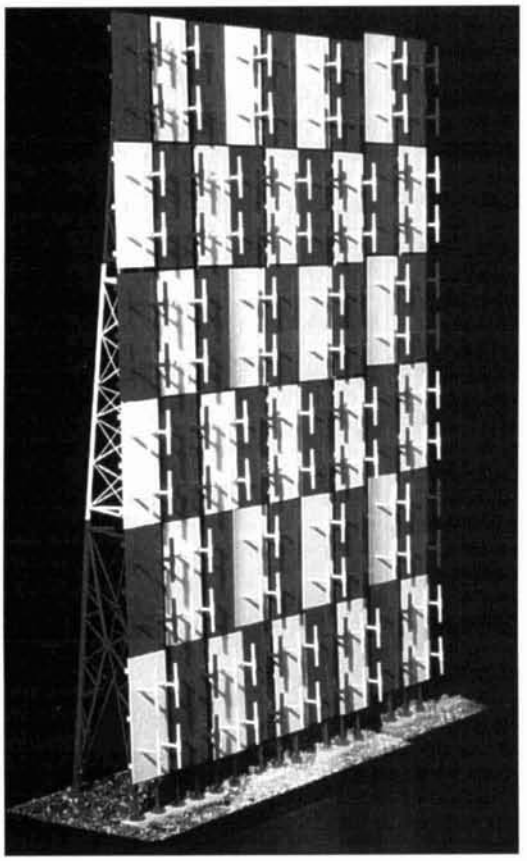

Photo D. Model of 30-dB high-gain receiving antenna.

The approximation is valid for an electrically short, vertical antenna (height $H \leq \lambda/10$). For a quarterwave antenna field strength is about 5 percent higher, which usually has no practical importance. Note that the losses—for instance into the ground—and the efficiency of matching networks must be taken into account.

Although, strictly speaking, **Equation 7** is only valid for ideally conducting plane surfaces, the results are often precise enough for the following cases:

- with highly conductive subsurfaces; for example, sea water,
- at very low frequencies,
- at relatively short distances (however, within the far-field region).

The deviation from more accurate methods can be seen in the propagation curves from CCIR Recommendation 368[5] shown later, in which the approximate value is plotted as a limit; that is, the inverse distance curve.

The *subsurface that is actually present* for ground-wave propagation sometimes leads to distinctive differences from the behavior already discussed. Its characteristics are described by the complex dielectric constant $\underline{\varepsilon}_r$:

$$\underline{\varepsilon}_r = \varepsilon_r - j\,60\,\sigma\lambda \qquad \textbf{(8)}$$

The complex dielectric constant is derived from the real dielectric constant, ε_r, and an expression from the product of conductivity in S/m and wavelength λ in m. Low conductivity can be compensated for, purely in terms of propagation, by choosing lower frequencies.

Values for frequently occurring types of ground from **Reference 7** and CCIR Recommendation 524[5] are summarized in **Table 1**. If the spread of conductivity between the extreme cases of desert and sea water is considered, it is possible to obtain an idea of how different the achievable range of the ground wave can be in both cases. A correlation can be observed between dielectric constant ε_r and conductivity σ of the types of ground surfaces that are usually encountered; this is shown in **Figure 20**.[8]

The field pattern of a ground wave with finite conductivity of the subsurface is shown in **Figure 21**. Whereas the lines of electric flux are vertical to the conducting surface as $\sigma \rightarrow \infty$ (**Figure 19**), a field vector which is slanted with respect to the ground is obtained in this case. It enables the vertical component $\underline{E}_z$ and a radical component $\underline{E}_r$ to be derived in the direction of propagation. The ratio of these field-strength components can be obtained from the complex dielectric constant:

$$\frac{\underline{E}_z}{\underline{E}_r} \approx \sqrt{\underline{\varepsilon}_r} \qquad \textbf{(9)}$$

The two components are not in phase, so the field vector describes an ellipse in a plane defined by the vertical and the direction of propagation. The ellipse in this case is not the usual elliptically polarized wave in which the plane of the field ellipse is *perpendicular* to the direction of propagation. For more detailed treatment of this see CCIR 879.[5]

With large wavelengths—i.e., very low frequencies— $\underline{E}_z/\underline{E}_r$ yield large values. The deviation from the ideal case is small and the field-strength vector is only slightly slanted at the boundary of the surface. In the lower shortwave band, on the other hand, a distinctive slant is obtained with the usual types of ground. The direction of propagation of the wave is perpendicular to this slant and at an angle to the ground. In **Figure 21** the wave can be seen to penetrate the ground, where it is attenuated by heating losses. Thus, the field strength decreases with increasing distance more quickly than in propagation over an ideally conducting surface.

In calculations, this is taken into account by the Sommerfeld attenuation function $F_{(\rho)}$[2,3], by which the values obtained with **Equation 7** are multiplied:

$$E = E_0 \cdot F(\rho) \qquad \textbf{(10)}$$

The correction factor, F, is dependent on the magnitude of ρ, which is called the numerical distance. This takes into account the distance with reference to the wavelength, d/λ, and the complex factor, $\underline{C}$, to be discussed later.

$$\rho = j \cdot \frac{\pi \cdot d}{\lambda} \cdot \underline{C} \qquad \textbf{(11)}$$

Calculation of ground-wave field strength is somewhat time-consuming with real ground surfaces, as the Sommerfeld attenuation function is really dependent on distance. Practical evaluation is therefore usually carried out by using the graphs given in CCIR Report 368.[5] For example, **Figure 22** shows the field strength resulting from propagation of a ground wave across sea water. The abscissa is divided logarithmically; a large distance can therefore be shown with good resolution of smaller and middle distances. Each curve is marked with the associated frequency and wavelength. The limit curve of the field strength at ideal conductivity according to **Equation 7** (the inverse distance curve) is also plotted. As one can be see, it is only at relatively large distances or higher frequencies that field strength starts to show markedly lower values than with ideal conductivity. For ground-wave communications over open sea, good range can be expected, dependent in each case on transmitter power, transmitting and receiving antennas, receiver, bandwidth and required signal/noise ratio, as well as the operating frequency for larger distances as shown by **Figure 22**.

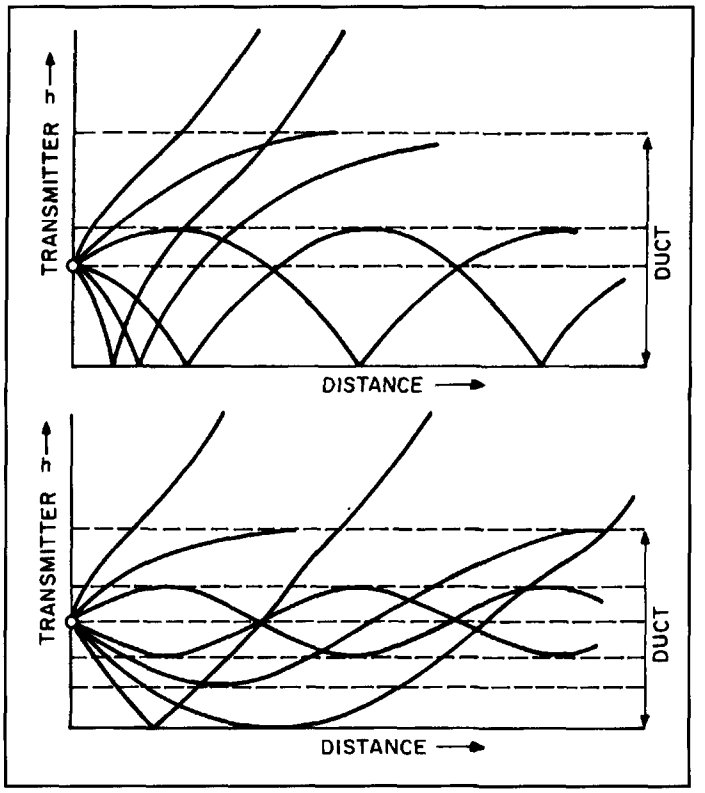

Figure 17. Duct propagation: top, surface duct; bottom, elevated duct.

A much more critical situation is illustrated by the field-strength curves over land in **Figure 23**. Although propagation is over pastural ground with good conductivity, the occurrences of distinctive differences from the limit (inverse distance) curve of **Equation 7** can readily be seen even at short distances in the shortwave band. Thus considerably lower field strength will be achieved across land with the same transmitting equipment specifications than across water, and this leads to a corresponding reduction in range. The larger spacing between the field-strength curves in **Figure 23** compared to **Figure 22** clearly indicates that the choice of frequency for ground-wave communications over land is especially important, and explains why only the lower frequencies of the radiotelephony and shortwave band (RT and HF band) are used in practice. For instance, the field strength at 5 MHz and a distance of only some tens of kilometers is already about one order of magnitude below the inverse distance curve. Operation at 1.5 MHz, for example, yields about 23 dB more received field strength at a distance of 50 km than for 5 MHz. The corresponding factor of 200 in the power is not fully available since, especially at night,

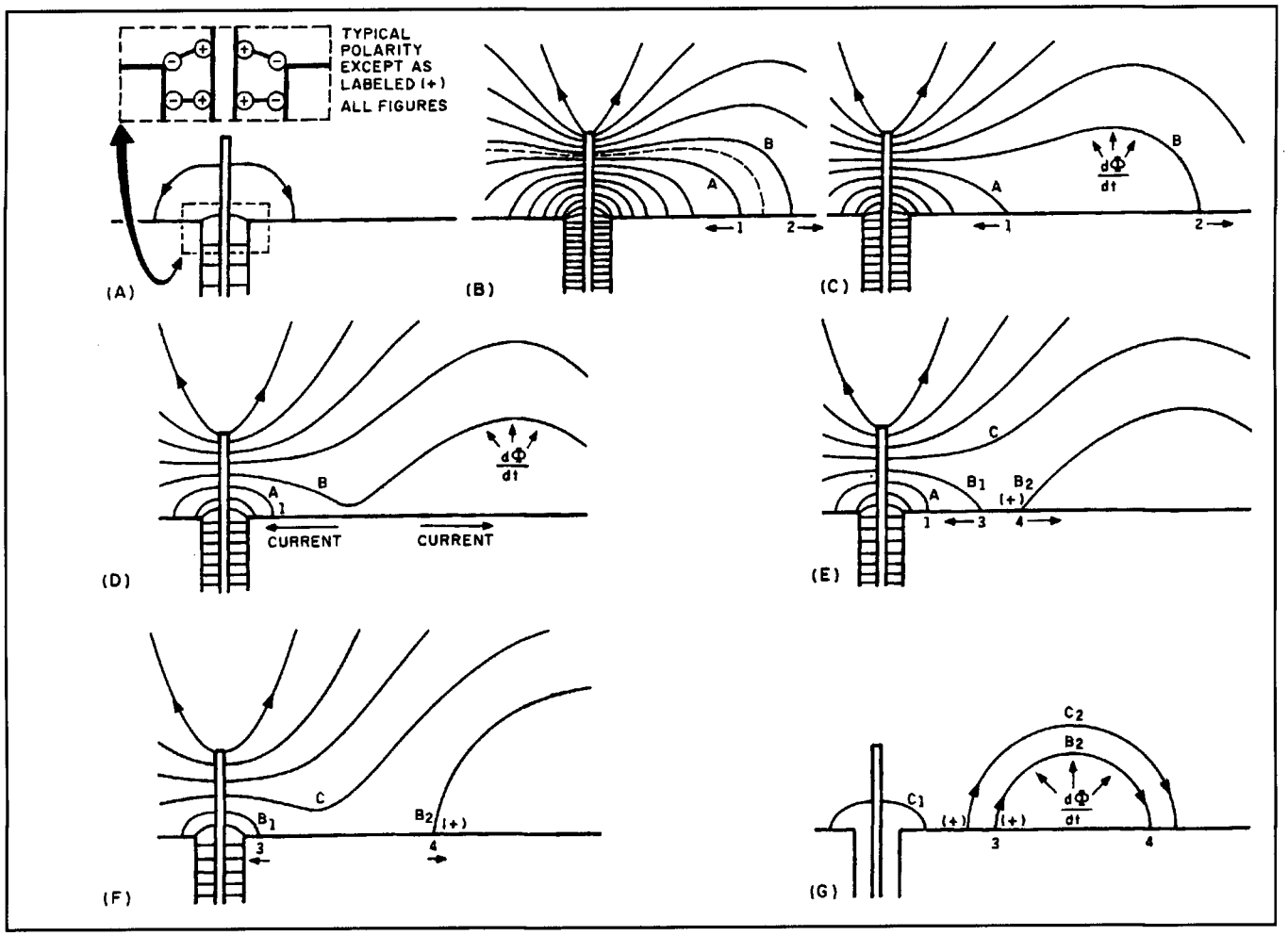

Figure 18. Wave separation from electrically short vertical antenna.

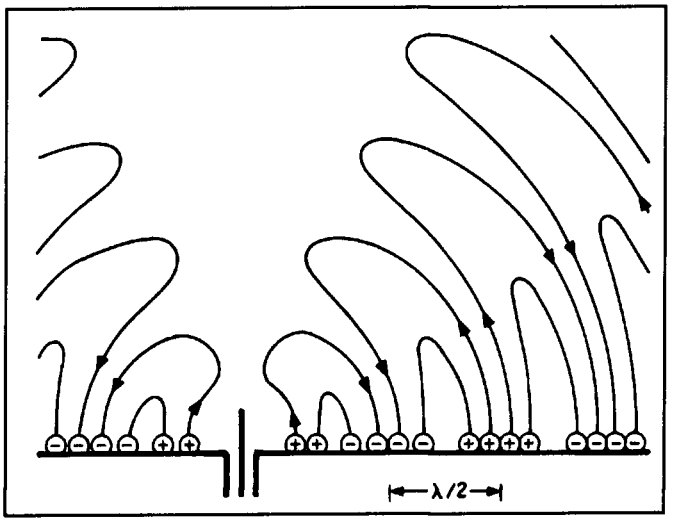

Figure 19. Field pattern of ground waves under ideal conditions.

atmospheric noise increases at lower frequencies. With electrically short antennas the transformation losses in tuning circuits increase towards lower frequencies. The fully automatic Antenna Tuning Unit FK 859[9] from Rohde & Schwarz sets standards with regard to high efficiency and helps, also from the point of view of wave propagation, to produce greater optimization of radio communication links.

Taking into account these partly opposing tendencies, **Figure 24**[10] shows that a change of frequency from 3 to 1.5 MHz for ground-wave communication over land leads to a considerable increase in range.

Polarization of ground wave

Equation 11 for the numerical distance contains a complex correction factor $\underline{C}$, which corresponds to the influence of the real ground. This factor is dependent on the polarization of the wave. For vertical polarization it is:

$$\underline{C}_v = \frac{\underline{\varepsilon}_r - 1}{\underline{\varepsilon}_r^2} \approx \frac{1}{\underline{\varepsilon}_r} \qquad (12)$$

and for horizontal polarization:

$$\underline{C}_H = \underline{\varepsilon}_r - 1 \approx \underline{\varepsilon}_r \qquad (13)$$

when $|\underline{\varepsilon}_r| \gg 1$.

Types of ground	Relative dielectric constants	Ground conductivity (S/m)
Clays	35 to 60	$3x10^{-2}$ to $3x10^{-1}$
Loam, marl, sandy clay	20 to 40	$2x10^{-2}$ to $2x10^{-1}$
Clayey sands, clayey slate	12 to 40	$3x10^{-3}$ to $3x10^{-2}$
Sand, gravel, sandstone, gypsum, limestone	4 to 40	10^{-4} to 10^{-2}
Crystalline rock	7 to 15	10^{-5} to $3x10^{-3}$
Rock salt, anhydrite	4 to 7	10^{-6} to $3x10^{-4}$
Ores	10 to 30	10^{-2} to 10^{4}
$\mu_r = 1$ (apart from a few exceptions with ores)		

Table 1. Values for frequently occurring types of ground.[7]

If typical values for the complex dielectric constant of the ground are substituted, propagation attenuation is obtained for horizontal polarization that is several powers of ten greater than attenuation for vertically polarized waves. Because of the higher horizontal attenuation, horizontally polarized waves die away after a short distance. As a result, a practical application has yet to be found for such waves. Ground-wave communication is therefore carried out exclusively with vertically polarized waves. Rod antennas as in **Photo E** are used to launch such waves; more recently, integrated antenna systems as in **Figure 25**[11] have also come into use and have proved to be very efficient in land-mobile applications.

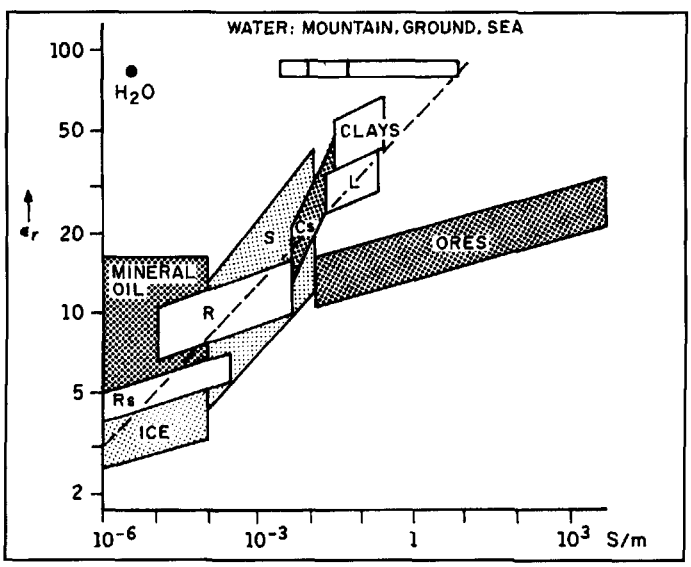

Figure 20. Ground conductivity, σ and relative dielectric constant, ε_r, for various types of ground and water. L: loam, marl, sandy clays; Cs: clayey sand, clayey slate; S: sand, gravel, sandstone, gypsum, limestone; R: crystalline rock; Rs: rock salt, anhydrite.

Nonhomogeneous propagation paths

Ground-wave communication between two ships on the high seas is one of the few cases where it is possible to assume a homogenous subsurface of known parameters. Across land it is more usual for conductivity to change along the propagation path. Conditions that need particular attenuation are land/sea transitions. Peninsulas and islands can cause considerable attenuation and "radio shadows" behind islands are well-known with ground-wave propagation and understandable under the circumstances. Calculation of paths with differing constants is possible; methods are given in Annex II of CCIR Recommendation 368[5] and **Figure 26** shows an example.

Calculations for this type of radio link, although relatively elaborate, can be solved using the computers available today. Greater difficulties are often experienced in obtaining data for calculations—especially data concerning conductivity and dielectric constant, which are affected by the water content of the ground.[8] The tendency of a wave at relatively low ground-wave frequencies to penetrate the ground to a considerable depth must also be taken into consideration. **Figure 27** provides examples of this taken from CCIR Recommendation 527.[5]

Propagation of skywaves via ionosphere

The major importance of the radiotelephony and shortwave bands (RT and HF bands) in radiocommunication stems from the possibility

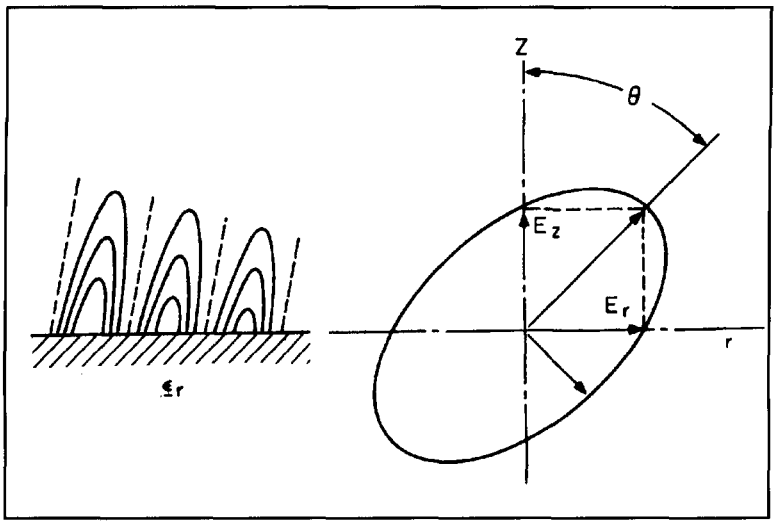

Figure 21. Field pattern and field vector for ground wave over nonideal surface.

they offer of being able to establish a radio link via the ionosphere to any point on earth. The very low cost of shortwave communication links in comparison to the alternatives—transmission via satellite, submarine cable, or a chain of relay sites across land—is only *one* reason for the upward trend in shortwave communications presently observed around the world. For many professional users, it is much more important that the risk of interference to and interruption of a communication link in a crisis is lower with shortwave than with the available alternatives. Even tropospheric scatter links, whose immunity to jamming make them the next choice after skywave communications via the ionosphere, are considered to have a higher risk factor because of the large antennas that are necessary. In the case of the HF band, spare antennas for emergencies may be held in readiness and quickly and easily installed.

Propagation paths

With a suitable choice of frequency and departure angle of the radiated waves, refraction back to the earth's surface takes place in the ionosphere, as illustrated in **Figure 28**. The maximum distance that can be covered with a single hop of this kind is between 2000 and 3000 km. The refracted waves can then be reflected from the surface of the earth. If the angle of incidence to the surface is low, the loss is negligible—even when the conductivity of the ground in the reflection area is low. The wave reflected from the ground is thus launched for the next hop, which will be over approximately the same distance, provided that the ionospheric layer that is in use is at the same height above the ground at the new point of refraction.

This is not always necessarily the case. On

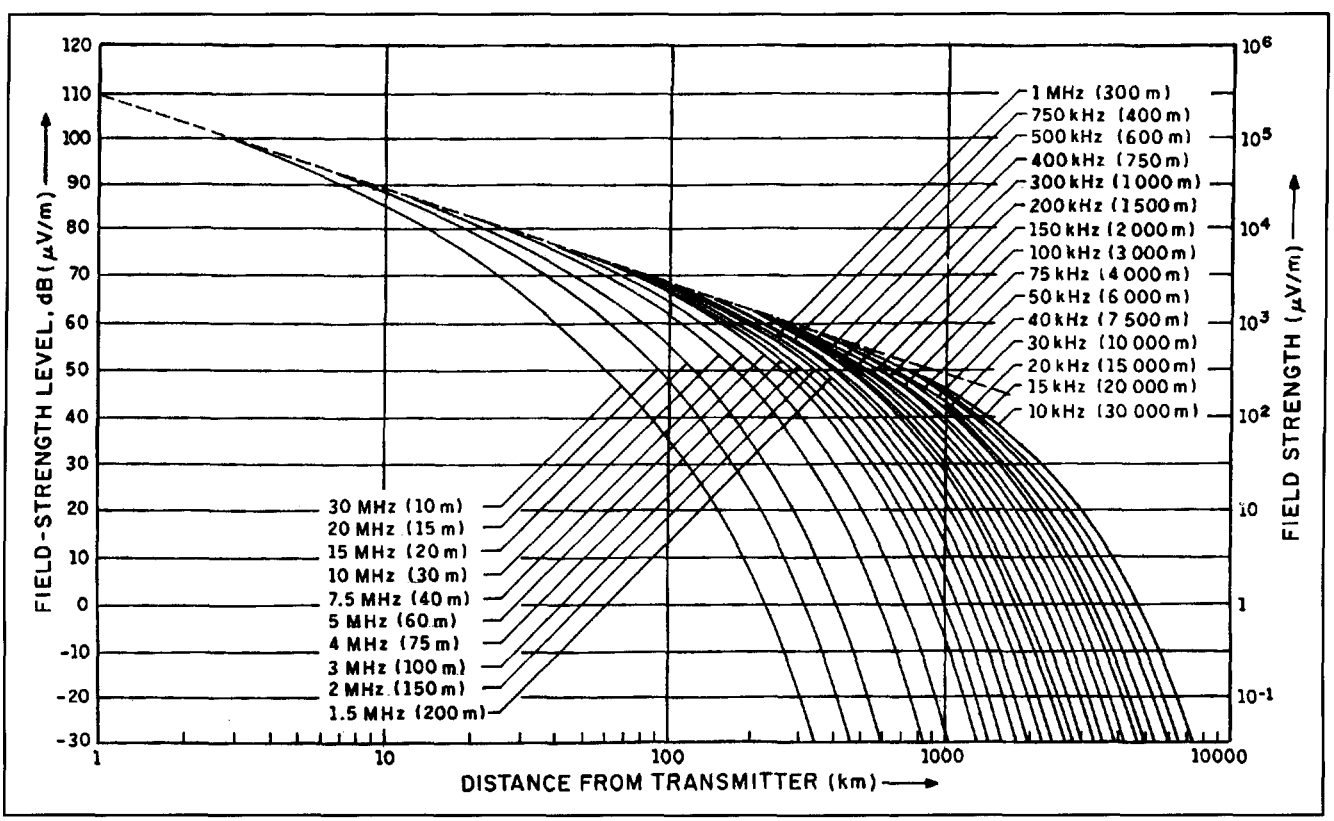

Figure 22. Ground-wave propagation curves over sea water (input power 1 kW into electrically short lossless vertical antenna).

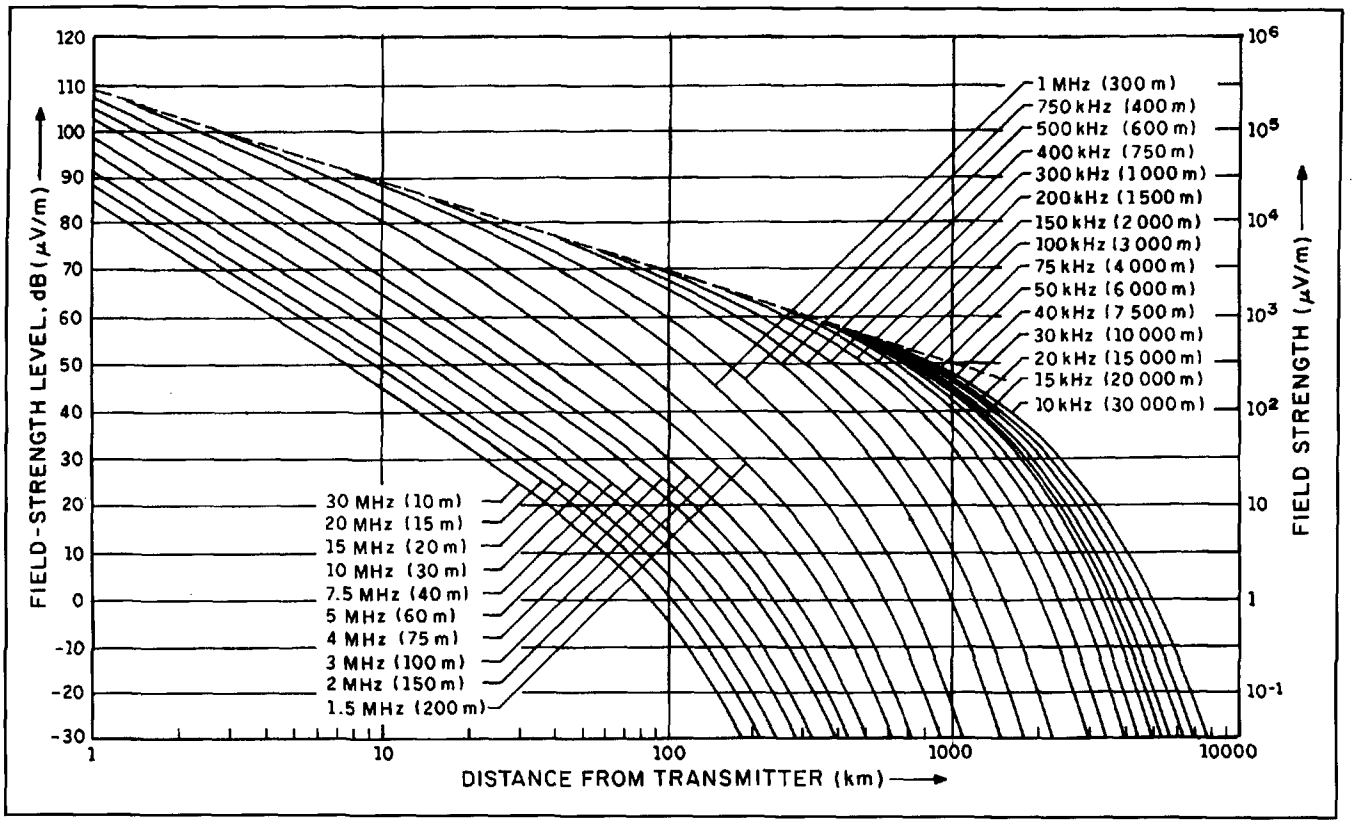

Figure 23. Ground-wave propagation curves over land (well conducting pastural ground). Data as in *Figure 22*.

long east-west radio links in particular, it may happen that the day-to-night transition is crossed. With the *F_2 layer* of the ionosphere, which because of its great altitude is most important for long-distance traffic, there are considerable differences in height above ground—especially in summer, as shown in **Figure 29**. The most noticeable day-to-night change occurs in the *E layer*. This layer is formed at a lower level than the F_2 layer, where there is greater particle concentration in the atmosphere. As a result, much shorter free-path lengths are available, over which a charged particle can travel before colliding with a particle of opposite polarity and recombining to form an electrically neutral molecule. With the setting of the sun, the energy source for ionization disappears; however, this condition is a requirement for the refraction of electromagnetic waves in the ionosphere. The number of newly split and recombined particles is no longer as high as during daylight hours. Shortly after sunset the E layer is no longer sufficiently ionized to refract shortwaves, and radio transmission is no longer possible in the HF band via the E layer.

The relationships are different with the F_2 layer. Due to the greater height of the layer, as a result of the much lower concentration of particles in the air, the free-path lengths of the charged particles are considerably greater. The time for recombination is therefore much longer. Even when the energy source has disappeared, a high enough level of ionization is retained until the next sunrise to cause refraction of the waves. Thus, as opposed to the E layer, the F_2 layer is available both by day and night for shortwave transmissions.

Figure 29 also shows the D layer, which is at an even lower height above the surface of the earth. For the same reason as the E layer, the D layer is only present during daylight hours for waves in the RT and HF bands. The waves are not refracted back to the surface of the earth by the D layer, but are attenuated as they pass through. Consequently, the attenuation—unlike many other forms of attenuation that can be observed—decreases with increasing frequency. The practical result is well-known to listeners of broadcast radio. During the day, medium wave stations can only be received in a narrowly defined radius via ground wave. Once the D layer disappears with the onset of night, skywave transmission is possible via the ionosphere and stations can be received over a much greater distance.

Other layers of the ionosphere play a lesser part. The *sporadic E layer* E_S lies some tens of kilometers above the E layer. As its name suggests, it occurs irregularly and is still not possi-

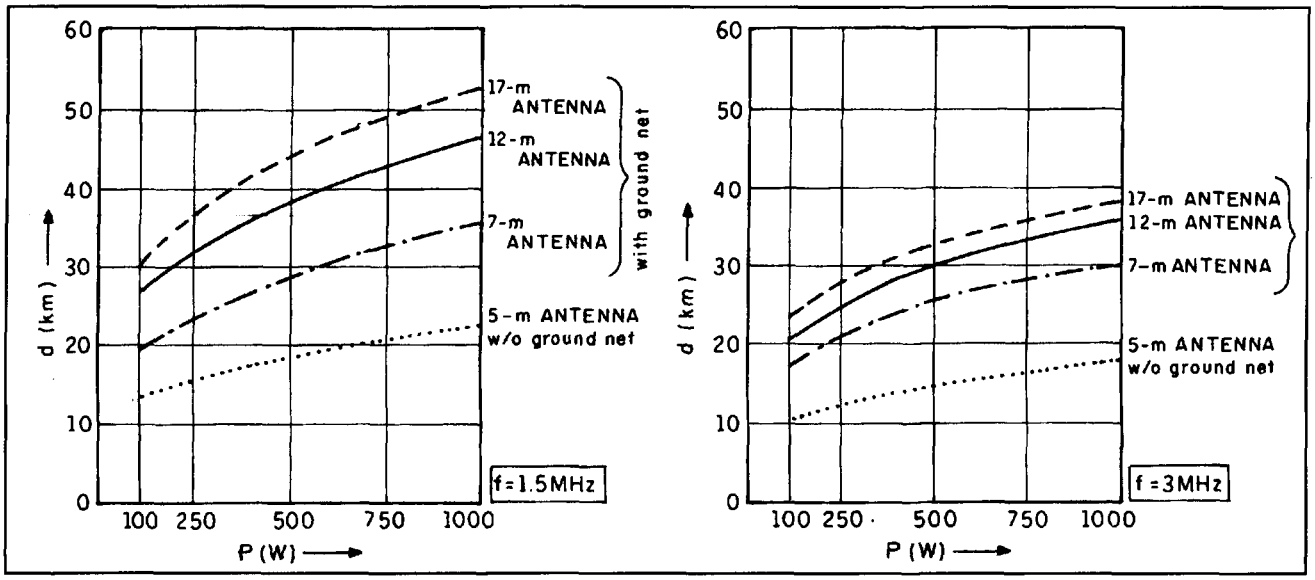

Figure 24. Effect of frequency on range of ground wave d (bandwidth 3 kHz; night).

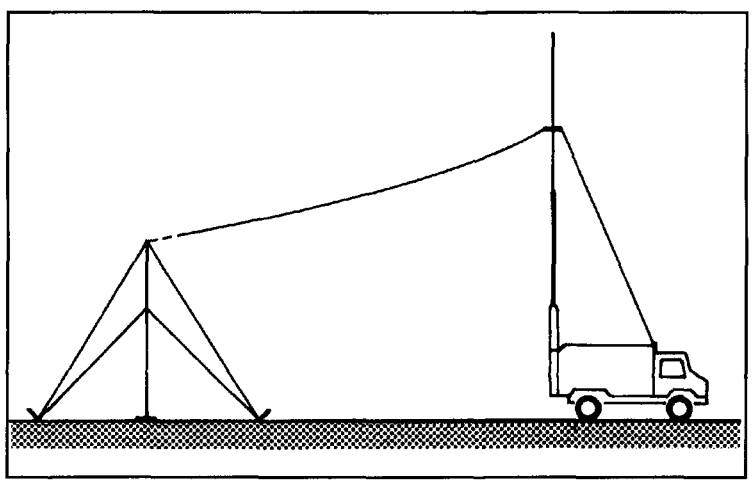

Figure 25. Vertical antenna (shown upper right) of HF Antenna System AK 501 is used for ground-wave communications.

ble to predict with any reliability. The *F_1 layer* lies between the E_S and F_2 layers and is of little importance compared to the F_2 layer.

The curved propagation paths in **Figure 28** are not suitable for graphic representation. So, as in the case of tropospheric propagation, a simplified approach was sought. **Figure 30** shows how the approximately straight propagation paths below the ionosphere can be extended to intersect at a higher point in the ionosphere. **Figure 31** shows a graph based on this principle. The corresponding distance and height scales also account for the curvature of the earth. The path of the wave from the transmitting antenna to the "reflecting" ionosphere, given by the corresponding height above ground on the ordinate, and back to the receiving site can now be plotted as a straight line. Charts of this type can be utilized for the propagation paths which normally occur. Special cases, like the upper example in **Figure 28** where waves are channeled over long distances by the ionosphere as in a waveguide, must be treated separately.

Frequency bands

The band between 1.5 and 30 MHz is the range of frequencies often called on for sky-wave transmission at shortwave. In specific cases, only part of this frequency band is suitable for a given link. If a wave is radiated towards the zenith, and is incident perpendicular to the ionosphere, reflection is possible up to an upper limit, or *critical frequency* f_c. **Figure 32** shows critical frequencies measured over a 24-hour period.[3] With the F_2 layer (left), a very noticeable dependence on the time of day can be seen. At night, the majority of the measured values found in the example presented are between 2.5 and 3.5 MHz. In the morning there is a sharp increase to the daylight frequencies between 6.5 and 8 MHz. In the late afternoon a fall to the lower nighttime frequencies can be observed.

For the E layer (right) an equally noticeable dependence on the time of day can be seen; however, the critical frequencies on the whole are distinctly lower. Since the E layer is only sufficiently ionized during the time of the sun's radiation, a characteristic response is obtained starting with the morning sunrise and ending with the darkness of evening. The extent to which the refraction capability of the E layer follows the sun can be seen from a comparison of the critical frequencies for December and June; it can clearly be seen how much shorter a winter day is at a latitude of about 50 degrees north than a summer day.

In practical applications the wave is not perpendicularly incident with the ionosphere, but

at an angle formed by the distance between the transmitting and receiving sites and the height of the reflecting layer of the ionosphere above the surface of the earth. In this case, the angle of incidence Ψ in **Figure 30**—the angle at which a wave enters the ionosphere and also leaves it after refraction—is particularly important because it determines the maximum usable frequency f_l at which waves will be refracted back to the receiving site.[2] The association is described in the first approximation by the *secant law*:

$$f_l = \frac{f_c}{\cos\Psi} = f_c \cdot \sec\Psi \qquad \textbf{(14)}$$

If a wave is beamed into the ionosphere at an angle of incidence Ψ, the frequency just reflected increases with increasing Ψ. This has far-reaching consequences for the association between the distance to be covered and the usable frequency band. The longer the radio link, the more acute the angle of departure of the waves needs to be. This is achieved by shaping the radiation pattern of the transmitting antenna,[12,13] so the required distance can be covered in a minimum number of hops. The shortest possible path for the waves, as well as the minimum number of passages through the D layer and reflections from the surface of the earth, generally lead to the lowest possible attenuation.

The secant law provides a usable approximation of the suitable frequencies for small to middling angles of incidence (high-angle radiation). However, for waves with lower angles of incidence, that is to say long-range traffic, the calculations yield frequencies that are arbitrarily high. This is contrary to practical experience. In this case, a more accurate approach must be used that takes into consideration the earth's radius r_E. Together with the maximum height h of the propagation path, a modified secant law is obtained:

$$f_l = \frac{f_c}{\sqrt{1 - \frac{\sin^2\Psi}{(1 + h/r_E)^2}}} \qquad \textbf{(15)}$$

The maximum value of f_l/f_c for the F_2 layer is about 3 in practice, and approximately 5 for the E layer.

The expression *MUF* (maximum usable frequency) for the upper frequency of skywave transmission is widely used. It should be noted, however, that besides the classic MUF there are a series of MUF definitions.[2] For instance, the "operational MUF" also takes scattering processes in the ionosphere into account, allowing certain types of transmission at frequencies above the classic MUF.

Photo E. HF Rod Antennas HA 175 on fast patrol boat.

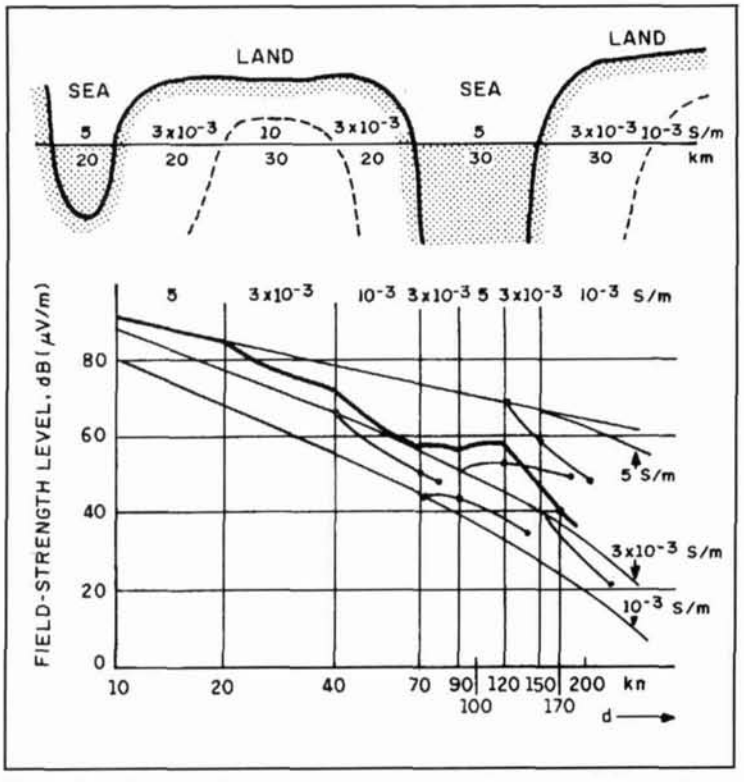

Figure 26. Field strength of ground waves over nonhomogeneous propagation paths (antenna data as for *Figure 22*; f = 700 kHz).

The *LUF* (lowest usable frequency) designates the limit of transmission of the lower frequencies.[14] LUF is mainly obtained from the square-law increase of attenuation in the D layer, with decreasing frequency, and is the limit at which a given minimum field strength at the receiving site is achieved with the available transmitter power. **Figure 33** shows a typi-

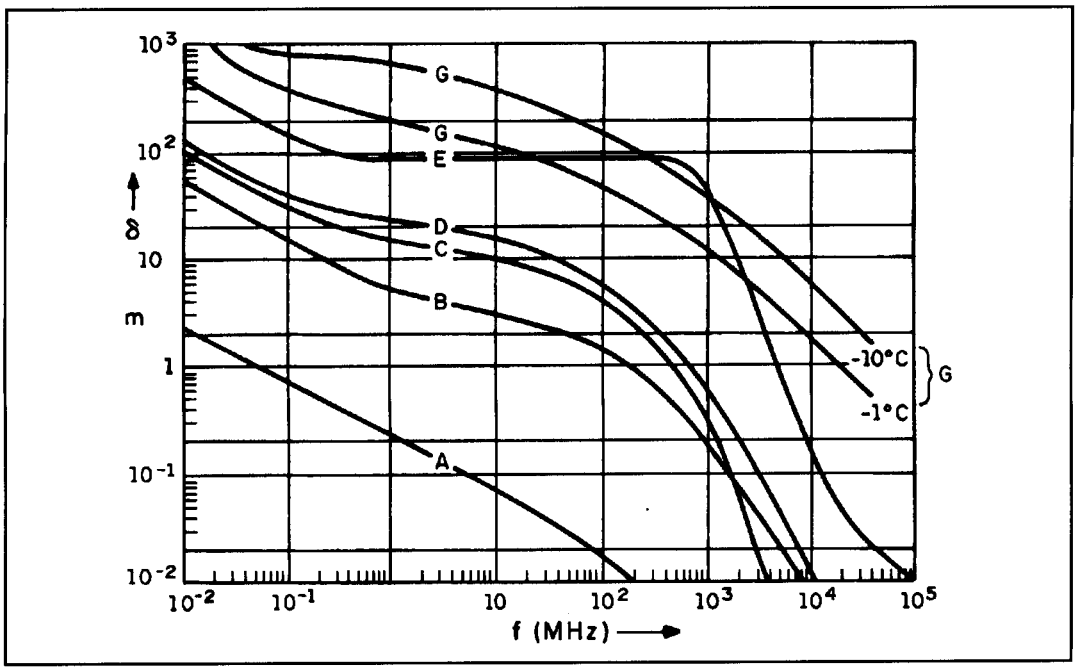

Figure 27. Penetration depth δ of ground waves into differing subsurfaces. (A) sea water, (B) moist ground, (C) fresh water, (D) average dry earth, (E) very dry earth, (G) fresh-water ice.

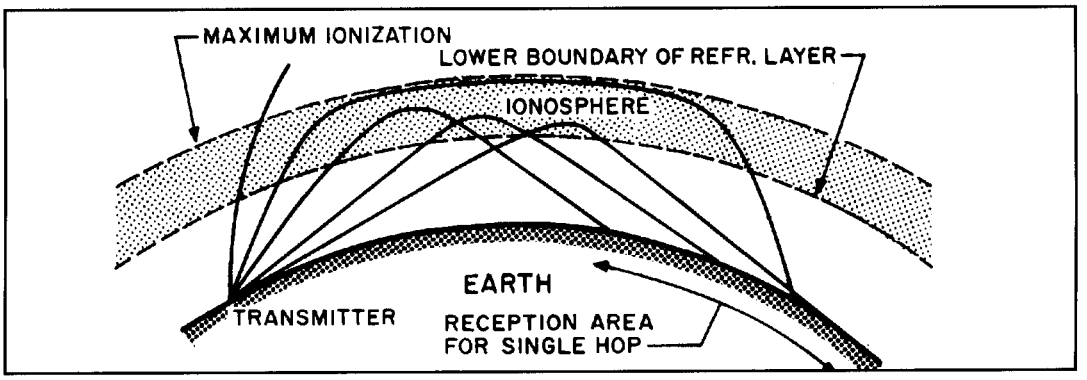

Figure 28. Skywave propagation via ionosphere.

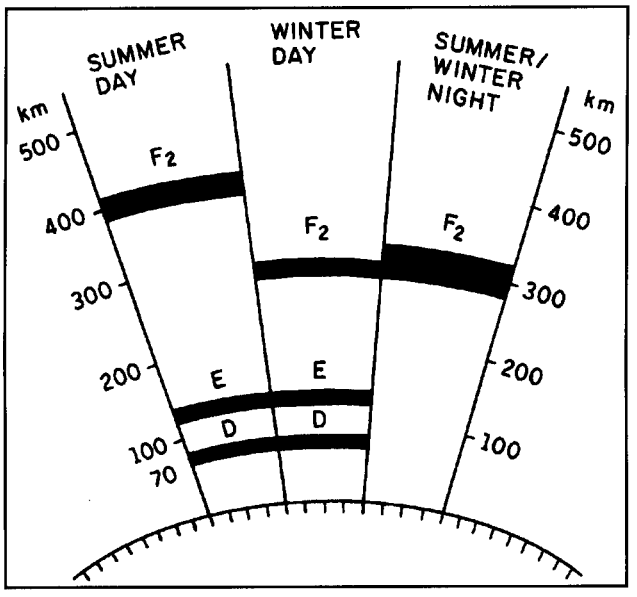

Figure 29. The more important layers of the ionosphere.

cal response of the transmission frequency band over a one-day period, extracted from a printout of a frequency forecast.

Field-strength calculations

Sky-wave transmission by way of the ionosphere is a complicated process. Dependence on time of day and season of the year, as well as the eleven-year sunspot cycle, needs to be taken into consideration. The geographic location where the refraction takes place is also important, because the bending of the waves in the plasma of the ionosphere is also affected by the earth's magnetic field. Unlike the other cases we've looked at, a rule of thumb for field strength cannot be given this time. Even rough estimates require a considerable amount of time for calculation. Field-strength calculations of high accuracy are performed using the comput-

er program developed and optimized by the "Ionosphere" research group of the Deutsche Bundespost at its telecommunications engineering center. In reliability forecasts not only are field-strength calculations performed, but the noise level at the receiving site is also taken into account. This enables a statistical forecast to be made of the probability that a desired signal/noise ratio will be achieved or exceeded. An example of a typical forecast printout is shown in **Figure 34**.

This forecast printout contains particularly valuable data for the antenna designer: under "ANGL.RX", the angle of elevation is given at which the radio waves are received. The evaluation of a number of similar forecast calculations enables elevation-angle statistics to be drawn up that can be used for optimizing the vertical radiation pattern of shortwave antennas. By taking into account radio links at various times of the day and seasons of the year with different relative sunspot numbers, as well as the various geographic regions, and evaluating all these parameters over a broad range of different distances, one will obtain the cloud of dots shown on the frequency versus elevation graph in **Figure 35**. This provides the highest REL value (reliability); i.e., the maximum link reliability under the given conditions.

The association between optimal angle of elevation and frequency shows a characteristic response, which can be described by statistical parameters. The parameter on the individual curves in **Figure 36** shows the percentage probability that the optimum elevation angle is less than the value on the ordinate. Note that in this example only small to medium elevation angles occur. This is due to the choice of distances to be covered, which in this case are for special areas of application for midrange distance and long-haul traffic.

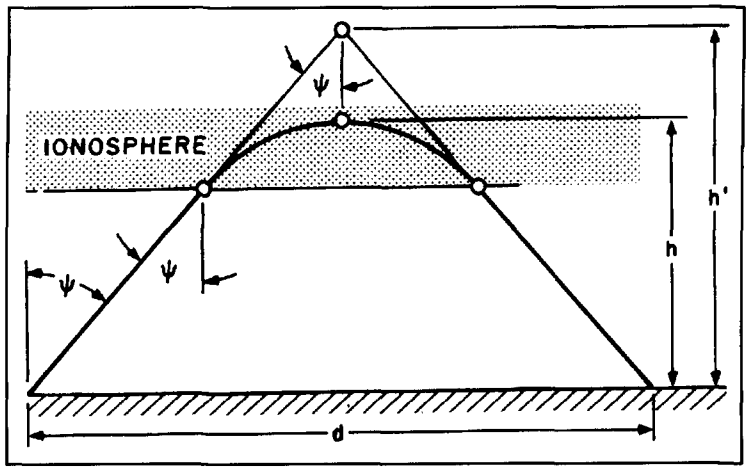

Figure 30. Apparent height of reflection point and actual propagation path on waves in ionosphere.

Using the elevation angle statistics, it is possible to derive the "desired vertical radiation pattern" for antennas to ensure maximum link reliability. For instance, if the range of angles between the 10 and 90 percent curves in **Figure 36** is selected, 80 percent of all cases that occur are covered. If a section is made through the

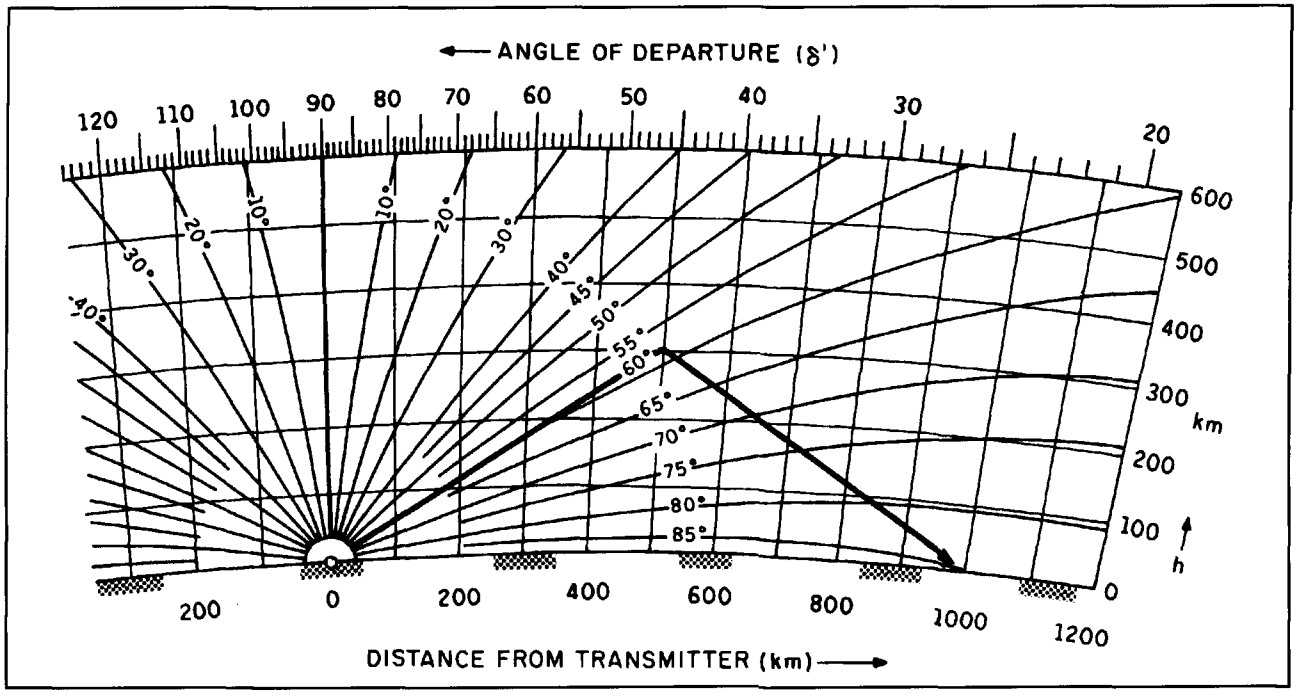

Figure 31. Chart for determining propagation paths via ionosphere.

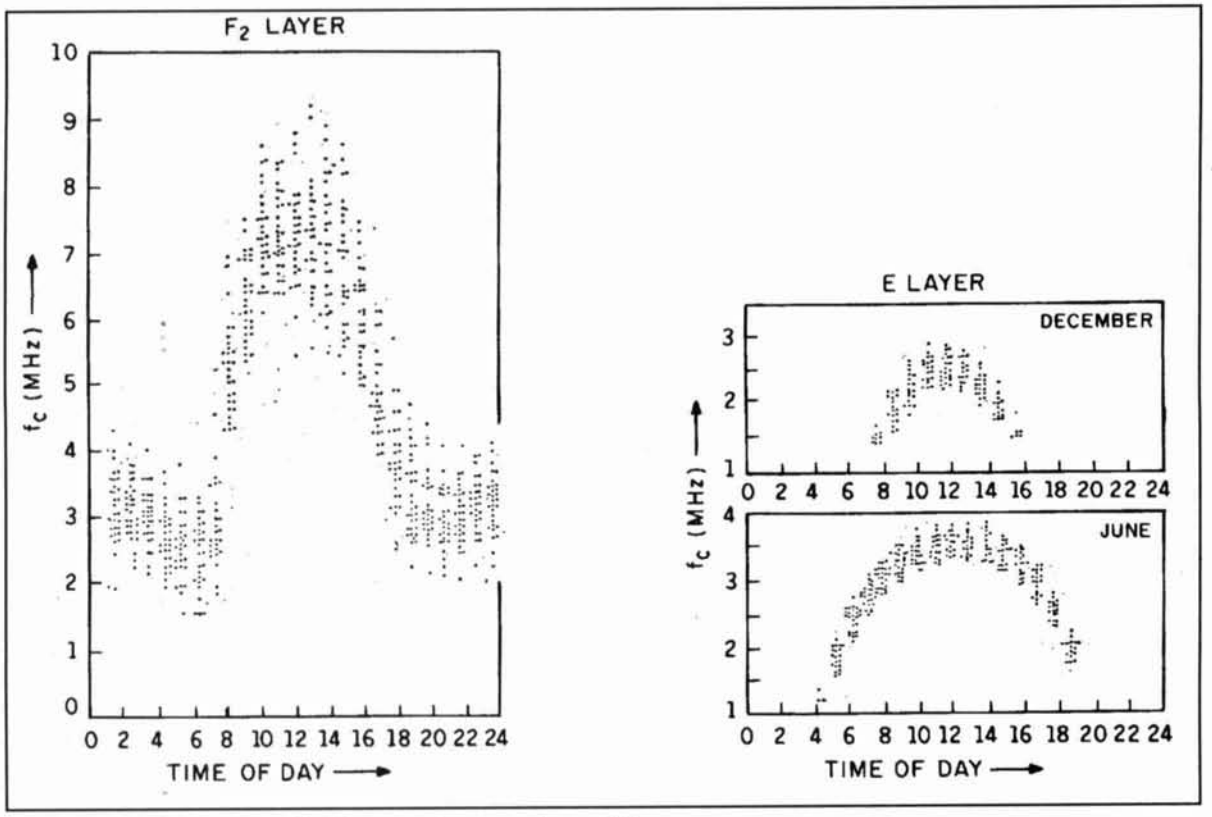

Figure 32. Typical values of critical frequency f_c measured over 24 hours.

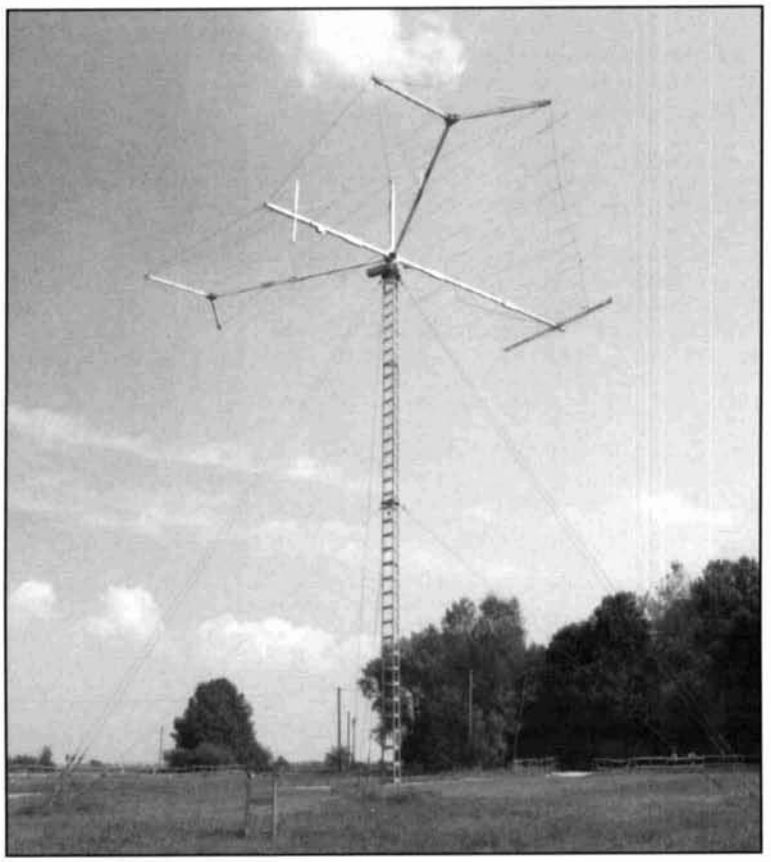

Photo F. Rotatable Log-periodic Antenna AK 853.

family of curves at frequency f_1 as in **Figure 37A**, the probability distribution of the elevation angle can be obtained as shown in **Figure 37B**. It makes sense to fix the maximum of the desired vertical radiation pattern at the departure angle that occurs most frequently. For the 25 and 75 percent values, for instance, a reduction of 1 dB with reference to the maximum may be permitted. The 10 and 90 percent boundary curves can be defined as 3-dB points of the antenna lobe width as in **Figure 37C**. The radiation pattern obtained has been derived purely from wave propagation; up to now there has been no mention of any physical antenna at all.

The possibility of fulfilling such requirements derived from wave propagation is, of course, of interest. **Figure 38** shows the vertical radiation pattern of a horizontal logarithmic-periodic antenna from Rohde & Schwarz, like the one in **Photo F**.[15] The shaded section of the pattern in **Figure 38** corresponds to the elevation-angle range between 10 and 90 percent boundary curves in **Figure 36**. With antennas designed by R&S, the position of the maximum, as well as the width of the vertical main lobe, are matched very accurately to the necessary elevation angle range required by the wave-propagation analysis. Reports on the success of this type of antenna in practical radiocommunications have confirmed the correctness of such an approach.

A horizontal logarithmic-periodic antenna, with appropriate design parameters, can be used in every kind of sky-wave transmission—from high-angle radiation to long-range traffic—as shown by the vertical radiation patterns in **Figure 38**. This feature, in addition to the broadband input impedance and the high efficiency, is one of the main reasons why this type of antenna forms the, technically, most efficient solution to universal applications in the RT and HF bands.

High-angle radiation and its demands on antennas

From the point of view of wave propagation, high-angle radiation is of great importance to a large number of users. The wave must be radiated at very steep angles of elevation so that, after refraction in the ionosphere, it reaches a receiving site located relatively near (e.g., 80 km) the transmitting site (see **Figure 28**). High-angle radio links via the ionosphere can be implemented reliably if one considers that, if the angle of incidence Ψ in **Equation 14** is small, the cosine of Ψ tends to a value of 1. Thus the classic MUF is very close to the critical frequency; i.e., at lower values than are usual for medium and long-range distances.

The second requirement is that the vertical radiation patterns of transmitting and receiving antennas be matched to the application. The majority of failures in high-angle transmission are due to the often repeated attempt to use the "most convenient" radiator; that is, a rod or whip antenna. The reason this is doomed to failure at twilight and during the nighttime is shown in **Figure 39**: the vertical radiation pattern at the steep elevation angle required in this case has a null! While daytime coverage of medium distances is possible via the E layer, a vertical antenna does not radiate enough energy at the steep elevation angle required at night for F_2 transmission because of the greater altitude of the layer. Therefore, simply devised tests and small-scale trials during the day are of no value if the radio link is to be used at different times of the day. Carefully planned trials take this into account, just as they avoid a trial period that is too short in view of the pronounced time dependence of sky-wave propagation.[14]

Shortwave transmission with a vertical antenna always leads to coverage like that in **Figure 40**. Here, a limited area around the transmitting site is covered by ground waves and the depressed vertical pattern enables long distances to be covered well. In between there is a "skip zone," usually several hundred kilometers wide depending on the radiated power, that cannot be reached with a vertical antenna; this

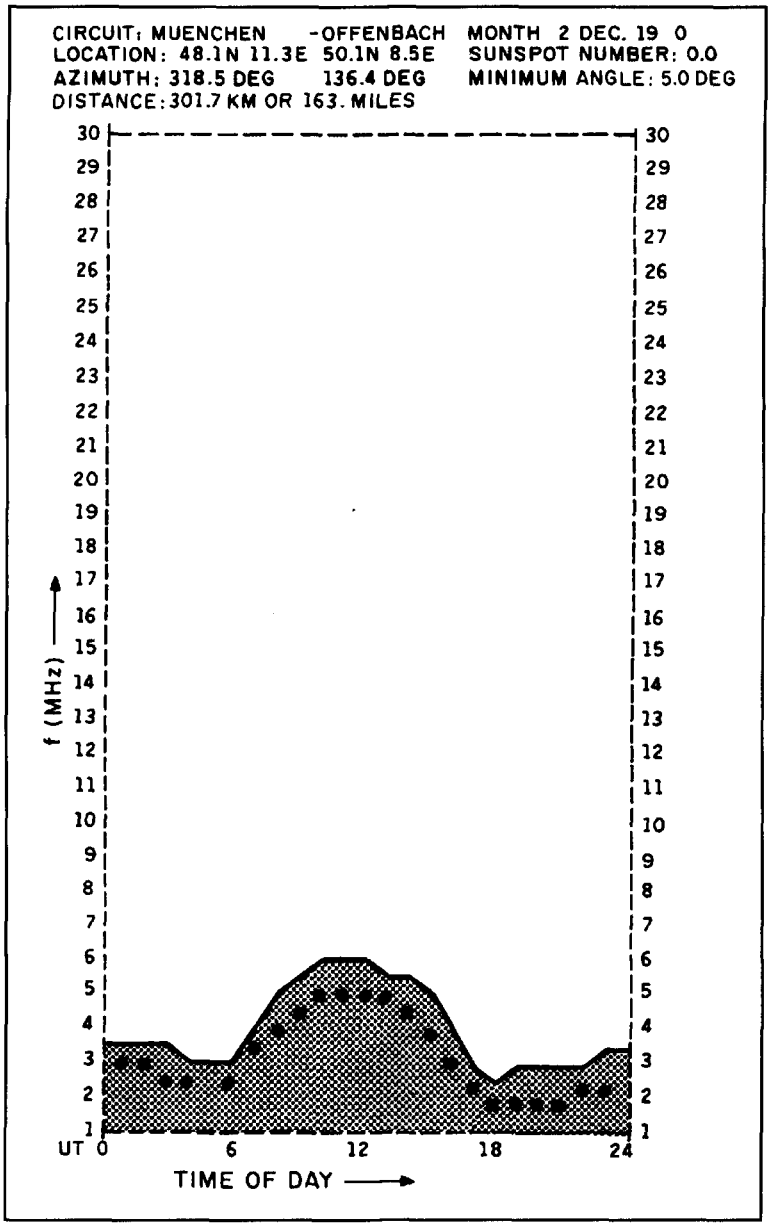

Figure 33. Variation over 24 hours of transmitting frequency range (MUF); optimum working frequency indicated by +.

is valid both for transmission and reception. In this case, a high-angle radiation antenna is required with a three-dimensional radiation pattern as shown in **Figure 41**.

The tunable HF Dipole HX 002 from Rohde & Schwarz with its integrated tuning unit[16] meets these requirements and occupies a minimum of space (**Photo G**). For applications in mobile-radio systems, and to provide a solution to this coverage task, Rohde & Schwarz developed mobile HF Antenna System AK 501,[11] which not only includes the vertical antenna for ground-wave coverage mentioned above but also—and despite relatively small dimensions—a high-angle radiator with full capabili-

CIRCUIT : MUENCHEN - OFFENBACH MONTH : DEC. 19 0
LOCATION: 48.1N 11.3E 50.1N 8.5E SUNSPOT NUMBER: 0.0
AZIMUTH : 318.5 DEG 136.4 DEG MINIMUM ANGLE : 5.0 DEG
DISTANCE: 301.7 KM OR 163. NAUT.MILES BANDWIDTH : 0.400 KHZ
REQ.S/N : 10.0 DB MAN-MADE-NOISE: RURAL

POWER	0.40	0.40	0.40	0.40	0.40	0.40	0.40	0.40	0.40	0.40	0.40	KW
TX-ANT.	0.1	1.0	2.0	2.8	3.5	4.0	4.5	5.0	6.0	6.5	7.0	DB
RX-ANT.	0.1	1.0	2.0	2.8	3.5	4.0	4.5	5.0	6.0	6.5	7.0	DB
	1.6	1.8	2.0	2.5	3.0	3.5	4.0	5.0	6.0	7.5	9.0	FREQU.
UT MUF												
11 6.1												ANGL.RX
-	0.99	0.99	0.99	0.99	0.99	-	-	-	-	-	-	35.2 E
0.50	-	-	-	-	-	0.99	0.99	0.94	0.55	0.01	-	55.8 F
0.50	0.99	0.99	0.99	0.99	0.99	0.99	0.99	0.94	0.55	0.01	-	F.DAYS
34.	-4.	3.	9.	18.	24.	29.	32.	37.	41.	26.	-	DBU
48.	22.	29.	34.	42.	48.	52.	54.	58.	61.	45.	-	SIG.DBU
-1.	10.	9.	7.	4.	2.	0.	-1.	-4.	-5.	-10.	-	NOI.DBU
49.	12.	20.	27.	38.	46.	51.	56.	61.	66.	55.	-	S/N DB
0.50	0.27	0.41	0.54	0.88	0.98	0.99	0.99	0.93	0.55	0.01	-	REL.MUF

Figure 34. Forecast printout for reliability calculations of radio link.

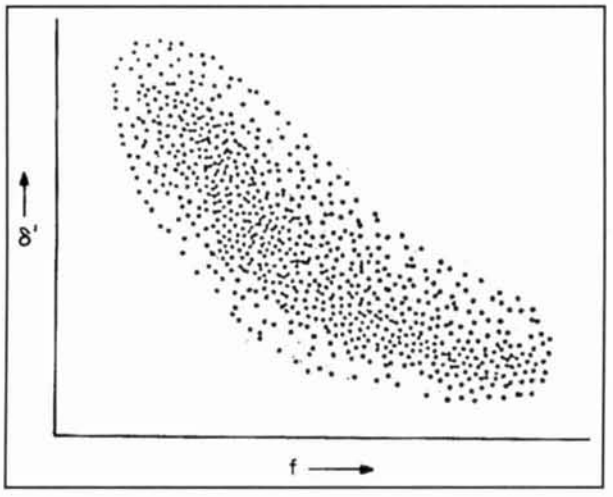

Figure 35. Elevation angles as cloud of dots.

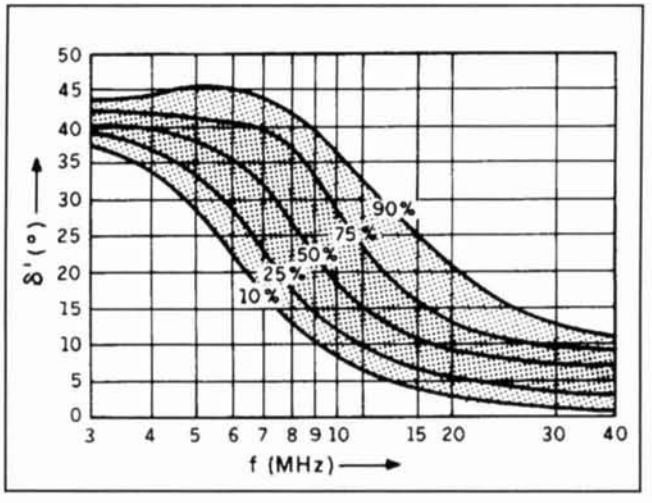

Figure 36. Statistics of elevations angles.

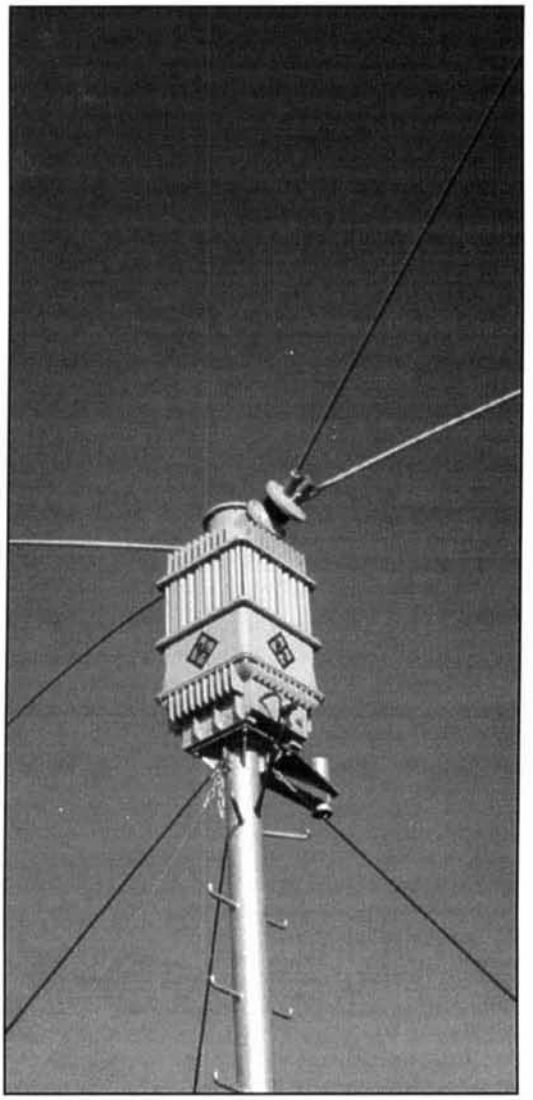

Photo G. HF Dipole HX 002 with integrated 1-kW ATU.

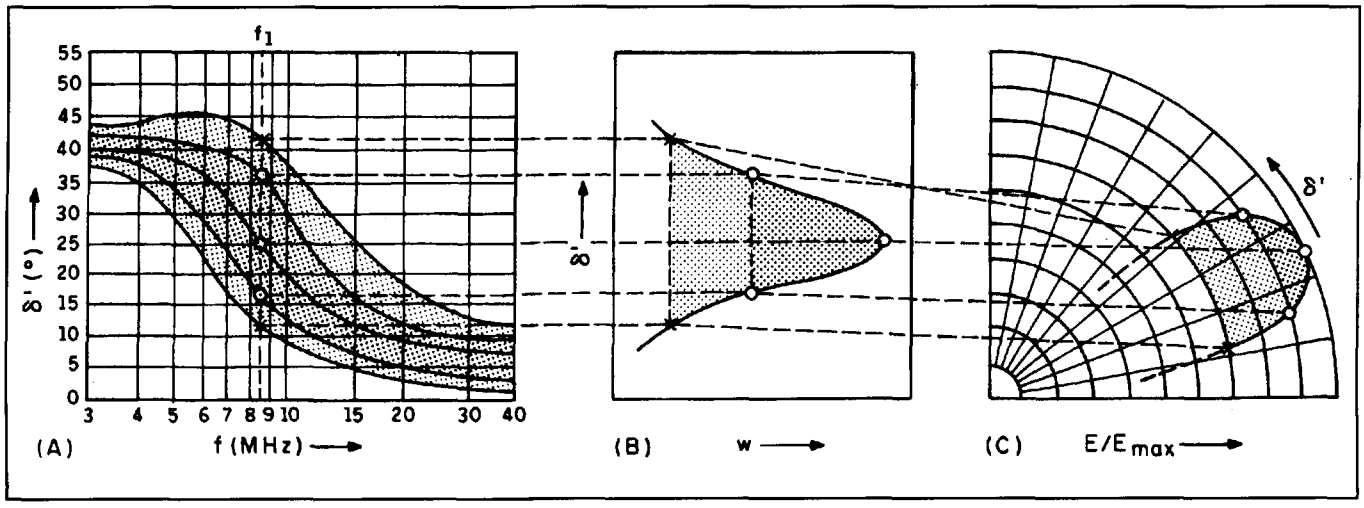

Figure 37. Association between statistics of elevation angles and desired vertical radiation pattern.

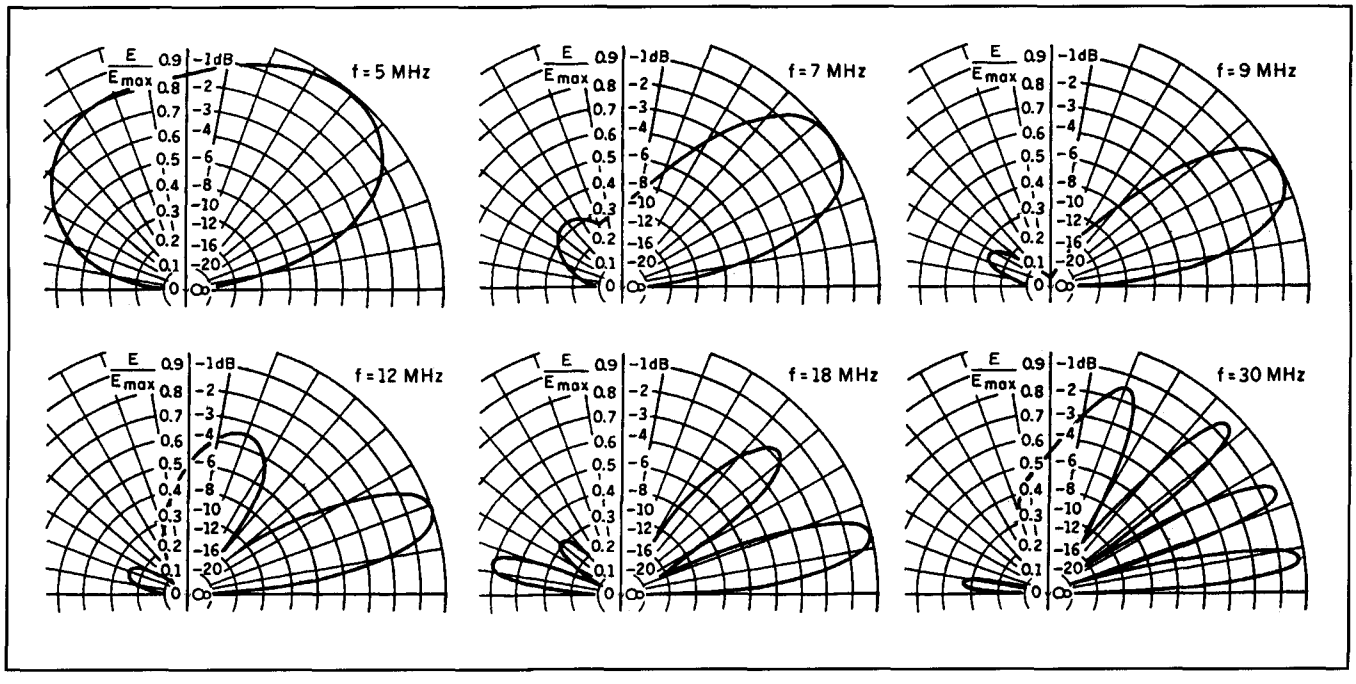

Figure 38. Vertical radiation patterns of horizontal Log-periodic Antenna AK 851 optimized using wave-propagation analysis.

ties[17] as shown in **Figure 42**. This antenna has proved itself in many practical solutions.

Conclusion

This article has shown that electromagnetic-wave propagation has considerable practical importance in all areas of radiocommunication. It is not something that has relevance only for research institutes or universities, but can be used in day-to-day radiocommunications work.■

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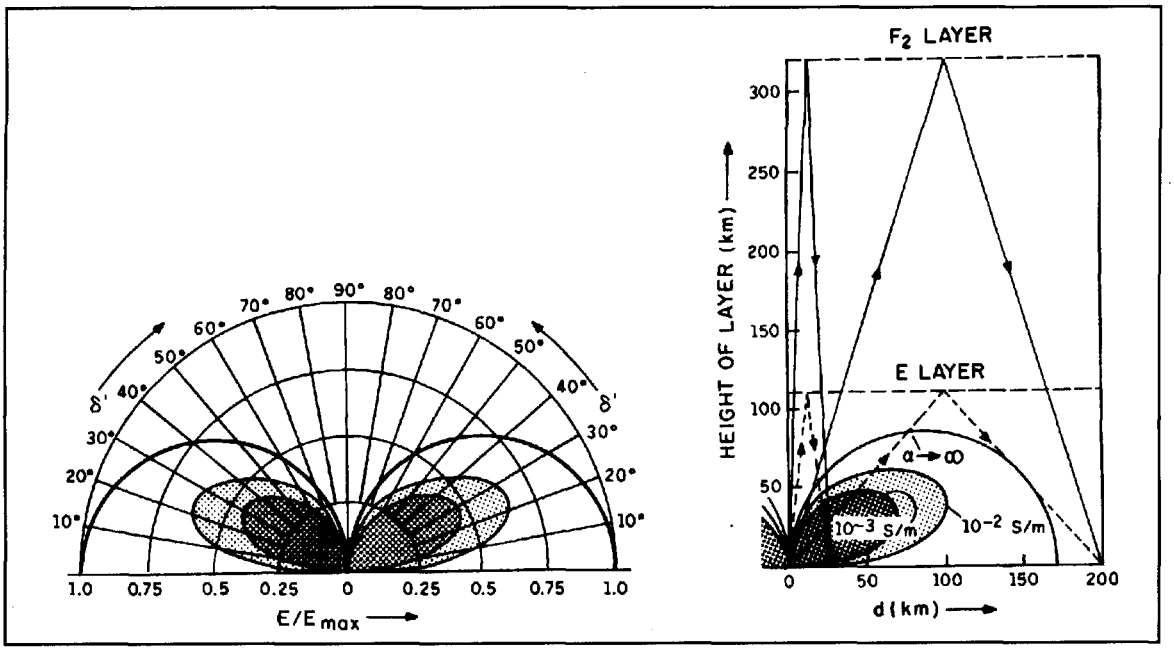

Figure 39. Vertical radiation pattern of λ/4 vertical antenna and transmission path for high-angle radiation.

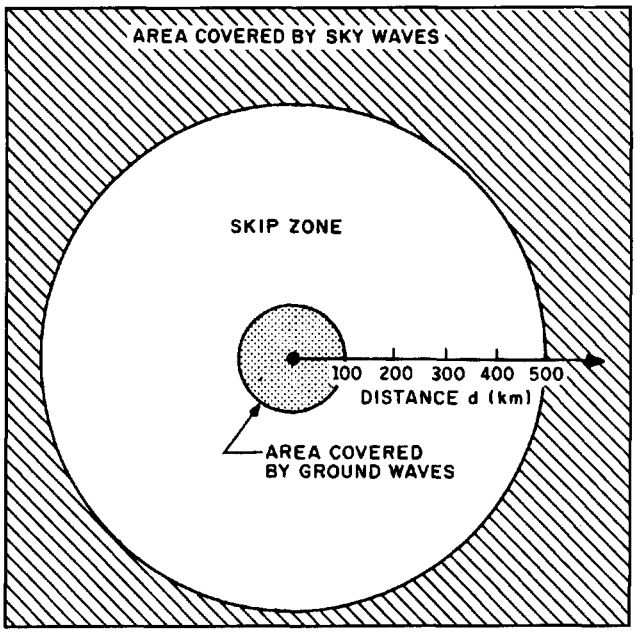

Figure 40. "Skip zone" in shortwave coverage with vertical antenna.

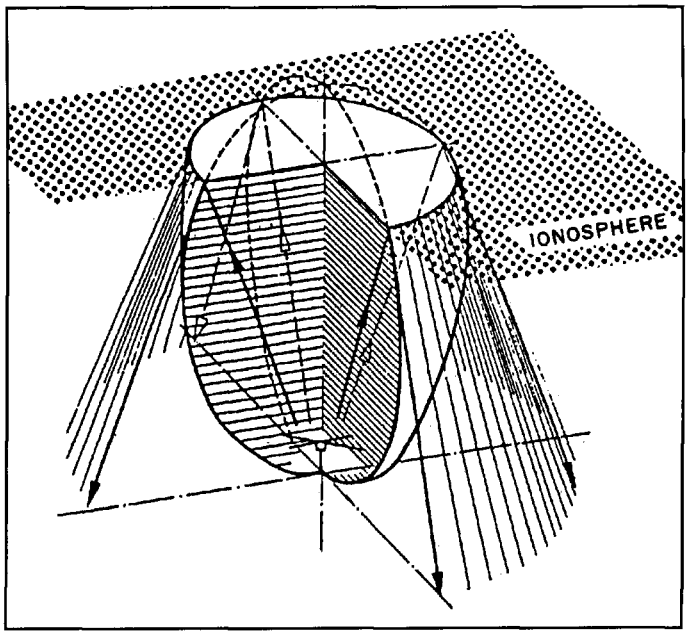

Figure 41. Coverage without gaps using special high-angle antennas.

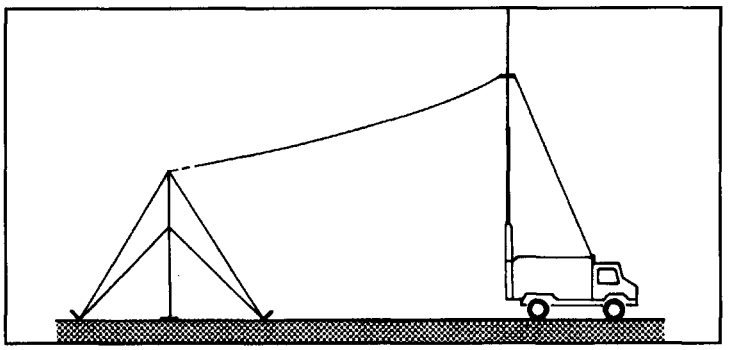

Figure 42. High-angle radiator (running from insulator on the left, up to antenna, and down to truck) of Antenna System AK 501 for mobile operations.

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THE SOLAR SPECTRUM

A portable VLF receiver and loop antenna system

In one way or another, most of us are familiar with that old bugaboo—local interference. Depending on its signature and the type of measurements that are affected, various methods have been devised to seek out the source of such noise. Recently, our colleague and collaborator in Koniz, Switzerland, Dr. Walter S. Moos, encountered this problem while recording ionospheric disturbances caused by solar flares. I've asked Walter to describe the details of the portable VLF receiver system he developed in an attempt to resolve the situation. (Text, graphics, and photographs courtesy of Walter Moos.) Walter:

A.J. Stokes and P.O. Taylor have published several papers on VLF receiver systems which are suitable for the detection of sudden ionospheric disturbances.[1,2,3] Since I have monitored various other atmospheric parameters for many years, it was a simple matter for me to include SID measurements in my program. My first receiver was constructed in close adherence to the above authors' instructions, and in fact, used some components which they supplied. Their help in this regard is herewith gratefully acknowledged.

Unfortunately, after a rather successful beginning, considerable difficulty arose when an unknown source of disturbance interfered with the measurements, in the form of serious rhythmic excursions of the recordings at certain times during the day. The interference displayed about 5 to 10 second pulsations with an amplitude three to ten times the normal signal level, superimposed on a 19.6 kHz (GBZ) trace. Thus, a good portion of the recordings were rendered unusable.

Efforts to locate the source of this noise by the Swiss equivalent of the U.S. Federal Communications Commission, the electric power company, and the author are still underway. Thus far, a location determination by triangulation has been attempted with the authority's car-mounted equipment, which however, has proved to be too heavy for a search of off-street locations. Moreover, little such equipment is available nowadays, since interest in VLF radio has dwindled in Switzerland.

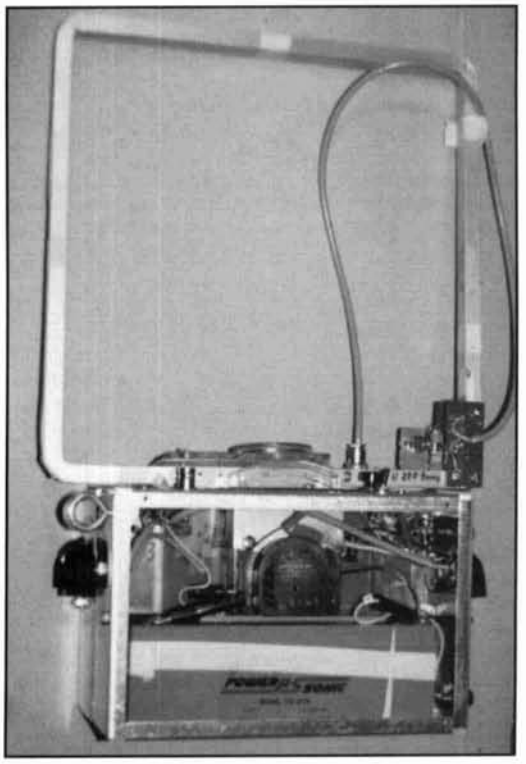

Photo A. The Moos portable VLF receiver/loop antenna system.

To facilitate this search and also for eventual

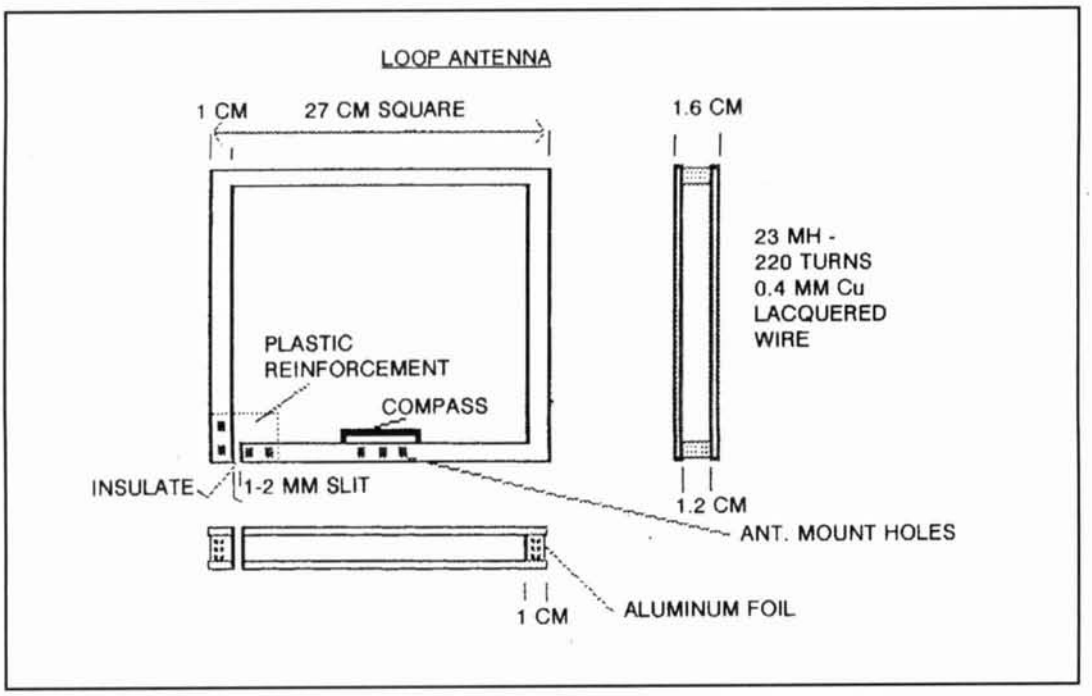

Figure 1. Loop antenna schematic and details.

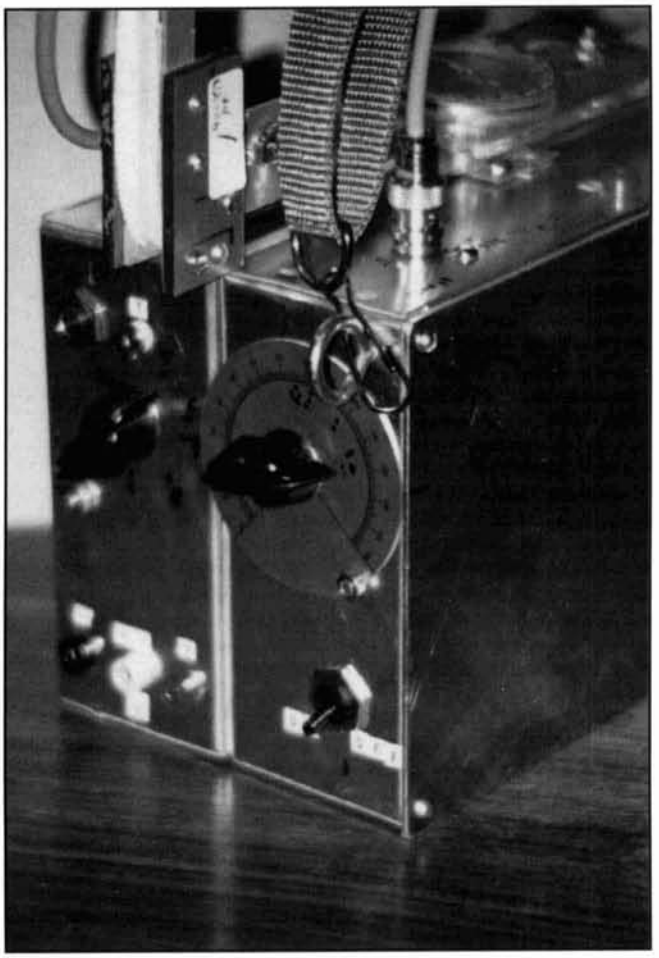

Photo B. Potentiometers, various jacks, etc., installed in the combined box unit described in text.

use as an additional VLF receiver, I constructed a unit that is portable and battery operated (**Photo A**), allowing the system to operate uninterrupted for a week or more. The basic construction details follow.

Loop antenna

A 1-meter section of aluminum "U-Profile" with 1-inch centimeter wide sides was purchased at the local hardware store. The Profile was formed into a U-shaped frame about 25 by 25 centimeters, with the open side facing outward. Aluminum Profile can be bent and formed quite easily by carefully heating the area to be bent to a temperature that will just begin to char a match stick. The forming must be carried out slowly, with the heat continuously distributed over the surrounding area out to about 5 centimeters away from the bend itself. During this process, it is helpful to clamp the remainder of the Profile into a vice, insulating it with some odd pieces of ceramic.

The frame must be joined in such a way that a thin opening—perhaps 1 to 2 millimeters wide—is obtained where its ends meet. Remove all sharp edges and burrs with sandpaper, and cover the opening with insulation tape or good quality plastic tape. Next, drill (3 millimeters) and tap both sides and legs of the U-shaped opening, and prepare two 2 to 3 millimeter thick plastic plates which can be screwed to the juncture. The plastic squares serve as reinforcements that retain the shape of

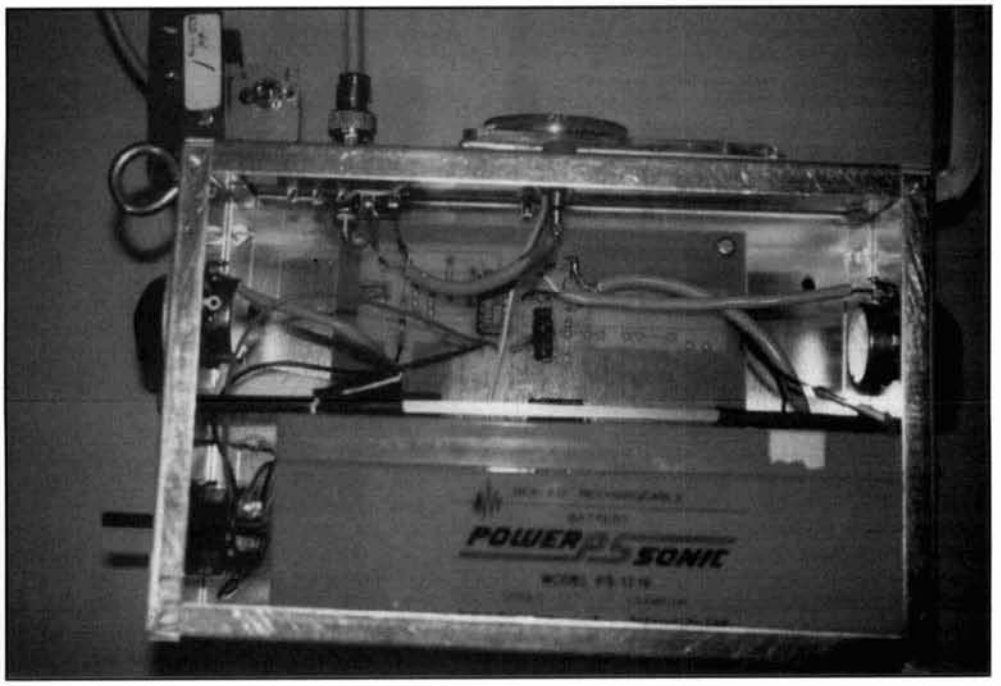

Photo C. The Panasonic battery installed in the cabinet.

the antenna frame. At the same time, drill two additional holes to allow the antenna wire ends to be attached along with a shunt condenser across the wire coil.

To construct the coil, I used 0.4 millimeter diameter lacquered copper wire. After the winding is completed within the frame, the Profile's open side is enclosed with ordinary kitchen aluminum foil held in place with a good self-adhesive tape. I used 220 turns of wire, resulting in 23 mH. At the bottom-center portion (antenna rotating center), I mounted an

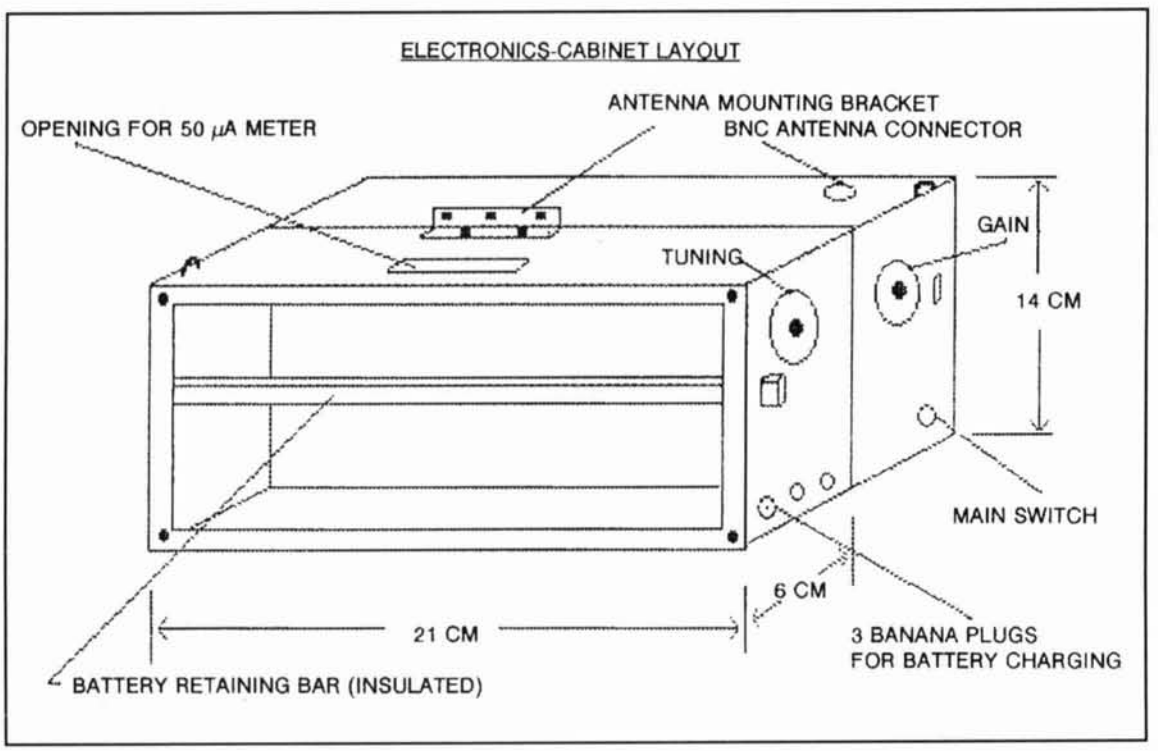

Figure 2. General electronics layout.

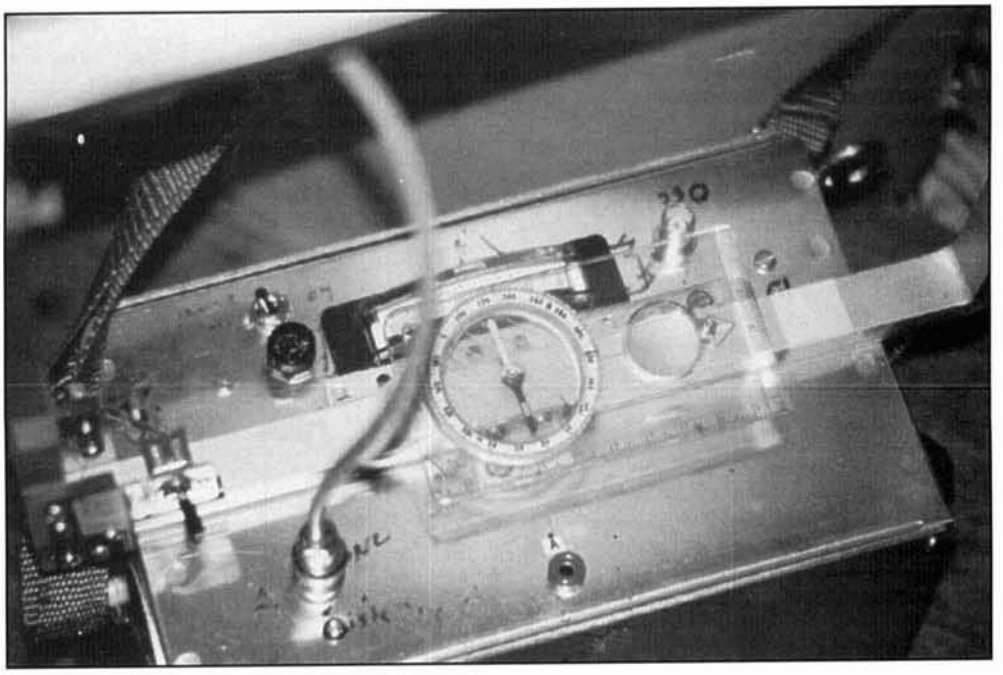

Photo D. Details of the antenna mount, meter, and compass.

old plastic Boy Scout compass, which aids in the precise location of a VLF station or interference source. The antenna is connected to the rest of the circuit via a short length of shielded cable and BNC connector. Details are shown in **Figure 1**.

VLF receiver and preamplifier

Two "gyrator" style printed circuit boards (see **References 2** and **3**), with the diodes and capacitors of the rectifier portion removed, are employed in this design. I purchased two standard, commercially available aluminum boxes (each 21 x 6 x 14 centimeters) and attached them back-to-back. Appropriate openings for banana plugs, potentiometers, main power switch, etc. were drilled into the combined unit as shown in **Photo B** and diagrammed in **Figure 2**.

The 3-millimeter brass bars serve only to hold the batteries in position, but should be insulated to avoid possible short circuits. I use

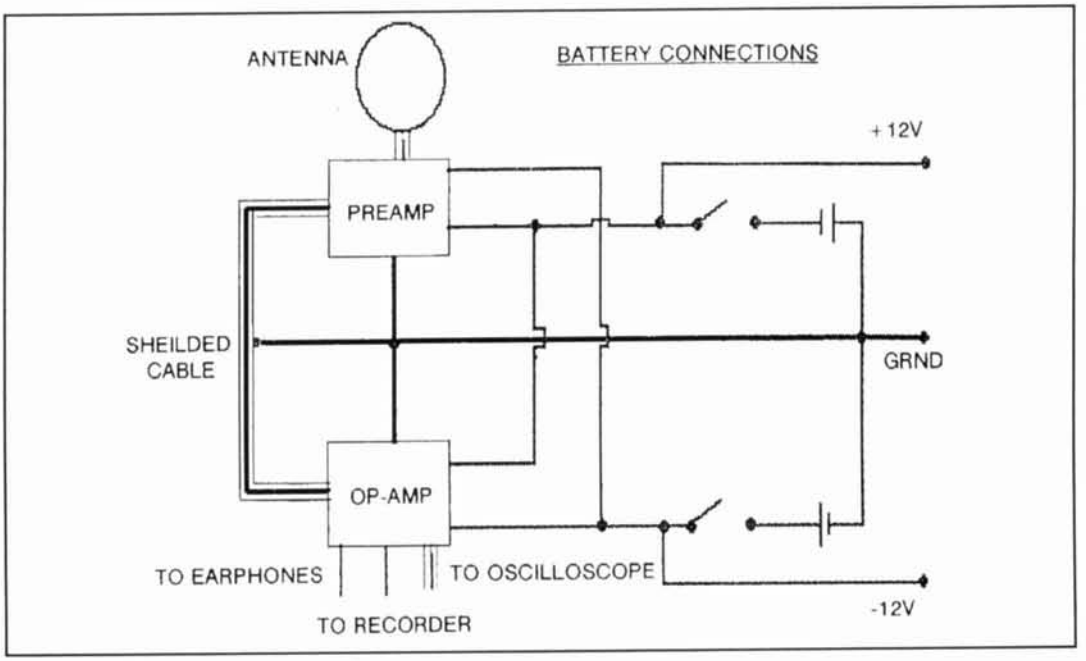

Figure 3. Battery connection schematic.

Panasonic "Power Sonic" 12-volt batteries (Model PS 12 19), but others of similar dimensions and capacity will serve as well (**Photo C**).

On top of the housing, I used old "Mechano" corner pieces to mount the antenna, and made an opening into one box large enough to install a flat-faced, 50-μA meter (**Photo D**). Three banana plugs serve for connecting the batteries to a charging device, with the main switch set in the off position (**Figure 3** - left). An external trickle charger could also be connected for longer-term operation.

Another helpful addition provides a connection for small earphones; the ears are very sensitive when locating a station, and for determining signal strength while turning the unit to the optimum direction. One can also add another BNC connector to pin 12 of the operational amplifier and follow traces on an oscilloscope. A jack-type connector allows for attaching the unit to an external strip-chart recorder using a short section of cable with the built-in meter set to off.

It is very important that the connection between the preamplifier housed in the one box and the second op amp input is made only with well-shielded cable, and that all ground connections are adequately secured. Any openings between the two boxes should be kept to a minimum size.

Finally, two hooks installed in the unit make it possible to attach a carrying strap of convenient length. I checked operation using a borrowed two-channel strip-chart recorder, tracing the signal from GBZ simultaneously with my stationary receiver and the portable unit. Of course, a small portable computer—perhaps a battery-powered laptop—could be equipped with an A-D converter and used to record the data while "on the road." To date, I have been able to tune the portable receiver to the following stations from my location in Switzerland:

kHz	Location
16.8	France (FUB)
19.0	England (GQD)
19.6	England (GBZ)
21.4	USA (NSS)
24.0	USA (NAA)*

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Switzerland

Change of Internet address for NASA lightning project

In the last issue, we called attention to Marshall Space Flight Center engineer Otha Vaughan's request for readers with VLF monitoring stations to participate in the collection of data related to the Mesoscale Lightning Experiment, and highlighted the World Wide Web address for material related to that project. Shortly thereafter, the URL was changed to:

http://wwwghcc.msfc.nasa.gov:5678/

Internet users who are interested in these phenomena should use the new address. Various means of contacting Otha directly appeared in the original column and remain the same.

In a separate but related development, NASA recently announced that researchers have captured on videotape the first conclusive evidence that the mysterious flashes of red light called "red sprites," which extend up to 90 kilometers above electrical thunderstorms, are not limited to the United States.

A research team from the Geophysical Institute of the University of Alaska, Fairbanks, recorded the unusual flashes above thunderstorms near the equator in South America last February and March. Previously, they had seen the recently discovered sprites above storms only in the U.S., leading some scientists to question whether or not they occur in other parts of the world.

Geophysical Institute researchers D. Sentman, G. Wescott, and D. Osborne used special low-light-level cameras aboard a Westwind-2 jet aircraft to record the brief flashes. The flights—part of a NASA-sponsored investigation into the phenomenon—were coordinated with the Peruvian Air Force.

In form and in visual appearance, the sprites over South America look similar to flashes recorded by the team over storms in the central U.S. last summer. About 500 sprites were recorded last June and July, many on color video for the first time. None of the sprite groups seen this winter over South America were as large or as intense in color as some of the larger groups recorded over the U.S.

Less intense thunderstorms may have contributed to the smaller number and desultory appearance of the sprites above South America. In the southwestern-central U.S., the storms form along a quickly moving frontal system, but the convective storms in South America are nearly stationary; they tend to grow in place, develop slowly into large systems, and then dissipate.

As Otha reported in the spring issue, airline pilots and others also have reported seeing blue or greenish columns propagate upward at great speed from the top of the thunderstorms. ■

*VLF station NAA (24.0 kHz, Cutler, Maine)—off-air for a number of weeks—appears to have resumed operation.

Westcott and Sentman were the first to report the video capture and characteristics of these "blue jets" from 1994 flights over the U.S. However, no blue jets were observed over South America.

Under the right conditions, sprites can also be seen from the ground after dark with the unaided eye. To encourage observers to report sprite sightings around the world, Sentman is also establishing a Sprite Watcher's Homepage on the Internet's World Wide Web. The homepage will give brief information about sprites, the conditions needed to view sprites from the ground or air, and simple directions to follow when recording an observation. All public sightings will be incorporated into a scientific database, and then displayed on a global map for Web users. Individuals interested in participating in this facet of lightning research can access the Sprite Watchers Homepage at the following URL:

http://elf.gi.alaska.edu/

Researchers from government laboratories, universities, and Federal agencies will continue their investigations of sprites and similar thunderstorm phenomena this summer. The study will be undertaken using radio frequencies, radar, and other measuring techniques. Again, we urge VLF radio enthusiasts to contact Otha Vaughan at MSFC for details on participation in this endeavor. ■

REFERENCES

1. P.O. Taylor and A.J. Stokes, "Recording Solar Flares Indirectly," *Communications Quarterly*, Summer 1991.
2. "A Shielded, Tuned-loop Antenna with Pre-Amplifier," *SID Technical Bulletin*, Volume 5, Number 2, April 1994.
3. A.J. Stokes, "A Gyrator Tuned VLF Receiver," *Communications Quarterly*, Spring 1994.

PRODUCT INFORMATION

Telex Announces New Digital Rotator Control Unit

Telex Communications, Inc. has introduced a new Digital Control Unit for Hy-Gain rotators. The DCU-1 Pathfinder is designed to be used with the Ham IV and T2X Tailtwister and is also backwards compatible with any 8-wire Hy-Gain rotator such as the Ham II or Ham III.

The DCU-1 Pathfinder incorporates features such as digital bearing readouts to 1 degree, motor slowdown and eight-second automatic brake delay. It also has automatic calibration and selectable center of rotation

According to Telex, the two "most exciting" features are the six user programmable memory presets and RS-232 computer control capability. The six programmable presets allow the user to pre-program their six favorite beam headings. The RS-232 computer compatibility of the DCU-1 Pathfinder will allow rotator control from the user's computer as well as having a serial port pass-through capability that allows the user to also connect their radio, terminal TNC, modem, etc. to this serial port. A 60-line Basic program for rotor control is included in the manual and a free package of information for application software developers is available from the factory.

The DCU-1 and rotator are available in both 110 and 220 VAC versions, operating on 50/60 Hz.

Interested amateurs should contact their favorite Hy-Gain dealer for price and availability information. The manufacturer's suggested retail for the T2X-D (DCU-1 and T2X Tailtwister) is $799.99; the Ham IV-D (DCU-1 and Ham IV combination) is $749.99; and the DCU-1 by itself is $519.60.

AEA's New KKCOM Package

Advanced Electronic Applications, Inc. has introduced an editing and control accessory for the KK-1 Keyboard Keyer. The KKCOM package consists of application software and a serial interface cable which connects the KK-1 to a PC computer.

AEA's KKCOM gives users the ability to compose and edit buffer contents right on the computer screen. Keyer parameters such as speed, sidetone pitch, serial number, and paddle sense can also be displayed and changed from the KKCOM software. KKCOM also allows the contents of the KK-1 Keyer, including messages and settings, to be quickly saved to a file and reloaded later. Customizing the KK-1's battery-packed message buffers is made easier using the KKCOM. Create and save one buffer or any number of buffers up to twelve. Once buffers are loaded into the KK-1 they will stay in the battery-backed buffer memory until you decide to change them again.

For more information, contact: Advanced Electronic Applications, Inc., P.O. Box C2160, Lynnwood, WA 98036 (phone: 206-774-5554; fax: 206-775-2340).

Rick Littlefield, *K1BQT*
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QUARTERLY DEVICES

Build dummy loads and resistive RF networks with these power-film Resistors

Most of us would like to have a low-cost source for 50-ohm transmitting loads—for the lab, service shop, or for use in new products. However, lab-quality loads that handle moderate power over a wide frequency range are usually anything but low in cost! Building your own loads can also be tricky, especially at VHF and above. But, all is not lost! In this edition of "Quarterly Devices," we'll look into a family of heat-sinkable film resistors from Caddock Electronics, Inc. that may offer a convenient solution.

Packaging

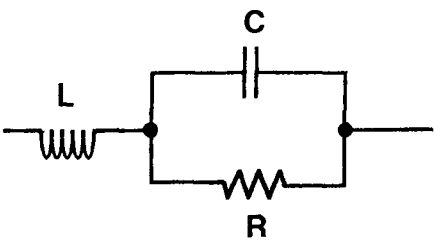
The elimination of heat is one of the toughest problems associated with the construction of RF loads from axial resistors. Stuffing a round resistor body into a tight cylindrical heatsink introduces unwanted capacitance. Distributing power over several resistors introduces parasitic inductance. Immersing a resistor in liquid coolant introduces chemicals and expensive packaging. It seems like there's no simple way to get the job done without a compromise!

Rather than struggling with axial-lead packaging, Caddocks' solution to the cooling problem involved changing the form of the resistor itself—into a plastic tab-mount package. This approach has been used successfully for years by the semiconductor industry to encase low-cost power transistors and regulators. Inexpensive cases like the TO-220 can dissipate up to 50 watts of power when used in conjunction with a good heatsink, and will support devices operating well into the VHF region.

The Caddock Electronics line of tab-mount devices are called Kool-Tab® Power Film Resistors. These come in several power ratings, with resistances ranging from 0.02 ohms to 10 k. This broad selection makes Kool-Tab

Resistor Style	MP 800 Series Model No.	Power Rating at +25°C Case Temp	Resistance Range	Package Description
TO-126 Style Package	MP 808	8 Watts	0.02Ω to 10K	Thermally Conductive Molded Package
	MP 825	25 Watts	0.02Ω to 10K	Copper Heat Sink Integral in the Molded Package
TO-220 Style Package	MP 816	16 Watts	0.10Ω to 10K	Thermally Conductive Molded Package
	MP 850	50 Watts	0.20Ω to 10K	Copper Heat Sink Integral in the Molded Package
TO-220 with Metal Mounting Tab	MP 820	20 Watts	10.0Ω to 10K	Metal Mounting Tab
	MP 821	20 Watts	0.02Ω to 9.99Ω	Metal Mounting Tab

Table 1. Available wattage ratings for the MP 800-series devices.



Model	Configuration	Equivalent L	Equivalent C	VSWR = 1.1*
MP 820-50Ω	Isolated tab	7 nH	1 pF	>150 MHz
MP 820-50Ω	Tab grounded	2.3 nH	1.5 pF	>400 MHz
MP 850-50Ω	Isolated tab	8 nH	1.7 pF	>200 MHz
MP 850-50Ω	Tab grounded	5.5 nH	4.5 pF	>100 MHz

* varies with resistor value

Figure 1. Equivalent circuit for MP 800-series resistors at RF.

devices useful for a wide variety of applications. However, I'll focus primarily on their use in RF-power loads.

Kool-Tab MP 800-series devices

Kool-Tab devices were first developed in 1988 to provide a reliable high-power density resistor in a familiar and widely used package. They are rated in the same manner as TO-126 and TO-220 semiconductors, and use the same off-the-shelf heatsinks, thermal interfaces, and hardware. You can even use the same formulas for calculating heatsink size and power level to ensure safe operation.

As shown in **Table 1**, MP 800-series devices are available in several wattage ratings. The MP 808 and MP 816 (8 and 16 watts) are encased in all-plastic packaging. In the MP 820, MP 821, MP 825, and MP 850, the resistor element is bonded to an internal copper plate that becomes the metal mounting surface of the device. This spreads the heat and ensures low thermal resistance.

MP 820s and MP 850s are often selected for RF applications because of their low parasitic inductance and capacitance, high power dissipation, and reliable performance at high frequencies. All MP 800-series devices offer 1500-volt isolation between the resistor element and mounting surface. Kool-Tab resistors are normally specified at ± 1 percent tolerance. However, tolerances ranging from 0.5 to 20 percent may be available—or may be manufactured—by special order.

RF characteristics

The RF characteristics of MP 800 series resistors permit their use in 50-ohm RF loads at frequencies up to 500-MHz and above. An equivalent circuit for MP 820s and MP 850s falling between 10 and 100 ohms is shown in **Figure 1**. This circuit assumes a lead length extending 0.2-inches from the body of the resistor and a F_O (frequency of operation) of 500 MHz and below. According to Caddock Electronics application notes, additional lead length will increase para-

	MP 820	MP 850
No heatsink	2.25 W	2.25 W
Bolted to PC board (maximum 110°C)	1.70 W	1.70 W
Small clip-on (Thermalloy 6094)	2.9/3.1 W	2.9/3.1 W
Small finned HS (Therm. 6038 or 6072)	6.3/8.5 W	6.5/11 W
Large extruded HS (Therm. 6098)	6.5/7 W	7.5/12 W
5" x 7" x 2" Aluminum chassis	14 W	25 W

Table 2. Safe operating levels at 25 degrees C with free-air convection or 100-LFM (linear foot per minute) air flow. This data assumes adherence to prescribed mounting practice

sitic inductance at the rate of approximately 1 nH per 0.1 inch. In some applications, stray lead inductance may be used to tune out parasitic capacitance, yielding improved high-end VSWR. The formula $L=R^2C$ provides a guideline for optimum inductance. Caddock Electronics claims that 50-ohm values of the MP 820, MP 825, and MP 850 will exhibit very low VSWR through 500 MHz. Some customers report using these devices in RF-power applications as high as 1-GHz with careful packaging and minor compensation for parasitic reactance.

Power handling and cooling

Caddock Electronics applications literature provides detailed rating information for continuous duty and intermittent high-power pulsed operation—along with derating curves for ambient temperature, heat-sink selection charts, and other useful formulae and data relating to thermal performance. Consult these notes any time you are engineering MP 800 series devices into critical applications. If you have specific questions that go beyond the scope of these reference materials, Caddock will also provide in-house engineering assistance and product support.

Be sure to heatsink RF loads generously—TO-220 and TO-126 devices handle only a small fraction of their rated power without external cooling. Sufficient cooling depends on two important factors: the size of the heatsink, and the temperature of the air surrounding that heatsink. As a rule of thumb, the guidelines that are presented in **Table 2** will help you determine a safe average power level—at room temperature—for the type of sink you're using. When the ambient air temperature around the heatsink rises above 25 degrees C (room temperature), a linear derating to 0 watts at 150 degrees C will apply.

In addition to thermal considerations, the mode of RF transmission may impact upon rating. Safe power levels for continuous-carrier modes like CW, FSK, AM and FM, can be based upon the device's published DC rating. For example, if your load uses a 20-watt resistor, you can base RF power-handling on this wattage—and then derate as needed to compensate for elevated ambient temperature or for limited heatsink capacity. For SSB modes, you can multiply you load's CW rating times 1.4 to determine a safe rating in Watts-PEP. Ratings for pulsed signals or data modes employing amplitude shifts should be based upon average transmitter output power. However, keep in mind that extremely high amplitude pulses may damage or destroy the resistive element without physically overheating the case—or without exceeding the load's average power rating.

F_O (frequency of operation) may also be a factor in rating, due to skin effect—the tendency of RF energy to concentrate on the outside of the resistive element at very high frequencies. Tests indicate there's no need to derate MP800 series resistors for skin effect below 500 MHz, but that some derating may be required above 500 MHz.

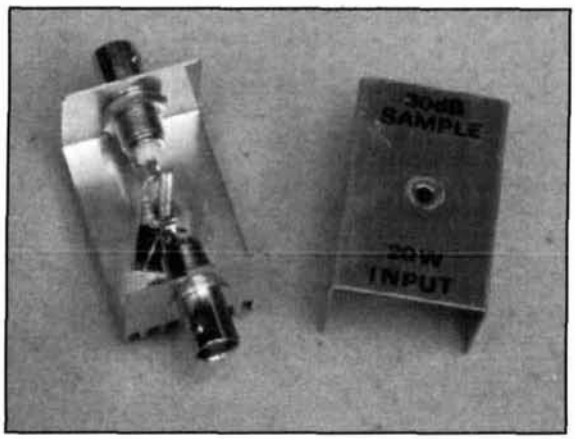


Photo A. Fifty-ohm dummy load using an MP 820 resistor.

In the end, case temperature is your best indicator of safe long-term operation. As a rule, if the resistor's case heats above 100 degrees C, this is a positive indicator that your load either needs more heatsink area—or requires derating to a more realistic power level. This is not to say you can't exceed the load's CW power rating for short periods of time, but use caution! Most MP 800-series devices will tolerate a 50 percent overload for up to 5 seconds, and the MP 820 and MP 850 will tolerate a 100 percent overload for the same time period. If you should overheat a load while conducting tests, allow it to cool down—then check DC resistance before reusing. If the device's resistance has increased measurably in value, replace it.

Mechanical mounting considerations

Careful mounting is especially important when installing a TO-126 or TO-220 device. Here are some guidelines to ensure best cooling performance:

- Deburr the mounting hole and confirm the surrounding surface area is totally flat.
- Mount devices with machine screws only. Sheet metal screws can burr the mounting hole during installation.
- Cover the tab's thermal contact surface with a light even coating of zinc-silicon heat sink compound (do not over apply).
- Use a compression-style washer over the tab—preferably a star or conical (Belleville) type. This will distribute downward pressure

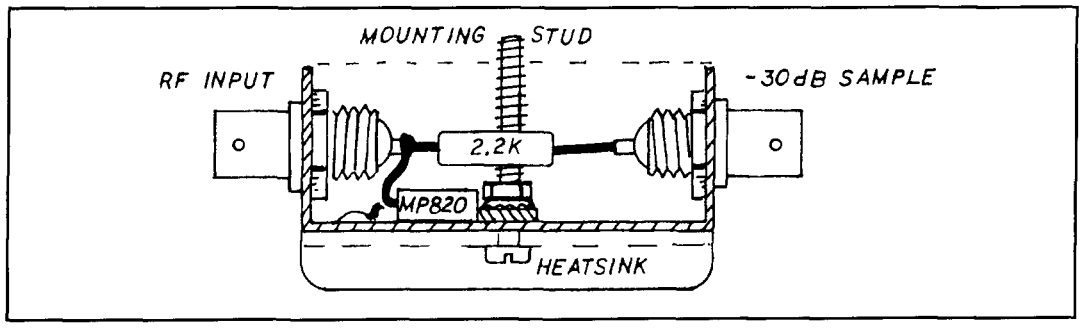


Figure 2. Layout of MP 820 dummy load.

evenly. Do not use a split-ring lock washer.
• Do not overtighten—this can distort the metal tab or fracture the thermoplastic case. Undertightening causes poor thermal contact. Use 6 to 8 in-lbs of torque for best results.
• If you use a thermal pad, choose one with favorable UHF dielectric properties—such as mica.

Building an RF load

In order to gain first-hand experience with these devices, I built a small load using a MP 820 sample provided by Caddock Electronics (see **Photo A**). To make the load more useful, I included a –30 dB resistive sampling tap to provide a means for safely connecting a frequency counter or spectrum analyzer.[1]

To encase the load, I laid out a pattern for a 3/4 x 1-1/8 x 1-3/4 inch U-box on a scrap of copper stock. I then cut it to size, drilled holes for two BNC connectors, and bent the box into shape. After shaping, I drilled a 1/8-inch hole directly in the center to mount the resistor and heatsink. At the same time, I cut and bent a lid to match the U-box. For added cooling, I cut a small aluminum heatsink from a scrap of extruded stock. I then drilled a 1/8-inch hole in the center of both the box lid and heatsink extrusion for mounting purposes.

To install the MP 820, I aligned it with the mounting hole, and soldered one resistor lead to the box base. I then coated the resistor tab and heatsink base with heatsink compound and secured both items in place using a single 1-1/8-inch no. 4-40 screw. Next, I mounted the two BNC jacks and connected the remaining 50-ohm resistor lead to the center pin of the closest jack (this became the RF-input port). Finally, I connected a 2.2-k 1/4-watt resistor from the RF-input jack to –30 dB sample jack (see **Figure 2**). Construction took less than an hour from start to finish.

Testing

To test the load initially, I borrowed the facilities at Austin Antenna Company. I checked the load on an HP8753C Network Analyzer. The SWR and Smith Chart printouts are shown in **Figures 3A** and **B**. I then checked the sample-tap, using a precision signal generator and spectrum analyzer. The sample port measured –31 dB below the input, and sample amplitude remained virtually flat from 10 to 500 MHz (the range of my generator). Finally, I connected a 50-MHz transmitter and applied a 10-watt carrier to the load. After a prolonged period of time, the heatsink and case were only mildly warm to the touch—suggesting safe operation at that power level. Some RF leakage was detected, but this was purely a function of my casual box design. For a more RF-tight enclosure, the lid should be bonded to the box at some point along all six contact surfaces.

Other options

In addition to building simple single-resistor loads, you can also use MP 800 series devices to meet other RF-design objectives. For example, you can construct larger loads by using two 100-ohm MP 850s mounted end-to-end and connected in parallel. With sufficient heatsinking, this configuration should handle up to 100 watts RF power with only minimal degradation in high-frequency performance.

You could also construct a variety of 50-ohm in-line RF power attenuators using either a pi- or T-network configuration. For best performance, mount attenuator devices close together on a copper plate to minimize parasitic inductance. The plate will serve as a heat spreader, evenly distributing heat over the base of the heatsink.

Finally, MP 800 devices should work well in manufactured products requiring built-in resistive networks or terminations, including hybrids, diplexers, and transverters.

Conclusion

For low-to-medium RF power loads and

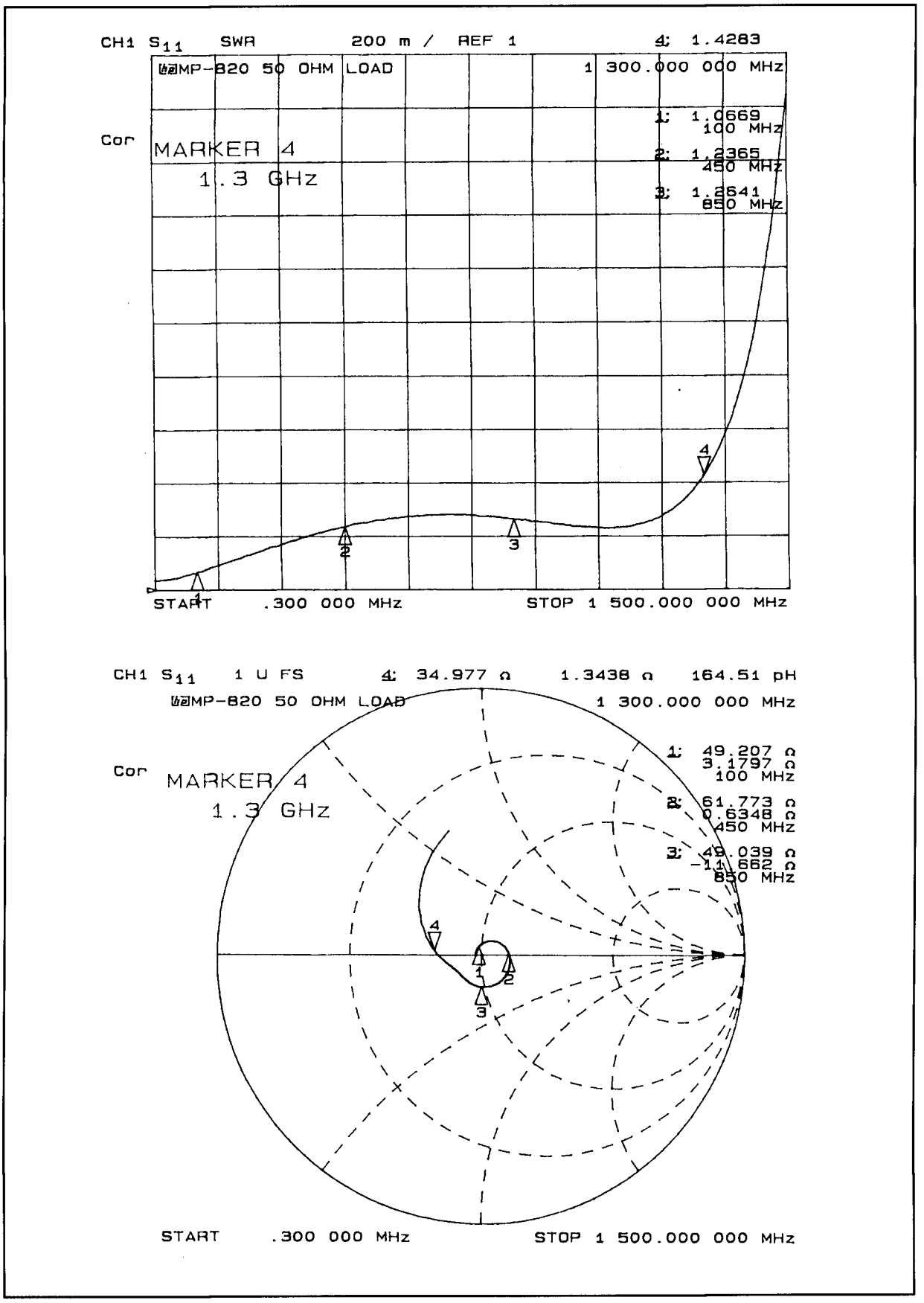


Figure 3: (A) VSWR plot of dummy load made with MP 820 resistor. (B) Smith Chart plot of load. Tests courtesy of Austin Antenna Co., Gonic, New Hampshire.

resistive networks with F_O through 500 MHz, Caddock Electronics non-inductive MP 800-series resistors appear to provide an excellent alternative to axial-style film resistors. The MP 850 is available from Allied Electronics for under $8 in single quantity, and all models are available direct from the factory with no minimum billing. The 1000-piece lot price for MP 820 is $2.05 each, and $2.99 each for MP 850 (1 percent tolerance). A number of standard resistance values useful for RF design are available from stock. For data sheets, technical support, or literature, contact Bob Bristol at (503)496-0700. Samples are available for prototype evaluation. In addition to Kool-Tab resistors, Caddock Electronics also manufactures a wide range of precision non-inductive axial resistors and SIP resistor networks.

If you come up with an innovative RF application for MP 800-series devices and are willing to share it, we'd like to hear from you. If accepted, your idea will appear in the Tech Notes section of *Communications Quarterly*—and we will pay our standard author's fee for your manuscript. Contact the editorial office at (603) 664-2515 for details. ■

REFERENCE
1. Rick Littlefield, K1BQT, "A –30 dB, 0 to 500 MHz Signal Sampler for Your Spectrum Analyzer," *Communications Quarterly*, Summer 1994, page 8.

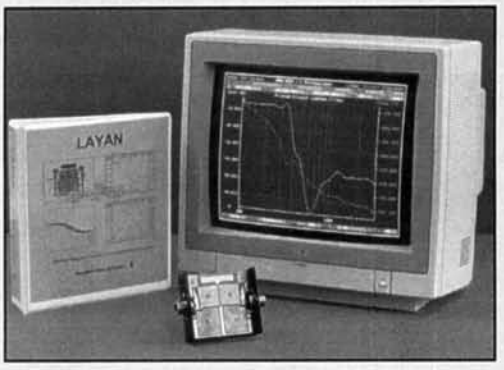
PRODUCT INFORMATION

A Major Breakthrough in Electromagnetic Simulation

Number One Systems, Ltd. has launched an Electromagnetic Circuit Simulation program to help design engineers from fully understanding the behavior of their PCB-based designs.

The LAYAN program brings electromagnetic simulation within the reach of every engineer. LAYAN fits between the EASY-PC Professional XM schematic and PCB Layout tool and ANALYSER III Professional, a Linear Analog Circuit Simulator. LAYAN examines the PCB layout, noting all the track positions, pad dimentsions and copper areas. The program then extracts an equivalent electrical circuit from the physical layout data and combines this with the theoretical circuit so that the behavior of the entire system can be simulated.

Existing users of EASY-PC Professional XM and ANALYSER III Professional can buy LAYAN for less than £500. The complete set of three programs, working together as a fully integrated Electronic Design System, is available for less than £1000. For further information, contact David Hull, Number One Systems Ltd., Harding Way, Somersham Road, St. Ives, Huntingdon, Cambridgeshire PE17 4WR, ENGLAND. Phone: 0480 461778; Fax: 0480 494042; Int'l phone: +44 480 461778; Fax: +44 480 494042.



RFM Develops Data Book for Products

RF Monolithics, Inc. has published the first comprehensive data book detailing its 210 standard products.

The company manufactures and markets radio frequency (RF) components and subassemblies based on surface acoustic wave (SAW) technology. The 528-page data book, available at no cost from RFM, details the company's three major product lines—low-power wireless, high-frequency timing and telecommunications. Included in the data book are contact persons, addresses, telephone and facsimile numbers for RFM's Dallas headquarters; ist operations facility in Woking, England; and its European sales office outside of Paris. The data book also lists the names, addresses and phone numbers of all RFM's sales representatives worldwide. Also featured are short-form catalogs, individual RFM product specifications and case dimensions. Application notes and article reprints are provided to support each of RFM's three product lines.

The data book can be obtained by contacting RFM's headquarters, 4441 Sigma Road, Dallas, TX 75244; phone 214-233-2903; fax 214-387-8148. In Europe, contact RFM's European Operations Facility at P.O. Box 40, Woking, Surrey GU22 7YU, United Kingdom; phone 44 1483 747077; fax 44 1483 747471.

The New Propagation Handbook Released

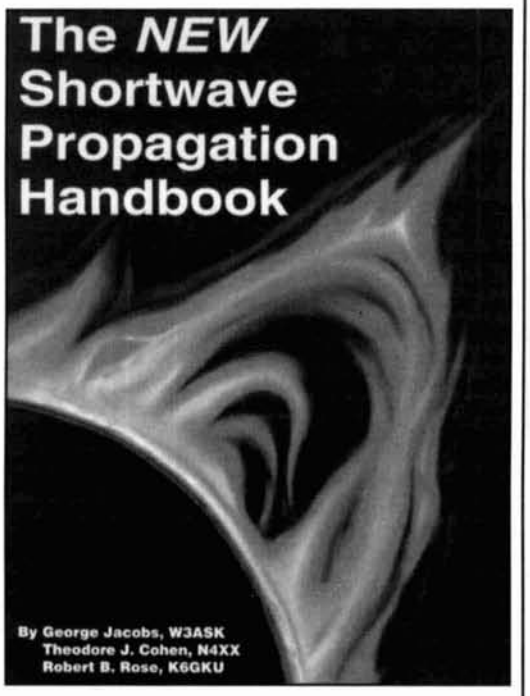
The New Propagation Handbook by George Jacobs, W3ASK; Theodore J. Cohen, N4XX; and Robert B. Rose, K6GKU made its debut this April. The book is a new, revised, updated version of *The Propagation Handbook* and is published by CQ Communications, Inc.

This revised handbook offers amateur radio enthusiasts the ability to make the most productive use of the radio spectrum. Because so many changes have taken place since the first propagation handbook was published, the authors felt it was time for a new comprehensive guide. One of the first changes they made was to update the earlier information on solar phenomena and sunspot cycle behavior. The references and data sources have been expanded, and, in keeping with the times, extensive material has been included on various microcomputer programs that are available for use in developing customized ionospheric circuit analyses. The authors' intent was "to make *The NEW Shortwave Propagation Handbook* the single most comprehensive source available to the amateur, CBer, or SWL interested in optimizing his or her use and enjoyment of the HF spectrum."

The handbook is composed of seven information-packed chapters that cover all you need to know about propagation, sunspots, and the ionosphere. Learn about:

- "Principles of Ionospheric Propagation"
- "Sunspots and the Sunspot Cycle"
- "Sunspot Cycle Predictions"
- "Do-It-Yourself Propagation Predictions and Master Propagation Charts"
- "Ionospheric Forecasts"
- "HF Propagation Prediction Programs;" and
- "Unusual HF and VHF Ionospheric Propagation"

The New Shortwave Propagation Handbook should give an extra edge to anyone interested in communicating



worldwide, in understanding skip interference, or in receiving radio broadcasts from around the world. A wide variety of photographs, figures, and tables (not to mention all of the Master Propagation Charts found in Chapter 4), provide additional information, as well as reinforce the information found in the text.

With all the information in this handbook, it's likely to become *the* resource book found in shacks around the world, and *the* guide that amateurs use to master the art of shortwave radio propagation.

The New Shortwave Propagation Handbook is available from CQ for $19.95 each plus $4 shipping and handling. To order your copy, call toll-free 1-800-853-9797; or mail your order to CQ Communications, Inc., 76 N. Broadway, Hicksville, NY 11801; fax 516-681-2926. Checks, money orders, Visa, Mastercard, AMEX, and Discover accepted.

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A FEW WORDS ABOUT DSP

The DSP-3's designer explains some of the digital signal processing applications and algorithms useful in ham radio!

DSP can be used to perform many different kinds of signal processing functions, including some that would be impractical to attempt with analog circuitry. Plummeting costs of DSP components (CPUs and converters) have brought DSP well within the budget of the typical amateur.

Filters based on delay lines are hard to construct with analog circuitry because analog LC delay lines are themselves generally complex structures. However in DSP, delay lines are easily made out of memory. The advantage of this kind of filter is that it can be made linear phase, which means superior performance for CW and FSK signals.

Analog filters are generally difficult to make variable (especially high order filters), and even more difficult to make self-modifying or adaptive. In DSP, on the other hand, adjustable and even adaptive filters are quite easy. Unlike the fixed, single bandwidth, single frequency filters such as the crystal filter in your rig, DSP filters can be easily designed as multipurpose devices. There are *fifty-eight* DSP-based IF filters in the Watkins-Johnson HF-1000 receiver![1] These filters are all created by the same hardware; however, different coefficients are loaded in software.

All of these applications can be performed in software. Little or no cost increase is incurred because "it's only *software*!" (Yes, I should know better than to say such things; my wife Sandy, N6SMF, is a software engineer!)

Digital filters can be made much more accurate and are 100 percent repeatable. Unlike analog circuitry, there's no "component tolerance" problem.

DSP basics

DSP begins by first converting the signal we want to process to a series of numbers or samples. The filtering function is then performed numerically. A DSP chip executes a software algorithm, and the result is another sequence of numbers. The output samples are converted back to an analog waveform that is then routed to a conventional audio amplifier.

The FIR filter

For DSP-based filtering, one of the most useful structures is the finite impulse response, or FIR, filter (see **Figure 1**). The samples enter the tapped delay line. The blocks labeled "Z^{-1}" represent a time delay of one sampling period. At every tap, the signal values are multiplied by a coefficient K_m. The resulting products are summed to form the filter output. With proper choice of coefficients, lowpass, bandpass, highpass, bandstop, and other kinds of filters can be implemented.

In many applications, the filter coefficients are fixed, resulting in a fixed non-varying filter for such things as CW bandpass filtering, FSK filters, etc.

FIR filters are usually designed to be *linear*

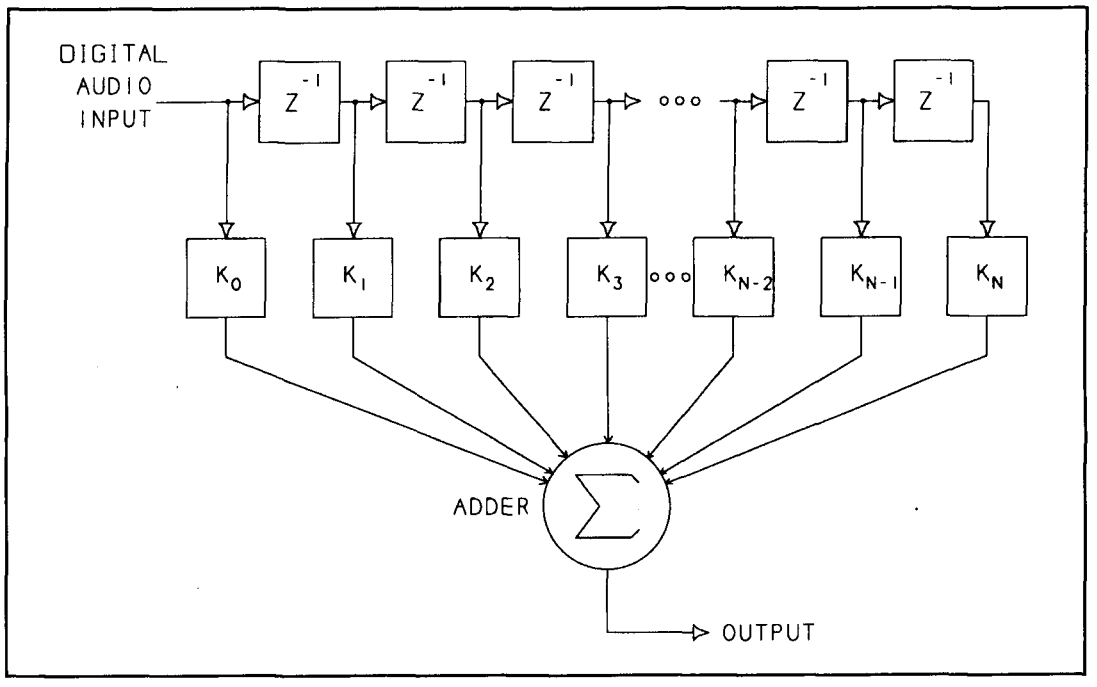

Figure 1. Finite impulse response filter block diagram.

phase. This means that the time delay through the filter is the same for all frequency components. (This is definitely not the case for most analog filters.) The advantage of linear phase is that FIR filters have less ringing and unwanted transient effects than equivalent analog (nonlinear phase filters).

(All bandpass filters ring, by the way. They do so in order to provide selectivity.)

Have you ever wondered why many DSP CW filters have bandwidths of 200, 100, or even 50 Hz, while crystal filters in transceivers are usually 400 Hz wide? One reason is that for a given amount of ringing, DSP-type FIR filters can be made narrower. By the same token, for the same amount of detected pulse distortion, linear phase FIR filters for FSK can be made narrower than equivalent analog filters, allowing them to reject more interference than their analog cousins.

FIR filter tradeoffs

FIR filters are similar to analog filters in on way—the more complex the filter, the better the selectivity. The 1.8 and 2.1 kHz narrow SSB filters in the W9GR DSP-3 are FIR types. I had the option of either making these filters razor sharp "brick wall" types (to mix metaphors), or to compromise the selectivity by combining the narrow filter with a denoiser and/or an automatic notch function. I chose to make them extremely sharp. I figured that the filters will be used when the dominant interference isn't noise, but a strong adjacent channel signal. In that case, it would be desirable to eliminate as much of the interference as possible—hence we would have an extremely sharp filter, without simultaneous noise reduction.

Adaptive filtering

Fixed FIR filters are often useful, but in other applications we may want to have the filter alter its characteristics in response to changing input or interference signals. Remember, DSP filters are easy to change, and we have the power of a processor on hand to help us decide how to make the filter self-modifying or adaptive under varying conditions.

The FIR filter above forms the basis for adaptive filtering based on the Widrow-Hoff Least Mean Square (LMS) algorithm.[2,3,4] But before I tell you how that works, I'll describe another technique that achieves basically the same result but in a less roundabout fashion. Once you understand "spectral subtraction," it will be easier to understand the LMS algorithm.

Spectral subtraction

One technique useful for DSP-based noise reduction is called "Spectral Subtraction."[5] Spectral subtraction is based on the use of the discrete Fourier transform (DFT) and its inverse, the inverse DFT. These transforms convert a signal from the time domain to the frequency domain and vice versa.

Spectral subtraction begins by taking the DFT of an input signal. This takes our time domain sequence of input samples, and converts it to an array of values that represent the energy present in our input signal as a function of *frequency* rather than time. If we were to immediately perform an inverse DFT on this *frequency domain* data, we would get our original *time domain* samples back again.

Instead of immediately returning to the time domain, spectral subtraction first manipulates the frequency domain DFT coefficients in a way that reduces noise. The assumption is that the desired signal isn't uniformly distributed throughout the frequency domain, but concentrated in a small number of frequencies.

Spectral subtraction further assumes that the noise *is* approximately uniformly distributed throughout the frequency spectrum.

In this method of differentiation between desired signals and undesired noise, spectral subtraction looks at the energy at each frequency. If it is lower than a certain threshold, that frequency component is assumed to be due to noise and is set to zero. This is shown in **Figure 2.** The resulting modified frequency domain data is then applied to the inverse DFT to produce a new time domain signal, hopefully with less perceived noise.

Spectral subtraction is used today in some DSPs. However, it has the following problems:

1. Time delay required to perform the forward and inverse transforms is a fraction of a second. This makes it difficult to rapidly tune receivers to look for signals. By the time you hear a signal at the output of the DSP, you may have already tuned past it. It also makes very rapid contest exchanges impossible.

2. Audio "tones" can be produced in some implementations. These appear as a series of seemingly random beeps with a length related to the number of samples in the DFT. They are caused by the algorithm's imperfect ability to distinguish between signal and noise in the frequency domain.[6]

3. The algorithm can only react or adapt in discrete time blocks (DFT length).

4. More DSP computing power is required than necessary for the LMS algorithm.

5. FFT/DFT artifacts such as windowing effects may adversely affect performance.

Although spectral subtraction has some disadvantages, under a number of conditions it may provide the best performance. The JPS NIR-12 DSP, for example, provides *both* spectral subtraction and LMS noise reduction (which JPS refers to as "adaptive peaking"). This combination allows the DSP user to choose the optimum noise reduction method for each set of circumstances.

To summarize, spectral subtraction looks for the most significant spectral lines in a signal and allows them to pass. Wideband noise signals that are well dispersed in the frequency domain tend to be rejected. Spectral subtraction does this in a very straightforward way—by transforming the signal to the frequency domain, throwing away the smaller Fourier

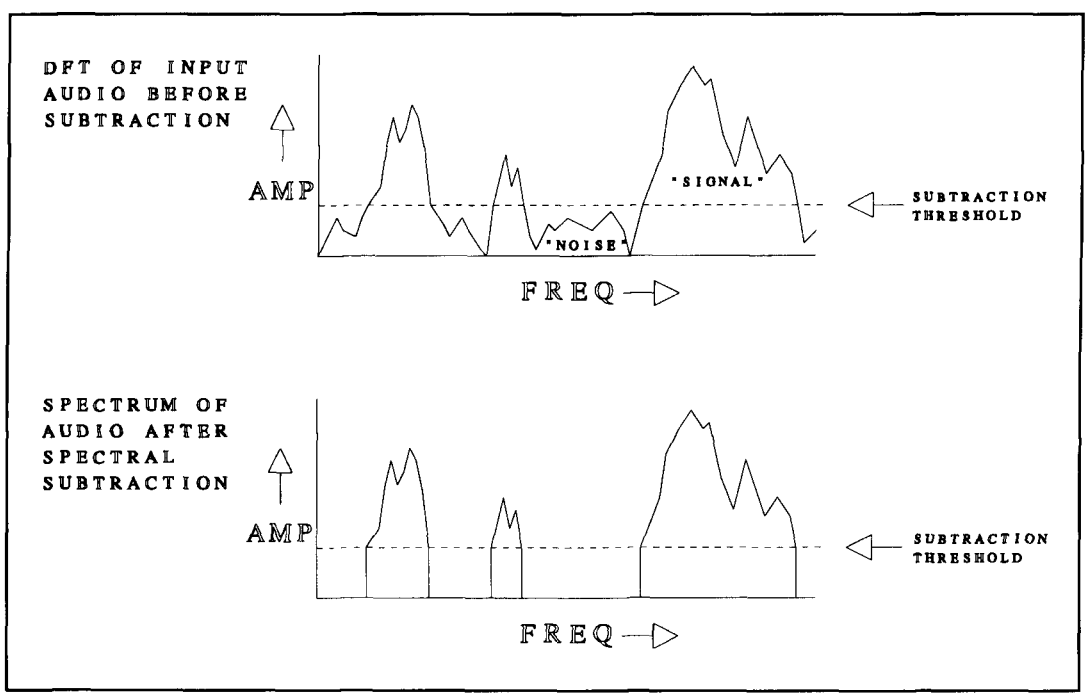

Figure 2. Spectral subtraction.

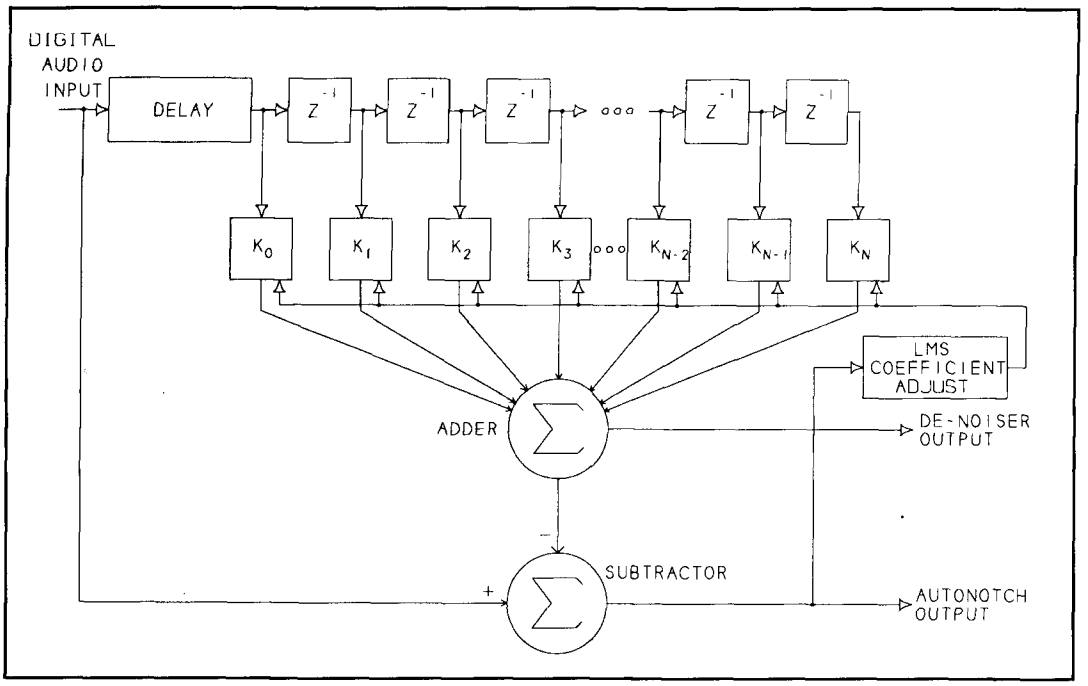

Figure 3. Widrow-Hoff LMS algorithm block diagram.

coefficients, then performing the inverse transform to return to the time domain.

The LMS algorithm

The Widrow-Hoff LMS algorithm (commonly abbreviated "LMS algorithm") does basically the same thing, but in a more roundabout fashion. In the noise reduction configuration, LMS also emphasizes signals that have much of their energy in a few dominant spectral lines, while reducing wideband noise type signals. However, in LMS, there is no transformation to and from the frequency domain. Thus, a major computational inefficiency is avoided, and the pitfalls associates with frequency domain transformations are avoided.

The LMS algorithm, like spectral subtraction, discriminates for or against signals as a function of how repetitive they are. This is another way of specifiying to what extent the signal energy is concentrated in a number of discrete spectral lines.

Although there are forms of the LMS algorithm that require both a corrupted signal plus noise input and a separate noise input, the form of the LMS algorithm most useful in amateur applications only requires a single input consisting of the desired signal plus noise and/or inference (see **Figure 3**). In this form of the algorithm, LMS compares the waveform of the input signal with a delayed version of itself that has been screened by a FIR filter. LMS subtracts the delayed and filtered signal from the input, and continually adjusts the FIR filter coefficients to try to force the output of the subtractor to zero.

If the input signal is a pure sinewave, then every cycle of the input signal is exactly the same as every successive cycle, and it's possible to choose FIR filter coefficients so the output of the subtractor is zero.

On the other hand, if the signal is white noise, it has no generally repetitive characteristic whatsoever because the white noise is random. It won't be possible to find a set of FIR coefficients to null out the noise at the subtractor output. In this case, a well-designed system will tend to generate a FIR filter whose coefficients are all zeros or nearly zero. Ideally, the FIR filter output would be zero.

Thus, the output of the FIR filter itself tends to contain primarily the repetitive signal components, and it tends to reject the nonrepetitive, noise-type signals.

If we take our output from the subtractor, we have the system topology for an automatic notch filter. On the other hand, if we take our output from the FIR filter, we have a noise reduction system.

Depending on whether we're trying to design a noise reduction filter or an automatic notch filter (or both) based on the LMS algorithm, there are a number of parameters that will affect the performance of our DSP. Delay line length, FIR filter order, sampling rate, LMS convergence coefficient, and the rate at which we decay our FIR filter coefficients are the various "handles" we have on the LMS algorithm. In addition, some experimenters have devel-

oped modifications to the basic algorithms to achieve certain objectives or improvements such as "making the adaptation itself adaptive."

Adaptive filter design: It's an art!

If we were only trying to design a fixed CW filter, for example, we could use a fairly objective design process. We could decide what bandwidth we want, the amount of tolerable ripple in the passband and the stopband, filter order, transition bandwidth (related to shape factor), and so on. All of these things are objective values to which we can assign numbers.

However, designing an adaptive filter using either spectral subtraction or LMS is not an objective process. I know of no way to put a number on "intelligibility improvement" that can be directly and objectively related to adaptive filter parameters. Subjective listening tests are required to evaluate performance. In other words, adaptive filter design is an art.

Adaptive filter design involves using a wide variety of real signals, real ears, and carefully tweaking the various parameters associated with the particular algorithm.

Because it is an art, there will be some variation in performance from one programmer to the next, resulting in different output signals from different digital signal processors.

My own preference is to try to optimize the DSP performance on the weakest and noisiest signals, while at the same time not degrading the intelligibility of strong signals.

Artifacts are a pitfall in the design of LMS filters. In LMS noise reduction filters, audible resonances are undesirable artifacts. This may sound like reverberation or "flanging," where resonances move through the signal—changing the frequency response and the signal timbre.

In spectral subtraction, similar effects can occur. Also, there can be seemingly random "bleeps" and "bloops" (audio tones). These are called "annoying musical tones."[6]

There are automatic notch designs that attack and remove the fundamental component of female voices. Some male voices also have significant voice energy removed. Poorly designed automatic notch filters may remove a significant portion of the desired signal's correlated energy, leaving the remaining signal sounding raspy or distorted. It is not nonlinear distortion being produced in this case; rather it is the emphasis of uncorrelated signal components that sound like distortion.

Other DSP functions

I hate to leave unused code space go to waste. After I did some code size reduction on the software/firmware in the DSP-3 design, there was a fair amount of space left for more DSP functions. So, I decided to put something useful in that space. What I came up with were two new functions—DTMF ("Touch Tone") and CTCSS ("PL") tone decoding. These functions let you check out your VHF/UHF gear, find out what PL tones are being used for repeaters, and so on. They also satisfy my passion for using every last available byte to do something useful.

Summary

DSP filters have brought an enormous amount of powerful signal processing to radio amateurs. Even though some advanced signal processing techniques are used in these devices, they do not have to be costly. The W9GR DSP-3 kit is an inexpensive way for amateurs to get started using advanced DSP algorithms. And it draws on the long amateur tradition of building it ourselves! ■

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PRODUCT INFORMATION

New Network Analyzers from Hewlett-Packard

Hewlett-Packard Company's new 300 MHz network analyzers focus on the final testing and plating process in the production test of crystal filters and resonators.

Specifications include: Frequency: 10 kHz to 300 MHz with 1-mHz resolution; Power: –9 to +11 dBm (standard) and –48 to +22 dBm (option 010); IFBW: 10 Hz to 30 kHz (1, 1.5, 2, 3, 4, 5, 8 steps); Dynamic range: 120 dB (IFBW = 1 kHz, 5100A); Dynamic accuracy: ±0.05 dB, ±0.3 deg; Measurement speed: 0.04 ms/point (IFBW = 30 kHz); Display: Monochrome LCD.

The HP E5100A network analyzer is available at prices starting at $15,000; the HP E5100B is available from $12,000. For more information, contact Hewlett-Packard, Direct Marketing Organization, P.O. Box 58059, MS51L-SJ, Santa Clara, CA 95051-8059.

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QUARTERLY REVIEW

Quantics W9GR DSP-3: affordable technology fights QRM!

First described in the September 1992 issue of *QST* magazine, the W9GR Digital Signal Processor was an immediate success. Over 2500 of these units have been shipped worldwide. Dave, W9GR, took note of the countless suggestions offered by users of the DSP-1 and DSP-2, and came with the DSP-3. Both the hardware and software used in the DSP-3 are significant improvements over that used in the original DSP-1.

Most hams are probably familiar with the more common analog audio filters, like the Autek QF-1A. I've used an Autek QF-1A audio filter on my ham receivers for several years, and until I was introduced to DSP technology, I had never operated without it! Audio filters are accessories that are intended to enhance receiver operation, and to overcome, to some degree, minor deficiencies that may be present in some models. For example, my JRC NRD 515, an otherwise fine receiver, does suffer from some degree of IF filter blow by, and filters that are a tad wide for serious SSB or CW work. This results in annoying high frequency monkey chatter from stations operating too close the to station I'm trying to work.

I've noticed that opinions on the utility of audio filters are subjective—some hams swear by audio processors, others feel they offer no noticeable improvement. A case in point: my QF-1A was a gift from a friend who felt it did nothing to enhance his operating pleasure. Regardless of the filtering method used, there are limitations! Filtering ideally belongs before the gain producing stages in a receiver, preceding the AGC detector. When using an outboard audio filter to produce a narrower bandwidth than is provided by the receiver IF, a strong signal within the IF bandwidth, yet outside of the audio filter bandwidth, may swamp the receiver AGC—reducing the desired signal's level to an almost unreadable level. The same problems can occur in CW, or when using notch filtering to reduce unwanted carriers in SSB operation.

Both analog and DSP filters can offer sharp

Mode 10. 400-Hz CW filter, 100-Hz bandwidth
passband 350 –450 Hz @ –3 dB
<270 Hz attenuated >60 dB; >530 Hz attenuated >60 dB.

Mode 11. 400-Hz CW filter, 50-Hz bandwidth
passband 375 –425 Hz @ –3dB
<320 Hz attenuated >50dB; >480 Hz attenuated >50 dB.

Mode 12. 600-Hz CW filter, 100-Hz bandwidth
passband 550 –650 Hz @ –3dB
<470 Hz attenuated >60dB; >730 Hz attenuated >60 dB.

Mode 13. 750-Hz CW filter, 200-Hz bandwidth
passband 650 –850 Hz @ –3 dB
<575 Hz attenuated >60 dB; >925 Hz attenuated >60 dB.

Mode 14. 750-Hz CW filter, 100-Hz bandwidth
passband 700 –800 Hz @ –3 dB
<620 Hz attenuated >60 dB; >880 Hz attenuated >60 dB.

Mode 15. 750-Hz CW filter, 50-Hz bandwidth
passband 725 –775 Hz @ –3 dB
<670 Hz attenuated >50 dB: >830 Hz attenuated >50 dB.

Mode 16. 1000-Hz CW filter, 100-Hz bandwidth
passband 950 –1050 Hz @ –3 dB
<870 Hz attenuated >60 dB; 1130 Hz attenuated >60 dB.

Table 1. CW filters, with various center frequencies and bandwidths, offered with the DSP-3.

filter bandwidths for CW and sideband operation. The main difference is where analog filters rely on hardware to produce the filter elements (discrete op amps, precision resistors and capacitors), the approach taken by DSP units is purely software derived. Of course, there is hardware in a DSP unit, but unlike analog filters, the hardware isn't modified to create a new filter. Changes are made solely by software control of microprocessor-based hardware.

Most hams accept filter "ringing" as an unfortunate tradeoff when using sharper filter positions—something that can result from poorly designed narrow bandwidth SSB/CW filters at audio or IF frequencies. While "ringing" is a natural and expected result of bandwidth reduction, it is greatly exaggerated by filters exhibiting nonlinear phase responses. The W9GR DSP units use linear phase FIR filters for CW. Even at the sharpest 50-Hz selection, ringing isn't a problem.

The hardware in the DSP-3 is based on the Texas Instrument 16-bit TMS320P15 DSP CPU chip. Because DSP CPUs, like other CPUs, are digital devices, a Texas Instrument TCM320AC39 audio codec chip performs the necessary analog-to-digital conversion for the CPU, and conversely reconstructs the digital output of the CPU back to analog audio. Output from the codec chip is serial; this is converted to parallel data for loading into the CPU by two 74xx299 shift registers. The codec chip also limits the audio response to a maximum of 3.2 kHz, via an internal SCF lowpass filter, to prevent aliasing due to the DSP CPU's sampling rate of 7700 Hz. Additional SCF filtering helps with reconstruction of the analog audio from the digital data stream. The 13-bit precision allows for 80 dB of dynamic range—a large improvement over the 50 dB limitation of the original DSP-1 hardware! This equates to a greater ability to ferret out weak CW signals buried in strong QRM, and offers a large amount of headroom in SSB operation, where the typical S/N rarely exceeds 30 dB.

Adaptive filtering

Adaptive filtering is a powerful feature of DSP. In this software "mode" the filter actually adapts itself to the noise or QRM environment. It works by forming narrow bandwidth notch filters around the most repetitive interfering tones within the receiver's passband, effectively removing them from the output. Because heterodynes caused by carriers are the most repetitive in nature, they are easily dealt with by the DSP-3 adaptive notch filter position.

Noise reduction is somewhat more difficult. The repetitiveness of speech signals falls somewhere between constant carriers (completely repetitive) and noise (completely unrepetitive.) The trick is to favor the somewhat repetitive speech signal and reject the unrepetitive noise. To make this distinction, the input SNR should be 6 dB or better for optimum performance. Impulse, or ignition noise, is best handled by a noise blanker. The DSP-3's ability to eliminate carriers on SSB is amazing. Sixty dB over S9 interference from stations tuning up on 20-meter DX stations was completely eliminated by my DSP-3 in a matter of milliseconds! I was curious to see if the adaptive noise filter would work on FM to eliminate noise on less than full quieting signals. I was informed by W9GR that not only is the DSP-3 is very effective for noise elimination, some West Coast repeaters use DSP-3s for this purpose!

SSB filters

The DSP-3 offers two SSB positions: 2.1 and 1.8 kHz. The 2.1-kHz FIR filter has a passband of 300 to 2100 Hz with less than 0.3 dB of ripple. The stopband attenuation of this filter is spectacular, with attenuations of greater than 50 dB at 150 Hz, and 70 dB at 2300 Hz!

Mode 6. RTTY filter. Intended for 170-Hz FSK signals, Baudot, Amtor.

BIO=0 North American RTTY tones 2125/2295 Hz
passband 2075 –2345 Hz ±0.3 dB ripple
<1875 Hz attenuated >60 dB; 2545 Hz attenuated >60 dB.

BIO=1 European RTTY tones 1275/1445 Hz
passband 1225 –1495 Hz ±0.3 dB ripple
<1025 Hz attenuated >60 dB; >1695 attenuated >60 dB.

Mode 7. HF-packet filter, SSTV filter. Intended for
200-Hz shift 1200-baud HF-packet FSK signals, or SSTV signals.

BIO=0 HF packet.
passband 1550 –1850 Hz ±0.4 dB ripple
<1350 Hz attenuated >65 dB;>2050 Hz attenuated >65 dB.

BIO=1 SSTV.
passband 1075 –2350 Hz ±0.3 dB ripple
<875 Hz attenuated >70 dB; >2550 Hz attenuated >60 dB.

Table 2. Switch settings and filter specifications.

Performance of the narrower 1.8-kHz filter is equally impressive. With a passband of 300 to 1800 Hz, this filter is 50 dB down at 150 Hz and 70 dB at 2100 Hz. Note that as tradeoff for this brickwall filtering, adaptive noise or notch filtering is not available in the SSB filter positions. Cascaded adaptive noise or notch filters could have been added to the SSB filter modes, but with degraded sideband filter performance. The available ROM area in the TI CPU also becomes a limiting factor for the software needed to perform cascaded filter operations of this nature.

CW filters

Quantics offers several CW filters, with various center frequencies and bandwidths (see **Table 1**). They are all 229th order FIR filters.

Each CW filter center frequency is tunable in a downward direction by 30 percent. This is done at "power on" with the appropriate filter selected. Each filter will tune downward (producing a slowly descending audio note at center Fo) until stopped by the operator at the desired point. For example, the 1000 Hz center Fo filter may be tuned to as low as 700 Hz. This is volatile; the filter reverts to the preprogrammed value when power is removed.

One benefit of the narrow filters is noise reduction. The narrower the bandwidth, the quieter the receiver became. The DSP CW filters were more effective in noise reduction than was using the JRC 400 Hz BW CW filter with the DSP-3 noise reduction (Mode 2).

Bargraph display

The DSP-3 includes a 10-segment bargraph display. For most modes, the display is a peak reading meter with delayed decay, each display LED corresponds to a 3 dB input change for a total dynamic display range of 30 dB. The ideal volume setting is such that the majority of LEDs will be illuminated on the audio peaks of the loudest signal. This is optimum for best D/A and A/D precision.

Internal AGC

On the rear panel of the DSP-3, you'll find switch S4. This is the BIO (for binary I/O) switch. The main DSP functions of the DSP-3 are controlled by the 16 position switch, with 12 noon being mode 1, increasing in a clockwise rotation. Because the DSP-3 features eighteen DSP functions, the BIO switch is used to activate special applications. For example, in modes 1, 2, and 3 (1. combined automatic notch and denoiser, 2. denoiser, and 3. automatic notch, respectively) placing the BIO switch to 1 will activate the digital DSP AGC in the DSP-3. So long as one LED is lit, the audio will rise to full level in the presence of a signal. This AGC function is entirely DSP based with no analog processing of the audio. The AGC attack time is virtually instantaneous; the release time is 30 seconds. One can force the AGC to reset by momentarily selecting another mode.

For the ultimate in noise reduction, the

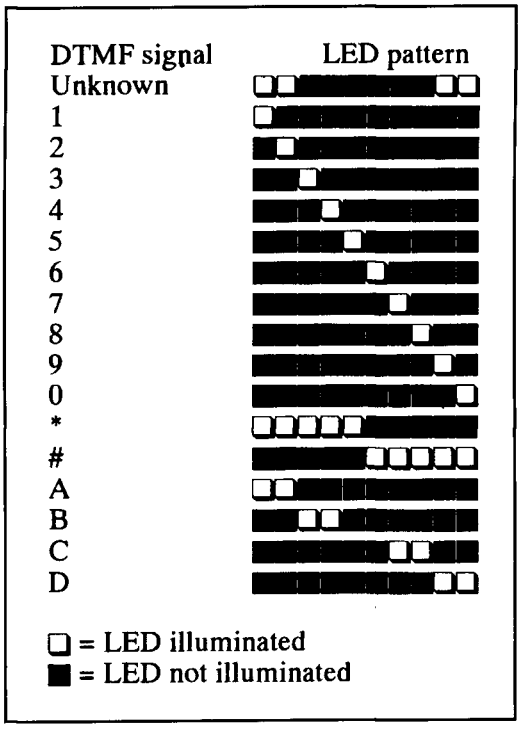

Table 3. LED illumination guide for DTMS decoding.

designer recommends mode 2 versus mode 1 for somewhat greater effectiveness. The noise reduction feature is a bit more subtle than what you observe when using the brickwall filters or notch filter—these are far more dramatic in use. In tests, I noticed some reduction in background noise on 75-meter SSB, but the noise reduction effect was more pronounced on ten meters (about 6 dB by ear alone)—no doubt due to the different characteristics of the noise spectra. All in all, the noise reduction does work, but don't expect miracles or you'll be disappointed. VHF operators may gain some fatigue reduction using the noise reduction feature.

RTTY, HF packet and SSTV filters

These filters are selected by modes 6 and 7 and the BIO switch. All are 100th-order linear phase filters. See **Table 2** for switch settings and filter specifications.

DTMF and CTCSS

FM VHF operators and scanner buffs will be interested to know that the DSP-3 can decode DTMF and CTCSS tones. Mode switch position 8 is for DTMF, position 9 is for CTCSS. Using a unique bargraph readout (see **Table 3**) the DSP-3 can decode and display the value of any of the standard 16 ATT Touchtone tone groups. Up to sixteen sequential digits may be stored, played back, and displayed via an internal FIFO register. Playback is at about a one digit per second rate, with decode display.

CTCSS, or subaudible "continuous tone-coded squelch system" (PL for you Motorola FM radio fans), decoding is available via mode 9. Any of the standard 38 tones can be detected and displayed for as long as the tone is present (see **Table 4**). The receiver used must be able to pass audio down to at least 50 Hz without attenuation for this feature to work!

Custom software

The programming for the DSP-3 is supplied in preprogrammed nonvolatile memory burned into the TI DSP CPU chip. For those advanced experimenters who want to try their hand at DSP programming, the CPU may be replaced with a TMS320C10 or TMS320C15. These processors have no provisions for internal program memory storage. This is provided for on the DSP-3 pc board by socket positions that allow adding two high-speed 57C43 EPROMS for custom software. Those wishing to delve deeper into the mathematics and programming algorithms for DSP should note **References 1** and **2**.

Besides the front panel 16-position mode switch and rear panel BIO selector, there is a front panel power push off/on switch, and a push off/on switch that allows bypassing the DSP-3. A front panel 1/8-inch stereo jack allows headphone operation. The pc board may be jumpered to allow for either stereo or monaural earphones to operate from this jack.

The DSP-3 kit

I recommend this kit for experienced builders. The many soldering points, most with tight 0.1 inch spacings, require good eyes, a steady hand, and good soldering skills. The instruction booklet is complete, but make sure you read and understand it before attempting to assemble the unit.

The pc board is silk-screened, but in some cases the markings may be covered by the components once they're installed. This is true of the polarity markings for the electrolytic capacitors. It's important to make sure you've inserted them right the first time! You *must* be able to read resistor and capacitor values; give yourself plenty of time! The markings on 10 k and 1 k resistors can begin to look alike to tired eyes. The resistors aren't presorted, so using a digital ohmmeter to verify values is a good idea. The entire kit assembly is broken down into 24 steps. Most of the assembly instructions deal with inserting the smallest components first,

working up to the larger components last. Decision points are reached where jumpers may or may not be installed on the board. There are no drawings—you must read the component value and search the silk-screen for its location. There is a precheck procedure that I strongly suggest you adhere to before installing any ICs. Not surprisingly, Quantics reports the vast majority of returned kits don't work due to faulty soldering or improper component insertion. For the faint of heart, assembled DSP-2 units are available.*

An experienced electronic technician might be able to assemble this kit in under three hours, but I advise allowing the better part of a weekend day, and frequent rest breaks, to assemble yours.

There's a brief troubleshooting section, and the operating directions are complete and thorough. I heartily recommend the optional case—it's made of steel and lends enough weight to the unit to keep it in place while you are manipulating switches, etc. The case is painted a handsome grey, with silk-screened lettering and is well made. Power is taken from an external 12-volt DC source.

My impressions

There were no surprises here. The unit performed as well as I had hoped. I gave the unit a good workout on the lower HF bands for several evenings. The CW filters are excellent—signals not in their bandwidth just aren't there! Also, it was nice to be free of the regenerative "howl" I had experienced when using analog audio filters set for extremely sharp CW bandwidths. Ditto for the SSB filters. My first impressions of the noise reduction modes were given earlier; but as I get more comfortable using the DSP-3 in the 75-meter DX window, I am now becoming more aware of the way signals seem to "standout" of the noise when the noise filter mode is in use.

For my own purposes, I would have preferred the designer offered a few more wider-bandwidth CW positions—200 Hz or lower isn't always needed or desired for casual CW work. Considering that the CW filters are tunable, it seemed somewhat redundant to offer repetitive bandwidth values with differing center frequencies. I would have liked to have adaptive filtering available in the SSB modes for noise or carrier rejection; however, I can also accept the design tradeoff made for no-compromise brickwall filtering. Again, I can't help thinking that one or two of those redundant CW filter positions could be used for "weaker" SSB filters that would allow cascading some adaptive filtering for noise or carrier reduction. This would be a plus, in my opinion.

CTCSS signal	LED pattern
NO CTCSS	
67.0	
71.9	
74.4	
77.0	
79.7	
82.5	
85.4	
88.5	
91.5	
94.8	
97.4	
100.0	
103.5	
107.2	
110.9	
114.8	
118.8	
123.0	
127.3	
131.8	
136.5	
141.3	
146.2	
151.4	
156.7	
162.2	
167.9	
173.8	
179.9	
186.2	
192.8	
203.5	
210.7	
218.1	
225.7	
233.6	
241.8	
250.3	

□ = LED illuminated
■ = LED not illuminated

Table 4. Illumination guide for CTCSS decoding.

One change I did make was to decrease the length of the plastic shaft of the mode switch—enough to permit the knob to sit closer to the front panel. Even so, the knob skirt indicator is small, the dial markings are petite, and it's hard at times for my old eyes to focus on exactly which mode the knob has selected. This was a major pain at times. The BIO switch has a fair-

* An assembled W9GR II DSP unit is available from j-Com. The DSP-2 is similar to the DSP-3, but with fewer DSP modes. The j-Com is unit comes in a nice extruded enclosure and is priced at $299.00. j-com accepts credit cards and may be reached at (716) 924-0422.

ly long, threaded shaft. I added a second "backing" nut to limit the distance the switch protruded through the cabinet housing.

All-in-all, I found the DSP-3 a good value; a useful station accessory for the serious HF/VHF operator. The future of DSP lies in IF signal processing, and perhaps, someday soon, all receivers will implement DSP technology at the final IF frequency. Until that time, we have the DSP-3.

For more information, or to order the Quantics DSP 3, write to: Quantics, P.O. Box 2163, Nevada City, California 95959-2163. The DSP-3 kit is priced at $149.95; the optional DSP-3 enclosure is $19. ■

REFERENCES
1. Dave Hershberger, W9GR, "Low-Cost Digital Signal Processing for the Radio Amateur," *QST*, September 1992.
2. Dr. Steven E. Reyer, WA9VNJ and David L. Hershberger, W9GR, "Using the LMS Algorithm for QRM and QRN reduction," *QEX*, September 1992.

Corrections

Please don't ground me!

Clint Bowman, W9GLW, sent us the following correction to his article "A Low-Power 20-Meter Transceiver" (Winter 1995, page 69):

"Pin 14 of the MC1648P integrated circuit in the VFO schematic found on page 71 should not be grounded. Instead, it should be connected to the 5-volt positive source, and bypassed to ground through a 0.01 ceramic capacitor."

Good upbringing outs

Once again from W4RNL:

"You probably already know this, but your table of contents gives me credit for both the Moxon article ("Modeling and Understanding Small Beams, Part 2," Spring 1995) and the vertical study. I would love to take credit for the extra good work, but honesty says I have to report this for a correction in the next issue (doggone childhood training and genes)."

Thanks, L.B.! The article in question, "Understanding Elevated Vertical Antennas," was written by Bill Shanney, *KJ6GR*, and appears on page 71 of our Spring 1995 issue.

Sorry for the mix-up, Bill.

The dreaded typo

"There's a dreaded typo in the Beta match discussion on page 51, Equation 3 (Technical Conversations, Winter 1995*)," wrote L.B. Cebik, W4RNL. The equation should read:*

Xa = **Ra** SQR (Ro-Ra/Ra)

PRODUCT INFORMATION

New W9GR DSP-3 DSP Kit Introduced

Quantics has introduced a new digital signal processor (DSP) kit for amateur use. The DSP-3 kit incorporates many suggestions and requests made by users of the original W9GR DSP-1 kit. This new kit has 18 DSP functions selected by a rotary switch, including various combinations of noise (QRN) reduction and heterodyne removal (automatic notch filtering), a DTMF tone decoder with memory, a CTCSS tone decoder with memory, seven tunable CW filters, and various filters for FSK, SSTV, and narrow SSB. A 13 bit converter chip provides the wide dynamic range necessary to filter out weak signals amidst strong QRM. A LED display indicates audio level, or in the tone decoder modes it indicates DTMF and CTCSS tones. The kit operates from a 12 volt DC power source (not included) and connects between your rig and speaker. A custom metal cabinet is also available.

The DSP-3 kit is priced at $149, and the optional metal cabinet is $19. California residents must add 7.25% sales tax. Shipping and handling is $7 in the USA and Canada. Quantics deals by mail order only.

For further information, write to Quantics, P.O. Box 2163, Nevada City, CA 95959-2163.

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P.O. Box 307
Meriden, New Hampshire 03770
CompuServe: 72510,2302
MCI Mail ID: 407-8117

QUARTERLY COMPUTING

Building a PC toolkit

Unless you can afford to buy one of those fully loaded 552-piece briefcase toolkits featured in better mail-order catalogs, chances are you've grown your ham radio toolkit one piece at a time. In fact, over the years you've probably assembled several special-purpose tool collections.

Looking around the house and garage, I discovered that I own seven—a portable toolbox stuffed with generic components for electronic fix-it visits to family members, a more elaborate ham shack kit housed in a machinist's chest, a tool tray for household maintenance, smaller kits for the truck and car, a few tools in the fishing tackle box, and a belt pouch containing a multimeter and a couple of screwdrivers for hamfest take-apart missions.

When you acquire a personal computer, you quickly discover that a PC's operating system and its accessories also resemble a toolkit—one that lets you create and manipulate files, view and edit text, and more.

However, you may also discover that the operating system's tools don't "feel right." They're clumsy to use, require too many keystrokes, or lack capabilities you need. Before long, you'll find yourself building toolkits composed of programs and utilities instead of screwdrivers and pliers.

In this column, we'll look at a couple of categories of programs that fit nicely into the amateur radio software toolbox. You can find shareware and commercial versions of both types, and while the specific products I discuss are oriented toward IBM-compatible PCs, similar programs exist for Apple's Macintosh family of computers.

Personal Information Managers

Clever programmers know that while computers process *data*, people use *information*. Indeed, many so-called "killer apps"—programs like Lotus' 1-2-3 that open a new chapter in computer history and earn fabulous profits for their creators—define new methods of turning data into information.

At the other extreme, novice programmers frequently spend their time writing yet another checkbook-balancing or recipe-management program that forces its users to convert information back into data in a format convenient for the PC.

In between, a class of products called Personal Information Managers, or PIMs for short, have earned modest success by offering users flexible manipulation of data into information while maintaining a relaxed attitude toward data entry.

Think of a PIM as a multiway merger of a lab notebook, a box of filecards, a calendar, and a reminder timer. Add a search engine that can locate any text or numbers you've entered, and your PIM is complete.

Actually, you can use database software to create your own PIM—a somewhat daunting chore made easier by the advent of object-oriented languages such as Microsoft's *Visual Basic* and Borland's *Delphi*, and databases like Claris' *Filemaker Pro for Windows*. However, most of us would rather spend our free time on amateur radio, and so we'll purchase a shareware or commercial PIM.

In the commercial PIM category, Micro Logic's *Info Select for Windows* rates as my personal favorite. Also available in a DOS version, *Info Select* uses a windows-and-stack metaphor to keep track of your data. A "window" defines an electronic sheet of paper onto which you type your data, and a "stack" consists of a group of windows.

What goes into a window? Just about any

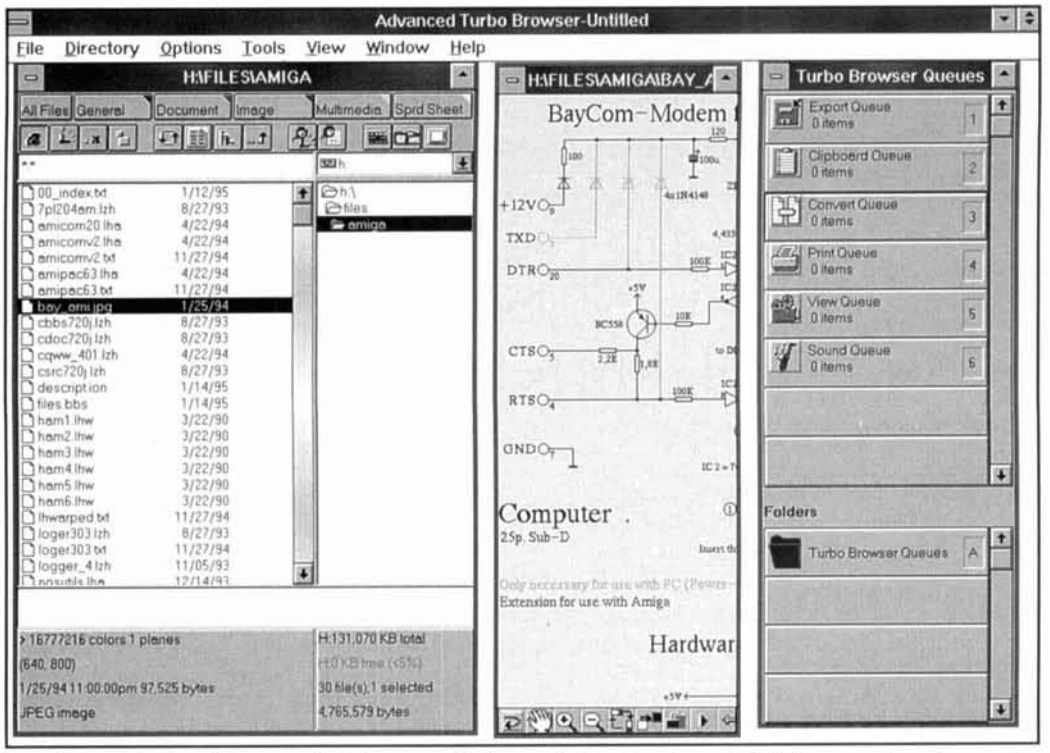

Figure 1. ***Info Select's*** **windows can include automatic time-and-date stamping and link phone numbers to a modem for automatic dialing. Scroll bars on individual windows allow access to overflow text.**

block of text. Think of *Info Select* as an electronic collection point for all of those unstructured scraps of paper bearing important notes that clutter your physical desktop. Right now, my desktop includes a publishing schedule, a reminder to buy some antenna-support line, a note to call N1JHF regarding some RF modules, and a look-for list of a dozen or so components. In free moments, I type these sticky-backed bits of paper into *Info Select* windows.

While *Info Select* doesn't impose a data structure, you can create predefined forms (e.g., a mailing label for QSL cards) to contain data. For general notetaking, you need not specify a format—simply type data into a window in free-form fashion.

For instance, you can use *Info Select* as a QSO logger. Each QSO occupies a window, but you're not limited to fitting callsign, signal report and QTH into specific places as many QSO loggers demand. Instead, you enter data in whatever order it arrives. Beyond the callsign exchange, no two hams carry on a QSO in exactly the same way.

If you're using a Windows-based callbook (e.g., Walnut Creek 's CD-ROM based *QRZ II*), you can look up a callsign and use Windows' copy-and-paste capabilities to capture a mailing address from *QRZ* and insert it into an *Info Select* window. Once you've added several windows to your QSO stack, you can unleash *Info Select's* powerful "neural" search engine to locate, say, all callsigns containing a W1 prefix.

Onscreen, a display field represents each window in the current stack as a dot. As you type each character in your search request, you'll note that the number of dots diminishes as the search engine sets aside extraneous records. You can specify wild-card characters if you're unsure of a callsign's inner characters.

After a few keystrokes, you're left with a clutch of "filtered" windows that either match, or come close to matching, your search request. You can step through the filter's output until you find the callsign you're looking for.

Other uses for *Info Select*—and PIMs in general—include keeping track of your electronic components inventory, logging phone calls, and scheduling appointments. **Figure 1** shows a few of the 338 windows currently forming my amateur radio stack on *Info Select*.

Like all database programs, *Info Select* requires self-discipline for best results. If you make an effort to enter data, the program's value grows as its contents increase. Fortunately, *Info Select's* designers have made data entry relatively easy. On the downside, you do need to keep a PC nearby and under power to make maximum use of *Info Select* or any PIM. The usual caveats about making regular backup copies of your PIM's database apply, too.

If you're unsure of whether a PIM will work for you, Micro Logic offers *Info Select Express*, a full-featured version that's limited to a single

55,000-character data stack (approximately 1,000 windows), for $19.95—a bargain if your needs are modest.

Capable of storing a maximum of 32 million characters, the full-featured version of *Info Select for Windows* sports a list price of $149 and is available as a competitive upgrade for $99—street prices are considerably lower.

Browsers

After a while, your PC's hard disk drive shares a certain resemblance to the HF radio spectrum as monitored by a general-coverage SW receiver. Tune across most bands, and you'll hear a variety of oddly modulated signals—FAX, AMTOR, and RTTY—that you'd like to decipher, but may lack the specialized equipment to do so. Likewise, your PC's hard disk drive will gradually accumulate a mixture of icons, images, sound clips, and text files in dozens of unfamiliar formats. If you own software capable of creating a given format, you can open and examine any file of the same format—just as your sense of hearing accommodates demodulated audio and CW transmissions.

However, if you don't own the appropriate software, you're in the same position as you are when you listen to a RTTY signal—you can't recover any information using the resources at hand. Also, while many programs such as word processors will translate a few competitors' formats, chances are the format you need is the one that's omitted.

The obvious approach to viewing files—obtaining a shareware or commercial viewer for each specialized format—increases clutter on your hard disk and becomes expensive after a while. Software designers have responded to file-format proliferation by creating software viewers or browsers capable of opening and displaying files of many formats.

As **Table 1** shows, *Advanced Turbo Browser* by Pacific Gold Coast Corp. takes several steps toward achieving the goal of a truly universal file-examination toolkit. Because prospective users may lack familiarity with file extensions, *Turbo Browser* lists many formats by name. Selecting a format "filters" extraneous files from directory listings for easier viewing.

To simplify file selection, *Turbo Browser* offers a powerful range of file-finding sorting options. You can sort by name, size, file type, or creation date. You can locate files by searching for creation dates and file sizes and file contents. In the latter mode, you can order a "fuzzy search" that looks for an exact match and then searches for a filename or text string that's in error by one or more characters.

You can preview files in several ways. Clicking on a filename with the mouse's right button displays the file's contents in the pre-

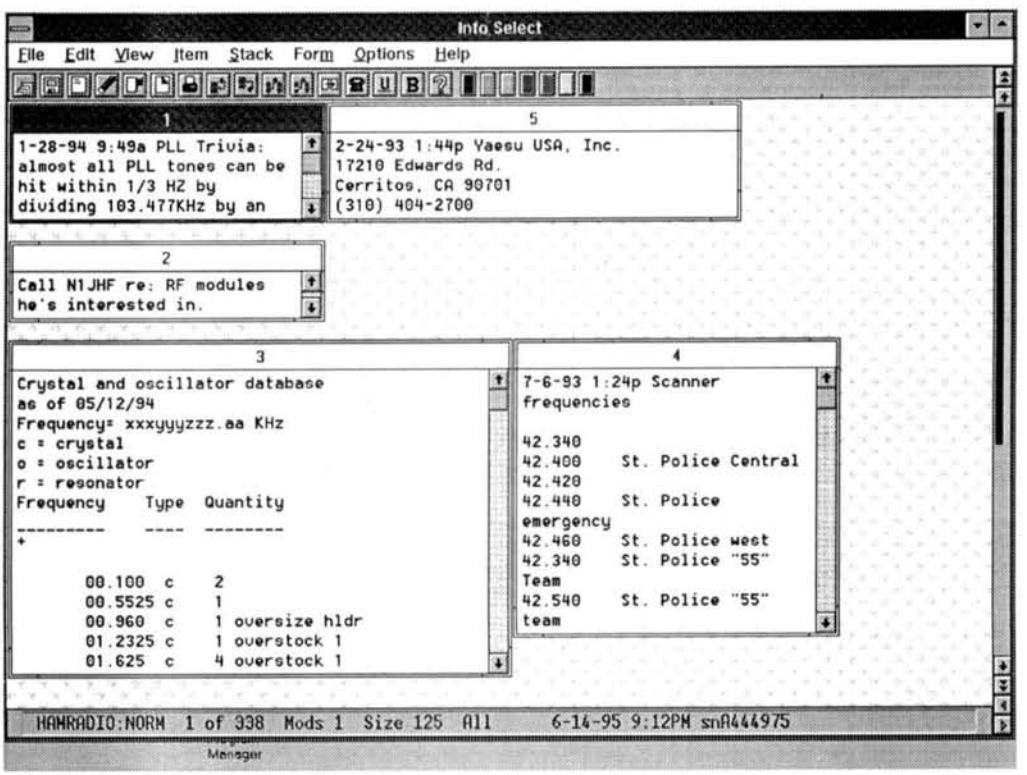

Figure 2. ***Advanced Turbo Browser*** **displays a list of files on Walnut Creek's *QRZ* amateur radio CD-ROM. The center panel shows a portion of the selected file—a JPEG-encoded schematic of the BayCom packet-radio modem as implemented for the Amiga PC.**

Browser Filter Categories	File Extension
General Subcategory	
Windows Icon	*.ICO
Windows Clipboard	*.CLP
Windows Metafile	*.WMF
Computer Graphics Metafile	*.CGM
Encapsulated PostScript	*.EPS
WordPerfect Graphics	*.WPG
Text-file Subcategory	
ASCII Text	*.TXT
AMI Pro	*.SAM
Microsoft Word	*.DOC
WordPerfect	*.WP?
Windows Write	*.WRI
PostScript	*.PS
Image-file Subcategory	
Windows Bitmap	*.BMP
Color Wheel	*.CWD
Graphic Interchange Format	*.GIF
Joint Photo Experts Group (JPEG)	*.JPG
Kodak PhotoCD	*.PCD
PCX	*.PCX
PICT	*.PIC
Targa	*.TGA
TIFF	*.TIF
Multimedia Subcategory	
Video For Windows	*.AVI
Animator Pro	*.FL?
QuickTime For Windows	*.MOV
Wave Sound	*.WAV
MIDI Sound	*.MID
Spreadsheet Subcategory	
Lotus 1-2-3	*.WK?
Microsoft Excel	*.XLS
Borland Quattro Pro	*.WB?

Table 1. The Advanced Turbo Browser.

view window. Files unrecognized by any of *Turbo Browser's* routines appear as ASCII characters, or in hexadecimal format. You can also select a group of files and use the mouse pointer to drag-and-drop the group into any of *Turbo Browser's* six queues. These predefined file folders allow you to copy or export files to another disk drive or directory, dispatch files to the Windows Clipboard, convert, print, view, and play sound files. When viewing files, you can select a "slide show" mode that sequences through the viewing queue.

Figure 2 shows a graphics file selection within a file directory (left panel), a section of the file in a preview window (center panel), and a list of predefined file-manipulation queues (right panel).

In addition to file display, *Turbo Browser* lets you create custom folders and program groups containing files or Windows desktop accessory programs that you use frequently. You can also reduce keystrokes by assigning "shortcuts" to directories or deeply buried subdirectories.

From a preview list, you can launch files by double-clicking the left mouse button while pointing at a filename. Note that while *Turbo Browser* can view files, it's up to you to provide application software that actually uses a given filetype. If *Turbo Browser* doesn't find a program that recognizes the file's extension, it allows you to select an alternative application.

Once you have viewed an image, you can convert it into an alternative graphic format or print the image either via *Turbo Browser* or through a selected application. You can also convert a word-processed document into plain ASCII text; this is a big bonus if you lack the original application.

A few rough spots surfaced during my review. If you exit from the program and save program settings without leaving Program, Access Control, or File Control windows active, you'll view a mysterious gray screen when you next activate *Turbo Browser*. My call to Pacific Gold Coast's tech support line was returned promptly along with an explanation.

Also, if you import GIF files created on a Macintosh, *Turbo Browser* won't allow you to view the files until you remove the GIF87a or GIF89a headers the Macintosh embeds in the files. You can use Dave Seah's excellent STRIPGIF freeware utility (available on most BBSes) to remove the headers and write PC-viewable GIF files.

I've omitted descriptions of several of *Turbo Browser's* many tricks. There's much more to explore within this complex program, including an audio CD player. The program includes an extensive on-line help file, and the 114-page printed manual covers topics in greater depth, but would benefit from editing to remove the passive voice.

Will *Advanced Turbo Browser* help you manage your PC's burgeoning file collection? Yes, and at a list price of $89.95 it's a good value; but, be prepared to spend an evening or two learning the program's extensive collection of features.

To order *Info Select for Windows*, contact Micro Logic at P.O. Box 70, Hackensack, NJ 07602, by phone at(800) 342-5930 or (201) 342-6518, or by fax at (201) 342-0370.

You can obtain a copy of *Advanced Turbo Browser* from Pacific Gold Coast Corp., 15 Glen St., Glen Cove, New York 11542, or call (516) 759-3011. You can also contact the company via fax at (516) 759-3014, or America On-line (Go Browser), CompuServe (102023,636), or the Internet (gobrowser@aol.com). ■

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INSTRUMENTS FOR ANTENNA DEVELOPMENT AND MAINTENANCE

Part 2: Signal generators

As I mentioned in Part 1 of this series, the work of developing and testing a new antenna is reduced significantly if you have the right tools. Let's take a look at the signal generator, with an eye to how to choose and use one.

Signal generators

Figure 1 shows the basic elements of a signal generator—a useful piece of test equipment that produces a voltage of known level at a known frequency. This frequency accuracy depends on the oscillator calibration, and on drift. An external digital meter is necessary for precise work. Amplitude accuracy depends on the accuracy of the RF voltmeter and on the attenuator. A good design provides both the pre- and post-attenuator outputs, typically 1.0 and zero to 0.1 volt. Frequency swept generators use a low-frequency form of FM modulation, linear versus time, or sine wave—both typically synchronized to the power line frequency. Power line filtering, or the double shielding of the oscillator as in the best designs, is not shown here.

As **Figure 1** illustrates, signal generators are really quite simple. As was discussed in more detail above, the key elements include an oscillator for the desired frequency, some method of indicating the signal level, and an attenuator for setting the level.

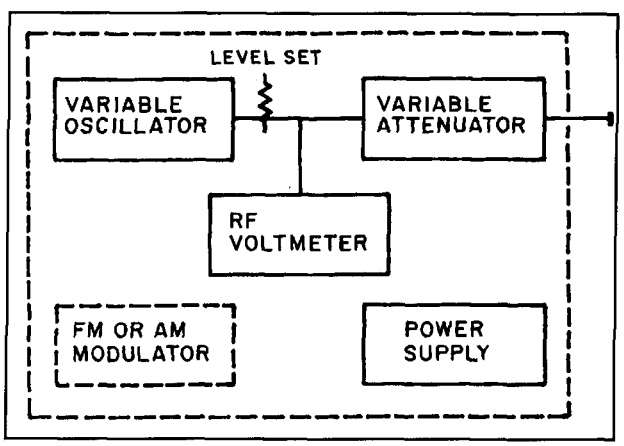
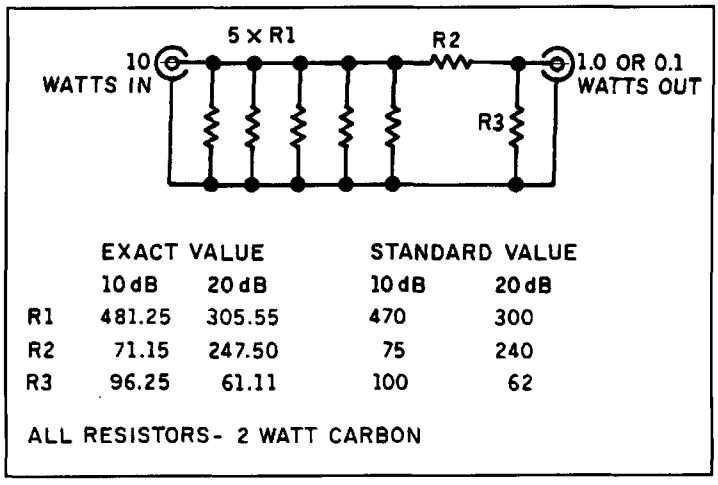


Figure 1. Elements of a signal generator.

Most station transmitters contain at least two of these features, if not all three, and have the great advantage of being readily available. Also, your station transmitter is likely to be the continued terminal of your antenna system. For this reason, the bulk of amateur antenna work makes use of the station transmitter.

However, despite the convenience, a transmitter has its drawbacks. First, a good number of units won't cover an adequate frequency



	EXACT VALUE		STANDARD VALUE	
	10 dB	20 dB	10 dB	20 dB
R1	481.25	305.55	470	300
R2	71.15	247.50	75	240
R3	96.25	61.11	100	62

ALL RESISTORS- 2 WATT CARBON

Figure 2. Transmitter power reducer.

range. As a result, it's often difficult to set the output power to the low level that's needed for many measurements and required to prevent unnecessary interference. Second, solid-state transmitters don't like the widely varying loads inherent in much antenna work. Finally, it's often inconvenient to move the transmitter from its station location to the work area.

Some of these problems can be avoided by using a power attenuator between the transmitter and the antenna. This device decreases transmitter levels for antenna measurements. The design in **Figure 2** shows components of two values. These are intended to reduce a level of 10 watts input to either 1 or 0.1 watt output to a matched load. I chose 10 watts because it's a level found on most power meters.

The attenuator can be mounted with the resistors in oil, or can be placed in a perforated metal case for good air circulation. Such a unit will reduce interference, keep solid-state transmitters happy, and provide a source of known impedance.

Frankly, it's rarely worthwhile to build a signal generator. With a little luck, you can find low-cost used units at hamfests. When searching for a suitable device, don't forget to look at the older tube-type generators. While they may need some replacement capacitors and a good cleaning, their ruggedness provides adequate compensation for the work you'll put into them.

There are some wide-range new units available at reasonable cost. Look for a constant level output terminal at about 1 volt, to connect to a digital frequency meter. While signal generators have calibrated dials, the resolution is usually not sufficient for productive antenna work. In addition, the output of many generators is low, so a simple instrument such as a SWR meter may not work. *The ARRL Handbook* describes amplifiers that can raise the power to necessary levels. Don't omit the output filtering of these amplifiers, because harmonics affect the accuracy of many measurements.

If you must homebrew a generator, **Figure 3** shows a simple design that offers fair performance. The unit is a single-IC generator for RF work; values for 14 MHz are shown. The coil must have good Q, and an all crystal design is easier to build. Output may be increased by replacing the three parallel open collector inverters with a high-voltage type, the 50-ohm resistor being fed from 12 to 25 volts.

This homebrew generator is intended for use over a single band, and provides several fixed crystal frequencies in addition to variable frequency. Because the transistor output is a square wave (or nearly so), the output filter is a necessity.

The units discussed above are basically single frequency generators. Spread frequency generators are also useful in antenna work. The noise

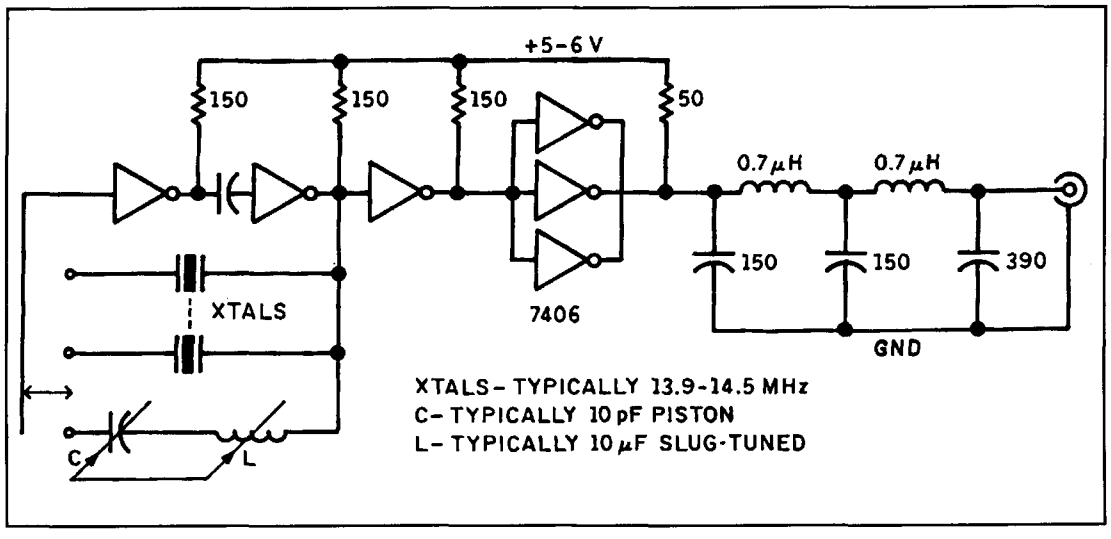


Figure 3. Simple signal generator.

generator and swept frequency generator are two common types. While the single frequency system can use broadband detectors like a simple RF voltmeter (except for the harmonic error problem), the broadband type must use a frequency selective detector/indicator. The most common of these is the station receiver.

The circuit for a simple noise generator appears in **Figure 4**. This circuit assumes a DC current path through the external circuit. If not present, a resistor equal to the line impedance may be shunted across the output terminals. While the circuit is intended for relative measurement, the noise output can be calibrated by comparing it to the signal from a good RF signal generator.

Precision diodes are occasionally available at hamfest tables. The Microwave 1N21 diode shown isn't the only kind that can be used, but it has high output. The other common type is the low-voltage Zener diode. Low-frequency coverage is set by the size of the coupling capacitor and the isolating choke: the values shown are good for all HF bands. The upper limit is determined by the stray inductance of the coupling capacitor and stray capacitance to ground. With normal components and open construction, usefulness extends through the UHF. In use, the noise generator replaces the oscillator and the selective voltmeter (receiver) replaces the detector in **Figure 1**.

Early swept-frequency oscillators used a mechanically varied capacitor to cover the swept range. World War II versions are sometimes seen at hamfests. Current designs use a

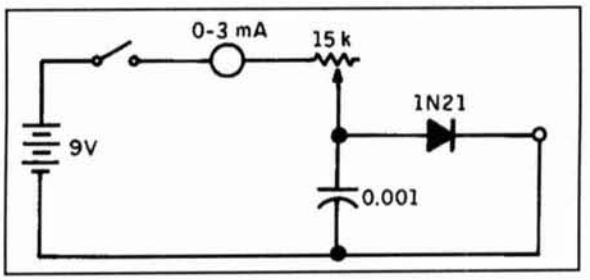


Figure 4. Diode noise generator.

voltage variable capacitor—a diode designed for high inter-element capacitance. Generators usually provide an output voltage synchronized to the instantaneous frequency, but a few supply only a trigger. Either is used to control the detector indicator, a CRT being common. The detector itself must be broadband; for example, a simple RF voltmeter, or a SWR meter. The generators are often found at hamfests, disguised as television sweep generators.

Conclusion

As I've shown here, the signal generator can be a very valuable part of an antenna workshop. However, one needs not build such a device from scratch, as many used generators are available at hamfests.

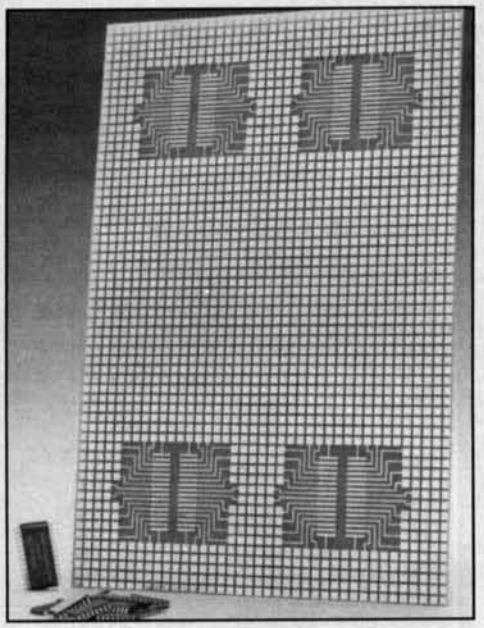
In my next installment of "Instruments for Antenna Development and Maintenance," I'll look as SWR meters and some precision measurements to replace SWR. ■

PRODUCT INFORMATION

New Surface Mount Prototype Board

ECode Systems, Inc. has introduced Surfboard; a unique surface mount prototype board that offers variable size SOIC pads for design and prototyping flexibility. The pad layout allows mounting of both wide (400 mil) and narrow (300 mil) body SIOC styles while accommodating up to 32 pins. The 100 mil matrix of 80 mil squares is specifically designed for easy mounting of chip resistors, capacitors, inductors, SOT style semiconductors and even DIP packages.

Surfboard is manufactured in the USA and utilizes components such as FR4 (fiberglass epoxy laminate) and incorporates solder plating over 1 oz. copper on both sides of the board. Prices are as follows: 1 @ $19.95/ea.; 2–24 @ $16.95/ea.; and 25 or more @ $14.95/ea. For more information, contact ECode Systems, Inc., 7050 North Wilder Road, Phoenix, AZ 85021; phone 602-870-8063; fax 602-371-8736.



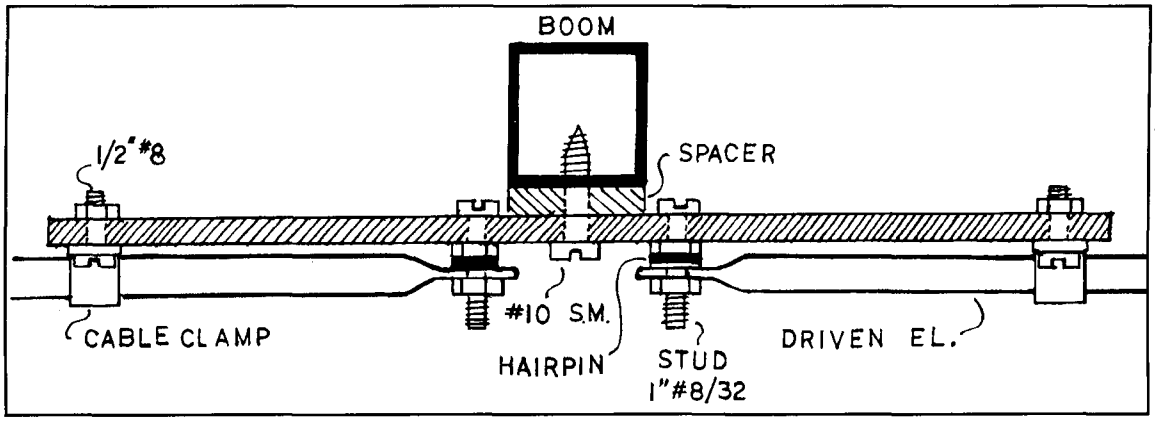


Figure 4. Driven element assembly.

dle and mounted on an insulated center block. For this, cut two identical lengths of tubing measuring 54-3/4 inches. Use a vice to flatten one end of each section back 1/2 inch, then drill a 3/16-inch mounting hole 1/4 inch from the end.

When mounting elements, install the reflector closest to the U-bolt mounting holes (the back of the antenna). Center the tubing in the hole in the boom and hold it firmly in place. Using a 1/8-inch bit, drill through the boom and element. Pin the reflector in place with a 1/2 inch long no. 6 screw and secure with a nut (use **ONLY** stainless-steel hardware—other types will corrode or rust quickly when exposed to weather). Tighten firmly, but avoid crushing the tubing. Repeat this procedure to install the director.

Cut the driven-element center block from 3/16 or 1/4-inch Plexiglas™ (many glass-replacement shops carry this material, and most save remnants). *Do not* substitute a plastic with unknown RF insulating properties! Make the block's dimensions 2-1/4 by 10 inches. Drill the specified mounting-hole pattern using a 3/16-inch bit, as shown in **Figure 2**. Also, cut two 1/2 by 1-inch Plexiglas™ spacers and drill a 3/16-inch hole in the center of each.

To install the center block, position it at the exact center of the boom. Using the block as a pattern, drill two 7/64-inch pilot holes on the bottom surface of the boom—on centerline—to correspond with the block's mounting holes. Then, secure the block and its spacers to the boom with 3/4-inch no. 10 sheet metal screws. Make sure it seats at a right angle to the boom line before tightening. Finally, install 1-inch no. 8-32 element mounting studs in the block, as shown in the driven element assembly.

The antenna's hairpin inductor is made from 1/8-inch brass rod, as illustrated in **Figure 3**. Cut the rod to length, so the completed hairpin measures 1-1/2 inches in width and 5-3/8 inches in length (measure from the spade-lug mounting centers to the end of the loop). Crimp and solder the lugs in place to ensure good

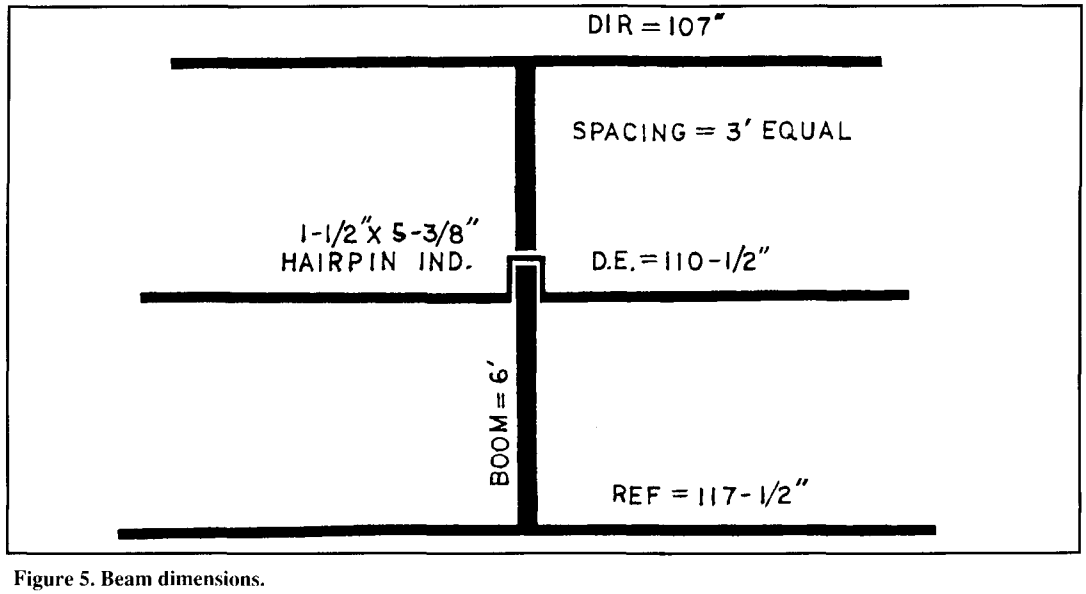


Figure 5. Beam dimensions.

electrical contact. Solder a 2-inch bus wire to form a center tap on the hairpin; you'll connect this to the boom later on.

Begin final assembly by installing the hairpin on the driven-element mounting studs (see **Figure 4**). Use a flat washer above and below each spade lug, and position the loop-portion of the hairpin forward—toward the director end of the antenna. Next, install the two driven element sections on top of the hairpin and hold them in place with no. 8 nuts turned finger tight. To secure element sections more permanently, install a 5/16-inch (1/4 inch will do) plastic cable clamp on each, and fasten these to the center block with no. 8 hardware. (If you use 1/4-inch cable clamps, you'll need to fatten the tubing by winding a few turns of electrical tape under the clamp.) Confirm that the driven element forms a 90-degree angle to the boom, then tighten all hardware. If further alignment is required, gently bend the affected element as needed. **Figure 5** shows dimensions.

To complete assembly, drill a small pilot hole in the boom—on centerline—at the loop end of the hairpin. Install a no. 6 sheet metal screw and use it to bond the hairpin's center tap to the boom. Position the hairpin parallel with the boom, keep the center tap lead short as possible, and remove any excess bus wire.

Feedline and Balun

For coax runs up to 60 feet, RG-8/mini is probably the most practical choice for low-power operation at 50-MHz. The antenna's feedpoint is balanced, so you'll need to add a balun to decouple the feedline from the driven element. For simplicity, I fabricated a W2DU—or Maxwell—style choke balun using ferrite sleeves. Other balun types may be slightly more effective in critical applications, but the choke balun works well and represents the most practical solution for a project of this scope. For RG-8/mini feedline or RG-59 stacking harness, slip three FB-43-5621s over the cable and secure them at the feedpoint with tape. For RG-58, install three FB-43-6301s in the same manner. Ferrite sleeves are available via mail order from Amidon Associates at (310)763-5770.

Connect the coax directly to the element mounting studs using no. 8 lugs installed on 3/4-inch pigtails. For weather protection, apply two or three heavy coats of contact cement to exposed coax braid. When you install the antenna on its mast, tape the feedline securely to the mast below the balun sleeves to provide stress relief. Also, use TV mounting hardware with boom-reinforcement plates to protect the aluminum boom from being crushed during mounting. (Consumer-grade TV antennas also use 1-inch square boom stock, so reinforced TV hardware will work extremely well.)

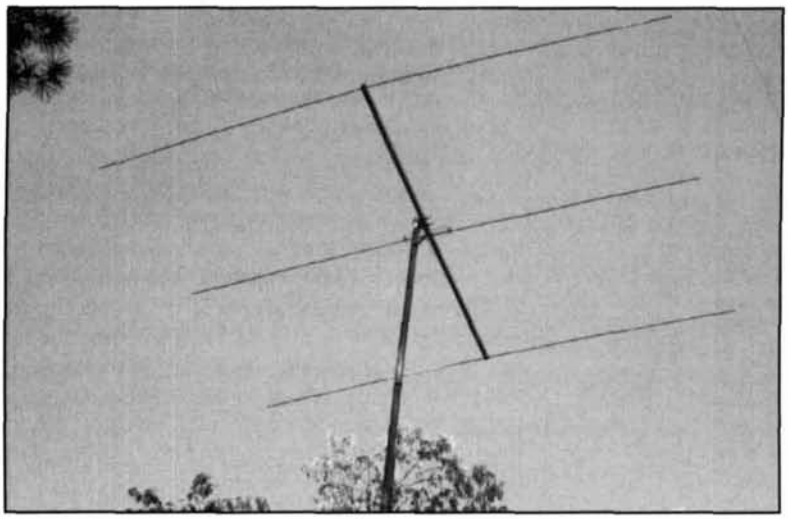


Photo A. Stacked featherweight Yagis on light-weight aluminum mast at my QTH.

Performance

I used a MFJ-259 Antenna Analyzer to measure VSWR during initial adjustments to the prototype. The driven element and hairpin tuned without difficulty, and I set F_R (resonant frequency) at 50.250—slightly above the SSB calling frequency. I did this for a couple of reasons. First, best gain and front-to-back typically occur slightly below F_R. Second, when water collects on the elements, F_R tends to migrate downward (better to have resonance move toward the SSB operating range rather than away). When optimized and permanently mounted 22-feet above the roof, the antenna's VSWR was flat at 50.250 and well under 1.2:1 at 50.100. The elements aren't adjustable; however, if you follow the prescribed measurements during construction, no pruning should be necessary.

Once the antenna was mounted in place, I performed tests for front-to-back ratio and gain using what limited resources I had available. To check front-to-back ratio, I installed a step attenuator at my receiver and placed a low-power 50.125-MHz signal source several hundred yards from the mast. Using the attenuator to match S-meter readings, I rotated the antenna to compare front and back sensitivity. Backlobe rejection measured 15 dB, with much deeper sidelobes.

To check forward gain, I installed a reference dipole below the Yagi on the same mast—then used my attenuator and signal source to compare the two. The Yagi provided a 7 to 8 dB advantage, but this shouldn't be interpreted as the antenna's forward gain in dBd. During these tests, the reference dipole was only 8-feet above the roof, where unwanted reflections could easily skew the results. In the real world, 6 dBd gain is probably tops for a 3-element Yagi on a 6-foot boom.

From a subjective standpoint, the project was

a total success. After operating for several months with one featherweight Yagi, I now have two stacked twelve feet apart on a light-weight aluminum mast (see **Photo A**). Despite my marginal VHF location, they're great performers—allowing me to enjoy everything the band has to offer! Best of all, at less than two pounds apiece, I don't worry about heavy antennas crashing through our roof when the snow flies and the "Montreal Express" rips through New England on those cold winter nights.

For those of you who already have a large antenna farm, or don't like to work 6 meters, we have a different kind of project. Jay Craswell, WBØVNE, took a Collins Military radio and modified it for use on the 80 and 40-meter bands.

Get On the Air With a "Cheap" Collins Rig

Jay Craswell, WBØVNE/AAV5TH
Previously ran in Electric Radio, January 1995

I'm not quite sure how I got involved in the conversion project for the PRC-47. Suffice it to say the "magic" of Collins radio was not lost on me. However, the cost of collecting Collins gear put this particular project in the realm of daydream. However one day I was talking with a friend about radio, and he mentioned that he has a lot of fun with military surplus stuff. One piece that caught my attention was an HF SSB rig for 2 to 12 MHz SSB made by Collins. It sounded interesting, so I did some research and found that several places had them for sale. Prices varied wildly. The highest was over $200 and the cheapest, from TARTAN Electronics, was only $49.95.

I got talking with Gene the owner and he confided to me, "I musta bought 10 zillion pounds of 'em!" (All gone by now, I believe.) I kidded him about that. "Ten zillion pounds, eh? How many radios is that? Five or six?" Fortunately, Gene is a good-natured guy and we got to be good friends.

When we turned our attention to the conversion project, the problems, as far as we could see, were: (a) the rig was USB only, (b) it was 28 volts DC or 115 volts AC 400 cycle, and (c) the only way to connect an antenna was through a weird screw-in whip antenna. I ordered a pair of radios and dissected the schematic. The power supply wasn't a problem, as I had access to an old computer power supply with the proper output. The antenna (to me) seemed to be a case for some machine shop time. The only problem that remained was the restriction to USB. Although USB is fine if you're 100 percent into MARS or CAP, it's not really great for hamming the 80 and 40-meter phone bands.

How it works

The PRC-47 uses a mechanical filter scheme. Technically, swapping the USB mechanical fil-

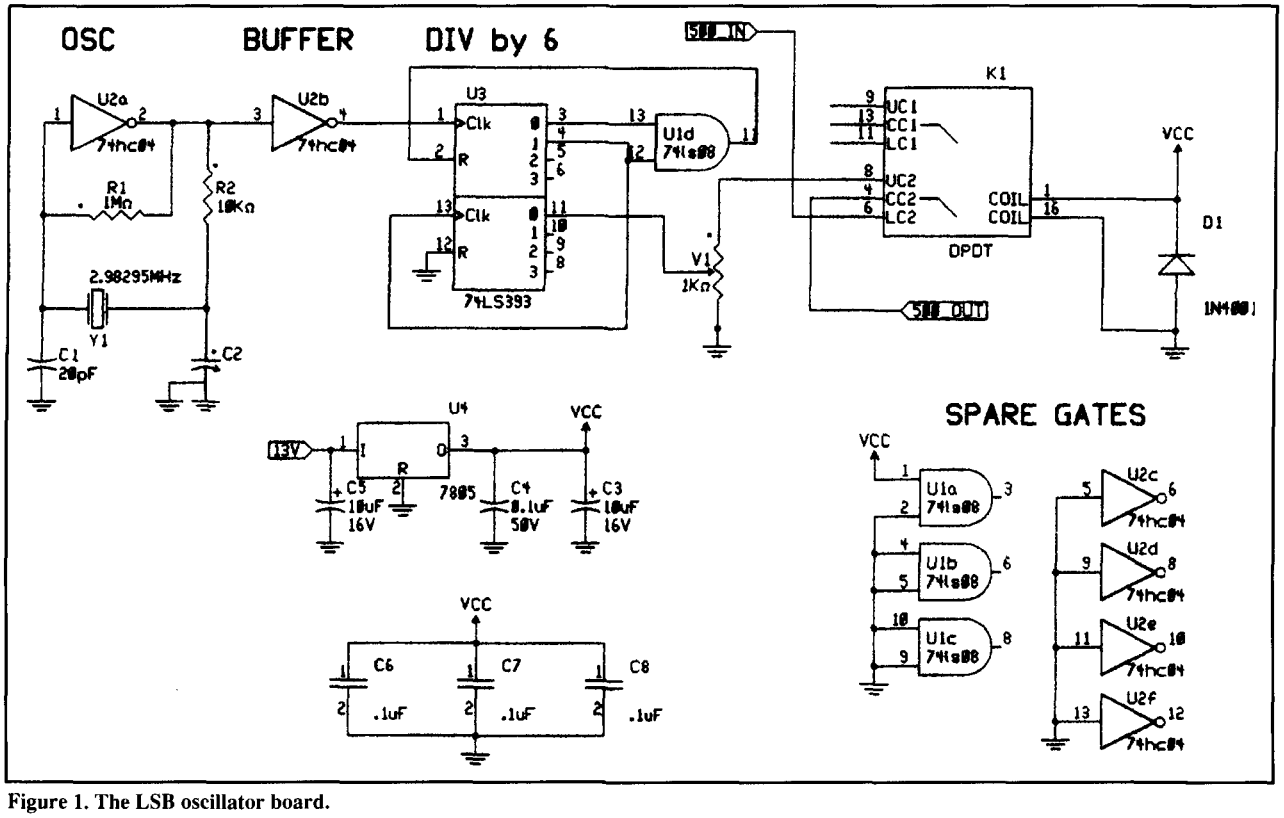


Figure 1. The LSB oscillator board.

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SL-11R	•	•	7	11	2⅝ × 7 × 9¾	12
SL-11S	•	•	7	11	2⅝ × 7⅝ × 9¾	12
SL-11R-RA		•	7	11	4¾ × 7 × 9¾	13

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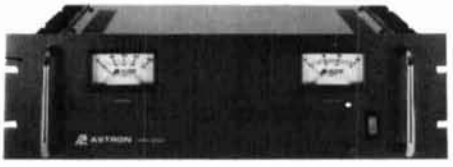
RS-4L	3	4	3½ × 6⅛ × 7¼	6
RS-5L	4	5	3½ × 6⅛ × 7¼	7

RM SERIES



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RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
RM-60A	50	55	7 × 19 × 12½	60
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50
RM-60M	50	55	7 × 19 × 12½	60

RS-A SERIES

MODEL RS-7A

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A		•	2.5	3	3 × 4¾ × 5¾	4
RS-4A	•	•	3	4	3¾ × 6½ × 9	5
RS-5A		•	4	5	3½ × 6⅛ × 7¼	7
RS-7A	•	•	5	7	3¾ × 6½ × 9	9
RS-7B	•	•	5	7	4 × 7½ × 10¾	10
RS-10A	•	•	7.5	10	4 × 7½ × 10¾	11
RS-12A	•	•	9	12	4½ × 8 × 9	13
RS-12B		•	9	12	4 × 7½ × 10¾	13
RS-20A	•	•	16	20	5 × 9 × 10½	18
RS-35A	•	•	25	35	5 × 11 × 11	27
RS-50A	•		37	50	6 × 13¾ × 11	46
RS-70A	•		57	70	6 × 13¾ × 12⅛	48

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46
RS-70M	57	70	6 × 13¾ × 12⅛	48

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MODEL VS-35M

- Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES

MODEL RS-12S

- Built in speaker

MODEL	Colors Gray	Colors Black	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	•	•	5	7	4 × 7½ × 10¾	10
RS-10S	•	•	7.5	10	4 × 7½ × 10¾	12
RS-12S	•	•	9	12	4½ × 8 × 9	13
RS-20S	•	•	16	20	5 × 9 × 10½	18
SL-11S	•	•	7	11	2¾ x 7⅝ x 9¾	12

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

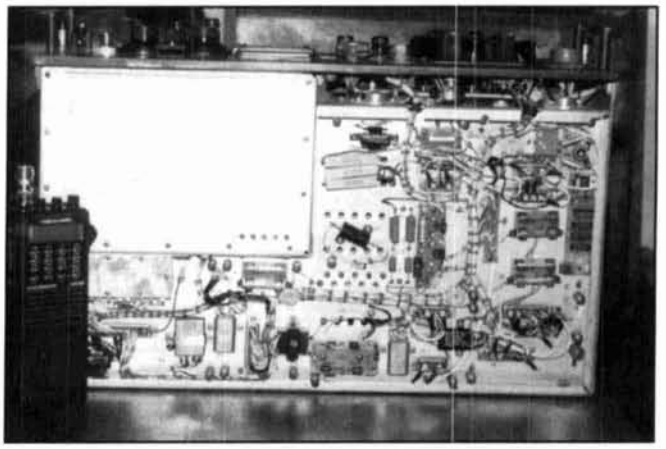

Photo A. Bottom view. The LSB board is on the middle right, next to the fuse blocks. (Photo by WB0VNE.)

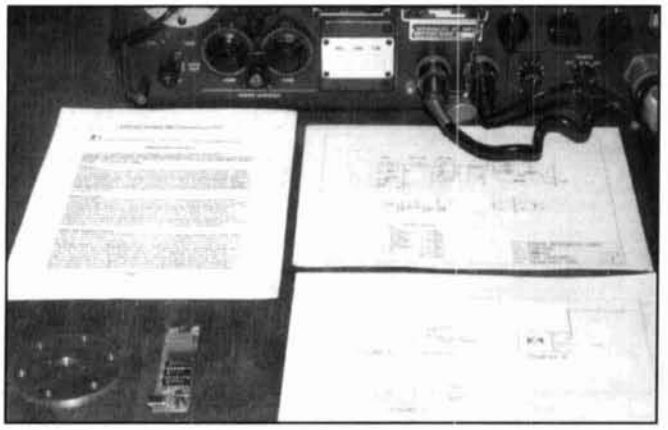

Photo B. The PRC-47 kit includes the new antenna panel, which uses the SO-239 connector. (Photo by WB0VNE.)

ter for a LSB filter would work, but the cost of a Collins 500-kHz LSB filter made it prohibitive. Besides, I wanted USB and LSB! I decided to inject a different carrier frequency into the balanced modulator and make the USB filter function as a LSB filter!

The radio normally uses a 500-kHz oscillator frequency. I did some experimenting with my DOVER DigiVFO (0 to 20 MHz) card and tried different frequencies to see how the audio sounded at different points. I found that the LSB sounded pretty good with an injection frequency of 497 kHz. Once I determined this, I went to the junkbox to see what parts I had on hand to make a more permanent gadget that would fit inside the radio. Crystals for 497 MHz are rare and expensive. I decided to start with a higher frequency and divide down with some TTL parts (see **Figure 1**). The 74HC04 runs the crystal up to TTL levels. The 74LS393 and 74LS08 provide the desired divide by 6 needed with the crystals I could get my hands on. My local parts supplier had 2.983-MHz crystals (close enough) available and I grabbed those. The pot allows you to adjust the output to match the normal 500-kHz input.

I laid out the pc board using the *CIRCAD for Hams* program. As a matter of fact, I drew all my schematics with this program. It sure beats the heck out of my old drafting table! Boards for the project are available as a kit, or as blank boards with documentation.*

The final board fits neatly inside the rig with some double-sided tape (**Photo A**). The relay and switch let you flip back and forth between USB and LSB. Instead of gouging a new hole in the radio, I unscrewed the battery test switch and inserted my USB/LSB switch. I taped up the battery test switch and tucked it behind the front panel. A local machine shop made new antenna panels, so I could use SO-239 connectors instead of the weird screw-on device (**Photo B**). Of course, everything is 100 percent restorable should I need to return the PRC-47 to its pristine condition.

On the air

With the standard handset and external speaker, I felt a little like Radar O'Reilly talking to his pal Skippy. My first contact was with Jerry, W9JIF, and Dave, W9OC, on 40-meter phone. Because of the new injection frequency, the dial was 3 kHz lower than the actual frequency (no big deal; just add 3!). We exchanged 5–9 signal reports, and surprisingly the audio with the HT33 handset was pretty good. I worked a number of other stations; a few commented that a faint 400-cycle power supply noise was audible. One fellow insisted I was in an airliner because of the noise! Another guy I talked to somehow knew I was on a Collins rig. My buddy Roger, WA0VLL, attributes this to the distinctive sound of SSB through a Collins mechanical filter.

Some miscellaneous notes and a little history on the PRC-47

Gene from TARTAN Electronics said the "Fightin' Marines" were the last owners of these rigs, so you know they saw some action. All of this gear is Vietnam era stuff, so it's not ancient history. In fact, the expensive crystal mixing scheme is still pretty fantastic with respect to today's rigs and all the accompanying phase noise and CPU hash. Let's not forget that the magical Collins is still unbeatable!

Some of these units may have salt water damage. Gene had a few such rigs that he cured by tossing them out. If you purchase some of these radios, be sure to request units *without water damage*.

All in all, this was a fun project. I'm proud to say that one of the rigs is now being used by the Farmington, Minnesota Civil Air Patrol for its HF communications. ■

*Kits are available as follows: USB/LSB Semi Kit (with antenna panel), $49; USB/LSB Raw Board, $19; Antenna Panel, $9. Include $3.00 for 2-day air shipping. Send orders to: Jay Craswell, 321 West 4th Street, Jordan, Minnesota 55352-1313. Phone: (612)492-3913.

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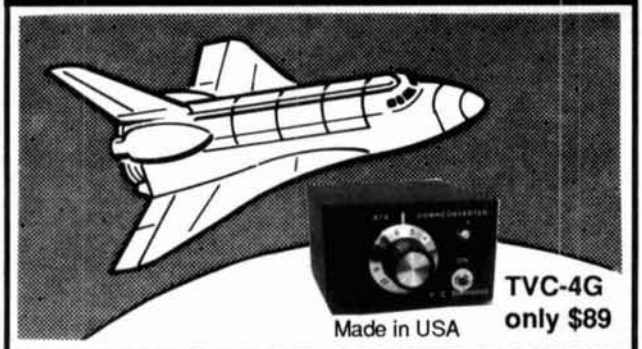

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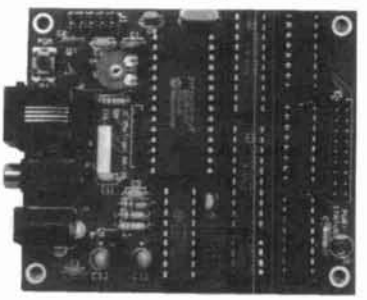

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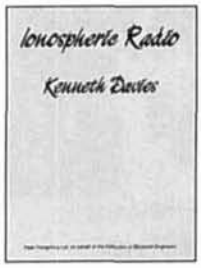

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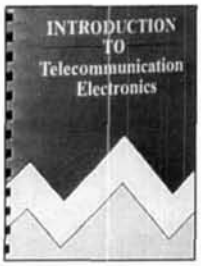

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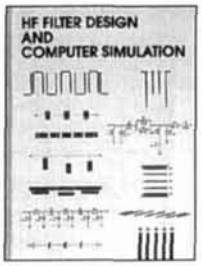

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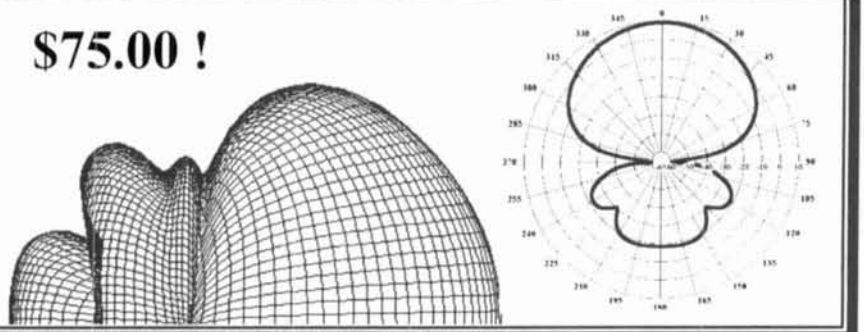

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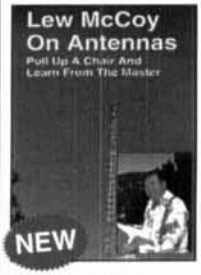

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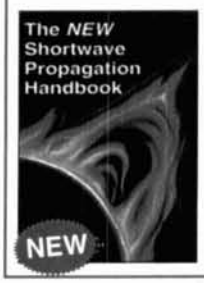

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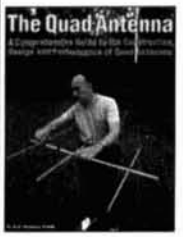

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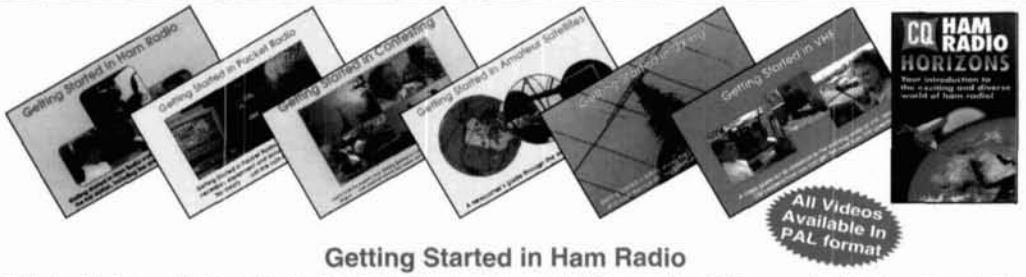

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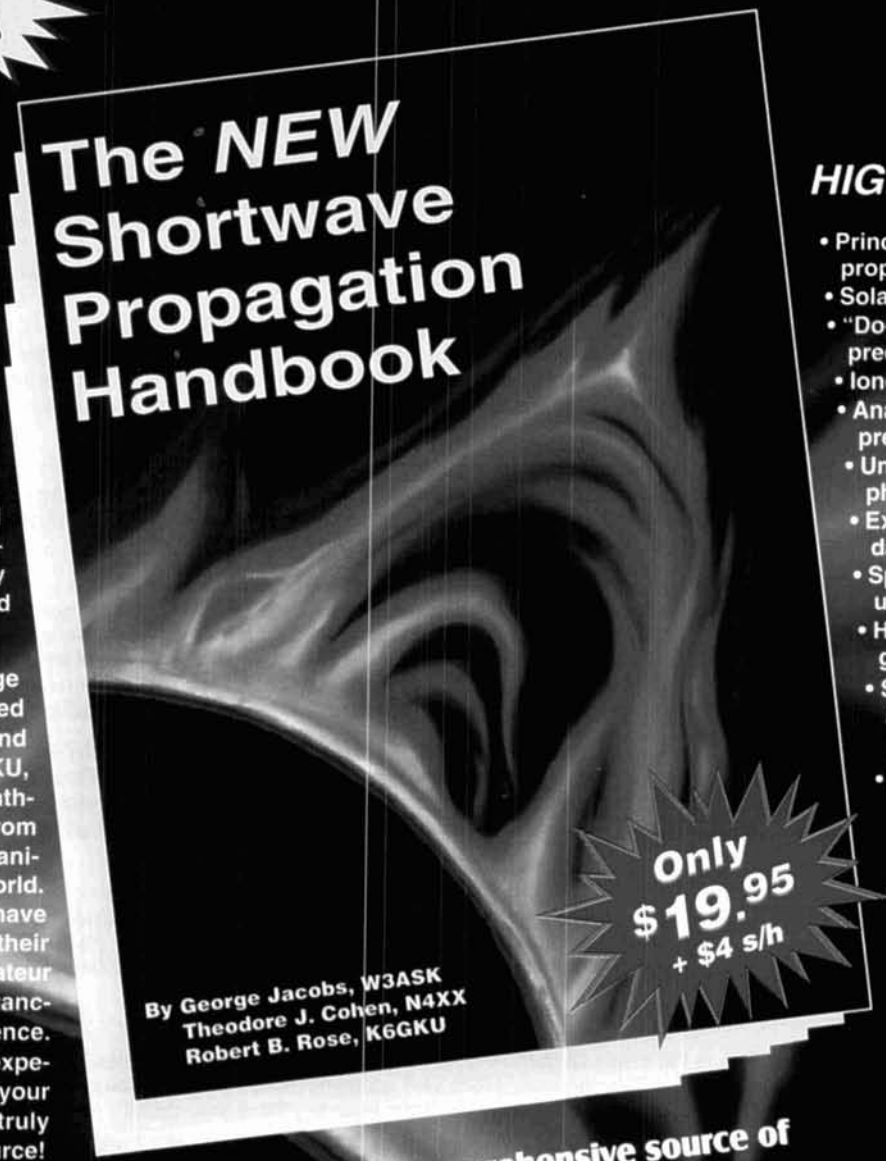

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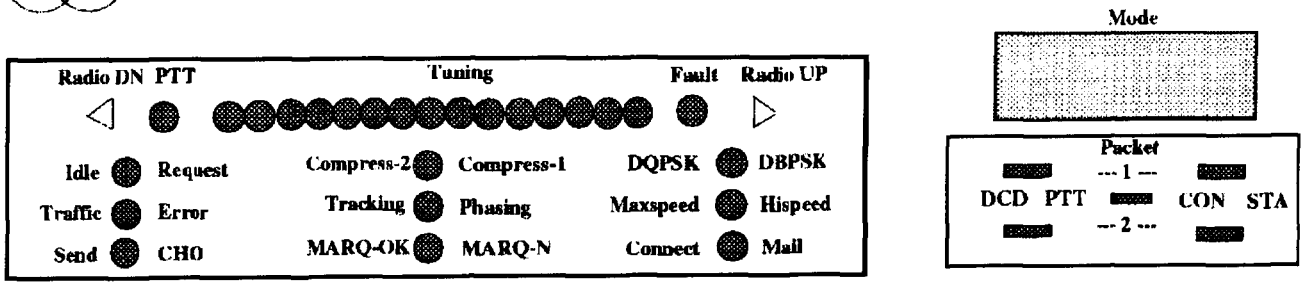
PacComm PTC-II Multimode DSP Communication Controller



You've been hearing about it for several years.... Now it is available. PACTOR-II.

The PTC-II is a new multi-mode controller and "communications platform" which contains very powerful and flexible hardware and firmware.

It is being built in the United States by PacComm under license from S.C.S., the group that developed both the original PACTOR and PACTOR-II.

The PTC-II offers the most robust HF digital protocol available to radio amateurs, but it should not be overlooked that the PTC-II is configurable as a triple-port multimode controller supporting packet data rates of 1200 and 9600 bps and numerous other modes.

What is PACTOR-II?

Like PACTOR-I, PACTOR-II is a communication system. There is a unique protocol, carefully designed and optimized for excellent HF performance, new and powerful modems realized via Digital Signal Processing, and a hardware platform to allow the firmware to realize its full operating potential.

PACTOR-II is the name of the new protocol including modulation specifications. PTC-II is the name of this specific hardware realization of the PACTOR-II architecture.

PACTOR-II offers all the following advantages:

- A step-synchronous ARQ protocol.
- Full support of memory ARQ.
- Independent of sideband; no mark/space convention.
- Center frequency adjustable between 400 and 2600 Hz to exactly match your radio's filters.
- Long-path capability for worldwide connectivity.
- Full compatibility with PACTOR-I (the original PACTOR system), AMTOR, and RTTY.
- All-mode mailbox with up to 32 megabytes of storage.
- Automatic switching between Level-1 (PACTOR-I) and Level-2 (PACTOR-II) at contact initiation.
- Occupies a bandwidth of under 500 Hz - use your 500 Hz CW filters.
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- Differential Phase Shift Keying with two continuously transmitted carriers. 100 symbols per second.
- DBPSK modulation yields 200 bps (uncompressed).
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- Powerful Forward Error Correction (FEC):
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  - » Viterbi decoding using soft decision point. Coding rate varies between 1/2 and 7/8.
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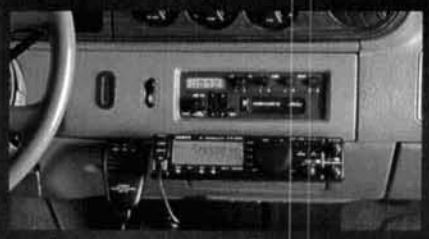
As a base station, the compact full function FT-900AT includes direct keypad entry for pinpoint accuracy during quick band/frequency changes. Other features you'll like include CW keyer with front panel speed adjustment, speech processor, twin stacking VFOs, and IF Shift and Notch. No competitor offers this! Bonuses, such as signal strength, power output, SWR and ALC digital meters, add value to the FT-900AT, and the proven duct-flow cooling system provides excellent long-term transmit power output reliability and frequency stability. For ease of use, Yaesu's exclusive Omni-Glow display enhances viewing in any light condition. And, since the high speed antenna tuner is built-in, it means less clutter in your shack.



The FT-900AT controls mount almost anywhere in your car, truck or camper. 100 Watt RF deck can mount in trunk, or under seat.

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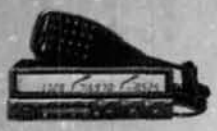

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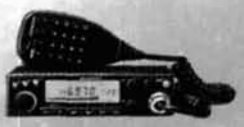

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